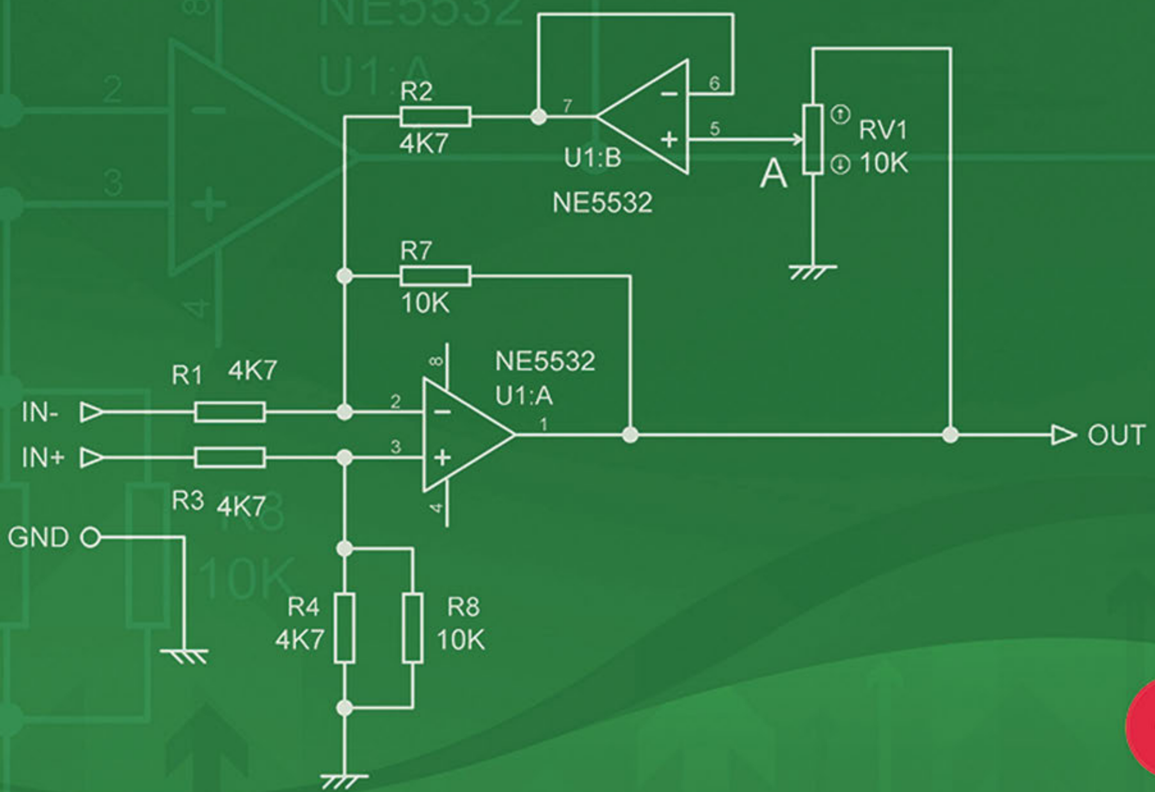
DOUGLAS SELF

SMALL SIGNAL AUDIO DESIGN

SECOND EDITION



Small Signal Audio Design

This much expanded second edition of ***Small Signal Audio Design*** is the essential and unique guide to the design of high-quality analogue circuitry for preamplifiers, mixing consoles, and many other signal-processing devices. You will learn to use inexpensive and readily available parts to obtain state-of-the-art performance in all the vital parameters of noise, distortion, crosstalk, etc. This practical handbook provides an extensive repertoire of circuit blocks from which almost any type of audio system can be built.

Essential points of theory that determine practical performance are lucidly and thoroughly explained, with the mathematics at an absolute minimum. Virtually every page reveals nuggets of specialized knowledge not found elsewhere. Douglas' background in design for manufacture ensures he keeps a wary eye on the cost of things.

Learn how to:

- Make amplifiers with apparently impossibly low noise
- Design discrete circuitry that can handle enormous signals with vanishingly low distortion
- Use ordinary bipolar transistors to make amplifiers with an input impedance of more than 50 Megohms
- Transform the performance of low-cost-opamps, and how to make filters with very low noise and distortion
- Make incredibly accurate volume controls
- Make a huge variety of audio equalisers
- Make magnetic cartridge preamplifiers that have noise so low it is limited by basic physics
- Sum, switch, clip, compress, and route audio signals effectively
- Build reliable power-supplies, with many practical ways to keep both the noise and the cost down

This much enlarged second edition is packed with new information, including completely new chapters on:

- Opamps for low voltages (down to 3.3 V)
- Moving-magnet inputs: archival and non-standard equalisation, for 78s etc.
- Moving-magnet inputs: discrete transistor circuitry
- Moving-magnet inputs: noise and distortion
- Balance and width controls
- Headphone amplifiers, including Class-A designs

There is also new material on: using multiple components to improve accuracy, ultra-linear discrete opamps, RIAA optimisation, the Baxandall volume control, distributed volume controls, loudness controls, the ideal balance-control law, instrumentation amplifier inputs, ground-cancelling outputs, zero-impedance outputs, and system control by microcontrollers.

This book includes numerous circuit blocks with component values so you can build them at once and easily adapt them to your particular requirements. It is lavishly illustrated with diagrams and graphs, and full of practical measurements on real circuitry using state-of-art testgear.

Douglas Self studied engineering at Cambridge University, then psychoacoustics at Sussex University. He has spent many years working at the top level of design in both the professional audio and hifi industries, and has taken out a number of patents in the field of audio technology. He currently acts as a consultant engineer in the field of audio design.

Small Signal Audio Design

Second Edition

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Dedication

To Julie, with all my love

Preface

Scientia potentia est

'Another damned thick book! Always scribble, scribble, scribble! Eh, Mr. Gibbon?'

Attributed to Prince William Henry, Duke of Gloucester, in 1781 upon receiving the second volume of *The History of the Decline and Fall of the Roman Empire* from its author.

This book deals with small-signal audio design; the amplification and control of audio in the analogue domain, where the processing is done with opamps or discrete transistors, usually working at a nominal level of a volt or less. It constitutes a major update of the first edition, being some 50% longer. 'Small-signal design' is the opposite term to the 'large-signal design' which in audio represents power amplifiers driving loudspeakers, rather than the electricity distribution grid or lightning.

There is unquestionably a need for high-quality analogue circuitry. For example, a good microphone preamplifier needs a gain range from 0 to +80 dB if it is to get any signal it is likely to encounter up to a workable nominal value. There is clearly little prospect of ever being able to connect an A-to-D converter directly to a microphone. The same applies to other low-output transducers such as moving-coil and moving-magnet phono cartridges. If you are starting at line level, and all you need is a simple but high-quality tone control, there is little incentive to convert to digital via a relatively expensive ADC, perform the very straightforward arithmetic manipulations in the digital domain, then go back to analogue via a DAC; there is also the need to implement the actual controls as rotary encoders and have those overseen by a microcontroller. All digital processing involves some delay, because it takes time to do the calculations; this is called the latency and can cause serious problems if more than one signal path is involved.

The total flexibility of digital signal processing certainly allows greater scope – you might contemplate how to go about implementing a one-second delay in the analogue domain, for example – but there are many times when greater quality or greater economy can be obtained by keeping the signal analogue. Sometimes analogue circuitry connects to the digital world, and so a complete chapter of this book deals with the subtleties of analogue/digital interfacing.

Therefore analogue circuitry is often the way to go. This book describes how to achieve high performance without spending a lot of money. As was remarked in a review of my recent book *Active Crossover Design*, duplicating this performance in the digital domain is not at all a trivial business. You can, of course, start off in analogue, and when you have identified the filter slopes, equalisation curves, and what-not that you want it is relatively easy to move it over to the DSP world.

I have devoted the first few chapters to the principles of high-quality, small-signal design, moving on to look closely at first hifi preamplifiers, and then mixing consoles. These two genres were chosen partly because they are of wide interest in themselves, but also because they use a large number of different functional blocks, with very little overlap between them. They cover a wide range of circuit functions that will be useful for all kinds of audio systems. You will find out how to adapt or design these building-blocks for audio, and how to put them together to form a system without bad things happening due to loading or interaction. You should then be able to design pretty much anything in this field.

In the pursuit of high quality at low cost, there are certain principles that pervade this book. Low-impedance design reduces the effects of Johnson noise and current noise without making voltage noise worse; the only downside is that low impedance requires an opamp capable of driving it effectively, and sometimes more than one. The most ambitious application of this approach so far has been in the ultra-low noise Elektor 2012 Preamplifier.

Another principle is that of using multiple components to reduce the effects of random noise. This may be electrical noise, in which case the outputs of several amplifiers are averaged (very simply, with a few resistors) and the noise from them is partially cancelled. Multiple amplifiers are also very useful for driving the low impedances just mentioned. Alternatively, it may be numerical noise, such as tolerances in a component value – making up the required value with multiple parts in series or parallel also makes errors partially cancel. This technique has its limits because of the square-root way it works; four amplifiers or components are required to half the noise, sixteen to reduce it to a quarter and so on.

There is also the principle of ‘optimisation’, in which each circuit block is closely scrutinised to see if it is possible to improve it by a bit more thinking. One example is the optimisation of RIAA equalisation networks. There are four ways to connect resistors and capacitors to make an RIAA network, and I have shown that one of them requires smaller values of expensive precision capacitors than the others. This new finding is presented in detail in Chapter 8, along with related techniques of optimising resistor values to get convenient capacitor values.

In many places, hybrid amplifiers combining the virtues of discrete active devices and opamps are used. If you put a bipolar transistor before an opamp, you get lower noise but the loop gain of the opamp means the distortion is as good as the opamp alone. This is extremely useful for making microphone amplifiers and virtual-earth summing amplifiers. If you reverse the order, with an opamp followed by bipolar transistors, you can drive much heavier loads, with the opamp gain once again providing excellent linearity. This latter technology, among others, is explained in a brand new chapter on headphone amplifiers.

So much has been added that it is difficult to summarise it, but the new material includes:

- An increasing demand for 5 V and 3.3 V single-rail audio with good performance has led to a whole new chapter on low-voltage opamps. Likewise there is new information on analogue switching with low supply rails.
- The material on moving-magnet and moving-coil amplifiers has been much expanded to include non-standard replay equalisation (for 78s, wax cylinders, etc.), more on moving-magnet noise, the ultimate limits on moving-coil noise performance, specialised filtering, and more. Mind you, the fact that it needs four whole chapters to cover the process of extracting a reasonable signal from a record groove indicates to me that there is something amiss with the whole concept.
- There is much more on discrete transistor circuitry, especially input stages for moving-magnet cartridges, and discrete opamps with ultra-low distortion.
- Active volume controls, especially the Baxandall control, are covered in greater depth. Loudness controls are currently unfashionable but the thinking behind them is intriguing. I include a possible solution to the mystery of why almost everyone disliked them, when consensus of any sort is rare in the hifi business.
- Balance controls now have their own chapter: passive, active, and switched types are covered, plus the technology for true constant-volume balance systems.
- The tone-control chapter is much expanded, and includes my new split-drive configuration which makes it practical to use 1 kΩ pots in a low-impedance design, giving much lower noise. There is a new design that gives variable-frequency HF and LF control in one stage, and a new type of switched-frequency LF EQ.
- Instrumentation amplifiers have long been praised for giving good common-mode rejection, but this has been hard to exploit in audio. I demonstrate how to do it, and get improved noise at the same time.
- There is much more on the ingenious but little-known technology of ground-cancelling outputs, showing how they can give a noise advantage over conventional balanced interconnections. Cunning ways of substantially reducing line output transformer distortion at near-zero cost are described.
- One of the many new chapters is devoted to headphone amplifiers, including hybrid types and a discrete Class-A design with ultra-low distortion.

- In the field of mixing console design, there is more on routing systems, balanced virtual-earth summing amplifiers, and level indication, including the Log Law Level LED or LLLL, which gives much more level information from a single LED than just on/off.

- The chapter on interfacing with the digital domain now includes the use of housekeeping microcontrollers for muting, input selection, IR decoding, and so on.

However, what you most emphatically will *not* find here is any truck with the religious dogma of audio subjectivism – the directional cables, the oxygen-free copper, the World War One vintage triodes still spattered with the mud of the Somme, and all the other depressing paraphernalia of pseudo- and anti-science. I have spent more time than I care to contemplate in double-blind listening tests – properly conducted ones, with rigorous statistical analysis – and every time the answer was that if you couldn't measure it you couldn't hear it. Very often if you could measure it you still couldn't hear it. However, faith-based audio is not going away any time soon because few people (apart, of course, from the unfortunate customers) have any interest in it so doing; you can bet your bottom diode on that. If you want to know more about my experiences and reasoning in this area, there is a full discussion in my book *Audio Power Amplifier Design.*

A good deal of thought and experiment has gone into this book, and I dare to hope that I have moved analogue audio design a bit further forward. I hope you find it useful. I hope you enjoy it too.

All suggestions for the improvement of this book that do not involve its combustion will be gratefully received. My email address can be found on the front page of my website at www.douglas-self.com.

To the best of my knowledge no supernatural assistance was received in the making of this book.

Further information, and PCBs, kits and built circuit boards of some of the designs described here, such as phono input stages and complete preamplifiers, can be found at www.signaltransfer.freeuk.com

Douglas Self

London, December 2013

Acknowledgments

My heartfelt thanks go to

Gareth Connor of *The Signal Transfer Company* for unfailing encouragement, providing the facilities with which some of the experiments in this book were carried out, and with much appreciation of our long collaboration in the field of audio.

Acronyms

ADC	Analog-to-digital converter
AFL	After-fade listen
AGS	Active gain stage
BFMA	Balanced-feedback microphone amplifier
BJT	Bipolar junction transistor
CFA	Current feedback amplifier
CFP	Complementary feedback pair
CM	Common mode
CMOS	Complementary metal oxide semiconductor
CMRR	Common-mode rejection ratio
CRM	Control-room monitor
DAC	Digital-to-analog converter
EF	Emitter-follower
EIN	Equivalent input noise
EQ	Equalization
ESR	Equivalent series resistance
ETP	Electrolytic tough pitch
FET	Field-effect transistor
FS	Full scale
GC	Ground canceling
HF	High frequency
IC	Integrated circuit
IDC	Insulation-displacement connector
JFET	Junction field-effect transistor
LED	Light-emitting diode
LF	Low frequency
MC	Moving-coil
MM	Moving-magnet
MOSFET	Metal oxide semiconductor field-effect transistor
NF	Noise figure
NFB	Negative feedback
OTA	Operational transconductance amplifier
PA	Public address
PCB	Printed-circuit board
PFL	Prefade listen
PGA	Programmable gain amplifier
PPM	Peak programme meter
PSRR	Power-supply rejection ratio
RF	Radio frequency
RIAA	Recording Industry Association of America
RTF	Return to flat
SIP	Solo in place
SM	Surface mount
TH	Through hole
THD	Total harmonic distortion
VAS	Voltage-amplifier stage
VCA	Voltage-controlled amplifier
VCVS	Voltage-controlled voltage source

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CHAPTER 1

Basics

Signals

An audio signal can be transmitted either as a voltage or a current. The construction of the universe is such that almost always the voltage mode is more convenient; consider for a moment an output driving more than one input. Connecting a series of high-impedance inputs to a low-impedance output is simply a matter of connecting them in parallel, and if the ratio of the output and input impedances is high there will be negligible variations in level. To drive multiple inputs with a current output it is necessary to have a series of floating current-sensor circuits that can be connected in series. This can be done [1], as pretty much anything in electronics can be done, but it requires a lot of hardware, and probably introduces performance compromises. The voltage-mode connection is just a matter of wiring.

Obviously, if there's a current, there's a voltage, and vice versa. You can't have one without the other. The distinction is in the output impedance of the transmitting end (low for voltage-mode, high for current-mode) and in what the receiving end responds to. Typically, but not necessarily, a voltage input has a high impedance; if its input impedance was only 600 Ω, as used to be the case in very old audio distribution systems, it is still responding to voltage, with the current it draws doing so a side issue, so it is still a voltage amplifier. In the same way, a current input typically, but not necessarily, has a very low input impedance. Current outputs can also present problems when they are not connected to anything. With no terminating impedance, the voltage at the output will be very high, and probably clipping heavily; the distortion is likely to crosstalk into adjacent circuitry. An open-circuit voltage output has no analogous problem.

Current-mode connections are not common. One example is the Krell Current Audio Signal Transmission (CAST) technology which uses current-mode to interconnect units in the Krell product range. While it is not exactly audio, the 4–20 mA current loop format is widely used in instrumentation. The current-mode operation means that voltage drops over long cable runs are ignored, and the zero offset of the current (i.e. 4 mA = zero) makes cable failure easy to detect – if the current is zero, you have a broken cable.

The old DIN interconnection standard was a form of current-mode connection in that it had voltage output via a high output impedance, of 100 kΩ or more. The idea was presumably that you could scale the output to a convenient voltage by selecting a suitable input impedance.

The drawback was that the high output impedance made the amount of power transferred very small, leading to a poor signal-to-noise ratio. The concept is now wholly obsolete.

Amplifiers

At the most basic level, there are four kinds of amplifier, because there are two kinds of signal (voltage and current) and two types of port (input and output). The handy word 'port' glosses over whether the input or output is differential or single ended. Amplifiers with differential input are very common – such as all opamps and most power amps – but differential outputs are rare and normally confined to specialised telecoms chips.

To summarise the four kinds of amplifier:

TABLE 1.1 The four types of amplifier

Amplifier type	Input	Output	Application
Voltage amplifier	Voltage	Voltage	General amplification
Transconductance amplifier	Voltage	Current	Voltage control of gain
Current amplifier	Current	Current	?
Transimpedance amplifier	Current	Voltage	Summing amplifiers, DAC interfacing

Voltage amplifiers

These are the vast majority of amplifiers. They take a voltage input at a high impedance and yield a voltage output at a low impedance. All conventional opamps are voltage amplifiers in themselves, but they can be made to perform as any of four kinds of amplifier by suitable feedback connections. Figure 1.1a shows a high-gain voltage amplifier with series voltage feedback. The closed-loop gain is (R1 + R2)/R2.

Transconductance amplifiers

The name simply means that a voltage input (usually differential) is converted to a current output. It has a transfer ratio $A = I_{out}/V_{in}$, which has dimensions of I/V or conductance, so it is referred to as a transconductance amplifier. It is possible to make a very simple, though not very linear, voltage-controlled amplifier with transconductance technology – differential-input operational transconductance amplifier (OTA) ICs have an extra pin that gives voltage-control of the transconductance, which when used with no negative feedback gives gain control (see Chapter 19 for details). Performance falls well short of that required for quality hifi or professional audio. Figure 1.1b shows an OTA used without feedback – note the current-source symbol at the output.

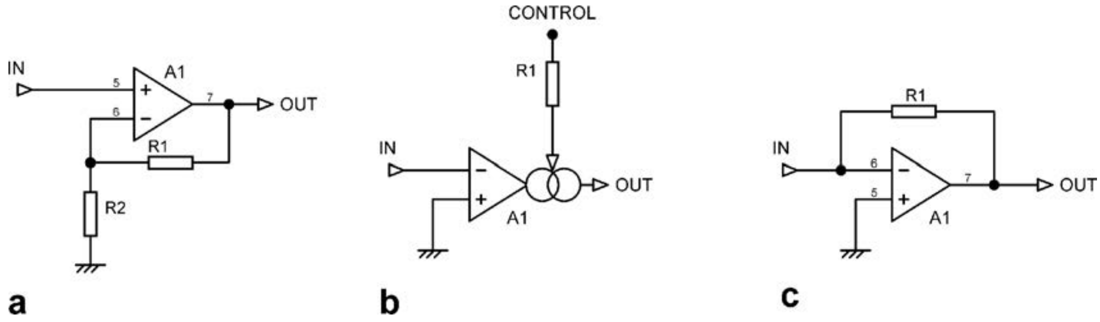

Figure 1.1: a) A voltage amplifier, b) a transconductance amplifier, c) a transimpedance amplifier

Current amplifiers

These accept a current in, and give a current out. Since, as we have already noted, current-mode operation is rare, there is not often a use for a true current amplifier in the audio business. They should not be confused with current feedback amplifiers (CFAs) which have a voltage output, the 'current' bit referring to the way the feedback is applied in current-mode [2]. The bipolar transistor is sometimes described as a current amplifier, but it is nothing of the kind. Current may flow in the base circuit but this is just an unwanted side-effect. It is the *voltage* on the base that actually controls the transistor.

Transimpedance amplifiers

A transimpedance amplifier accepts a current in (usually single-ended) and gives a voltage out. It is sometimes called an I–V converter. It has a transfer ratio $A = V_{out}/I_{in}$, which has dimensions of V/I or resistance. That is why it is referred to as a transimpedance or transresistance amplifier. Transimpedance amplifiers are usually made by applying shunt voltage feedback to a high-gain voltage amplifier. An important use is as virtual-earth summing amplifiers in mixing consoles (see Chapter 17). The voltage amplifier stage (VAS) in most power amplifiers is a transimpedance amplifier. They are used for I–V conversion when interfacing to DACs with current outputs (see Chapter 21). Transimpedance amplifiers are sometimes incorrectly described as 'current amplifiers'.

Figure 1.1c shows a high-gain voltage amplifier transformed into a transimpedance amplifier by adding the shunt voltage feedback resistor R1. The transimpedance gain is simply the value of R1, though it is normally expressed in V/µA rather than Ohms.

Negative feedback

Negative feedback is one of the most useful and omnipresent concepts in electronics. It can be used to control gain, to reduce distortion and improve frequency response, and to set input and output impedances, and one feedback connection can do all these things at the same time.

Negative feedback comes in four basic modes, as in the four basic kinds of amplifier. It can be taken from the output in two different ways (voltage or current feedback) and applied to the amplifier input in two different ways (series or shunt). Hence there are four combinations.

However, unless you're making something exotic like an audio constant-current source, the feedback is always taken as a voltage from the output, leaving us with just two feedback types, series and shunt, both of which are extensively used in audio. When series feedback is applied to a high-gain voltage amplifier, as in Figure 1.1a, the following statements are true:

- Negative feedback reduces voltage gain.
- Negative feedback increases gain stability.
- Negative feedback increases bandwidth.
- Negative feedback increases amplifier input impedance.
- Negative feedback reduces amplifier output impedance.
- Negative feedback reduces distortion.
- Negative feedback does not directly alter the signal-to-noise ratio.

If shunt feedback is applied to a voltage amplifier to make a transimpedance amplifier, as in Figure 1.1c, all the above statements are still true, except since we have applied shunt rather series negative feedback, the input impedance is reduced.

The basic feedback relationship in Equation 1.1 is dealt with at length in any number of textbooks, but it is of such fundamental importance that I feel obliged to include it here. The open-loop gain of the amplifier is A, and β is the feedback fraction, such that if in Figure 1.1a R1 is 2 kΩ and R2 is 1 kΩ, β is $\frac{1}{3}$. If A is very high, you don't even need to know it – the 1 on the bottom becomes negligible, and the As on top and bottom cancel out, leaving us with a gain of almost exactly three.

$$\frac{V_{out}}{V_{in}} = \frac{A}{1 + A\beta} \qquad \text{(Equation 1.1)}$$

Negative feedback can however do much more than stabilising gain. Anything unwanted occurring in the amplifier, be it distortion, DC drift or almost any of the other ills that electronics is heir to, is also reduced by the negative feedback factor (NFB factor for short). This is equal to:

$$\text{NFB factor} = \frac{1}{1 + A\beta} \qquad \text{(Equation 1.2)}$$

What negative feedback cannot do is improve the noise performance. When we apply feedback, the gain drops and the noise drops by the same factor, leaving the signal-to-noise ratio the same. Negative feedback and the way it reduces distortion is explained in much more detail in one of my other books [3].

Nominal signal levels and dynamic range

The absolute level of noise in a circuit is not of great significance in itself; what counts is how much greater the signal is than the noise – in other words, the signal to noise ratio. An important step in any design is the determination of the optimal signal level at each point in the circuit. Obviously a real signal, as opposed to a test sinewave, continuously varies in amplitude, and the signal level chosen is purely a nominal level. One must steer a course between two evils:

- If the signal level is too low, it will be contaminated unduly by noise.
- If the signal level is too high there is a risk it will clip and introduce severe distortion.

The wider the gap between them the greater the dynamic range. You will note that the first evil is a certainty, while the second is more of a statistical risk. The consequences of both must be considered when choosing a level, and, if the best possible signal-to-noise is required in a studio recording, then the internal level must be high, and if there is an unexpected overload you can always do another take. In live situations it will often be preferable to sacrifice some noise performance to give less risk of clipping. The internal signal levels of mixing consoles are examined in detail in Chapter 12.

If you seek to increase the dynamic range, you can either increase the maximum signal level or lower the noise floor. The maximum signal levels in opamp-based equipment are set by the voltage capabilities of the opamps used, and this usually means a maximum signal level of about 10 Vrms or +22 dBu. Discrete transistor technology removes the absolute limit on supply voltage and allows the voltage swing to be at least doubled before the supply rail voltages get inconveniently high. For example, ±40 V rails are quite practical for small-signal transistors and permit a theoretical voltage swing of 28 Vrms or +31 dBu. However, in view of the complications of designing your own discrete circuitry, and the greater space and power it requires, those nine extra dB of headroom are dearly bought. You must also consider the maximum signal capabilities of stages downstream – they might get damaged.

The dynamic range of human hearing is normally taken as 100 dB, ranging from the threshold of hearing at 0 dB SPL to the usual 'jack hammer at 1 m' at +100 dB SPL; however hearing damage is generally reckoned to begin with long exposures to levels above +80 dB SPL. There is, in a sense, a physical maximum to the loudest possible sound. Since sound is composed of cycles of compression and rarefaction, this limit is reached when the rarefaction creates a vacuum, because you can't have a lower pressure than that. This corresponds to about +194 dB SPL. I thought that this would probably be instantly fatal for a human being, but a little research shows that stun grenades generate +170 to +180 dB, so maybe not. It is certainly possible to get asymmetrical pressure spikes higher than +194 dB but it is not clear that this can be defined as sound.

Compare this with the dynamic range of a simple piece of cable. Let's say it has a resistance of 0.5 Ω; the Johnson noise from that will be −155 dBu. If we comply with the European Low Voltage Directive, the maximum voltage will be 50 V_{peak} = 35 Vrms = +33 dBu, so the dynamic range is 155 + 33 = 188 dBu, which, purely by coincidence, is close to the maximum sound level of 194 dB SPL.

Gain structures

There are some very basic rules for putting together an effective gain structure in a piece of equipment. Like many rules, they are subject to modification or compromise when you get into a tight corner. Breaking them reduces the dynamic range of the circuitry, either by worsening the noise or restricting the headroom; whether this is significant depends on the overall structure of the system and what level of performance you are aiming at. Three simple rules are:

1. Don't amplify then attenuate.
2. Don't attenuate then amplify.
3. The signal should be raised to the nominal internal level as soon as possible to minimise contamination with circuit noise.

There are exceptions. For an example, see Chapter 12 on moving-coil disc inputs, where attenuation after amplification does not compromise headroom because of a more severe headroom limit downstream.

Amplification then attenuation

Put baldly it sounds too silly to contemplate, but it is easy to thoughtlessly add a bit of gain to make up for a loss somewhere else, and immediately a few dB of precious and irretrievable headroom are gone for good. This assumes that each stage has the same power rails and hence the same clipping point, which is usually the case in opamp circuitry.

Fig 1.2a shows a system with a gain control designed to keep 10 dB of gain in hand. In other words, the expectation is that the control will spend most of its working life set somewhere around its '0 dB' position where it introduces 10 dB of attenuation, as is typically the case for a fader on a mixer. To maintain the nominal signal level at 0 dBu we need 10 dB of gain, and a +10 dB amplifier (Stage 2) has been inserted just before the gain control. This is not a good decision. This amplifier will clip 10 dB before any other stage in the system, and introduces what one might call a headroom bottleneck.

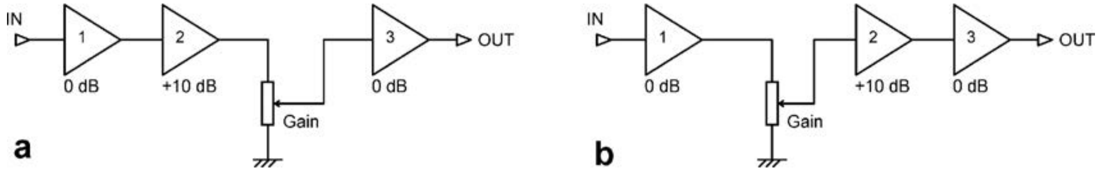

Figure 1.2: a) Amplification then attenuation. Stage 2 will always clip first, reducing headroom, b) attenuation then amplification. The noise from Stage 2 degrades the S/N ratio. The lower the gain setting, the worse the effect

There are exceptions. The moving-coil phono head-amp described in Chapter 8 appears to flagrantly break this rule, as it always works at maximum gain even when this is not required. But when considered in conjunction with the following RIAA stage, which also has considerable gain, it makes perfect sense, for the stage gains are configured so that the second stage always clips first, and there is actually no loss of headroom.

Attenuation then amplification

In Figure 1.2b the amplifier is now after the gain control, and noise performance rather than headroom suffers. If the signal is attenuated, any active device will inescapably add noise in restoring the level. Any conventional gain-control block has to address this issue. If we once more require a gain variable from +10 dB to off, i.e. $-\infty$ dB, as would be typical for a fader or volume control, then usually the potentiometer is placed before the gain stage as in Figure 1.2b because, as a rule, some loss in noise performance is more acceptable than a permanent 10 dB reduction in system headroom. If there are options for the amplifier stages in terms of a noise/cost trade-off (such as using the 5532 versus a TL072) and you can only afford one low-noise stage then it should be Stage 2. If all stages have the same noise performance this configuration is 10 dB noisier than the previous version when gain is set to 0 dB.

Raising the input signal to the nominal level

Getting the incoming signal up to the nominal internal level in one jump is always preferable as it gives the best noise performance. Sometimes it has to be done in two amplifier stages; typical examples are microphone preamps with wide gain ranges and phono preamps that insist on performing the RIAA equalisation in several goes (these are explored in their respective chapters). In these cases the noise contribution of the second stage may no longer be negligible.

Consider a signal path which has an input of −10 dBu and a nominal level of 0 dBu. The first version has an input amplifier with 10 dB of gain followed by two unity-gain circuit blocks A and B. All circuit blocks are assumed to introduce noise at −100 dBu. The noise output for the first version is −89.2 dBu. Now take a second version of the signal path that has an input amplifier with 5 dB of gain, followed by block A, another amplifier with 5 dB of gain, then block B. The noise output is now −87.5 dB, 1.7 dB worse, due to the extra amplification of the noise from block A. There is also more hardware, and the second version is clearly an inferior design.

Active gain controls

The previous section should not be taken to imply that noise performance must always be sacrificed when a gain control is included in the signal path. This is not so. If we move beyond the idea of a fixed-gain block, and recognise that the amount of gain present can be varied, then less gain when the maximum is not required will reduce the noise generated. For volume-control purposes it is essential that the gain can be reduced to near-zero, though it is not necessary for it to be as firmly 'off' as the faders or sends of a mixer.

An active volume-control stage gives lower noise at lower volume settings because there is less gain. The Baxandall active configuration also gives excellent channel balance as it depends solely on the mechanical alignment of a dual linear pot – all mismatches of its electrical characteristics are cancelled out, and there are no quasi-log dual slopes to induce anxiety. Active gain controls are looked at in depth in Chapter 9.

Noise

Noise here refers only to the random noise generated by resistances and active devices. The term is sometimes used to include mains hum, spurious signals from demodulated RF and other non-random sources, but this threatens confusion and I prefer to call the other unwanted signals 'interference'. In the one case we are striving to minimise the random variations arising in the circuit itself, in the other we are trying to keep extraneous signals out, and the techniques are wholly different.

When noise is referred to in electronics it means white noise unless it is specifically labelled as something else, because that is the form of noise that most electronic processes generate. There are two elemental noise mechanisms which make themselves felt in all circuits and active devices. These are Johnson noise and Shot noise, which are both forms of white noise. Both have Gaussian probability density functions. These two basic mechanisms generate the noise in both BJTs and FETs, though in rather different ways.

There are other forms of noise that originate from less fundamental mechanisms such as device processing imperfections which do not have a white spectrum: examples are $1/f$ (flicker) noise and popcorn noise. These noise mechanisms are described later in this chapter.

Non-white noise is given a colour which corresponds to the visible spectrum – thus red noise has a larger low-frequency content than white noise, while pink is midway between the two.

White noise has equal power in equal absolute bandwidth, i.e. with the bandwidth measured in Hz. Thus there is the same power between 100 and 200Hz as there is between 1100 and 1200Hz. It is the type produced by most electronic noise mechanisms [4].

Pink noise has equal power in equal ratios of bandwidth, so there is the same power between 100 and 200Hz as there is between 200 and 400Hz. The energy per Hz falls at 3 dB per octave as frequency increases. Pink noise is widely used for acoustic applications like room equalisation and loudspeaker measurement as it gives a flat response when viewed on a third-octave or other constant-percentage-bandwidth spectrum analyzer [5].

Red noise has energy per Hz falling at 6 dB per octave rather than 3. It is important in the study of stochastic processes and climate models, but has little application in audio. The only place you are likely to encounter it is in the oscillator section of analogue synthesisers. It is sometimes called Brownian noise as it can be produced by Brownian motion, hence its alternative name of random-walk noise. Brown here is a person and not a colour [6].

Blue noise has energy per Hz rising at 3 dB per octave. Blue noise is used for dithering in image anti-aliasing, but has, as far as I am aware, no application to audio. The spectral density of blue noise (i.e. the power per Hz) is proportional to the frequency. It appears that the light-sensitive cells in the retina of the mammalian eye are arranged in a pattern that resembles blue noise [7]. Great stuff, this evolution.

Violet noise has energy per Hz rising at 6 dB per octave (I imagine you saw that one coming). It is also known as 'differentiated white noise' as a differentiator circuit has a frequency response rising at 6 dB per octave. It is sometimes called purple noise. It has no known audio use.

Grey noise is pink noise modified by a psychoacoustic equal loudness curve, such as the inverse of the A-weighting curve, to give the perception of equal loudness at all frequencies.

Green noise really does exist, though not in the audio domain. It is used for stochastic half-toning of images, and consists of binary dither patterns composed of homogeneously distributed minority pixel clusters. I think we had better leave it there.

Johnson noise

Johnson noise is produced by all resistances, including those real resistances hiding inside transistors (such as r_{bb}, the base spreading resistance). It is not generated by the so-called intrinsic resistances, such as r_e, which is an expression of the V_{be}/I_c slope and not a physical resistance at all. Given that Johnson noise is present in every circuit, and often puts a limit on noise performance, it is a bit surprising that it was not discovered until 1928 by John B. Johnson at Bell Labs [8].

The rms amplitude of Johnson noise is easily calculated with the classic formula:

$$v_n = \sqrt{4kTRB} \qquad \text{(Equation 1.3)}$$

where v_n is the rms noise voltage, T is absolute temperature in °K, B is the bandwidth in Hz, k is Boltzmann's constant and R is the resistance in Ohms.

The only thing to be careful with here (apart from the usual problem of keeping the powers of ten straight) is to make sure you use Boltzmann's constant (1.380662×10^{-23}), and *not* the Stefan-Boltzmann constant (5.67×10^{-08}) which relates to black-body radiation, and will give some spectacularly wrong answers. Often the voltage noise is left in its squared form for ease of summing with other noise sources. Table 1.2 gives a feel for how resistance affects the magnitude of Johnson noise. The temperature is 25 °C and the bandwidth is 22 kHz.

Johnson noise theoretically goes all the way to daylight, but in the real world is ultimately band-limited by the shunt capacitance of the resistor. Johnson noise is not produced by circuit reactances – i.e. pure capacitance and inductance. In the real world, however, reactive components are not pure, and the winding resistances of transformers can produce significant

TABLE 1.2 Resistances and their Johnson noise

Resistance (Ω)	Noise voltage (μV)	Noise voltage (dBu)	Application
1	0.018	−152.2	Moving-coil cartridge impedance (low output)
3.3	0.035	−147.0	Moving-coil cartridge impedance (medium output)
10	0.060	−142.2	Moving-coil cartridge impedance (high output)
47	0.13	−135.5	Line output isolation resistor
100	0.19	−132.2	Output isolation or feedback network
150	0.23	−130.4	Dynamic microphone source impedance
200	0.27	−129.2	Dynamic microphone source impedance (older)
600	0.47	−124.4	The ancient matched-line impedance
1000	0.60	−122.2	A nice round number
2500	0.95	−118.2	Worst-case output impedance of 10 kΩ pot
5000	1.35	−115.2	Worst-case output impedance of 20 kΩ pot
12500	2.13	−111.2	Worst-case output impedance of 50 kΩ pot
25000	3.01	−108.2	Worst-case output impedance of 100 kΩ pot
1 Meg (10^6)	19.0	−92.2	Another nice round number
1 Giga (10^9)	190	−62.2	As used in capacitor microphone amplifiers
1 Tera (10^{12})	1900	−32.2	Insulation testers read in Tera-Ohms
1 Peta (10^{15})	19,000	−2.2	OK, it's getting silly now

Johnson noise; this is an important factor in the design of moving-coil cartridge step-up transformers. Capacitors with their very high leakage resistances approach perfection much more closely, and the capacitance has a filtering effect. They usually have no detectable effect on noise performance, and in some circuitry it is possible to reduce noise by using a capacitive potential divider instead of a resistive one [9].

The noise voltage is of course inseparable from the resistance, so the equivalent circuit is of a voltage source in series with the resistance present. While Johnson noise is usually represented as a voltage, it can also be treated as a Johnson noise current, by means of the Thevenin-Norton transformation, which gives the alternative equivalent circuit of a current-source in shunt with the resistance. The equation for the noise current is simply the Johnson voltage divided by the value of the resistor it comes from:

$$i_{\mathrm{n}} = v_{\mathrm{n}}/R$$

When it is first encountered, this ability of resistors to generate electricity from out of nowhere seems deeply mysterious. You wouldn't be the first person to think of connecting a small electric motor across the resistance and getting some useful work out – and you wouldn't be the first person to discover it doesn't work. If it did, then by the First Law of Thermodynamics (the law of conservation of energy) the resistor would have to get colder, and such a process is flatly forbidden by . . . the Second Law of Thermodynamics. The Second Law is no more negotiable than the First Law, and it says that energy cannot be extracted by simply cooling down one body. If you could, it would be what thermodynamicists call a Perpetual Motion Machine of the Second Kind, and they are no more buildable than the common sort of perpetual motion machine.

It is interesting to speculate what happens as the resistor is made larger. Does the Johnson voltage keep increasing until there is a hazardous voltage across the resistor terminals? Obviously not, or picking up any piece of plastic would be a lethal experience. Johnson noise comes from a source impedance equal to the resistor generating it, and this alone would prevent any problems. Table 1.2 ends with a couple of silly values to see just how this works: the square root in the equation means that you need a peta-ohm resistor (1×10^{15} Ω) to reach even 600 mVrms of Johnson noise. Resistors are made up to at least 100 GΩ but peta-ohm resistors (PΩ?) would really be a minority interest.

Shot noise

It is easy to forget that an electric current is not some sort of magic fluid, but is actually composed of a finite, though usually very large, number of electrons, so current is in effect quantized. Shot noise is so called because it allegedly sounds like a shower of lead shot being poured onto a drum, and the name emphasises the discrete nature of the charge-carriers.

TABLE 1.3 How shot noise varies with current

Current (DC)	Current noise (nA_{rms})	Fluctuation (%)	R (Ω)	Voltage noise (μV)	Voltage noise (dBu)
1 pA	0.000084	8.4	100	8.4×10^{-6}	−219.3
1 nA	0.0026	0.27	100	0.000265	−189.3
1 μA	0.084	0.0084	100	0.0084	−159.3
1 mA	2.65	0.00027	100	0.265	−129.3
1 A	84	0.000008	100	8.39	−99.3

Despite the picturesque description, the spectrum is still that of white noise, and the noise current amplitude for a given steady current is described by a surprisingly simple equation (as Einstein said, the most incomprehensible thing about the universe is that it is comprehensible) that runs thus:

$$\text{Noise current } i_n = \sqrt{ZqI_{\text{dc}}B} \qquad \text{(Equation 1.4)}$$

where q is the charge on an electron (1.602×10^{-19} Coulomb), I_{dc} is the mean value of the current and B is the bandwidth examined.

As with Johnson noise, often the shot noise is left in its squared form for ease of summing with other noise sources. Table 1.3 helps to give a feel for the reality of shot noise. As the current increases, the shot noise increases too, but more slowly as it depends on the square root of the DC current; therefore the *percentage* fluctuation in the current becomes less. It is the small currents which are the noisiest.

The actual level of shot noise voltage generated if the current noise is assumed to flow through a 100 Ohm resistor is rather low, as the last column shows. There are few systems which will be embarrassed by an extra noise source of even −99 dBu unless it occurs right at the very input. To generate this level of shot noise requires 1 Amp to flow through 100 Ohms, which naturally means a voltage-drop of 100 V and a 100 Watts of power dissipated. These are not often the sort of circuit conditions that exist in preamplifier circuitry. This does not mean that shot noise can be ignored completely, but it can usually be ignored unless it is happening in an active device where the noise is amplified.

1/*f* noise (flicker noise)

This is so-called because it rises in amplitude proportionally as the frequency examined falls. Unlike Johnson noise and shot noise, it is not a fundamental consequence of the way the universe is put together, but the result of imperfections in device construction.

$1/f$ noise appears in all kinds of active semiconductors, and also in some resistors. As frequency falls, the $1/f$ noise amplitude stays level down to the $1/f$ corner frequency, after which it rises at 6 dB/octave. For a discussion of flicker noise in resistors see Chapter 2.

Popcorn noise

This form of noise is named after the sound of popcorn being cooked, rather than eaten. It is also called burst noise or bistable noise, and is a type of low frequency noise that appears primarily in integrated circuits, appearing as low level step changes in the output voltage, occurring at random intervals. Viewed on an oscilloscope, this type of noise shows bursts of changes between two or more discrete levels. The amplitude stays level up to a corner frequency, at which point it falls at a rate of $1/f^2$. Different burst-noise mechanisms within the same device can exhibit different corner frequencies. The exact mechanism is poorly understood, but is known to be related to the presence of heavy-metal ion contamination, such as gold. As for $1/f$ noise, the only measure that can be taken against it is to choose an appropriate device. Like $1/f$ noise, popcorn noise does not have a Gaussian amplitude distribution.

Summing noise sources

When random noise from different sources is summed, the components do not add in a 2 + 2 = 4 manner. Since the noise components come from different sources, with different versions of the same physical processes going on, they are uncorrelated and will partially reinforce and partially cancel, so root-mean-square (rms) addition holds, as shown in Equation 1.5. If there are two noise sources with the same level, the increase is 3 dB rather than 6 dB. When we are dealing with two sources in one device, such as a bipolar transistor, the assumption of no correlation is slightly dubious, because some correlation is known to exist, but it does not seem to be enough to cause serious calculation errors.

$$V_{ntot} = \sqrt{(V_{n1}^2 + V_{n2}^2 + \ldots)} \qquad \text{(Equation 1.5)}$$

Any number of noise sources may be summed in the same way, by simply adding more squared terms inside the square root, as shown by the dots. When dealing with noise in the design process, it is important to keep in mind the way that noise sources add together when they are not of equal amplitude. Table 1.4 on the following page shows how this works in decibels. Two equal voltage noise sources give a sum of +3 dB, as expected. What is notable is that when the two sources are of rather unequal amplitude, the smaller one makes very little contribution to the result.

TABLE 1.4 The summation of two uncorrelated noise sources

dB	dB	dB sum
0	0	+3.01
0	−1	+2.54
0	−2	+2.12
0	−3	+1.76
0	−4	+1.46
0	−5	+1.19
0	−6	+0.97
0	−10	+0.41
0	−15	+0.14
0	−20	+0.04

If we have a circuit in which one noise source is twice the rms amplitude of the other (a 6 dB difference), then the quieter source only increases the rms-sum by 0.97 dB, a change barely detectable on critical listening. If one source is 10 dB below the other, the increase is only 0.4 dB, which in most cases could be ignored. At 20 dB down, the increase is lost in measurement error. This mathematical property of uncorrelated noise sources is exceedingly convenient, because it means that in practical calculations we can neglect all except the most important noise sources with minimal error. Since all semiconductors have some variability in their noise performance, it is rarely worthwhile to make the calculations to great accuracy.

Noise in amplifiers

There are basic principles of noise design that apply to all amplifiers, be they discrete or integrated, single-ended or differential. Practical circuits, even those consisting of an opamp and two resistors, have multiple sources of noise. Typically one source of noise will dominate, but this cannot be taken for granted and it is essential to evaluate all the sources and the ways that they add together if a noise calculation is going to be reliable. Here I add the complications one stage at a time.

Figure 1.3 shows that most useful of circuit elements, the perfect noiseless amplifier (these seem to be unaccountably hard to find in catalogues). It is assumed to have a definite gain A without bothering about whether it is achieved by feedback or not, and an infinite input impedance. To emulate a real amplifier, noise sources are concentrated at the input, combined into one voltage noise source and one current noise source. These can represent any number of actual noise sources inside the real amplifier. Figure 1.3 shows two ways of drawing the same situation.

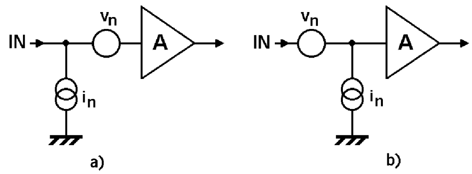

Figure 1.3: The noise sources of a perfect amplifier. The two circuits are exactly equivalent

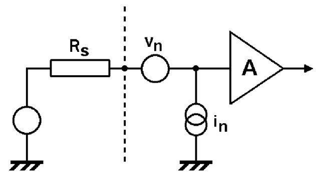

Figure 1.4: The perfect amplifier and noise sources with a signal source now connected

It does not matter on which side of the voltage source the current source is placed; the 'perfect' amplifier has an infinite input impedance, and the voltage source a zero impedance, so either way all of the current noise flows through whatever is attached to the input.

Figure 1.4 shows the first step to a realistic situation, with a signal source now connected to the amplifier input. The signal source is modelled as a perfect zero-impedance voltage source with added series resistance R_s. Many signal sources are modelled accurately enough for noise calculations in this way. Examples are low impedance dynamic microphones, moving-coil phono cartridges, and most electronic outputs. In others cases, such as moving-magnet phono cartridges and capacitor microphone capsules, there is a big reactive component which has a major effect on the noise behaviour and cannot be ignored or treated as a resistor. The magnitude of the reactances tends to vary from one make to another, but fortunately the variations are not great enough for the circuit approach for optimal noise to vary greatly. It is pretty clear that a capacitor microphone will have a very high source impedance at audio frequencies, and will need a special high-impedance preamplifier to avoid low frequency roll-off (see Chapter 13 for details). It is perhaps less obvious that the series inductance of a moving-magnet phono cartridge becomes the dominating factor at the higher end of the audio band, and designing for the lowest noise with the 600 Ω or so series resistance alone will give far from optimal results. This is dealt with in Chapter 7.

There are two sources of voltage noise in the circuit of Figure 1.4.

1. The amplifier voltage noise source v_n at the input.
2. The Johnson noise from the source resistance Rs.

These two voltage sources are in series and sum by rms-addition as they are uncorrelated.

There is only one current noise component – the amplifier noise current source i_n across the input. This generates a noise voltage when its noise current flows through Rs (it cannot flow into the amplifier input because we are assuming an infinite input impedance). This third source of voltage noise is also added in by rms-addition, and the total is amplified by the voltage gain A and appears at the output. The noise voltage at the input is the equivalent input noise (EIN). This is impossible to measure, so the noise at the amplifier output is divided by A to get the EIN. Having got this, we can compare it with the Johnson noise from the source resistance Rs; with a noiseless amplifier there would be no difference, but in real life the EIN will be higher by a number of dB which is called the noise figure (NF). This gives a concise way of assessing how noisy our amplifier is and if it is worth trying to improve it. Noise figures very rarely appear in hifi literature, probably because most of them wouldn't look very good. For the application of noise figures to phono cartridge amplifiers, see Chapters 8 and 11.

Noise in bipolar transistors

An analysis of the noise behaviour of discrete bipolar transistors can be found in many textbooks, so this is something of a quick summary of the vital points. Two important transistor parameters for understanding noise are r_{bb}, the base spreading resistance, and r_e, the intrinsic emitter resistance. r_{bb} is a real physical resistance – what is called an *extrinsic* resistance. The second parameter, r_e, is an expression of the V_{be}/I_c slope and not a physical resistance at all, so it is called an *intrinsic* resistance.

Noise in bipolar transistors, as in amplifiers in general, is best dealt with by assuming a noiseless transistor with a theoretical noise voltage source in series with the base and a theoretical noise current source connected from base to emitter. These sources are usually described simply as the 'voltage noise' and the 'current noise' of the transistor.

Bipolar transistor voltage noise

The voltage noise v_n is made up of two components:

1. The Johnson noise generated in the base spreading resistance r_{bb}.
2. The collector current (I_c) shot noise creating a noise voltage across r_e, the intrinsic emitter resistance.

These two components can be calculated from the equations given earlier, and rms-summed thus:

Voltage noise density $v_n = \sqrt{4kTr_{bb} + 2(kT)^2/(qI_c)}$ in $\mathrm{V}/\sqrt{\mathrm{Hz}}$ (usually $\mathrm{nV}/\sqrt{\mathrm{Hz}}$) (Equation 1.6) where k is Boltzmann's constant (1.380662×10^{-23}), q is the charge on an electron (1.602×10^{-19} Coulomb), T is absolute temperature in °K, I_c is the collector current and r_{bb} is the base resistance in Ohms.

The first part of this equation is the usual expression for Johnson noise, and is fixed for a given transistor type by the physical value of r_{bb}, so the lower this is the better. The only way you can reduce this is by changing to another transistor type with a lower R_{bb} or using paralleled transistors. The absolute temperature is a factor; running your transistor at 25 °C rather than 125 °C reduces the Johnson noise from r_{bb} by 1.2 dB. Input devices usually run cool but this may not be the case with moving-coil preamplifiers where a large I_c is required, so it is not impossible that adding a heatsink would give a measurable improvement in noise.

The second (shot noise) part of the equation decreases as collector current I_c increases; this is because as I_c increases, r_e decreases proportionally, following $r_e = 25/I_c$ where I_c is in mA. The shot noise, however, is only increasing as the square root of I_c, and the overall result is that the total v_n falls – though relatively slowly – as collector current increases, approaching asymptotically the level of noise set by the first part of the equation. There is an extra voltage noise source resulting from flicker noise produced by the base current flowing through r_{bb}; this is only significant at high collector currents and low frequencies due to its 1/*f* nature, and is not usually included in design calculations unless low frequency quietness is a special requirement.

Bipolar transistor current noise

The current noise i_n, is mainly produced by the shot noise of the steady current I_b flowing through the transistor base. This means it increases as the square root of I_b increases. Naturally I_b increases with I_c. Current noise is given by:

$$\text{Current noise density } i_n = \sqrt{2qI_b} \text{ in A}/\sqrt{\text{Hz}} \text{ (usual values are in pA)} \qquad \text{(Equation 1.7)}$$

where q is the charge on an electron and I_b is the base current.

So, for a fixed collector current, you get less current noise with high-beta transistors because there is less base current.

The existence of current noise as well as voltage noise means it is not possible to minimise transistor noise just by increasing the collector current to the maximum value the device can take. Increasing I_c reduces voltage noise, but it increases current noise, as in Figure 1.5. There is an optimum collector current for each value of source resistance, where the contributions are equal. Because both voltage and current noise are proportional to the square root of I_c, they change slowly as it alters, and the combined noise curve is rather flat at the bottom. There is no need to control collector current with great accuracy to obtain optimum noise performance.

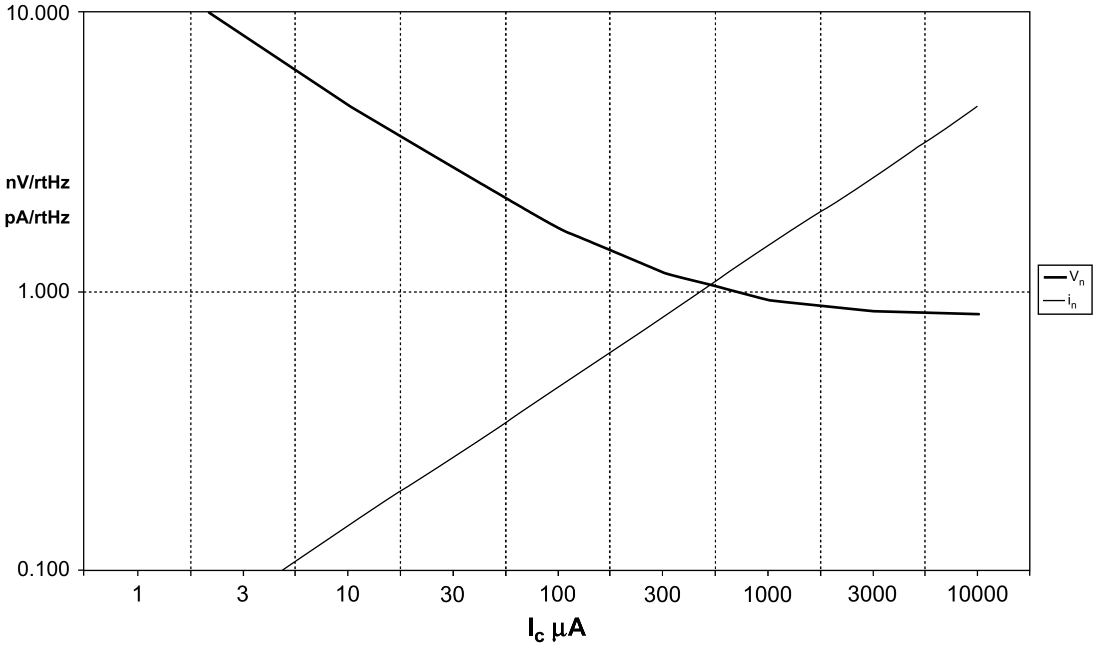

Figure 1.5: How voltage noise density V_n and current noise density I_n vary with collector current I_c in a generic transistor. As I_c increases voltage noise asymptotes to a limit while current noise continuously increases

I must emphasise that this is a simplified noise model. In practice both voltage and current noise densities vary with frequency. I have also ignored 1/*f* noise. However, it gives the essential insight into what is happening and leads to the right design decisions so we will put our heads down and press on.

A quick example shows how this works. In a voltage amplifier we want the source impedances seen by the input transistors to be as low as possible, to minimise Johnson noise, and to minimise the effects of current noise. If we are lucky it may be as low as 100 Ω. How do we minimise the noise from a single transistor faced with a 100 Ω source resistance?

We assume the temperature is 25 °C, the bandwidth is 22 kHz, and the r_{bb} of our transistor is 40 Ω (why don't they put this on spec sheets anymore?). The h_{fe} (beta) is 150. Set I_c to 1 mA, which is plausible for an amplifier input stage, step the source resistance from 1 to 100,000 Ω in decades, and we get Table 1.5.

Column 1 shows the source resistance, and Column 2 the Johnson noise density it generates by itself. Factor in the bandwidth, and you get Columns 3 and 4 which show the voltage in nV and dBu respectively. Column 5 is the noise density from the transistor,

TABLE 1.5 The summation of Johnson noise from the source resistance with transistor noise

1 R_s (Ω)	2 R_s Johnson (nV/√Hz)	3 R_s Johnson BW (nV)	4 R_s Johnson BW (dBu)	5 Transistor noise incl. in R_s (nV/√Hz)	6 Transistor noise plus R_s Johnson (nV/√Hz)	7 Noise in BW (nV)	8 Noise in BW (dBu)	9 Noise figure (dB)
1	0.128	19.0	−152.2	0.93	0.94	139.7	−134.9	17.3
10	0.406	60.2	−142.2	0.93	1.02	150.9	−134.2	8.0
100	1.283	190.3	−132.2	0.94	1.59	236.3	−130.3	1.9
1000	4.057	601.8	−122.2	1.73	4.41	654.4	−121.5	0.7
10000	12.830	1903.0	−112.2	14.64	19.46	2886.9	−108.6	3.6
100000	40.573	6017.9	−102.2	146.06	151.59	22484.8	−90.7	11.4

the rms-sum of the voltage noise and the voltage generated by the current noise flowing in the source resistance. Column 6 gives total noise density when we sum the source resistance noise density with the transistor noise density. Factor in the bandwidth again, and the resultant noise voltage is given in Columns 7 and 8. Column 9 gives the noise figure (NF), which is the amount by which the combination of transistor and source resistance is noisier than the source resistance alone. In other words, it tells how close we have got to perfection, which would be a noise figure of 0 dB. The results for the 100 Ω source show that the transistor noise is less than the source resistance Johnson noise; there is little scope for improving things by changing transistor type or operating conditions.

The results for the other source resistances are worth looking at. The lowest noise output (−134.9 dBu) is achieved by the lowest source resistance of 1 Ω, as you would expect, but the NF is very poor at 17.3 dB, because the r_{bb} at 40 Ω is generating a lot more noise than the 1 Ω source. This gives you some idea why it is hard to design quiet moving-coil head amplifiers. The best noise figure, and the closest approach to theoretical perfection is with a 1000 Ω source, attained with a *greater* noise output than 100 Ω. As source resistance increases further, NF worsens again; a transistor with I_c = 1 mA has relatively high current noise and performs poorly with high source resistances.

Since I_c is about the only thing we have any control over here, let's try altering it. If we increase I_c to 3 mA we find that for a 100 Ω source resistance, our amplifier is only a marginal 0.2 dB quieter. See Table 1.6, which skips the intermediate calculations and just gives the output noise and NF.

At 3 mA the noise with a 1 Ω source is 0.7 dB better, due to slightly lower voltage noise, but with 100 kΩ noise is higher by no less than 9.8 dB as the current noise is much increased.

TABLE 1.6 How input device collector current affects noise output and noise figure

	I_c = 3 mA		I_c = 10 mA		I_c = 10 mA,2SB737		I_c = 100 uA	
R_s (Ω)	Noise (dBu)	NF (dB)	Noise (dBu)	NF (dB)	Noise (dBu)	NF (dB)	Noise (dBu)	NF (dB)
1	−135.6	16.6	−135.9	16.3	−145.9	6.3	−129.9	22.3
10	−134.8	7.4	−135.1	7.1	−140.9	1.3	−129.7	12.5
100	−130.5	1.7	−130.3	1.9	−131.5	0.7	−127.9	4.3
1000	−120.6	1.6	−118.5	3.7	−118.6	3.6	−121.5	0.7
10k	−105.3	6.4	−100.7	11.4	−100.7	11.4	−111.6	0.6
100k	−86.2	16.0	−81.0	21.2	−81.9	21.2	−98.6	3.6

If we increase I_c to 10 mA, this makes the 100 Ω noise worse again, and we have lost that slender 0.2 dB improvement.

At 1 Ω the noise is 0.3 dB better, which is not exactly a breakthrough, and for the higher source resistances things worsen again, the 100 kΩ noise increasing by another 5.2 dB. It therefore appears that a collector current of 3 mA is actually pretty much optimal for noise with our 100 Ω source resistance.

If we now pluck out our ‘ordinary’ transistor and replace it with a specialised low r_{bb} part like the much-lamented 2SB737, with its superbly low r_{bb} of 2 Ω, the noise output at 1 Ω plummets by 10 dB, showing just how important low r_{bb} is for moving-coil head amplifiers. The improvement for the 100 Ω source resistance is much less at 1.0 dB.

If we go back to the ordinary transistor and reduce I_c to 100 μA, we get the last two columns in Table 1.6. Compared with I_c = 3 mA, noise with the 1 Ω source worsens by 5.7 dB, and with the 100 Ω source by 2.6 dB, but with the 100 kΩ source there is a hefty 12.4 dB improvement, due to reduced current noise. Quiet BJT inputs for high source impedances can be made by using low collector currents, but JFETs usually give better noise performance under these conditions.

The transistor will probably be the major source of noise in the circuit, but other sources may need to be considered. The transistor may have a collector resistor of high value, to optimise the stage gain, and this naturally introduces its own Johnson noise. Most discrete transistor amplifiers have multiple stages, to get enough open-loop gain for linearisation by negative feedback, and an important consideration in discrete noise design is that the gain of the first stage should be high enough to make the noise contribution of the second stage negligible. This can complicate matters considerably. Precisely the same situation prevails in an opamp, but here someone else has done the worrying about second-stage noise for you; if you’re not happy with it all you can do is pick another type of opamp.

Noise in JFETs

JFETs operate completely differently to bipolar transistors, and noise arises in different ways. The voltage noise in JFETs arises from the Johnson noise produced by the channel resistance, the effective value of which is the inverse of the transconductance (g_m) of the JFET at the operating point we are looking at. An approximate but widely accepted equation for this noise is:

$$\text{Noise density } e_n = \sqrt{4kT\frac{2}{3g_m}} \text{ in V}/\sqrt{\text{Hz}} \text{ (usually nV}/\sqrt{\text{Hz}}) \qquad \text{(Equation 1.8)}$$

where k is Boltzmann's constant (1.380662×10^{-23}) and T is absolute temperature in °K.

FET transconductance goes up proportionally to the square root of drain current I_d. When the transconductance is inserted into the equation above, it is again square-rooted, so the voltage noise is proportional to the fourth root of drain current, and varies with it very slowly. There is thus little point in using high drain currents

The only current noise source in a JFET is the shot noise associated with the gate leakage current. Because the leakage current is normally extremely low, the current noise is very low, which is why JFETs give a good noise performance with high source resistances. However, don't let the JFET get hot, because gate leakage doubles with each 10 °C rise in temperature; this is why JFETs can actually show *increased* noise if the drain current is increased to the point where they heat up.

The g_m of JFETs is rather variable, but at $I_d = 1$ mA ranges over about 0.5 to 3 mA/V (or mMho) so the voltage noise density varies from 4.7 to 1.9 nV/$\sqrt{\text{Hz}}$. Comparing this with Column 5 in Table 1.5, we can see that the BJTs are much quieter except at high source impedances, where their current noise makes them noisier than JFETs.

However, if you are prepared to use multiple devices, the lowest possible noise may be given by JFETs, because the voltage noise falls faster than the effect of the current noise rising, when more devices are added. A low-noise laboratory amplifier design by Samuel Groner achieves a spectacularly low noise density of 0.39 nV/$\sqrt{\text{Hz}}$ by using eight paralleled JFETs [10].

Noise in opamps

The noise behaviour of an opamp is very similar to that of a single input amplifier, the difference being that there are now two inputs to consider, and usually more associated resistors.

An opamp is driven by the voltage difference between its two inputs, and so the voltage noise can be treated as one voltage v_n connected between them. See Figure 1.6, which shows a differential amplifier.

Opamp current noise is represented by two separate current generators i_n+ and i_n-, one in parallel with each input. These are assumed to be equal in amplitude and not correlated with each

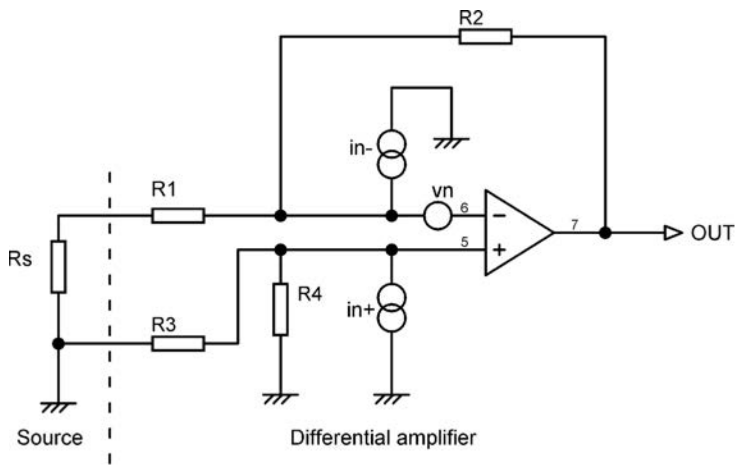

Figure 1.6: The noise sources in an opamp differential amplifier circuit

other. It is also assumed that the voltage and current noise sources are likewise uncorrelated, so that rms-addition of their noise components is valid. In reality, things are not quite so simple, and there is some correlation, and the noise produced can be slightly higher than calculated. In practice the difference is small compared with natural variations in noise performance.

Calculating the noise is somewhat more complex than for the simple amplifier of Figure 1.4. You must:

1. Calculate the voltage noise from the voltage noise density.
2. Calculate the two extra noise voltages resulting from the noise currents flowing through their associated components.
3. Calculate the Johnson noise produced by each resistor.
4. Allow for the noise gain of the circuit when assessing how much each noise source contributes to the output.
5. Add the lot together by rms-addition.

There is no space to go through a complete calculation, but here is a quick example:

Suppose you have an inverting amplifier like that in Figure 1.9a below. This is simpler because the non-inverting input is grounded, so the effect of i_n+ disappears, as it has no resistance to flow through and cannot give rise to a noise voltage. This shunt-feedback stage has a 'noise gain' that is greater than the signal gain. The input signal is amplified by -1, but the voltage noise source in the opamp is amplified by 2, because the voltage noise generator is amplified as if the circuit was a series-feedback gain stage.

Low-noise opamp circuitry

The rest of this chapter deals with designing low-noise opamp circuitry, dealing with opamp selection and the minimisation of circuit impedances. It also shows how adding more stages can actually make the circuitry quieter. This sounds somewhat counter-intuitive, but as you will see, it is so.

When you are designing for low noise, it is obviously important to select the right opamp, the great divide being between bipolar and JFET inputs. This chapter concentrates mainly on using the 5532, as it is not only a low-noise opamp with superbly low distortion, but also a low-cost opamp, due to its large production quantities. There are opamps with lower noise, such as the AD797 and the LT1028, but these are specialised items and the cost penalties are high. The LT1028 has a bias-cancellation system that increases noise unless the impedances seen at each input are equal, and since audio does not need the resulting DC precision, it is not useful. The new LM4562 is a dual opamp with somewhat lower noise than the 5532, but at present it also is much more expensive.

The AD797 runs its bipolar input transistors at high collector currents (about 1 mA) which reduces voltage noise but increases current noise. The AD797 will therefore only give lower noise for rather low source resistances; these need to be below 1 kΩ to yield benefit for the money spent. There is much more on opamp selection in Chapters 4 and 14.

Noise measurements

There are difficulties in measuring the low noise levels we are dealing with here. The Audio Precision System 1 test system has a noise floor of −116.4 dBu when its input is terminated with a 47 Ω resistor. When it is terminated in a short circuit, the noise reading only drops to −117.0 dBu, demonstrating that almost all the noise is internal to the AP and the Johnson noise of the 47 Ω resistor is much lower. The significance of 47 Ω is that it is the lowest value of output resistor that will guarantee stability when driving the capacitance of a reasonable length of screened cable; this value will keep cropping up.

To delve below this noise floor, we can subtract this figure from the noise we measure (on the usual rms basis) and estimate the noise actually coming from the circuit under test. This process is not very accurate when circuit noise is much below that of the test system, because of the subtraction involved, and any figure below −120 dBu should be regarded with caution. Cross-checking against the theoretical calculations and SPICE results is always wise; in this case it is essential.

We will now look at a number of common circuit scenarios and see how low-noise design can be applied to them.

How to attenuate quietly

Attenuating a signal by 6 dB sounds like the easiest electronic task in the world. Two equal-value resistors to make up a potential divider, and *voila*! This knotty problem is solved. Or is it?

To begin with, let us consider the signal going into our divider. Wherever it comes from, the source impedance is not likely to be less than 50 Ω. This is also the lowest output impedance setting for most high-quality signal generators (though it's 40 Ω on my AP SYS-2702). The Johnson noise from 50 Ω is −135.2 dBu, which immediately puts a limit – albeit a very low one – on the performance we can achieve. The maximum signal handling capability of opamps is about +22 dBu, so we know at once our dynamic range cannot exceed 135 + 22 = 155 dB. This comfortably exceeds the dynamic range of human hearing, which is about 130 dB if you are happy to accept 'instantaneous ear damage' as the upper limit.

In the scenario we are examining, there is only one variable – the ohmic value of the two equal resistors. This cannot be too low or the divider will load the previous stage excessively, increasing distortion and possibly reducing headroom. On the other hand, the higher the value, the greater the Johnson noise voltage generated by the divider resistances that will be added to the signal, and the greater the susceptibility of the circuit to capacitative crosstalk and general interference pickup. The trade-off is examined in Table 1.7.

What happens when our signal with its −135.2 dBu noise level encounters our 6 dB attenuator? If it is made up of two 1 kΩ resistors, the noise level at once jumps up to −125.2 dBu, as the effective source resistance from two 1 kΩ resistors in parallel is 500 Ω. 10 dB of signal-to-noise ratio is irretrievably gone already, and we have only deployed two passive components. There will no doubt be more active and passive circuitry downstream, so things can only get worse.

TABLE 1.7 Johnson noise from 6 dB resistive divider with different resistor values (bandwidth 22 kHz, temperature 25 °C)

Divider R value	Divider R_{eff}	Johnson noise (dBu)	Relative noise (dB)
100 Ω	50 Ω	−135.2	−27.0
500 Ω	250 Ω	−128.2	−20.0
1 kΩ	500 Ω	−125.2	−17.0
5 kΩ	2500 Ω	−118.2	−10.0
10 kΩ	5 kΩ	−115.2	−7.0
50 kΩ	25 kΩ	−108.2	0 reference
100 kΩ	50 kΩ	−105.2	+3.0

However, a potential divider made from two 1 kΩ resistors in series presents an input impedance of only 2 kΩ, which is too low for most applications. 10 kΩ is normally considered the minimum input impedance for a piece of audio equipment in general use, which means we must use two 5 kΩ resistors, and so we get an effective source resistance of 2.5 kΩ. This produces Johnson noise at −118.2 dBu, so the signal-to-noise ratio has been degraded by another 7 dB simply by making the input impedance reasonably high.

In some cases 10 kΩ is not high enough, and a 100 kΩ input impedance is sought. Now the two resistors have to be 50 kΩ, and the noise is 10 dB higher again, at −108.2 dBu. That is a worrying 27 dB worse than our signal when it arrived.

If we insist on an input impedance of 100 kΩ, how can we improve on our noise level of −108.2 dBu? The answer is by buffering the divider from the outside world. The output noise of a 5532 voltage-follower is about −119 dBu with a 50 Ω input termination. If this is used to drive our attenuator, the two resistors in it can be as low as the opamp can drive. The 5532 has a most convenient combination of low noise and good load-driving ability, and the divider resistors can be reduced to 500 Ω each, giving a load of 1 kΩ and a generous safety margin of drive capability (pushing the 5532 to its specified limit of a 500 Ω load tends to degrade its superb linearity by a small but measurable amount). See Figure 1.7.

The noise from the resistive divider itself has now been lowered to −128.2 dBu, but there is of course the extra −119 dBu of noise from the voltage-follower that drives it. This however is halved by the divider just as the signal is, so the noise at the output will be the rms sum of −125 dBu and −n−128.2 dBu, which is −123.3 dBu. A 6 dB attenuator is actually the worst case, as it has the highest possible source impedance for a given total divider resistance. Either more or less attenuation will mean less noise from the divider itself.

So, despite adding active circuitry that intrudes its own noise, the final noise level has been reduced from −108.2 to −123.3 dBu, an improvement of 15.1 dB.

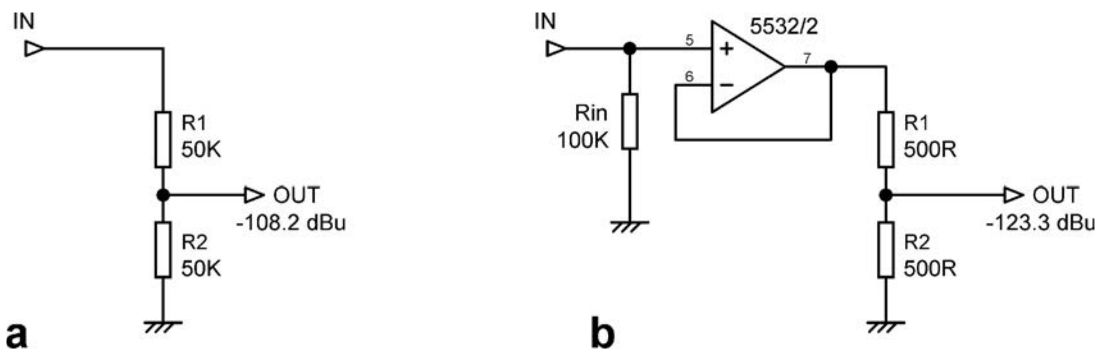

Figure 1.7: Two 6 dB attenuators with a 100 kΩ input impedance: a) simple attenuator with high resistor values, b) buffered attenuator with low resistor values. Despite the extra noise from the 5532 voltage follower this version is 15 dB quieter

How to amplify quietly

OK, we need a low-noise amplifier. Let's assume we have a reasonably low source impedance of 300 Ω, and we need a gain of four times (+12 dB). Figure 1.8a shows a very ordinary circuit using half a 5532 with typical values of 3 kΩ and 1 kΩ in the feedback network, and the noise output measures as −105.0 dBu. The Johnson noise generated by the 300 Ω source resistance is −127.4 dBu, and amplifying that by a gain of four gives −115.4 dBu. Compare this with the actual −105.0 dBu we get, and the noise figure is 10.4 dB – in other words the noise from the amplifier is three times the inescapable noise from the source resistance, making the latter essentially negligible. This amplifier stage is clearly somewhat short of noise-free perfection, despite using one of the quieter opamps around.

We need to make things quieter. The obvious thing to do is to reduce the value of the feedback resistances; this will reduce their Johnson noise and also reduce the noise produced in them by the opamp current noise generators. Figure 1.8b shows the feedback network altered to 360 Ω and 120 Ω, adding up to a load of 480 Ω, pushing the limits of the lowest resistance the opamp can drive satisfactorily. This assumes of course that the next stage presents a relatively light load so that almost all of the driving capability can be use to drive the negative-feedback network; keeping tiny signals free from noise can involve throwing some serious current about. The noise output is reduced to −106.1 dBu, which is only an improvement of 1.1 dB, and only brings the noise figure down to 9.3 dB, leaving us still a long way from what is theoretically attainable. However, at least it cost us nothing in extra components.

If we need to make things quieter yet, what can be done? The feedback resistances cannot be reduced further, unless the opamp drive capability is increased in some way. An output stage made of discrete transistors could be added, but it would almost certainly compromise the low distortion we get from a 5532 alone. For one answer see the next section on ultra-low noise design.

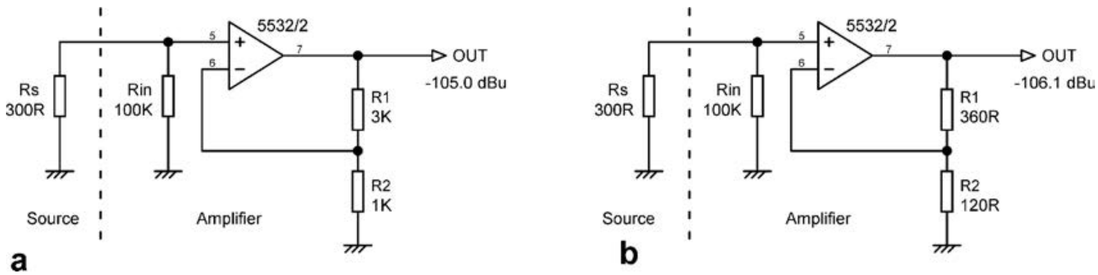

Figure 1.8: 4× amplifier a) with 'normal' feedback resistances, b) with low-impedance feedback arm resistances. Noise is only reduced by 1.1 dB

How to invert quietly

Inverting a signal always requires the use of active electronics (OK, you *could* use a transformer). Assume that an input impedance of 47 kΩ is required, along with a unity-gain inversion. A straightforward inverting stage as shown in Figure 1.9a will give this input impedance and gain only if both resistors are 47 kΩ. These relatively high value resistors contribute Johnson noise and exacerbate the effect of opamp current noise. Also the opamp is working at a noise gain of two times so the noise output is high at −101.4 dBu.

The only way to improve this noise level is to add another active stage. It sounds paradoxical – adding more non-silent circuitry to reduce noise – but that's the way the universe works. If a voltage-follower is added to the circuit given in Figure 1.9b, then the resistors around the inverting opamp can be greatly reduced in value without reducing the input impedance, which can now be pretty much as high as we like. The 'Noise buffered' column in Table 1.8 shows that if R1 and R2 are reduced to 2.2 kΩ, the total noise output is lowered by 8.2 dB, which is very useful improvement. If R1 and R2 are further reduced to 1 kΩ, which is perfectly practical with a 5532's drive capability, the total noise is reduced by 9.0 dB compared with the 47 kΩ case. The 'Noise unbuffered' column gives the noise output

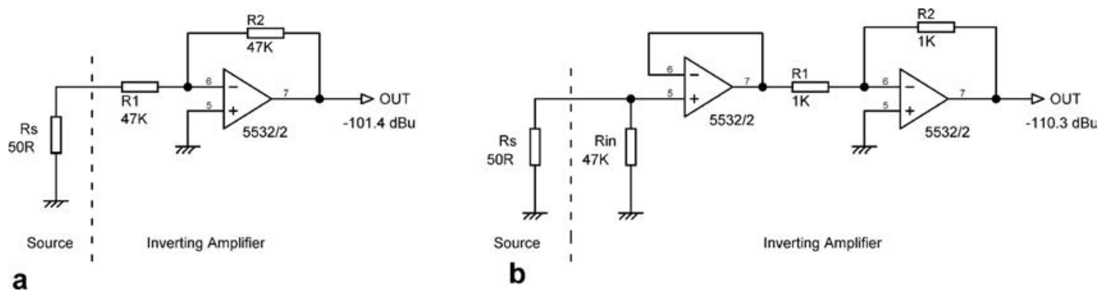

Figure 1.9: The noise from an inverter with 47 kΩ input impedance a) unbuffered, b) buffered and with low-value resistors

TABLE 1.8 Measured noise from simple inverter and buffered inverter (5532)

R value (Ω)	Noise unbuffered (dBu)	Noise buffered (dBu)	Noise reduction with reference to 47 kΩ case (dB)
1k	−111.0	−110.3	9.0
2k2	−110.1	−109.5	8.2
4k7	−108.9	−108.4	7.1
10k	−106.9	−106.6	5.3
22k	−104.3	−104.3	3.0
47k	−101.4	−101.3	0 reference

with specified R value but without the buffer, demonstrating that adding the buffer does degrade the noise slightly, but the overall result is still far quieter than the unbuffered version with 47 kΩ resistors. In each case the circuit input is terminated to ground via 50 Ω.

How to balance quietly

The design of low and ultra-low noise balanced amplifiers is thoroughly examined in Chapter 14 on line inputs.

Ultra low-noise design with multipath amplifiers

Are the above circuit structures the ultimate? Is this as low as noise gets? No. In the search for low-noise, a powerful technique is the use of parallel amplifiers with their outputs summed. This is especially useful where source impedances are low and therefore generate little noise compared with the voltage noise of the electronics.

If there are two amplifiers connected, the signal gain increases by 6 dB due to the summation. The noise from the two amplifiers is also summed, but since the two noise sources are completely uncorrelated (coming from physically different components) they partially cancel and the noise level only increases by 3 dB. Thus there is an improvement in signal-to-noise ratio of 3 dB. This strategy can be repeated by using four amplifiers, in which case the signal-to-noise improvement is 6 dB. Table 1.9 shows how this works for increasing numbers of amplifiers.

TABLE 1.9 How noise performance improves with multiple amplifiers

No. of amplifiers	Noise reduction (dB)
1	0 reference
2	−3.01
3	−4.77
4	−6.02
5	−6.99
6	−7.78
7	−8.45
8	−9.03
12	−10.79
16	−12.04
32	−15.05
64	−18.06
128	−21.07
256	−24.58

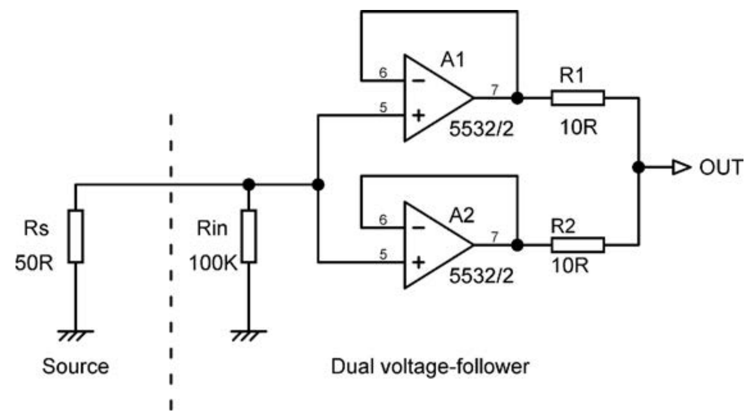

Figure 1.10: A double voltage-follower or buffer. The noise of this simple circuit is below that of the leading test equipment available

In practice the increased signal gain is not useful, and an active summing amplifier would compromise the noise improvement, so the output signals are averaged rather than summed as shown in Figure 1.10. The amplifier outputs are simply connected together with low-value resistors; so the gain is unchanged but the noise output falls. The amplifier outputs are nominally identical, so very little current should flow from one opamp to another. The combining resistor values are so low that their Johnson noise can be ignored.

Obviously there are economic limits on how far you can take this sort of thing. Unless you're measuring gravity waves or something equally important, 256 parallel amplifiers is probably not a viable choice.

Be aware that this technique does not give any kind of fault redundancy. If one opamp turns up its toes, the low value of the averaging resistors means the whole stage will stop working.

Ultra low-noise voltage buffers

The multiple-path philosophy works well even with a minimally simple circuit such as a unity-gain voltage buffer. Table 1.10 gives calculated results for 5532 sections (the noise output is too low to measure reliably even with the best test gear) and shows how the noise output falls as more opamps are added. The distortion performance is not affected.

The 10 Ω output resistors combine the opamp outputs, and limit the currents that would flow from output to output as a result of DC offset errors. AC gain errors here will be very

TABLE 1.10 Noise from parallel-array buffers using 5532 sections

Number of opamps	Calculated noise out (dBu)
1	−120.4
2	−123.4
3	−125.2
4	−126.4

small indeed as the opamps have 100% feedback. If the output resistors were raised to 47 Ω they would as usual give HF stability when driving screened cables or other capacitances, but the total output impedance is usefully halved to 23.5 Ω. Another interesting bonus of this technique is that we have doubled the output drive capability; this stage can easily drive 300 Ω. This can be very useful when using low-impedance design to reduce noise in the following stage.

Ultra low-noise amplifiers

We now return to the problem studied earlier; how to make a really quiet amplifier with a gain of four times. We saw that the minimum noise output using a single 5532 section and a 300 Ω source resistance was −106.1 dBu, with a not particularly impressive noise figure of 9.3 dB. Since almost all the noise is being generated in the amplifier rather than the source resistance, the multiple-path technique should work well here. And it does.

There is however, a potential snag that needs to be considered. In the previous section, we were combining the outputs of voltage followers, which have gains very close indeed to unity because they have 100% negative feedback and no resistors are involved in setting the gain. We could be confident that the output signals would be near identical and unwanted currents flowing from one opamp to the other would be small despite the low value of the combining resistors.

The situation here is different: the amplifiers have a gain of four times, so there is a smaller negative feedback factor to stabilise the gain, and there are two resistors with tolerances that set the closed-loop gain for each stage. We need to keep the combining resistors low to minimise their Johnson noise, so things might get awkward. It seems reasonable to assume that the feedback resistors will be 1% components. Considering the two-amplifier configuration in Figure 1.11, the worst case would be to have R1A 1% high and R2A 1% low in one amplifier, while the other had the opposite condition of R1B 1% low and R2B 1% high. This highly unlikely state of affairs gives a gain of 4.06 times

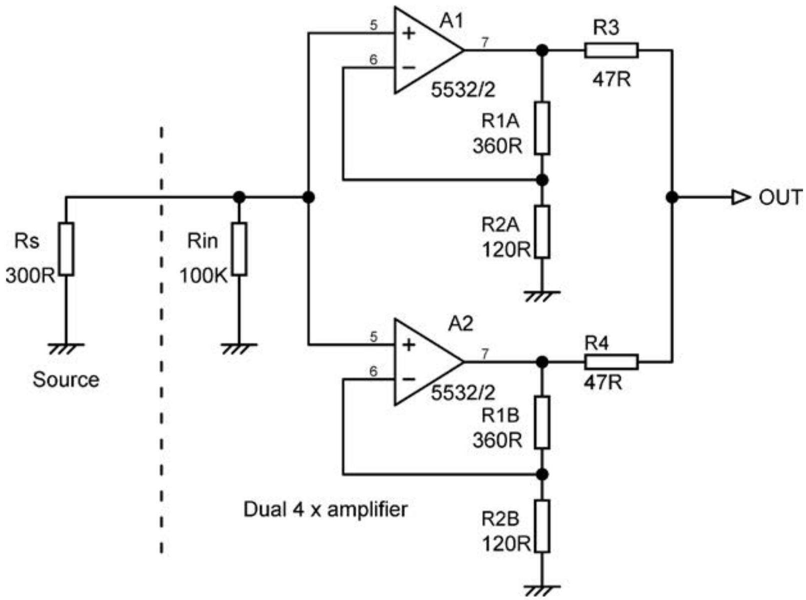

Figure 1.11: A 4× amplifier using two opamps to reduce noise by approaching 3 dB

in the first amplifier and 3.94 times in the second. Making the further assumption of a 10 Vrms maximum voltage swing, we get 10.15 Vrms at the first output and 9.85 Vrms at the second, both applied to the combining resistors, which here are set at 47 Ω. The maximum possible current flowing from one amplifier output into the other is therefore 0.3 V/(47 Ω + 47 Ω) which is 3.2 mA; in practice it will be much smaller. There are no problems with linearity or headroom, and distortion performance is indistinguishable from that of a single opamp.

Having reassured ourselves on this point, we can examine the circuit of Figure 1.11, with two amplifiers combining their outputs. This reduces the noise at the output by 2.2 dB. This falls short of the 3 dB improvement we might hope for because of a significant Johnson noise contribution from source resistance, and doubling the number of amplifier stages again only achieves another 1.3 dB improvement. The improvement is greater with lower source resistances; the measured results with 1, 2, 3 and 4 opamps for three different source resistances are summarised in Table 1.11.

The results for 200 Ω and 100 Ω show that the improvement with multiple amplifiers is greater for lower source resistances, as these resistances generate less Johnson noise of their own.

TABLE 1.11 Noise from multiple amplifiers with 4× gain

R_s (Ω)	No. of opamps	Noise out (dBu)	Improvement (dB)
300	1	−106.1	0 reference
300	2	−108.2	2.2
300	3	−109.0	2.9
300	4	−109.6	3.5
200	1	−106.2	0 reference
200	2	−108.4	2.2
200	3	−109.3	3.1
200	4	−110.0	3.8
100	1	−106.3	0 reference
100	2	−108.7	2.4
100	3	−109.8	3.5
100	4	−110.4	3.9

Multiple amplifiers for greater drive capability

We have just seen that the use of multiple amplifiers with averaged outputs not only reduces noise but increases the drive capability proportionally to the number of amplifiers employed. This is highly convenient as heavy loads need to be driven when pushing hard the technique of low-impedance design. There is more on this in Chapter 15, relating to low-impedance Baxandall tone-controls.

Using multiple amplifiers gets difficult when the stage has variable feedback to implement gain control or tone control. In this case the configuration in Figure 1.12 doubles the drive capability in a fool-proof manner – I have always called it ‘mother’s little helper’. A1 may be enmeshed in as complicated a circuit as you like, but unity-gain buffer A2 will robustly carry out its humble duty of sharing the load. This is unlikely to give any noise advantage, as most of the noise will presumably come from the more complex circuitry around A1.

It is assumed that A1 has load-driving capabilities equivalent to those of A2. This approach is more parts-efficient than simply putting a multiple-buffer like that in Figure 1.10 after A1; that would make no use of the drive capability of A1. This technique was used to drive the input of a Baxandall volume control using 1 kΩ pots in the Elektor 2012 preamplifier design [11].

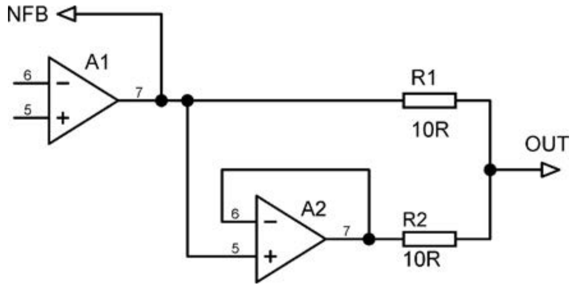

Figure 1.12: Mother's little helper. Using unity-gain buffer A2 to double the drive capability of any opamp stage

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CHAPTER 2

Components

Conductors

It is easy to assume, when wrestling with electronic design, that the active devices will cause most of the trouble. This, like so much in electronics, is subject to Gershwin's Law; 'it ain't necessarily so'. Passive components cannot be assumed to be perfect, and while their shortcomings are rarely discussed, in polite company, they are all too real. In this chapter I have tried to avoid repeating basic stuff that can be found in many other places, to allow room for information that goes deeper.

Normal metallic conductors, such as copper wire, show perfect linearity for our purposes, and as far as I am aware, for everybody's purposes. Ohm's Law was founded on metallic conductors, after all; not resistors, which did not exist as we know them at the time. George Simon Ohm published a pamphlet in 1827, entitled 'The Galvanic Circuit Investigated Mathematically', while he was a professor of mathematics in Cologne. His work was not warmly received, except by a perceptive few; the Prussian minister of education pronounced that 'a professor who preached such heresies was unworthy to teach science'. This is the sort of thing that happens when politicians try to involve themselves in science, and in that respect we have progressed little since then.

Although the linearity is generally effectively ideal, metallic conductors will not be perfectly linear in some circumstances. Poorly-made connections between oxidised or otherwise contaminated metal parts are capable of generating harmonic distortion at the level of several per cent, but this is a property of the contact interface rather than the bulk material, and usually means that the connection is about to fail altogether. A more subtle danger is that of magnetic conductors – the soft iron in relay frames causes easily detectable distortion at power-amplifier current levels.

From time to time some of the dimmer audio commentators speculate that metallic conductors are actually a kind of 'sea of micro-diodes', and that non-linearity can be found if the test signal levels are made small enough. This is both categorically untrue and physically impossible. There is no threshold effect for metallic conduction. I have myself added to the mountain of evidence on this, by measuring distortion at very low signal levels [1]. Renardsen has some more information at [2].

Copper and other conductive elements

Copper is the preferred metal for conducting electricity in almost all circumstances. It has the lowest resistance of any metal but silver, is reasonably resistant to corrosion, and can be made mechanically strong; it's wonderful stuff. Being a heavy metal, it is unfortunately not that common in the earth's crust, and so is expensive compared with iron and steel. It is however cheap compared with silver. The price of metals varies all the time due to changing economic and political factors, but at the time of writing silver was 100 times more expensive than copper by weight. Given the same cross-section of conductor, the use of silver would only reduce the resistance of a circuit by 5%. Despite this, silver connection wire has been used in some very expensive hifi amplifiers; output impedance-matching transformers wound with silver wire are not unknown in valve amplifiers. Since the technical advantages are usually negligible such equipment is marketed on the basis of indefinable subjective improvements. The only exception is the moving-coil step-up transformer, where the use of silver in the primary winding might give a measurable reduction in Johnson noise.

Table 2.1 gives the resistivity of the commonly-used conductors, plus some insulators to give it perspective. The difference between copper and quartz is of the order of 10 to the 25, an enormous range that is not found in many other physical properties. There are several reasonably conductive metals that are lighter than copper, but their higher resistivity means they require larger cross-sections to carry the same current, so copper is always used when space is limited, as in electric motors, solenoids, etc. However, when size is not the primary constraint the economics work out differently. The largest use of non-copper conductors is probably in the transmission line cables that are strung between pylons. Here minimal weight is more important than minimal diameter, so the cables have a central steel core for strength, surrounded by aluminium conductors.

It is clear that simply spending more money does not automatically bring you a better conductor; gold is a somewhat poorer conductor than copper, and platinum, which is even more expensive, is worse by a factor of six. Another interesting feature of this table is the relatively high resistance of mercury, nearly 60 times that of copper. This often comes as a surprise; people seem to assume that a metal of such high density must be very conductive, but it is not so. There are many reasons for not using mercury-filled hoses as loudspeaker cables, and their conductive inefficiency is just one. The cost and the insidiously poisonous nature of the metal are two more. Nonetheless, it is reported that the Hitachi Cable company has experimented with speaker cables made from polythene tubes filled with mercury. There appear to have been no plans to put such a product on the market. RoHS compliance might be a problem.

We also see that the resistivity of solder is high compared with that of copper – nine times higher if you compare copper with the 60/40 tin/lead solder. This is unlikely to be a problem as the thickness of solder the current passes through in a typical joint is very small. There are many formulations of lead-free solder, with varying resistivities, but all are high compared with copper.

TABLE 2.1 Properties of conductors and non-conductors

Material	Resistivity ρ (Ω–m)	Temperature coefficient per degree C	Electrical usage
Silver	1.59×10^{-8}	0.0061	conductors
Copper	1.72×10^{-8}	0.0068	conductors
Gold	2.2×10^{-8}	0.0041	inert coatings
Aluminium	2.65×10^{-8}	0.00429	conductors
Tungsten	5.6×10^{-8}	0.0045	lamp filaments
Iron	9.71×10^{-8}	0.00651	barreters*
Platinum	10.6×10^{-8}	0.003927	electrodes
Tin	11.0×10^{-8}	0.0042	coatings
Mild steel	15×10^{-8}	0.0066	busbars
Solder (60:40 tin/lead)	15×10^{-8}	0.006	soldering
Lead	22×10^{-8}	0.0039	storage batteries
Manganin (Cu, Mn, Ni)**	48.2×10^{-8}	0.000002	resistances
Constantan (Cu, Ni)**	$49 - 52 \times 10^{-8}$	+/−0.00002	resistances
Mercury	98×10^{-8}	0.0009	relays
Nichrome (Ni, Fe, Cr alloy)	100×10^{-8}	0.0004	heating elements
Carbon (as graphite)	$3 - 60 \times 10^{-5}$	−0.0005	brushes
Glass	$1 - 10000 \times 10^{9}$	. . .	insulators
Fused quartz	More than 10^{18}	. . .	insulators

* A barreter is an incredibly obsolete device consisting of thin iron wire in an evacuated glass envelope. It was typically used for current regulation of the heaters of RF oscillator valves, to improve frequency stability.

** Constantan and manganin are resistance alloys with moderate resistivity and a low temperature coefficient. Constanan is preferred as it has a flatter resistance/temperature curve and its corrosion resistance is better.

The metallurgy of copper

Copper is a good conductor because the outermost electrons of its atoms have a large mean free path between collisions. The electrical resistivity of a metal is inversely related to this electron mean free path, which in the case of copper is approximately 100 atomic spacings.

Copper is normally used as a very dilute alloy known as electrolytic tough pitch (ETP) copper, which consists of very high purity metal alloyed with oxygen in the range of 100 to 650 ppm. In view of the wide exposure that the concept of oxygen-free copper has had in the audio business, it is worth underlining that the oxygen is deliberately alloyed with the copper to act as a scavenger for dissolved hydrogen and sulphur, which become water and sulphur dioxide. Microscopic bubbles form in the mass of metal but are completely eliminated during hot rolling. The main use of oxygen-free copper is in conductors exposed to a hydrogen atmosphere at high temperatures. ETP copper is susceptible to hydrogen embrittlement in these circumstances, which arise in the hydrogen-cooled alternators in power stations.

Gold and its uses

As stated above, gold has a higher resistivity than copper, and there is no incentive to use it as the bulk metal of conductors, not least because of its high cost. However it is very useful as a thin coating on contacts because it is almost immune to corrosion, though it is chemically attacked by fluorine and chlorine (if there is a significant amount of either gas in the air then your medical problems will be more pressing than your electrical ones). Other electrical components are sometimes gold-plated simply because the appearance is attractive. A carat (or karat) is a 1/24 part, so 24-carat gold is the pure element, while 18-carat gold contains only 75% of the pure metal.18-carat gold is the sort usually used for jewellery as it retains the chemical inertness of pure gold but is much harder and more durable; the usual alloying elements are copper and silver.

18-carat gold is widely used in jewellery and does not tarnish, so it is initially puzzling to find that some electronic parts plated with it have a protective transparent coating which the manufacturer claims to be essential to prevent blackening. The answer is that if gold is plated directly onto copper, the copper diffuses through the gold and tarnishes on its surface. The standard way of preventing this is to plate a layer of nickel onto the copper to prevent diffusion, then plate on the gold. I have examined some transparent-coated gold-plated parts and found no nickel layer; presumably the manufacturer finds the transparent coating is cheaper than another plating process to deposit the nickel. However, it does not look as good as bare gold.

Cable and wiring resistance

Electrical cable is very often specified by its cross-sectional area and current-carrying capacity, and the resistance per meter is seldom quoted. This can however be a very important parameter for assessing permissible voltage drops and for predicting the crosstalk that will be introduced between two signals when they unavoidably share a common ground conductor. Given the resistivity of copper from Table 2.1, the resistance R of L metres of cable is simply:

$$R = \frac{resistivity \cdot L}{area} \qquad \text{(Equation 2.1)}$$

Note that the area, which is usually quoted in catalogues in square millimetres, must be expressed here in square metres to match up with the units of resistivity and length. Thus 5 metres of cable with a cross-sectional area of $1.5 mm^2$ will have a resistance of:

$$(1.72 \times 10^{-8}) \times 5/(0.0000015) = 0.057 \text{ Ohms}$$

This gives the resistance of our stretch of cable, and it is then simple to treat this as part of a potential divider to calculate the voltage drop down its length.

PCB track resistance

It is also useful to be able to calculate the resistance of a PCB track for the same reasons. This is slightly less straightforward to do; given the smorgasbord of units that are in use in PCB technology, determining the cross-sectional area of the track can present some difficulty.

In the USA and the UK, and probably elsewhere, there is inevitably a mix of metric and imperial units on PCBs, as many important components come in dual in-line packages which are derived from an inch grid; track widths and lengths are therefore very often in thousands of an inch, universally (in the UK at least) referred to as 'thou'. Conversely, the PCB dimensions and fixing-hole locations will almost certainly be metric as they interface with a world of metal fabrication and mechanical CAD that (except the USA) went metric many years ago. Add to this the UK practice of quoting copper thickness in ounces (the weight of a square foot of copper foil) and all the ingredients for dimensional confusion are in place.

Standard PCB copper foil is known as one ounce copper, having a thickness of 1.4 thou (= 35 microns). Two-ounce copper is naturally twice as thick; the extra cost of specifying it is small, typically around 5% of the total PCB cost, and this is a very simple way of halving track resistance. It can of course be applied very easily to an existing design without any fear of messing up a satisfactory layout. Four-ounce copper can also be obtained but is more rarely used and is therefore much more expensive. If heavier copper than two-ounce is required, the normal technique is to plate two-ounce up to three-ounce copper. The extra cost of this is surprisingly small, in the region of 10% to 15%.

Given the copper thickness, multiplying by track width and length gives the cross-sectional area. Since resistivity is always in metric units, it is best to convert to metric at this point, so the table below gives area in square millimetres. This is then multiplied by the resistivity, not forgetting to convert the area to metres for consistency. This gives the 'resistance' column in the table, and it is then simple to treat this as part of a potential divider to calculate the usually unwanted voltage across the track.

For example, if the track in question is the ground return from an 8 Ω speaker load, this is the top half of a potential divider while the track is the bottom half, (I am of course ignoring here the fact loudspeakers are not purely resistive loads) and a quick calculation gives the fraction of the input voltage found along the track. This is expressed in the last column of Table 2.2 as attenuation in dB. This shows clearly that loudspeaker outputs should not have common return tracks or the interchannel crosstalk will be dire.

It is very clear from this table that relying on thicker copper on your PCB as a means of reducing path resistance is not very effective. In some situations it may be the only recourse, but in many cases a path of much lower resistance can be made by using 32/02 cable soldered between the two relevant points on the PCB.

TABLE 2.2 Thickness of copper cladding and the calculation of track resistance

Weight (oz)	Thickness (thou)	Thickness (micron)	Width (thou)	Length (inch)	Area (mm²)	Resistance (Ω)	Attenuation ref 8 Ω (dB)
1	1.38	35	12	3	0.0107	0.123	−36.4
1	1.38	35	50	3	0.0444	0.029	−48.7
2	2.76	70	12	3	0.0213	0.061	−42.4
2	2.76	70	50	3	0.0889	0.015	−54.7
4	5.52	140	50	3	0.178	0.0074	−60.7

TABLE 2.3 PCB track current capacity for a permitted temperature rise

	+10 °C		+20 °C		+30 °C	
	Copper weight		Copper weight		Copper weight	
Track width (thou)	1 oz (A)	2 oz (A)	1 oz (A)	2 oz (A)	1 oz (A)	2 oz (A)
10	1.0	1.4	1.2	1.6	1.5	2.2
15	1.2	1.6	1.3	2.4	1.6	3.0
20	1.3	2.1	1.7	3.0	2.4	3.6
25	1.7	2.5	2.2	3.3	2.8	4.0
30	1.9	3.0	2.5	4.0	3.2	5.0
50	2.6	4.0	3.6	6.0	4.4	7.3
75	3.5	5.7	4.5	7.8	6.0	10.0
100	4.2	6.9	6.0	9.9	7.5	12.5
200	7.0	11.5	10.0	16.0	13.0	20.5
250	8.3	12.3	12.3	20.0	15.0	24.5

PCB tracks have a limited current capability because excessive resistive heating will break down the adhesive holding the copper to the board substrate, and ultimately melt the copper. This is normally only a problem in power amplifiers and power supplies. It is useful to assess if you are likely to have problems before committing to a PCB design, and Table 2.3, based on MIL-standard 275, gives some guidance.

Note that Table 2.3 applies to tracks on the PCB surface only. Internal tracks in a multi-layer PCB experience much less cooling, and need to be about three times as thick for the same temperature rise. This factor depends on laminate thickness and so on, and you need to consult your PCB vendor.

Traditionally, overheated tracks could be detected visually because the solder mask on top of them would discolour to brown. I am not sure if this still applies with modern solder mask materials, as in recent years I have been quite successful in avoiding overheated tracking.

PCB track-to-track crosstalk

The previous section described how to evaluate the amount of crosstalk that can arise because of shared track resistances. Another crosstalk mechanism is caused by capacitance between PCB tracks. This is not very susceptible to calculation, so I did the following experiment to put some figures to the problem.

Figure 2.1 shows the setup; four parallel conductors, 1.9 inches long on a standard piece of 0.1 inch pitch prototype board were used as test tracks. These are perhaps rather wider than the average PCB track, but one must start somewhere. The test signal was applied to track A, and track C was connected to a virtual-earth summing amplifier A1.

The tracks B and D were initially left floating. The results are shown as Trace 1 in Figure 2.2; the coupling at 10 kHz is −65 dB, which is worryingly high for two tracks 0.2 inch apart. Note that the crosstalk increases steadily at 6 dB per octave, as it results from a very small capacitance driving into what is effectively a short circuit.

It has often been said that running a grounded screening track between two tracks that are susceptible to crosstalk has a beneficial effect, but how much good does it really do? Grounding track B, to place a screen between A and C, gives Trace 2 and has only improved matters by 9 dB; not the dramatic effect that might be expected from screening. The reason, of course, is that electric fields are very much three-dimensional, and if you could see the

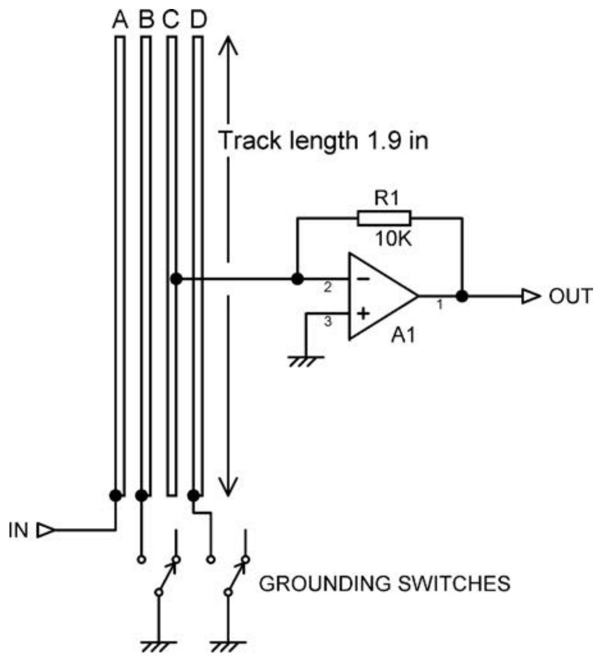

Figure 2.1: Test circuit for measuring track-to-track crosstalk on a PCB

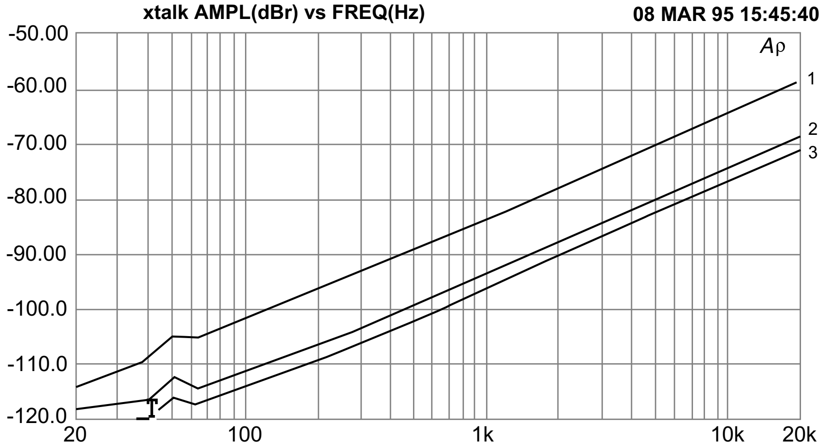

Figure 2.2: Results of PCB track-to-track crosstalk tests

electrostatic 'lines of force' that appear in physics textbooks you would notice they arch up and over any planar screening such as a grounded track. It is easy to forget this when staring at a CAD display. There are of course two-layer and multi-layer PCBs, but the visual effect on a screen is still of several slices of 2-D. As Mr Spock remarked in one of the Star Trek films, 'He's intelligent, but not experienced. His pattern indicates two-dimensional thinking.'

Grounding track D, beyond receiving track C, gives a further improvement of about 3 dB (Trace 3); this would clearly not happen if PCB crosstalk was simply a line-of-sight phenomenon.

To get more effective screening than this you must go into three dimensions too; with a double-sided PCB you can put one track on each side, with ground plane opposite. With a 4-layer board it should be possible to sandwich critical tracks between two layers of ground plane, where they should be safe from pretty much anything. If you can't do this and things are really tough you may need to resort to a screened cable between two points on the PCB; this is of course expensive in assembly time. If components, such as electrolytics with their large surface area, are talking to each other you may need to use a vertical metal wall, but this costs money. A more cunning plan is to use electrolytics not carrying signal, such as rail decouplers, as screening items.

The internal crosstalk between the two halves of a dual opamp is very low according to the manufacturer's specs. Nevertheless, avoid having different channels going through the same opamp if you can because this will bring the surrounding components into close proximity, and will permit capacitive crosstalk.

Impedances and crosstalk: a case history

Capacitive crosstalk between two opamp outputs can be surprisingly troublesome. The usual isolating resistor on an opamp output is 47 Ω, and you might think that this impedance is so low that the capacitive crosstalk between two of these outputs would be completely negligible, but . . . you would be wrong.

A stereo power amplifier had balanced input amplifiers with 47 Ω output isolating resistors included to prevent any possibility of instability although the opamps were driving only a few cm of PCB track rather than screened cables with their significant capacitance. Just downstream of these opamps was a switch to enable bi-amping by driving both left and right outputs with the left input. This switch and its associated tracking brought the left and right signals into close proximity, and the capacity between them was not negligible.

Crosstalk at low frequencies (below 1 kHz) was pleasingly low, being better than −129 dB up to 70Hz, which was the difference between the noise floor and the maximum signal level (the measured noise floor was unusually low at −114 dBu because each input amplifier was a quadruple noise cancelling type as described in Chapter 18, and that figure includes the noise from an AP System 1). At higher frequencies things were rather less gratifying, being −96 dB at 10 kHz, as shown by the '47R' trace in Figure 2.3. In many applications this would be more than acceptable, but in this case the highest performance possible was being sought.

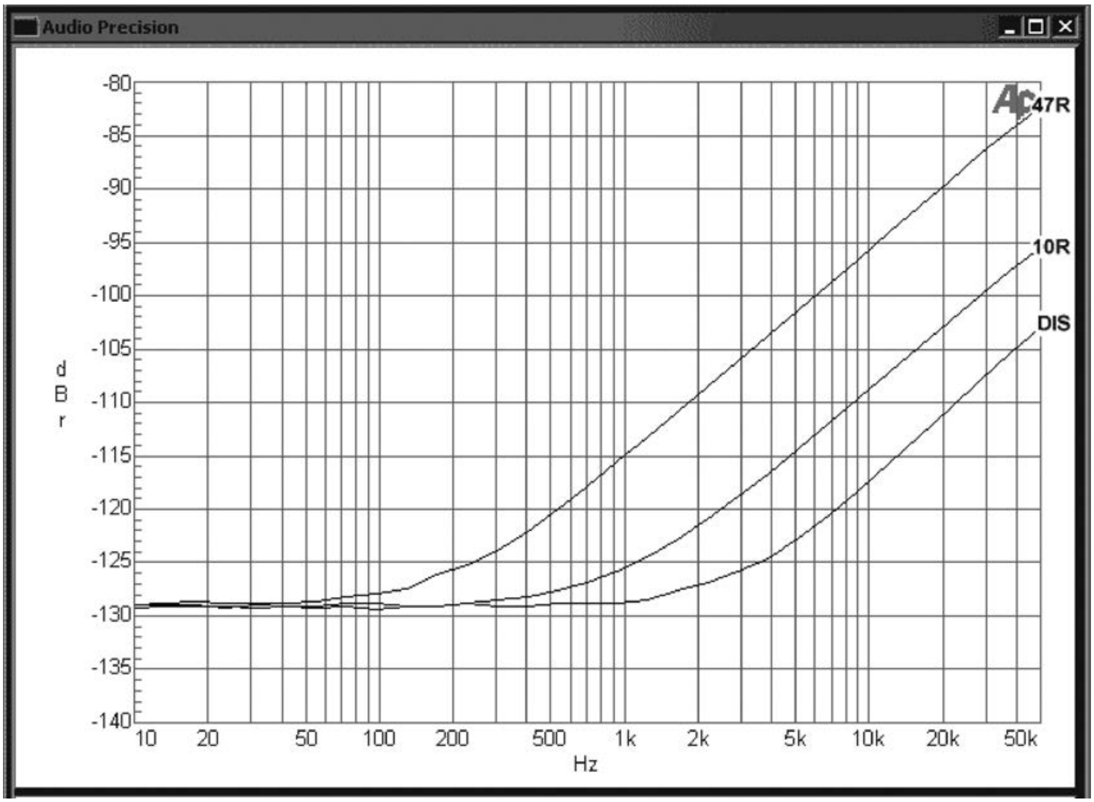

Figure 2.3: Crosstalk between opamp outputs with 47 Ω and 10 Ω output isolating resistors

I therefore decided to reduce the output isolating resistors to 10 Ω, so the inter-channel capacitance would have less effect (checks were done at the time and all through the prototyping and pre-production process to make sure that this would be enough resistance to ensure opamp stability – it was). This handily reduced the crosstalk to −109 dB at 10 kHz, an improvement of 13 dB at zero cost. This is the ratio between the two resistor values.

The third trace marked ‘DIS’ shows the result of removing the isolating resistor from the speaking channel, so no signal reached the bi-amping switch. As usual, this reveals a further crosstalk mechanism, at about −117 dB, for reducing crosstalk is proverbially like peeling onions. There is layer after layer, and even strong men are reduced to tears.

Resistors

In the past there have been many types of resistor, including some interesting ones consisting of jars of liquid, but only a few kinds are likely to be met with now. These are usually classified by the kind of material used in the resistive element, as this has the most important influence on the fine details of performance. The major materials and types are shown in Table 2.4.

These values are illustrative only, and it would be easy to find exceptions. As always, the official data sheet for the component you have chosen is the essential reference. The voltage coefficient is a measure of linearity (lower is better) and its sinister significance is explained later.

It should be said that you are most unlikely to come across carbon composition resistors in modern signal circuitry, but they frequently appear in vintage valve equipment so they are included here. They also live on in specialised applications such as switch-mode snubbing circuits, where their ability to absorb a high peak power in a mass of material rather than a thin film is very useful.

TABLE 2.4 Characteristics of resistor types

Type	Resistance tolerance (%)	Temperature coefficient (ppm/°C)	Voltage coefficient (ppm)
Carbon composition	±10	+400 to −900	350
Carbon film	±5	−100 to −700	100
Metal film	±1	+100	1
Metal oxide	±5	+300	variable but too high
Wirewound	±5	±70 to ±250	1

Carbon film resistors are currently still sometimes used in low-end consumer equipment, but elsewhere have been supplanted by the other three types. Note from Table 2.4 that they have a significant voltage coefficient.

Metal film resistors are now the usual choice when any degree of precision or stability is required. These have no non-linearity problems at normal signal levels. The voltage coefficient is usually negligible.

Metal oxide resistors are more problematic. Cermet resistors and resistor packages are metal oxide, and are made of the same material as thick-film SM resistors. Thick-film resistors can show significant non-linearity at opamp-type signal levels, and should be kept out of high-quality signal paths.

Wirewound resistors are indispensable when serious power needs to be handled. The average wirewound resistor can withstand very large amounts of pulse power for short periods, but in this litigious age component manufacturers are often very reluctant to publish specifications on this capability, and endurance tests have to be done at the design stage; if this part of the system is built first then it will be tested as development proceeds. The voltage coefficient is usually negligible.

Resistors for general PCB use come in both through-hole and surface-mount types. Through-hole (TH) resistors can be any of the types tabled above; surface-mount (SM) resistors are always either metal film or metal oxide. There are also many specialised types; for example, high-power wirewound resistors are often constructed inside a metal case that can be bolted down to a heatsink.

Through-hole resistors

These are too familiar to require much description; they are available in all the materials mentioned above; carbon film, metal film, metal oxide, and wirewound. There are a few other sorts, such as metal foil, but they are restricted to specialised applications. Conventional through-hole resistors are now almost always 250 mW 1% metal film. Carbon film used to be the standard resistor material, with the expensive metal film resistors reserved for critical places in circuitry where low temperatures and an absence of excess noise were really important, but as metal film got cheaper so it took over many applications.

TH resistors have the advantage that their power and voltage rating greatly exceed those of surface-mount versions. They also have a very low voltage coefficient, which for our purposes is of the first importance. On the downside, the spiral construction of the resistance element means they have much greater parasitic inductance; this is not a problem in audio work.

Surface-mount resistors

Surface-mount resistors come in two main formats, the common chip type, and the rarer (and much more expensive) MELF format.

Chip surface-mount (SM) resistors come in a flat tombstone format, which varies over a wide size range; see Table 2.5.

MELF surface-mount resistors have a cylindrical body with metal endcaps, the resistive element is metal film, and the linearity is therefore as good as conventional resistors, with a voltage coefficient of less than 1 ppm. MELF is apparently an acronym for 'Metal ELectrode Face-bonded' though most people I know call them 'Metal Ended Little Fellows' or something quite close to that.

Surface-mount resistors may have thin-film or thick-film resistive elements. The latter are cheaper and so more often encountered, but the price differential has been falling in recent years. Both thin film and thick film SM resistors use laser trimming to make fine adjustments of resistance value during the manufacturing process. There are important differences in their behaviour.

Thin film (metal film) SM resistors use a nickel-chromium (Ni-Cr) film as the resistance material. A very thin Ni-Cr film of less than 1 um thickness is deposited on the aluminium oxide substrate by sputtering under vacuum. Ni-Cr is then applied onto the substrate as conducting electrodes. The use of a metal film as the resistance material allows thin film resistors to provide a very low temperature coefficient, much lower current noise and vanishingly small non-linearity. Thin film resistors need only low laser power for trimming (one-third of that required for thick film resistors) and contain no glass-based material. This prevents possible micro-cracking during laser trimming and maintains the stability of the thin film resistor types.

TABLE 2.5 The standard surface-mount resistor sizes with typical ratings

Size L × W	Maximum power dissipation	Maximum voltage (V)
2512	1 W	200
1812	750 mW	200
1206	250 mW	200
0805	125 mW	150
0603	100 mW	75
0402	100 mW	50
0201	50 mW	25
01005	30 mW	15

Thick film resistors normally use ruthenium oxide (RuO_2) as the resistance material, mixed with glass-based material to form a paste for printing on the substrate. The thickness of the printing material is usually 12 μm. The heat generated during laser trimming can cause micro-cracks on a thick film resistor containing glass-based materials which can adversely affect stability. Palladium/silver (PdAg) is used for the electrodes.

The most important thing about thick-film surface-mount resistors from our point of view is that they do not obey Ohm's Law very well. This often comes as a shock to people who are used to TH resistors, which have been the highly linear metal film type for many years. They have much higher voltage coefficients than TH resistors, at between 30 and 100 ppm. The non-linearity is symmetrical about zero voltage and so gives rise to third-harmonic distortion. Some SM resistor manufacturers do not specify voltage coefficient, which usually means it can vary disturbingly between different batches and different values of the same component, and this can have dire results on the repeatability of design performance.

Chip-type surface-mount resistors come in standard formats with names based on size, such as 1206, 0805, 0603 and 0402. For example, 0805, which used to be something like the 'standard' size, is 0.08 in by 0.05 in; see Table 2.5. The smaller 0603 is now more common. Both 0805 and 0603 can be placed manually if you have a steady hand and a good magnifying glass.

The 0402 size is so small that the resistors look rather like grains of pepper; manual placing is not really feasible. They are only used in equipment where small size is critical, such as mobile phones. They have very restricted voltage and power ratings, typically 50 V and 100 mW. The voltage rating of TH resistors can usually be ignored, as power dissipation is almost always the limiting factor, but with SM resistors it must be kept firmly in mind.

Recently, even smaller surface-mount resistors have been introduced; for example several vendors offer 0201, and Panasonic and Yageo offer 01005 resistors. The latter are truly tiny, being about 0.4 mm long; a thousand of them weigh less than a twentieth of a gram. They are intended for mobile phones, palmtops, and hearing aids; a full range of values is available from 10 Ω to 1 MΩ (jumper inclusive). Hand placing is really not an option.

Surface-mount resistors have a limited power-dissipation capability compared with their through-hole cousins, because of their small physical size. SM voltage ratings are also restricted, for the same reason. It is therefore sometimes necessary to use two SM resistors in series or parallel to meet these demands, as this is usually more economic than hand-fitting a through-hole component of adequate rating. If the voltage rating is the issue then the SM resistors will obviously have to be connected in series to gain any benefit.

Resistor accuracy

As noted in Table 2.4, the most common tolerance for metal film resistors today is 1%. If you want a closer tolerance then the next that is readily available is 0.1%; while there is considerable variation in price, roughly speaking the 0.1% resistors will be ten times as expensive.

Resistors are widely available in the E24 series (24 values per decade) and the E96 series (96 values per decade). There is also the E192 series (you guessed it, 192 values per decade) but this is less freely available. Using the E96 or E192 series means you have to keep a lot of different resistor values in stock; when non-E24 values are required it is usually more convenient to use a series or parallel combination of two E24 resistors.

Using two or more resistors to make up a desired value has a valuable hidden benefit. If it is done correctly it will actually increase the average accuracy of the total resistance value so it is *better* than the tolerance of the individual resistors; this may sound paradoxical but it is simply an expression of the fact that random errors tend to cancel out if you have a number of them. This also works for capacitors, and indeed any parameter that is subject to random variations, but for the time being we will focus on the concrete example of multiple resistors. Note that this assumes that the mean (i.e. average) value of the resistors is accurate. It is generally a sound assumption as it is much easier to control a single value such as the mean in a manufacturing process than to control all the variables that lead to scatter about that mean. This is confirmed by measurement.

Component values are usually subject to a Gaussian distribution, also called a normal distribution. It has a familiar peaky shape, not unlike a resonance curve, showing that the majority of the values lie near the central mean, and that they get rarer the further away from the mean you look. This is a very common distribution, cropping up wherever there are many independent things going on that affect the value of a given component. The distribution is defined by its mean, and its standard deviation, which is the square-root of the sum of the squares of the distances from the mean – the RMS-sum, in other words. Sigma (σ) is the standard symbol for standard deviation. A Gaussian distribution will have 68.3% of its values within $\pm 1\ \sigma$, 95.4% within $\pm 2\ \sigma$, 99.7% within $\pm 3\ \sigma$, and 99.9% within $\pm 4\ \sigma$. This is illustrated in Figure 2.4, where the X-axis is calibrated in numbers of standard deviations on either side of the central mean value.

If we put two equal-value resistors in series, or in parallel, (see Figures 2.5a and 2.5b) the total value has proportionally a narrower distribution than that of the original components. The standard deviation of summed components is the sum of the squares of the individual standard deviations, as shown in Equation 2.2. σ_{sum} is the overall standard deviation, and σ_1 and σ_2 are the standard deviations of the two resistors in series.

$$\sigma_{sum} = \sqrt{(\sigma_1)^2 + (\sigma_2)^2} \qquad \text{(Equation 2.2)}$$

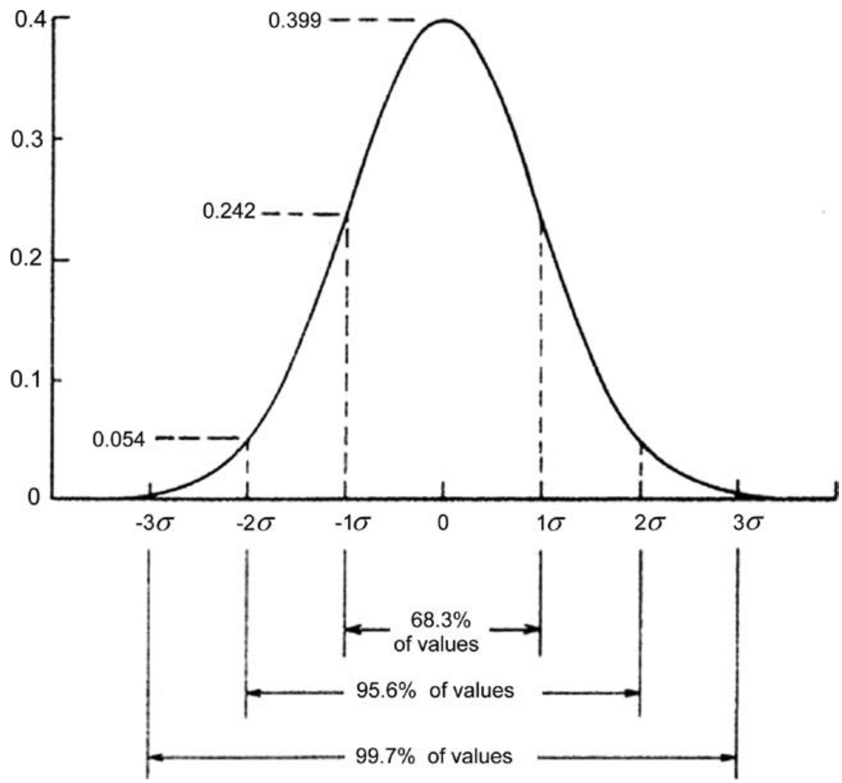

Figure 2.4: A Gaussian (normal) distribution with the X-axis marked in standard deviations on either side of the mean. The apparently strange value for the height of the peak is actually carefully chosen so the area under the curve is one

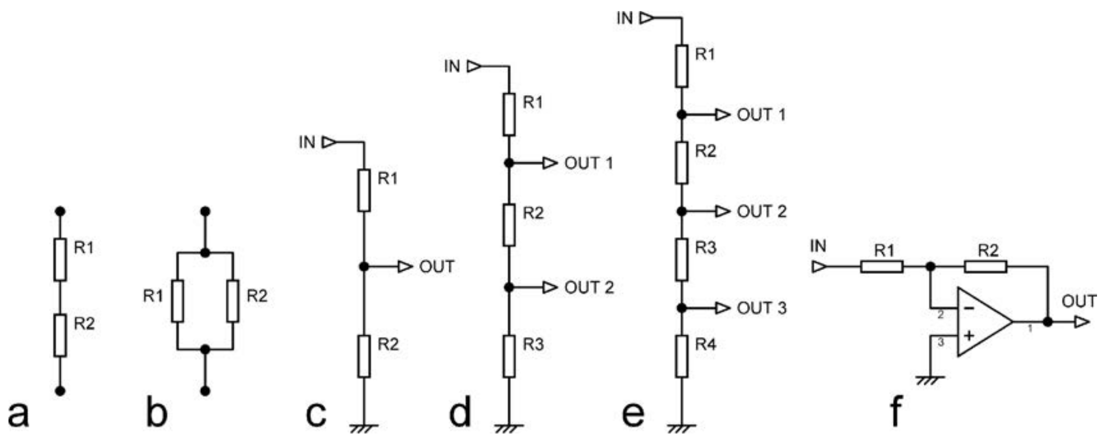

Figure 2.5: Resistor combinations: a) series, b) parallel, c) one-tap divider, d) two-tap divider, e) three tap divider, f) inverting amplifier

Thus if we have four 100 Ω 1% resistors in series, the standard deviation of the total resistance increases only by the square root of 4, that is 2 times, while the total resistance has increased by 4 times; thus we have made a 0.5% close-tolerance 400 Ω resistor, for only twice the price, whereas a 0.1% resistor would be at least ten times the price and may give more accuracy than we need. There is a happy analogue here with the use of multiple amplifiers to reduce electrical noise; we are using essentially the same technique of rms-summation to reduce 'statistical noise'.

You may object that putting four 1% resistors in series means that the worst-case errors can be four times as great. This is obviously true – if they are all 1% low, or 1% high, the total error will be 4%. But the probability of this occurring is actually very, very small indeed. The more resistors you combine, the more the values cluster together in the centre of the range.

The mathematics for series resistors is very simple (see Equation 2.2) but in other cases gets complicated very quickly. It also holds for two parallel resistors as in Figure 2.5b, though this is mathematically much less obvious. I verified it by the use of Monte-Carlo methods [3]. A suitable random number generator is used to select two resistor values, and their combined value is calculated and recorded. This is repeated many times (by computer, obviously) and then the mean and standard deviation of all the accumulated numbers is recorded. This will never give the *exact* answer, but it will get closer and closer as you make more trials. For the series and parallel cases the standard deviation is $1/\sqrt{2}$ of the standard deviation for a single resistor. If you are not wholly satisfied that this apparently magical improvement in average accuracy is genuine, seeing it happen on a spreadsheet makes a convincing demonstration.

In an Excel spreadsheet, random numbers with a uniform distribution are generated by the function RAND(), but random numbers with a Gaussian distribution and specified mean and standard deviation can be generated by the function NORMINV(). Let us assume we want to make an accurate 20 kΩ resistance. We can simulate the use of a single 1% tolerance resistor by generating a column of Gaussian random numbers with a mean of 20 and a standard deviation of 0.2; we need to use a lot of numbers to smooth out the statistical fluctuations, so we generate 400 of them. As a check we calculate the mean and standard deviation of our 400 random numbers using the AVERAGE() and STDEV() functions. The results will be very close to 20 and 0.2 but not identical, and will change every time we hit the F9 recalculate key as this generates a new set of random numbers. The results of five recalculations are shown in Table 2.6, demonstrating that 400 numbers are enough to get us quite close to our targets.

To simulate two 10 kΩ resistors of 1% tolerance in series we generate two columns of 400 Gaussian random numbers with a mean of 10 and a standard deviation of 0.1. We then set up a third column which is the sum of the two random numbers on the same row, and if we calculate the mean and standard deviation using AVERAGE() and STDEV() again, we

TABLE 2.6 Mean and standard deviation of five batches of 400 Gaussian random resistor values

Mean KΩ	Standard deviation
20.0017	0.2125
19.9950	0.2083
19.9910	0.1971
19.9955	0.2084
20.0204	0.2040

TABLE 2.7 Mean and standard deviation of five batches of 400 Gaussian 10 kΩ 1% resistors, two in series

Mean KΩ	Standard deviation
19.9999	0.1434
20.0007	0.1297
19.9963	0.1350
20.0114	0.1439
20.0052	0.1332

TABLE 2.8 Mean and standard deviation of five batches of 400 Gaussian 5 kΩ 1% resistors, four in series

Mean KΩ	Standard deviation
20.0008	0.1005
19.9956	0.0995
19.9917	0.1015
20.0032	0.1037
20.0020	0.0930

find that the mean is still very close to 20 but the standard deviation is reduced on average by the expected factor of $\sqrt{2}$. The result of five trials is shown in Table 2.7. Repeating this experiment with two 40 kΩ resistors in parallel gives the same results.

If we repeat this experiment by making our 20 kΩ resistance from a series combination of four 5 kΩ resistors of 1% tolerance we generate four columns of 400 Gaussian random numbers with a mean of 5 and a standard deviation of 0.05. We sum the four numbers on the same row to get a fifth column, and calculate the mean and standard deviation of that. The result of five trials is shown in Table 2.8. The mean is very close to 20 but the standard deviation is now reduced on average by a factor of $\sqrt{4}$, which is 2.

TABLE 2.9 The improvement in value tolerance with number of equal-value parts

Number of equal-value parts	Tolerance reduction factor
1	1.000
2	0.707
3	0.577
4	0.500
5	0.447
6	0.408
7	0.378
8	0.354
9	0.333
10	0.316

I think this demonstrates quite convincingly that the spread of values is reduced by a factor equal to the square root of the number of the components used. The principle works equally well for capacitors or indeed any quantity with a Gaussian distribution of values. The downside is the fact that the improvement depends on the square root of the number of equal-value components used, which means that big improvements require a lot of parts and the method quickly gets unwieldy. Table 2.9 demonstrates how this works; the rate of improvement slows down noticeably as the number of parts increases. The largest number of components I have ever used in this way for a production design is five. Constructing a 0.1% resistance from 1% resistors would require a hundred of them, and is hardly a practical proposition. It would cost more than just buying a 0.1% resistor.

You may be wondering what happens if the series resistors used are *not* equal. If you are in search of a particular value the method that gives the best resolution is to use one large resistor value and one small one to make up the total, as this gives a very large number of possible combinations. However, the accuracy of the final value is essentially no better than that of the large resistor. Two equal resistors, as we have just demonstrated, give a $\sqrt{2}$ improvement in accuracy, and near-equal resistors give almost as much, but the number of combinations is very limited, and you may not be able to get very near the value you want. The question is, how much improvement in accuracy can we get with resistors that are some way from equal, such as one resistor being twice the size of the other?

The mathematical answer is very simple; even when the resistor values are not equal, the overall standard deviation is still the rms-sum of the standard deviations of the two resistors, as shown in Equation 2.2 above; σ_1 and σ_2 are the standard deviations of the two resistors in series. Note

TABLE 2.10 The improvement in value tolerance with number of equal-value parts

Series resistor values Ω	Resistor ratio	Standard deviation
20K single		0.2000
19.9K + 100	199:1	0.1990
19.5K + 500	39:1	0.1951
19K + 1K	19:1	0.1903
18K + 2K	9:1	0.1811
16.7K + 3.3K	5:1	0.1700
16K + 4K	4:1	0.1649
15K + 5K	3:1	0.1581
13.33K + 6.67K	2:1	0.1491
12K + 8K	1.5:1	0.1442
11K + 9K	1.22:1	0.1421
10K + 10K	1:1	0.1414

that this equation is only correct if there is no correlation between the two values; this is true for two separate resistors but would not hold for two film resistors on the same substrate.

Since both resistors have the same percentage tolerance, the larger of the two has the greater standard deviation, and dominates the total result. The minimum total deviation is thus achieved with equal resistor values. Table 2.10 shows how this works; using two resistors in the ratio 2:1 or 3:1 still gives a worthwhile improvement in average accuracy.

The entries for 19.5k + 500 and 19.9k + 100 demonstrate that when one large resistor value and one small are used to get a particular value, its accuracy is very little better than that of the large resistor alone.

Other resistor combinations

So far we have looked at serial and parallel combinations of components to make up one value, as in Figure 2.5. Other important combinations are the resistive divider in Figure 2.5c, (frequently used as the negative-feedback network for non-inverting amplifiers) and the inverting amplifier in Figure 2.5f where the gain is set by the ratio R2/R1. All resistors are assumed to have the same tolerance about an exact mean value.

I suggest it is not obvious whether the divider ratio of Figure 2.5c, which is R2/(R1 + R2), will be more or less accurate than the resistor tolerance, even in the simple case with R1 = R2. However, the Monte-Carlo method shows that in this case partial cancellation of errors still occurs and the division ratio is more accurate by a factor of $\sqrt{2}$.

This factor depends on the divider ratio, as a simple physical argument shows:

- If the top resistor R1 is zero, then the divider ratio is obviously one with complete accuracy, the resistor values are irrelevant, and the output voltage tolerance is zero.
- If the bottom resistor R2 is zero, there is no output and accuracy is meaningless, but if instead R2 is very small compared with R1 then the R1 completely determines the current through R2, and R2 turns this into the output voltage. Therefore the tolerances of R1 and R2 act independently, and so the combined output voltage tolerance is worse by their rms-sum $\sqrt{2}$.

Some more Monte-Carlo work, with 8000 trials per data point, revealed that there is linear relationship between accuracy and the 'tap position' of the output between R1 and R2, as shown in Figure 2.6. With R1 = R2 the tap is at 50%, and accuracy improved by a factor of $\sqrt{2}$, as noted above. With a tap at about 30% (R1 = 7 kΩ, R2 = 3 kΩ) the accuracy is the same as the resistors used. This 'tap' concept is *not* applicable to potentiometers as the two sections of the pot are not uncorrelated.

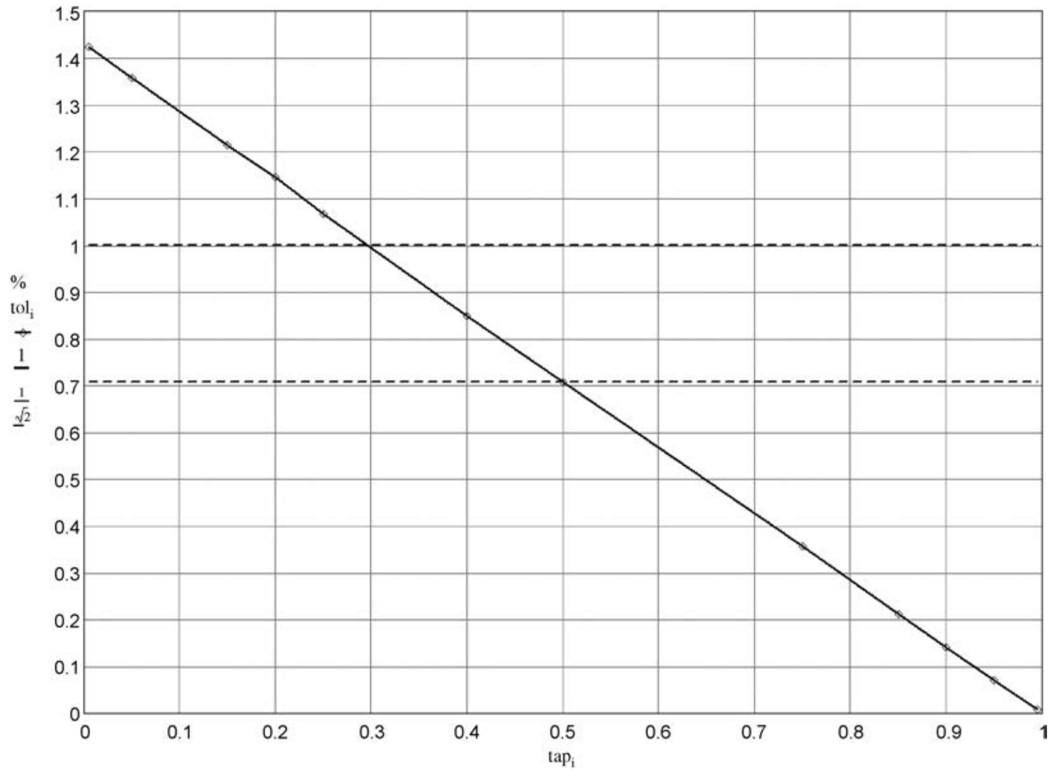

Figure 2.6: The accuracy of the output of a resistive divider made up with components of the same tolerance varies with tap position and the divider ratio.

The two-tap divider (Figure 2.5d) and three-tap divider (Figure 2.5e) were also given a Monte-Carlo-ing, though only for equal resistors. The two-tap divider has an accuracy factor of 0.404 at OUT 1 and 0.809 at OUT 2. These numbers are very close to $\sqrt{2}/2\sqrt{3}$ and $\sqrt{2}/\sqrt{3}$ respectively. The three-tap divider has an accuracy factor of 0.289 at OUT 1, 0.500 at OUT 2, and 0.864 at OUT 3. The middle figure is clearly 1/2 (twice as many resistors as a one-tap divider so $\sqrt{2}$ times more accurate) while the first and last numbers are very close to $\sqrt{3}/6$ and $\sqrt{3}/2$ respectively. It would be helpful if someone could prove analytically that the factors proposed are actually correct.

For the inverting amplifier of Figure 2.5f, the accuracy of the gain is always $\sqrt{2}$ worse than the tolerance of the two resistors, assuming the tolerances are equal. The nominal resistor values have no effect on this. We therefore have the interesting situation that a non-inverting amplifier will always be equally or more accurate in its gain than an inverting amplifier. So far as I know this is a new result.

Resistor value distributions

At this point you may be complaining that this will only work if the resistor values have a Gaussian (also known as normal) distribution with the familiar peak around the mean (average) value. Actually, it is a happy fact that this effect does *not* assume that the component values have a Gaussian distribution, as we shall see in a moment. An excellent account of how to handle statistical variations to enhance accuracy is in [4]. This deals with the addition of mechanical tolerances in optical instruments, but the principles are just the same.

You sometimes hear that this sort of thing is inherently flawed, because, for example, 1% resistors are selected from production runs of 5% resistors. If you were using the 5% resistors, then you would find there was a hole in the middle of the distribution; if you were trying to select 1% resistors from them, you would be in for a very frustrating time as they have already been selected out, and you wouldn't find a single one. If instead you were using the 1% components obtained by selection from the 5% population, you would find that the distribution would be much flatter than Gaussian and the accuracy improvement obtained by combining them would be reduced, although there would still be a definite improvement.

However, don't worry. In general this is not the way that components are manufactured nowadays, though it may have been so in the past. A rare contemporary exception is the manufacture of carbon composition resistors [5] where making accurate values is difficult, and selection from production runs, typically with a 10% tolerance, is the only practical way to get more accurate values. Carbon composition resistors have no place in audio circuitry, because of their large temperature and voltage coefficients and high excess noise, but they live on in specialised applications such as switch-mode snubbing circuits, where their ability to absorb high peak power in bulk material rather than a thin film is useful, and in RF circuitry where the inductance of spiral-format film resistors is unacceptable.

So, having laid that fear to rest, what is the actual distribution of resistor values like? It is not easy to find out, as manufacturers are not exactly forthcoming with this sort of sensitive information, and measuring thousands of resistors with an accurate DVM is not a pastime that appeals to all of us. Any nugget of information in this area is therefore very welcome.

Hugo Kroeze [6] reported the result of measuring 211 metal film resistors from the same batch with a nominal value of 10 kΩ and 1% tolerance. He concluded that:

1. The mean value was 9.995k Ω (0.05% low).
2. The standard deviation was about 10 Ω, i.e. only 0.1%. This spread in value is surprisingly small (the resistors were all from the same batch, and the spread across batches widely separated in manufacture date might have been less impressive).
3. All resistors were within the 1% tolerance range.
4. The distribution appeared to be Gaussian, with no evidence that it was a subset from a larger distribution.

I decided to add my own morsel of data to this. I measured 100 ordinary metal film 1kΩ resistors of 1% tolerance from Yageo, a Chinese manufacturer, and very tedious it was too. I used a recently-calibrated 4.5 digit meter.

1. The mean value was 997.66 Ω (0.23% low).
2. The standard deviation was 2.10 Ω, i.e. 0.21%.
3. All resistors were within the 1% tolerance range. All but one was within 0.5%, with the outlier at 0.7%.
4. The distribution appeared to be Gaussian, with no evidence that it was a subset from a larger distribution.

These are only two reports, and it would be nice to have more confirmation, but there seems to be no reason to doubt that the mean value is very well controlled, and the spread is under good control as well. The distribution of resistance values appears to be Gaussian with no evidence of the most accurate specimens being selected out. Whenever I have attempted this kind of statistical improvement in accuracy, I have always found that the expected benefit really does appear in practice.

The uniform distribution

As I mentioned earlier, improving average accuracy by combining resistors does not depend on the resistance value having a Gaussian distribution. Even a batch of resistors with a uniform distribution gives better accuracy when two of them are combined. A uniform

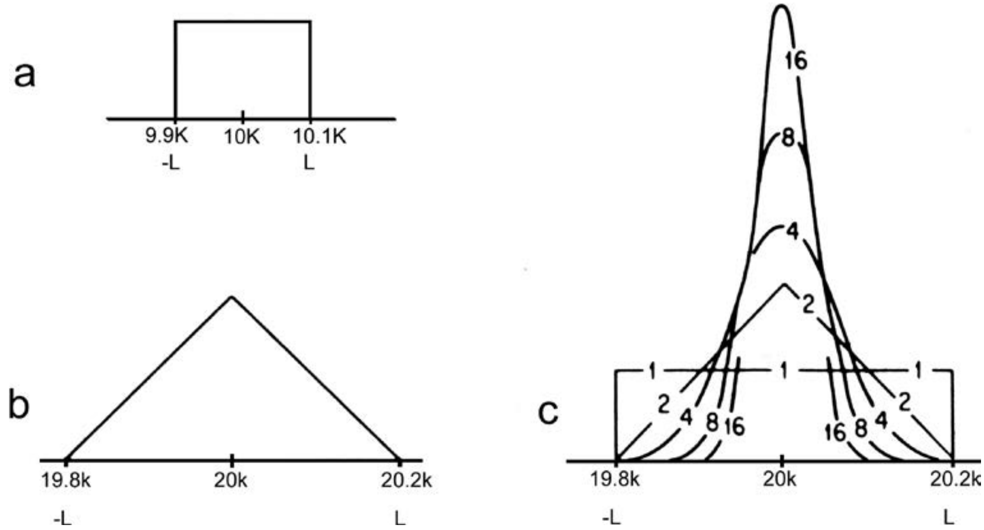

Figure 2.7: How a uniform distribution of values becomes a Gaussian (normal) distribution when more component values are summed (after Smith)

distribution of component values may not be likely but the result of combining two or more of them is highly instructive, so stick with me for a bit.

Figure 2.7a shows a uniform distribution that cuts off abruptly at the limits L and −L, and represents 10 kΩ resistors of 1% tolerance. We will assume again that we want to make a more accurate 20 kΩ resistance. If we put two of the uniform-distribution 10 kΩ resistors in series, we get not another uniform distribution, but the triangular distribution shown in Figure 2.7b. This shows that the total resistance values are already starting to cluster in the centre; it is possible to have the extreme values of 19.8 kΩ and 20.2 kΩ, but it is very unlikely.

Figure 2.7c shows what happens if we use more resistors to make the final value; when four are used the distribution is already beginning to look like a Gaussian distribution, and as we increase the number of components to 8 and then 16, the resemblance becomes very close.

Uniform distributions have a standard deviation just as Gaussian ones do. It is calculated from the limits L and −L as in Equation 2.3. Likewise, the standard deviation of a triangular distribution can be calculated from its limits L and −L as in Equation 2.4

$$\sigma = \frac{1}{\sqrt{3}}L \qquad \text{(Equation 2.3)}$$

$$\sigma = \frac{1}{\sqrt{6}}L \qquad \text{(Equation 2.4)}$$

Applying Equation 2.3 to the uniformly-distributed 10 kΩ 1% resistors in Figure 2.7a, we get a standard deviation of 0.0577. Applying Equation 2.4 to the triangular distribution of 20 kΩ resistance values in Figure 2.7b, we get 0.0816. The mean value has doubled, but the standard deviation has less than doubled, so we get an improvement in average accuracy; the ratio is $\sqrt{2}$, just as it was for two resistors with a Gaussian distribution. This is also easy to demonstrate with Monte-Carlo methods on a spreadsheet.

Resistor imperfections

It is well-known that resistors have inductance and capacitance, and vary somewhat in resistance with temperature. Unfortunately there are other less obvious imperfections, such as excess noise and non-linearity; these can get forgotten because parameters describing how bad they are are often omitted from component manufacturer's data sheets.

Being components in the real world, resistors are not perfect examples of resistance and nothing else. Their length is not infinitely small and so they have series inductance; this is particularly true for the many kinds that use a spiral resistive element. Likewise, they exhibit stray capacitance between each end, and also between the various parts of the resistive element. Both effects can be significant at high frequencies, but can usually be ignored below 100 kHz unless you are using very high or low resistance values.

It is a sad fact that resistors change their value with temperature. Table 2.4 shows some typical temperature coefficients. This is not likely to be a problem in audio applications, where the temperature range is small and extreme precision is not required unless you are designing measurement equipment. Carbon film resistors are markedly inferior to metal film in this area.

Resistor excess noise

All resistors, no matter what their resistive material or mode of construction, generate Johnson noise. This is white noise, its level being determined solely by the resistance value, the absolute temperature, and the bandwidth over which the noise is being measured. It is based on fundamental physics and is not subject to negotiation. In some cases it places the limit on how quiet a circuit can be, though the noise from active devices is often more significant. Johnson noise is covered in Chapter 1.

Excess resistor noise refers to the fact that some kinds of resistor, with a constant voltage drop imposed across them, generate excess noise in addition to its inherent Johnson noise. According to classical physics, passing a current through a resistor should have no effect on its noise behaviour; it should generate the same Johnson noise as a resistor with no steady current flow. In reality, some types of resistors do generate excess noise when they have a DC voltage across them. It is a very variable quantity, but is essentially proportional to the

TABLE 2.11 Resistor excess noise

Type	Noise (μV/V)
Carbon film TH	0.2 − 3
Metal oxide TH	0.1 − 1
Thin film SM	0.05 − 0.4
Bulk metal foil TH	0.01
Wirewound TH	0

Note: Wirewound resistors are normally considered to be completely free of excess noise.

DC voltage across the component, a typical spec being '1 μV/V' and it has a 1/*f* frequency distribution. Typically it could be a problem in biasing networks at the input of amplifier stages. It is usually only of interest if you are using carbon or thick film resistors- metal film and wirewound types should have very little excess noise. 1/*f* noise does not have a Gaussian amplitude distribution, which makes it difficult to assess reliably from a small set of data points. A rough guide to the likely specs is given in Table 2.11.

The level of excess resistor noise changes with resistor type, size, and value in Ohms; here are the relevant factors:

- Thin film resistors are markedly quieter than thick film resistors; this is due to the homogeneous nature of thin film resistive materials which are metal alloys such as nickel-chromium deposited on a substrate. The thick film resistive material is a mixture of metal (often Ruthenium) oxides and glass particles; the glass is fused into a matrix for the metal particles by high temperature firing. The higher excess noise levels associated with thick film resistors are a consequence of their heterogeneous structure, due to the particulate nature of the resistive material. The same applies to carbon-film resistors where the resistive medium is finely-divided carbon dispersed in a polymer binder.
- A physically large resistor has lower excess noise than a small resistor. In the same resistor range, the highest wattage versions have the lowest noise. See Figure 2.8.
- A low ohmic value resistor has lower excess noise than a high ohmic value. Noise in μV per V rises approximately with the square root of resistance. See Figure 2.8 again.
- A low value of excess noise is associated with uniform constriction-free current flow; this condition is not well met in composite thick film materials. However, there are great variations among different thick film resistors. The most readily apparent relationship is between noise level and the amount of conductive material present. Everything else being equal, compositions with lower resistivity have lower noise levels.

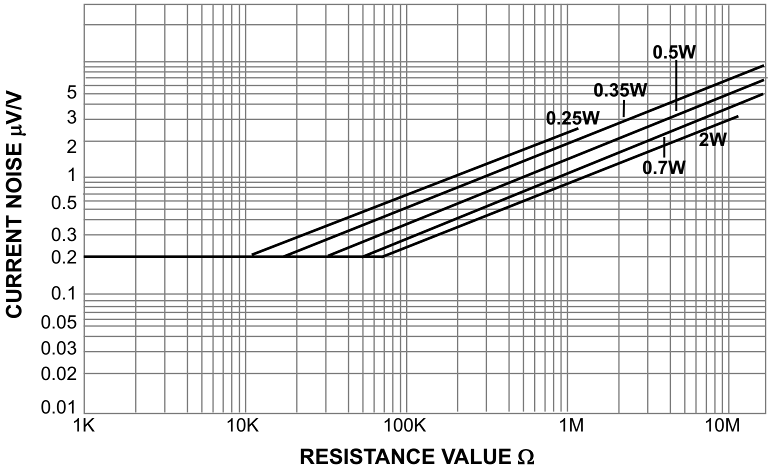

Figure 2.8: The typical variation of excess resistor noise with ohmic value and physical size; this is for a range of carbon film resistors. The flat part of the plot represents the measurement floor, not a change in noise mechanism

- Higher resistance values give higher excess noise since it is a statistical phenomenon related to the total number of charge carriers available within the resistive element; the fewer the total number of carriers present, the greater will be the statistical fluctuation.

Traditionally at this point in the discussion of excess resistor noise, the reader is warned against using carbon composition resistors because of their very bad excess noise characteristics. Carbon composition resistors are still made – their construction makes them good at handling pulse loads – but are not likely to be encountered in audio circuitry.

One of the great benefits of dual-rail opamp circuitry is that it is noticeably free of resistors with large DC voltages across them. The offset voltages and bias currents are far too low to cause trouble with resistor excess noise. However, if you are getting into low-noise hybrid discrete/opamp stages, such as the MC head amplifier in Chapter 12, you might have to consider it.

To get a feel for the magnitude of excess resistor noise, consider a 100 kΩ 1/4W carbon film resistor with 10 V across it. This, from the graph above, has an excess noise parameter of about 0.7 µV/V and so the excess noise will be of the order of 7 µV, which is −101 dBu. This definitely could be a problem in a low-noise preamplifier stage.

Resistor non-linearity

Ohm's Law strictly is a statement about metallic conductors only. It is dangerous to assume that it also invariably applies to 'resistors' simply because they have a fixed value of resistance marked on them; in fact resistors – whose main *raison d'etre* is packing a lot of controlled resistance in a small space – do not always adhere to Ohm's Law very closely. This is a distinct difficulty when trying to make low-distortion circuitry.

Resistor non-linearity is normally quoted by manufacturers as a voltage coefficient, usually the number of parts per million (ppm) that the resistor value changes when one volt is applied. The measurement standard for resistor non-linearity is IEC 6040.

Through-hole metal film resistors show perfect linearity at the levels of performance considered here, as do wirewound types. The voltage coefficient is less than 1 ppm. Carbon film resistors are quoted at less than 100 ppm; 100 ppm is however enough to completely dominate the distortion produced by active devices, if it is used in a critical part of the circuitry. Carbon composition resistors, probably of historical interest only, come in at about 350 ppm, a point that might be pondered by connoisseurs of antique equipment. The greatest area of concern over non-linearity is thick-film surface-mount resistors, which have high and rather variable voltage coefficients; more on this below.

Table 2.12 (calculated with SPICE) gives the THD in the current flowing through the resistor for various voltage coefficients when a pure sine voltage is applied. If the voltage coefficient is significant this can be a serious source of non-linearity.

A voltage-coefficient model generates all the odd-order harmonics, at a decreasing level as order increases. No even-order harmonics occur as the model is symmetrical. This is covered in much more detail in my *Active Crossover* book [7].

My own test setup is shown in Figure 2.9. The resistors are usually of equal value, to give 6 dB attenuation. A very low-distortion oscillator that can give a large output voltage is necessary; the results in Figure 2.10 were taken at a 10 Vrms (+22 dBu) input level. Here

TABLE 2.12 Resistor voltage coefficients and the resulting distortion at +15 and +20 dBu

Voltage coefficient (ppm)	THD at +15 dBu (%)	THD at +20 dBu (%)
1	0.00011	0.00019
3	0.00032	0.0005
10	0.0016	0.0019
30	0.0032	0.0056
100	0.011	0.019
320	0.034	0.06
1000	0.11	0.19
3000	0.32	0.58

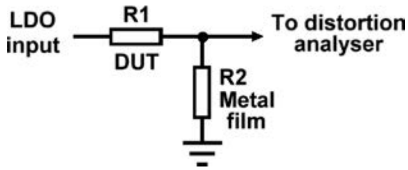

Figure 2.9: Test circuit for measuring resistor non-linearity. The not-under-test resistor R2 in the potential divider must be a metal-film type with negligible voltage coefficient

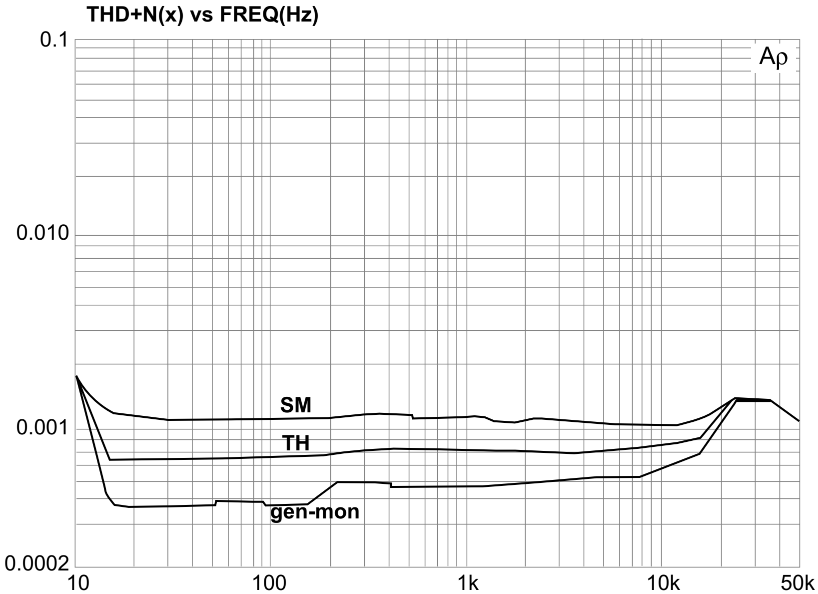

Figure 2.10: SM resistor distortion at 10 Vrms input, using 10 kΩ 0805 thick-film resistors

thick-film SM and through-hole resistors are compared. The 'Gen-mon' trace at the bottom is the record of the analyzer reading the oscillator output and is the measurement floor of the AP System 1 used. The TH plot is higher than this floor, but this is not due to distortion. It simply reflects the extra Johnson noise generated by two 10 kΩ resistors. Their parallel combination is 5 kΩ, and so this noise is at −115.2 dBu. The SM plot, however, is higher again, and the difference is the distortion generated by the thick-film component.

For both thin-film and thick-film SM resistors, non-linearity increases with resistor value, and also increases as the physical size (and hence power rating) of the resistor shrinks. The thin-film versions are much more linear; see Figures 2.11 and 2.12. Sometimes it is appropriate to reduce the non-linearity by using multiple resistors in series. If one resistor is replaced by two with the same voltage coefficient in series, the THD in the current flowing is halved. Similarly, three resistors reduces THD to a third of the original value. There are obvious

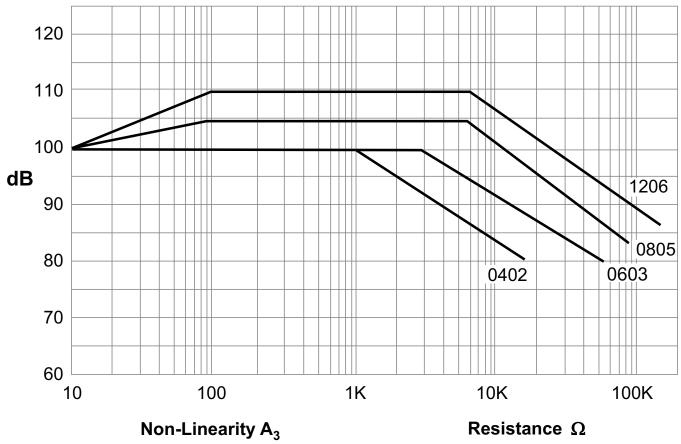

Figure 2.11: Non-linearity of thin-film surface-mount resistors of different sizes. THD is in dB rather than percent

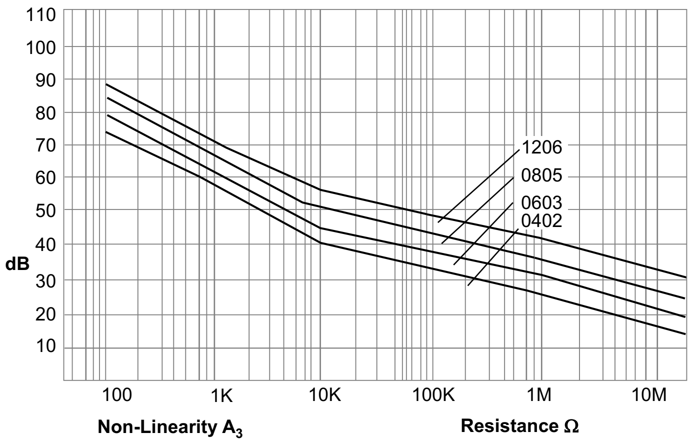

Figure 2.12: Non-linearity of thick-film surface-mount resistors of different sizes

economic limits to this sort of thing, but it can be useful in specific cases, especially where the voltage rating of the resistor is a limitation.

Capacitors

Capacitors are diverse components. In the audio business their capacitance ranges from 10 pF to 100,000 µF, a ratio of 10 to the tenth power. In this they handily outdo resistors, which usually vary from 0.1 Ω to 10M Ω, a ratio of only 10 to the eighth power. However, if you include the 10 GΩ bias resistors used in capacitor microphone head amplifiers, this range increases to ten to the eleventh. There is however a big gap between the 10 MΩ resistors, which are used in DC servos, and 10 GΩ microphone resistors; I am not aware of any audio applications for 1 GΩ resistors.

Capacitors also come in a wide variety of types of dielectric, the great divide being between electrolytic and non-electrolytic types. Electrolytics used to have *much* wider tolerances than most components, but things have recently improved and ±20% is now common. This is still wider than for typical non-electrolytics which are usually ±10% or better.

This is not the place to reiterate the basic information about capacitor properties, which can be found from in sources. I will simply note that real capacitors fall short of the ideal circuit element in several ways, notably leakage, Equivalent Series Resistance (ESR), dielectric absorption and non-linearity.

Capacitor leakage is equivalent to a high value resistance across the capacitor terminals, and allows a trickle of current to flow when a DC voltage is applied. It is usually negligible for non-electrolytics, but is much greater for electrolytics.

ESR is a measure of how much the component deviates from a mathematically pure capacitance. The series resistance is partly due to the physical resistance of leads and foils, and partly due to losses in the dielectric. It can also be expressed as tan-δ (tan-delta). Tan-delta is the tangent of the phase angle between the voltage across and the current flowing through the capacitor.

Dielectric absorption is a well-known phenomenon; take a large electrolytic, charge it up, and then make sure it is fully discharged. Use a 10 Ω WW resistor across the terminals rather than a screwdriver unless you're not too worried about either the screwdriver or the capacitor. Wait a few minutes, and the charge will partially reappear, as if from nowhere. This 'memory effect' also occurs in non-electrolytics to a lesser degree; it is a property of the dielectric, and is minimized by using polystyrene, polypropylene, NPO ceramic, or PTFE dielectrics. Dielectric absorption is invariably simulated by a linear model composed of extra resistors and capacitances; nevertheless dielectric absorption and distortion correlate across the different dielectrics.

Capacitor non-linearity is undoubtedly the least known of these shortcomings. A typical RC low-pass filter can be made with a series resistor and a shunt capacitor, and if you examine the output with a distortion analyzer, you may find to your consternation that the circuit is not linear. If the capacitor is a non-electrolytic type with a dielectric such as polyester, then the distortion is relatively pure third harmonic, showing that the effect is symmetrical. For a 10 Vrms input, the THD level may be 0.001% or more. This may not sound like much but it is substantially greater than the mid-band distortion of a good opamp. Capacitor non-linearity is dealt with at greater length below.

Capacitors are used in audio circuitry for three main functions, where their possible non-linearity has varying consequences:

1. Coupling or DC blocking capacitors. These are usually electrolytics, and if properly sized have a negligible signal voltage across them at the lowest frequencies of interest. The properties of the capacitor are pretty much unimportant unless current levels are high; power amplifier output capacitors can generate considerable mid-band distortion [8]. Much nonsense has been talked about mysterious coupling capacitor properties, but it is all nonsense. For small-signal use, as long as the signal voltage across the capacitor is kept low, non-linearity is not normally detectable. The capacitance value is non-critical, as it has to be, given the wide tolerances of electrolytics.

2. Supply filtering or decoupling capacitors. These are electrolytics if you are filtering out supply rail ripple, etc. and non-electrolytics, usually around 100 nF, when the task is to keep the supply impedance low at high frequencies and so keep opamps stable. The capacitance value is again non-critical.

3. Setting time-constants, for example the capacitors in the feedback network of an RIAA amplifier. This is a much more demanding application than the other two. Firstly, the actual value is now crucially important as it defines the accuracy of the frequency response. Secondly, there is by definition significant signal voltage across the capacitor and so non-linearity can be a serious problem. Non-electrolytics are normally used; sometimes an electrolytic is used to define the lower end of the bandwidth, but this is a bad practice likely to introduce distortion at the bottom of the frequency range. Small value ceramic capacitors are used for compensation purposes.

In subjectivist circles it is frequently asserted that electrolytic coupling capacitors (if they are permitted at all) should be bypassed by small non-electrolytics. There is no sense in this; if the main coupling capacitor has no signal voltage across it, the extra capacitor can have no effect.

Capacitor non-linearity examined

When attempting the design of linear circuitry, everyone knows that inductors and transformers with ferromagnetic core material can be a source of non-linearity. It is however less obvious that capacitors and even resistors can show non-linearity and generate some unexpected and very unwelcome distortion. Resistor non-linearity has been dealt with earlier in this chapter; let us examine the shortcomings of capacitors.

The definitive work on capacitor distortion is a magnificent series of articles by Cyril Bateman in *Electronics World* [9]. The authority of this work is underpinned by Cyril's background in capacitor manufacture (the series is long because it includes the development of a low-distortion THD test set in the first two parts).

Capacitors generate distortion when they are actually implementing a time-constant- in other words, when there is a signal voltage across them. The normal coupling or DC-blocking capacitors have no significant signal voltage across them, as they are intended to pass all the information through, not to filter it or define the system bandwidth. Capacitors with no signal across them do not generally produce distortion at small-signal current levels. This was confirmed for all the capacitors tested below. However electrolytic types may do so at power amplifier levels where the current through them is considerable, such as in the output coupling capacitor of a power amplifier [8].

Non-electrolytic capacitor non-linearity

It has often been assumed that non-electrolytic capacitors, which generally approach an ideal component more closely than electrolytics, and have dielectrics constructed in a totally different way, are free from distortion. It is not so. Some non-electrolytics show distortion at levels that is easily measured, and can exceed the distortion from the opamps in the circuit. Non-electrolytic capacitor distortion is primarily third harmonic, because the non-polarised dielectric technology is basically symmetrical. The problem is serious, because non-electrolytic capacitors are commonly used to define time-constants and frequency responses (in RIAA equalisation networks, for example) rather than simply for DC-blocking.

Very small capacitances present no great problem. Simply make sure you are using the C0G (NP0) type, and so long as you choose a reputable supplier, there will be no distortion. I say 'reputable supplier' because I did once encounter some allegedly C0G capacitors from China that showed significant non-linearity [10].

Middle-range capacitors, from 1nF to 1μF, present more of a problem. Capacitors with a variety of dielectrics are available, including polyester, polystyrene, polypropylene, polycarbonate and polyphenylene sulphide, of which the first three are the most common (note that what is commonly called 'polyester' is actually polyethylene terephthalate, PET).

Figure 2.13 shows a simple low-pass filter circuit which, in conjunction with a good THD analyzer, can be used to get some insight into the distortion problem; it is intended to be representative of a real bit of audio circuitry. The values shown give a pole frequency, or −3 dB roll-off point, at 710 Hz. Since it might be expected that different dielectrics give

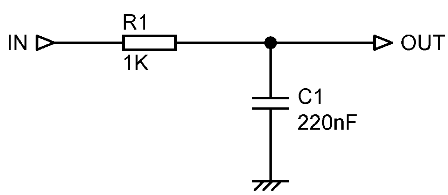

Figure 2.13: Simple low-pass test circuit for non-electrolytic capacitor distortion

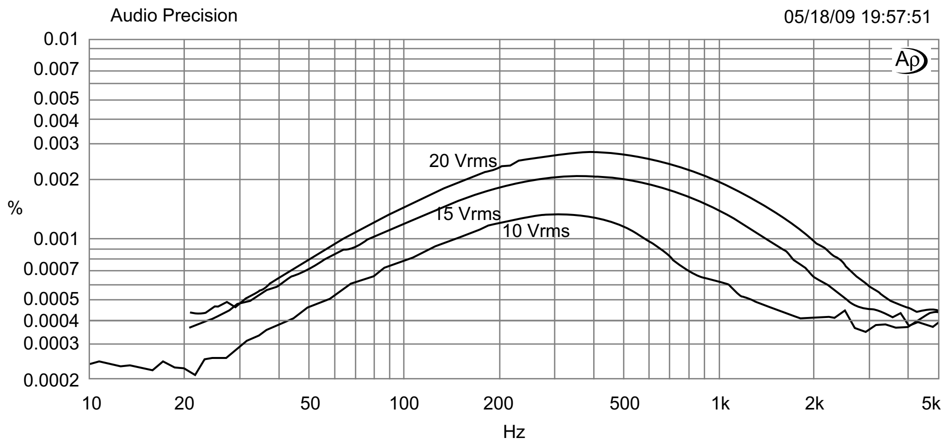

Figure 2.14: Third-harmonic distortion from a 220 nF 100 V polyester capacitor, at 10, 15 and 20 Vrms input level, showing peaking around 400 Hz

different results (and they definitely do) we will start off with polyester, the smallest, most economical, and therefore the most common type for capacitors of this size.

The THD results for a microbox 220 nF 100 V capacitor with a polyester dielectric are shown in Figure 2.14, for input voltages of 10, 15 and 20 Vrms. They are unsettling.

The distortion is all third-harmonic and peaks at around 300 to 400 Hz, well below the pole frequency, and even with input limited to 10 Vrms, will exceed the non-linearity introduced by opamps such as the 5532 and the LM4562. Interestingly, the peak frequency changes with applied level. Below the peak, the voltage across the capacitor is constant but distortion falls as frequency is reduced, because the increasing impedance of the capacitor means it has less effect on a circuit node at a 1 kΩ impedance. Above the peak, distortion falls with increasing frequency because the low-pass circuit action causes the voltage across the capacitor to fall.

The level of distortion varies with different samples of the same type of capacitor; six of the above type were measured and the THD at 10 Vrms and 400 Hz varied from 0.00128% to 0.00206%. This puts paid to any plans for reducing the distortion by some sort of cancellation method. The distortion can be seen in Figure 2.15 to be a strong function of level, roughly tripling as the input level doubles. Third harmonic distortion normally quadruples for doubled level, so there may well be an unanswered question here. It is however clear that reducing the voltage across the capacitor reduces the distortion. This suggests that if cost is not the primary consideration, it might be useful to put two capacitors in series to halve the voltage, and the capacitance, and then double up this series combination to restore the original capacitance, giving the series-parallel arrangement in Figure 2.11. The results are shown in Table 2.13, and once more it can be seen that halving the level has reduced distortion by a

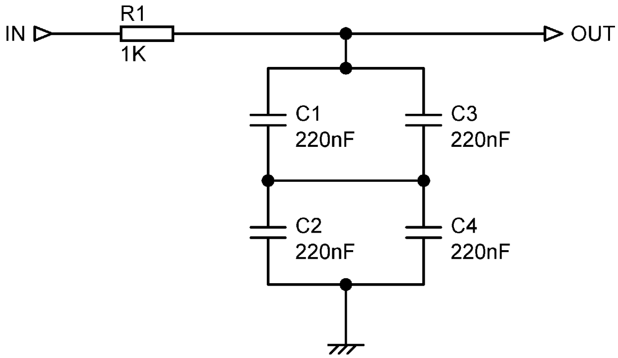

Figure 2.15: Reducing capacitor distortion by series-parallel connection

TABLE 2.13 The reduction of polyester capacitor distortion by series-parallel connection

Input level (Vrms)	Single capacitor (%)	Series-parallel capacitors (%)
10	0.0016	0.00048
15	0.0023	0.00098
20	0.0034	0.0013

factor of three rather than four. The series-parallel arrangement obviously has limitations in terms of cost and PCB area occupied, but might be useful in some cases.

Clearly polyester gives significant distortion, despite its extensive use in audio circuitry of all kinds.

An unexpected complication was that every time a sample was re-measured, the distortion was lower than before. I found a steady reduction in distortion over time if a test signal was left applied; 9 Vrms at 1 kHz halved the THD over 11 hours. This is a semi-permanent change as some of the distortion returns over time when the signal is removed. This effect may be of little practical use, but it does demonstrate that polyester capacitors are more complicated than they look. For much more on this see [11].

The next dielectric we will try is polystyrene. Capacitors with a polystyrene dielectric are extremely useful for some filtering and RIAA-equalisation applications because they can be obtained at a 1% tolerance at up to 10 nF at a reasonable price. They can be obtained in larger sizes at an unreasonable, or at any rate much higher, price.

The distortion test results are shown in Figure 2.16 for a 4n7 2.5% capacitor; the series resistor R1 has been increased to 4.7 kΩ to keep the −3 dB point inside the audio band, and it is now at 7200 Hz. Note that the THD scale has been extended down to a

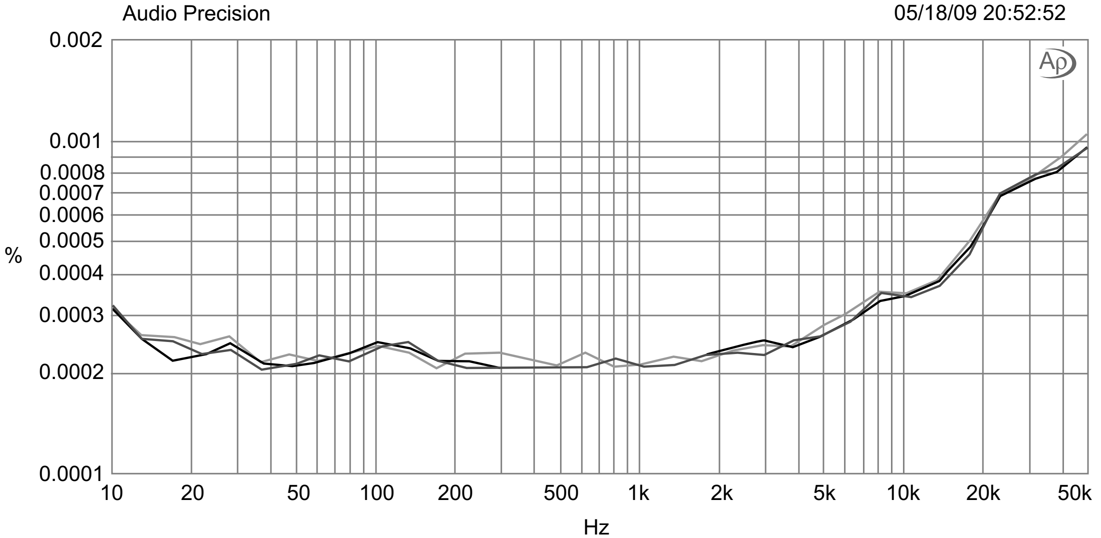

Figure 2.16: The THD plot with three samples of 4n7 2.5% polystyrene capacitors, at 10 Vrms input level. The reading is entirely noise

subterranean 0.0001%, and if it was plotted on the same scale as Figure 2.14 it would be bumping along the bottom of the graph. Figure 2.16 in fact shows no distortion at all, just the measurement noise floor, and the apparent rise at the HF end is simply due to the fact that the output level is decreasing, because of the low-pass action, and so the noise floor is relatively increasing. This is at an input level of 10 Vrms, which is about as high as might be expected to occur in normal opamp circuitry. The test was repeated at 20 Vrms, which might be encountered in discrete circuitry, and the results were the same. No measurable distortion.

The tests were done with four samples of 10 nF 1% polystyrene from LCR at 10 Vrms and 20 Vrms, with the same results for each sample. This shows that polystyrene capacitors can be used with confidence; this is in complete agreement with Cyril Bateman's findings [12].

Having settled the issue of capacitor distortion below 10 nF, we need now to tackle capacitor values greater than 10 nF. Polyester having proven unsatisfactory, the next most common dielectric is polypropylene, and I might as well say at once that it was with considerable relief that I found these were effectively distortion-free in values up to 220 nF. Figure 2.17 shows the results for four samples of a 220 nF 250 V 5% polypropylene capacitor from RIFA. Once more the plot shows no distortion at all, just the noise floor, the apparent rise at the HF end being increasing relative noise due to the low-pass roll-off. This is also in agreement with Cyril Bateman's findings. Rerunning the tests at 20 Vrms gave the same result – no distortion. This is very pleasing, but there is a downside. Polypropylene capacitors

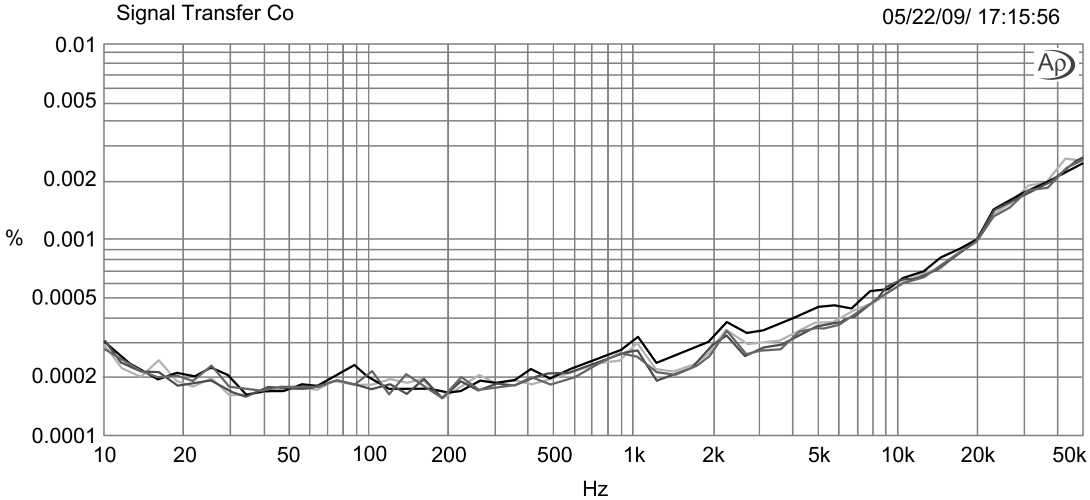

Figure 2.17: The THD plot with four samples of 220 nF 250 V 5% polypropylene capacitors, at 10 Vrms input level. The reading is again entirely noise

of this value and voltage rating are physically much larger than the commonly used 63 or 100 V polyester capacitor, and more expensive.

It was therefore important to find out if the good distortion performance was a result of the 250 V rating, and so I tested a series of polypropylene capacitors with lower voltage ratings from different manufacturers. Axial 47 nF 160 V 5% polypropylene capacitors from Vishay proved to be THD-free at both 10 Vrms and 20 Vrms. Likewise, microbox polypropylene capacitors from 10 nF to 47 nF, with ratings of 63 V and 160 V from Vishay and Wima proved to generate no measurable distortion, so the voltage rating appears not to be an issue. This finding is particularly important because the Vishay range has a 1% tolerance, making them very suitable for precision filters and equalisation networks. The 1% tolerance is naturally reflected in the price.

The only remaining issue with polypropylene capacitors is that the higher values (above 100nF) appear to be currently only available with 250 V or 400 V ratings, and that means a physically big component. For example, the EPCOS 330 nF 400V 5% part has a footprint of 26 mm by 6.5 mm, with a height of 15 mm. One way of dealing with this is to use a smaller capacitor in a capacitance multiplication configuration, so a 100 nF 1% component could be made to emulate 470 nF. It has to be said that the circuitry for this is only straightforward if one end of the capacitor is connected to ground.

When I first started looking at capacitor distortion, I thought that the distortion would probably be lowest for the capacitors with the highest voltage rating. I therefore tested some RF-suppression X2 capacitors, rated at 275 Vrms, which translates into a peak or

DC rating of 389 V. These are designed to be connected directly across the mains and therefore have a thick and tough dielectric layer. For some reason manufacturers seem to be very coy about saying exactly what the dielectric material is, normally describing them simply as 'film capacitors'. A problem that surfaced immediately is that the tolerance is 10 or 20%, not exactly ideal for precision filtering or equalisation. A more serious problem, however, is that they are far from distortion-free. Four samples of a 470 nF X2 capacitor showed THD between 0.002% and 0.003% at 10 Vrms. Clearly a high voltage rating alone does not mean low distortion.

Electrolytic capacitor non-linearity

Cyril Bateman's series in *Electronics World* [13] included two articles on electrolytic capacitor distortion. It proved to be a complex subject, and many long-held assumptions (such as 'DC biasing always reduces distortion') were shown to be quite wrong. Distortion was in general a good deal higher than for non-electrolytic capacitors

My view is that electrolytics should never, ever, under any circumstances, be used to set time-constants in audio. There should be a time-constant early in the signal path, based on a non-electrolytic capacitor, that determines the lower limit of the bandwidth, and all the electrolytic-based time-constants should be much longer so that the electrolytic capacitors can never have significant signal voltages across them and so never generate measurable distortion. There is of course also the point that electrolytics have large tolerances, and cannot be used to set accurate time-constants anyway.

However, even if you obey this rule, you can still get into deep trouble. Figure 2.18 shows a simple high-pass test circuit designed to represent an electrolytic capacitor in use for coupling or DC-blocking. The load of 1 kΩ is the sort of value that can easily be encountered if you are using low-impedance design principles. The calculated −3 dB roll-off point is 3.38 Hz, so the attenuation at 10 Hz, at the very bottom of the audio band, will be only 0.47 dB; at 20 Hz it will be only 0.12 dB, which is surely a negligible loss. As far as frequency response goes, we are doing fine. But . . . examine Figure 2.19, which shows the measured distortion of this arrangement. Even if we limit ourselves to a 10 Vrms level, the distortion at 50 Hz is 0.001%, already above that of a good opamp. At 20 Hz it has risen to 0.01%, and by

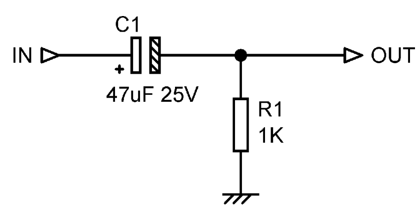

Figure 2.18: High-pass test circuit for examining electrolytic capacitor distortion

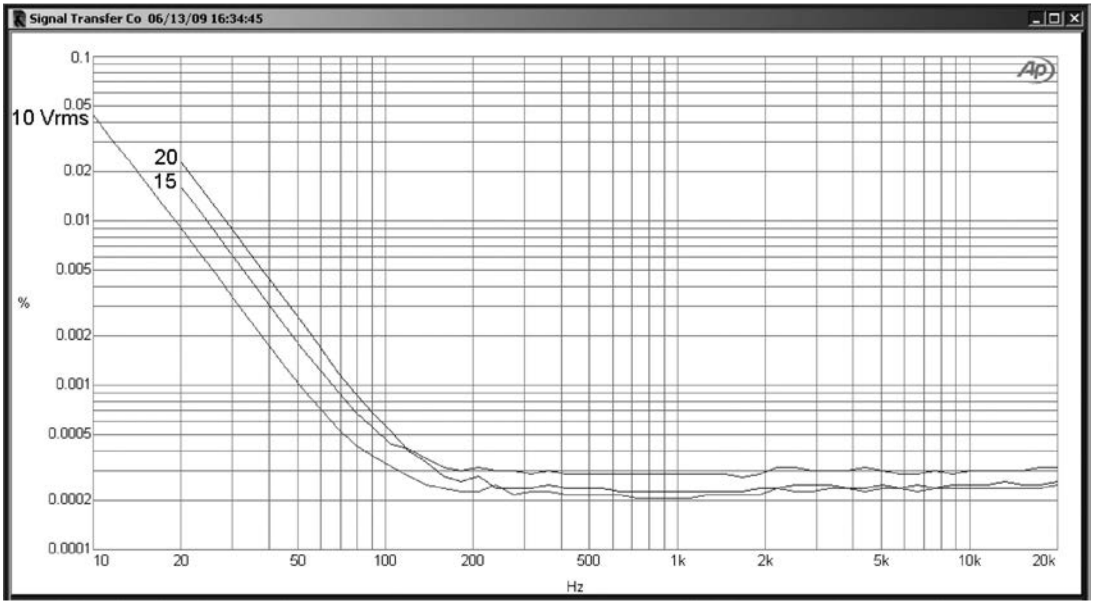

Figure 2.19: Electrolytic capacitor distortion from the circuit in Figure 2.13. Input level 10, 15 and 20 Vrms

10 Hz a most unwelcome 0.05%. The THD is increasing by a ratio of 4.8 times for each octave fall in frequency, in other words increasing faster than a square law. The distortion residual is visually a mixture of second and third harmonic, and the levels proved surprisingly consistent for a large number of 47 μF 25 V capacitors of different ages and from different manufacturers.

Figure 2.19 also shows that the distortion rises rapidly with level; at 50 Hz going from an input of 10 Vrms to 15 Vrms almost doubles the THD reading. To underline the point, consider Figure 2.20, which shows the measured frequency response of the circuit with 47 μF and 1 KΩ; note the effect of the capacitor tolerance on the real versus calculated figures. The roll-off that does the damage, by allowing an AC voltage to exist across the capacitor, is very modest indeed – less than 0.2 dB at 20 Hz.

Having demonstrated how insidious this problem is, how do we fix it? Changing capacitor manufacturer is no help. Using 47 μF capacitors of higher voltage does not work – tests showed there is very little difference in the amount of distortion generated. An exception was the sub-miniature style of electrolytic, which was markedly worse.

The answer is simple – just make the capacitor bigger in value. This reduces the voltage across it in the audio band, and since we have shown that the distortion is a strong function of the voltage across the capacitor, the amount produced drops more than proportionally. The result is seen in Figure 2.21, for increasing capacitor values with a 10 Vrms input.

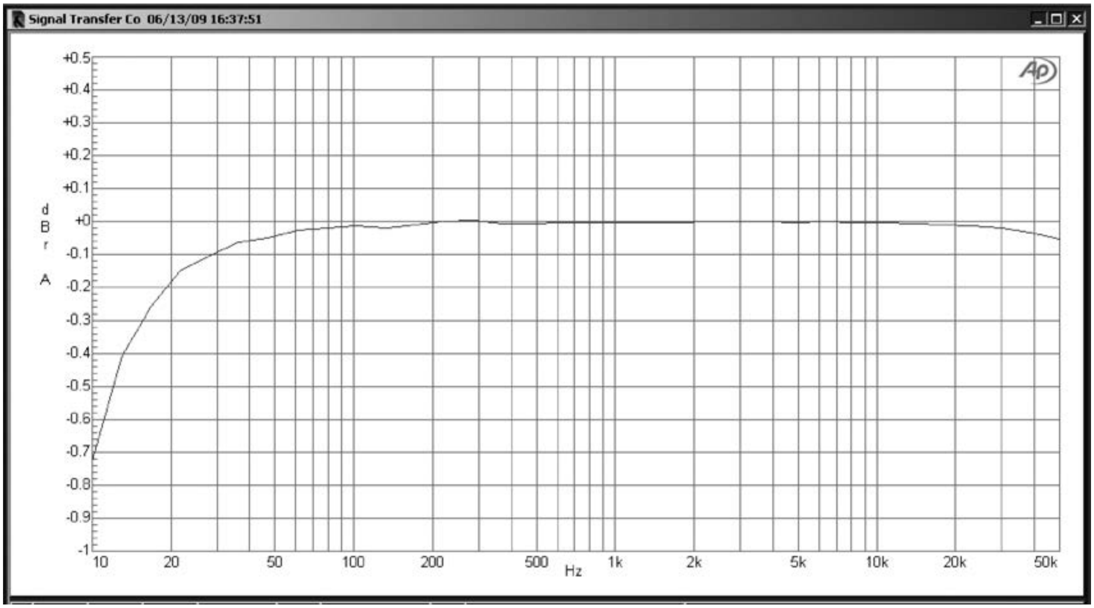

Figure 2.20: The measured roll-off of the high-pass test circuit for examining electrolytic capacitor distortion

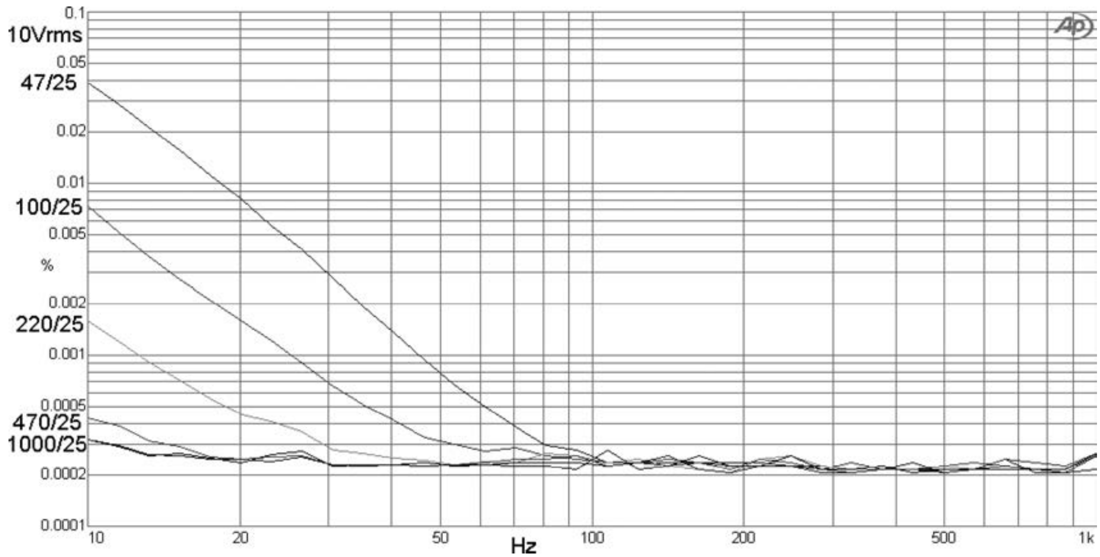

Figure 2.21: Reducing electrolytic capacitor distortion by increasing the capacitor value. Input 10 Vrms

Replacing C1 with a 100 µF 25 V capacitor drops the distortion at 20 Hz from 0.0080% to 0.0017%, an improvement of 4.7 times; the voltage across the capacitor at 20 Hz has been reduced from 1.66 Vrms to 790 mVrms. A 220 µF 25 V capacitor reduces the voltage across itself to 360 mV, and gives another very welcome reduction to 0.0005% at 20 Hz, but it is necessary to go to 1000 µF 25 V to obtain the bottom trace, which is indistinguishable from the noise floor of the AP-2702 test system. The voltage across the capacitor at 20 Hz is now only 80 mV. From this data, it appears that the AC voltage across an electrolytic capacitor should be limited to below 80 mVrms if you want to avoid distortion. I would emphasise that these are ordinary 85 °C rated electrolytic capacitors, and in no sense special or premium types.

This technique can be seen to be highly effective, but it naturally calls for larger and somewhat more expensive capacitors, and larger footprints on a PCB. This can be to some extent countered by using capacitors of lower voltage, which helps to bring back down the CV product and hence the can size. I tested 1000 µF 16 V and 1000 µF 6V3 capacitors, and both types gave exactly the same results as the 1000 µF 25 V part in Figure 2.21, with useful reductions in CV product and can size. This does of course assume that the capacitor is, as is usual, being used to block small voltages from opamp offsets to prevent switch clicks and pot noises rather than for stopping a substantial DC voltage.

The use of large coupling capacitors in this way does require a little care, because we are introducing a long time-constant into the circuit. Most opamp circuitry is pretty much free of big DC voltages, but if there are any, the settling time after switch-on may become undesirably long.

More information on capacitor distortion in specific applications can be found in Chapters 3 and 8.

Inductors

For several reasons, inductors are unpopular with circuit designers. They are relatively expensive, often because they need to be custom-made. Unless they are air-cored (which limits their inductance to low values) the core material is a likely source of non-linearity. Some types produce substantial external magnetic fields, which can cause crosstalk if they are placed close together, and similarly they can be subject to the induction of interference from other external fields. In general they deviate from being an ideal circuit element much more than resistors or capacitors.

It is rarely, if ever, essential to use inductors in signal-processing circuitry. Historically they were used in tone controls, before the Baxandall configuration swept all before it, and their last applications were probably in mid EQ controls for mixing consoles and in LCR filters for graphic equalisers. These too were gone by the end of the seventies, being replaced by active filters and gyrators, to the considerable relief of all concerned (except inductor manufacturers).

The only place where inductors are essential is when the need for galvanic isolation, or enhanced EMC immunity, makes input and output transformers desirable, and even then they need careful handling; see Chapters 18 and 10 on line-in and line-out circuitry.

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CHAPTER 3

Discrete transistor circuitry

This chapter deals with small-signal design using discrete transistors, specifically BJT's. Many things found in standard textbooks are skated over quickly. It concentrates on audio issues, and gives information that I do not think appears anywhere else, including the distortion behaviour of various configurations.

Why use discrete transistor circuitry?

Circuitry made with discrete transistors is not obsolete. It is appropriate when:

- A load must be driven to higher voltages than an opamp can sustain between the supply rails. Opamps are mostly restricted to supply voltages of ±18 or ±20 Volts. Hybrid-construction amplifiers, typically packaged in TO3 cans, will operate from rails as high as ±100 V but they are very expensive, and not optimised for audio use in parameters like crossover distortion. Discrete circuitry provides a viable alternative, but it must never be forgotten that excessive output signal levels may damage opamp equipment. Hifi is very rarely equipped with input voltage protection.
- A load requires more drive current, because of its low impedance, than an opamp can provide without overheating or current-limiting; e.g. any audio power amplifier.
- The best possible noise performance is required. Discrete bipolar transistors can outperform opamps, particularly with low source resistances, say 500 Ω or less. The commonest examples are moving-coil head amps and microphone preamplifiers. These almost invariably use a discrete input device or devices, with the open-loop gain (for linearity) and load-driving capability provided by an opamp which may itself have fairly humble noise specs.
- The best possible distortion performance is demanded. Most opamps have Class-B or AB output stages, and many of them (though certainly not all) show clear crossover artefacts on the distortion residual. A discrete opamp can dissipate more power than an IC, and so can have a Class-A output stage, sidestepping the crossover problem completely.
- When it would be necessary to provide a low-voltage supply to run just one or two opamps. The cost of extra transformer windings, rectifiers, reservoirs, and regulators will

buy a lot of discrete transistors. For example, if you need a buffer stage to drive a power amplifier from a low impedance, it may be more economical, and save space and weight, to use a discrete emitter-follower running from the same rails as the power amplifier. In these days of auto-insertion fitting the extra parts on the PCB will cost very little.

- Purely for marketing purposes, as you think you can mine a vein of customers that don't trust opamps.

When studying the higher reaches of discrete design, the most fruitful source of information is paradoxically papers on analogue IC design. This applies with particular force to design with BJTs. The circuitry used in ICs can rarely be directly adapted for use with discrete semiconductors, because some features such as multiple collector transistors and differing emitter areas simply do not exist in the discrete transistor world; it is the basic principles of circuit operation that can be useful. A good example is a paper by Erdi, dealing with a unity-gain buffer with a slew rate of 300 V/μs [1]. Another highly informative discourse is by Barry Hilton, [2] which also deals with a unity-gain buffer.

A little caution is required when a discrete stage may be driving not another of its own kind on the same supply rail, but opamp-based circuitry that is likely to be running off no more than ±18 V, as peak signal levels may drive the opamp inputs outside their rail voltages and cause damage. This normally only applies to discrete output stages but should be kept in mind whenever the supply rail is higher than +36 V for single rail and ±18 V for dual rail.

Bipolars and FETs

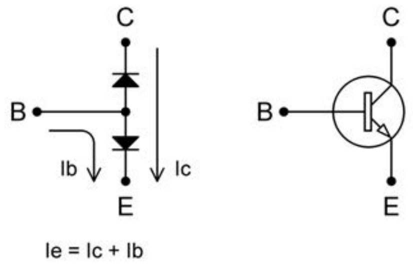
This chapter only deals with bipolar transistors. Their high transconductance and predictable operation make them far more versatile than FETs. The highly-variable V_g of a FET can be dealt with by expedients such as current-source biasing, but the low gain, which means low feedback and poor linearity, remains a problem. FETs have their uses when super-high input impedances are required, and an example of a JFET working with an opamp that provides loop gain can be found in Chapter 17 on microphone amplifiers.

Bipolar junction transistors

There is one thing to get straight first:

The bipolar junction transistor is a voltage-operated device.

What counts is the base-emitter voltage, or V_{be}. Certainly a BJT needs base current to flow for it to operate, but this is really an annoying imperfection rather than the basis of operation. I appreciate this may take some digesting; far too many discussions of transistor action say



Ebers-Moll equation

$$Ie = Is\left(e^{\frac{Vbe}{Vt}} - 1\right)$$

Is is the reverse saturation current of the base-emitter diode (of the order of 10^{-15} to 10^{-12} Amp)

Vbe is the base-emitter voltage

Vt is the thermal voltage kT/q (approx 26 mV at 25C room temperature).

k is Boltsmann's constant 1.381x 10^{-23} Joule/K
T is absolute temperature in K
q is the electron charge 1.6022 x 10^{-19} Coulomb

Figure 3.1: Current flow through a bipolar transistor, and the fundamental transistor equation

something like 'a small current flowing into the base controls a much larger current flowing into the collector'. In fact the only truly current-operated amplifying device that comes to mind is the Hall-effect multiplier, and you don't come across those every day. I've certainly never seen one used in audio- could be a market niche there.

Transistor operation is thus: if the base is open-circuit, then no collector current flows, as the collector-base junction is effectively a reverse-biased diode, as seen in Figure 3.1. There is a little leakage through from the collector to the emitter, but with modern silicon BJTs you can usually ignore it.

When the base is forward biased by taking it about 600 mV above the emitter, charge carriers are launched into the base region. Since the base region is narrow, the vast majority shoot through into the collector, to form the collector current I_c. Only a small proportion of these carriers are snared in the base and become the base current I_b, which is clearly a result and not the cause of the base-emitter voltage. I_b is normally just a nuisance.

The transistor equation

Every bipolar transistor obeys the Ebers-Moll transistor equation shown in Figure 3.1 with startling accuracy over nine or ten decades of I_c, which is a pretty broad hint that we are looking at the fundamental mechanism. In contrast, beta varies with I_c, temperature, and just about everything else you can think of. The collector current is to a first approximation independent of collector voltage – in other words it is a current-source output. The qualifications to this are:

1. This only holds for V_{ce} above, say, 2 volts.
2. It is not a perfect current-source; even with a high V_{ce}, I_c increases slowly with V_{ce}. This is called the Early Effect, after Jim Early, [3] and has nothing to do with timing or punctuality. It is a major consideration in the design of stages with high voltage gain. The same effect when the transistor is operated in reverse mode – a perversion that will not concern us here – has sometimes been called the Late Effect. Ho-ho.

Beta

Beta (or h_{fe}) is the ratio of the base current I_b to the collector current I_c. It is not a fundamental property of a BJT. Never design circuits that depend on beta, unless of course you're making a transistor tester.

Here are some of the factors that affect beta. This should convince you that it is a shifty and thoroughly untrustworthy parameter:

- Beta varies with I_c. First it rises as I_c increases, reaching a broad peak, then it falls off as I_c continues to increase.
- Beta increases with temperature. This seems to be relatively little known. Most things, like leakage currents, get worse as temperature increases, so this makes a nice change.
- Beta is lower for high-current transistor types.
- Beta is lower for high V_{ceo} transistor types. This is a major consideration when you are designing the small-signal stages of power amplifiers with high supply rails.
- Beta varies widely between nominally identical examples of the same transistor type.

A very good refutation of the beta-centric view of BJTs is given by Barrie Gilbert in [4].

Unity-gain buffer stages

A buffer stage is used to isolate two portions of circuitry from each other. It has a high input impedance and low output impedance; typically it prevents things downstream from loading things upstream. The use of the word 'buffer' normally implies 'unity-gain buffer' because otherwise we would be talking about an amplifier or gain stage. The gain with the simpler discrete implementations is in fact slightly less than one. The simplest discrete buffer circuit-block is the one-transistor emitter-follower; it is less than ideal both in its mediocre linearity, and its asymmetrical load-driving capabilities. If we permit ourselves another transistor, the complementary feedback-pair (CFP) configuration gives better linearity. Both versions can have their load-driving performance much improved by replacing the emitter resistor with a constant-current source or a push-pull Class-A output arrangement.

If the CFP stage is not sufficiently linear, the next stage in sophistication is to combine an input differential pair with an output emitter-follower, using three transistors. This arrangement, often called the Schlotzaur configuration [5], can be elaborated until it gives a truly excellent distortion performance. Its load-driving capability can be enhanced in the same as the simpler configurations.

The simple emitter-follower

The simplest discrete circuit-block is the one transistor emitter-follower. This count of one does not include extra transistors used as current-sources, etc, to improve load-driving ability. It does not have a gain of exactly one, but it is usually pretty close.

Figure 3.2 shows a simple emitter-follower with a 2k7 emitter resistor Re, giving a quiescent current of 8.8 mA with ±24 V supply rails. Biasing is by a high-value resistor R1 connected to 0V. Note the polarity of the output capacitor; the output will sit at about −0.6 V due to the V_{bc} drop, plus a little lower due to the voltage drop caused by the base current I_b flowing through R1.

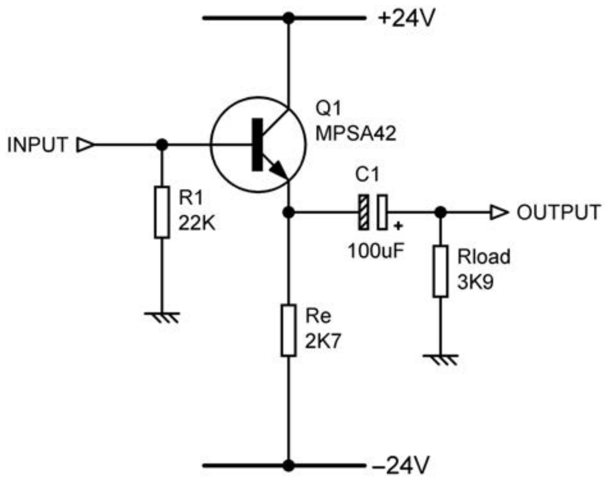


Figure 3.2: The simple emitter-follower circuit running from ±24 V supply rails

The input impedance is approximately that of the emitter resistor in parallel with an external loading on the stage multiplied by the transistor beta:

$$R_{in} = \beta(R_e \| R_{load}) \qquad \text{(Equation 3.1)}$$

Don't expect the output impedance to be as low as a opamp with plenty of NFB. The output impedance is approximately that of the source resistance divided by beta:

$$R_{out} = R_s/(\beta) \qquad \text{(Equation 3.2)}$$

The gain of an emitter-follower is always slightly less than unity, because of the finite transconductance of the transistor. Essentially the intrinsic emitter resistance r_e (not to be confused with the physical component Re) forms a potential divider with the output load. It is simple to work out the small-signal gain at a given operating point. The value of r_e is given by $25/I_c$ (for I_c in mA) Since the value of r_e is inversely proportional to I_c, it varies with large signals, and it is one cause of the rather imperfect linearity of the simple emitter-follower. Heavier external loading increases the modulation of I_c, increases the gain variation, and so increases distortion.

Figure 3.2 shows the emitter-follower with an AC-coupled load. Its load driving capability is not very good. While the transistor can, within limits, source as much current as required into the load, the current-sinking ability is limited by the emitter resistor Re, which forms a potential divider with the load resistance R_{load}. With the values shown here, negative clipping occurs at about −10 V severely limiting the maximum output amplitude. The circuit shows a high-voltage low-beta transistor type running from ±24 V rails, to give worst-case performance results and to exploit the ability of discrete circuitry to run from high-voltage rails.

The simple emitter-follower has several factors that affect its distortion performance:

- Distortion is reduced as the emitter resistor is reduced, for a given load impedance.
- Distortion is reduced as the DC bias level is raised above the mid-point.
- Distortion increases as the load impedance is reduced.
- Distortion increases monotonically with output level.
- Distortion does NOT vary with the beta of the transistor. This statement assumes low-impedance drive, which may not be the case for an emitter-follower used as a buffer. If the source impedance is significant then beta is likely to have a complicated effect on linearity and not always for the worse.

Emitter-follower distortion is mainly second harmonic, except when closely approaching clipping. This is entirely predictable, as the circuit is asymmetrical. Only symmetrical

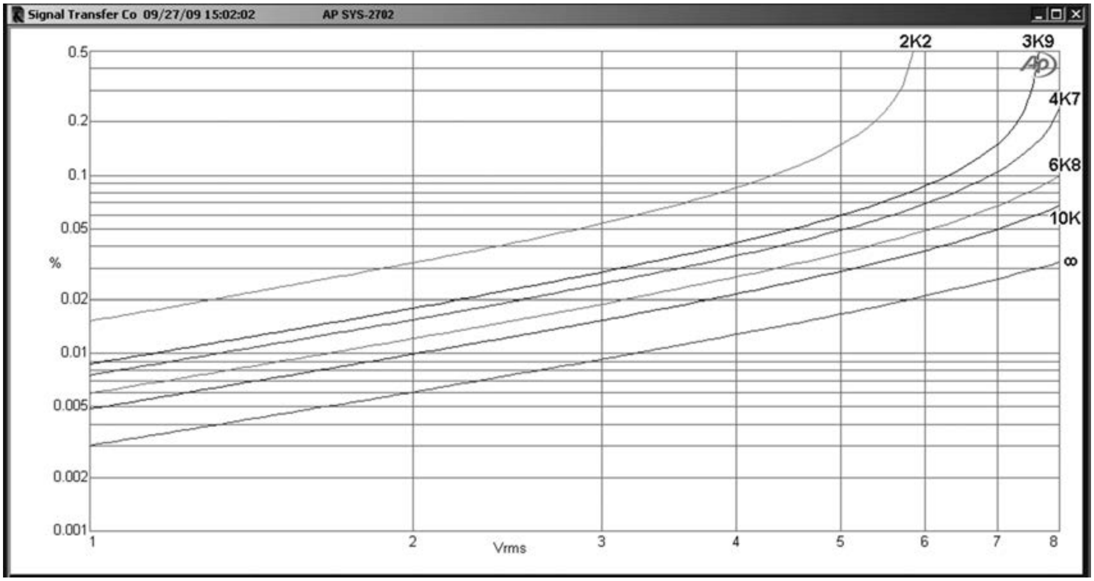


Figure 3.3: How various external loads degrade the linearity of the simple emitter-follower. Re = 2k7

configurations, such as the differential pair, restrict themselves to generating odd harmonics only, and then only when they are carefully balanced [6]. Symmetry is often praised as desirable in an audio circuit, but this is subject to Gershwin's Law: 'it ain't necessarily so'. Linearity is what we want in a circuit, and symmetry is not necessarily the best way to get it.

Because the distortion is mostly second harmonic, its level is proportional to amplitude, as seen in Figure 3.3. At 2 Vrms with no load it is about 0.006%, rising to 0.013% at 4 Vrms. External loading always makes the distortion worse, and more rapid as the amplitude approaches the clipping point; the THD is more than doubled from 0.021% to 0.050% at 6 Vrms, just by adding a light 6k8 load. For these tests Re was 2k7, as in Figure 3.2. For both the EF and CFP circuits, distortion is flat across the audio band so no THD/frequency plots are given.

Insight into what's happening can be gained by using SPICE to plot the incremental gain over the output swing, as in Figure 3.4. As loading increases the curvature of the gain characteristic becomes greater for a given voltage swing. It is obvious that the circuit is much more linear on the positive side of 0 V explaining why emitter-followers give less distortion when biased above the mid-point. This trick can be very useful if the full output swing is not required.

Most amplifier stages are biased so the quiescent output voltage is at the mid-point of the operating region, to allow the maximum symmetrical voltage swing. However, the asymmetry of the simple emitter-follower's output current-capability means that if there is significant loading, a greater symmetrical output swing is often possible if the stage is biased positive of 0 V.

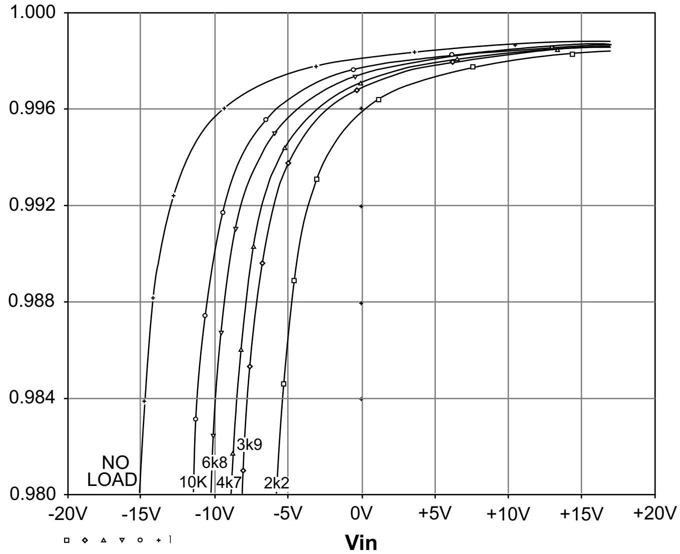


Figure 3.4: Incremental gain of the circuit in Figure 3.2, with different loads. A distortionless circuit would have constant gain and so give a horizontal line. SPICE simulation

If the output is loaded with 2k2, negative clipping occurs at −8 V, which allows a maximum output amplitude of only 5.6 Vrms. The unloaded output capability is about 12 Vrms. If the bias point is raised from 0 V to +5 V the output capability becomes roughly symmetrical and the maximum loaded output amplitude is increased to 9.2 Vrms. Don't forget to turn the output capacitor around.

The measured noise output of this stage with a 40 Ω source resistance is a commendably low −122.7 dBu (22 Hz–22 kHz) but with a base-stopper resistor (see below) of 1 kΩ this degrades to −116.9 dBu. A higher stopper of 2k7 gives −110.4 dBu.

The constant-current emitter-follower

The simple emitter-follower can be greatly improved by replacing the sink resistor Re with a constant-current source, as shown in Figure 3.5. The voltage across a current-source does not (to a good approximation) affect the current through it, so if the sink current is large enough a load can be driven to the full voltage swing in both directions.

The current source Q2 is biased by D1, D2. One diode cancels the V_{be} drop of Q2, while the other sets up 0.6 V across the 100 Ω resistor R2, establishing the quiescent current at 6 mA. The 22 kΩ resistor R3 in turn biases the diodes. This simple bias system works quite adequately if the supply rails are regulated, but might require filtering if they are not. The 22 kΩ value is non-critical; so long as the diode current exceeds the I_b of Q2 by a reasonable factor (say ten times) there will be no problem.

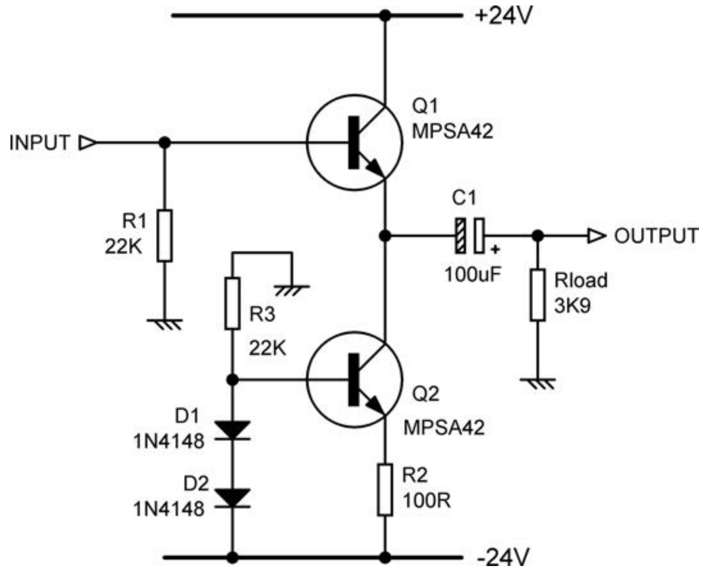


Figure 3.5: Emitter-follower with a constant-current source replacing the emitter resistor

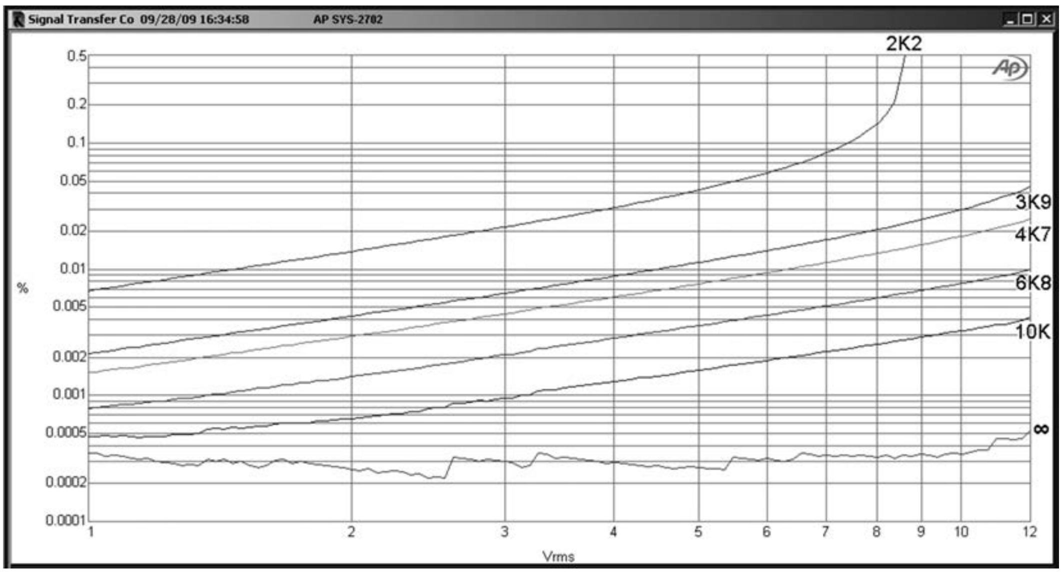


Figure 3.6: How various loads degrade the linearity of the current-source emitter-follower. Quiescent current 6 mA. Note scale change compared with Figure 3.3

Figure 3.6 shows that distortion is much reduced. With no external load, the 0.013% of the simple emitter-follower (at 4Vrms) has become less than 0.0003%, the measurement system noise floor. This is because the amount by which the Q1 collector current is modulated is very much less. The linearity of this emitter-follower is still degraded by increasing loading, but to a much lesser extent; with a significant external load of 4k7, the 0.036% of the simple emitter-follower (at 4Vrms) becomes 0.006%. The steps in the bottom (no load) trace are artefacts of the AP SYS-2702 measuring system.

The noise performance of this stage is exactly as for the simple emitter-follower above.

The push-pull emitter-follower

This is an extremely useful and trouble-free form of push-pull output; I have used it many times in preamplifiers, mixers, etc. I derived the notion from the valve-technology White cathode-follower, described by Nelson-Jones in a long-ago *Wireless World* [7]. The original reference is a British Patent taken out by Eric White in 1940 [8].

Figure 3.7 shows a push-pull emitter-follower. When the output is sourcing current, there is a voltage drop through the upper sensing resistor R5, so its lower end goes downwards in voltage. This is coupled to the current-source Q2 through C3, and tends to turn it off. Likewise, when the current through Q1 falls, Q2 is turned on more. This is essentially a negative-feedback loop with an open-loop gain of unity, and so by simple arithmetic the current variations in Q1, Q2 are halved, and this stage can sink twice the current of the constant-current version described above, while running at the same quiescent current.

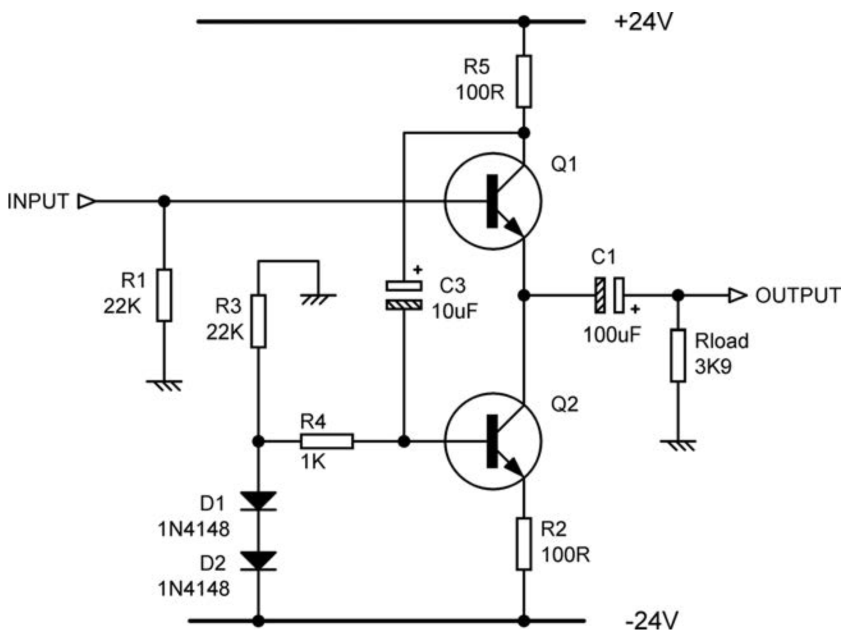


Figure 3.7: Circuit of push-pull emitter-follower. Quiescent current still 6 mA as before, but the load-driving capability is twice as great

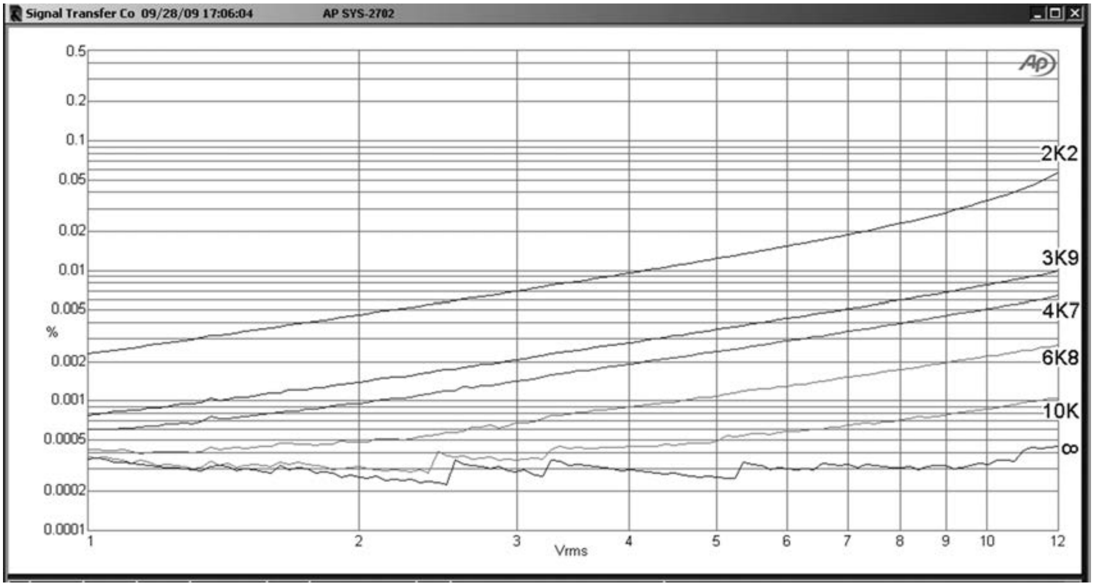


Figure 3.8: How loading degrades linearity of the push-pull emitter-follower. Loads from 10 kΩ to 2k2. Quiescent current 6 mA

The effect of loading on linearity is once again considerably reduced, and only one resistor and one capacitor have been added.

This configuration needs fairly clean supply rails to work, as any upper-rail ripple or disturbance is passed directly through C3 to the current-source, modulating the quiescent current and disrupting the operation of the circuit.

Push-pull action further improves the linearity of load driving; the THD with a 4k7 external load is halved from 0.006% for the constant-current version (4 Vrms) to 0.003%, at the same quiescent current of 6 mA, as seen in Figure 3.8. This is pretty good linearity for such simple circuitry.

Emitter-follower stability

The emitter-follower is about as simple as an amplifier gets, and it seems highly unlikely that it could suffer from obscure stability problems. However, it can, and often does. Emitter-followers are liable to RF oscillation when fed from an inductive source impedances. This oscillation is in the VHF region, usually in the area 100–400 MHz and will be quite invisible on the average oscilloscope; however a sure sign of this problem is unusually high distortion that varies strongly when the transistor is touched with a probing finger. One way to stop this is to put a 'base-stopper' resistor directly in series with the base. This should come after the bias resistor to prevent loss of gain. Depending on the circuit conditions, the resistor may be

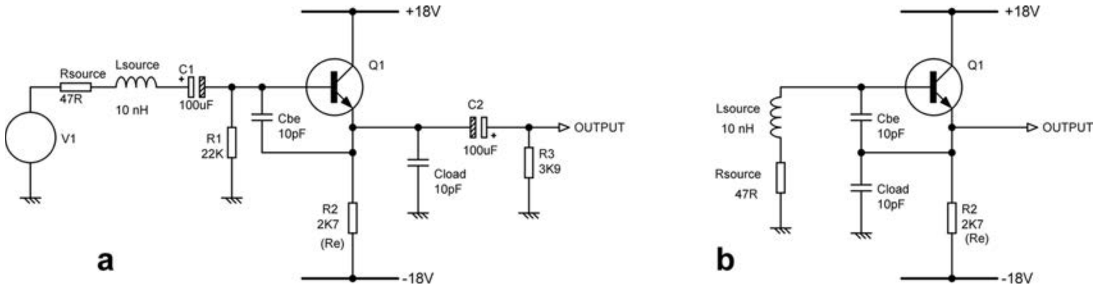


Figure 3.9: Emitter-follower oscillation: the effective circuit at a) bears a startling resemblance to the Colpitts oscillator at b)

as low as 100 Ω or as high as 2k7. The latter generates −119.6 dBu of Johnson noise which in itself is inconvenient in low-noise circuitry, but the effects of the transistor noise current flowing through it are likely to be even worse. Base-stopper resistors are not shown in the following diagrams to aid clarity but you should always be aware of the possible need for them. This also applies to the CFP configuration which is equally, if not more, susceptible to the problem.

The instability is due to the fact that the typical emitter-follower is fed from a source with some inductance, and has some capacitive loading, even if it is only due to stray capacitance, as in Figure 3.9a, where the transistor internal base-emitter capacitance C_{be} is included. If this is redrawn as in Figure 3.9b, it is the classic circuit of a Colpitts oscillator.

For more information on this phenomenon see Feucht [9] and de Lange [10].

CFP emitter-followers

The simple emitter-follower is lacking both in linearity and load-driving ability. The first shortcoming can be addressed by adding a second transistor to increase the negative feedback factor by increasing the open-loop-gain. This also allows the stage to be configured to give voltage gain, as the output and feedback point are no longer inherently the same. This arrangement is usually called the 'complementary feedback pair' (hereafter CFP) though it is sometimes known as the Szilaki configuration. This circuit can be modified for constant-current or push-pull operation exactly as for the simple emitter-follower.

Figure 3.10 shows an example. The emitter resistor Re is the same value as in the simple emitter-follower to allow meaningful comparisons. The value of R4 is crucial to good linearity, as it sets the I_c of the first transistor, and determines its collector loading. The value of 3k3 shown here is a good compromise.

This circuit is also susceptible to emitter-follower oscillation, particularly if it sees some load capacitance, and will probably need a base-stopper. If 1 kΩ does not do

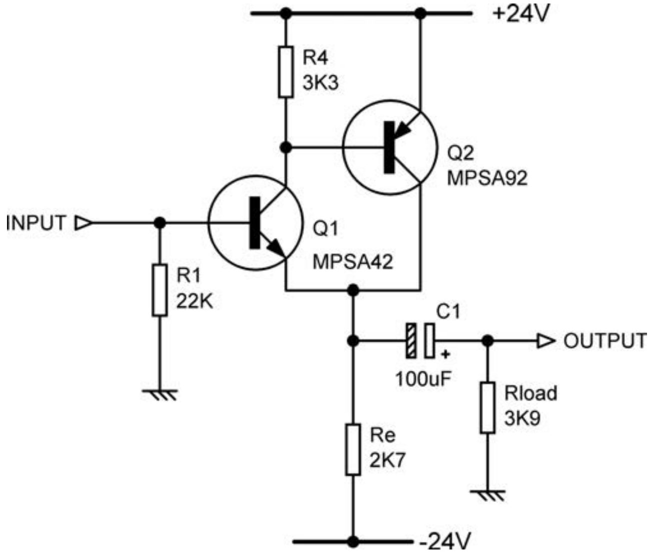


Figure 3.10: The CFP emitter-follower. The single transistor is replaced by a pair with 100% voltage feedback to the emitter of the first transistor

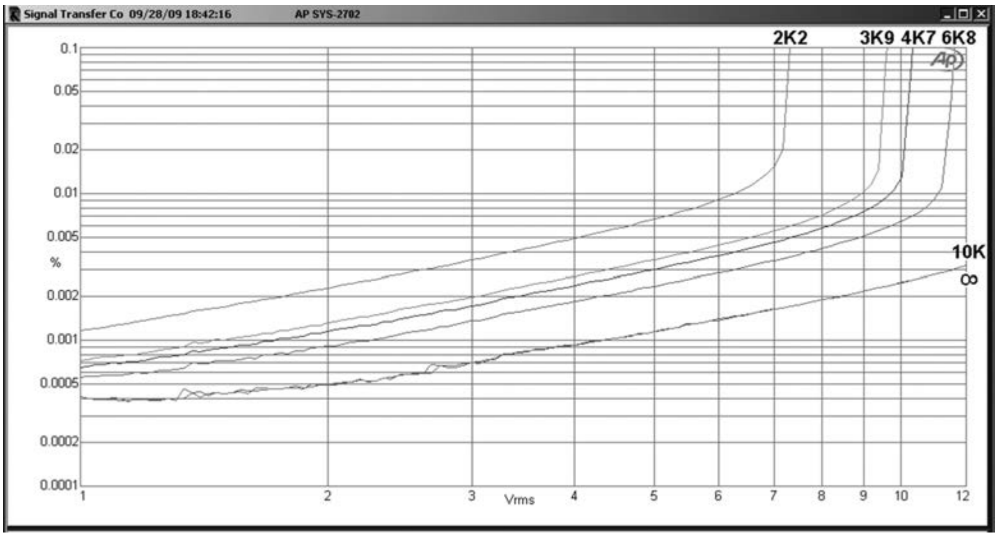


Figure 3.11: How loading effects the distortion of a CFP emitter-follower. THD at 6 Vrms, 6k8 load is only 0.003% compared with 0.05% for the simple EF. Re is 2K7.

the job, try adding a series output resistor of 100 Ω close to the stage to isolate it from load capacitance.

Figure 3.11 shows the improved linearity; Figure 3.12 is the corresponding SPICE simulation. The measured noise output with a 100 Ω base-stopper is −116.1 dBu.

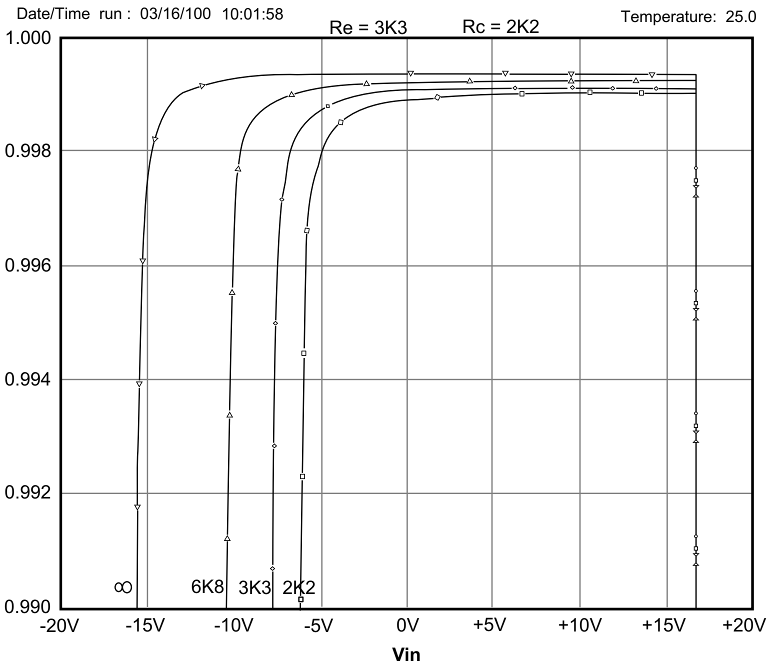


Figure 3.12: SPICE simulation of the circuit in Figure 3.10, for different load resistances. The curves are much flatter than those in Figure 3.4, even though the vertical scale has been expanded

If we replace Re with a 6 mA current-source, as in Figure 3.13, we once more get improved linearity and load-driving capability, as shown in Figure 3.14. The 6 Vrms, 6k8 THD is now only just above the noise at 0.0005% (yes, three zeros after the point. Three-transistor circuitry can be rather effective).

Converting the constant-current CFP to push-pull operation as in Figure 3.15 gives another improvement in linearity and load driving. Figure 3.16 shows that now only the results for 2k2 and 3k9 loading are above the measurement floor.

Improved unity-gain buffers

There is often a need for a unity-gain buffer with very low distortion. If neither the simple emitter-follower, made with one transistor, nor the CFP configuration with two transistors are adequately linear, we might ponder the advantages of adding a third transistor to improve performance, without going to the complexity of the discrete opamps described later in this chapter (in our transistor count we are ignoring current-sources used to create the active output loads that so improve linearity into significant external loading).

One promising next step is a three-transistor configuration that is often called the Schlotzaur configuration; see Feucht [5] and Staric and Margan [11]. The single input transistor in the

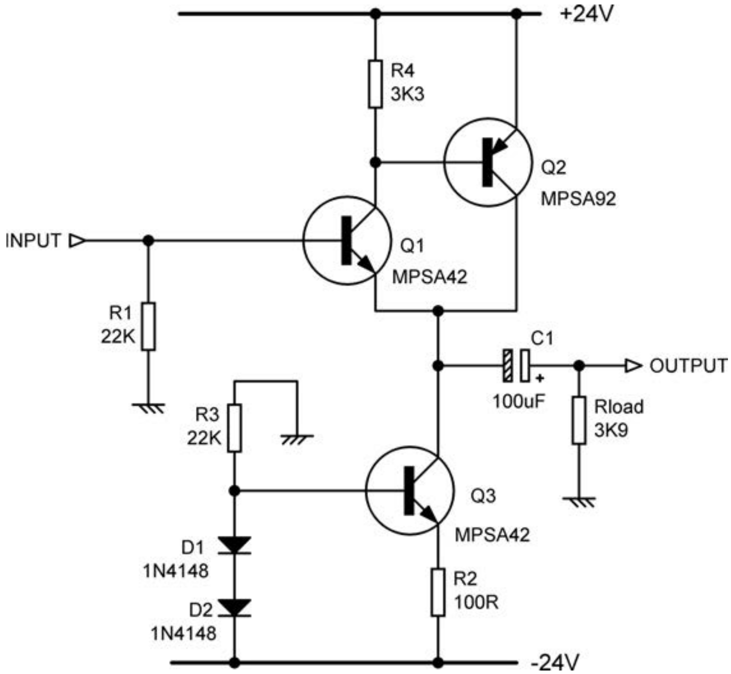


Figure 3.13: Constant-current CFP follower. Once more the resistive emitter load is replaced by a constant-current source to improve current-sinking

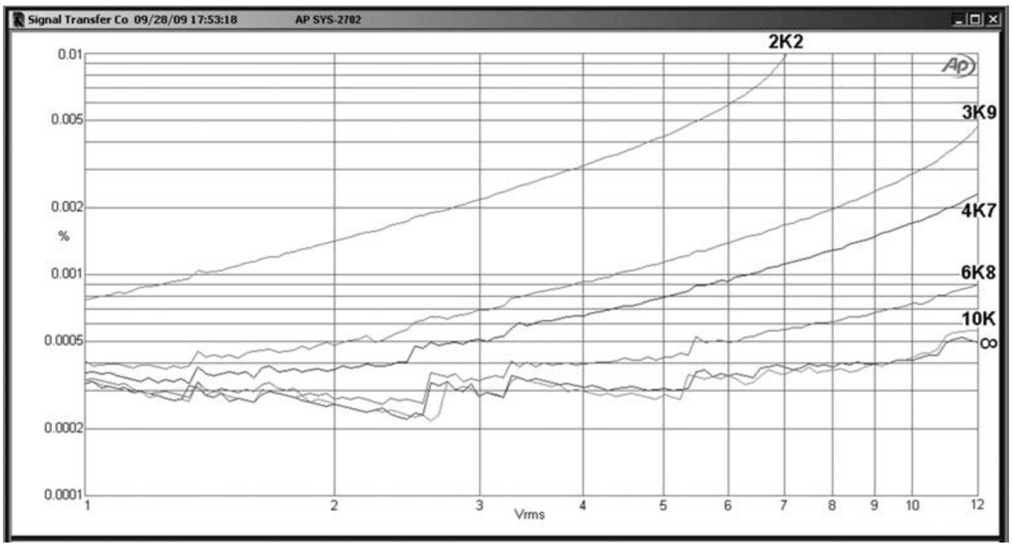


Figure 3.14: Distortion and loading effects on the CFP emitter-follower with a 6 mA current-source. The steps on the lower traces are artefacts caused by the measurement system gain-ranging as it attempts to measure the THD of pure noise. Note change of scale

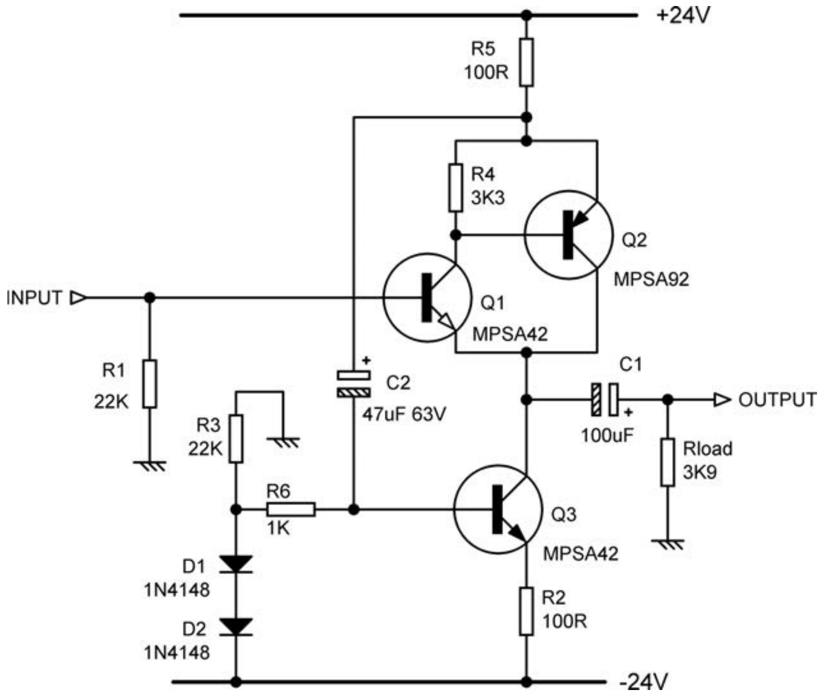


Figure 3.15: Circuit of a push-pull CFP follower. This version once more gives twice the load-driving capability for no increase in standing current

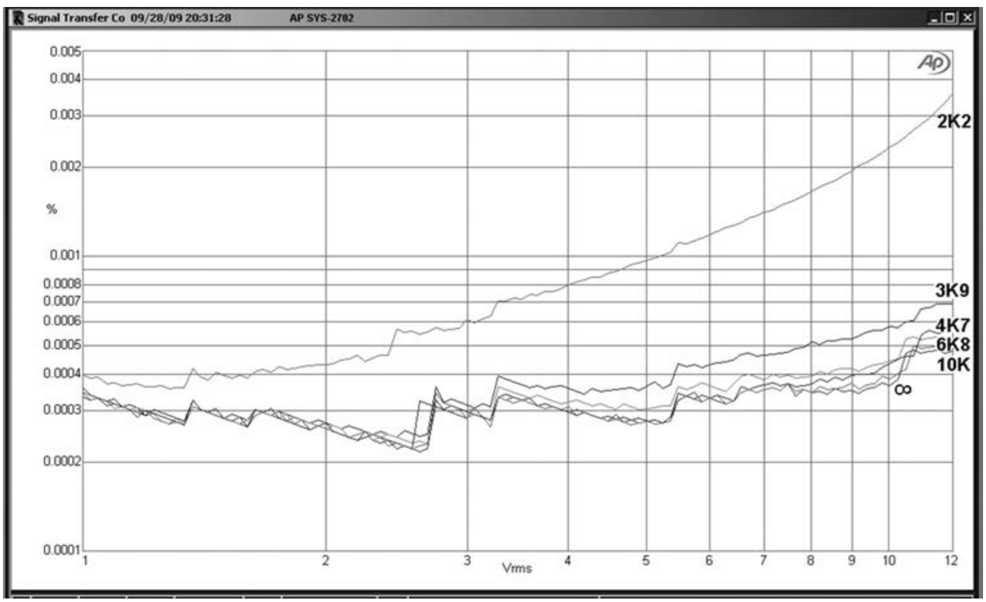


Figure 3.16: Distortion and loading effects on the push-pull CFP emitter-follower, still with 6 mA of quiescent current

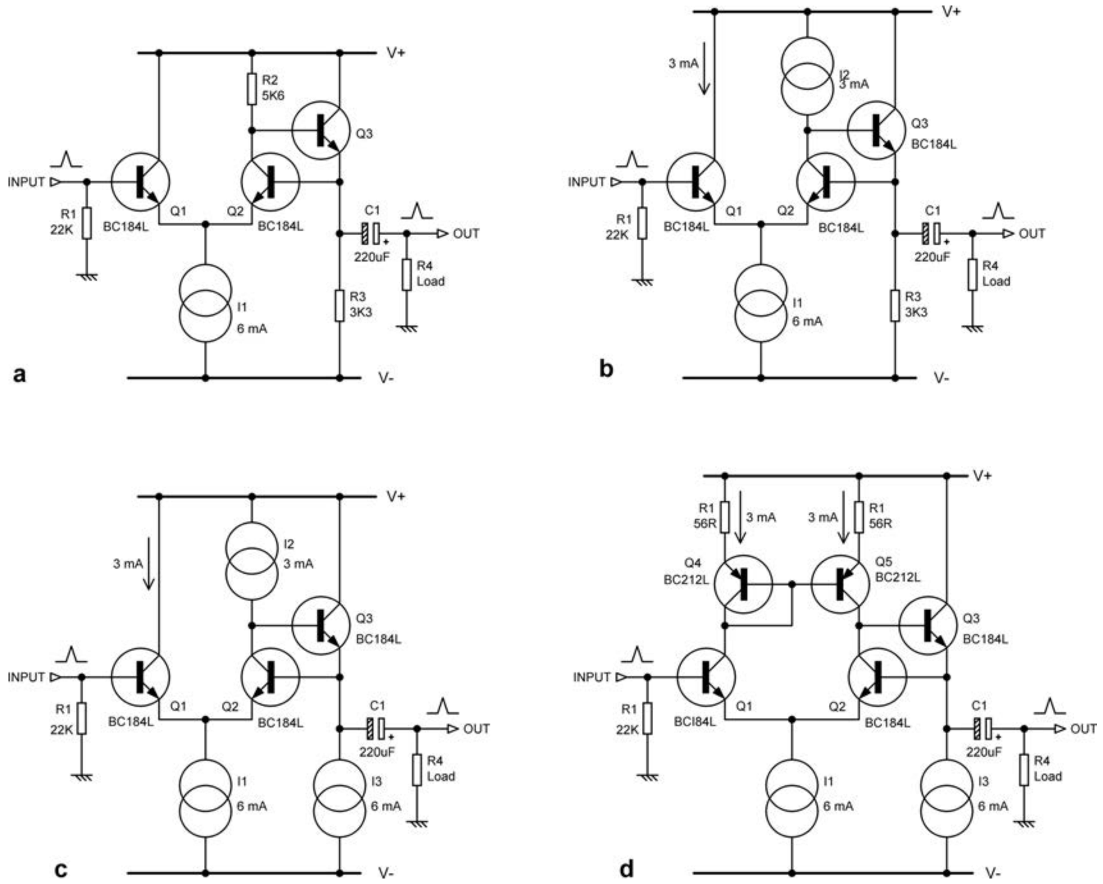


Figure 3.17: Developing a unity-gain buffer design, which replaces the single input transistor with a long-tail-pair: a) simple Schlotzaur circuit, b) collector load R2 replaced with current-source, c) output current-source added, d) inserting current-mirror in input pair collectors

CFP emitter-follower is now replaced with a long-tail-pair with 100% feedback to Q2, as in Figure 3.17a. Because LTPs have the property of cancelling out their even-order distortion, [6] you might expect a considerable improvement. You would be wrong – running on ±15 V rails we get 0.014% at 6 Vrms unloaded, almost flat across the audio band. The linearity is much inferior to the CFP emitter-follower. The open-loop gain, determined by measuring the error voltage between Q1, Q2 bases, is 249 times.

It's not a promising start, but we will persist! Replace the collector load R2 with a current-source of half the value of the tail current-source as in Figure 3.17b, and the linearity is transformed, yielding 0.00075% at 6 Vrms unloaded on ±15 V rails. Increasing the rails to ±18 V gives 0.00066% at 6 Vrms, and a further increase to ±24 V as high-voltage operation is part of what discrete design is all about, gives 0.00042% at 6 Vrms (unloaded). It's the increase in the positive rail that gives the improvement. Reducing the measurement bandwidth from 80 kHz to 22 kHz for a 1 kHz signal eliminates some noise and gives a truer figure of 0.00032% at 6 Vrms. The open-loop gain is increased to 3400 times.

Adding loading to the buffer actually has very little effect on the linearity, but its output capability is clearly limited by the use of R3 to sink current, just as for a simple emitter-follower. Replacing R3 with a 6 mA constant-current source, as for previous circuits, much improves drive capability and also improves linearity somewhat. See Figure 3.17c. On ±30 V rails we get a reading of 0.00019% at 5 Vrms with a 2k2 load, and that is mostly the noise in a 22 kHz bandwidth.

Finally we remove current-source I2 and replace it with a simple current-mirror in the input pair collectors, as in Figure 3.17d, in the pious hope that the open-loop gain will be doubled and distortion halved; On ±30 V rails there is a drop in THD from 0.00019% to 0.00017% at 5 Vrms with a 2k2 load, (22 kHz bandwidth) but that is almost all noise and I am pushing the limits of even the magnificent Audio Precision SYS-2702. This goes to show that there are other ways of designing low-distortion circuitry apart from hefting a bucket of opamps.

Figure 3.18 shows the distortion plot at 5 Vrms. Using an 80 kHz bandwidth so the HF end is meaningful means the readings are higher, and virtually all noise below 10 kHz. It also gives another illustration of the distortion generated by under-sized coupling capacitors. C1 started as 22 μF, but you can see that 220 μF is required to eliminate distortion at 10 Hz with a 2k2 load.

SPICE analysis shows that the collector currents of the input pair Q1, Q2 are somewhat unbalanced by the familiar base-current errors of a simple current-mirror. Replacing the

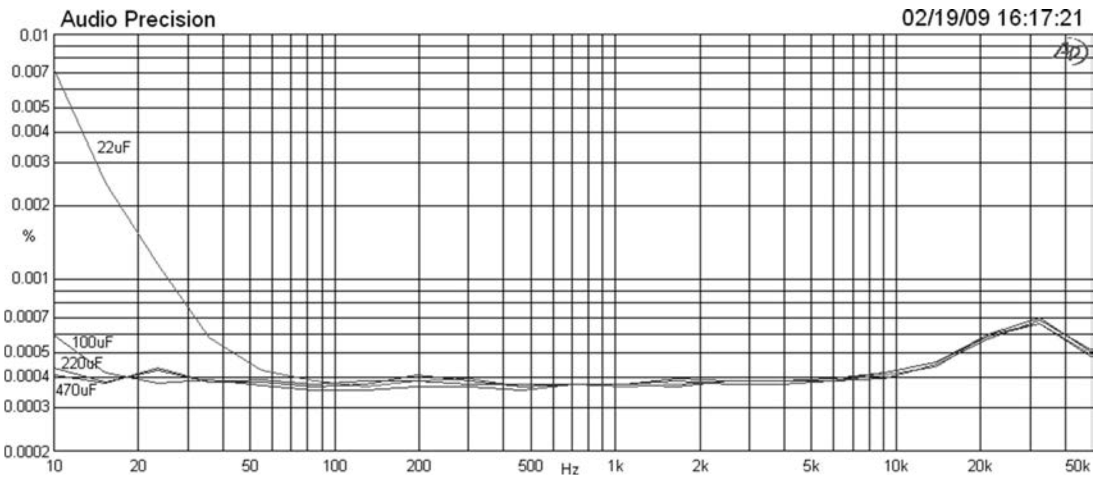


Figure 3.18: THD plot for the final version of the buffer, at 5 Vrms with a 2k2 load. The rising LF curves illustrate the distortion generated by under-sized output capacitors. Bandwidth 80 kHz

simple current-mirror with the well-known Wilson improved mirror might get us further improvement, if we could measure it. Work in progress . . .

The circuit we have now could still be regarded as a much-enhanced emitter-follower, but it is probably more realistic to consider it as a two-stage discrete opamp.

Gain stages

This section covers any discrete transistor stage that can give voltage gain. It may be as simple as a single transistor or as complex as an opamp implemented with discrete components. A single transistor can give voltage gain in either series or shunt mode.

One-transistor shunt-feedback gain stages

Single-transistor shunt-feedback gain stages are inherently inverting, and of very poor linearity by modern standards. The circuit in Figure 3.19 is inevitably a collection of compromises. The collector resistor R4 should be high in value to maximise the open-loop gain; but this reduces the collector current of Q1, and thus its transconductance, and hence reduces open-loop gain once more. The collector resistor must also be reasonably low in value as the collector must drive external loads directly. Resistor R2, in conjunction with R3, sets the operating conditions. This stage has only a modest amount of shunt feedback via R3, and the input can hardly be called a virtual earth. However such circuits were once very common in low-end discrete preamplifiers, back in the days when the cost of an active device was a serious matter. Such stages are still occasionally found doing humble jobs like driving VU meters, but the cost advantage over a opamp section is small, if it exists at all.

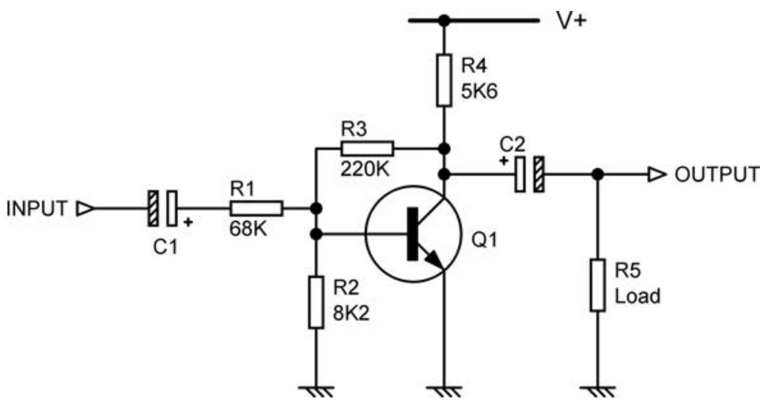


Figure 3.19: Circuit of single-transistor gain stage, shunt-feedback version. +24 V rail

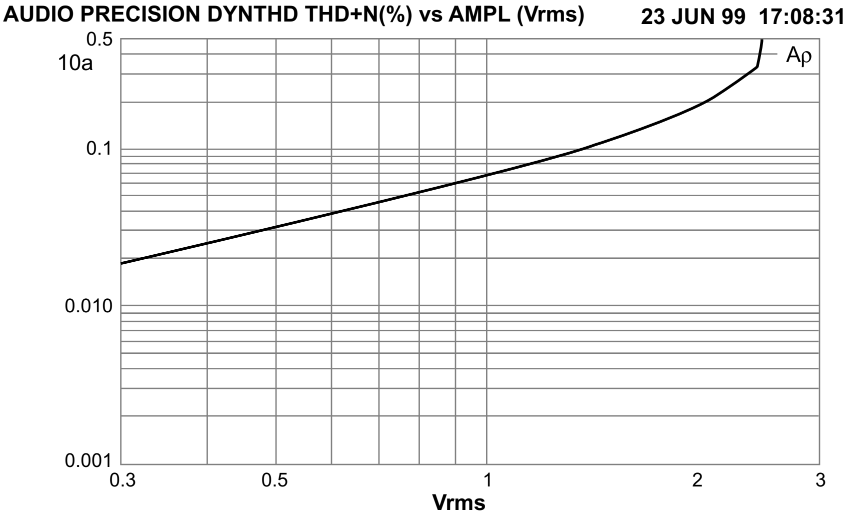


Figure 3.20: Single-transistor shunt-feedback gain stage, distortion versus level. +24 V rail. Gain is 2.3 times

The gain of the stage in Figure 3.19 is, at a first look, 220k/68k = 3.23 times, but the actual gain is 2.3x (with no load) due to the small amount of open-loop gain available. The mediocre distortion performance that must be expected even with this low gain is shown in Figure 3.20.

One-transistor series-feedback gain stages

Single-transistor series-feedback gain stages are made by creating a common-emitter amplifier with a feedback resistor that gives series voltage feedback to the emitter. The gain is the ratio of the collector and emitter resistors, if loading is negligible. Note that unlike opamp-based series-feedback stages, this one is inherently inverting. To make a non-inverting stage with a gain more than unity requires at least two transistors, because of the inversion in a single common-emitter stage.

This very simple stage shown in Figure 3.21 naturally has disadvantages. The output impedance is high, being essentially the value of the collector resistor R3. The output is neither good at sinking or sourcing current.

The distortion performance as seen in Figure 3.22 is indifferent, giving 0.3% THD at 1 Vrms out. Compare this with the shunt-feedback one-transistor gain stage above, which gives 0.07% under similar conditions.

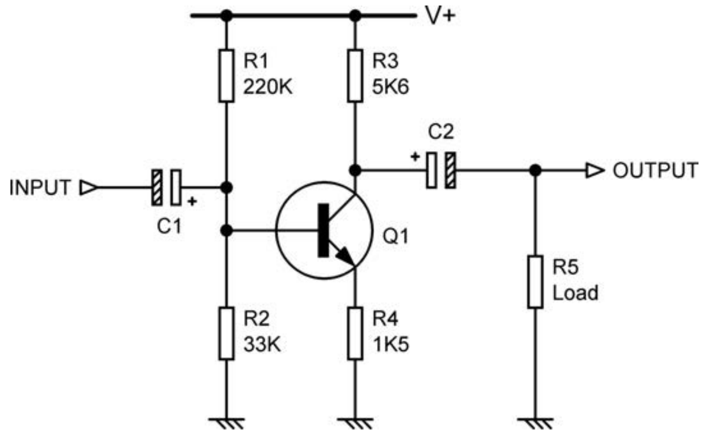


Figure 3.21: Circuit of single-transistor series-feedback gain stage. +24 V rail

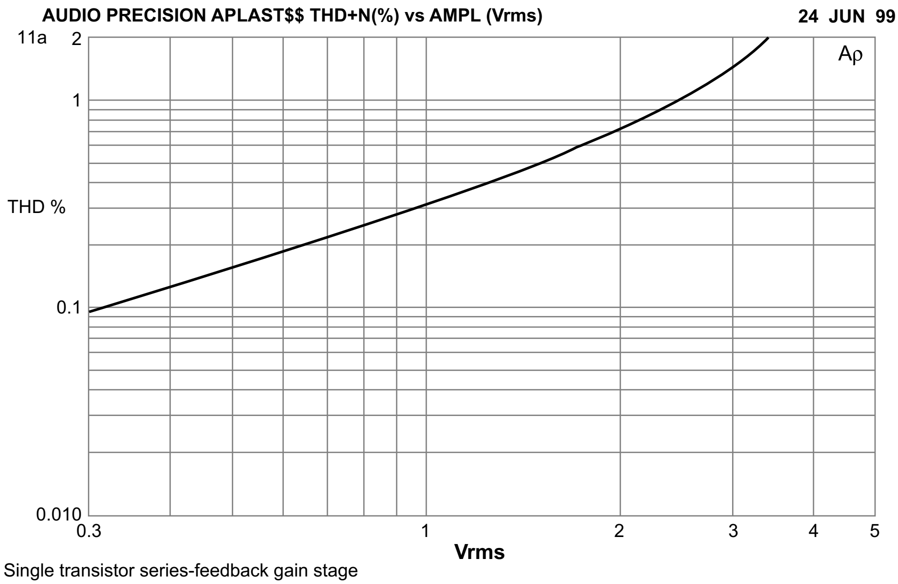


Figure 3.22: Single-transistor series-feedback gain stage, distortion versus input level. +24 V rail. Gain is three

Two-transistor shunt-feedback gain stages

Before the advent of the opamp, inverting stages were required for tone controls and virtual-earth summing amplifier. The one-transistor amplifier stage already described is deficient in distortion and load-driving capability. A much better amplifier can be made with two

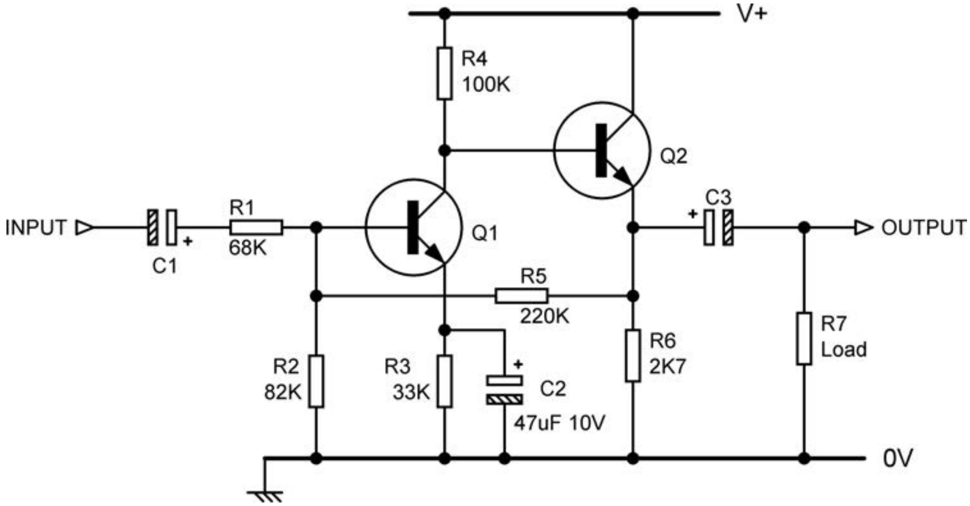


Figure 3.23: Two-transistor gain stage with shunt-feedback. +24 V rail. Gain is three

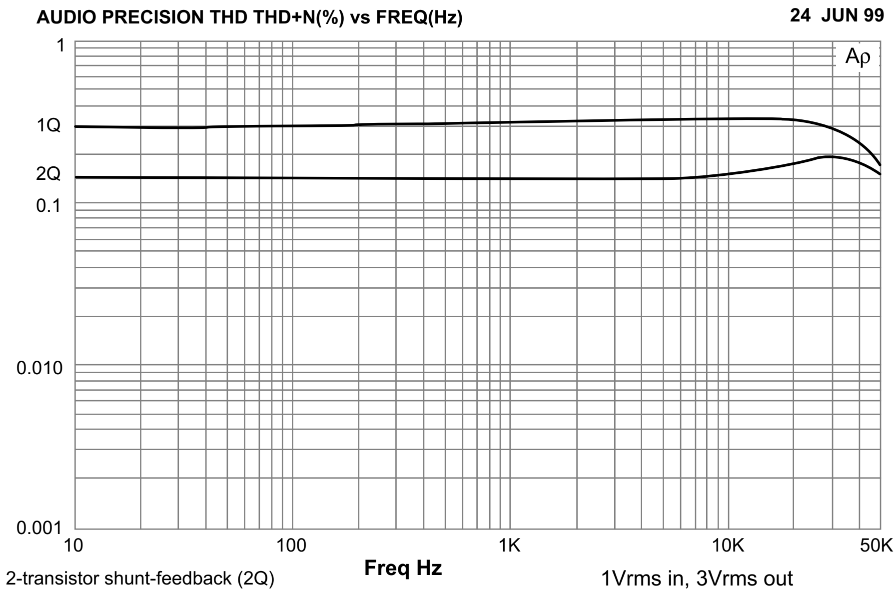


Figure 3.24: Distortion of two-transistor shunt stage versus freq (2Q). Distortion of one-transistor version (1Q) is also shown. The two-transistor version is only twice as good, which seems a poor return for the extra active device. +24 V rail

transistors, as in Figure 3.23. The voltage gain is generated by Q1, which has a much higher collector resistor R4 and so much higher gain. This is possible because Q2 buffers it from external loading, and allows a higher NFB factor. See Figure 3.24 for the distortion performance.

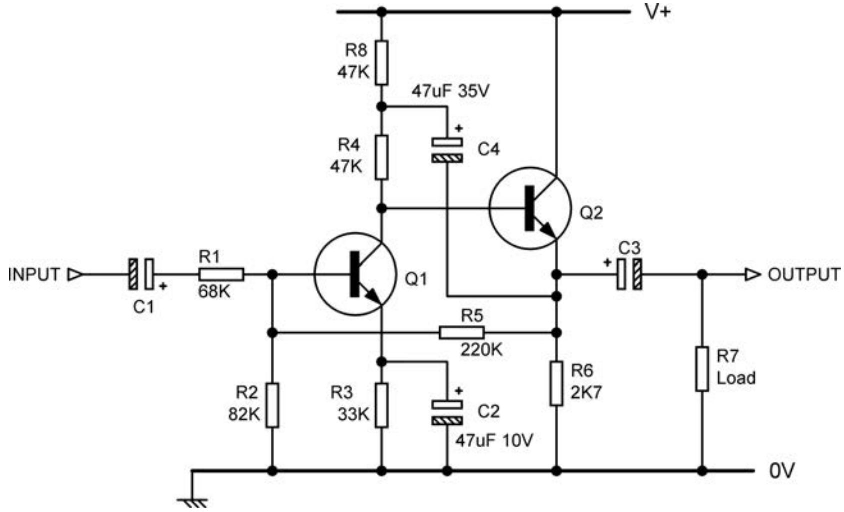


Figure 3.25: Two-transistor shunt-feedback stage, with bootstrapping added to the first stage to improve linearity. Q1 collector current is 112 μA. +24 V rail

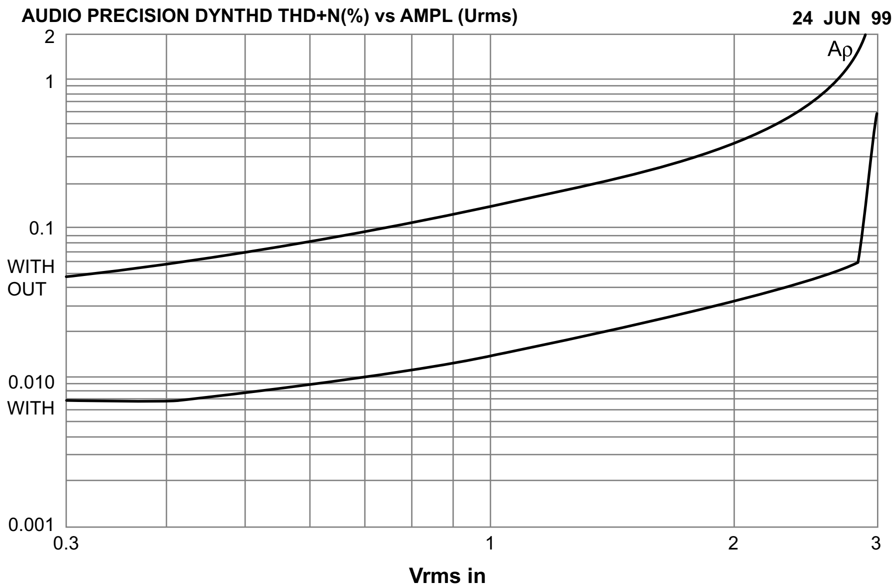


Figure 3.26: Distortion of the two-transistor shunt stage versus level, with and without bootstrapping. X-axis is input level; output is three times this. +24 V rail

With the addition of bootstrapping, as shown in Figure 3.25, the two-transistor stage has its performance transformed. Figure 3.26 above shows how THD is reduced by a factor of ten; 0.15% at 1 Vrms in, 3 Vrms out becomes 0.015%, which is much more respectable. The improvement is due to the increased voltage-gain of the first stage giving a higher NFB factor. THD is still approximately proportional to level, as the distortion products are mainly second harmonic. Clipping occurs abruptly at 2.9 Vrms in, 8.7 Vrms out; abrupt clipping onset is characteristic of stages with a high NFB factor. Stages like this were commonly used as virtual-earth summing amplifiers in mixing consoles before acceptable opamps were available at reasonable cost.

Figure 3.27 shows that the distortion is further reduced to 0.002% at 3 Vrms out if the impedance of the input and feedback networks is reduced by ten times. SPICE simulation confirms that this is because the signal currents flowing in R1 and R5 are now larger compared with the non-linear currents drawn by the base of Q1. There is also a noise advantage.

Figure 3.28 shows how output drive capability can be increased, as before, by replacing R6 with a 6 mA current-source. The input and feedback resistors R1, R5 have again been scaled down by a factor of 10. Push-pull operation can also be simply implemented as before. The EIN of this version is −116 dBu.

The constant-current and push-pull options can also be added to more complex discrete stages. Some good examples can be found in a preamplifier design of mine [12].

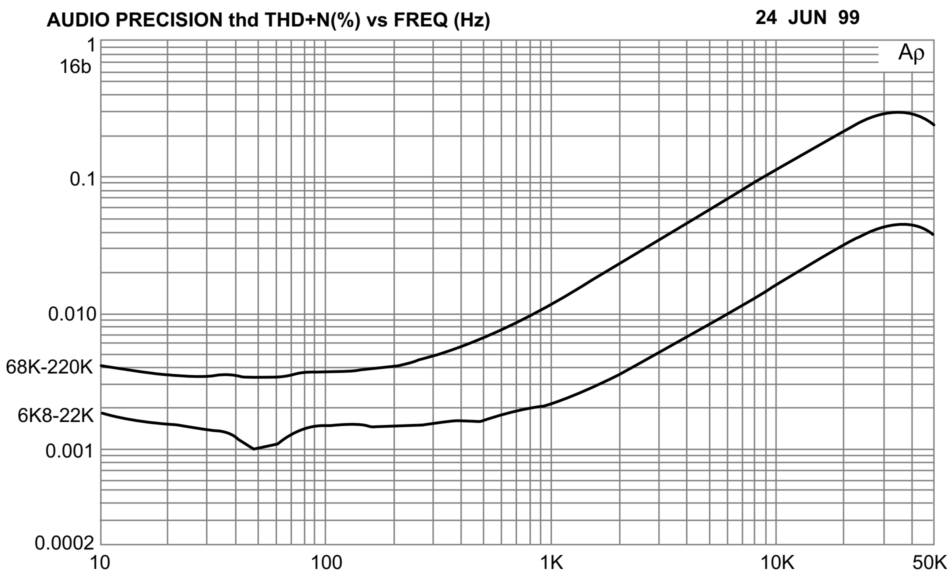


Figure 3.27: The distortion is reduced by a further factor of at least five if the impedance of the input and feedback networks is reduced by ten times. +24 V rail. Output 3 Vrms

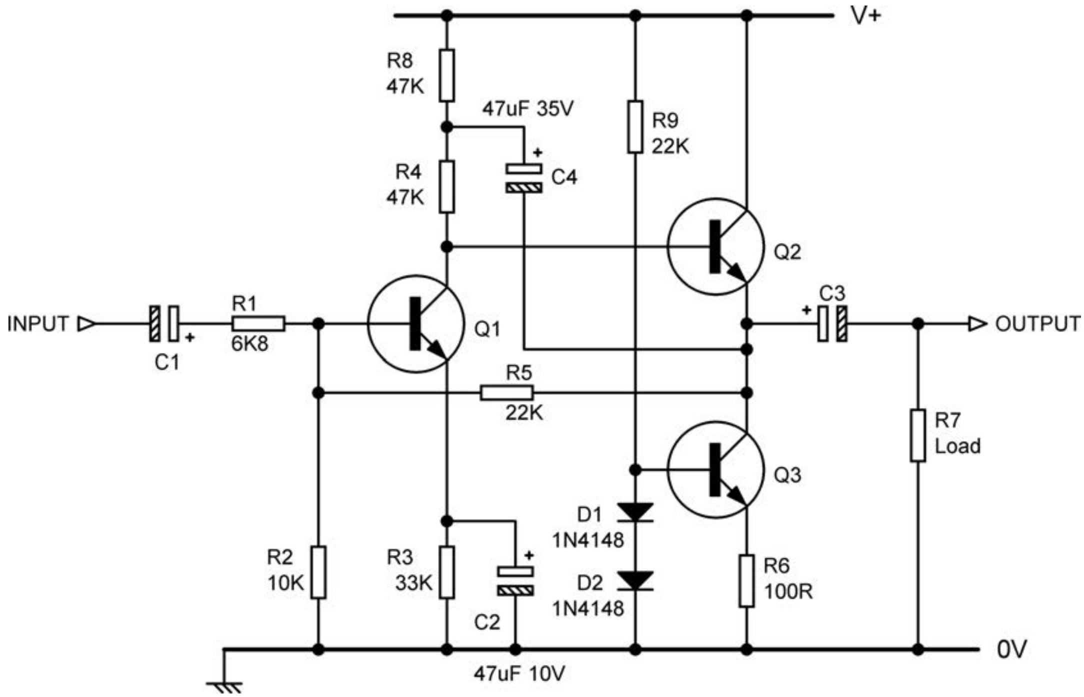


Figure 3.28: Two-transistor bootstrapped shunt-feedback configuration, with low-impedance feedback network, and with current-source output to enhance load driving capability. +24 V rail

Two-transistor shunt-feedback stages: improving linearity

As is usual with discrete configurations, the quickest way to improve linearity is to crank up the supply voltage, if that is feasible. This obviously increases the power consumed by the circuitry, but this is not normally a major issue. Figure 3.29 shows the powerful affect of increasing the rail voltage on both the LF and HF distortion regimes. It is interesting to note that this simple configuration shows HF distortion (mainly second harmonic) increasing at 6 dB/octave with frequency, despite the simplicity of the circuit and the apparent absence of compensation capacitance. Open-loop gain does fall off at HF due to the internal C_{bc} of Q1, but the major source of HF distortion is the non-linear nature of this C_{bc} which varies strongly in capacitance with the voltage across it. There is more on this vital point in the later section on discrete opamps.

Table 3.1 summarises how the distortion at 1 kHz drops quickly at first, as the supply voltage is increased, but the improvement slows down at higher voltages. The input was 1 Vrms and the output 3.23 Vrms. These measurements were all taken with the emitter-resistor R6 at 2k7, as shown in Figure 3.35. R1 was 6k8 and R5 was 22 kΩ, as for Figure 3.27 and in Figure 3.28 above.

TABLE 3.1 Reduction in distortion at 1 kHz for 3.23 Vrms out versus supply rail voltage

Supply voltage (V)	THD 1 kHz (%)	THD reduction ratio ref +24 V
+24	0.00435	1.00
+30	0.00324	0.74
+36	0.00262	0.60
+40	0.00228	0.52
+45	0.00201	0.46
+48	0.00188	0.43

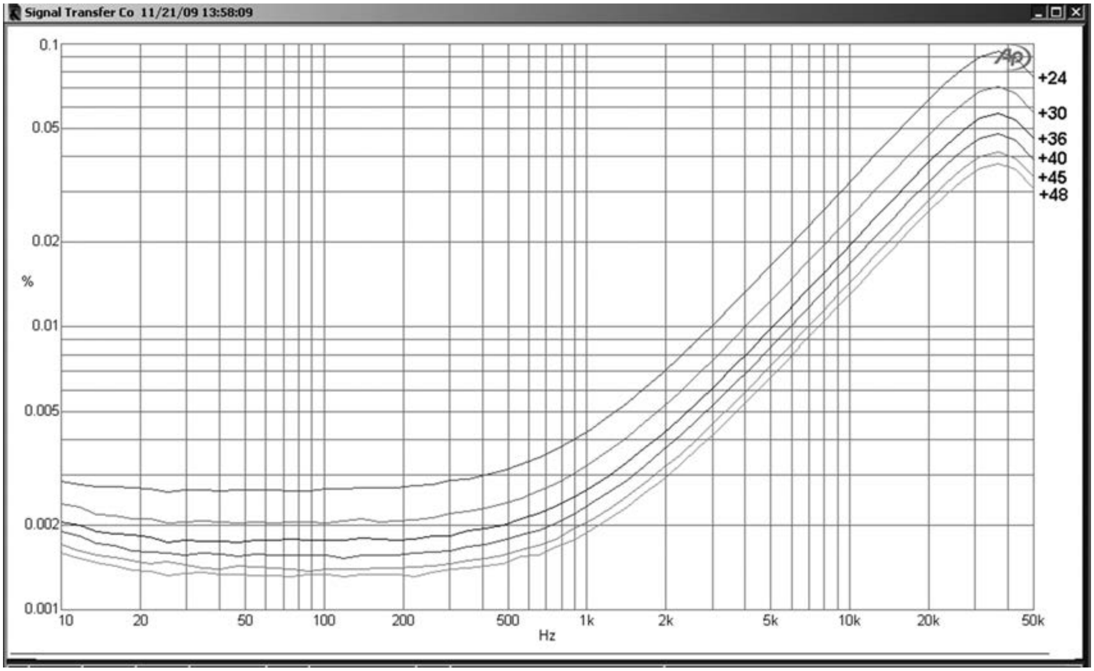


Figure 3.29: Distortion performance of two-transistor bootstrapped shunt-feedback stage, with varying supply voltage. Output = 3.23 Vrms

In this configuration there are two stages, both of which have a certain curvature to their in/out characteristics. With discrete design you have complete control over both of them, and there is always the possibility that you might be able to reduce distortion by altering the curvatures so that some degree of cancellation occurs in the distortion produced by each stage. This will be most effective on second-harmonic distortion.

A convenient way to alter the curvature of the second stage is to vary the value of the emitter resistor R6 in Figure 3.25. The results of doing this are shown in Table 3.2; the input level

TABLE 3.2 Reduction in distortion at 1 kHz for 3.23 Vrms out on changing R6 and supply rail voltage

Supply voltage (V)	R6 value	THD 1 kHz (%)	THD reduction ratio
+24	3k3	0.00466	1.09
+24	2k7	0.00435	1.00 (reference)
+24	2k2	0.00378	0.88
+24	2k0	0.00360	0.84
+30	2k0	0.00287	0.67
+36	2k0	0.00228	0.53
+40	2k0	0.00204	0.48
+45	2k0	0.00179	0.42
+48	2k0	0.00167	0.39

was 1 Vrms and the output level 3.23 Vrms. The measurements in this table were all taken with R6 set to 2k7, as shown in Figure 3.25.

The second row shows the standard results, as obtained from Figure 3.25 with R6 set to 2k7. Our first experiment is to raise its value to 3k3, but this is clearly a step in the wrong direction as THD increases from 0.00435% to 0.00466%. We therefore try reducing R6 to 2k2, and get an immediate improvement to 0.00378%. Pressing on further, we reduce R6 to 2k0, and THD falls again, but this time by a smaller amount, giving 0.00360%. It looks as if further decreases will yield diminishing benefits, and the increased standing-current in the emitter-follower may have serious consequences for power dissipation at higher supply voltages. Looking at the first four rows of the table, it is clear we get a useful increase in distortion performance simply by changing one component value, for no extra cost at all.

Since it appears we have gone about as far as we can in tweaking the output emitter-follower, we can consider a more radical step. As a general rule, any emitter-follower in a discrete transistor configuration can be inverted- in other words replaced by its complementary equivalent. This can be very useful when attempting to cancel the distortion from different stages, as just described. The result of this move is shown in Figure 3.30, where the output emitter follower Q2 is now a PNP device.

Unfortunately, we discover when we do the measurements that we have royally messed up an existing distortion cancellation rather than improved it, as shown in Table 3.3 (the 'THD ratio' is based on the original configuration; the second row of Table 3.2 is the reference). The distortion with a +24 V supply rail is almost doubled, and clearly we are reinforcing the curvature of the two stages rather than partially cancelling it. Increasing the supply voltage still improves the linearity, as it almost always will.

TABLE 3.3 THD versus supply voltage with inverted emitter-follower. R6 = 2k7

Supply voltage (V)	THD 1 kHz (%)	THD ratio
+24	0.00850	1.99
+30	0.00711	1.66
+36	0.00589	1.38
+40	0.00521	1.22

TABLE 3.4 THD versus R6 value with inverted emitter-follower. R6 = 2k7

Supply voltage (V)	R6 value	THD 1 kHz (%)	THD ratio
+24	2K2	0.00882	2.06
+24	2K7	0.00850	1.99
+24	3K3	0.00806	1.89

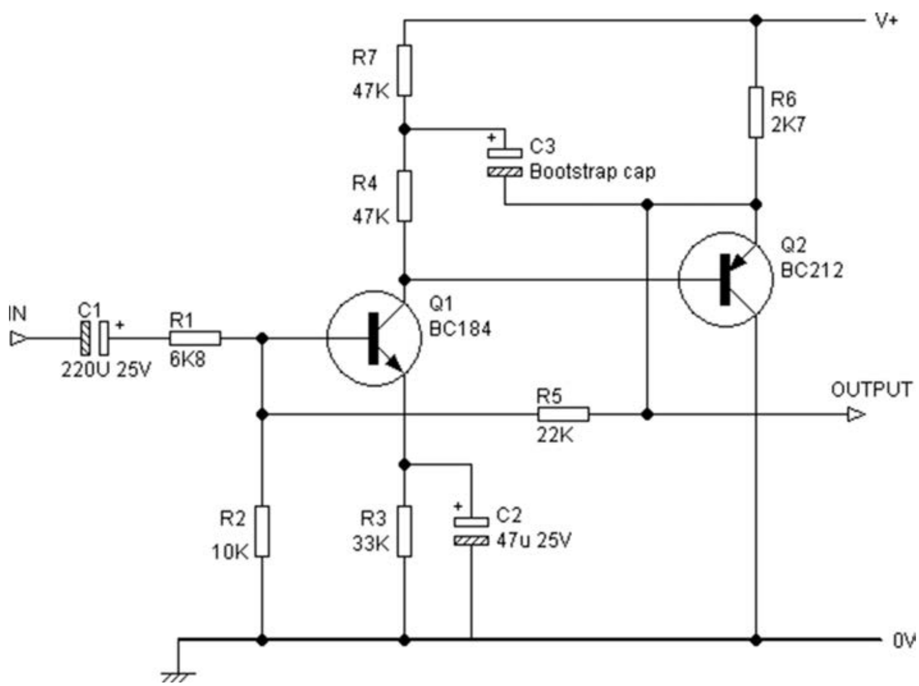


Figure 3.30: Two-transistor shunt-feedback amplifier with the emitter-follower Q2 inverted. Note R1 and R5 have the reduced values

We can again attempt to improve the linearity by modifying the value of the output emitter resistor R6, but it really doesn't help very much. The deeply unimpressive results are seen in Table 3.4. The 'THD ratio' reference is again the second row of Table 3.2.

Clearly in this case the conventional configuration with two NPN transistors is the superior one. It was and is the most common version encountered, so sometimes conventional is optimal.

Two-transistor shunt-feedback stages: noise

The noise output of the shunt-feedback circuit in Figure 3.28 with R1 = 6k8 and R5 = 22 kΩ is −99.5 dBu. To work out the equivalent input noise (EIN) we need to know the actual noise gain at which the circuit works, not the closed-loop gain. The latter is simply (R5/R1) but the apparent noise gain is higher at (R5/R1) +1, which evaluates as +12.5 dB, as it would be for the equivalent opamp circuit. The noise gain is actually rather higher, because we must allow for the presence of the biasing resistor R2. This raises the true noise gain to +16.2 dBu. In the typical application for this kind of circuit, that of virtual-earth summing amp, there would have been many input resistances R1, and the presence of R2 would have made relatively little difference to the overall noise performance.

Armed with the true noise gain, we can work out the EIN as −115.7 dBu, which is basically the noise performance of the first stage, as it implements all the voltage gain. It has to be said that I have so far made no attempt to optimise the noise performance of the circuit shown here. Increasing the first-stage collector current would probably reduce the noise with the feedback values shown.

Two-transistor shunt-feedback stages: bootstrapping

Choosing the right size for a bootstrap capacitor is not quite as straightforward as it appears. It looks as if quite a small value could be used because of the high impedance of the Q1 collector load. In Figures 3.25 and 3.28, the bootstrap capacitor C4 effectively sees only the 47 kΩ impedance of R8 to the supply rail, and a 2μ2 capacitor in conjunction with this gives −3 dB frequency of 1.54 Hz, which looks ample.

But it is not. Figure 3.31 shows that this value gives an enormous rise in LF distortion, reaching 0.020% at 10 Hz. The reason for this steep rise is that bootstrapping as a means of gain enhancement requires the bootstrapped point to *accurately* follow the output of the voltage stage, because even small deviations from unity-gain in the bootstrapping mean that the increase in Q1 collector impedance, and hence the overall open-loop gain, is seriously compromised. Note that there is no requirement for R4 and R8 to be the same value, but their sum, in conjunction with the voltage conditions, will define Q1 collector current. Figure 3.31 also shows that a 47 μF bootstrap capacitor is big enough to keep the distortion flat down to 10 Hz, and going to 100 μF gives no further improvement.

We saw earlier that inverting the output emitter-follower made the distortion worse rather than better. There is, however, an interesting consequence of that inversion. We can now use what might be called DC-bootstrapping, because no capacitor is used. The collector load of Q1 is a single resistor R4, and it is now connected between the base and emitter of the output emitter-follower Q2, as shown in Figure 3.32.

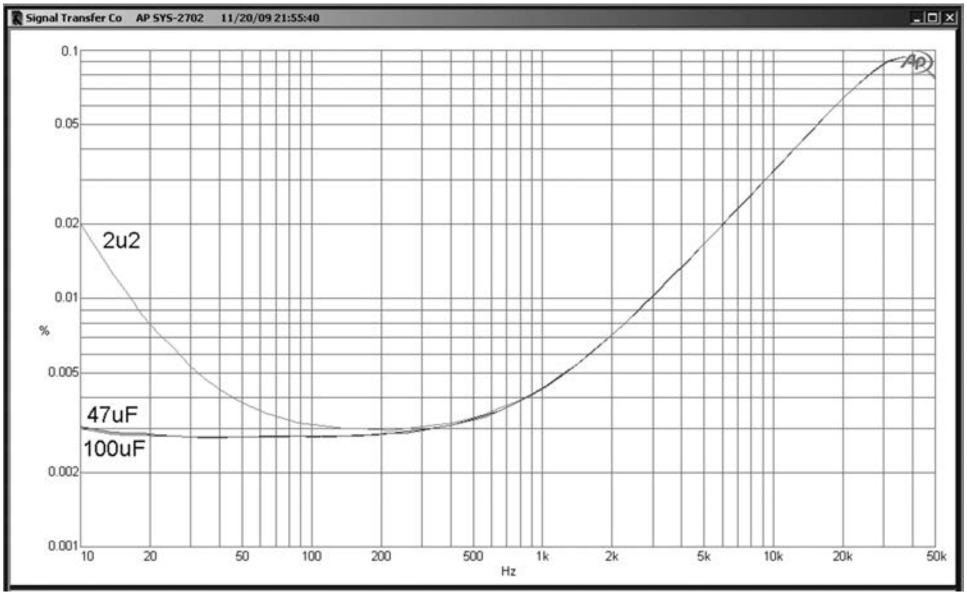


Figure 3.31: The 2µ2 trace shows the effect of using too small a bootstrap capacitor. Output 3 Vrms, no external load

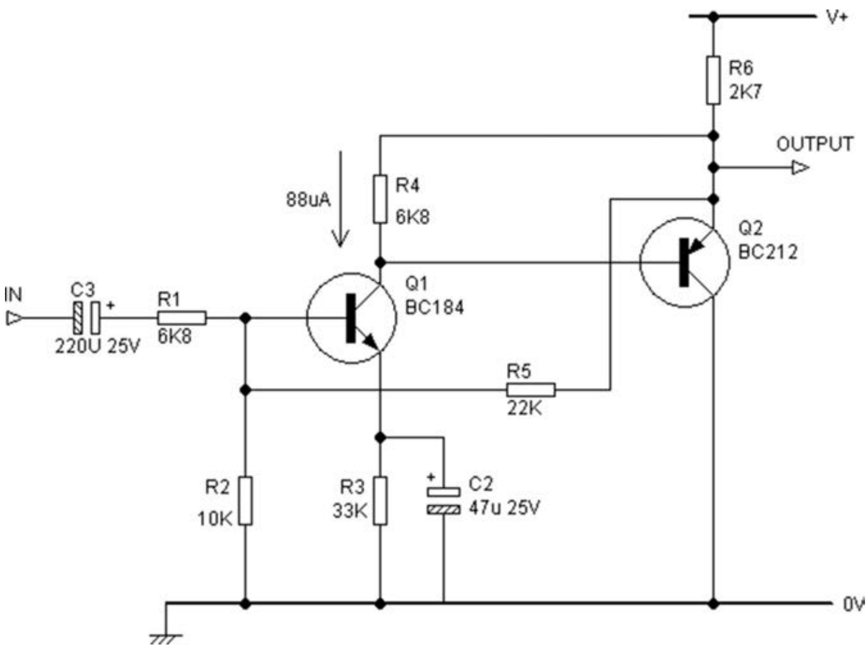


Figure 3.32: Two-transistor shunt-feedback amplifier with emitter-follower Q2 inverted and DC-bootstrapping of the Q1 load R4

TABLE 3.5 THD versus R4 and R6 values with inverted emitter-follower DC bootstrapping

Supply voltage (V)	R4 value	R6 value	THD 1 kHz (%)	THD ratio
+24	6k8	2k7	0.00694	1.62
+24	6k8	3k3	0.00705	1.65
+24	6k8	2k2	0.00686	1.61
+24	4k7	2k7	0.014	3.28
+24	10k	2k7	0.015	3.51

This technique eliminates the bootstrap capacitor and its potential problems, but unfortunately is less linear. The distortion performance is summarised in Table 3.5, where once more, the 'THD ratio' is based on the second row of Table 3.2 as the reference.

This configuration has better linearity than the inverted emitter-follower with capacitive bootstrapping, but it is still markedly worse than the original version of Figure 3.25. As Table 3.5 shows, tweaking R6 doesn't help much. We can also try changing the value of the Q1 collector load R4, as it determines the Q1 collector current, and is likely to have an important effect on circuit operation, but as the last two rows of Table 3.5 demonstrate, sadly we find that either increasing or decreasing it from 6k8 makes things a lot worse.

We have however saved two components – in particular an electrolytic capacitor has been eliminated, and these are the least reliable parts in the long-term, due to their tendency to dry out and drop in value.

Two-transistor shunt-feedback stages as summing amplifiers

Two-transistor shunt-feedback stages like those described above were used as virtual-earth summing amplifiers in mixers for many years. They were still in use in the early 1980s in mixer designs that were all-discrete to avoid the compromises in noise and distortion that came with the (affordable) opamps of the day. A two-transistor summing amplifier is simply the design of Figure 3.25, with a 22 kΩ feedback resistor, the 6k8 input resistor representing a summing resistor from a mixer channel. The junction of the two was the virtual-earth bus.

We have seen that the distortion of the two-transistor configuration is not stunning by modern standards, though it can be improved. Another important parameter for a summing amplifier is the level of the bus residual, which is the signal voltage actually existing on the virtual-earth bus. Ideally this would be zero, but in the real world it is non-zero because the summing amplifier has a finite open-loop gain. In simpler mixers this can have serious effects on inter-bus crosstalk; this is discussed in more detail in Chapter 22.

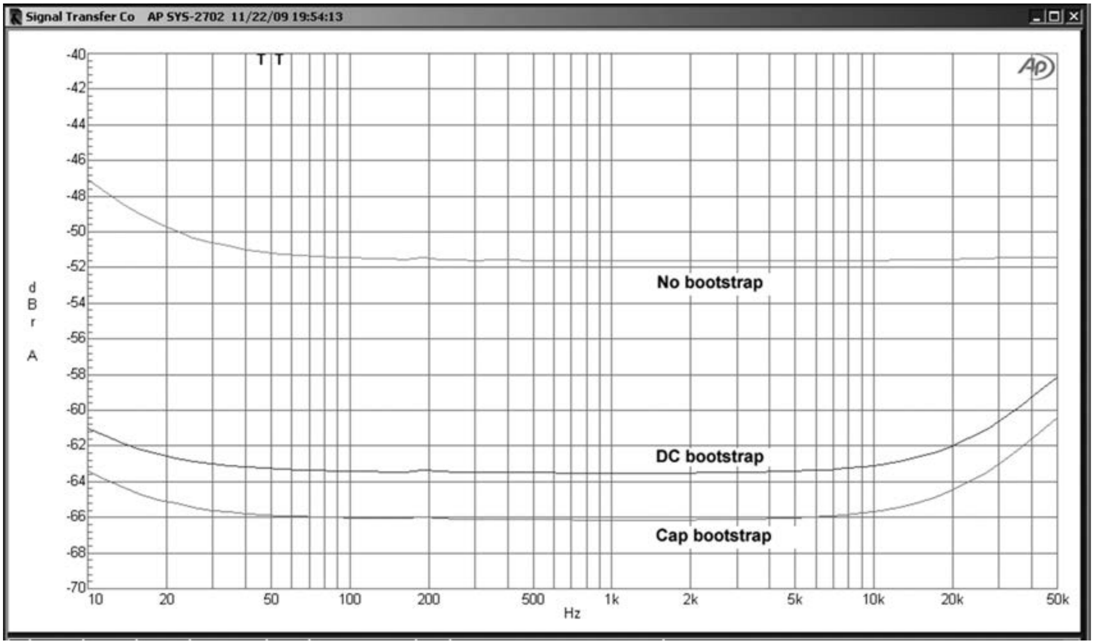


Figure 3.33: The bus residual for non-bootstrapped, capacitor-bootstrapped, and DC-bootstrapped two-transistor summing amplifiers. Measured relative to output voltage

Figure 3.33 shows the bus residual, relative to the summing amplifier output voltage, for three versions of the two-transistor configuration. With no bootstrapping, as in Figure 3.23, the bus residual is an unimpressive −52 dB. Adding conventional capacitor bootstrapping to increase the open-loop gain gives a much lower −66 dB, and DC-bootstrapping, as described in the previous section, gives us −64 dB, indicating it is slightly less effective at raising the open-loop gain. These two traces show a rise at the HF end which does not appear on the −52 dB trace; this is almost certainly due to the open-loop gain falling off at HF due to the C_{bc} of Q1. There is also a rise in bus residual at the LF end, looking the same for all three cases. I suspect increasing C2 would remove that.

Two-transistor series-feedback gain stages

The circuit in Figure 3.34 clearly has a close family relationship with the CFP emitter-follower, which is simply one of these stages configured for unity gain. The crucial difference here is that the output is separated from the input emitter, so the closed-loop gain is set by the R3-R2 divider ratio.

Only limited NFB is available, so closed-loop gains of two or three times are usually the limit.

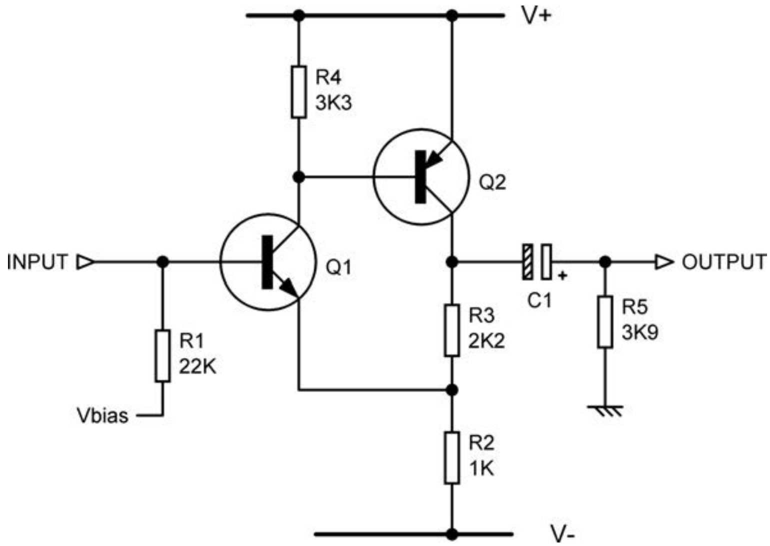


Figure 3.34: Two-transistor gain stage, series-feedback. +24 V rail. Gain once more approximately three

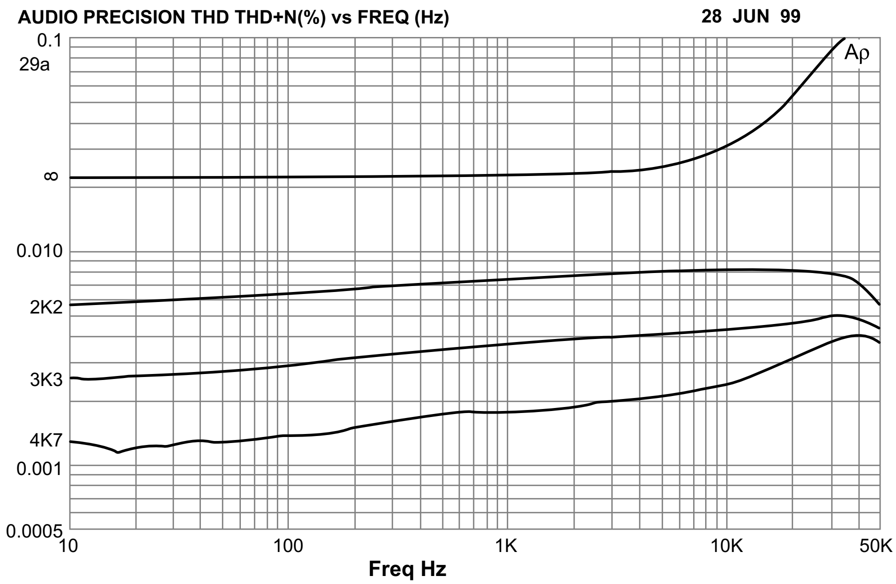


Figure 3.35: Two-transistor series-feedback stage. THD varies strongly with value of R4. +24 V rail

It is less easy to adapt this circuit to improve load-driving capability, because the feedback resistive divider must be retained. Figure 3.35 shows that distortion performance depends strongly on the value of R4, the collector load for Q1. The optimal value for linearity and noise performance is around 4k7.

Discrete opamp design

When the previously described circuits do not show enough linearity or precision for the job in hand, or a true differential input is required, it may be time to use an opamp made from discrete transistors. The DC precision will be much inferior to an IC opamp and the parts count high compared with the other discrete configurations we have looked at, but the noise and distortion performance can both be very good indeed. A good example of discrete opamp usage is a preamplifier I designed a while back, [13] though it must be said the opamps there could definitely be improved by following the suggestions in this section.

A typical, though non-optimal, discrete opamp is shown in Figure 3.36. The long-tail pair first stage Q2, Q3 subtracts the input and feedback voltages. It is a transconductance amplifier, i.e. it turns a differential voltage input into a current output, which flows into the second amplifier stage Q4, Q5. This is a transadmittance amplifier; current in becomes voltage out. Its input is at low impedance (a sort of virtual-earth) because of local NFB through dominant-pole capacitor C3, and so is well-adapted to accepting the output current from the first stage. The third stage is a Class-A emitter follower with current source, which isolates the second stage collector from external loads. The higher the quiescent current in this stage (here 6 mA as before) the lower the load impedance that can be driven symmetrically. The configuration is basically that of a power amplifier, with a simplified small-scale output

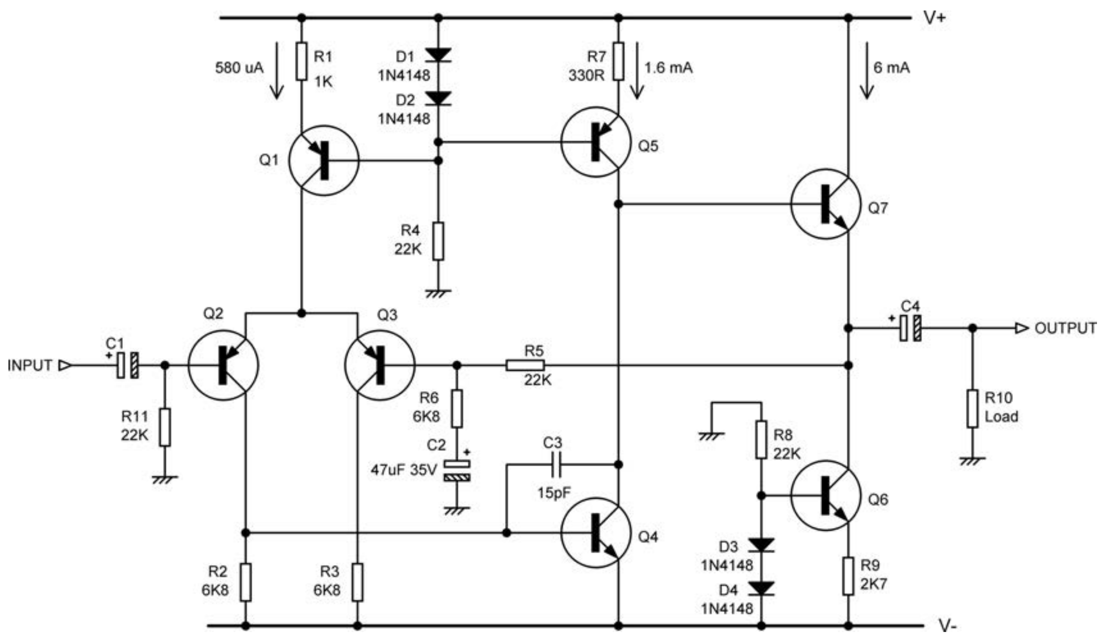


Figure 3.36: A typical discrete opamp circuit, with current-source output

stage; a very great deal more information than there is space for here can be found in my power amplifier book [14].

The dominant-pole capacitor C3 is shown here as 15 pF, which is often large enough for stable small-signal operation. In a power amplifier it would probably be 100 pF because of the greater phase shifts in a full-scale output stage with power transistors. The closed-loop gain is here set to about 4 times (+12.5 dB) by R5, R6.

The opamp in Figure 3.36 can be much improved by addressing the major causes of distortion in the circuit. These are:

- Input stage: The current/voltage gain (transconductance) curve of the input stage peaks broadly at the centre of its characteristic, where the collector currents of Q2, Q3 are equal, and this is its most linear area. The distortion is a relatively small amount of third-harmonic. If the input pair is operated away from this point there is a rapid rise in second-harmonic distortion, which quickly swamps the third harmonic.
- Second stage: The transistor Q4 in conjunction with C3 converts current to voltage. It has the full output swing on its collector, and so I call it the VAS (voltage amplifier stage). It has an internal collector-base capacitance C_{bc} which varies with the collector-base voltage, (V_{cb}) and so creates second-harmonic distortion at HF. Distortion at LF is at a much lower level and is due to Early effect in Q4.
- Output stage: The distortion of the Class-A output stage is negligible compared with that of the first two stages, providing it is not excessively loaded.

Discrete opamp design: the input stage

If the input pair is operated away from its balanced condition, there is a rapid (12 dB/octave) rise in second-harmonic THD with frequency due to the roll-off of open-loop gain by C3. As a result the value of R2 in Figure 3.36 is crucial. Global NFB establishes the V_{be} of Q4 (0.6 V) across R2, so it needs to be approximately half the value of R1, which has 0.6 V across it to set the input stage tail current. The value of R3 has almost no effect on the collector-current balance but is chosen to equalise the power dissipation in the two input transistors and so reduce DC drift. The critical importance of the value of R2 is demonstrated in Figure 3.37, where even small errors in R2 greatly increase the distortion. The collector currents of Q2, Q3 need to be matched to about 1% for minimum distortion.

Figure 3.36 has a small Miller capacitance C3 of 15 pF, and the large amount of NFB consequently available means that simply optimising R2 gives results almost indistinguishable from the testgear residual of an Audio Precision System-1. The circuit with R2 corrected is shown in Figure 3.38. An additional unrelated improvement shown is the inversion of the output stage so the biasing diodes D1, D2 can be shared, and the component count reduced.

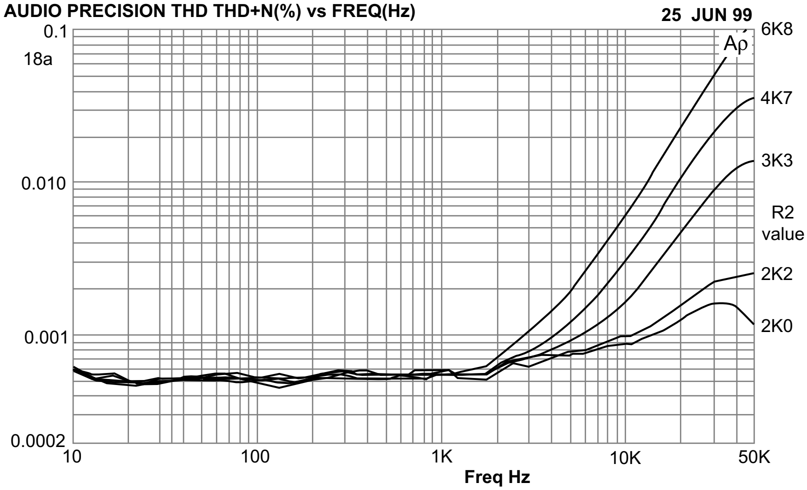
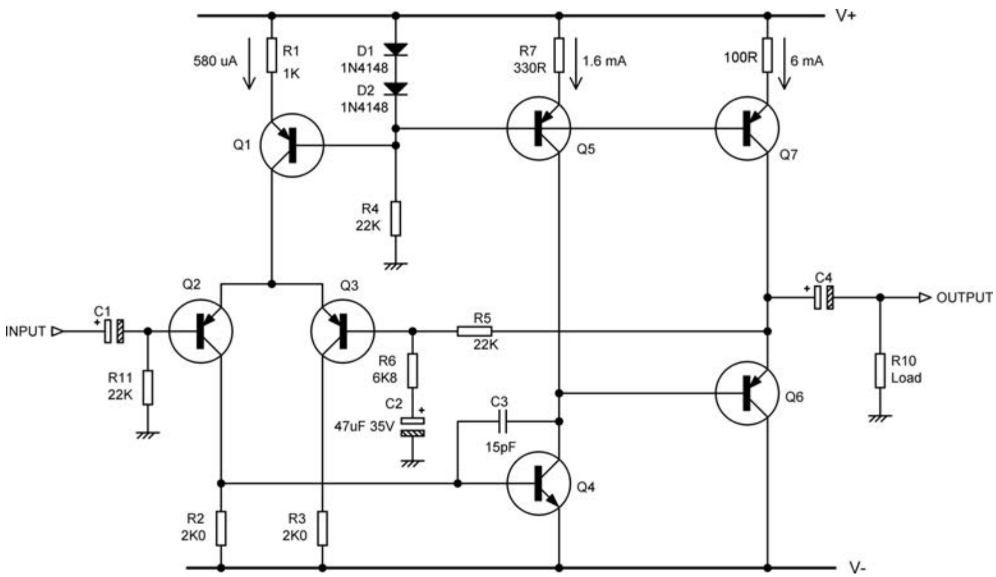


Figure 3.37: Reduction in high-freq distortion of Figure 3.36 as input stage approaches balance with R2 = 2 kΩ



Figiure 3.38: Improved discrete opamp, with input stage balanced, and output stage inverted to share the biasing diodes D1, D2. +20 dBu output, ±20 V rails

The good linearity obtained with the simple design of Figure 3.38 is only achieved because the closed-loop gain is moderate at four times, and the open-loop gain is high. In more demanding applications (including the front-end of power amplifiers) the closed-loop gain will be higher (often 23 times) and the open-loop gain lower due to a typical Miller capacitance of 100 pF. This places much greater demands on the basic linearity of the circuit, and further improvements are valuable:

- The tail current of the input pair Q2, Q3 is increased from 580 μA to 4 mA, increasing the transconductance of the two transistors by roughly eight times. The transconductance is then reduced back to its original value by inserting emitter degeneration resistors. Since the transconductance is now controlled much more by fixed resistors rather than the transistors, this greatly linearises the input stage.
- The collector resistors are replaced with a current-mirror that forces the collector currents of the input pair Q2, Q3 into accurate equality. This is more dependable than optimising R2, and also doubles the open-loop gain as the collector currents of both input transistors are now put to use.

These two enhancements give us the circuit shown in Figure 3.39. The input pair emitter degeneration resistors are R2, R3, and the current-mirror is Q4, Q5, which is itself degenerated by R4, R5 for greater accuracy. The distortion performance is the upper trace

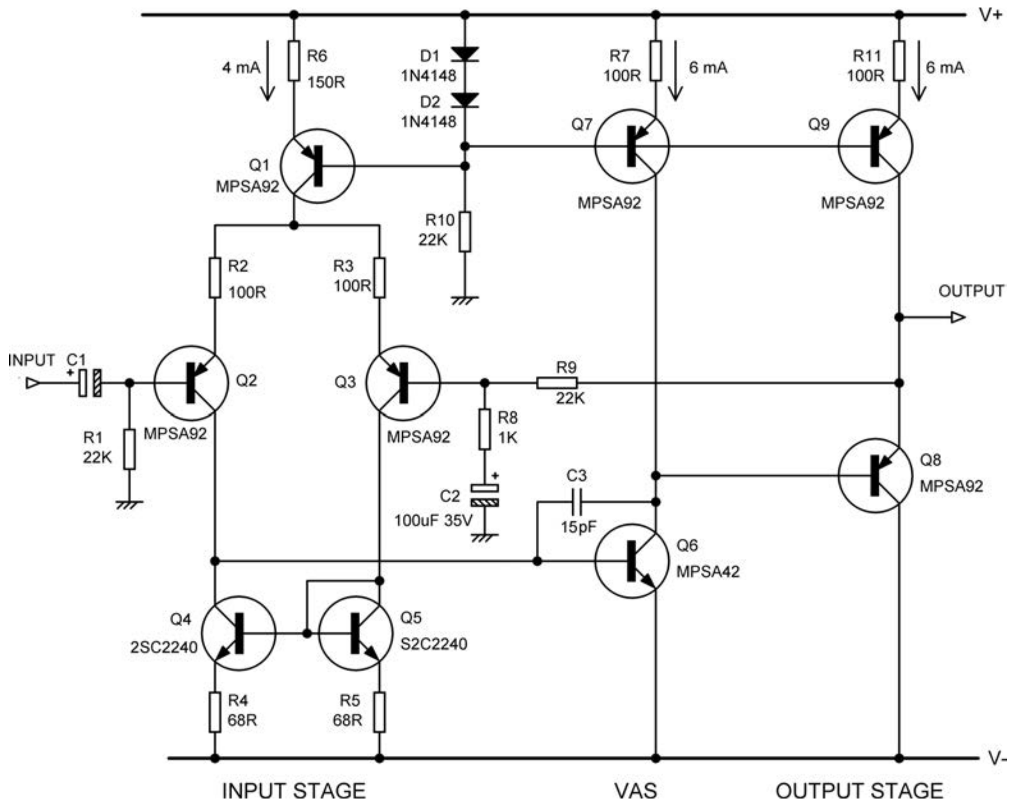


Figure 3.39: The improved discrete opamp with input pair degeneration R2, R3 and current-mirror Q4, Q5 added

in Figure 3.41; this looks pretty poor compared with the lower trace in Figure 3.37, but as explained above our discrete opamp is now working under much more demanding conditions of high closed-loop gain and low open-loop gain, much reducing the NFB factor. Thanks to our efforts on the first stage, the distortion seen is coming entirely from the second stage Q6, so we had better sort it out.

Discrete opamp design: the second stage

As noted above, the dominant cause of second-stage (VAS) distortion is the non-linear C_{bc} of the VAS transistor, now labelled Q6. The local feedback around it is therefore non-linear and second-harmonic distortion is generated. A simple but extremely effective cure for this is to add an emitter-follower Q10, inside the local feedback loop of C3, giving us Figure 3.40. The non-linear local feedback through Q6, C_{bc}, is now harmlessly absorbed by the low impedance at the emitter of Q10, and the linear component C3 alone controls the VAS behaviour. The result is the lower trace in Figure 3.41, which is indistinguishable from the output of the state-of-art Audio Precision SYS-2702. Similar results can be obtained by instead cascoding Q6, but more parts are required and the linearity is in general not quite so good.

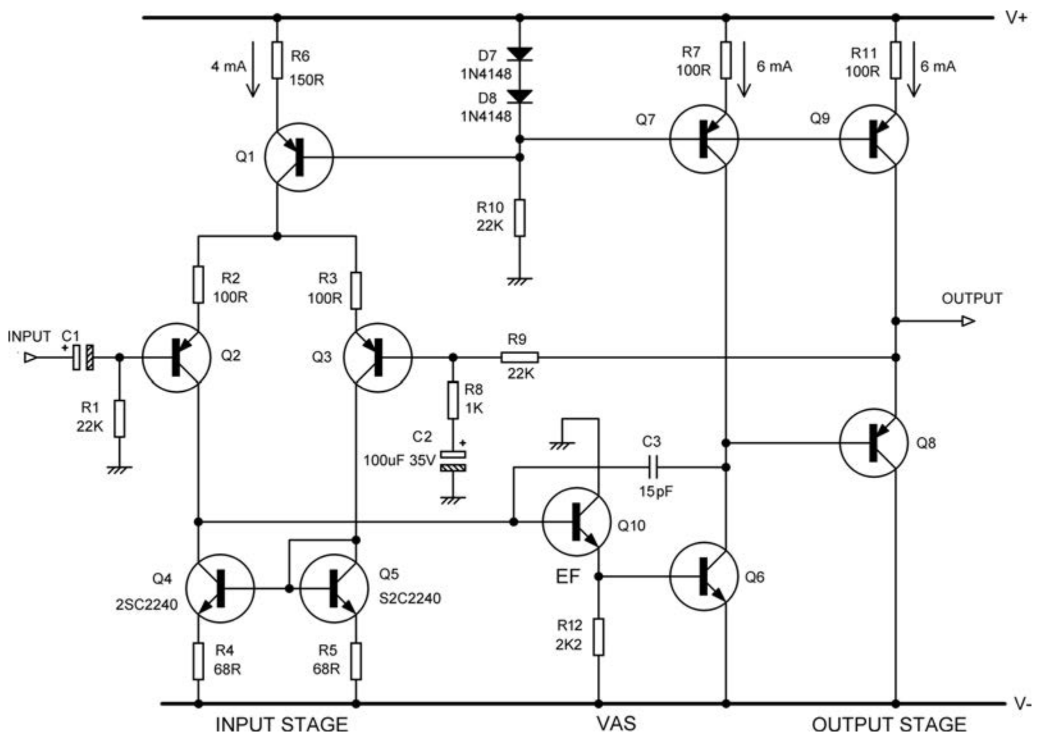


Figure 3.40: The improved discrete opamp of Figure 3.39 with emitter-follower Q10 added inside the VAS local feedback loop

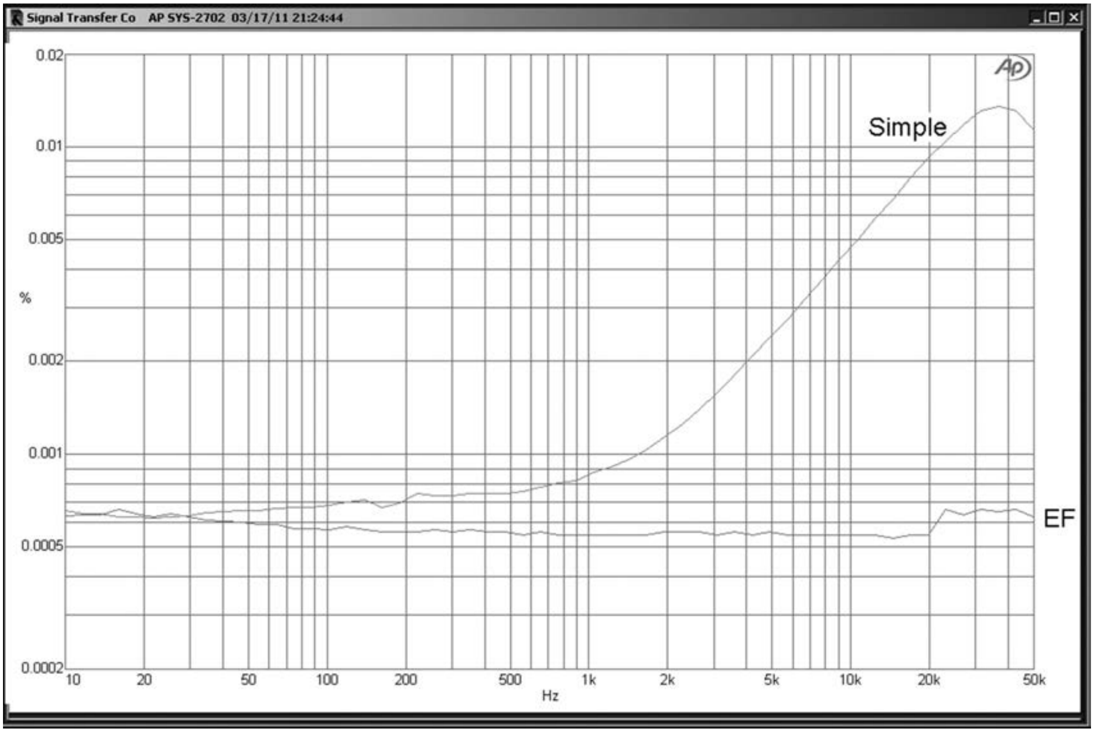


Figure 3.41: THD of the improved opamp circuit, with and without emitter-follower added to the VAS. +20 dBu output, ±20 V rails

In both Figures 3.39 and 3.40 the VAS current has been increased from 1.6 mA to 6 mA. This change has very little effect on the distortion performance, but much increases the positive slew-rate, marking it more nearly equal to the negative slew-rate. The transistors are all MPSA42/92 high-voltage low-beta types (with the exception of the current-mirror) to underline the fact that high-beta devices are not required for low distortion. Do not use MPSA42 in the current-mirror as it has an unusually high minimum V_{ce} for proper operation.

Discrete opamp design: the output stage

Figure 3.41 shows there is no need to worry about output stage distortion with moderate loading, as it is very low with even moderate levels of NFB. If the loading is so heavy that linearity deteriorates, the options are to increase the output stage standing current or to adopt a push-pull emitter-follower output stage, as described earlier in this chapter. The push-pull method is more efficient and will give better linearity.

For still heavier loads, a Class-AB output stage employing two complementary small-signal transistors can be used. The operation of such a stage is rather different from that of a power amplifier output stage, with crossover distortion and critical biasing being less of a problem. See Chapter 20 on headphone amplifiers for more on this.

High input impedance bipolar stages

Bipolar transistors are commonly thought of as low-impedance devices, but in fact they can be used to create amplifiers with extremely high input impedances. FETs are of course the obvious choice for high input impedance amplifiers, but their lack of transconductance is a drawback. This section was inspired by an article by T. D. Towers in 1968 that is still well worth reading [15]. The circuits that follow deliberately use high-voltage low-beta transistors as higher than normal rails are one of the reasons for using discrete circuitry. ±24 V rails are used. The impedances given were measured and also checked with SPICE.

Figure 3.42a shows a simple emitter-follower biased by R_{bias}, which at 100 kΩ is about as high as you would want it to be, because of the voltage drop due to the base current. That

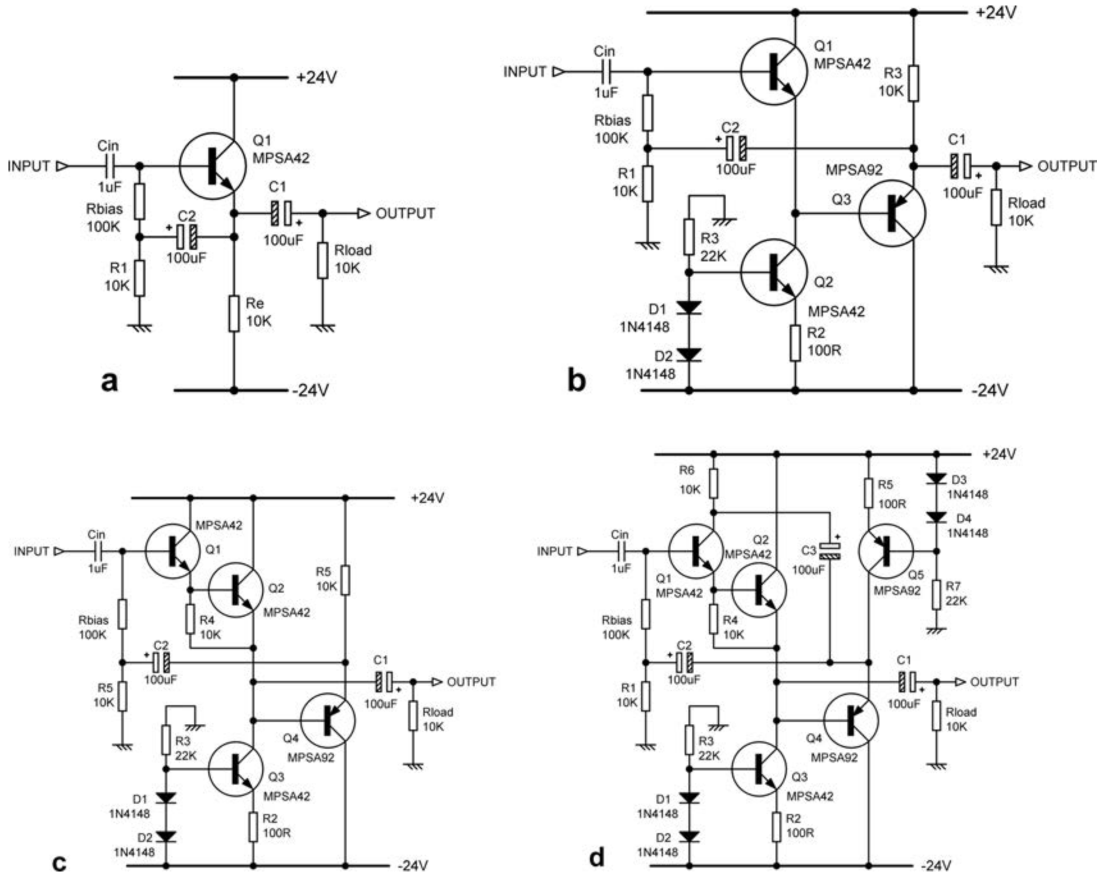


Figure 3.42: High-impedance input stages: a) Simple emitter-follower with bootstrapped biasing; 500 kΩ, b) emitter-follower with current-source and bootstrap driver stage; 6.1 MΩ, c) Darlington with current-source and bootstrap driver stage; 21 MΩ, d) Darlington with current-source and Q1 collector bootstrapped; 61 MΩ

limits the input impedance to 100 kΩ of course. The First Principle of high-impedance design is to bootstrap the bias resistor as in Figure 3.42a, which raises the input impedance to 500 kΩ. You may be wondering: why not more? One reason is that Q1 is only a simple emitter-follower and its voltage gain is distinctly less than one, limiting the efficacy of the bootstrapping.

The Second Principle of high-impedance design with simple discrete stages is that any loading on the output reduces the input impedance because, as noted earlier:

$$R_{in} = \beta(R_e \| R_{load}) \qquad \text{(Equation 3.3)}$$

and the bootstrap capacitor C2 is driving the load of R1 even if there is no external loading. Adding an external 10 kΩ R_{load} to represent the following stage reduces the input impedance to 350 kΩ.

Having noted that both R_e and R_{load} pull down the input impedance, we will take steps to increase their effective values. Figure 3.42b shows R_e made very high by replacing the emitter resistor with a current source Q2. R_{load} is made high by adding simple emitter-follower Q3 to drive the bootstrap and the output. This gives a 6.1 MΩ input impedance even when driving an external 10 kΩ load output.

A further increase in input impedance can be obtained by increasing the β term in Equation 3.3, by using a Darlington configuration, where one emitter-follower feeds another, as in Figure 3.42c. Q1 must have a reasonable collector current to operate at a good β; more than the base current of Q2. An emitter resistor to the negative rail would increase the loading and defeat the object, so R4 is used, with its lower end bootstrapped from the emitter of Q2. This technique is also used in power amplifier circuitry [16]. This gives an input impedance of 21 MΩ with an external 10 kΩ load. The output is now taken from Q2 emitter once more.

To significantly further raise the input impedance, we need to take on board the Third Principle of high-impedance design; bootstrap the input transistor collector, as in Figure 3.42d. The collector resistance r_c of Q1 and the base-collector capacitance c_{bc} are both effectively in parallel with the input; the former can also be regarded as Early Effect causing I_b to vary. Their effects are reduced by bootstrapping Q1 collector using R6 and C3 (note that collector bootstrapping cannot be used with a single-transistor stage [17]). The emitter follower Q4 also has its emitter resistor replaced by current source Q5 to make its gain nearer one for more effective bootstrapping and help with driving R1 and R6. The output point has also been shifted back to the second emitter-follower. The result is an input impedance of 60 MΩ, or 50 MΩ with an external 10 kΩ load.

We have achieved this using only five transistors, so the BJT is clearly not an inherently low-impedance device. There are many more technical possibilities if you need a really astronomical input impedance; the record in 1968 appears to have been no less than 20,000 MΩ [18].

The circuits in Figure 3.42 are not optimised for linearity. They all give substantially more distortion when fed from a high source impedance such as 1 MΩ, as the base currents drawn are not linear.

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CHAPTER 4

Opamps and their properties

Introduction

Audio design has for many years relied on a very small number of opamp types; the TL072 and the 5532 dominated the audio small-signal scene for many years. The TL072, with its JFET inputs, was used wherever its negligible input bias currents and low cost were important. For a long time the 5534/5532 was much more expensive than the TL072, so the latter was used wherever feasible in an audio system, despite its inferior noise, distortion, and load-driving capabilities. The 5534 was reserved for critical parts of the circuitry. Although it took many years, the price of the 5534 is now down to the point where you need a very good reason to choose any other type of opamp for audio work.

The TL072 and the 5532 are dual opamps; the single equivalents are TL071 and 5534. Dual opamps are used almost universally, as the package containing two is usually cheaper than the package containing one, simply because it is more popular.

There are however other opamps, some of which can be useful, and a selected range is covered here.

A very brief history of opamps

The opamp is today thought of as quintessentially a differential amplifier, responding to the difference of the input voltages while (hopefully) ignoring any common-mode component. The history of differential amplifiers goes back to that great man Alan Blumlein, and his 1936 patent [1] for a pair of valves with their cathodes connected to ground through a common resistor. However, the first valve-based operational amplifiers, i.e. those intended to be capable of performing a mathematical operation, were in fact not differential at all, having only one input. That had to be an inverting input, of course, so you could apply negative feedback.

The first opamp to get real exposure in the UK was the Fairchild uA709, designed by the great Bob Widlar and introduced in 1965. It was a rather awkward item which required quite complicated external compensation and was devoid of output short-circuit protection. One slip of the probe and an expensive IC was gone. It was prone to latch-up with high common-mode voltages and did not like capacitive loads. I for one found all this most discouraging, and gave up on the 709 pretty quickly. If you're going to quit, do it early, I say.

The arrival of the LM741 was a considerable relief. To my mind, it was the first really practical opamp, and it was suddenly possible to build quite complex circuitry with a good chance of it being stable, doing what it should do, and not blowing up at the first shadow of an excuse. I have given some details of it in this chapter for purely historical reasons. There is also an interesting example of how to apply the LM741 appropriately in Chapter 15.

The first IC opamps opened up a huge new area of electronic applications, but after the initial enthusiasm for anything new, the audio market greeted these devices with less than enthusiasm. There were good reasons for this. The LM741 worked reliably; the snag with using it for audio was the leisurely slew rate of 0.5 V/μs, which made full output at 20 kHz impossible. For a period of at least five years, roughly from 1972 to 1977, the only way to obtain good performance in a preamp was to stick with discrete transistor Class-A circuitry, and this became recognised as a mark of high quality. The advent of the TL072 and the 5532 changed this situation completely, but there is still sometimes marketing cachet to be gained from a discrete design.

An excellent and detailed history of operational amplifiers can be found in reference [2].

Opamp properties: noise

There is no point in regurgitating manufacturer's data sheets, especially since they are readily available on the internet. Here I have simply ranked the opamps most commonly used for audio in order of voltage noise (Table 4.1).

The great divide is between JFET input opamps and BJT input opamps. The JFET opamps have more voltage noise but less current noise than bipolar input opamps, the TL072 being particularly noisy. If you want the lowest voltage noise, it has to be a bipolar input. The difference however between a modern JFET-input opamp such as the OPA2134 and the old faithful 5532 is only 4 dB; but the JFET part is a good deal more costly. The bipolar AD797 seems to be out on its own here, but it is a specialised and expensive part. The LT1028 is not suitable for audio use for reasons described in Chapter 11. The LM741 has no noise specs on its data sheets, and the 20nV/√Hz is from measurements.

TABLE 4.1 Opamps ranked by voltage noise density (typical)

Opamp	en (nV/√Hz)	in (pA/√Hz)	Input device type	Bias cancel?
LM741	20	??	BJT	No
TL072	18	0.01	FET	No
OPA604	11	0.004	FET	No
NJM4556	8	Not spec'd	BJT	No
OPA2134	8	0.003	FET	No
OP275	6	1.5	BJT+FET	No
OPA627	5.2	0.0025	FET	No
5532A	5	0.7	BJT	No
LM833	4.5	0.7	BJT	No
MC33078	4.5	0.5	BJT	No
5534A	3.5	0.4	BJT	No
OP270	3.2	0.6	BJT	No
OP27	3	0.4	BJT	Yes
LM4562	2.7	1.6	BJT	No
AD797	0.9	2	BJT	No
LT1028	0.85	1	BJT	Yes

Both voltage and current noise increase at 6 dB/octave below the 1/*f* corner frequency, which is usually around 100 Hz. The only way to minimise this effect is to choose an appropriate opamp type.

Opamps with bias cancellation circuitry are normally unsuitable for audio use due to the extra noise this creates. The amount depends on circuit impedances, and is not taken into account in Table 4.1. The general noise behaviour of opamps in circuits is dealt with in Chapter 1.

Opamp properties: slew rate

Slew rates vary more than most parameters; a range of 100:1 is shown here in Table 4.2. The slowest is the LM741, which is the only type not fast enough to give full output over the audio band. There are faster ways to handle a signal, such as current-feedback architectures, but they usually fall down on linearity. In any case, a maximum slew rate greatly in excess of what is required appears to confer no benefits whatever.

TABLE 4.2 Opamps ranked by slew rate (typical)

Opamp	V/μs
LM741	0.5
OP270	2.4
OP27	2.8
NJM4556	3
MC33078	7
LM833	7
5532A	9
LT1028	11
TL072	13
5534A	13
OPA2134	20
LM4562	20
AD797	20
OP275	22
OPA604	25
OPA627	55

The 5532 slew rate is typically ±9 V/μs. This version is internally compensated for unity-gain stability, not least because there are no spare pins for compensation when you put two opamps in an 8-pin dual package. The single-amp version, the 5534, can afford a couple of compensation pins, and so is made to be stable only for gains of 3 times or more. The basic slew rate is therefore higher at ±13 V/μs.

Compared with power-amplifier specs, which often quote 100 V/μs or more, these speeds may appear rather sluggish. In fact they are not; even ±9 V/μs is more than fast enough. Assume you are running your opamp from ±18 V rails, and that it can give a ±17 V swing on its output. For most opamps this is distinctly optimistic, but never mind. To produce a full-amplitude 20 kHz sine wave you only need 2.1 V/μs, so even in the worst case there is a safety-margin of at least four times. Such signals do not of course occur in actual use, as opposed to testing. More information on slew-limiting is given in the section on opamp distortion.

Opamp properties: common mode range

This is simply the range over which the inputs can be expected to work as proper differential inputs. It usually covers most of the range between the rail voltages, with one notable exception. The data sheet for the TL072 shows a common-mode (CM) range that looks a bit curtailed at −12 V. This bland figure hides the deadly trap this IC contains for the unwary. Most opamps, when they hit their CM limits, simply show some sort of clipping. The TL072, however, when it hits its negative limit, promptly inverts its phase, so your circuit either latches up, or shows nightmare clipping behaviour with the output bouncing between the two supply rails. The positive CM limit is in contrast trouble-free. This behaviour can be especially troublesome when TL072s are used in high-pass Sallen and Key filters.

Opamp properties: input offset voltage

A perfect opamp would have its output at 0 V when the two inputs were exactly at the same voltage. Real opamps are not perfect and a small voltage difference – usually a few milliVolts – is required to zero the output. These voltages are large enough to cause switches to click and pots to rustle, and DC blocking is often required to keep them in their place.

The typical offset voltage for the 5532A is ±0.5 mV typical, ±4 mV maximum at 25 °C; the 5534A has the same typical spec but a lower maximum at ±2 mV. The input offset voltage of the new LM4562 is only ±0.1 mV typical, ±4 mV maximum at 25 °C.

Opamp properties: bias current

Bipolar input opamps not only have larger noise currents than their JFET equivalents, they also have much larger bias currents. These are the base currents taken by the input transistors. This current is much larger than the input offset current, which is the difference between the bias current for the two inputs. For example, the 5532A has a typical bias current of 200 nA, compared with a much smaller an input offset current of 10 nA. The LM4562 has a lower bias current of 10 nA typical, 72 nA maximum. In the case of the 5532/4 the bias current flows into the input pins as the input transistors are NPN.

Bias currents are a considerable nuisance, when they flow through variable resistors they make them noisy when moved. They will also cause significant DC offsets when they flow through high-value resistors.

It is often recommended that the effect of bias currents can be cancelled out by making the resistance seen by each opamp input equal. Figure 4.1a shows a shunt-feedback stage with a 22 kΩ feedback resistor. When 200 nA flows through this it will generate a DC offset of 4.4 mV, which is rather more than we would expect from the input offset voltage error.

If an extra resistance R_{compen}, of the same value as the feedback resistor, is inserted into the non-inverting input circuit then the offset will be cancelled. This strategy works well and is done almost automatically by many designers. However, there is a snag. The resistance R_{compen} generates extra Johnson noise, and to prevent this it is necessary to shunt the

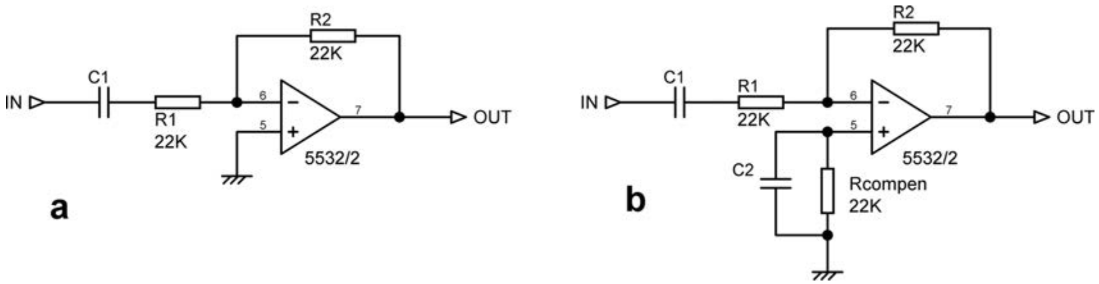

Figure 4.1: Compensating for bias current errors in a shunt-feedback stage. The compensating resistor must be bypassed by a capacitor C2 to prevent it adding Johnson noise to the stage

resistance with a capacitor, as in Figure 4.1b. This extra component costs money and takes up PCB space, so it is questionable if this technique is actually very useful for audio work. It is usually more economical to allow offsets to accumulate in a chain of opamps, and then remove the DC voltage with a single output blocking capacitor. This assumes that there are no stages with a large DC gain, and that the offsets are not large enough to significantly reduce the available voltage swing. Care must also be taken if controls are involved, because even a small DC voltage across a potentiometer will cause it to become crackly, especially as it wears.

FET input opamps have very low bias current at room temperature; however it doubles for every 10 degree Centigrade rise. This is pretty unlikely to cause trouble in most audio applications, but a combination of high internal temperatures and high-value pots could lead to some unexpected crackling noises.

Opamp properties: cost

While it may not appear on the datasheet, the price of an opamp is obviously a major factor in deciding whether or not to use it. Table 4.3 was derived from the averaged prices for 1+ and 25+ quantities across a number of UK distributors. In September 2009 (at the time of writing the First Edition) the cheapest popular opamps were the TL072 and the 5532, and these happened to come out at exactly the same price for 25+, so their price is taken as unity and used as the basis for the price ratios given.

Table 4.3 was compiled using prices for DIL packaging and the cheapest variant of each type. Price is per package and not per opamp section. It is obviously only a rough guide. Purchasing in large quantities or in different countries may change the rankings somewhat (even going from 1+ to 25+ causes some changes) but the basic look of things will not alter too much. One thing is obvious – the 5532 is one of the great opamp bargains of all time.

TABLE 4.3 Opamps ranked by price (2009) relative to 5532 and TL072

Type	Format	Price ratio 1+	Type	Price ratio 25+
LM833	Dual	1.45	5532	1.00
5532	Dual	1.64	TL072	1.00
MC33078	Dual	1.97	LM833	1.12
TL072	Dual	2.45	MC33078	1.27
OPA604	Single	5.09	TL052	2.55
OPA2134PA	Dual	5.55	OP275GP	3.42
TL052	Dual	5.76	OPA2134PA	4.45
OP275GP	Dual	7.18	OPA604	5.03
OP27	Single	8.67	OP27	6.76
LM4562	Dual	12.45	LM4562	9.06
AD797	Single	25.73	AD797	13.09
OP270	Dual	29.85	LT1028	17.88
LT1028	Single	30.00	OPA270	24.42
OPA627	Single	51.91	OPA627	48.42

Opamp properties: distortion

Relatively few discussions of opamp behaviour deal with non-linear distortion, perhaps because it is a complex business. Opamp 'accuracy' is closely related, but the term is often applied only to DC operation. Accuracy here is often specified in terms of bits, so '20-bit accuracy' means errors not exceeding one part in 2 to the 20, which is −120 dB or 0.0001%. Audio signal distortion is of course a dynamic phenomenon, very sensitive to frequency, and DC specs are of no use at all in estimating it.

Distortion is always expressed as a ratio, and can be quoted as a percentage, as number of decibels, or in parts per million. With the rise of digital processing, treating distortion as the quantization error arising from the use of a given number of bits has become more popular. Figure 4.2 hopefully provides a way of keeping perspective when dealing with these different metrics. There are several different causes of distortion in opamps. We will now examine them.

Opamp internal distortion

This is what might be called the basic distortion produced by the opamp you have selected. Even if you scrupulously avoid clipping, slew-limiting, and common-mode issues, opamps are not distortion free, though some types such as the 5532 and the LM4562 have very low levels. If distortion appears when the opamp is run with shunt feedback, to prevent

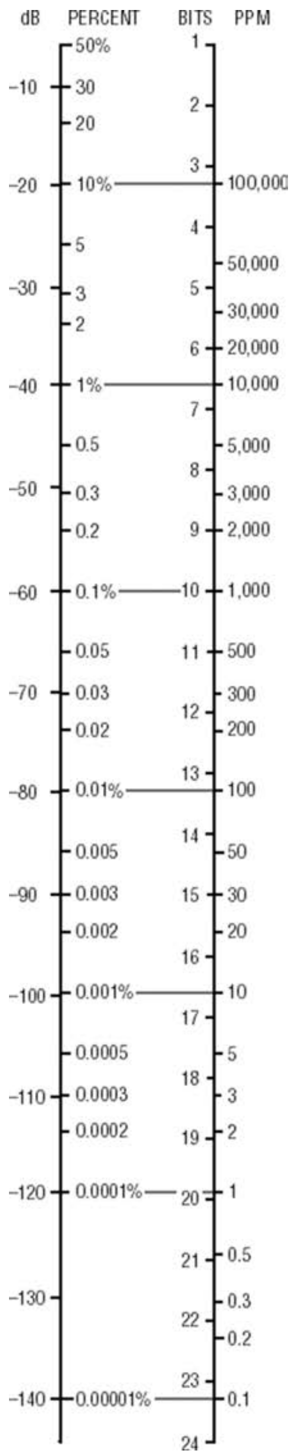

Figure 4.2: The relation between different ways of quoting THD - decibels, percentages, bit accuracy and parts per million

common-mode voltages on the inputs, and with very light output loading, then it is probably wholly internal and there is nothing to be done about it except pick a better opamp.

If the distortion is higher than expected, the cause may be internal instability provoked by putting a capacitative load directly on the output, or neglecting the supply decoupling. The classic example of the latter effect is the 5532, which shows high distortion if there is not a capacitor across the supply rails close to the package; 100 nF is usually adequate. No actual HF oscillation is visible on the output with a general-purpose oscilloscope, so the problem may be instability in one of the intermediate gain stages.

Slew rate limiting distortion

While this is essentially an overload condition, it is wholly the designer's responsibility. If users whack up the gain until the signal is within a hair of clipping, they should still be able to assume that slew-limiting will never occur, even with aggressive material full of high frequencies.

Arranging this is not too much of a problem. If the rails are set at the usual maximum voltage, i.e. ±18 V, then the maximum possible signal amplitude is 12.7 Vrms, ignoring the saturation voltages of the output stage. To reproduce this level cleanly at 20 kHz requires a minimum slew rate of only 2.3 V/μs. Most opamps can do much better than this, though with the OP27 (2.8 V/μs) you are sailing rather close to the wind. The old LM741 looks as though it would be quite unusable, as its very limited 0.5 V/μs slew rate allows a full output swing only up to 4.4 kHz.

Horrific as it may now appear, audio paths full of LM741s were quite common in the early 1970s. Entire mixers were built with no other active devices, and what complaints there were tended to be about noise rather than distortion. The reason for this is that full-level signals at 20 kHz simply do not occur in reality; the energy at the HF end of the audio spectrum is well-known to be much lower than that at the bass end.

This assumes that slew-limiting has an abrupt onset as level increases, rather like clipping. This is in general the case. As the input frequency rises and an opamp gets closer to slew-limiting, the input stage is working harder to supply the demands of the compensation capacitance. There is an absolute limit to the amount of current this stage can supply, and when you hit it the distortion shoots up, much as it does when you hit the supply rails and induce voltage clipping. Before you reach this point, the linearity may be degraded, but usually only slightly until you get close to the limit. It is not normally necessary to keep big margins of safety when dealing with slew-limiting. If you are employing the Usual Suspects in the audio opamp world – the TL072, the 5532 and the LM4562, with maximal slew rates of 13, 9 and 20 V/μs respectively, you are most unlikely to suffer any slew rate non-linearity.

Distortion due to loading

Output stage distortion is always worse with heavy output loading because the increased currents flowing exacerbate the gain changes in the Class-B output stage. These output stages are not individually trimmed for optimal quiescent conditions (as are audio power amplifiers) and so the crossover distortion produced by opamps tends to be both higher and can be more variable between different specimens of the same chip. Distortion increases with loading in different ways for different opamps. It may rise only at the high-frequency end, (e.g. the OP2277) or there may be a general rise at all frequencies. Often both effects occur, as in the TL072.

The lowest load that a given opamp can be allowed to drive is an important design decision. It will typically be a compromise between the distortion performance required and opposing factors such as number of opamps in the circuit, cost of load-capable opamps, and so on. It even affects noise performance, for the lower the load resistance an amplifier can drive, the lower the resistance values in the negative feedback can be, and hence the lower the Johnson noise they generate. There are limits to what can be done in noise-reduction by this method, because Johnson noise is proportional to the square-root of circuit resistance, and so improves only slowly as opamp loading is increased.

Thermal distortion

Thermal distortion is that caused by cyclic variation of the properties of the amplifier components due to the periodic release of heat in the output stage. The result is a rapid rise in distortion at low frequencies, which gets worse as the loading becomes heavier.

Those who have read my work on audio power amplifiers will be aware that I am highly sceptical – in fact totally sceptical – about the existence of thermal distortion in amplifiers built from discrete components [3]. The power devices are too massive to experience per-cycle parameter variations, and there is no direct thermal path from the output stage to the input devices. There is no rise, rapid or otherwise, in distortion at low frequencies in a properly designed discrete power amplifier.

The situation is quite different in opamps, where the output transistors have much less thermal inertia and are also on the same substrate as the input devices. Nonetheless, opamps do not normally suffer from thermal distortion; there is generally no rise in low frequency distortion, even with heavy output loading. Integrated-circuit power amplifiers are another matter, and the much greater amounts of heat liberated on the substrate do appear to cause serious thermal distortion, rising at 12 dB/octave below 50 Hz. I have never seen anything resembling this in any normal opamp.

Common-mode distortion

This is the general term for extra distortion that appears when there is a large signal voltage on both the opamp inputs. The voltage difference between these two inputs will be very small, assuming the opamp is in its linear region, but the common-mode (CM) voltage can be a large proportion of the available swing between the rails.

It appears to be by far the least understood mechanism, and gets little or no attention in opamp textbooks, but it is actually one of the most important influences on opamp distortion. It is simple to separate this effect from the basic forward-path distortion by comparing THD performance in series and shunt feedback modes; this should be done at the same noise gain. The distortion is usually a good deal lower for the shunt feedback case where there is no common mode voltage. Bipolar and JFET input opamps show different behaviour and they are treated separately below.

Bipolar input opamps

Figure 4.4 shows the distortion from a 5532 working in shunt mode with low-value resistors of 1 kΩ and 2k2 setting a gain of 2.2 times, at an output level of 5 Vrms. This is the circuit of Figure 4.3a with R_s set to zero; there is no CM voltage. The distortion is well below 0.0005% up to 20 kHz; this underlines what a superlative bargain the 5532 is.

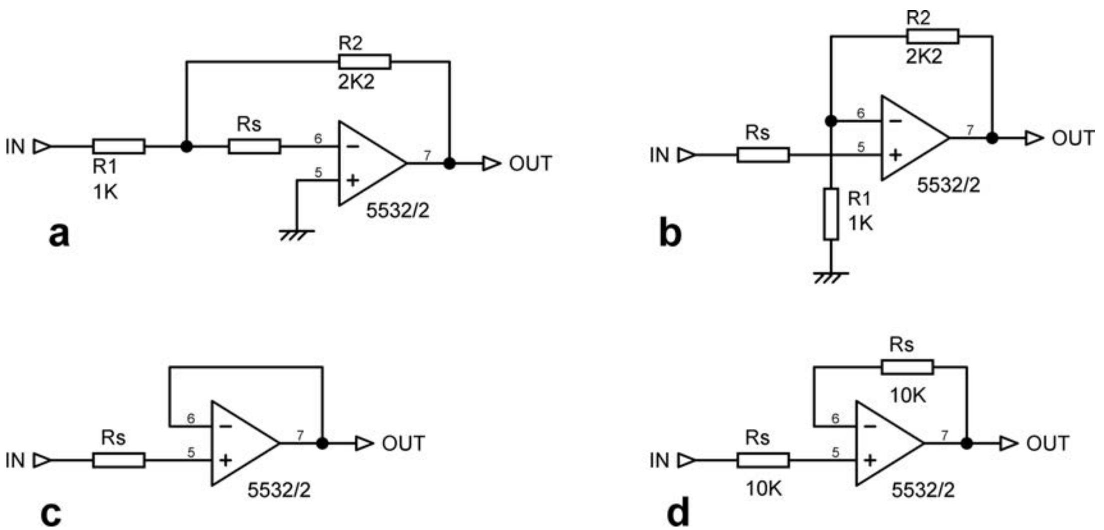

Figure 4.3: Opamp test circuits with added source resistance Rs: a) shunt, b) series, c) voltage-follower, d) voltage-follower with cancellation resistor in feedback path

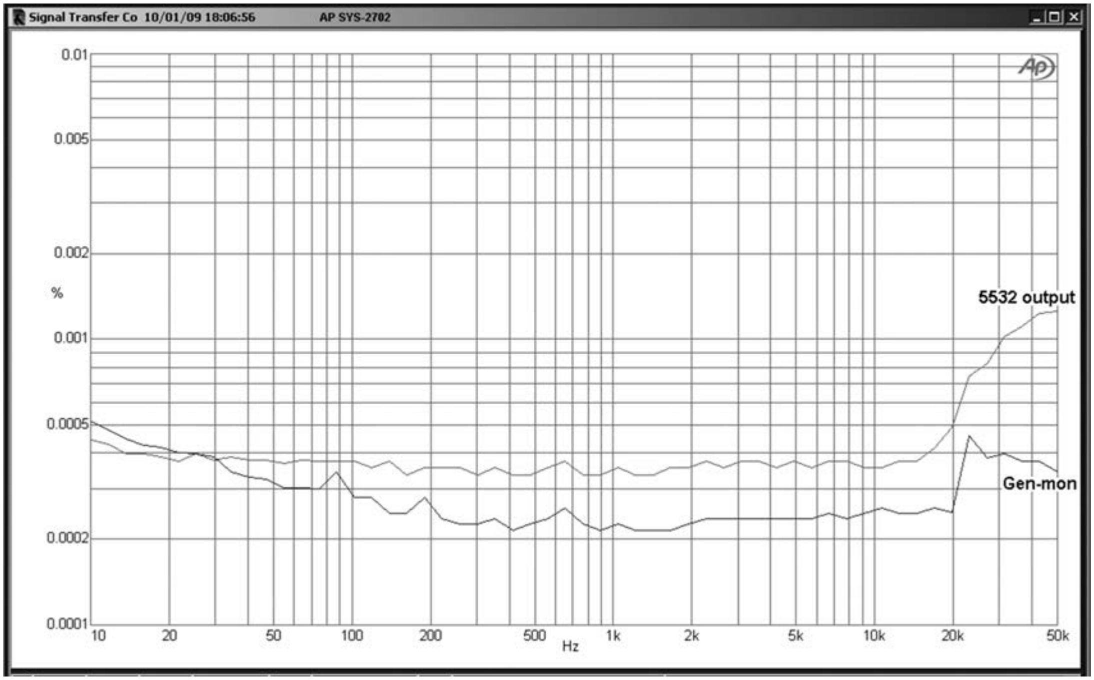

Figure 4.4: 5532 distortion in a shunt feedback circuit at 5 Vrms out. This shows the AP SYS-2702 output (lower trace) and the opamp output (upper trace). Supply ±18 V

Figure 4.5 shows the same situation but with the output increased to 10 Vrms (the clipping level on ±18 V rails is about 12 Vrms) and there is now significant distortion above 10 kHz, though it only exceeds 0.001% at 18 kHz.

This remains the case when Rs in Figure 4.3a is increased to 10 kΩ and 47 kΩ – the noise floor is higher but there is no real change in the distortion behaviour. The significance of this will be seen in a moment.

We will now connect the 5532 in the series feedback configuration, as in Figure 4.3b; note that the stage gain is greater at 3.2 times but the opamp is working at the same noise gain. The CM voltage is 3.1 Vrms. With a 10 Vrms output we can see in Figure 4.6 that even with no added source resistance the distortion starts to rise from 2 kHz, though it does not exceed 0.001% until 12 kHz. But when we add some source resistance Rs, the picture is radically worse, with serious mid-band distortion rising at 6 dB/octave, and roughly proportional to the amount of resistance added. We will note it is 0.0085% at 10 kHz with R_s = 47 kΩ.

The worst case for CM distortion is the voltage-follower configuration, as in Figure 4.3c, where the CM voltage is equal to the output voltage. Figure 4.7 shows that even with a

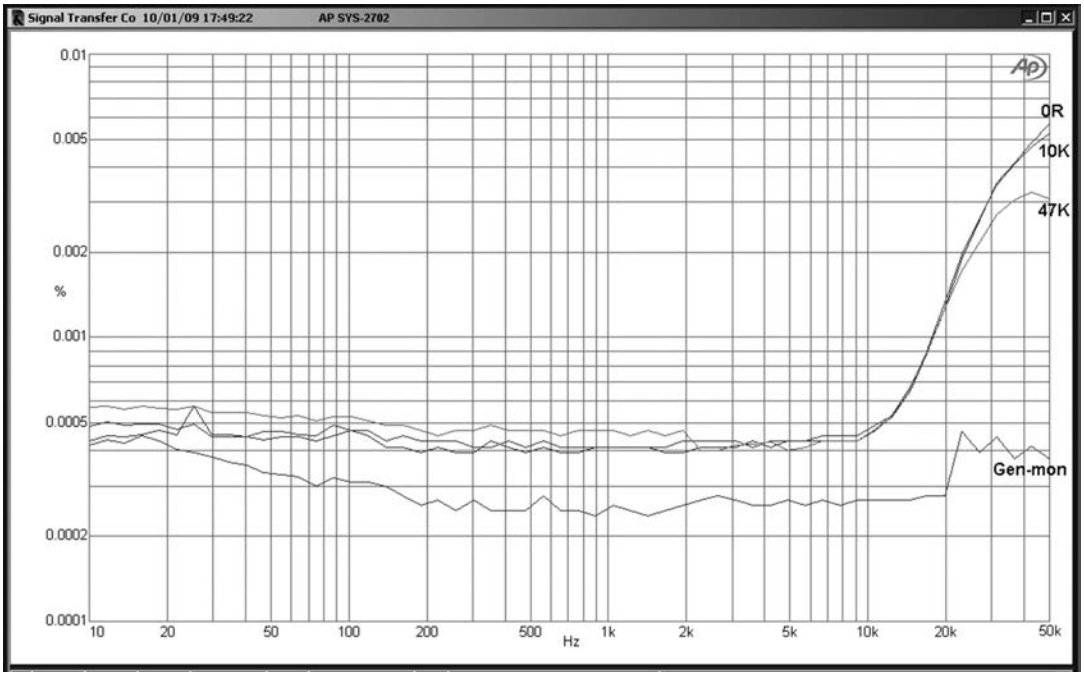

Figure 4.5: 5532 distortion in the shunt feedback circuit of Figure 4.3b. Adding extra resistances of 10 kΩ and 47 kΩ in series with the inverting input does not degrade the distortion at all, but does bring up the noise floor a bit. Test level 10 Vrms out, supply ±18 V

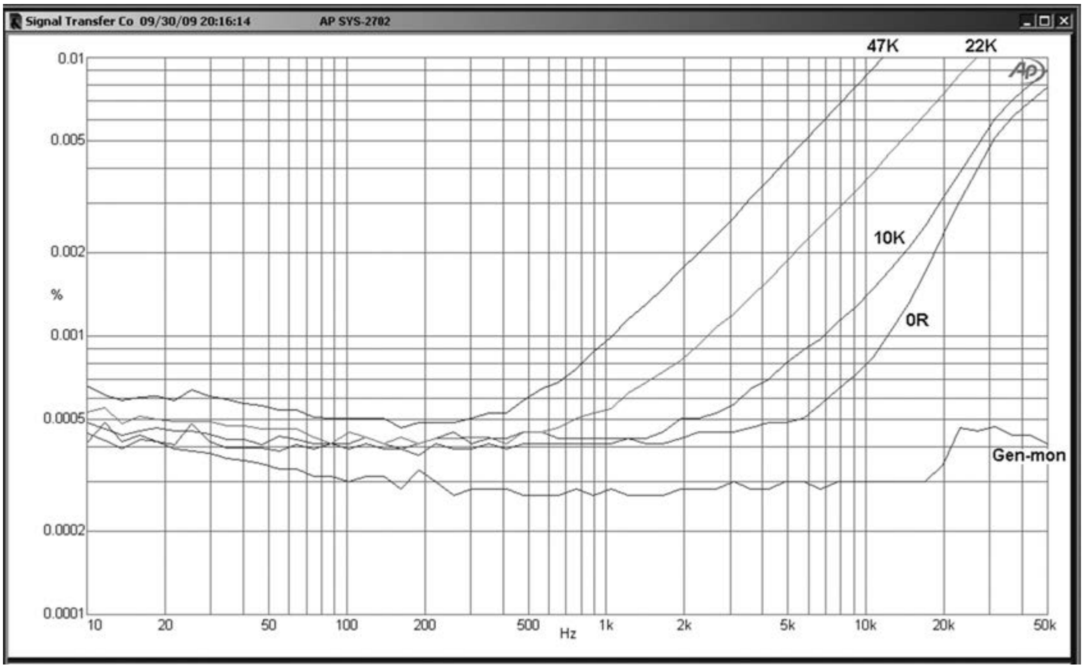

Figure 4.6: 5532 distortion in a series feedback stage with 2k2 and 1kΩ feedback resistors, and varying source resistances. 10 Vrms output

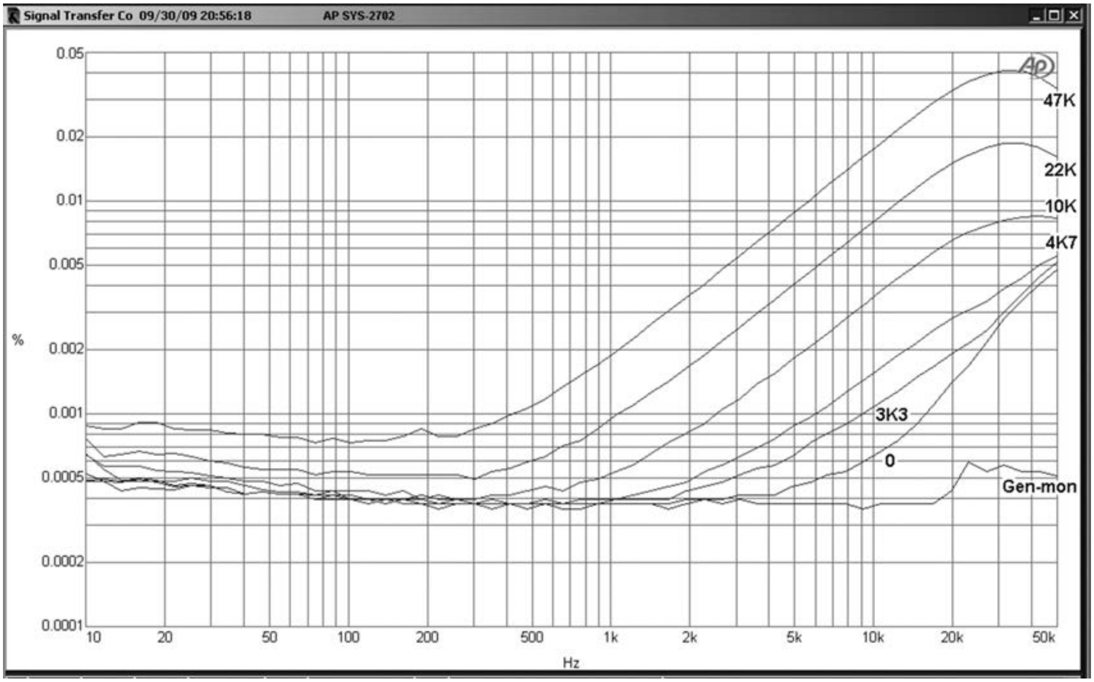

Figure 4.7: 5532 distortion in a voltage-follower circuit with a selection of source resistances. Test level 10 Vrms, supply ±18 V. The lowest trace is the analyser output measured directly, as a reference

CM voltage of 10 Vrms, the distortion is no greater than for the shunt mode. However, when source resistance is inserted in series with the input, the distortion mixture of second, third and other low-order harmonics increases markedly. It increases with output level, approximately quadrupling as level doubles. The THD is now 0.018% at 10 kHz with R_s = 47 kΩ, more than twice that of the series-feedback amplifier above, due to the increased CM voltage.

It would be highly inconvenient to have to stick to the shunt feedback mode, because of the phase inversion and relatively low input impedance that comes with it, so we need to find out how much source resistance we can live with. Figure 4.8 zooms in on the situation with resistance of 10 kΩ and below; when the source resistance is below 2k2, the distortion is barely distinguishable from the zero source resistance trace. This is why the low-pass Sallen and Key filters in Chapter 6 have been given series resistors that do not in total exceed this figure. Close examination reveals the intriguing fact that a 1 kΩ source actually gives *less* distortion than no source resistance at all, reducing THD from 0.00065% to 0.00055% at 10 kHz. Minor resistance variations around 1 kΩ make no difference. This must be due to the cancellation of distortion from two different mechanisms. It is hard to say whether it is repeatable enough to be exploited in practice.

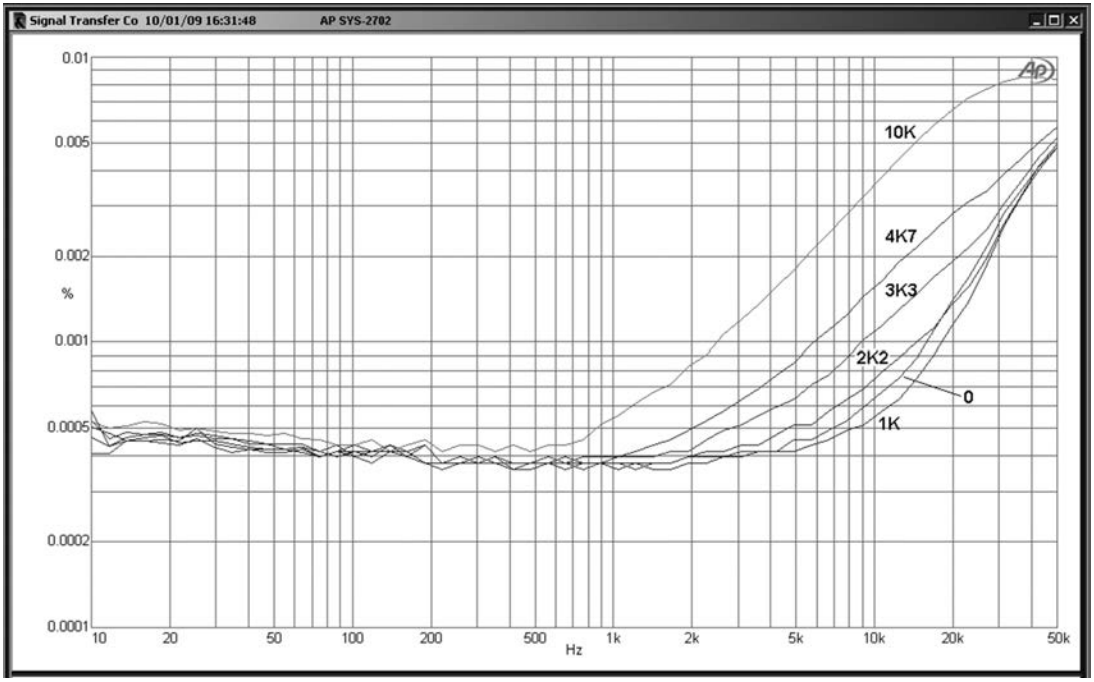

Figure 4.8: A closer look at 5532 distortion in a voltage-follower with relatively low source resistances; note that a 1 kΩ source resistance actually gives less distortion than none. Test level 10 Vrms, supply ±18 V

So, what's going on here? Is it simply due to non-linear currents being drawn by the opamp inputs? Audio power amplifiers have discrete input stages which are very simple compared with those of most opamps, and draw relatively large input currents. These currents show appreciable non-linearity even when the output voltage of the amplifier is virtually distortion-free, and if they flow through significant source resistances will introduce added distortion [4].

If this was the case with the 5532 then the extra distortion would manifest itself whenever the opamp was fed from a significant source resistance, no matter what the circuit configuration. But we have just seen that it only occurs in series-feedback situations; increasing the source resistance in a shunt-feedback does not perceptibly increase distortion. The effect may be present but if so it is very small, no doubt because opamp signal input currents are also very small, and it is lost in the noise.

The only difference is that the series circuit has a CM voltage of about 3 Vrms, while the shunt circuit does not, and the conclusion is that with a bipolar input opamp, you must have *both* a CM voltage and a significant source resistance to see extra distortion. The input stage of a 5532 is a straightforward long-tailed pair (see Figure 4.21 below) with a simple tail current source, and no fancy cascoding, and I suspect that Early effect operates on it when there is a large CM voltage, modulating the quite high input bias currents, and this is what causes the distortion. The signal input currents are much smaller, due to the high open-loop gain of the opamp, and as we have seen appear to have a negligible effect.

JFET opamps

FET input opamps behave differently from bipolar input opamps. Take a look at Figure 4.9, taken from a TL072 working in shunt and in series configuration with a 5 Vrms output. The circuits are as in Figure 4.3a and 4.3b, except that the resistor values have to be scaled up to 10 kΩ and 22 kΩ because the TL072 is nothing like so good at driving loads as the 5532. This unfortunately means that the inverting input is seeing a source resistance of $10k \| 22k = 6.9k$, which introduces a lot of common-mode (CM) distortion – five times as much at 20 kHz as for the shunt case. Adding a similar resistance in the input path cancels out this distortion, and the trace then is the same as the "Shunt" trace in Figure 4.9. Disconcertingly, the value that achieved this was not 6.9k, but 9k1. That means adding −113 dBu of Johnson noise, so it's not always appropriate.

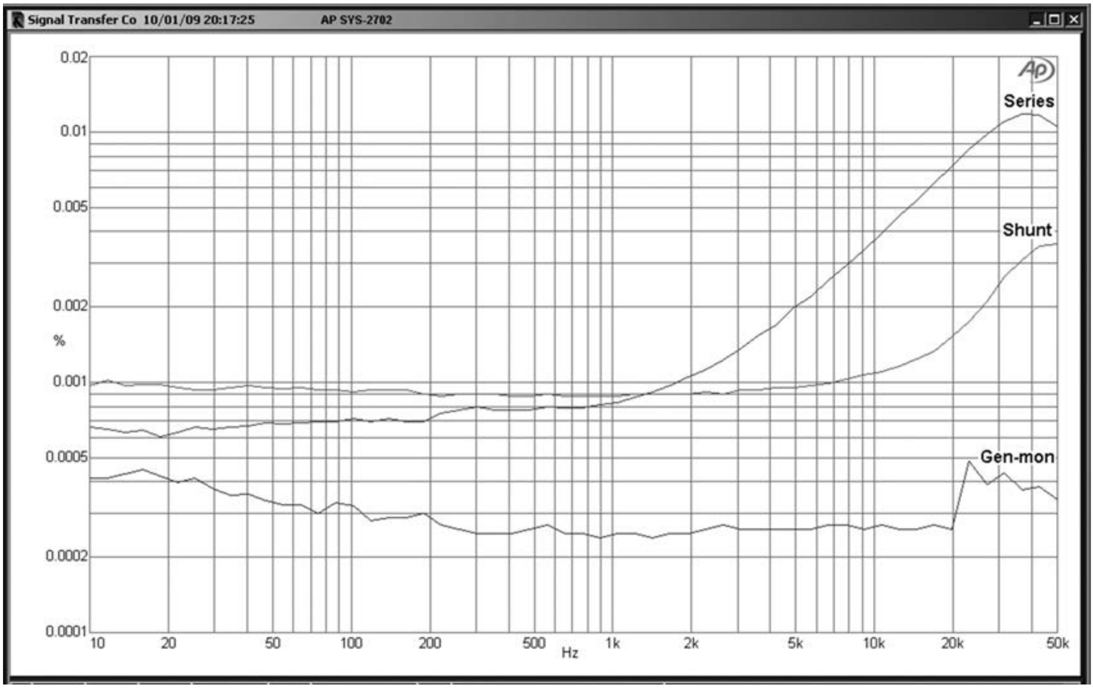

Figure 4.9: A TL072 shunt-feedback stage using 10 kΩ and 22 kΩ resistors shows low distortion. The series version is much worse due to the impedance of the NFB network, but it can be made the same as the shunt case by adding cancellation source resistance in the input path. No external loading, test level 5 Vrms, supply ±18 V

It's worth mentioning that the flat part of the shunt trace below 10 kHz is not noise, as it would be for the 5532; it is distortion.

A voltage-follower has no inconvenient medium-impedance feedback network, but it does have a much larger CM voltage. Figure 4.10 shows a voltage-follower working at 5 Vrms. With no source resistance the distortion is quite low, due to the 100% NFB, but as soon as a 10 kΩ source resistance is added we are looking at 0.015% at 10 kHz.

Once again, this can be cured by inserting an equal resistance in the feedback path of the voltage-follower, as in Figure 4.3d above. This gives the 'Cancel' trace in Figure 4.10. Adding resistances for distortion cancellation in this way has the obvious disadvantage that they introduce extra Johnson noise into the circuit. Another point is that stages of this kind are often driven from pot wipers, so the source impedance is variable, ranging between zero and one-quarter of the pot track resistance. Setting a balancing impedance in the other opamp input to a mid-value, i.e. one-eighth of the track resistance, should reduce the average amount of input distortion, but it is inevitably a compromise.

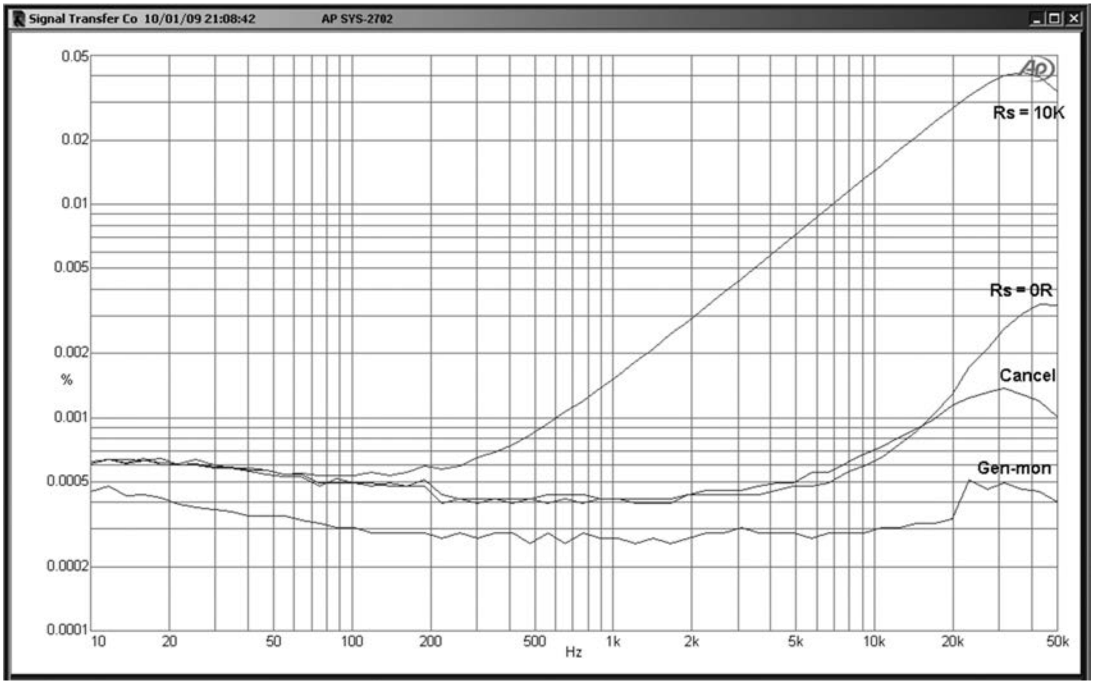

Figure 4.10: A TL072 voltage-follower working at 5 Vrms with a low source resistance produces little distortion (Rs = 0R), but adding a 10 kΩ source resistance makes things much worse (Rs = 10K). Putting a 10 kΩ resistance in the feedback path as well gives complete cancellation of this extra distortion. Supply ±18 V

With JFET inputs the problem is not the operating currents of the input devices themselves, which are negligible, but the currents drawn by the non-linear junction capacitances inherent in field-effect devices. These capacitances are effectively connected to one of the supply rails. For P-channel JFETs, as used in the input stages of most JFET opamps, the important capacitances are between the input JFETs and the substrate, which is normally connected to the V– rail. See Jung [5].

According to the Burr-Brown data sheet for the OPA2134, 'The P-channel JFETs in the input stage exhibit a varying input capacitance with applied CM voltage.' It goes on to recommend that the input impedances should be matched if they are above 2 kΩ.

Common-mode distortion can be minimised by running the opamp off the highest supply rails permitted, though the improvements are not large. In one test on a TL072 going from ±15 V to ±18 V rails reduced the distortion from 0.0045% to 0.0035% at 10 kHz.

Rail bootstrapping to reduce CM distortion

So what do you do if you need a really high-impedance low-distortion voltage-follower and you have a significant source resistance, but you don't want the added noise that would come from adding a cancellation resistor? We noted above that the non-linear input capacitances that cause the trouble with JFET opamp voltage-followers are effectively connected to the V– supply rail or substrate. This suggests a way to remove the problem; if the supply rails are bootstrapped so they go up and down with the inputs, the signal voltage across the non-linear input capacitances is zero, no current can flow through them, and no extra distortion is generated.

Figure 4.11a shows the idea. The resistor-Zener chain R4, D1, D2, R5 creates ±5 V rails which are moved up and down by opamp A4, and buffered by A2, A3. A1 expects reasonably low supply-rail impedances at HF, and attempting to run it directly from the outputs of A2, A3 does not work- the signal disappears in a fog of HF oscillation. The two resistors R2, R3 prevent this by isolating C1 from A2, A3 outputs, while capacitor C1 across A1 supply pins keeps the HF rail impedance low. Since the ±5 V rails of A1 have to remain inside the fixed ±15 V supply rails, the possible swing of the supplies is limited and the maximum output is reduced compared with a basic voltage-follower. The circuit of Figure 4.11a clips at 6.7 Vrms (1 kHz). This could be increased somewhat by using ±17 V or ±18 V fixed rails.

Figure 4.12 shows the result of basic bootstrapping while handling a 5 Vrms signal, which as we saw earlier, is enough to cause serious CM distortion. The increase in linearity is encouraging as the distortion is promptly halved.

Figure 4.13 however, shows that we can do better by adding C2, C3. These are in parallel with the effective slope resistance of the Zeners, and improve the accuracy of the rail bootstrapping. The lower trace marked WITH is indistinguishable from that of the testgear alone.

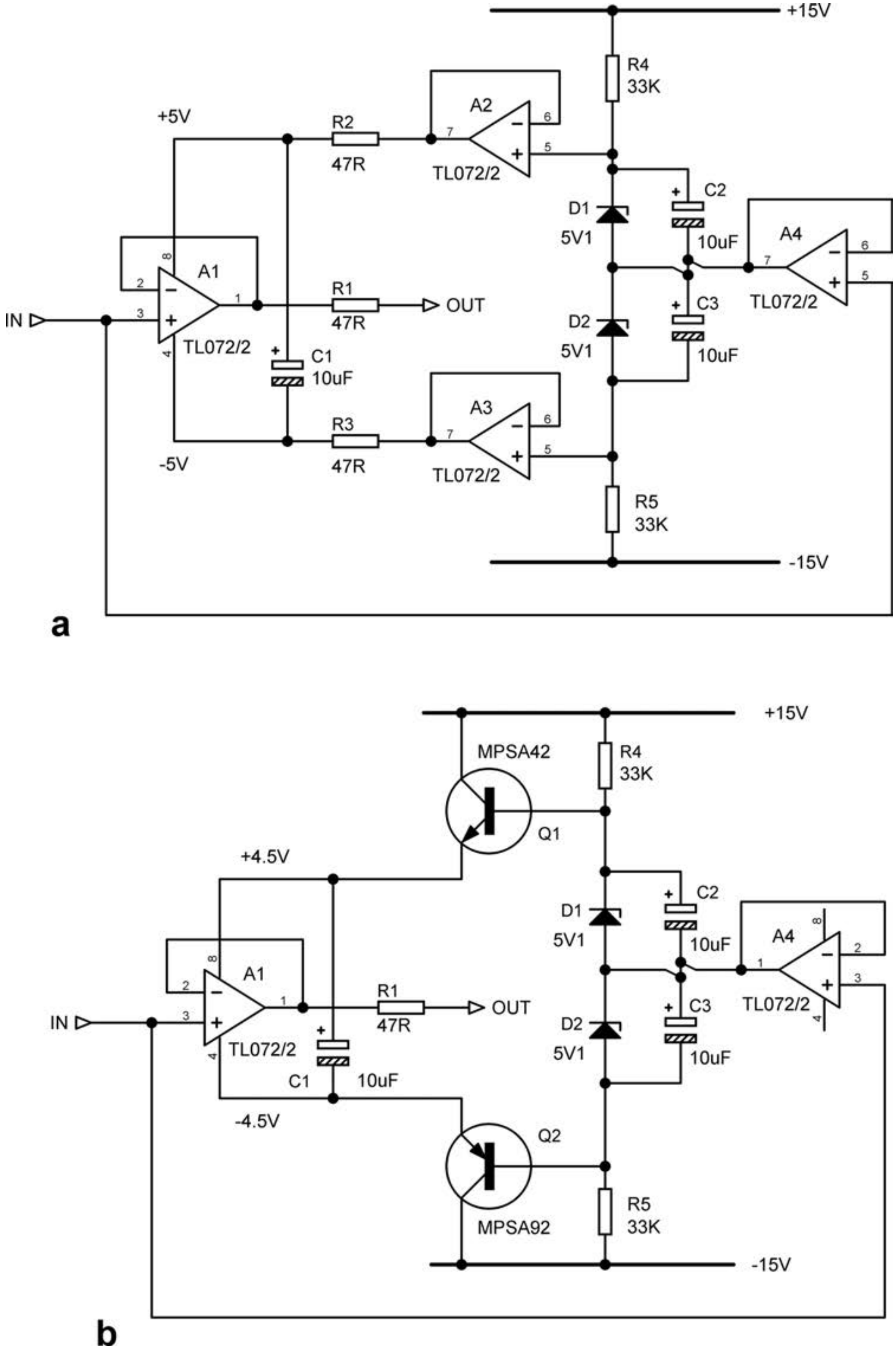

Figure 4.11: Bootstrapping the supply rails of voltage-follower A1 by moving them up and down with the input signal: a) using opamps, b) using transistors

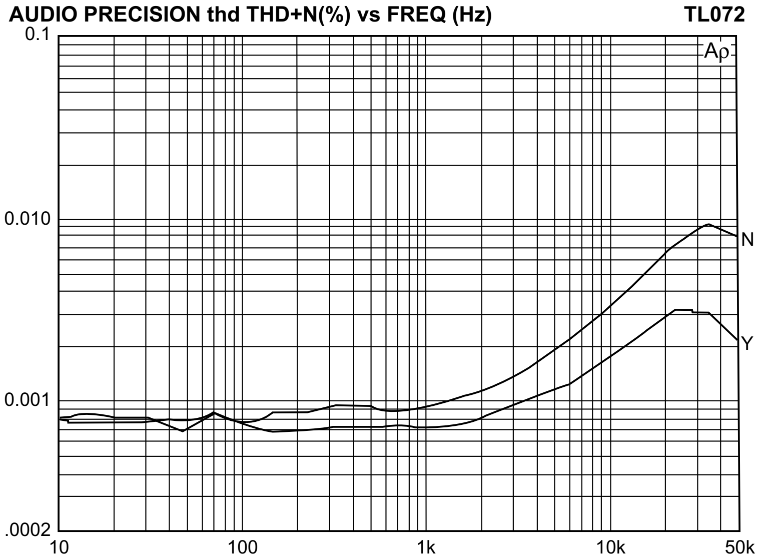

Figure 4.12: TL072 voltage-follower distortion with (Y) and without (N) rail bootstrapping. Test level 5 Vrms, supply ±15 V

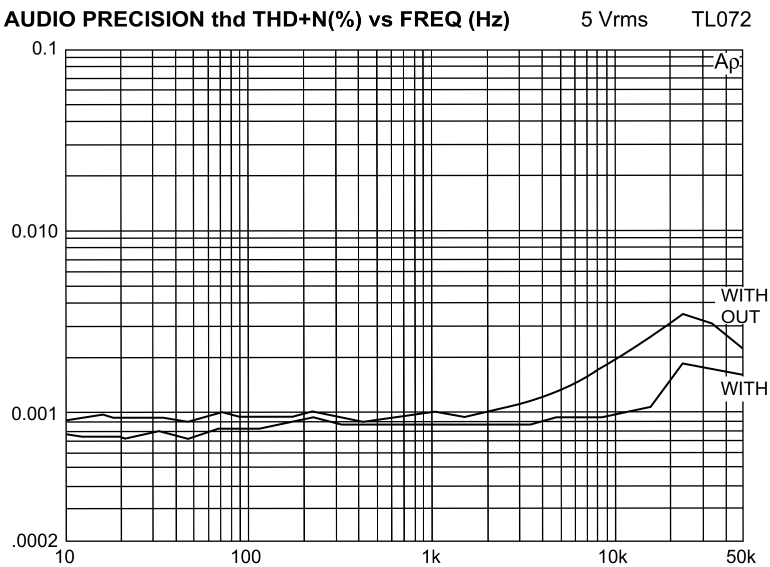

Figure 4.13: Rail bootstrapping is much enhanced by adding capacitors C2, C3. TL072, test level 5 Vrms, supply ±15 V. The "WITH" trace is essentially the distortion of the testgear alone

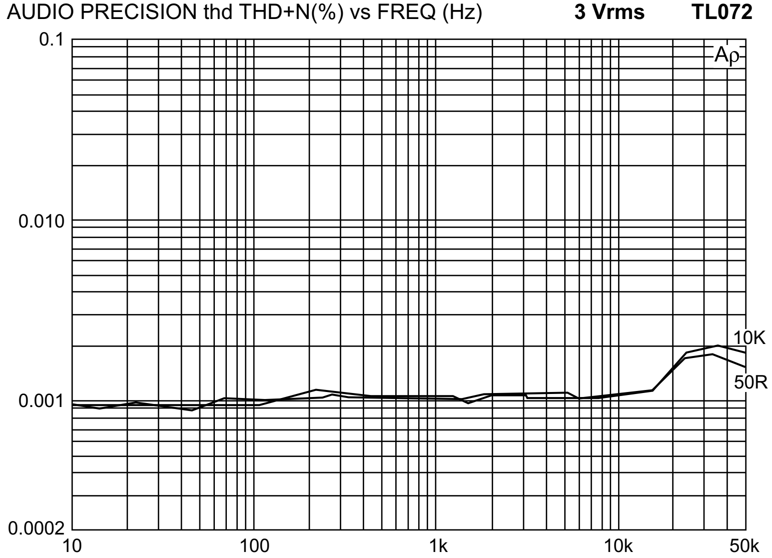

Figure 4.14: Voltage-follower distortion with 10 kΩ and 50 Ω source resistances, and no cancellation. Test level 3 Vrms, supply ±15 V.

Figures 4.12 and 4.13 were taken with near-zero source resistance, and show that internal CM distortion has been dealt with. But what happens when a 10 kΩ source resistance is reintroduced?

Figure 4.14 gives the answer: adding a 10 kΩ source resistance now makes almost no difference. Note that no cancellation resistor has been put in the feedback path.

The rail bootstrapping concept was also tested with the TL052 and the OPA2134 at 5 Vrms, and similar dramatic reductions in CM distortion were found.

Simpler rail bootstrapping

On contemplating Figure 4.11a, it may occur to you that using three opamps to make a friendly environment for one is a bit over-complex. You are quite right. It is always good to simplicate and add lightness when you can, and A2 and A3 can in fact be replaced by simple emitter followers with no detectable loss in performance, as in Figure 4.11b. The two 47 Ω resistors have been removed, but C1 is retained. This seems to be reliably stable. The total supply voltage to A1 has been reduced by two V_{be} drops, or 1.2 V; it could be restored by increasing the Zener voltages if required. The simpler version also uses less power as we no longer need to supply the quiescent currents of A2 and A3.

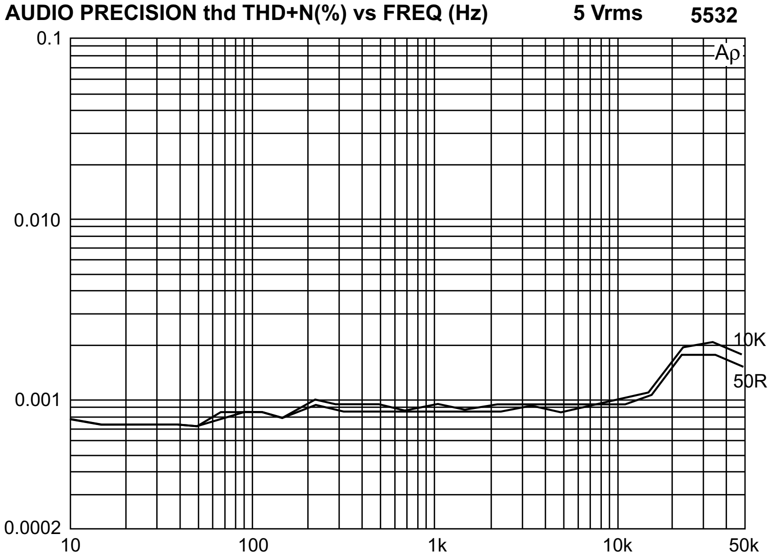

Figure 4.15: Showing that rail bootstrapping works for 5532 voltage-followers as well. 10 kΩ and 50 Ω source resistances, and no cancellation. Test level 5 Vrms, supply ±15 V

The attentive reader will recall that the troublesome non-linear capacitances are effectively connected to the substrate, which is usually the V− supply rail. Would it not be possible to bootstrap just that rail, and leave V+ connected to a fixed +15 V rail? It would; it works, but the results with an OPA2134, while more linear than a conventional voltage-follower, are worse than bootstrapping both rails. Before, we were keeping the magnitude of the A1 supply voltage substantially constant, although it was sailing up and down. If only one rail is bootstrapped, the actual supply voltage is being modulated, so it is hardly surprising that linearity suffers.

In the previous section we saw that CM distortion is also generated by bipolar input opamps, though by a different mechanism, so rail bootstrapping ought to work for these types of opamp as well. Figure 4.15 shows that it does. Adding a 10 kΩ source resistance now causes virtually no extra distortion.

Bootstrapping series-feedback JFET opamp stages

The voltage-follower is the worst case for CM distortion, as the full output voltage exists on both inputs. In contrast, it is the best case for output loading, as there is no resistive feedback network at all to drive- just a high-impedance input pin. Similarly, the shunt feedback amplifier is the best case for CM distortion as there is no significant signal on the inputs.

Series-feedback amplifier stages fall between these two cases. For a +10 dB amplifier stage, the signal on the inputs is one-third that of the output, and so the input distortion is less, but

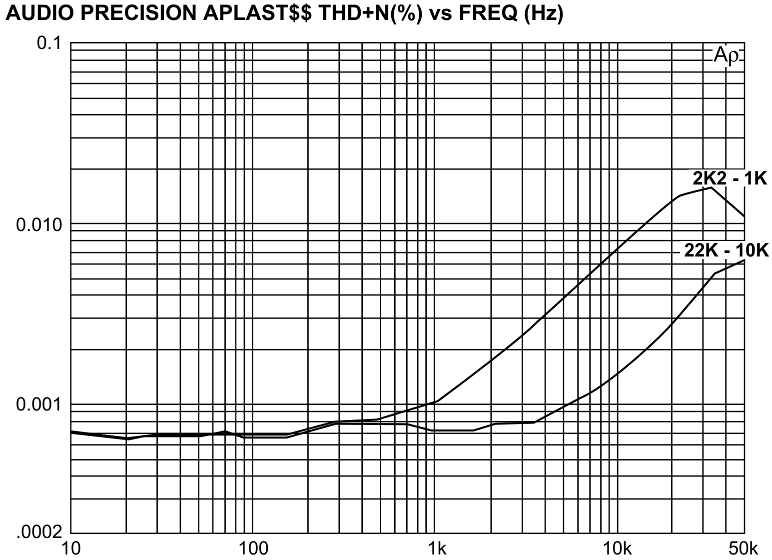

Figure 4.16: With the TL072, reducing the impedance of the negative feedback network may reduce input distortion, but output distortion more than makes up for it because of the extra loading. Upper trace 2k2−1 kΩ, lower trace 22 kΩ−10 kΩ in feedback network

still very definitely present, as we saw in Figures 4.6 and 4.9 above. Amplifier stages like this can have a mixture of distortion mechanisms. The impedance of the negative feedback network, as seen from the inverting input of the amplifiers, is 22 kΩ in parallel with 10 kΩ, i.e. 6.87 kΩ. We have seen above that this is enough to cause serious non-linearity unless the other input sees the same impedance, and it might be thought that reducing the impedance level of the negative feedback network would be a good way to deal with this, not least because it would minimise the Johnson noise produced by the network. Figure 4.16 shows that this does not work for the TL072; if the feedback network impedance is reduced by a factor of ten the distortion gets worse rather than better, due to the heavier loading on the output. Input distortion has been replaced by a larger amount of output distortion; this is not a good exchange. Lowering the NFB network impedance is however likely to be successful with JFET opamps having better load-driving capability than the TL072.

Rail bootstrapping is once more a possible answer. We drive the opamp supply rails up and down with the same signal as the *input* – not the output. The only modification required is to take the increased output swing into account by increasing the A1 supply voltage to ±10 V; see Figure 4.17. The Zeners have been replaced with simple resistive dividers. This works just as well, and is a ‘good thing’ as Zeners are more expensive than resistors. Figure 4.18 shows the excellent results.

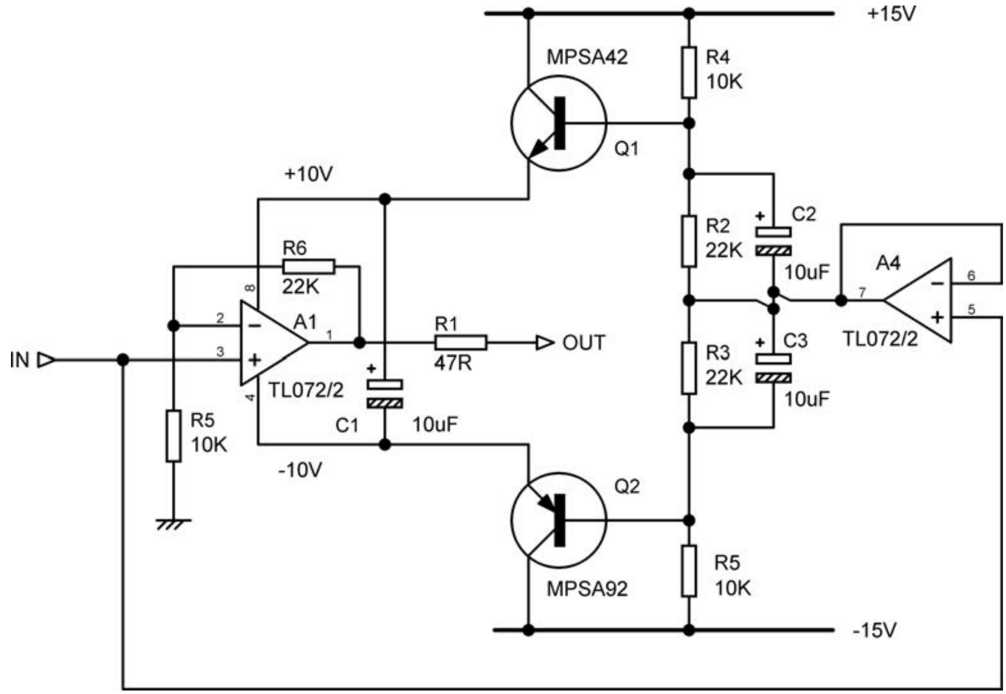

Figure 4.17: Bootstrapping the rails of a series-feedback amplifier from the input of A1. The Zeners have been replaced by resistors

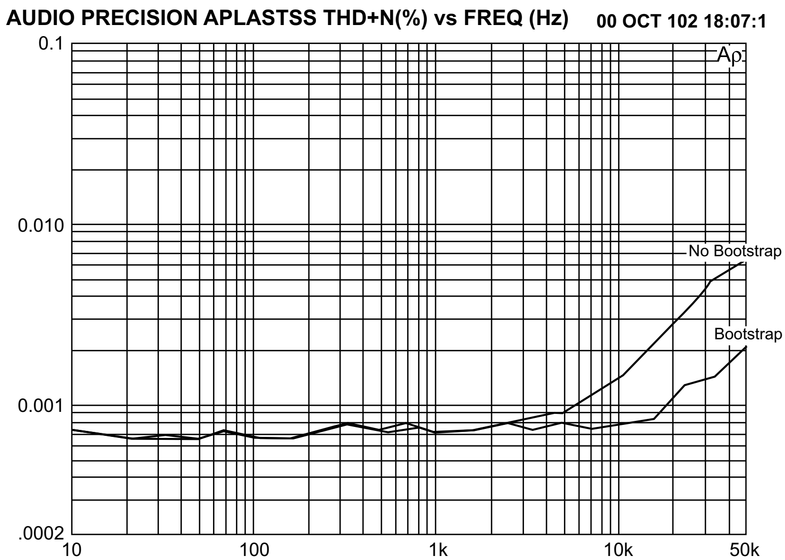

Figure 4.18: The benefit of bootstrapping the rails of a series-feedback amplifier with a gain of 3.2 times. The lower trace is essentially the THD from the test equipment

Selecting the right opamp

Until recently, the 5532 was pre-eminent. It is found in almost every mixing console, and in a large number of preamplifiers. Distortion is very low, even when driving 600 Ω loads. Noise is very low, and the balance of voltage and current noise in the input stage is well–matched to moving-magnet phono cartridges; in many applications discrete devices give no significant advantage. Large-quantity production has brought the price down to a point where a powerful reason is required to pick any other device.

The 5532 is not, however, perfect. It suffers common-mode distortion. It has high bias and offset currents at the inputs, as an inevitable result of using a bipolar input stage (for low noise) without any sort of bias-cancellation circuitry. The 5532 is not in the forefront for DC accuracy, though it's not actually that bad. The offset voltage spec is 0.5 mV typical, 4 mV max, compared with 3 mV typical, 6 mV max for the popular TL072. I have actually used 5532s to replace TL072s when offset voltage was a problem, but the increased bias current was acceptable.

With horrible inevitability, the very popularity and excellent technical performance of the 5532 has led to it being criticised by subjectivists who have contrived to convince themselves that they can tell opamps apart by listening to music played through them. This always makes me laugh, because there is probably no music on the planet that has not passed through a hundred or more 5532s on its way to the consumer.

The LM4562 represents a real advance on the 5532. It is however still a good deal more expensive, and is not perfect – it appears to be more easily damaged by excess common-mode voltages, and there is some evidence it is more susceptible to RF demodulation.

In some applications, such as low-cost mixing consoles, bipolar-style bias currents are a real nuisance because keeping them out of EQ pots to prevent scratching noises requires an inappropriate number of blocking capacitors. There are plenty of JFET input opamps around with negligible bias currents, but there is no obviously superior device that is the equivalent of the 5532. The TL072 has been used in this application for many years but its HF linearity is not first-class and distortion across the band deteriorates badly as output loading increases. However, the opamps in many EQ sections work in the shunt-feedback configuration with no CM voltage on the inputs, and this reduces the distortion considerably. When low bias currents are needed with superior performance then the OPA2134 is often a good choice, though it is at least four times as expensive as the TL072.

Opamps surveyed: BJT input types

The rest of this chapter looks at some opamp types and examines their performance, with the 5532 the usual basis for comparison. The parts shown here are not necessarily intended as audio opamps, though some, such as the OP275 and the OPA2134, were specifically designed as such. They have however all seen use, in varying numbers, in audio applications. Bipolar input opamps are dealt with first.

The LM741 opamp

The LM741 is only included here for its historical interest; in its day it was a most significant development, and to my mind, the first really practical opamp. It was introduced by Fairchild in 1968 and is considered a second-generation opamp, the 709 being first generation.

The LM741 had (and indeed has) effective short-circuit protection and internal compensation for stability at unity gain, and was much easier to make work in a real circuit than its predecessors. It was clear that it was noisy compared with discrete circuitry, and you sometimes had to keep the output level down if slew-limiting was to be avoided, but with care it was usable in audio. Probably the last place the LM741 lingered was in the integrators of state-variable EQ filters, where neither indifferent noise performance nor poor slewing capability is a serious problem; see Chapter 15 for more details on this application. The LM741 is a single opamp. The dual version is the LM747.

Figure 4.19 shows a region between 100 Hz and 4 kHz where distortion rises at 6 dB/octave. This is the result of the usual dominant-pole Miller compensation scheme. When slew-limiting begins, the slope increases and THD rises rapidly with frequency.

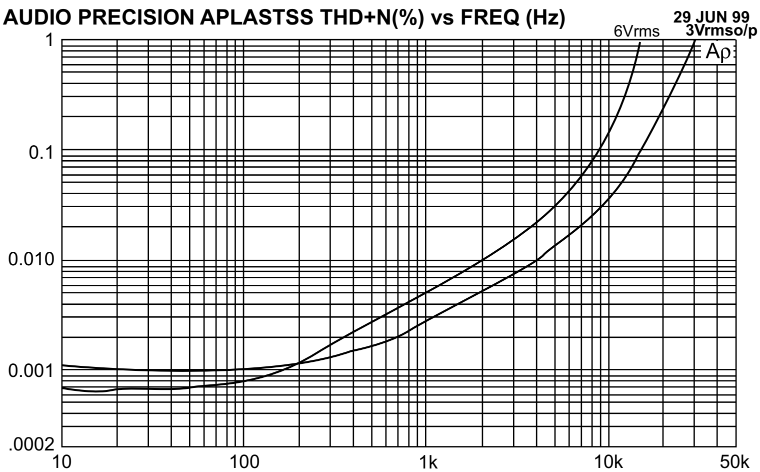

Figure 4.19: The THD performance of an LM741 working at a gain of 3×, on ±15 V rails, giving 3 Vrms and 6 Vrms outputs, with no load. At 6 Vrms, slew distortion exceeds 1% before 20 kHz is reached; there is visible slew-limiting in the waveform. THD is however very low at 100 Hz, due to the high NFB factor at low frequencies

The NE5532/5534 opamp

The 5532 is a low-noise, low distortion bipolar dual opamp, with internal compensation for unity-gain stability. The 5534 is a single version internally compensated for gains down to three times, and an external compensation capacitor can be added for unity-gain stability; 22 pF is the usual value. The 5532 achieves unity-gain stability by having degeneration resistors in the emitter circuits of the input transistors, to reduce the open-loop gain, and this is why it is noisier than the 5534.

The common-mode range of the inputs is a healthy ±13V, with no phase inversion problems if this is exceeded. It has a distinctly higher power consumption than the TL072, drawing approx 4 mA per opamp section when quiescent. The DIL version runs perceptibly warm when quiescent on ±17 V rails.

The 5534/5532 has bipolar transistor input devices. This means it gives low noise with low source resistances, but draws a relatively high bias current through the input pins. The input devices are NPN, so the bias currents flow into the chip from the positive rail. If an input is fed through a significant resistance then the input pin will be more negative than ground due to the voltage-drop caused by the bias current. The inputs are connected together with back-to-back diodes for reverse-voltage protection; and should not be forcibly pulled to different voltages. The 5532 is intended for linear operation, and using it as a comparator is not recommended.

As can be seen from Figure 4.20, the 5532 is almost distortion-free, even when driving the maximum 500 Ohm load. The internal circuitry of the 5532 has never been officially explained, but appears to consist of nested Miller loops that permit high levels of internal negative feedback. The 5532 is the dual of the 5534, and is much more commonly used than the single as it is cheaper per opamp and does not require an external compensation capacitor when used at unity gain.

The 5532/5534 is made by several companies, but they are not all created equal. Those by Fairchild, JRC, and ON-Semi have significantly lower THD at 20 kHz and above, and we're talking about a factor of two or three here.

The 5532 and 5534 type opamps require adequate supply-decoupling if they are to remain stable; otherwise they appear to be subject to some sort of internal oscillation that degrades linearity without being visible on a normal oscilloscope. The essential requirement is that the +ve and −ve rails should be decoupled with a 100 nF capacitor between them, at a distance of not more than a few millimetres from the opamp; normally one such capacitor is fitted per package as close to it as possible. It is *not* necessary, and often not desirable to have two capacitors going to ground; every capacitor between a supply rail and ground carries the risk of injecting rail noise into the ground.

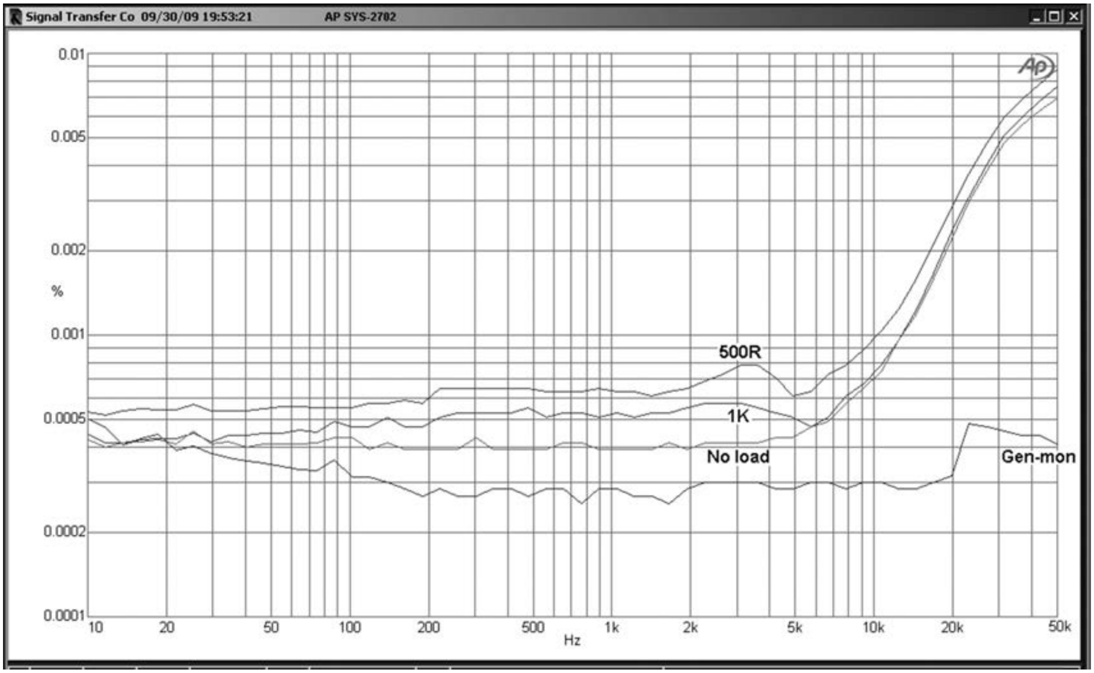

Figure 4.20: Distortion is very low from the 5532, though loading makes a detectable difference. Here it is working in series feedback mode at the high level of 10 Vrms with 500 Ω, 1 kΩ loads and no load. The Gen-mon trace is the output of the distortion analyser measured directly. Gain of 3.2 times. Supply ±18 V

Deconstructing the 5532

To the best of my knowledge, virtually nothing has been published about the internal operation of the 5532. This is surprising given its unique usefulness as a high-quality audio opamp. I believe the secret of the 5532's superb linearity is the use of nested negative feedback inside the circuit, in the form of traditional Miller compensation.

Figure 4.21 shows the only diagram of the internal circuitry that has been released; the component and node numbers are mine. It has been in the public domain for at least twenty years, so I hope no one is going to object to my impertinent comments on it. The circuit initially looks like a confusing sea of transistors, and there is even a solitary JFET lurking in there, but it breaks down fairly easily. There are three voltage-gain stages, plus a unity-gain output stage to increase drive capability. This has current-sensing overload protection.

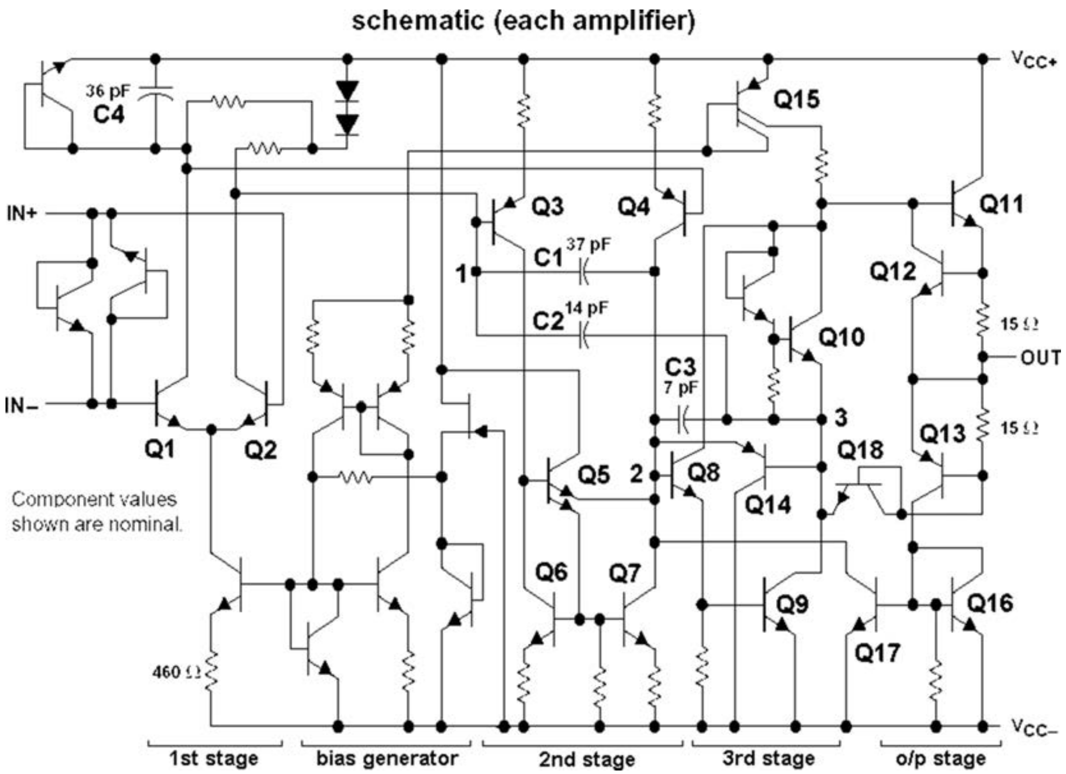

Figure 4.21: The internal circuitry of the 5534. Note absence of emitter resistors for Q1 and Q2

There is also a fairly complex bias generator which establishes the operating currents in the various stages.

In all conventional opamps there are two differential input signals which have to be subtracted to create a single output signal, and the node at which this occurs is called the 'phase-summing point'.

Q1, Q2 make up the input differential amplifier. They are protected against reverse-biasing by the diode-connected transistors across the input pins. Note there are no emitter degeneration resistors, which would linearise the input pair at the expense of degrading noise. Presumably high open-loop gain (note there are three gain stages, whereas a power amplifier normally only has two) means that the input pair is handling very small signal levels so its distortion is not a problem.

Q3, Q4 make up the second differential amplifier; emitter degeneration is now present. Phase summing occurs at the output of this stage at Node 2. C1 is the Miller capacitor around this stage, from Node 2 to Node 1. Q5, Q6, Q7 are a Wilson current-mirror which provides a driven current-source as the collector load of Q4. The function of C4 is obscure but it appears to balance C1 in some way.

The third voltage-amplifier stage is basically Q9 with split-collector transistor Q15 as its current-source load. Q8 increases the basic transconductance of the stage, and C3 is the Miller capacitor around it, feeding from Node 3 to Node 2- note that this Miller loop does not include the output stage. Things are a bit more complicated here as it appears that Q9 is also the sink half of the Class-B output stage. Q14 looks very mysterious as it seems to be sending the output of the third stage back to the input; possibly it's some sort of clamp to ensure clean clipping, but to be honest I haven't a clue. Q10 plus associated diode generates the bias for the class-B output stage, just as in a power amplifier.

The most interesting signal path is the semi-local Miller loop through C2, from Node 3 to Node 1, which encloses both the second and third voltage amplifiers; each of these have their own local Miller feedback, so there are two nested layers of internal feedback. This is probably the secret of the 5532's low distortion.

Q11 is the source side of the output stage, and as mentioned above, Q9 appears to be the sink. Q12, Q13 implement overcurrent protection. When the voltage drop across the 15 Ω resistor becomes too great, Q12 turns on and shunts base drive away from Q11. In the negative half-cycle, Q13 is turned on, which in turn activates Q17 to shunt drive away from Q8.

The biasing circuit shows an interesting point. Bipolar bias circuits tend not to be self-starting; no current flows anywhere until some flows somewhere, so to speak. Relying on leakage currents for starting is unwise, so here the depletion-mode JFET provides a circuit element that is fully on until you bias it off, and can be relied upon to conduct as the power rails come up from zero.

More information on the internals of the 5534 can be found in reference [6].

The LM4562 opamp

The LM4562 is a new opamp, which first became freely available at the beginning of 2007. It is a National Semiconductor product. It is a dual opamp – there is no single or quad version. It costs about ten times as much as a 5532.

The input noise voltage is typically 2.7nV/$\sqrt{\text{Hz}}$, which is substantially lower than the 4nV/$\sqrt{\text{Hz}}$ of the 5532. For suitable applications with low source impedances this translates into a useful noise advantage of 3.4 dB. The bias current is typically 10 nA, which is very low and would normally imply that bias-cancellation, with its attendant noise problems, was

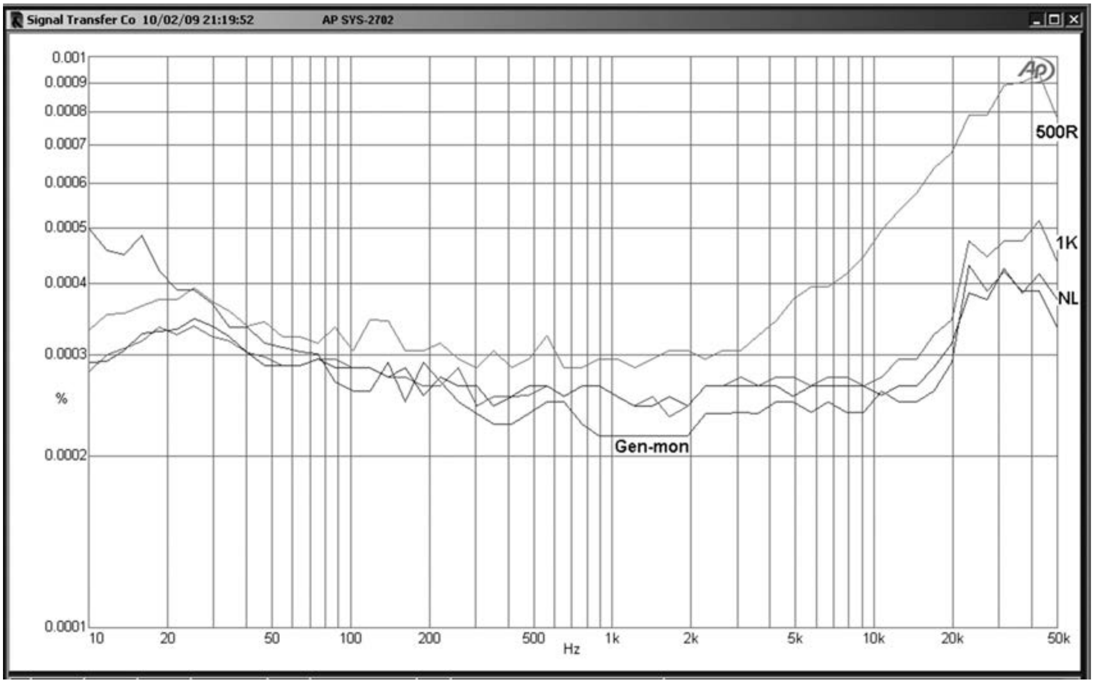

Figure 4.22: The LM4562 in shunt feedback mode, with 1 kΩ, 2k2 feedback resistors giving a gain of 2.2×. Shown for no load (NL) and 1 kΩ, 500 Ω loads. Note the vertical scale ends at 0.001% this time. Output level is 10 Vrms. ±18 V supply rails

being used. However, in my testing I have seen no sign of excess noise, and the data sheet is silent on the subject. No details of the internal circuitry have been released so far, and quite probably never will be.

It is not fussy about decoupling, and as with the 5532, 100 nF across the supply rails close to the package should ensure HF stability. The slew rate is typically ±20 V/µs, more than twice as quick as the 5532.

The first THD plot in Figure 4.22 shows the LM4562 working at a closed-loop gain of 2.2 times in shunt feedback mode, at a high level of 10 Vrms. The top of the THD scale is 0.001%, something you will see with no other opamp in this survey. The no-load trace is barely distinguishable from the AP SYS-2702 output, and even with a heavy 500 Ω load driven at 10 Vrms there is only a very small amount of extra THD, reaching 0.0007% at 20 kHz.

Figure 4.23 shows the LM4562 working at a gain of 3.2 times in series feedback mode, both modes having a noise gain of 3.2 times. There is little extra distortion from 500 Ω.

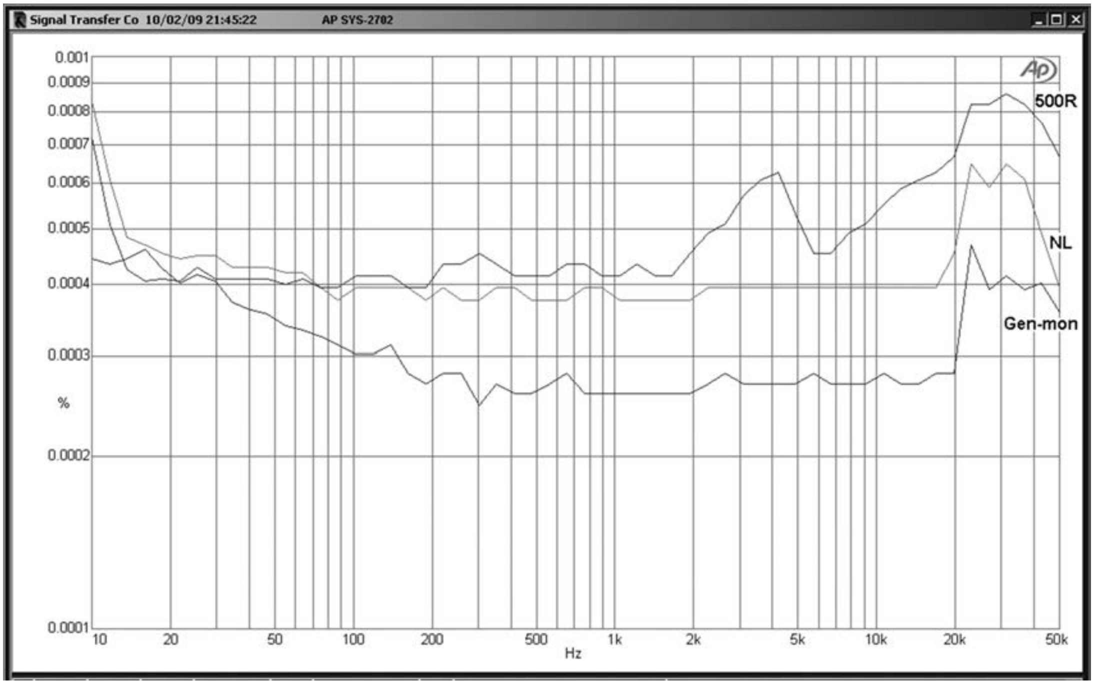

Figure 4.23: The LM4562 in series feedback mode, with 1 kΩ, 2k2 feedback resistors giving a gain of 3.2×. No load (NL) and 500 Ω load. 10 Vrms output. ±18 V supply rails

For Figures 4.22 and 4.23 the feedback resistances were 2k2 and 1 kΩ, so the minimum source resistance presented to the inverting input is 687 Ω. In Figure 4.24 extra source resistances were then put in series with the input path, (as was done with the 5532 in the section above on common-mode distortion) and this revealed a remarkable property of the LM4562 – it is much more resistant to common-mode distortion than the 5532. At 10 Vrms and 10 kHz, with a 10 kΩ source resistance the 5532 generates 0.0014% THD (see Figure 4.6) but the LM4562 gives only 0.00046% under the same conditions. I strongly suspect that the LM4562 has a more sophisticated input stage than the 5532, probably incorporating cascoding to minimise the effects of common-mode voltages.

Note that only the rising curves to the right represent actual distortion. The raised levels of the horizontal traces at the LF end is due to Johnson noise from the extra series resistance.

It has taken an unbelievably long time – nearly 30 years – for a better audio opamp than the 5532 to come along, but at last it has happened. The LM4562 is superior in just about every parameter, but it has much higher current noise. At present it also has a much higher price, but hopefully that will change.

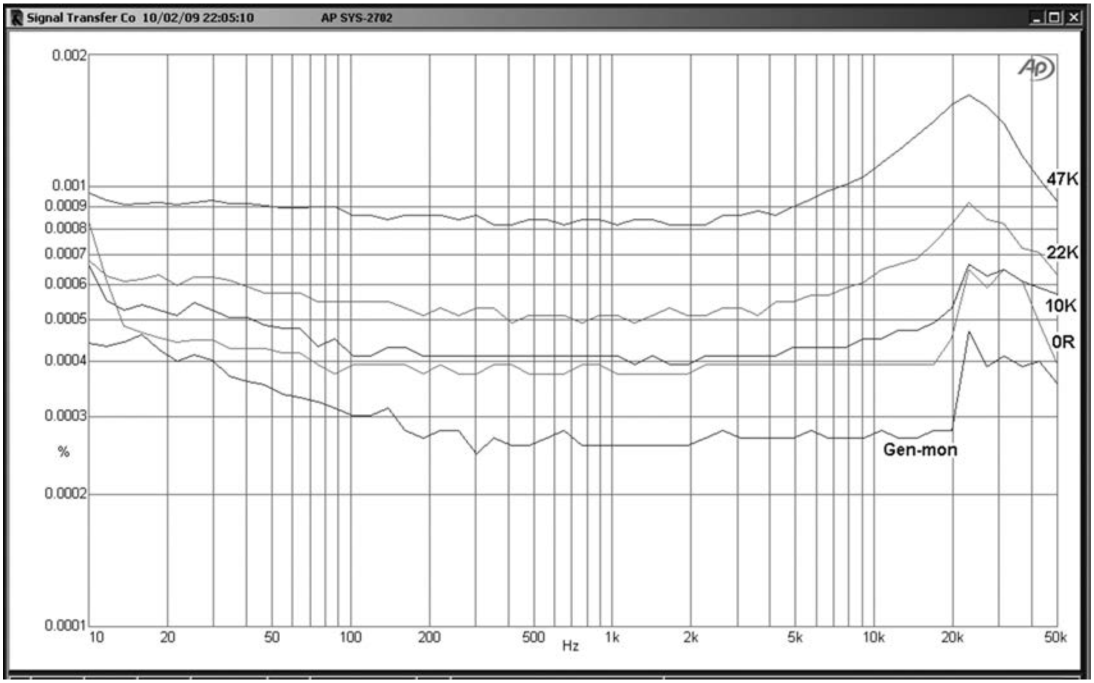

Figure 4.24: The LM4562 in series feedback mode, gain 3.2×, with varying extra source resistance in the input path. The extra distortion is much lower than for the 5532. 10 Vrms out, ±18 V supply rails

The AD797 opamp

The AD797 (Analog Devices) is a single opamp with very low voltage noise and distortion. It appears to have been developed primarily for the cost-no-object application of submarine sonar, but it works very effectively with normal audio – if you can afford to use it. The cost is something like 20 times that of a 5532. No dual version is available, so the cost ratio per opamp section is forty times.

This is a remarkably quiet device in terms of voltage noise, but current noise is correspondingly high due to the high currents in the input devices. Early versions appeared to be rather difficult to stabilise at HF, but the current product is no harder to apply than the 5532. Possibly there has been a design tweak, or on the other hand my impression may be wholly mistaken.

The AD797 incorporates an ingenious feature for internal distortion cancellation. This is described on the manufacturer's data sheet. Figure 4.25 shows that it works effectively.

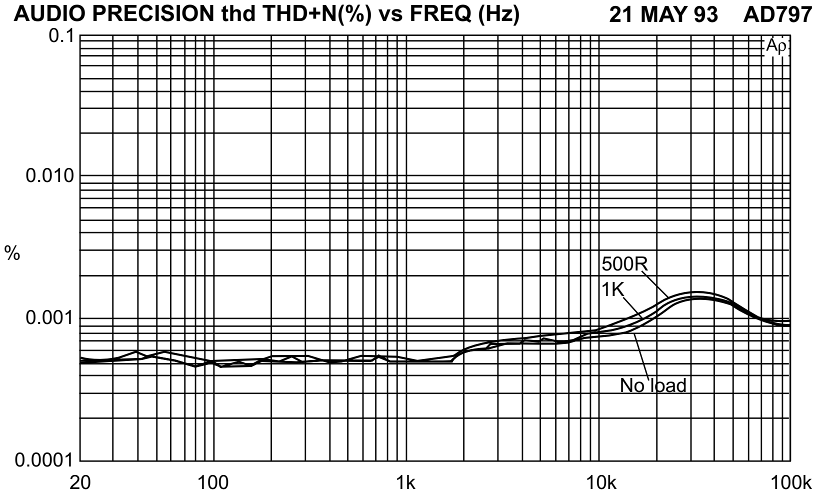

Figure 4.25: AD797 THD into loads down to 500 Ω, at 7.75 Vrms. Output is virtually indistinguishable from input. Series feedback, but no CM problems. Gain = 3.2×

The OP27 opamp

The OP27 from Analog Devices is a bipolar input, single opamp primarily designed for low noise and DC precision. It was not intended for audio use, but in spite of this it is frequently recommended for such applications as RIAA and tape head preamps. This is unfortunate, because while at first sight it appears that the OP27 is quieter than the 5534/5532, as the e_n is 3.2nV/$\sqrt{\text{Hz}}$ compared with 4nV/$\sqrt{\text{Hz}}$ for the 5534, in practice it is usually slightly noisier. This is because the OP27 is in fact optimised for DC characteristics, and so has input bias-current cancellation circuitry that generates common-mode noise. When the impedances on the two inputs are very different – which is the case in RIAA preamps – the CM noise does not cancel, and this appears to degrade the overall noise performance significantly.

For a bipolar input opamp, there appears to be a high level of common-mode input distortion, enough to bury the output distortion caused by loading; see Figures 4.26 and 4.27. It is likely that this too is related to the bias-cancellation circuitry, as it does not occur in the 5532.

The maximum slew rate is low compared with other opamps, being typically 2.8 V/µs. However, this is not the problem it may appear. This slew rate would allow a maximum amplitude at 20 kHz of 16 Vrms, if the supply rails permitted it. I have never encountered any particular difficulties with decoupling or stability of the OP27.

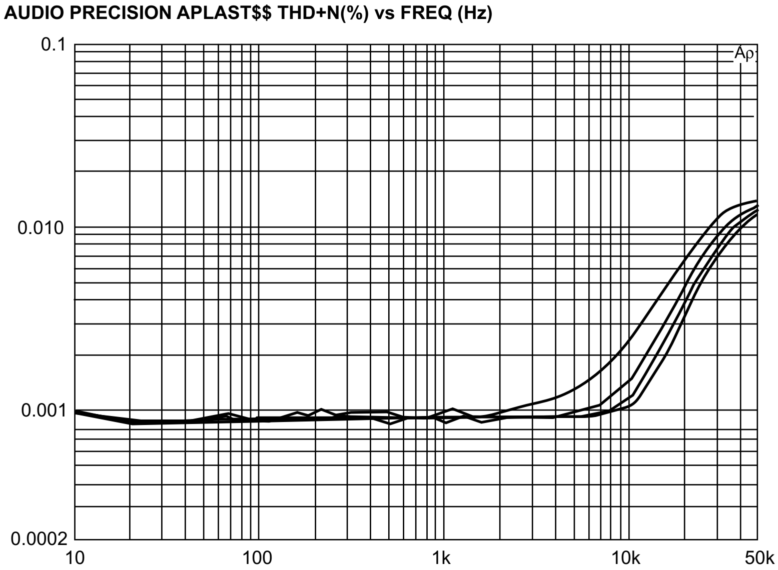

Figure 4.26: OP27 THD in shunt feedback mode with varying loads. This opamp accepts even heavy (1 kΩ) loading gracefully

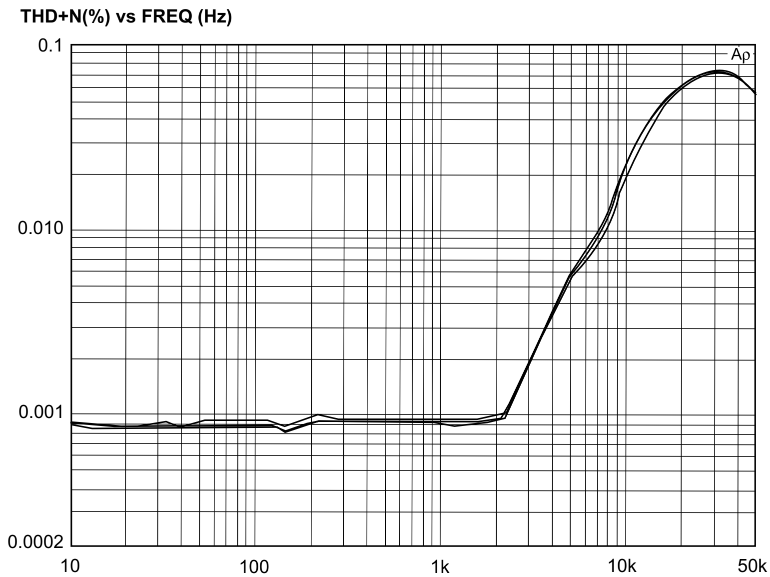

Figure 4.27: OP27 THD in series feedback mode. The common-mode input distortion completely obscures the output distortion

The OP270 opamp

The OP270 from Analog Devices is a dual opamp, intended as a 'very low noise precision operational amplifier' – in other words combining low noise with great DC accuracy. The input offset voltage is an impressively low 75 μV maximum. It has bipolar inputs with a bias current cancellation system; the presence of this is shown by the 15 nA bias current spec, which is 30 times less than the 500 nA taken by the 5534, which lacks this feature. It will degrade the noise performance with unequal source resistances, as it does in the OP27. The input transistors are protected by back-to-back diodes.

The OP270 distortion performance suffers badly when driving even modest loads (see Figures 4.28 and 4.29). The slew rate is a rather limited 2.4 V/μs, which is only just enough for a full output swing at 20 kHz. Note also that this is an expensive opamp, costing something like 25 times as much as a 5532; precision costs money. Unless you have a real need for DC accuracy, this part is not recommended.

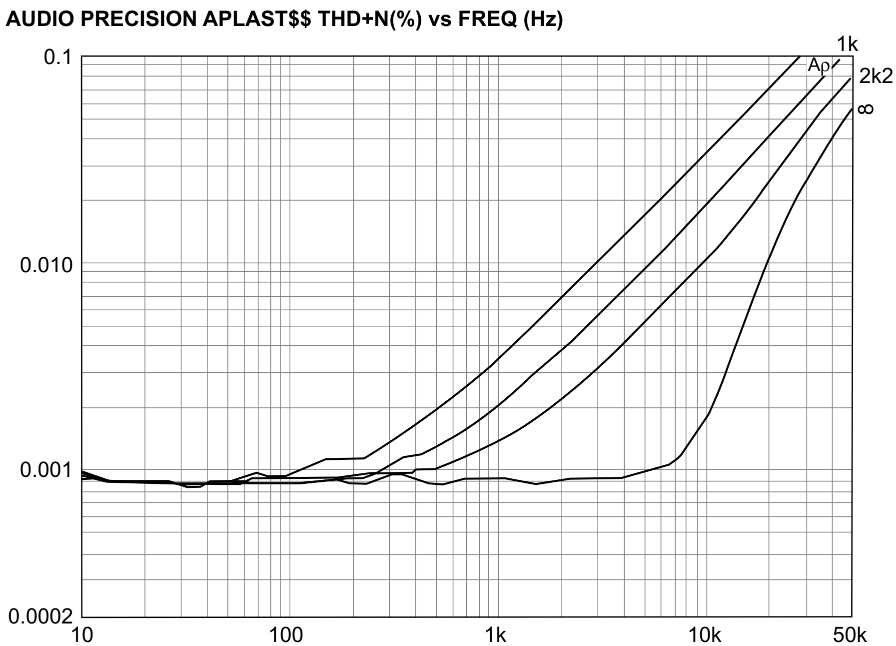

Figure 4.28: OP270 THD in shunt feedback mode. Linearity is severely degraded even with a 2k2 load

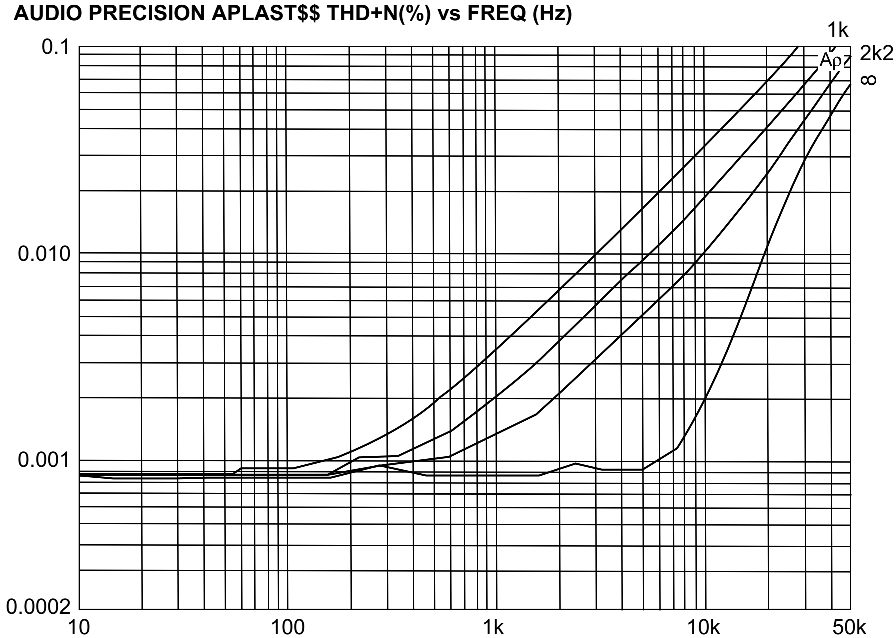

Figure 4.29: OP270 THD in series feedback mode. This looks the same as Figure 4.28 so CM input distortion appears to be absent

The OP275 opamp

The Analog Devices OP275 is one of the few opamps specifically marketed as an audio device. Its most interesting characteristic is the Butler input stage which combines bipolar and JFET devices. The idea is that the bipolars give accuracy and low noise, while the JFETs give speed and 'the sound quality of JFETs'. That final phrase is not a happy thing to see on a datasheet from a major manufacturer; the sound of JFETs (if any) would be the sound of high distortion. Just give us the facts, please.

The OP275 is a dual opamp; no single version is available. It is quite expensive, about six times the price of a 5532, and its performance in most respects is inferior. It is noisier, has higher distortion, and does not like heavy loads (see Figures 4.30 and 4.31). The CM range is only about two-thirds of the voltage between the supply rails, and I_{bias} is high due to the BJT part of the input stage. Unless you think there is something magical about the BJT/JFET input stage – and I am quite sure there is not – it is probably best avoided.

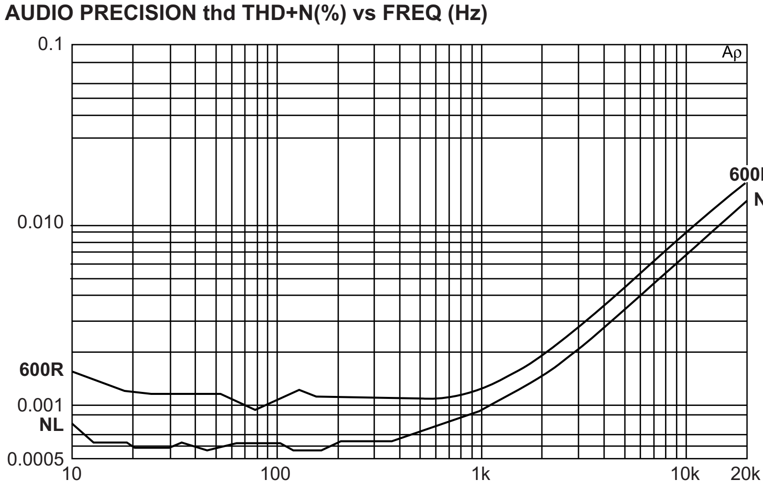

Figure 4.30: An OP275 driving 7.75 Vrms into No Load and 600 Ω. THD below 1 kHz is definitely non-zero with the 600 Ω load. Series feedback, gain 3.2×

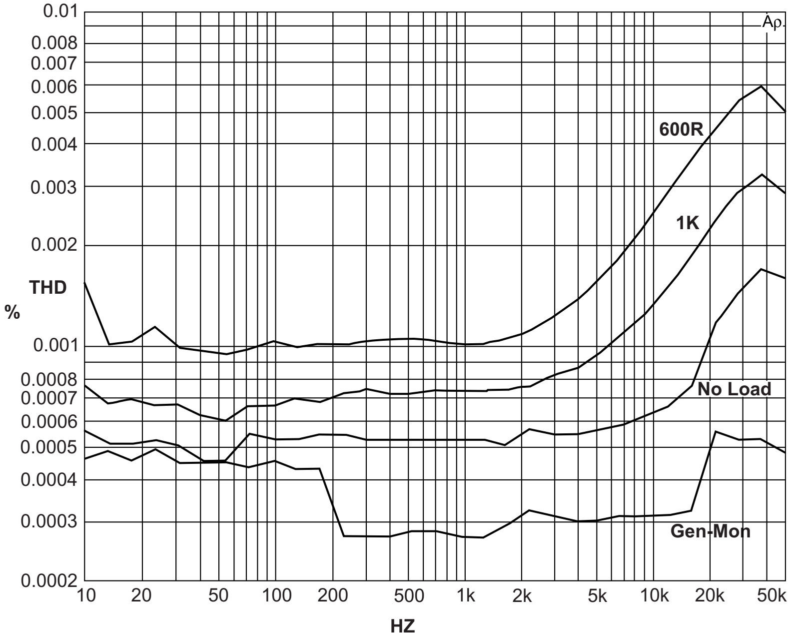

Figure 4.31: OP275 driving 5 Vrms into 1K and 600 Ohms. Shunt feedback, gain 2.2×, but note noise gain was set to 3.2× as for the series case. The Gen-mon trace shows the distortion of the AP System 2 generator; the steps at 200 Hz and 20 kHz are artefacts generated by internal range-switching

The THD at 10 kHz with a 600 Ω load is 0.0025% for shunt and 0.009% for series feedback; there is significant CM distortion in the input stage, which is almost certainly coming from the JFETs (I appreciate the output levels are not the same but I think this only accounts for a small part of the THD difference). Far from adding magical properties to the input stage, the JFETs seem to be just making it worse.

Opamps surveyed: JFET input types

Opamps with JFET inputs tend to have higher voltage noise and lower current noise than BJT input types, and therefore give a better noise performance with high source resistances. Their very low bias currents often allow circuitry to be simplified.

The TL072 opamp

The TL072 is one of the most popular opamps, having very high-impedance inputs, with effectively zero bias and offset currents. The JFET input devices give their best noise performance at medium impedances, in the range 1 kΩ–10 kΩ. It has a modest power consumption at typically 1.4 mA per opamp section, which is significantly less than the 5532. The slew rate is higher than for the 5532, at 13 V/µs against 9 V/µs. The TL072 is a dual opamp. There is a single version called the TL071 which has offset null pins.

However, the TL072 is not THD free in the way the 5532 is. In audio usage, distortion depends primarily upon how heavily the output is loaded. The maximum loading is a trade-off between quality and circuit economy, and I would put 2 kΩ as the lower limit. This opamp is not the first choice for audio use unless the near-zero bias currents (which allow circuit economies by making blocking capacitors unnecessary), the low price, or the modest power consumption are dominant factors.

It is a quirk of this device that the input common-mode range does not extend all the way between the rails. If the common mode voltage gets to within a couple of volts of the V– rail, the opamp suffers phase reversal and the inputs swap their polarities. There may be really horrible clipping, where the output hits the bottom rail and then shoots up to hit the top one, or the stage may simply latch up until the power is turned off.

TL072s are relatively relaxed about supply rail decoupling, though they will sometimes show very visible oscillation if they are at the end of long thin supply tracks. One or two rail-to-rail decoupling capacitors (e.g. 100 nF) per few centimetres is usually sufficient to deal with this, but normal practice is to not take chances, and allow one capacitor per package as with other opamps.

Because of common-mode distortion, a TL072 in shunt configuration is always more linear. In particular compare the results for 3k3 load in Figures 4.32 and 4.33. At heavier loadings the difference is barely visible because most of the distortion is coming from the output stage.

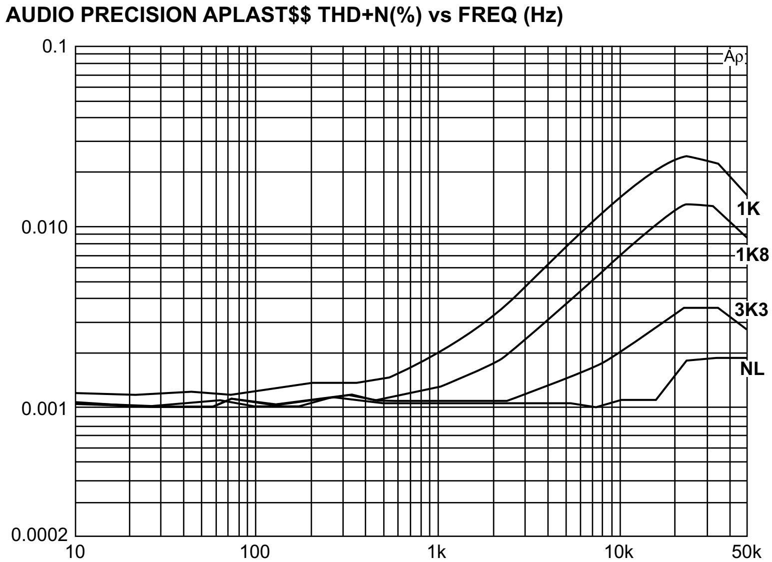

Figure 4.32: Distortion versus loading for the TL072, with various loads. Shunt feedback configuration eliminates CM input distortion. Output level 3 Vrms, gain 3.2×, rails ±15 V. No output load except for the feedback resistor. The no-load plot is indistinguishable from that of the testgear alone. Distortion always gets worse as the loading increases. This factor together with the closed-loop NFB factor determine the THD

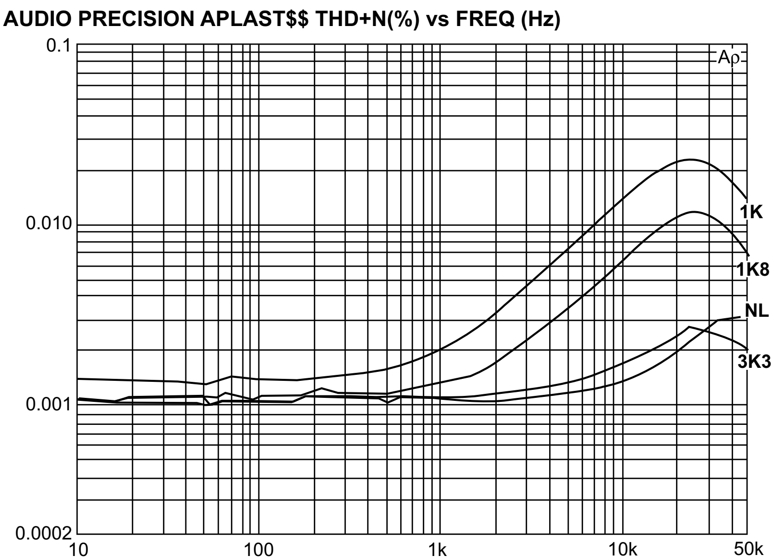

Figure 4.33: Distortion versus loading for the TL072, with various loads. Series feedback configuration Output level 3 Vrms, gain 3.2×, rails ±15 V. Distortion at 10 kHz with no load is 0.0015% compared with 0.0010% for the shunt configuration. This is due to the 1 Vrms CM signal on the inputs

TL072/71 opamps are prone to HF oscillation if faced with significant capacitance to ground on the output pin; this is particularly likely when they are used as unity-gain buffers with 100% feedback. A few inches of track can sometimes be enough. This can be cured by an isolating resistor, in the 47 to 75 Ω range, in series with the output, placed at the opamp end of the track.

The TL052 opamp

The TL052 from Texas Instruments was designed to be an enhancement of the TL072, and so is naturally compared with it. Most of the improvements are in the DC specifications. The offset voltage is 0.65 mV typical, 1.5 mV max, compared with the TL072's 3 mV typical, 10 mV max. It has half the bias current of the TL072. This is very praiseworthy, but rarely of much relevance to audio.

The distortion however *is* important, and this is worse rather than better. THD-performance is rather disappointing. The unloaded THD is low, as shown in Figure 4.34, in series feedback mode. As usual, practical distortion depends very much on how heavily the output is loaded. Figure 4.35 below shows that it deteriorates badly for loads of less than 4k7.

The slew rate is higher than for the TL072, (18 V/µs against 13 V/µs) but the lower figure is more than adequate for a full-range output at 20 kHz, so this enhancement is of limited interest. The power consumption is higher, typically 2.3 mA per opamp section, which is

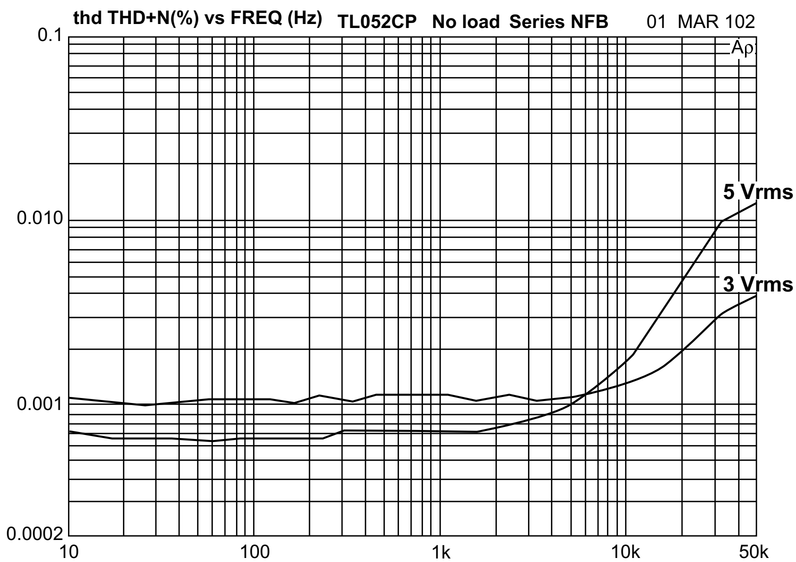

Figure 4.34: Distortion versus frequency at two output levels for the TL052CP, with no load. Series feedback

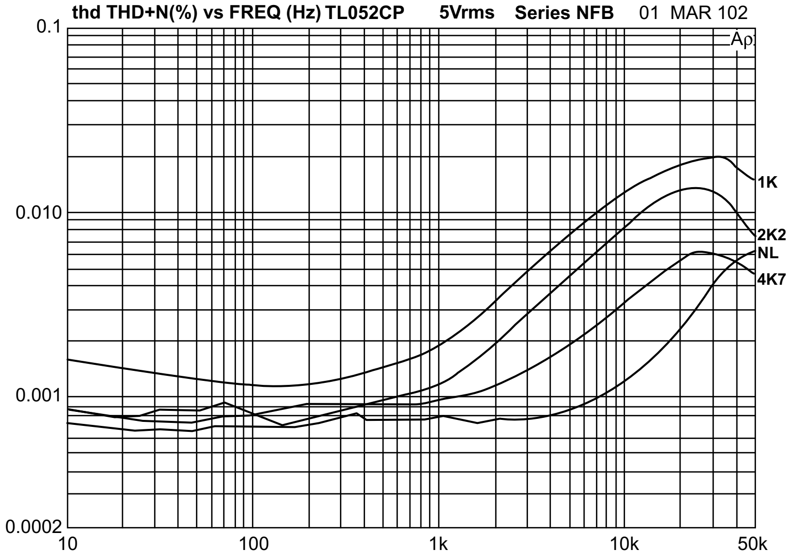

Figure 4.35: Distortion of the TL052 at 5 Vrms output with various loads. At 1 kΩ and 2k2 loading the residual is all crossover distortion at 1 kHz. Gain 3.2×, non-inverting. Series feedback

almost twice that of the TL072. Like the TL072, it is relatively relaxed about supply rail decoupling. In 2009 the TL052 cost at least twice as much as the TL072.

The OPA2134 opamp

The OPA2134 is a Burr-Brown product, the dual version of the OPA134. The manufacturer claims it has superior sound quality, due to its JFET input stage. Regrettably, but not surprisingly, no evidence is given to back up this assertion. The input noise voltage is $8\mathrm{nV}/\sqrt{\mathrm{Hz}}$, almost twice that of the 5532. The slew rate is typically ±20 V/μs, which is ample. It does not appear to be optimised for DC precision, the typical offset voltage being ±1 mV, but this is usually good enough for audio work. I have used it many times as a DC servo in power amplifiers, the low bias currents allowing high resistor values and correspondingly small capacitors.

The OPA2134 does not show phase-reversal anywhere in the common-mode range, which immediately marks it as superior to the TL072.

The two THD plots in Figures 4.36 and 4.37 show the device working at a gain of three times in both shunt and series feedback modes. It is obvious that a problem emerges in the series plot, where the THD is higher by about three times at 5 Vrms and 10 kHz. This distortion increases with level, which immediately suggests common-mode distortion in the input stage. Distortion increases with even moderate loading; see Figure 4.38.

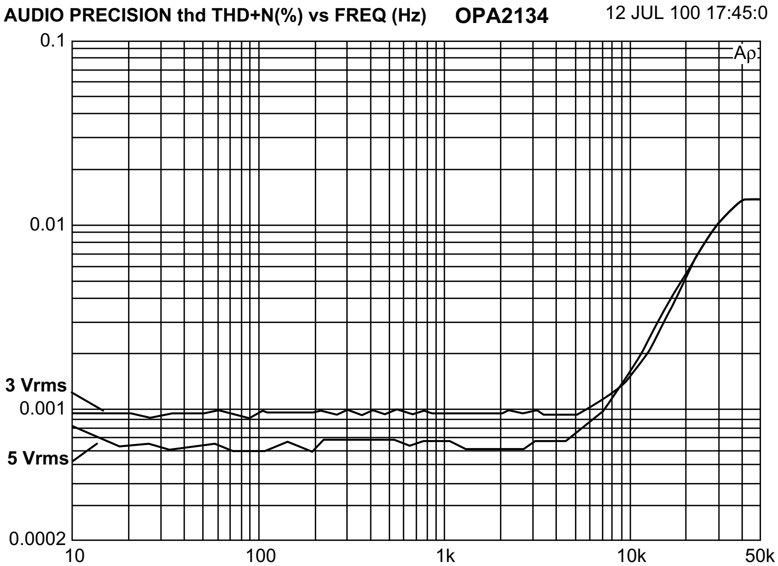

Figure 4.36: The OPA2134 working in shunt feedback mode. The THD is below the noise until frequency reaches 10 kHz; it appears to be lower at 5 Vrms simply because the noise floor is relatively lower

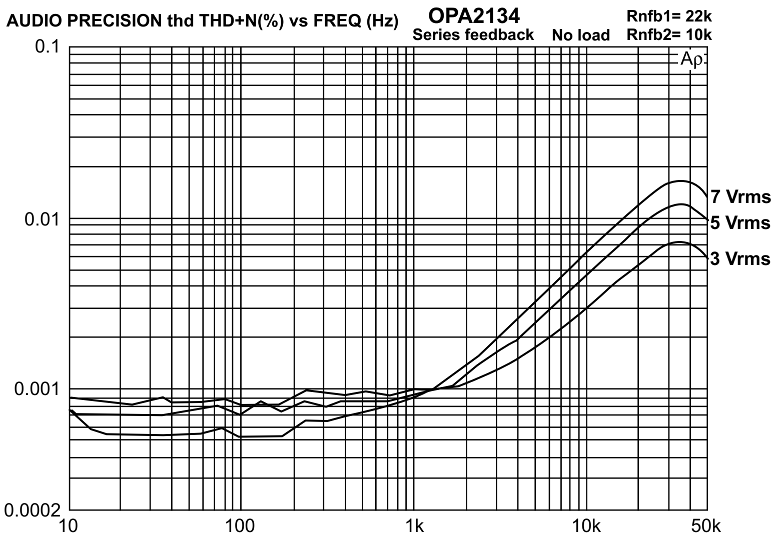

Figure 4.37: The OPA 2134 in series feedback mode. Note much higher distortion at HF

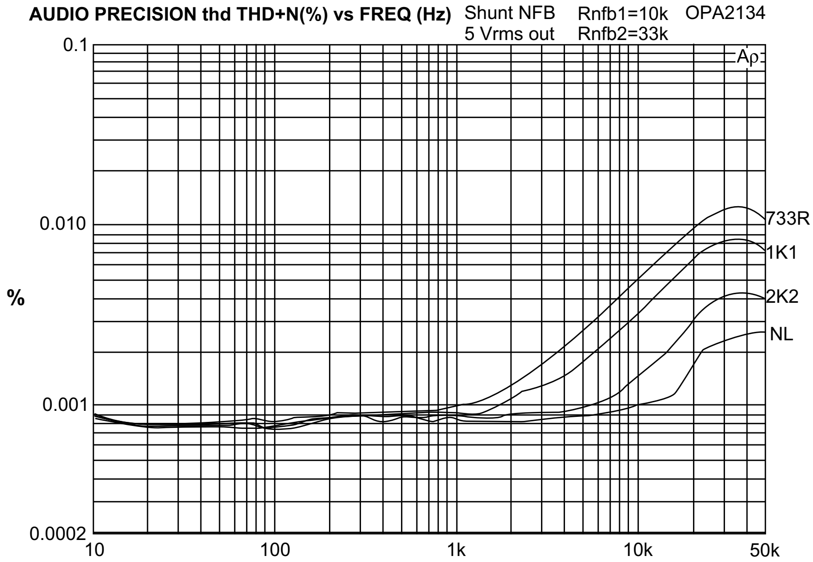

Figure 4.38: The OPA 2134 in shunt feedback mode (to remove input CM distortion) and with varying loads on the output. As usual, more loading makes linearity worse. 5 Vrms out, gain = 3.3×

This is a relatively modern and sophisticated opamp. When you need JFET inputs (usually because significant input bias currents would be a problem) this definitely beats the TL072; it is, however, four to five times more expensive.

The OPA604 opamp

The OPA604 from Burr-Brown is a single JFET-input opamp which has been specially designed to give low distortion. The simplified internal circuit diagram in the data sheet includes an enigmatic box intriguingly labelled 'Distortion Rejection Circuitry'. This apparently 'linearizes the open-loop response and increases voltage gain' but no details as to how are given; whatever is in there appears to have been patented so it ought to be possible to track it down. However, despite this, the distortion is not very low even with no load, (see Figure 4.39) and is markedly inferior to the 5532. The OPA604 is not optimised for DC precision, the typical offset voltage being ±1 mV. The OPA2604 is the dual version, which omits the offset null pins.

The data sheet includes a discussion that attempts to show that JFET inputs produce a more pleasant type of distortion than BJT inputs. This unaccountably omits the fact that the much higher transconductance of BJTs means that they can be linearised by emitter degeneration so that they produce far less distortion of whatever type than a JFET input [7]. Given that the

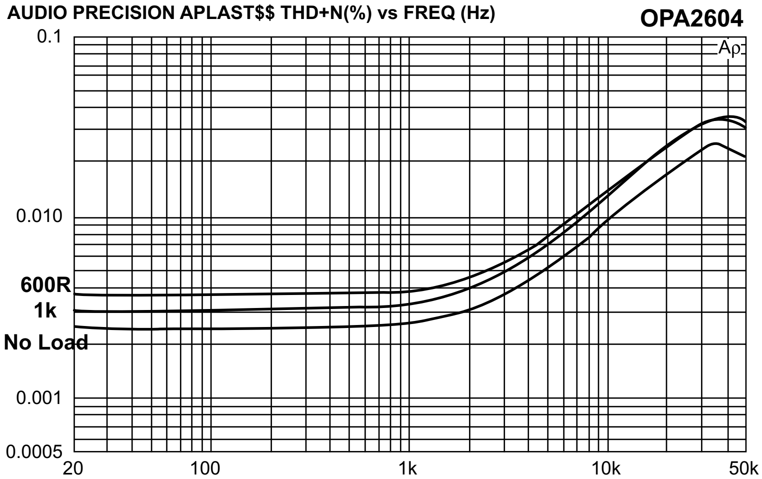

Figure 4.39: An OP2604 driving various loads at 7.75 Vrms. Series feedback, gain = 3.2×

OPA604 costs five times as much as a 5532, it is not very clear under what circumstances this opamp would be a good choice.

The OPA627 opamp

The OPA627 from Burr-Brown is a laser-trimmed JFET input opamp with excellent DC precision; the input offset voltage being typically ±100 μV. The distortion is very low, even into a 600 Ω load, though it is increased by the usual common-mode distortion when series feedback is used.

The OP627 is a single opamp and no dual version is available. The OPA637 is a decompensated version only stable for closed-loop gains of five or more. This opamp makes a brilliant DC servo for power amplifiers, if you can afford it; it costs about 50 times as much as a 5532, which is 100 times more per opamp section, and about 20 times more per opamp than the OPA2134, which is my usual choice for DC servo work.

The current noise i_n is very low, the lowest of any opamp examined in this book, apparently due to the use of Difet (dielectrically isolated JFET) input devices, and so it will give a good noise performance with high source resistances. Voltage noise is also very respectable at 5.2 nV/$\sqrt{\text{Hz}}$, only fractionally more than the 5532.

The series feedback case barely has more distortion than the shunt one, and only at the extreme HF end. It appears that the Difet input technology also works well to prevent input non-linearity and CM distortion. See Figures 4.40 and 4.41.

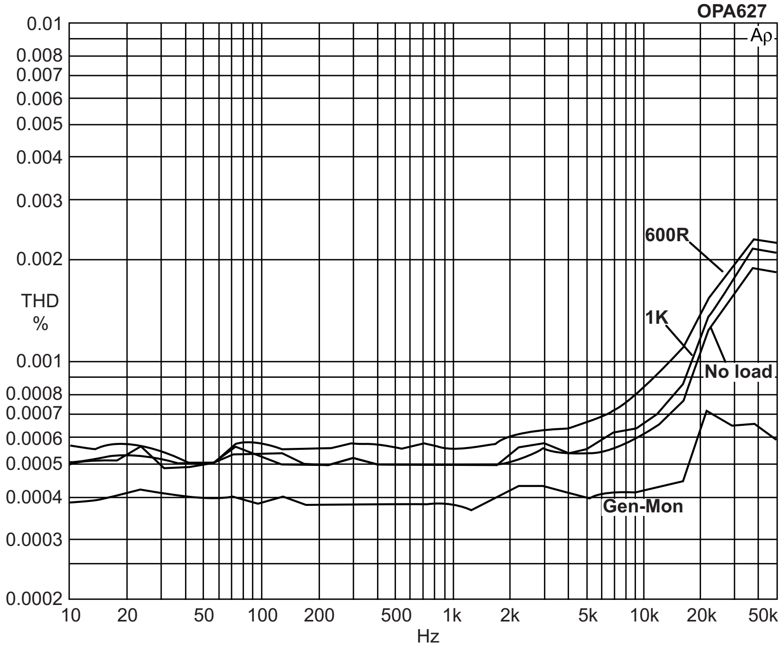

Figure 4.40: OP627 driving the usual loads at 5 Vrms. Series feedback, gain = 3.2×. Gen-mon is the testgear output

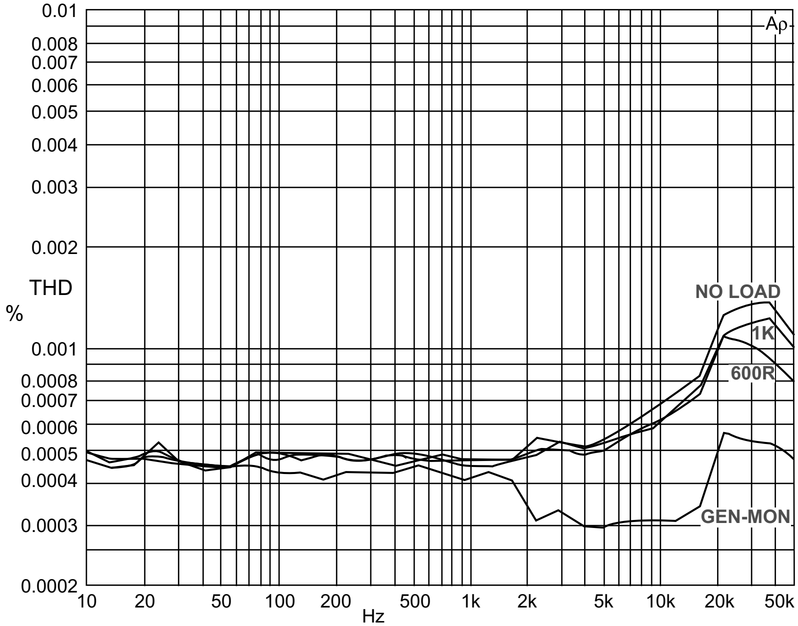

Figure 4.41: OP627 driving the usual loads at 5 Vrms. Shunt feedback, gain = 2.2× but noise gain = 3.2×. Gen-mon trace shows the distortion produced by the AP System 2 generator alone

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[5] Jung, W. *Opamp Applications Handbook*, Chapter 5, p. 399.
[6] Huijsing, J. H. *Operational Amplifiers: Theory and Design* (Kluwer Academic 2001), pp. 300–302.
[7] Self, D. *Audio Power Amplifier Design* 6th edn (Newnes 2013), p. 500.

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CHAPTER 5

Opamps for low voltages

High fidelity from low voltages

Nowadays there is considerable interest in audio circuitry that can run from +5 V, typically obtained from a USB port. It is expected that reasonable quality will be obtained; what might be called 'five-volt fidelity'. Since there is only a single +5 V rail available, the maximum possible peak voltage is clearly ±2.5 V, equivalent to 1.77 Vrms or +7.16 dBu (this calculation obviously ignores opamp output saturation voltages). A slew rate of only 0.31 V/µs is enough to give this maximum output at 20 kHz.

When the usual ±17 V rails are used, you get a maximum level of 12.02 Vrms or +23.81 dBu. Five-volt audio gives a maximum level that is 16.6 dB less, so the dynamic range is reduced by the same amount. If the same relative headroom is required, the nominal signal level will have to be correspondingly reduced by 16.6 dB, and the 5 V signal path will have a worse signal/noise ratio by 16.6 dB. Often a compromise between these two extremes is more appropriate. In this technology it is especially important to make sure that what dynamic range does exist is not compromised by deficiencies in design.

While +5 V was for many years the standard voltage for running digital circuitry, for some time now +3.3 V has been used to reduce power consumption and make larger processor ICs feasible. It is perhaps not very likely that applications will arise requiring quality audio circuitry to run off such a low supply, but the existence of opamps in Table 5.1 that are rated to operate down to 2.7 V and even 2.2 V shows that opamp manufacturers have this sort of thing in mind.

With a single +3.3 V rail, the maximum possible peak voltage is ±1.65 V, equivalent to 1.17 Vrms or +3.55 dBu, ignoring output saturation voltages. This is 3.61 dB less than for a +5 V rail, so the dynamic range issue is that much more acute. A slew rate of only 0.21 V/µs is enough to give the maximum output at 20 kHz.

The issues of +5 V operation are considered first, followed by those of +3.3 V operation.

TABLE 5.1 Low-voltage opamps, ranked in order of voltage noise density

Device	Supply voltage (V)	Voltage noise 1 kHz (nV/√Hz)	Current noise	CMRR type (dB)	Slew rate (V/μs)	Format	Price ratio 100-off
AD8022	5–24	2.3 nV/√Hz	1 pA/√Hz	−95	50	Dual	1.72
LM4562	5–34	2.7 nV/√Hz	1.6 pA/√Hz	−120	20	Dual	1.25
AD8656	2.7–5.5	4 nV/√Hz	No spec	−100	11	Single	1.00
AD8397	3–24	4.5 nV/√Hz	1.5 pA/√Hz	−90	53	Dual	1.69
OPA2365	2.2–5.5	4.5 nV/√Hz	4 fA/√Hz	−100	25	Dual	2.01
AD8066	5–24	7 nV/√Hz	0.6 fA/√Hz	−91	180	Dual	1.83
AD8616	2.7–5	10 nV/√Hz	0.05 pA/√Hz	−100	12	Dual	1.03
AD826	5–36	15 nV/√Hz	1.5 pA/√Hz	−80	350	Dual	2.50
AD823	3–36	16 nV/√Hz	1 fA/√Hz	−75	22	Dual	2.19

Prices are per package, not per opamp section. This puts the single AD8656 at a serious price disadvantage. Prices as at 2013.

Running opamps from a single +5 V supply rail

The vast majority of audio circuitry runs from dual supply rails; this has been the case ever since opamps became common in audio in the 1970s, and it was established that electrolytic capacitors could work satisfactorily and reliably without DC bias on them. When low voltages are used there is typically only one supply rail available, and it is necessary to bias the opamp outputs so they are halfway between the supply rail and ground. This is most economically done by an RC ‘V/2 generator’ which supplies bias to all the stages. This is sometimes called a ‘half-rail generator’. There is no necessity that the bias point be *exactly* half of the supply rail. Some opamps clip earlier in one direction than the other, and a slight adjustment up or down of the half-rail may allow a larger symmetrical output swing. Having all the signal-carrying parts of the circuit displaced from ground means that DC blocking capacitors are frequently required, to keep DC out of volume controls and switches, counteracting any parts saving there may be in the power supply due to the single rail.

Figure 5.1 shows series and shunt feedback stages biased by a common V/2 generator R10, R11, C10. The shunt stage has a voltage gain of 2.2 times, set by R2 and R3. Capacitor C1 performs DC-blocking at the input; R1 ensures that the DC level at the input cannot float up due to leakage through this capacitor. Likewise a blocking capacitor C2 is required to prevent DC flowing through the load.

The series feedback stage has a voltage gain of 3.2 times, which means it works at the same noise gain as the shunt stage (see Chapter 22 for an explanation of noise gain) so comparative tests can be made in which the only difference is the common-mode voltage on the series

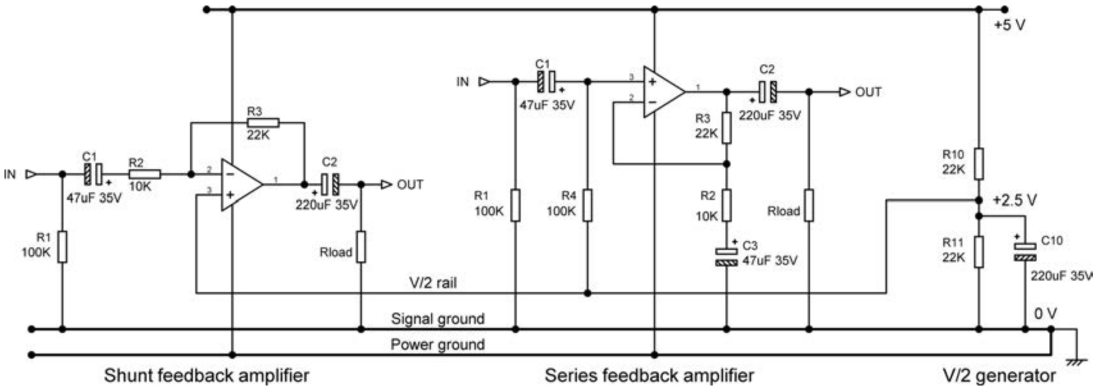

Figure 5.1: Opamps running from a single supply rail, with a half-rail biasing generator shared between the two stages

stage opamp input pins. The series circuit requires an extra blocking capacitor C3 to prevent DC flowing through the feedback network, and to reduce the gain to unity at zero frequency. An extra resistor R4 is needed to bias the stage input.

The V/2 generator typically uses a rather large capacitor to filter out disturbances on the supply rail. It is economical to share this with other stages, as shown in Figure 5.1. The low impedance of the filter capacitor should prevent interaction between stages, if they are not of high gain, but isolation will tend to fall with frequency. If a good crosstalk performance at LF is required in stereo circuitry it may be necessary to use separate V/2 generators for each channel.

One of the great advantages of dual-rail operation is that the opamp supply currents, which to some extent have the same half-wave rectified nature as those in a Class-B power amplifier, are inherently kept out of the signal ground. With single-rail operation the negative supply pin of an opamp will be at 0 V, but any temptation to connect it to signal ground must be stoutly resisted, or the half-wave rectified currents flowing will cause serious distortion. It is essential to provide a separate power ground, as shown in Figure 5.1, which only joins the signal ground back at the power supply.

Opamps for 5 V operation

Table 5.1 shows some candidates for low-voltage operation. The table has been restricted to those opamps capable of working from +5 V or lower. This therefore excludes the ubiquitous 5532 (with a minimum of 6 V) though it has been tested, with results given in the text. Those opamps that have been tested are shown in bold.

The NE5532 in +5 V operation

When an attempt is made to run the 5532 from a +5 V rail, the maximum output (with no load beyond the shunt feedback network) is only 845 mVrms (at 1% THD) compared with the 1.77 Vrms theoretically available. This is only 48% of the possible voltage swing – less than half – and the 5532 output stage is clearly not performing well at a task it was never intended to do. Shunt feedback was used to exclude any common-mode distortion in the opamp input stage, with a gain of 2.2 times, as in Figure 5.1. Distortion is high, even at lower levels such as 700 mVrms, but falls below the relatively high noise level at 500 mV out. Perhaps most importantly, using any IC outside of its specifications is very risky because one batch may work acceptably but another not at all.

The LM4562 in +5 V operation

The much newer LM4562 opamp has a supply specification reaching lower down to 5 V, so we expect better low-voltage performance. We get it too; the maximum output with no load is 1.16 Vrms (1% THD), a more reasonable 65% of the possible voltage swing, but this is still very inefficient compared with the same opamp working from ±17 V rails.

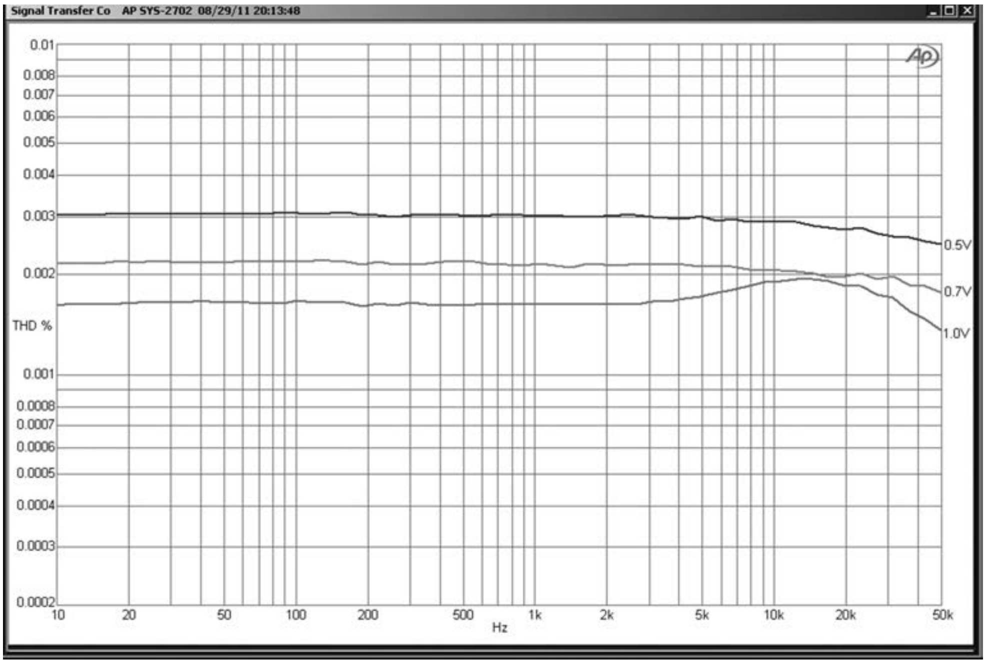

Figure 5.2: LM4562 THD with +5 V supply for output levels of 0.5 V, 0.7 V and 1.0 Vrms. No external load

The distortion performance for three output levels, with shunt feedback and a gain of 2.2 times as in Figure 5.1 is shown in Figure 5.2. The flat traces simply represent noise and no distortion is visible in the THD residual except around 10–20 kHz for a 1.0 Vrms output. You will note that the relative noise levels, and hence the minimum measurable THD levels, are considerably higher than for opamps on ±17 V rails

The LM4562 is thus a possibility for +5 V operation, but the 3.7 dB loss in maximum output caused by that 65% when the dynamic range is already being squeezed is very unwelcome. There is also the point that the opamp is working right on the lower limit of its supply specs, which is not ideal. I found some evidence that the noise performance was impaired compared with ±17 V rails, with a noticeable incidence of burst noise. The NE5532 is at least cheap, but the LM4562 is still relatively expensive. There may therefore be no economic disadvantage in using specialised low-voltage opamps which are also expensive but designed for the job. We will examine some of them now.

The AD8022 in +5 V operation

The AD8022 from Analog devices looks like a possible candidate, having an exceptionally low input noise density of 2.3 nV/$\sqrt{\text{Hz}}$ that should allow maximisation of the dynamic range. It is a bipolar opamp with a wide supply range from +5 V to +24 V for single-rail operation, or ±2.5 V to ±12 V for dual rails; we would therefore still be operating at the lower limit of supply voltage. It is fabricated using a high-voltage bipolar process called XFCB.

Firstly, the maximum output is 1.18 Vrms (no load, 1% THD), 67% of the available voltage swing; this is only a tiny improvement on the LM4562. The distortion performance, using shunt feedback and a gain of 2.2 times as in Figure 5.1, is not inspiring. According to the data sheet we can expect second harmonic at −80 dB, (0.01%) and third harmonic at −90 dB, (0.003%) with a 0.7 Vrms output into a 500 Ω load, flat with frequency. Figure 5.3 shows the measured results for various output levels; they are markedly worse than the data sheet suggests, for reasons unknown. Note that we have changed the THD scale radically to accommodate the much higher levels, and the flat traces here really do represent distortion rather than noise. Several samples were tested with near-identical results.

These distortion levels are too high for any sort of quality audio, and the AD8022 will not be a good choice in most cases. No tests were therefore done on how distortion varies with output loading. However, for some applications it may be worth bearing in mind that its input noise density is exceptionally low at 2.3 nV/$\sqrt{\text{Hz}}$.

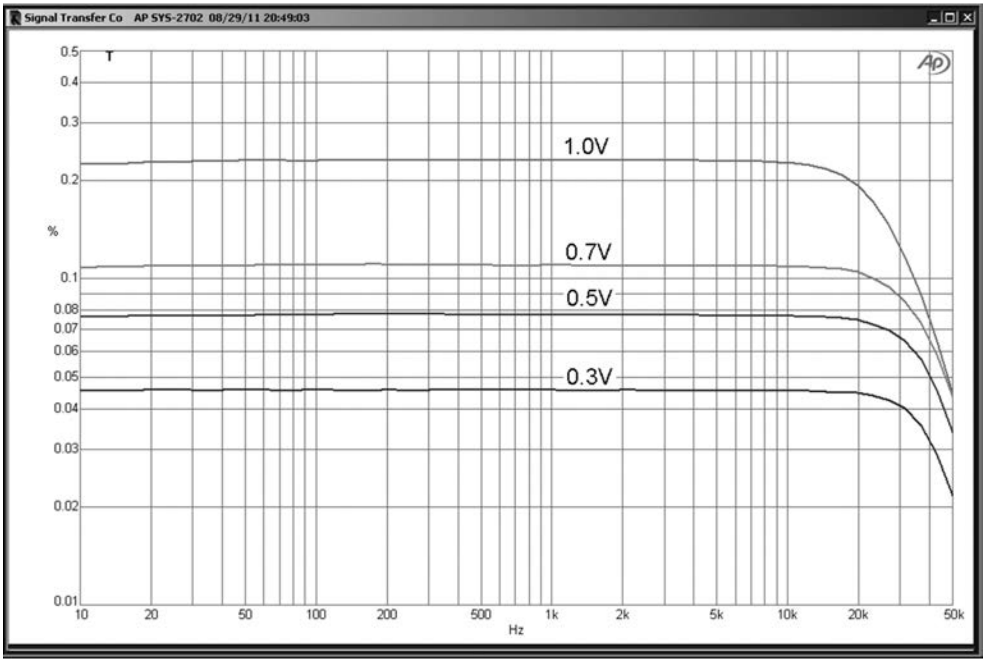

Figure 5.3: AD8022 THD with +5 V supply for output levels of 0.3 V, 0.5 V, 0.7 V and 1.0 Vrms. Shunt feedback, no external load. 80 kHz bandwidth

The AD8397 in +5 V operation

The AD8397 is another promising possibility, because it claims a very voltage-efficient output stage and good load-driving capabilities. The input noise density is somewhat higher at 4.5 $nV/\sqrt{Hz}$, slightly better than a 5532 and slightly worse than a 5534. It is a bipolar opamp with a wide supply range from 3 V to 24 V (for single-rail operation) or ±1.5 V to ±12 V for dual rails; this means that at +5 V we are not operating at the lower supply voltage limit, which increases our confidence. It is fabricated using a complementary bipolar high voltage process called XFCB-HV.

The first good result is that the AD3897 really is voltage-efficient. The maximum output with no external load is 1.83 Vrms at 1% THD. This is actually *more* than the theoretical output of 1.77 Vrms because the 1% THD criterion relies on some clipping to create the distortion. If we instead use as our output criterion the first hint of disturbance on the THD residual, we get exactly 1.77 Vrms, which is 100% voltage efficiency! Adding a 100 Ω load only reduces this to 1.72 Vrms.

The measured THD results, using shunt feedback and a gain of 2.2 times as in Figure 5.1, are shown in Figure 5.4 for various output levels (note that the noise gain is higher at 3.2×). The flat parts of the traces represent noise only, with detectable distortion only appearing above

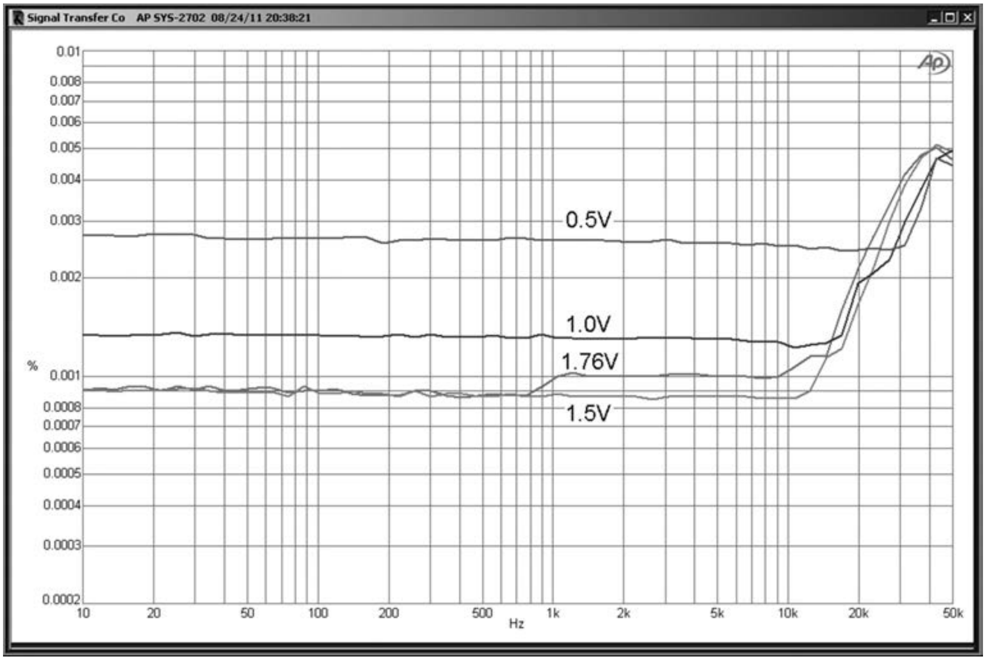

Figure 5.4: AD8397 THD with +5 V supply. Output levels of 0.5 V, 1.0 V, 1.5 V and 1.76 Vrms. Shunt feedback, no external load

10 kHz, except at 1.76 Vrms out. The measured relative noise levels obviously fall as the output voltage increases.

The unloaded distortion results are very encouraging, but how are they affected by an external load on the output? Hardly at all, as Figure 5.5 shows; the output level is close to the maximum at 1.5 Vrms, and the external loads go down to 220 Ω, well below the minimum 5532 load of 500 Ω. Shunt feedback was used.

The AD8397 data sheet refers to loads down to an impressively low 25 Ω, so we want to explore the linearity with loading a bit further. Figure 5.6 shows the measured THD results with No Load (NL) and external loads of 220 Ω, 150 Ω and 100 Ω, again at an output of 1.5 Vrms. Not until the load falls to 150 Ω do we start to see significant distortion appearing above 1 kHz, and this promises that this opamp will work well with low-impedance designs that will minimise noise and optimise the dynamic range.

So far, so good. The next thing to investigate is the distortion performance using series feedback, which puts a significant common-mode signal on the input pins. The test circuit used is the series stage in Figure 5.1, which has a gain of 3.2 times, but the same noise gain (3.2 times) as the shunt circuit. Figure 5.7 shows the measured THD results with No Load (NL)

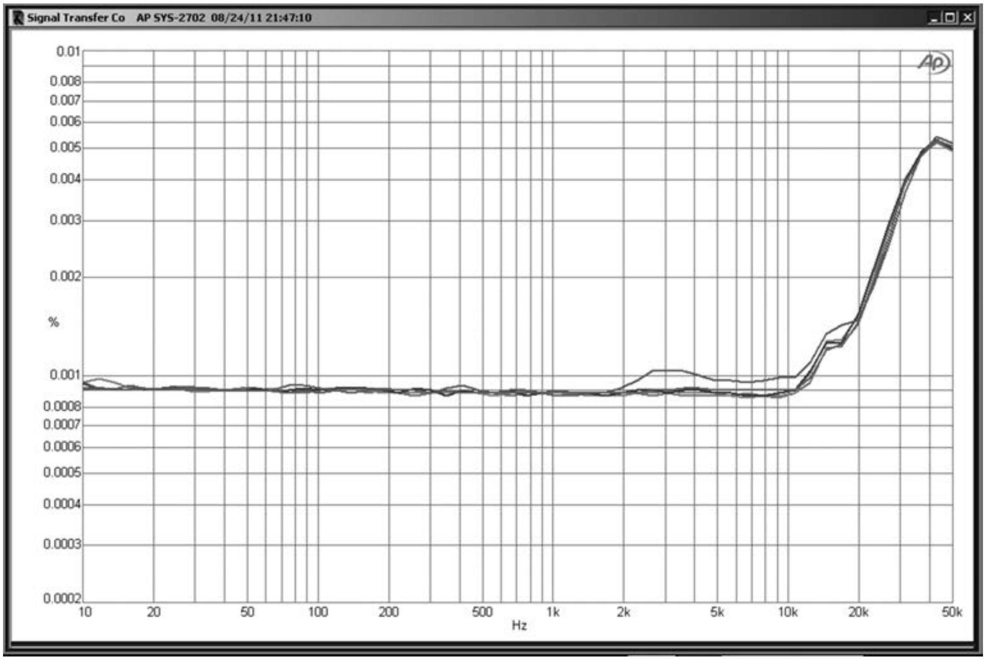

Figure 5.5: AD8397 THD with +5 V supply. External loads from 680 Ω down to 220 Ω at an output level of 1.5 Vrms. Shunt feedback

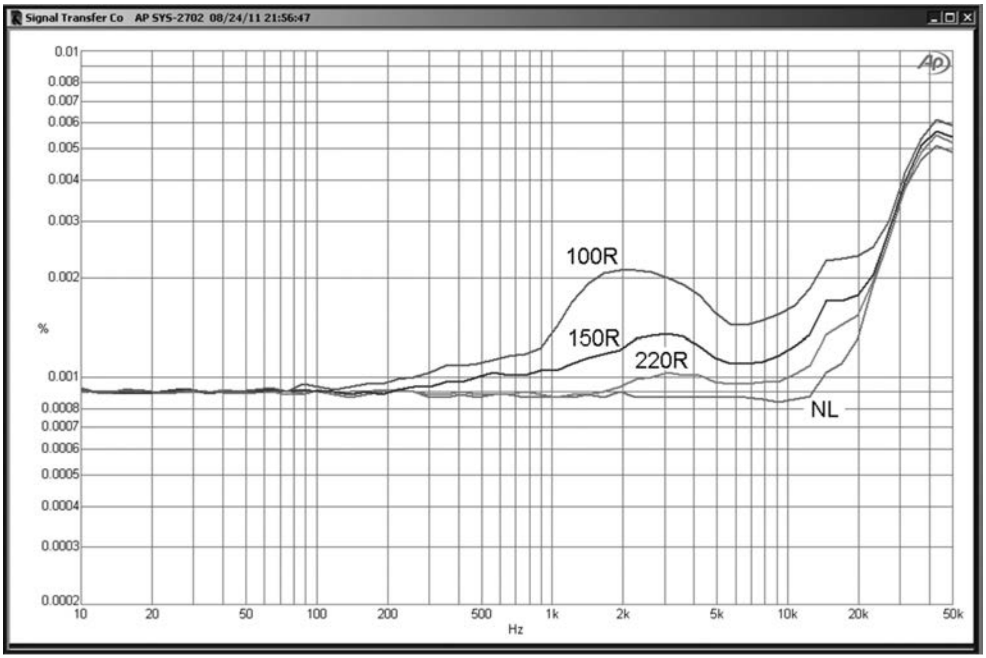

Figure 5.6: AD8397 THD with +5 V supply. With No Load (NL) and external loads of 220 Ω, 150 Ω and 100 Ω at an output level of 1.5 Vrms. Shunt feedback, gain = 2.2×

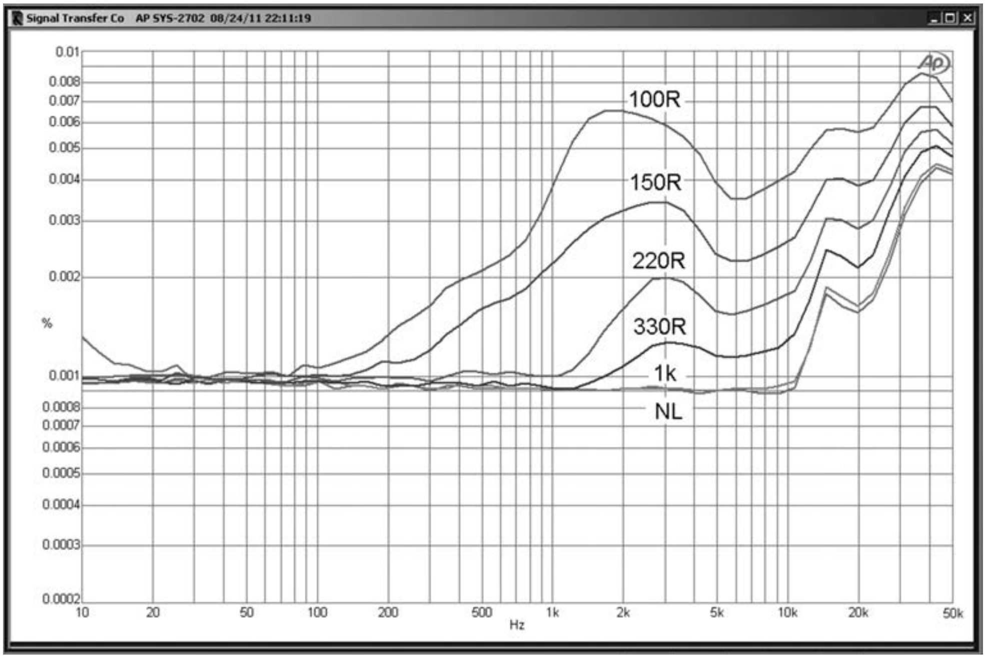

Figure 5.7: AD8397 THD with +5 V supply. With No Load (NL), and external loads of 330 Ω, 220 Ω, 150 Ω and 100 Ω at an output level of 1.5 Vrms. Series feedback, gain = 3.2×

and external loads of 220 Ω, 150 Ω and 100 Ω, again at an output of 1.5 Vrms. The distortion behaviour is clearly worse with series feedback for loads of 220 Ω or lower; care will be needed if low impedances are to be driven from a series feedback stage using an AD8397.

The worst-case for common-mode voltage is the voltage-follower configuration, where the full output voltage is present on the input terminals, and this exposes a limitation of the AD8397. The output stage does a magnificent job of using the whole available voltage swing, but the input stage does not quite have a full rail-to-rail capability. This means that voltage-follower stages, where the input and feedback signals are equal to the output signal, have a limitation on output level set by the input stage rather than the output stage. The distortion performance is also significantly worse than for the case of series feedback with a gain of 3.2 times. Figure 5.8 demonstrates that distortion rises quickly as the signal level exceeds 0.9 Vrms. If you want to make use of the full output swing then voltage-followers should be avoided. A relatively small amount of voltage gain – for example, 1.1 times – will reduce the signal levels on the input pins and allow the full output swing with low distortion.

Common mode distortion is worsened when the opamp is fed from a significant source resistance. Our final test is to drive a voltage-follower through a 4k7 resistance. Figure 5.9 shows that distortion is much increased compared with Figure 5.8. We had about 0.0025% THD with a 1.05 Vrms signal, but adding the 4k7 in series with the input has increased that to 0.007%.

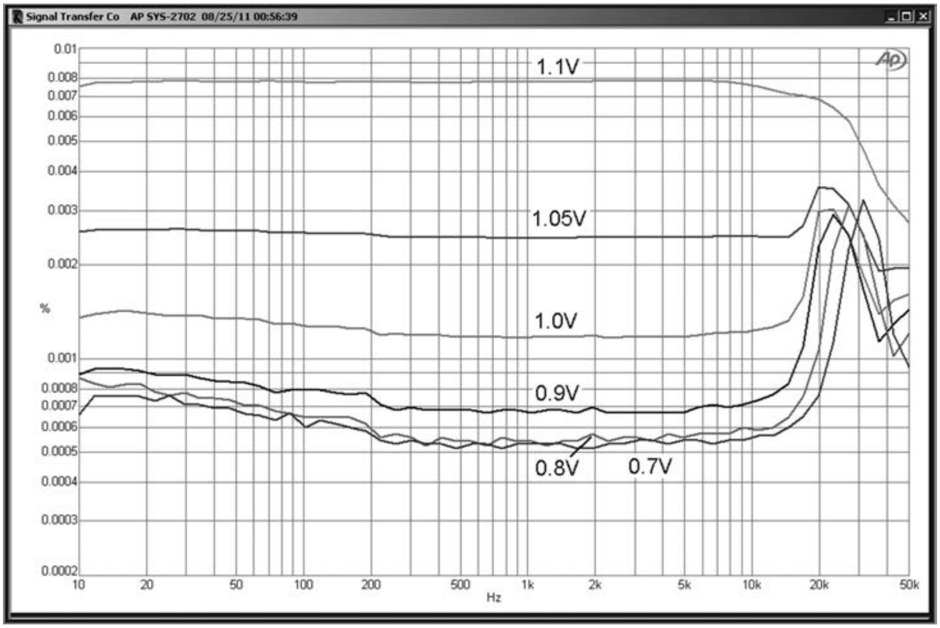

Figure 5.8: AD8397 voltage-follower THD with +5 V supply. No Load (NL) and input/output levels of 0.7 V, 0.8 V, 0.9 V, 1.0 V, 1.05 V, and 1.1 Vrms

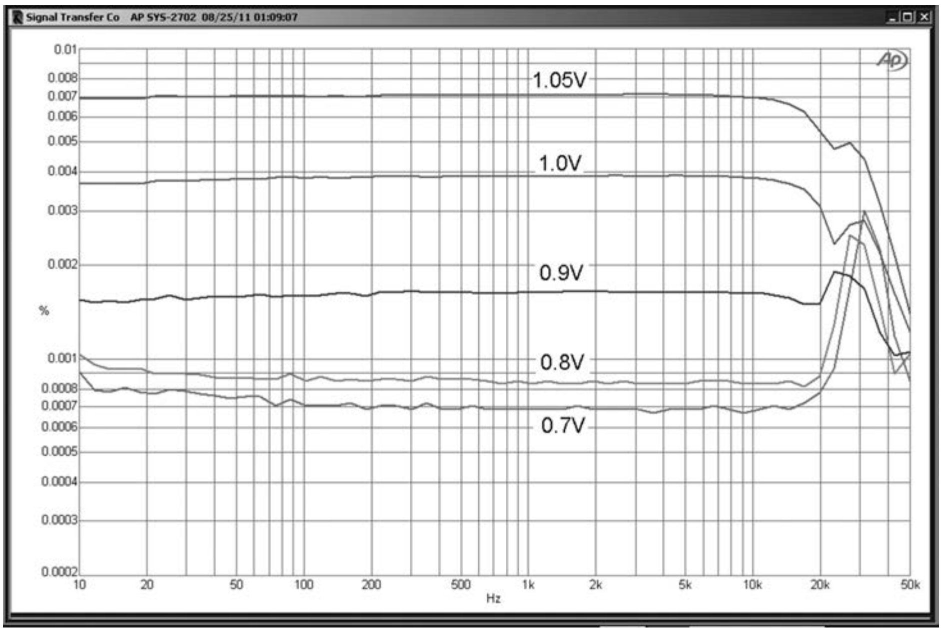

Figure 5.9: AD8397 voltage-follower THD with 4k7 series input resistor. +5V supply, No Load (NL) and input/output levels of 0.7 V, 0.8 V, 0.9 V, 1.0 V, and 1.05 Vrms

To conclude, the AD8397 is a useful opamp for +5 V operation; it is the best I have evaluated so far. It is however necessary to keep an eye on the effects of common-mode distortion and source resistance, and be aware of the voltage limits of the input stage.

Opamps for 3.3 V single-rail operation

Firstly, forget about the NE5532. Its maximum output on a +3.3 V rail (with appropriate half-rail biasing) is only 240 mVrms (1% THD). This is a meagre 20% of the available voltage swing and is clearly very inefficient. There is considerable distortion and the residual shows nasty crossover spikes of the sort associated with under-biased power amplifiers. Finally, using any IC so far outside of its specifications is very dangerous indeed – one batch may work acceptably but another batch not at all.

The LM4562 works rather better from +3.3 V, giving 530 mVrms (1% THD). This is a slightly more respectable 45% of the available voltage swing, but the objection to using it outside of its specifications remains. The THD at output levels up to 430 mVrms is around 0.01%, which is not very encouraging.

The AD8397, however, works well at +3.3 V. The maximum output is now 1.18 Vrms (1% THD). The measured THD results, using shunt feedback and a gain of 2.2 times as in Figure 5.1, are shown in Figure 5.10 for differing output levels. The flat parts of the traces represent noise

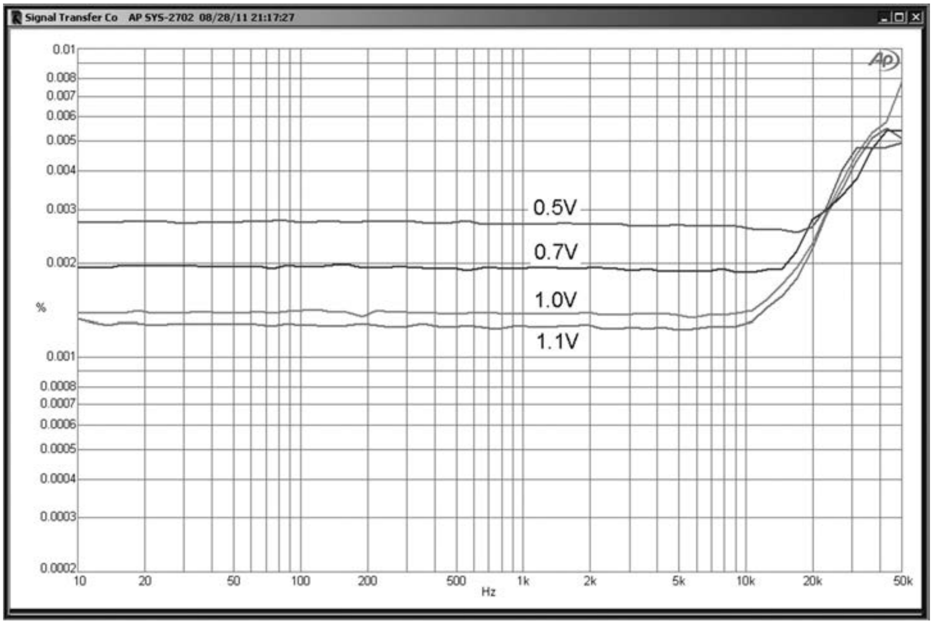

Figure 5.10: AD8397 THD with +3V3 supply. Output levels of 0.5 V, 0.7 V, 1.0 V, and 1.1 Vrms. Shunt feedback, no external load

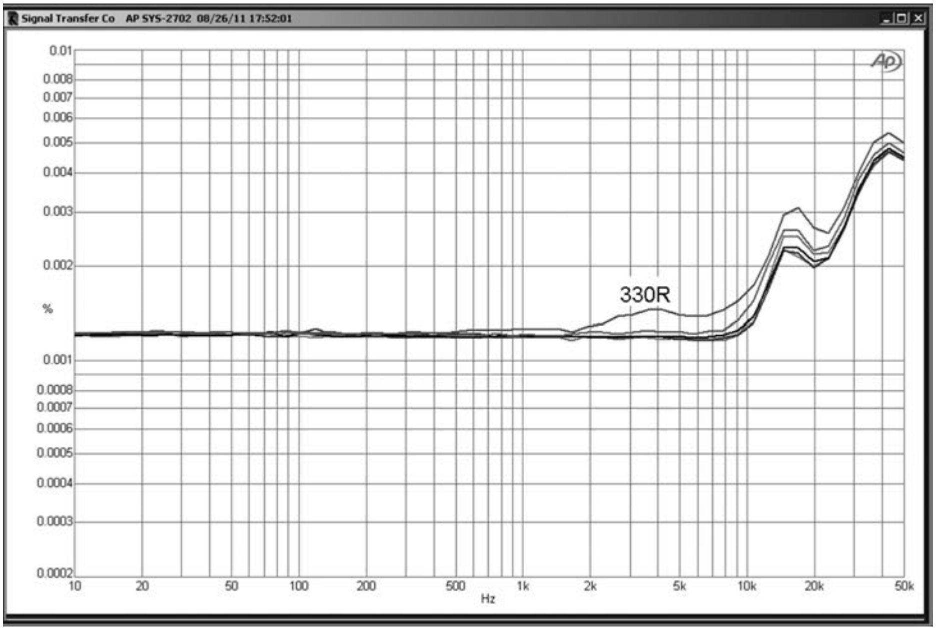

Figure 5.11: AD8397 THD with +3V3 supply. With No Load (NL) and external loads of 4k7, 2k2, 1k, 560 Ω, and 330 Ω at an output level of 1.1 Vrms. Shunt feedback, gain = 2.2×

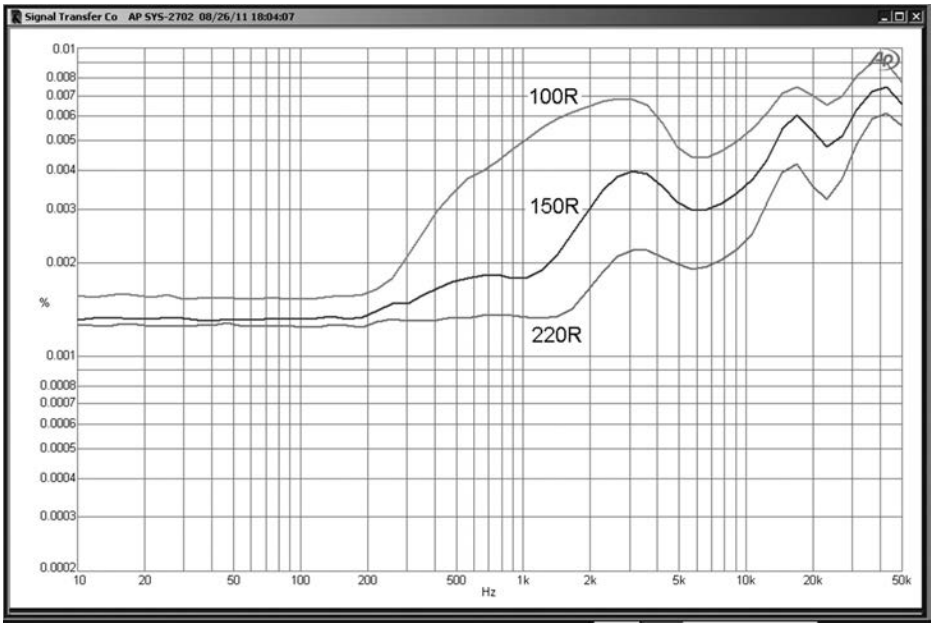

Figure 5.12: AD8397 THD with +3V3 supply. With external loads of 220 Ω, 150 Ω and 100 Ω at an output level of 1.1 Vrms. Shunt feedback, gain = 2.2×

only, with detectable distortion only appearing above 10 kHz. Compare this with the +5 V results in Figure 5.4 above; there is little difference, though the maximum test level is, of course, less.

At +3.3 V the AD8397 still works well with quite heavy loading. Figure 5.11 shows that loads lighter than 560 Ω have very little effect on the distortion performance, while 330 Ω causes detectable distortion only above 2 kHz. Figure 5.12 demonstrates that heavier loads than this lead to significant distortion. The AD8022 was not evaluated for +3.3 V operation as its minimum voltage is +5 V.

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CHAPTER 6

Filters

Introduction

Analogue filter design is an enormous subject, and it is of course quite impossible to cover even its audio aspects in a single chapter. A much, much more detailed account is given in my book *The Design of Active Crossovers*, which gives practical examples of just about every filter type you can think of [1]. There are several standard textbooks on filter theory [2] [3] [4], and there would be no point in trying to create another one here. This chapter instead aims to give information on audio applications not found in the standard textbooks.

Filter design is at the root highly mathematical, and it is no accident that all of the common filter characteristics such as Bessel, Gaussian, Chebyshev, and Legendre are named after mathematicians. A notable exception is the Butterworth characteristic, probably the most popular and useful characteristic of all; Stephen Butterworth was a British engineer [5]. Here however I am going to avoid the complexities of pole and zero placement etc. and concentrate on practical filter designs that can be adapted for different frequencies by simply scaling component values. Most filter textbooks give complicated equations for calculating the amplitude and phase response at any desired frequency. This is less necessary now access can be assumed to a simulator, which will give all the information you could possibly want, much more quickly and efficiently than wrestling with calculations. Free simulator packages can be downloaded. Filters are widely used in different kinds of audio system, and I make frequent references to them in other chapters.

Filters are either passive or active. Passive or LCR filters use only resistors, inductors, and capacitors. Active filters use resistors, capacitors, and gain elements such as opamps; active filter technology is usually adopted with the specific intent of avoiding inductors and their well-known limitations. Nevertheless, there are some applications where LCR filters are essential.

Passive filters

Passive filters do not use active electronics, and this is a crucial advantage in some applications. They are not subject to slew rate limiting, semiconductor non-linearity, or errors due to falling open-loop gain, and this makes them the best technology for roofing filters.

A roofing filter is one that stops out-of-band frequencies before they reach the first stage of electronics, and so prevents RF demodulation and slew limiting. A classic application is in the measurement of Class-D power amplifiers, which emit copious quantities of RF that will greatly upset audio measuring equipment. The answer, as described by Bruce Hofer [6] is a passive LCR roofing filter. There are many excellent text books that describe LCR filter design, such as [2] and [3], and I am not going into it here.

Active filters

Active filters do not normally use inductors as such, though configurations such as gyrators that explicitly model the action of an inductor are sometimes used. The active element need not be an opamp; the Sallen and Key configuration requires only a voltage follower, which in some cases can be a simple BJT emitter-follower. Opamps are usual nowadays, however. The rest of this chapter deals only with active filters.

Low-pass filters

Probably the most common use of low-pass filters is at the output of DACs to remove the high-frequency spurii that remain after oversampling; see Chapter 26 for more on this. They are also used to explicitly define the upper limit of the audio bandwidth in a system at, say, 50 kHz (see the example of a record-cutting amplifier in Chapter 8), though more often this is done by the casual accumulation of a lot of first-order roll-offs in succeeding stages. This is not a duplication of input RF filtering (i.e. roofing filtering) which, as described in Chapter 18, must be passive and positioned before the incoming signals encounter any electronics which can demodulate RF. Low-pass filters are also used in PA systems to protect power amplifiers and loudspeakers against ultrasonic oscillation in the system.

In what might be called The First Age of Vinyl, a fully-equipped preamplifier would certainly have had a switchable low-pass filter, called, with brutal frankness, the 'scratch' filter. This, having a slope of 12 or 18 dB/octave, faster than the tone-control stage, and rolling off at a higher frequency around 5–10 kHz, was aimed at suppressing, or at any rate dulling, record surface noise and the inevitable ticks and clicks; see Chapter 9 for much more on this. Interestingly, preamplifiers today, in The Second Age of Vinyl rarely have this facility, probably because it requires you to face up to the fact that the reproduction of music from mechanical grooves cut into vinyl is really not very satisfactory. Low-pass filters are of course essential to electronic crossovers for loudspeaker systems.

High-pass filters

High-pass filters are widely used. Preamplifiers for vinyl disk usage are commonly fitted with subsonic filtering, often below 10 Hz, to keep disturbances due to record warps and ripples from reaching the loudspeakers. Subsonic noise may affect the linearity of the speaker for the worse as the bass unit is often moved through a substantial part of its mechanical travel; this is particularly true for reflex designs with no cone loading at very low frequencies. If properly designed, this sort of filtering is considered inaudible by most people. In The First Age of Vinyl, preamplifiers sometimes were fitted with a ‘rumble filter’ which began operations at a higher frequency – typically 35 or 40 Hz – to deal with more severe disk problems, and would not be considered inaudible by anyone. Phono subsonic filtering is comprehensively dealt with in Chapter 8 on moving-magnet preamplifiers.

Mixer input channels very often have a switchable high-pass filter at a higher frequency again, usually 100 Hz. This is intended to deal with low-frequency proximity effect with microphones and general environmental rumblings. The slope required to do this effectively is at least 12 dB/octave, putting it outside the capabilities of the EQ section. Some mixer high-pass filters are third-order (18 dB/octave), and fourth-order (24 dB/octave) ones have been used occasionally. Once again, high-pass filters are used in electronic crossovers.

Combined low-pass and high-pass filters

When both subsonic and ultrasonic filters are required they can sometimes be economically combined into one stage using only one opamp, to give audio band definition. Filter combination is usually only practicable when the two filter frequencies are widely separated. There is more on this in Chapter 8.

Bandpass filters

Bandpass filters are principally used in mixing consoles and stand-alone equalisers (see Chapter 15). The Q required rarely exceeds a value of 5, which can be implemented with relatively simple active filters, such as the multiple-feedback type. Higher Qs or independent control of all the resonance parameters require the use of the more complex bi-quad or state-variable filters. Bandpass and notch filters are said to be ‘tuneable’ if their centre frequency can be altered relatively easily, say by changing only one component value.

Notch filters

Notch filters are mainly used in equalisers to deal with narrow peaks in the acoustic response of performance spaces, and in some electronic crossovers – see, for example, [7]. They can also be used to remove a single interfering frequency – once, a long time ago, I was involved with a product that had a slide-projector in close proximity to a cassette player. Hifi was not the aim, but even so, the enormous magnetic field from the projector transformer induced an unacceptable amount of hum into the cassette tape head. Mu-metal only helped a bit, and the fix was a filter that introduced notches at 50 Hz and 100 Hz, working on the '1 – bandpass' principle, of which more later.

All-pass filters

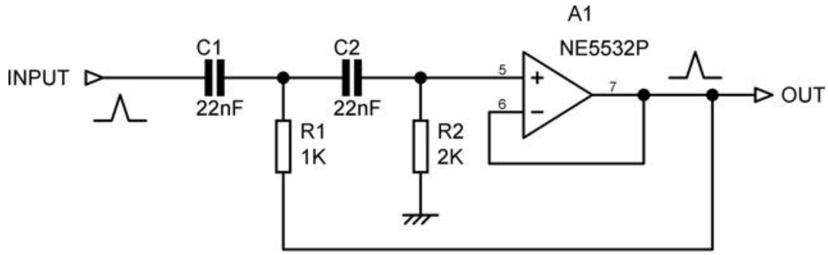
All-pass filters are so-called because they have a flat frequency response, and so pass all frequencies equally. Their point is that they have a phase-shift that *does* vary with frequency, and this is often used for delay correction in electronic crossovers. You may occasionally see a reference to an all-stop filter, which has infinite rejection at all frequencies. This is a filter designer's joke.

Filter characteristics

The simple second-order bandpass responses are basically all the same, being completely defined by centre frequency, Q, and gain. High-pass and low-pass filter characteristics are much more variable and are selected as a compromise between the need for a rapid roll-off, flatness in the passband, and a clean transient response. The Butterworth (maximally-flat) characteristic is the most popular for many applications. Filters with pass-band ripple, such as the Chebyshev or elliptical types, have not found favour for in-band filtering, such as in electronic crossovers, but were once widely used for applications like ninth-order anti-aliasing filters; such filters have mercifully been made obsolete by oversampling. The Bessel characteristic gives a maximally flat group delay (maximally linear phase response) across the passband, and so preserves the waveform of filtered signals, but it has a much slower roll-off than the Butterworth.

Sallen and Key filters

The Sallen and Key filter configuration was introduced by R. P. Sallen and E. L. Key of the MIT Lincoln Laboratory as long ago as 1955 [8]. It became popular as the only active element required is a unity-gain buffer, so in the days before opamps were cheap it could be effectively implemented with a simple emitter-follower.



$$f_0 = \frac{1}{2\pi C\sqrt{R_1 R_2}} \qquad Q = \frac{1}{2}\sqrt{\frac{R_2}{R_1}}$$

Figure 6.1: The classic second-order low-pass Sallen and Key filter. Cutoff frequency is 624 Hz. Q = 0.707 (critically damped)

Figure 6.1 shows a second-order low-pass Sallen and Key filter with a −3 dB frequency of 624 Hz and a Q of 0.707, together with the pleasingly simple design equations for cutoff (−3 dB) frequency f_0 and Q. It was part of a fourth-order Linkwitz-Riley electronic crossover for a three-way loudspeaker [9]. The main difference you will notice from textbook filters is that the resistor values are rather low and the capacitor values correspondingly high. This is an example of low-impedance design, where low resistor values minimise Johnson noise and reduce the effect of the opamp current noise and common-mode distortion. The measured noise output is −117.4 dBu. This is after correction by subtracting the testgear noise floor.

It is important to remember that a Q of 1 does not give the maximally-flat Butterworth response; you must use 0.707. Sallen and Key filters with a Q of 0.5 are used in second-order Linkwitz-Riley crossovers, but these are not favoured because the 12 dB/octave roll-off of the high-pass filter is not steep enough to reduce the excursion of a driver when a flat frequency response is obtained [9].

Sallen and Key low-pass filters have a lurking problem. When implemented with opamps, the response does not carry on falling for ever at the filter slope – instead it reaches a minimum and starts to come back up at 6 dB/octave. This is because the filter action relies on C1 seeing a low impedance to ground, and the impedance of the opamp output rises with frequency due to falling open-loop gain and hence falling negative feedback. When the circuit of Figure 6.1 is built using a TL072, the maximum attenuation is −57 dB at 21 kHz, rising again and flattening out at −15 dB at 5 MHz. This type of filter should not be used to reject frequencies well above the audio band; a low-pass version of the multiple feedback filter is preferred.

Low-pass filters used to define the top limit of the audio bandwidth are typically second-order with roll-off rates of 12 dB/octave; third-order 18 dB/octave filters are rather rarer, probably because there seems to be a general feeling that phase changes are more audible at the top end of the audio spectrum than the bottom. Either the Butterworth (maximally flat frequency response) or Bessel type (maximally flat group delay) can be used. It is unlikely that there is any real audible difference between the two types of filter in this application, as most of the action occurs above 20 kHz, but using the Bessel alignment does require compromises in the effectiveness of the filtering because of its slow roll-off. I will demonstrate.

The standard Butterworth filter in Figure 6.2a has its −3 dB point set to 50 kHz, and this gives a loss of only 0.08 dB at 20 kHz, so there is minimal intrusion into the audio band; see Figure 6.3. The response is a useful −11.6 dB at 100 kHz and an authoritative −24.9 dB at 200 kHz. C1 is made up of two 2n2 capacitors in parallel.

But let us suppose we are concerned about linear phase at high frequencies and we decide to use a Bessel filter. The only circuit change is that C1 is now 1.335 times as big as C2 instead

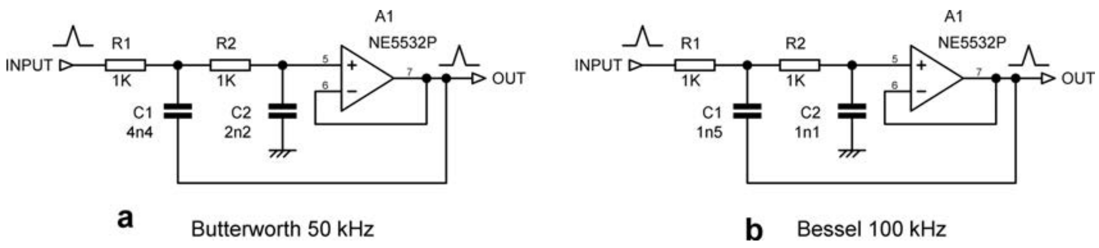


Figure 6.2: Second-order Sallen and Key low-pass circuits for ultrasonic filtering: a) Butterworth, b) Bessel. Both have a loss of less than 0.2 dB at 20 kHz

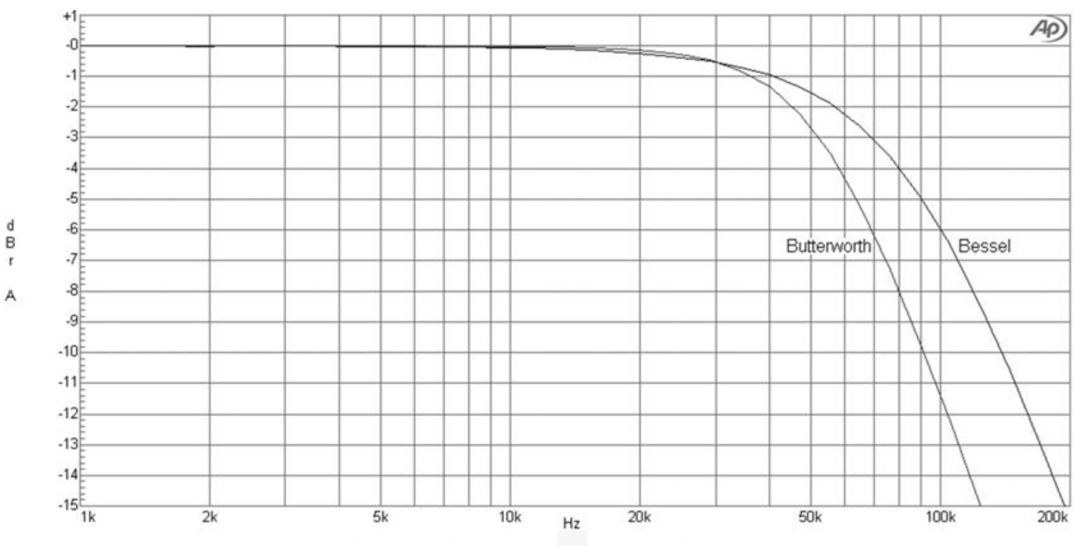


Figure 6.3: Frequency response of a 50 kHz Butterworth and a 72 kHz Bessel filter, as in Figure 6.2

of two times, but the response is very different. If we design for −3 dB at 50 kHz again, we find that the response is −0.47 dB at 20 kHz; a lot worse than 0.08 dB, and not exactly a stunning figure for your spec sheet. If we decide we can live with −0.2 dB at 20 kHz then the Bessel filter has to be designed for −3 dB at 72 kHz; this is the design shown in Figure 6.2b. Due to the inherently slower roll-off, the response is only down to −5.6 dB at 100 kHz, and −14.9 dB at 200 kHz, as seen in Figure 6.3; the latter figure is 10 dB worse than for the Butterworth. The measured noise output for both versions is −114.7 dBu using 5532s.

If we want to keep the 20 kHz loss to 0.1 dB, the Bessel filter has to be designed for −3 dB at 100 kHz, and the response is now only −10.4 dB down at 200 kHz, more than 14 dB less effective than the Butterworth. These results are summarised in Table 6.1:

TABLE 6.1 The frequency response of various ultrasonic filter options

Frequency (kHz)	Butterworth 50 kHz (dB)	Bessel 50 kHz (dB)	Bessel 72 kHz (dB)	Bessel 100 kHz (dB)
20	−0.08	−0.47	−0.2	−0.1
100	−11.6	−10.0	−5.6	−3.0
200	−24.9	−20.9	−14.9	−10.4

Discussions on filters always remark that the Bessel alignment has a slower roll-off, but often fail to emphasise that it is a *much* slower roll-off. You should think hard before you decide to go for the Bessel option in this sort of application.

It is always worth checking how the input impedance of a filter loads the previous stage. In this case, the input impedance is high in the pass-band, but above the roll-off point it falls until it reaches the value of R1, which here is 1 kΩ. This is because at high frequencies C1 is not bootstrapped, and the input goes through R1 and C1 to the low-impedance opamp output which is effectively at ground. Fortunately this low impedance only occurs at high frequencies, where one hopes the level of the signals to be filtered out will be low.

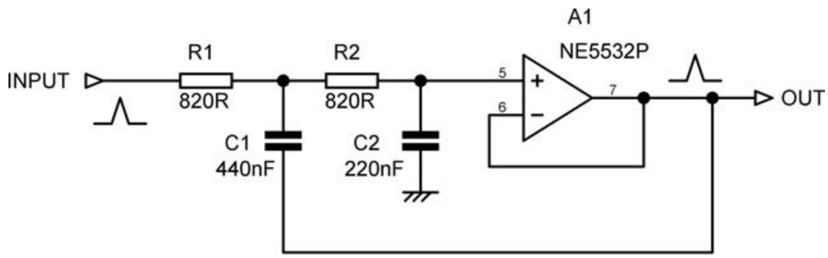
Another important consideration with low-pass filters is the balance between the R and C values in terms of noise performance. R1 and R2 are in series with the input and their Johnson noise will be added directly to the signal. Here the two 1 kΩ resistors together generate −119.2 dBu of noise (22 kHz bandwidth, 25 °C). The obvious conclusion is that R1 and R2 should be made as low in value as possible without causing excess loading, (and 1 kΩ is not a bad compromise) with C1, C2 scaled to maintain the desired roll-off frequency.

The need for specific capacitor ratios creates problems as capacitors are available in a much more limited range of values than resistors, usually the E6 series, running 10, 15, 22, 33, 47, 68. C1 or C2 often has to be made up of two capacitors in parallel.

Figure 6.4 shows a second-order high-pass Sallen and Key filter with a −3 dB frequency of 5115 Hz and a Q of 0.707, with its design equations; this was another part of the electronic crossover. The capacitors are equal while the resistors must have a ratio of two. The measured noise output of this filter is −115.2 dBu.

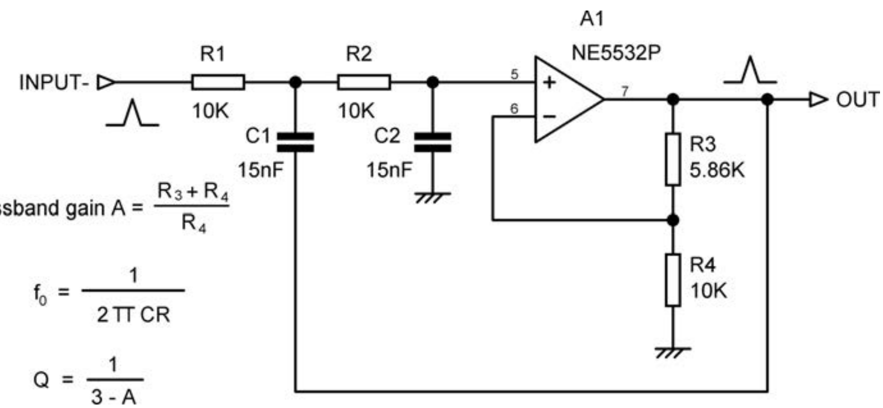
A variation on the low-pass Sallen and Key that can avoid capacitor ratio difficulties is shown in Figure 6.5. The unity-gain buffer is replaced with a voltage gain stage; the gain set by R3 and R4 must be 1.586 times (+4.00 dB) for a Q of 0.707. This allows C1 and C2 to be the same value. An equal-resistor-value high-pass filter can be made in exactly the same way.

To get a faster roll-off, we can use a third-order filter, which is a first-order filter (i.e. a simple RC time constant) cascaded with a second-order filter that has a Q of 1.0, causing a response peak. This peak combined with the slow first-order roll-off gives a flat passband and then



$$f_0 = \frac{1}{2\pi R\sqrt{C_1 C_2}} \qquad Q = \frac{1}{2}\sqrt{\frac{C_1}{C_2}}$$

Figure 6.4: The classic second-order high-pass Sallen and Key filter. Cutoff frequency is 5115 Hz. Q = 0.707 (critically damped)

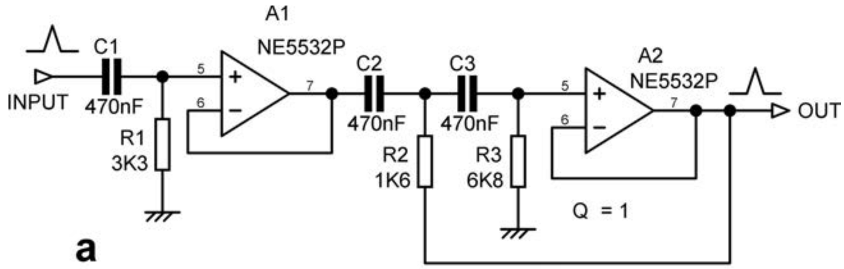


$$\text{Passband gain } A = \frac{R_3 + R_4}{R_4}$$

$$f_0 = \frac{1}{2\pi CR}$$

$$Q = \frac{1}{3 - A}$$

Figure 6.5: Equal-value second-order low-pass Butterworth filter with a cutoff frequency of 1061 Hz. Gain must be 1.586 times for maximally-flat response (Q = 0.707)



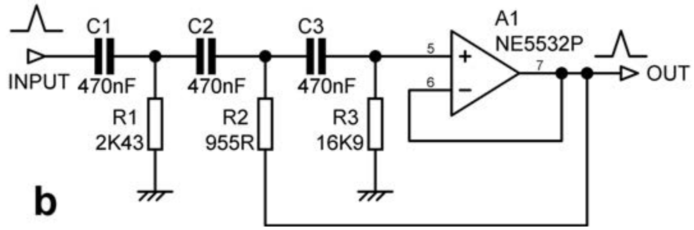


Figure 6.6: Two third-order Butterworth highpass filters with a cutoff frequency of 100 Hz

a steep roll-off. Figure 6.6a shows a third-order Butterworth filter with a −3 dB frequency of 100 Hz, as might be used at the front end of a mixer channel. It is built the obvious way, with a first-order filter R1, C1 followed by a unity-gain buffer to give low-impedance drive to the following second-order filter. R2 and R3 now have a ratio of 4 to obtain a Q of 1.0. E24 resistor values are shown and in this case they give an accurate response.

A more economical way to make a third-order filter is shown in Figure 6.6b, which saves an opamp section. The resistor values shown are the nearest E96 values to the mathematically exact numbers, and give an extremely accurate response. The nearest E24 values are R1 = 2k4, R2 = 910 Ω, and R3 = 16 kΩ, giving a very small peaking of 0.06 dB at 233 Hz. Errors due to the capacitor tolerances are likely to be larger than this. The distortion performance may not be as good as for Figure 6.6a.

Fourth–order filters are the steepest in normal use. They are made by cascading two second-order filters with Qs of 0.54 and 1.31. They can also be made in the same way as Figure 6.5b, but if so tend to be rather sensitive to component tolerances.

Distortion in Sallen and Key filters

When they have a signal voltage across them, many capacitor types generate distortion. This unwelcome phenomenon is described in Chapter 2. It afflicts not only all electrolytic capacitors, but also some types of non-electrolytic ones. If the electrolytics are being used

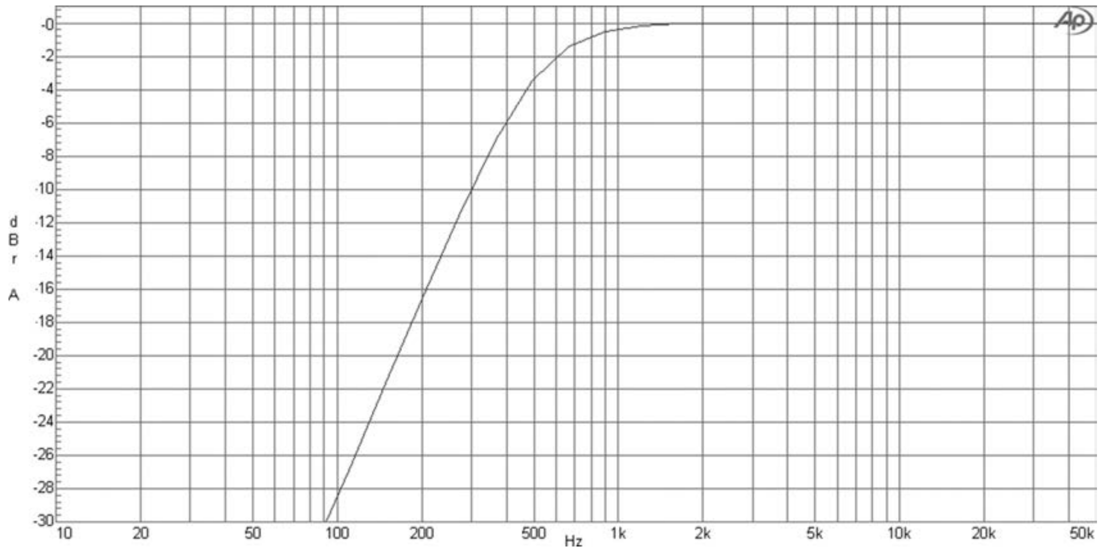


Figure 6.7: The frequency response of the second-order 520 Hz high-pass filter

as coupling capacitors, then the cure is simply to make them so large that they have a negligible signal voltage across them at the lowest frequency of interest; less than 80 mVrms is a reasonable criterion. This means they may have to be ten times the value required for a satisfactory frequency response.

However when non-electrolytics are used to set time-constants in filters they obviously must have substantial signal voltages across them and this simple fix is not usable. The problem is not a marginal one – the amounts of distortion produced can be surprisingly high. Figure 6.7 shows the frequency response of a conventional second-order Sallen and Key high-pass filter as seen in Figure 6.4, with a −3 dB frequency of 520 Hz. C1, C2 were 220 nF 100V polyester capacitors, with R1 = 1 kΩ and R2 = 2 kΩ. The opamp was a 5532. The distortion performance is shown by the upper trace in Figure 6.8; above 1 kHz the distortion comes from the opamp alone and is very low. However you can see it rising rapidly below 1 kHz as the filter begins to act, and it has reached 0.015 % by 100 Hz, completely overshadowing the opamp distortion; it is basically third-order. The input level was 10 Vrms, which is about as much as you are likely to encounter in an opamp system. The output from the filter has dropped to −28 dB by 100 Hz, and so the amplitude of the harmonics generated is correspondingly lower, but it still not a very happy outcome.

As explained in Chapter 2, polypropylene capacitors exhibit negligible distortion compared with polyester, and the lower trace in Figure 6.8 shows the improvement on substituting 220 nF 250 V polypropylene capacitors. The THD residual below 500 Hz is now pure noise, and the trace is only rising at 12 dB/octave because circuit noise is constant but

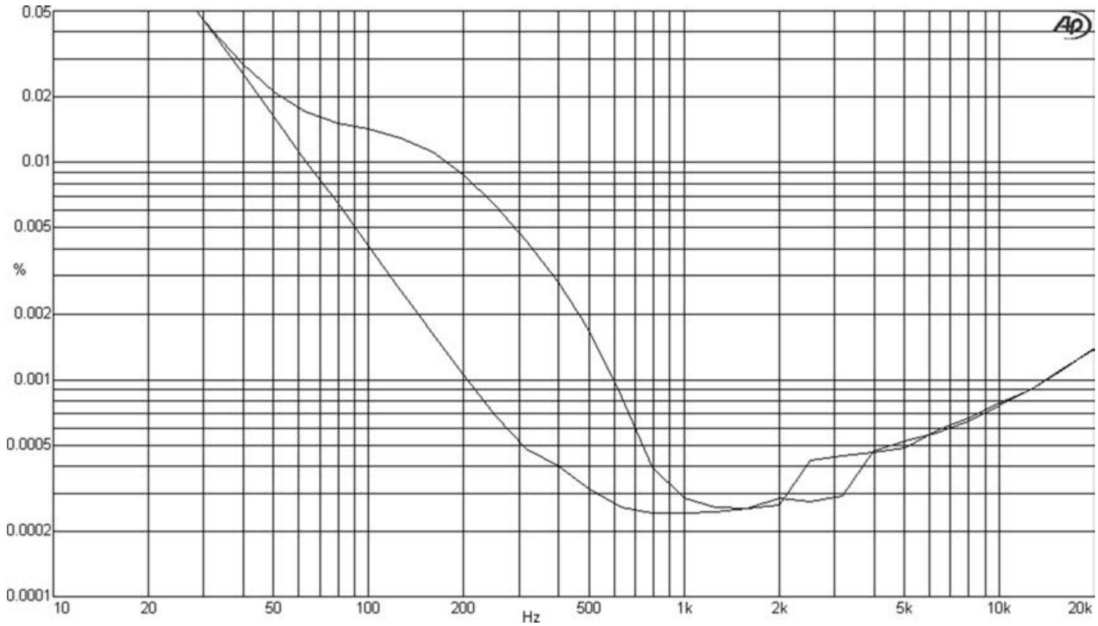


Figure 6.8: THD plot from the second-order 520 Hz high-pass filter; input level 10 Vrms. The upper trace shows distortion from polyester capacitors; the lower trace, with polypropylene capacitors, shows noise only

the filter output is falling. The important factor is the dielectric, not the voltage rating; 63 V polypropylene capacitors are also free from distortion. The only downside is that polypropylene capacitors are larger for a given CV product and more expensive.

Multiple-feedback bandpass filters

When a bandpass filter of modest Q is required, the multiple-feedback or Rauch type shown in Figure 6.9 has many advantages. The capacitors are equal and so can be made any preferred value. The opamp is working with shunt feedback and so has no common-mode voltage on the inputs, which avoids one source of distortion. It does however phase-invert, which can be inconvenient.

The filter response is defined by three parameters- the centre frequency f_0, the Q, and the passband gain (i.e. the gain at the response peak) A. The filter in Figure 6.8 was designed for $f_0 = 250$ Hz, Q = 2, and A = 1 using the equations given, and the usual awkward resistor values emerged. The resistors in Figure 6.7 are the nearest E96 value, and the simulated results come out as $f_0 = 251$ Hz, Q = 1.99, and A = 1.0024, which, as they say, is good enough for rock'n'roll.

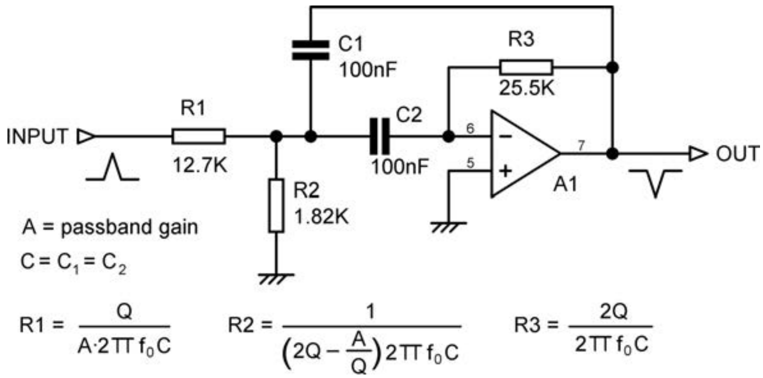


Figure 6.9: A bandpass multiple-feedback filter with f0 = 250 Hz, Q = 2 and a gain of 1

The Q of the filter can be quickly checked from the response curve as Q is equal to the centre frequency divided by the −3 dB bandwidth, i.e. the frequency difference between the two −3 dB points on either side of the peak. This configuration is not suitable for Qs greater than about 10, as the filter characteristics become unduly sensitive to component tolerances. If independent control of f_0 and Q are required the state-variable filter should be used instead.

Similar configurations can be used for low-pass and high-pass filters. The low-pass version does not depend on a low opamp output impedance to maintain stop-band attenuation at high frequencies, and avoids the 'oh no, it's coming back up again' behaviour of Sallen and Key low-pass filters.

Notch filters

There are many ways to dig a deep notch in your frequency response. The width of a notch is described by its Q; exactly as for a resonance peak, the Q is equal to the centre frequency divided by the −3 dB bandwidth, i.e. the frequency difference between the −3 dB points either side of the notch. The Q has no relation to the depth of the notch.

The best-known notch filter is the Twin-T notch network shown in Figure 6.10a, invented in 1934 by Herbert Augustadt [10]. The notch depth is infinite with exactly matched components, but with ordinary ones it is unlikely to be deeper than 40 dB. It requires ratios of two in component values which do not fit in well with preferred values, and when used alone has a Q of only ¼. It is therefore normally used with positive feedback via an opamp buffer A2, as shown. The proportion of feedback K and hence the Q-enhancement is set by R4 and R5, which here give a Q of 1. A great drawback is that the notch frequency can only be altered by changing three components, so it is not considered tuneable.

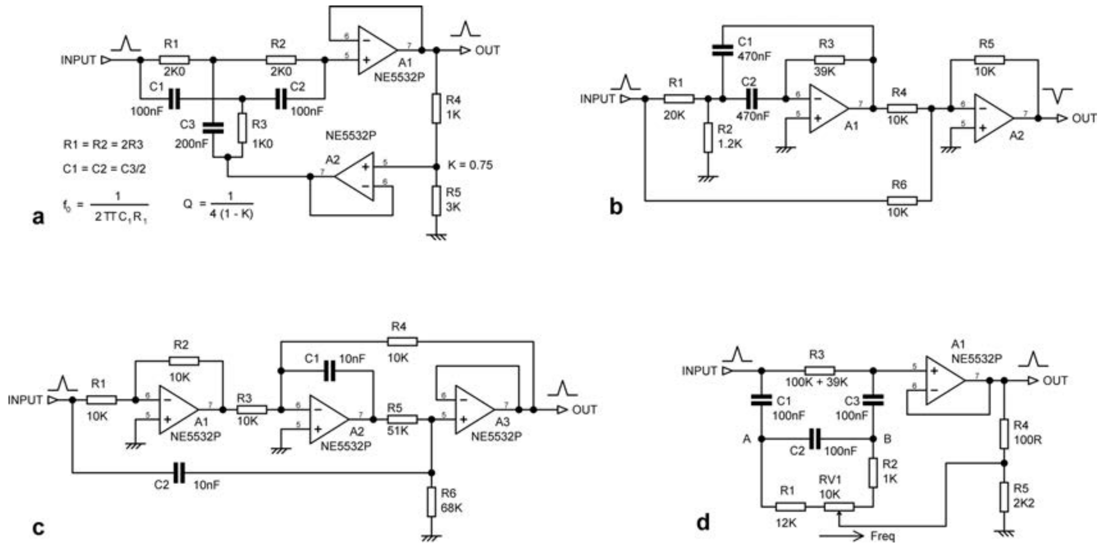


Figure 6.10: Notch filters: a) Twin-T with positive feedback, notch at 795 Hz and a Q of 1, b) '1-bandpass' filter with notch at 50 Hz, Q of 2.85, c) Bainter filter with notch at 700 Hz, Q of 1.29, d) Bridged-differentiator notch filter tuneable 80 to 180 Hz

Another way of making notch filters is the '1– bandpass' principle mentioned earlier. The input goes through a bandpass filter, typically the multiple-feedback type described earlier, and is then subtracted from the original signal. The accuracy of the cancellation and hence the notch depth is critically dependent on the mid-band gain of the bandpass filter. Figure 6.10b shows an example that gives a notch at 50 Hz with a Q of 2.85. The subtraction is performed by A2; note the output of the filter is phase-inverted. The multiple-feedback filter is designed for unity passband gain, but the use of E24 values as shown means that the actual gain is 0.97, limiting the notch depth to −32 dB. The value of R6 can be tweaked to deepen the notch; the nearest E96 value is 10.2 kΩ which gives a depth of −45 dB. The final output is inconveniently phase-inverted in the passband.

A notch filter that deserves to be better known is the Bainter filter [11], [12] shown in Figure 6.10c. It is non-inverting in the passband, and two out of three opamps are working at virtual earth and will give no trouble with common-mode distortion. An important advantage is that the notch depth does not depend on the matching of components, but only on the open-loop gain of the opamps, being roughly proportional to it. With TL072-type opamps the depth is from −40 to −50 dB. The values shown give a notch at 700 Hz with a Q of 1.29. The design equations can be found in reference [12].

A property of this filter that does not seem to appear in the textbooks is that if R1 and R4 are altered together, i.e. having the same values, then the notch frequency is tuneable with a good depth maintained, but the Q does change proportionally to frequency. To get a standard notch with equal gain either side of the crevasse, R3 must equal R4. R4 greater than R3 gives a low-pass notch, while R3 greater than R4 gives a high-pass notch; these responses are useful in crossover design [7].

The Bainter filter is usually shown with equal values for C1 and C2. This leads to values for R5 and R6 that are a good deal higher than other circuit resistances and this will impair the noise performance. I suggest that C2 is made ten times C1, i.e. 100 nF, and R5 and R6 are reduced by ten times to 5.1 kΩ and 6.8 kΩ; the response is unaltered and the Johnson noise much reduced.

But what about a notch filter that can be tuned with one control? Figure 6.10d shows a bridged-differentiator notch filter tuneable from 80 to 180 Hz by RV1. R3 must theoretically be six times the total resistance between A and B, which here is 138 kΩ, but 139 kΩ gives a deeper notch, about −27 dB across the tuning range. The downside is that Q varies with frequency from 3.9 at 80 Hz to 1.4 at 180 Hz.

Another interesting notch filter to look up is the Boctor, which uses only one opamp [13], [14]. Both the bi-quad and state-variable filters can be configured to give notch outputs.

When simulating notch filters, assessing the notch depth can be tricky. You need a lot of frequency steps to ensure you really have hit bottom with one of them. For example, in one run, 50 steps/decade showed a −20 dB notch, but upping it to 500 steps/decade revealed it was really −31 dB deep. In most cases, having a stupendously deep notch is pointless. If you are trying to remove an unwanted signal then it only has to alter in frequency by a tiny amount and you are on the side of the notch rather than the bottom and the attenuation is much reduced. The exception to this is the THD analyser, where a very deep notch (120 dB or more) is needed to reject the fundamental so very low levels of harmonics can be measured. This is achieved by continuously servo-tuning the notch so it is kept exactly on the incoming frequency.

Differential filters

At first an active filter that is also a differential amplifier, and thereby carries out an accurate subtraction, sounds like a very exotic creature. Actually they are quite common, and are normally based on the multiple-feedback filter described above. Their main application is DAC output filtering in CD payers and the like; more sophisticated versions are useful for data acquisition in difficult environments. Differential filters are dealt with in the section of Chapter 26 on interfacing with DAC outputs.

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CHAPTER 7

Preamplifier architectures

Some sort of preamplifier or control unit is required in all hifi systems, even if its only function is to select the source and set the volume. You could even argue that the source-selection switch could be done away with, if you are prepared to plug and unplug those rhodium-plated connectors, leaving a 'preamplifier' that basically consists solely of a volume control potentiometer in a box. Mind you don't catch your sackcloth on those heatsink corners, and try to keep the ashes away from the turntable.

I am assuming here that a selector switch will be required, and that gives us the 'passive preamplifier' (oxymoron alert!) in Figure 7.1a.

Passive preamplifiers

A device may have only one component but it does not follow that it is easy to design, even though the only parameter to decide is the resistance of the volume pot. Any piece of equipment that embodies its internal contradictions in its very name need to be treated with caution. The pot resistance of a 'passive preamplifier' cannot be too high because the output impedance, maximal at one quarter the track resistance when volume is set to −6 dB, will cause an HF roll-off in conjunction with the connecting cable capacitance. It also makes life difficult for those designing RF filters on the inputs of the equipment being driven, as described in Chapter 18.

On the other hand, if the volume pot resistance is too low the source equipment will suffer excessive loading. If the source is valve equipment, which does not respond well to even moderate loading, the problem starts to look insoluble.

If however, we can assume that our source equipment has a reasonable drive capability, we can use a 10 kΩ pot. Its maximum output impedance (at −6 dB) will then be 2.5 kΩ. The capacitance of most audio cable is 50–150 pF/metre, so with a 2.5 kΩ source impedance and 100 pF/metre cable, a maximum length of 5 metres is permissible before the HF loss hits the magic figure of −0.1 dB at 20 kHz. A very rapid survey of 'passive preamplifiers' in 2009 confirmed that 10 kΩ seems to be the most popular value. One model had a 20 kΩ potentiometer, and another had a 100 kΩ pot, which with its 25 kΩ source impedance would

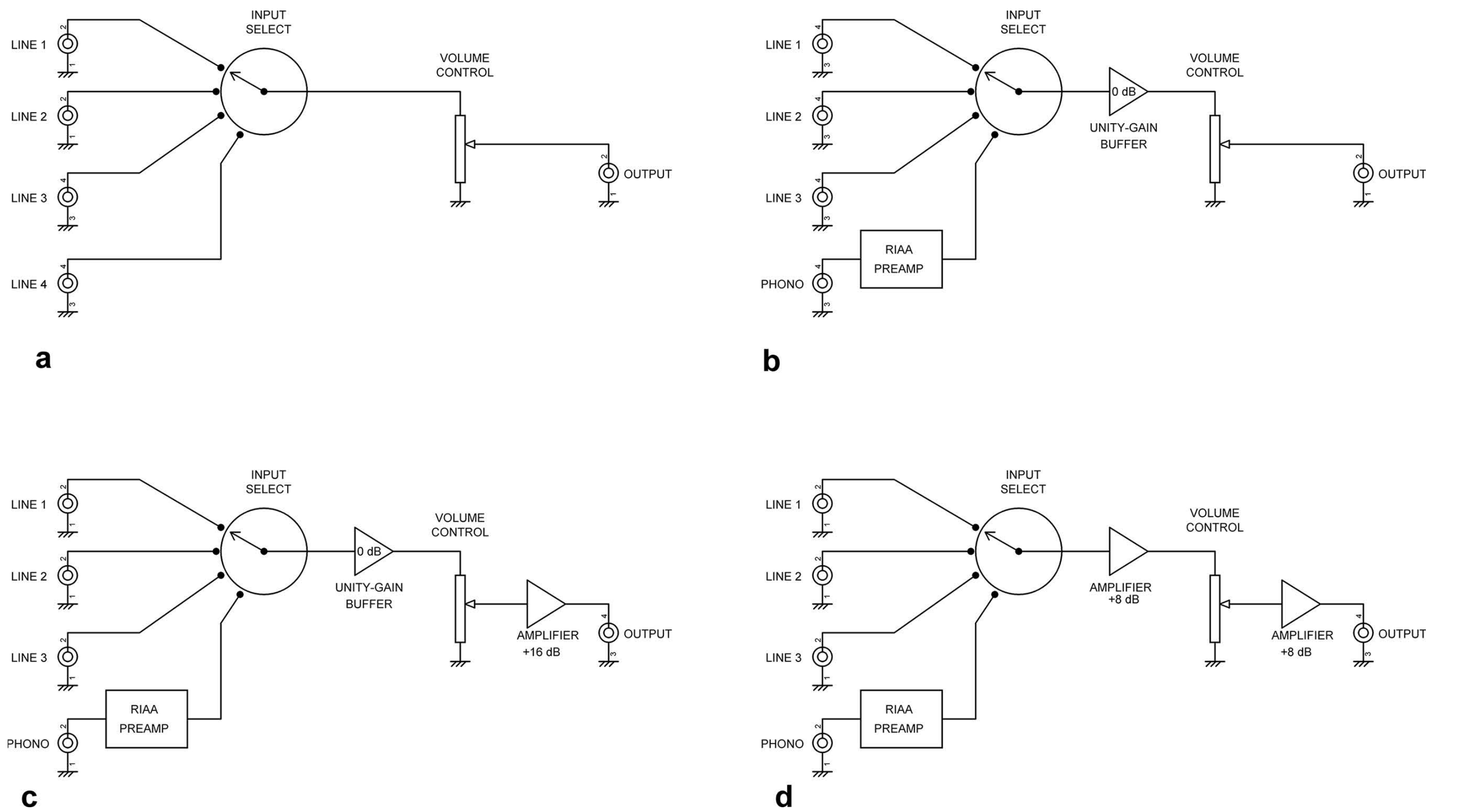

Figure 7.1: Preamplifier evolution: a) passive preamplifier, b) input buffer and phono amplifier added, c) amplification after the volume control added, d) amplification split into two stages, before and after volume control

hardly allow any cable at all. I suspect that the only reason such a pot can appear acceptable is because in normal use it is set well below −6 dB. For example, if a 100 kΩ pot is set to −15 dB, the output impedance is reduced to 14 kΩ, which would allow just under a metre of 100 pF/metre cable to be used with HF loss still limited to −0.1 dB at 20 kHz. A high output impedance also makes an interconnection more susceptible to interference unless it is totally screened at every point on its length.

Consider also that many power amplifiers have RC filters at the input, not so much for EMC immunity, but more as a gesture against what used to be called 'TID' but is actually just old-fashioned slew-limiting and highly unlikely in practice. These can add extra shunt capacitance to the input ranging from 100 pF to 1000 pF, apparently having been designed on the assumption of near-zero source impedance, and this can cause serious HF roll-offs.

There is at least one passive preamplifier on the market that controls volume by changing the taps on the secondary of a transformer; this should give much lower output impedances. There is more on this approach in Chapter 13 on volume controls.

Active preamplifiers

Once we permit ourselves active electronics, things get much easier. If a unity-gain buffer stage is added after the selector switch, as in Figure 7.1b, the volume pot resistance can be reduced to much less than 10 kΩ, while presenting a high impedance to the sources. If a 5532 is used there is no technical reason why the pot could not be as low as 1 kΩ, which will give a much more useable maximum output impedance of only 250 Ω, and also reduce Johnson noise by 10 dB. A phono preamplifier has also been added. Now we've paid for a power supply, it might as well supply something else.

This still leaves us with an 'amplifier' that has only unity line gain. Normally only CD players, which have an output of 2 Vrms, can fully drive a power amplifier without additional gain, and there are some high-power amplifiers that require more than this for full output. iPods appear to have a maximum output of 1.2 Vrms. Output levels for tuners, phono amps and so on vary but may be as low as 150 mVrms, while power amplifiers rarely have sensitivities lower than 500 mV. Clearly some gain would be a good thing, so one option is adding a gain stage after the volume control as in Figure 7.1c. The output level can be increased and the output impedance kept down to 100 Ω or lower.

This amplifier stage introduces its own difficulties. If its nominal output level with the volume control fully up is taken as 1 V rms for 150 mV in, which will let us drive most power amps to full output from most sources most of the time, we will need a gain of 6.7 times or 16.5 dB. If we decide to increase the nominal output level to 2 Vrms, to be sure of driving most if not all exotica to its limits, we need 22.5 dB. The problem is that the gain stage is amplifying its own noise at all volume settings, and amplifying a proportion of the Johnson noise of the pot whenever the wiper is off the zero stop. The noise performance will therefore deteriorate markedly at low volume levels, which are the ones most used.

Amplification and the gain-distribution problem

One answer to this difficulty is to take the total gain and split it so there is some before and some after the volume control, so there is less gain amplifying the noise at low volume settings. One version of this is shown in Figure 7.1d. The question is – how much gain before, and how much after? This is inevitably a compromise, and it might be called the gain-distribution problem. Putting more of the total gain before the volume control reduces the headroom as there is no way to reduce the signal level, while putting more after increases the noise output at low volume settings.

If you are exclusively using sources with a predictable output, of which the 2 Vrms from a CD player will be the maximum, the overload situation is well-defined, and if we assume that the pre-volume gain stage is capable of at least 8 Vrms out, so long as the pre-volume control gain is less than four times, there will never be a clipping problem. However, phono cartridges, particularly moving coil ones, which have a very wide range of sensitivities, produce much less predictable outputs after fixed-gain preamplification, and it is a judgement call as to how much safety margin is desirable.

As an aside, it's worth bearing in mind that even putting a unity-gain buffer before the volume control, which we did as the first step in preamp evolution, does place a constraint on the signal levels that can be handled, albeit at rather a high level of 8 to 10 Vrms depending on the supply rails in use. The only source likely to be capable of putting out such levels is a mixing console with the group faders fully advanced. There is also the ultimate constraint that a volume control pot can only handle so much power, and the manufacturers' ratings are surprisingly low, sometimes only 50 mW. This means that a 10 kΩ pot would be limited to 22 Vrms across it, and if you are planning to use lower resistance pots than this to reduce noise, their power rating needs to be kept very much in mind.

Whenever a compromise appears in engineering, you can bet that someone will try to find a way round it and get the best of both worlds. What can be done about the gain-distribution dilemma?

One possibility is the use of a special low-noise amplifier after the volume control, combined with a low resistance volume pot as suggested above. This could be done either by a discrete-device and opamp hybrid stage, or by using a multiple opamp array, as described in Chapter 1. It is doubtful if it is possible to obtain more than a 10 dB noise improvement by these means, but it would be an interesting project.

Another possible solution is the use of double gain controls. There is an input-gain control before any amplification stage which is used to set the internal level appropriately, thus avoiding overload, and after the active stages there is an output volume control, which gives the much-desired silence at zero volume (see Figure 7.2a). The input gain controls can be separate for each channel, so they double as a balance facility; this approach was used on the

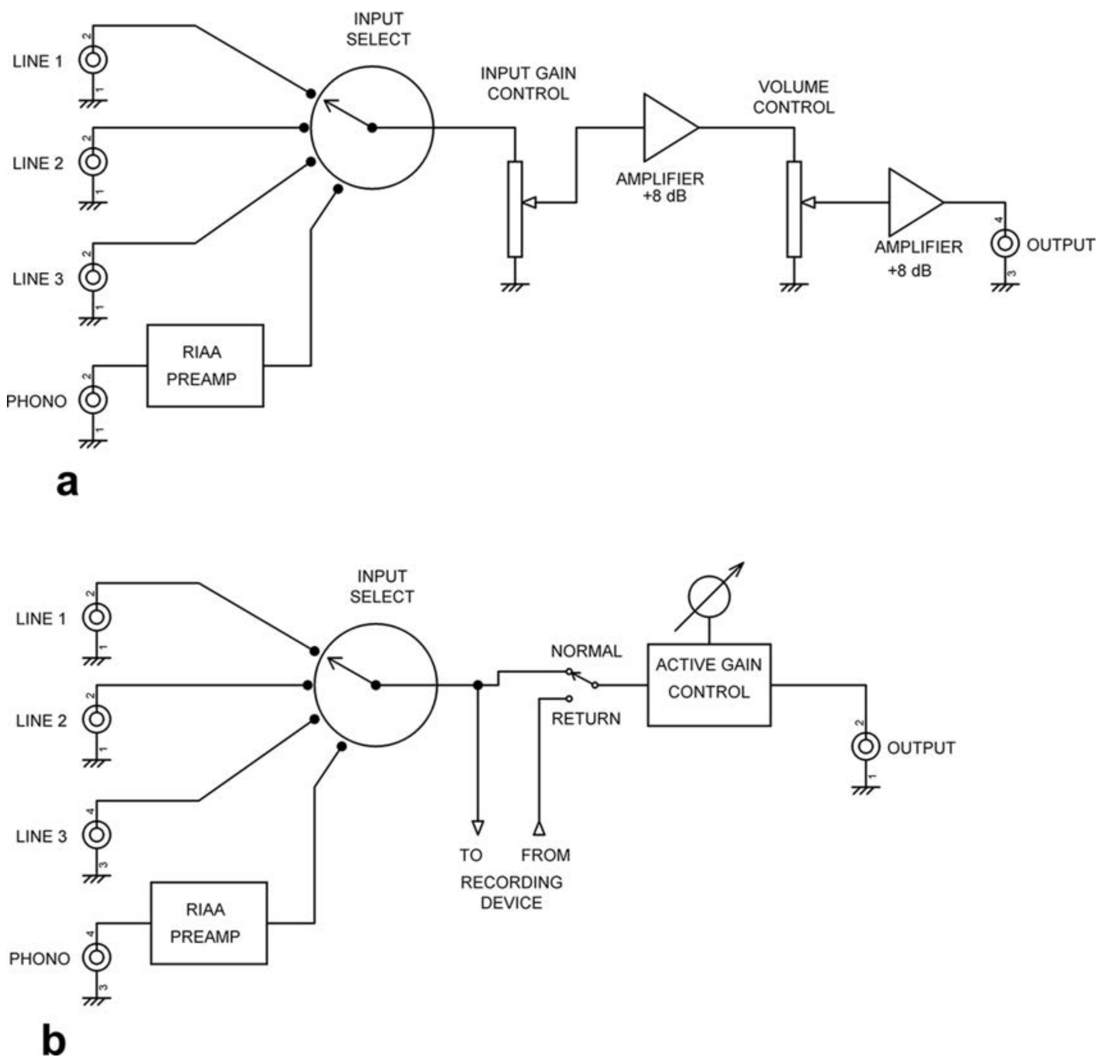

Figure 7.2: More preamp architectures: a) with input gain control and output volume control, b) with recording output and return input, and an active gain-control

Radford HD250 amplifier, and also in one of my early preamplifier designs [1]. This helps to offset the cost of the extra pot. However, having two gain controls is operationally rather awkward, and however attenuation and fixed amplification are arranged, there are always going to be some tradeoffs between noise and headroom. It could also be argued that this scheme does not make a lot of sense unless some means of metering the signal level after the input gain control is provided, so it can be set appropriately.

If the input and output gain controls are ganged together, to improve ease of operation at the expense of flexibility, this is sometimes called a distributed gain control.

Active gain controls

The noise/headroom compromise is completely avoided by replacing the combination of volume-control-and-amplifier with an active gain control, i.e. an amplifier stage whose gain is variable from near-zero to the required maximum (see Figure 7.2b). We get lower noise at gain settings below maximum, and we can increase that maximum gain so even the least sensitive power amplifiers can be fully driven, without impairing the noise performance at lower settings. We also get the ability to generate a quasi-logarithmic law from a linear pot, which gives excellent channel balance as it depends only on mechanical alignment. The only snags are that:

- most active gain controls phase-invert, though this can be corrected by suitable connection of a balanced input or balanced output stage, or a Baxandall tone control;
- the noise out is very low but not zero at zero volume as it would be with a passive pot since the noise gain does not fall below unity.

The technology of active gain controls is dealt with fully in Chapter 13.

Recording facilities

If a preamplifier is going to be used for recording the minimum requirement is an output taken from before the volume control, and there is usually also a dedicated input for a signal coming back from the recorder which can be switched to for checking purposes. Back when recording was done on tape, the return signal could be taken from the replay heads and gave assurance that recording had actually happened. Now that recording is done on hard disk machines or PCs the return signal normally only assures you that the signal has actually got there and back.

Much ingenuity used to be expended in designing switching systems so you could listen to one source while recording another, though it is rather doubtful how many people actually want to do this; it demands very high standards of crosstalk inside the preamplifier to make sure that the signal being recorded is not contaminated by another source.

Tone controls

Let us now consider adding tone controls. They have been unfashionable for a while, but this is definitely changing now. I think they are absolutely necessary, and it is a startling situation when, as frequently happens, anxious inquirers to hifi advice columns are advised to change their loudspeakers to correct excess or lack of bass or treble. This is an extremely expensive alternative to tone controls.

There are many possible types, as described in Chapter 15, but one thing they have in common is that they must be fed from a low-impedance source to give the correct boost/cut figures and predictable EQ curves. Another vital point is that most types, including the famous Baxandall configuration, phase-invert. Since there is now pretty much a consensus that all audio equipment should maintain absolute phase polarity for all input and outputs, this can be highly inconvenient.

However, as noted above, this phase inversion can very neatly be undone by the use of an active gain control, which also uses shunt feedback and so also phase-inverts. The tone-control can be placed before or after the active gain control, but if placed afterwards it generates noise that cannot be turned down. Putting it before the active gain control reduces headroom if boost is in use, but if we assume the maximum boost used is +10 dB, the tone-control will not clip until an input of 3 Vrms is applied, and domestic equipment rarely generates such levels. It therefore seems best to put the tone control before the active gain control. This is what I did in my most recent preamplifier designs [2], [3], [4], [5].

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CHAPTER 8

Moving-magnet inputs: levels and RIAA equalisation

Cartridge types

This chapter and those following deal with the design of preamplifiers to accept moving-magnet (MM) cartridge inputs, with their special loading requirements and need for RIAA equalisation. MM cartridges have been for many years less popular than moving coil (MC) cartridges (their preamps have their own chapter), but seem to be staging a comeback. However it would be unusual nowadays to design a phono input that accepted MM inputs only. There are several ways to design a combined MM/MC input, but the approach that gives the best results is to design an MM preamp that incorporates the RIAA equalisation, and put a flat-response low-noise head amplifier in front of it to get MC inputs up to MM levels. A large part of this chapter is devoted to the tricky business of RIAA equalisation, and so is equally relevant to MC input design. Almost all modern cartridges are of these two types, though Grado make a moving-iron (MI) series; this is essentially a variation on MM.

Ceramic cartridges were still very much around when I first got into the audio business, but they now seem to be a rare example of an obsolete audio technology that no one wants to revive. Ceramic cartridges have elements of Rochelle salt (in early versions, often called crystal pickups) or PZT (in ceramic versions) which generate electricity when flexed by the stylus. They look like a pure capacitance of around 200 pF to an amplifier input, and given a suitably high-impedance load, greater than 2 MΩ, they respond to stylus displacement rather than velocity. This led people to say that RIAA equalisation was not required, though this seems to overlook the presence of the 2-octave plateau in the middle of the RIAA characteristic. With 2 MΩ loading the output was in the range 200–600 mVrms, much higher than the output of MM cartridges. This did not necessarily simplify preamplifier circuitry because of the need to establish a 2 MΩ input impedance; there is no real difficulty in doing this even with low-beta BJTs, as described in Chapter 3, but a few more parts are needed. Alternatively the cartridge can be more heavily loaded and equalisation applied. More on the two philosophies of ceramic cartridge termination can be found in a 1969 article by Linsley-Hood [1] and there was a thorough discussion of the whole business by Burrows in 1970 [2], [3].

The tracking force for ceramic cartridges is usually higher than that of MM or MC, often being on the order of 3 to 5 grams, and the increased groove wear is one powerful reason for not using them today. Sonotone and Acos were major ceramic cartridge manufacturers. The first cartridge I ever owned was a Sonotone 9TA; I still have it. Like most of its kind, it was a turnover cartridge; in other words there were two styli, one on each side of the cantilever. One was for microgroove LPs and the other for coarse-groove 78s, and they were selected by turning over the cantilever with a little plastic tab. This type of cartridge is still sometimes used for transcribing old 78 rpm records. An excellent account of the history of Sonotone cartridges can be found in [4].

Strain-gauge cartridges also have a long history, and are still with us today. Those made by Sound-Smith [5] appear to have a good reputation, though I have no experience with them myself. They are also sensitive to stylus displacement, rather than stylus velocity (as MM and MC cartridges are). Naturally they require a specialised preamplifier, and the whole set-up is not cheap.

Capacitance pickups, aka 'FM pickups' or 'electrostatic pickups' consist of a small capacitor, one plate of which is wiggled by the stylus; the change in capacitance frequency-modulates an oscillator and the signal is decoded by standard FM-receiver technology. These were made by Weathers; the relevant patent appears to be US 4,489,278 granted to Paul Weathers in 1984. Stax also made them – their first capacitance cartridge was the mono CP-20 introduced in 1952, the stereo CPS-40 not appearing until 1962. In 1977, Stax introduced a new technology, the CP-Y cartridge having a permanent electret element and a head-amplifier IC built into the cartridge body. Shortly afterwards, their FM system was discontinued.

The vinyl medium

The vinyl disc as a medium for music delivery in its present form dates back to 1948 when Columbia introduced microgroove 33⅓ rpm LP records. These were followed soon after by microgroove 45 rpm records from RCA Victor. Stereo vinyl did not appear until 1958. The introduction of Varigroove technology, which adjusts groove spacing to suit the amplitude of the groove vibrations, using an extra look-ahead tape head to see what the future holds, allowed increases in groove packing density. This density rarely exceeded 100 grooves per inch in the 78 rpm format, but with Varigroove 180 to 360 grooves/inch could be used at 33⅓ rpm.

While microgroove technology was unquestionably a considerable improvement on 78 rpm records, any technology that is 60 years old is likely to show definite limitations compared with contemporary standards, and indeed it does. Compared with modern digital formats, vinyl has a restricted dynamic range, poor linearity (especially at the end of a side) and is very vulnerable to permanent and irritating damage in the form of scratches. Even with the greatest care, scratches are likely to be inflicted when the record is removed from its sleeve.

This action also generates significant static charges which attract dust and lint to the record surface. If not carefully removed this dirt builds up on the stylus and not only degrades the reproduction of high-frequency information today, but may also damage it in the future if it provokes mistracking.

Vinyl discs do not shatter under impact like the 78 shellac discs, but they are subject to warping by heat, improper storage, or poor manufacturing quality control. Possibly the worst feature of vinyl is that the stored material is degraded every time the disc is played, as the delicate high-frequency groove modulations are worn away by the stylus. When a good turntable with a properly balanced tone arm and correctly set-up low-mass stylus is used this wear process is relatively slow, but it nevertheless proceeds inexorably. The stylus suffers wear too.

However, for reasons that have very little to do with logic or common-sense, vinyl is still very much alive. Even if it is accepted that as a music-delivery medium it is technically as obsolete as wax cylinders, there remain many sizable album collections that it is impractical to replace with CDs and would take an interminable time to transfer to the digital domain. I have one of them. Disc inputs must therefore remain part of the audio designer's repertoire for the foreseeable future, and the design of the specialised electronics to get the best from the vinyl medium is still very relevant.

Spurious signals

It is not easy to find dependable statistics on the dynamic range of vinyl, but there seems to be general agreement that it is in the range 50 to 80 dB, the 50 dB coming from the standard quality discs, and the 80 dB representing direct-cut discs produced with quality as the prime aim. My own view is that 80 dB is rather optimistic.

The most audible spurious noise coming from vinyl is that in the mid frequencies, stemming from the inescapable fact that the music is read by a stylus sliding along a groove of finite smoothness. There is nothing that the designer of audio electronics can do about this. Scratches create clicks that have a large high-frequency content, and it has been shown that they can easily exceed the level of the audio [6]. It is important that such clicks do not cause clipping or slew-limiting, as this makes their subjective impact worse.

The signal from a record deck also includes copious amounts of low-frequency noise, which is often called rumble; it is typically below 30 Hz. This can come from several sources:

1. Mechanical noise generated by the motor bearings and picked up by the stylus/arm combination. These tend to be at the upper end of the low-frequency domain, extending up to 30 Hz or thereabouts. This is a matter for the mechanical designer of the turntable, as it clearly cannot be filtered out without removing the lower part of the audio spectrum.

2. Room vibrations will be picked up if the turntable and arm system is not well isolated from the floor. This is a particular problem in older houses where the wooden floors are not built to modern standards of rigidity, and have a perceptible bounce to them. Mounting the turntable shelf on the wall usually gives a major improvement; subsonic filtering is effective in removing room vibration.

3. Low frequency noise from disc imperfections. This is the worst cause of disturbances. They can extend as low as 0.55 Hz, the frequency at which a 33⅓ rpm disc rotates on the turntable, and is due to large-scale disc warps. Warping can also produce ripples in the surface, generating spurious subsonic signals up to a few Hertz at surprisingly high levels. These can be further amplified by a poorly controlled resonance of the cartridge compliance and the pickup arm mass. When woofer speaker cones can be seen wobbling – and bass reflex designs with no cone loading at very low frequencies are the worst for this – disc warps are usually the cause. Subsonic filtering is again effective in removing this.

(As an aside, I have heard it convincingly argued that bass reflex designs have only achieved their current popularity because of the advent of the CD player, with its greater bass signal extension but lack of subsonic output.)

Some fascinating data on the subsonic output from vinyl was given by Tomlinson Holman in [7] which show that the highest warp signals occur in the 2 to 4 Hz region, being some 8 dB less at 10 Hz. By matching these signals with a wide variety of cartridge-arm combinations, he concluded that to accommodate the very worst cases, a preamplifier should be able to accept not less than 35 mVrms in the 3–4 Hz region. This is a rather demanding requirement, driven by some truly diabolical cartridge-arm setups that accentuated subsonic frequencies by up to 24 dB.

Since the subsonic content generated by room vibrations and disc imperfections tends to cause vertical movements of the stylus, the resulting electrical output will be out of phase in the left and right channels. The use of a central mono subwoofer system that sums the two channels will provide partial cancellation, reducing the amount of rumble that is reproduced. It is, however, still important to ensure that subsonic signals do not reach the left and right speakers.

Other problems with vinyl

The reproduction of vinyl involves other difficulties apart from the spurious signals mentioned above:

- Distortion is a major problem. It is pretty obvious that the electromechanical processes involved are not going to be as linear as we now expect our electronic circuitry to be. Moving-magnet and moving-coil cartridges add their own distortion, which can reach 1 to 5% at high levels.

- Distortion gets worse as the stylus moves from the outside to the inside of the disc. This is called 'end of side distortion' because it can be painfully obvious in the final track. It occurs because the modulation of the inner grooves is inevitably more compressed than those of the outer tracks, due to the constant rotational speed of a turntable. I can well recall buying albums, and discovering to my chagrin that a favourite track was the last on a side.

- It is a notable limitation of the vinyl process that the geometry of the recording machine and that of the replay turntable do not match. The original recordings are cut on a lathe where the cutting head moves in a radial straight line across the disc. In contrast, almost all turntables have a pivoting tone arm about 9 inches in length. The pickup head is angled to reduce the mismatch between the recoding and reply situations, but this introduces side forces on the stylus and various other problems, increasing the distortion of the playback signal. A recent article in *Stereophile* [8] shows just how complicated the business of tone arm geometry is. SME produced a 12-inch arm to reduce the angular errors – I have one and it is a thing of great beauty, but I must admit I have never put it to use.

- The vinyl process depends on a stylus faithfully tracking a groove. If the groove modulation is excessive, with respect to the capabilities of the cartridge/arm combination, the stylus loses contact with the groove walls and rattles about a bit. This obviously introduces gross distortion, and is also very likely to damage the groove.

- A really disabling problem is 'wow' – the cyclic pitch change resulting from an off-centre hole. Particularly bad examples of this used to be called 'swingers' because they were so eccentric that they could be visibly seen to be rotating off-centre. I understand that nowadays the term means something entirely different, and relates to an activity which sounds as though it could only be a distraction from critical listening.

Maximum signal levels on vinyl

There are limits to the signal level possible on a vinyl disc, and they impose maxima on the signal that a cartridge and its associated electronics will be expected to reproduce. The exact values of these limits may not be precisely defined, but the way they work sets the ways in which maximum levels vary with frequency, and this is of great importance.

There are no variable gain controls on RIAA inputs, because implementing an uneven but very precisely controlled frequency response and a suitably good noise performance are quite hard enough without adding variable gain as a feature. No doubt it could be done, but it would not be easy, and the general consensus is that it is not necessary. The overload margin, or headroom, is therefore of considerable importance, and it is very much a case of the more the merrier when it comes to the numbers game of 'specmanship'. The issue can get a bit involved, as a situation with frequency-dependant vinyl limitations and frequency-dependant

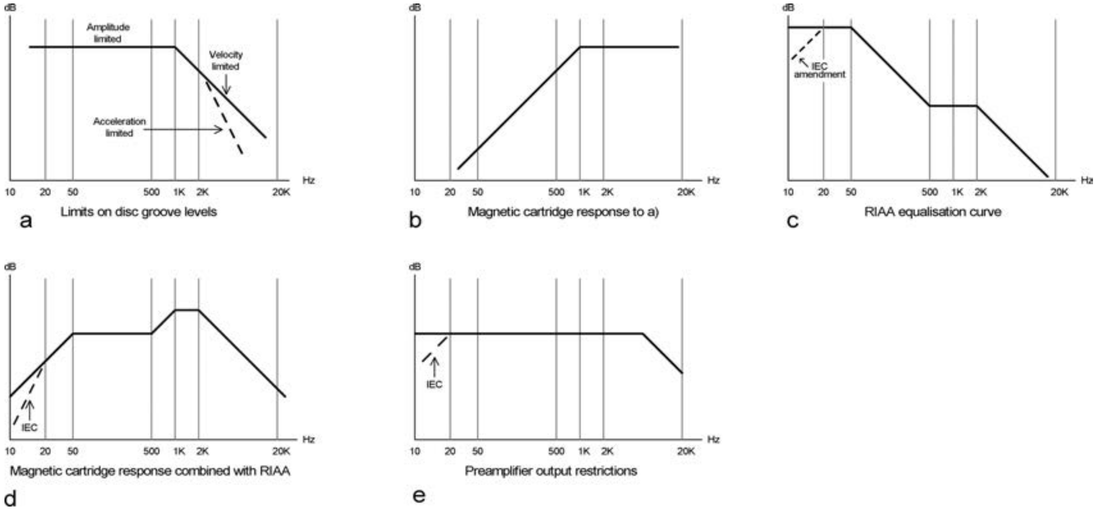

Figure 8.1: a) The levels on a vinyl disc, b) the cartridge response combined with the disc levels, c) the RIAA curve, d) the RIAA combined with curve b, e) possible preamplifier output restrictions

gain is often further complicated by a heavy frequency-dependant load in the shape of the feedback network, which can put its own limit on amplifier output at high frequencies. Let us first look at the limits on the signal levels which stylus-in-vinyl technology can deliver. In the diagrams above the response curves have been simplified to the straight-line asymptotes.

Figure 8.1a shows the physical groove amplitudes that can be put onto a disc. From subsonic up to about 1 kHz, groove amplitude is the constraint. If the sideways excursion is too great, the groove spacing will need to be increased to prevent one groove breaking into another, and playing time will be reduced. Well before actual breakthrough occurs, the cutter can distort the groove it has cut on the previous revolution, leading to ‘pre-echo’ in quiet sections, where a faint version of the music you are about to hear is produced. Time travel may be fine in science fiction but it does not enhance the musical experience. The ultimate limit to groove amplitude is set by mechanical stops in the cutter head.

There is an extra limitation on groove amplitude: out-of-phase signals cause vertical motion of the cutter, and if this becomes excessive it can cause it to cut either too deeply into the disc medium and dig into the aluminium substrate, or lose contact with the disc altogether; an excessive vertical component can also upset the playback process, especially when low tracking forces are used – in the worst case the stylus can be thrown out of the groove completely. To control this problem, the stereo signal is passed through a matrix that isolates the L-R vertical signal, which is then amplitude limited. This potentially reduces the perceived stereo separation at low frequencies, but there appears to be a general consensus that the effect is not audible. The most important factor in controlling out-of-phase signals is the panning of bass instruments (which create the largest cutter amplitudes) to the centre of the stereo stage. This approach is still advantageous with digital media as it means that there are two channels of amplification to reproduce the bass information rather than one.

From about 1 kHz up to the ultrasonic regions, the limit is groove velocity rather than amplitude. If the disc cutter head tries to move sideways too quickly compared with its relative forward motion, the back facets of the cutter destroy the groove that has just been cut by its forward edges.

On disc replay, there is a third restriction – that of stylus acceleration, or to put it another way, groove curvature. This sets a limit on how well a stylus of a given size and shape can track the groove. Allowing for this at cutting time places an extra limitation on signal level, shown by the dotted line in Figure 8.1a. The severity of this restriction depends on the stylus shape; an old-fashioned spherical type with a tip diameter of 0.7 mil requires a roll-off of maximum levels from 2 kHz, while a (relatively) modern elliptical type with 0.2 mil effective diameter postpones the problem to about 8 kHz. The limit however still remains.

Thus disc-cutting and playback technology put at least three limits on the maximum signal level. This is not as bad a problem as it might be, because the distribution of amplitude with frequency for music is not flat with frequency; there is always more energy at LF than HF. This is especially true of the regrettable phenomenon known as rap music. For some reason there seems to be very little literature on the distribution of musical energy versus frequency, but a very rough rule is that levels can be expected to be fairly constant up to 1 kHz and then fall by something like 10 dB/octave. The end result is that despite the limits on disc levels at HF, it is still possible to apply a considerable amount of HF boost which, when undone at replay, reduces surface noise problems. At the same time the LF levels are cut to keep groove amplitude under control. Both functions are implemented by applying the inverse of the familiar RIAA replay equalisation at cutting time. More on the limitations affecting vinyl levels can be found on Jim Lesurf's website [9].

A reaction to the limitations of the usual seven-inch single was the twelve-inch single, which appeared in the mid-1970s, before CDs arrived. The much greater playing area allowed greater groove spacing and higher recording levels. I bought several of these, in the 45 rpm format, and I can testify that the greater groove speed gave a much clearer and less distorted high end, definitely superior to 33 rpm LPs.

Having looked at the limitations on the signal levels put onto disc, we need to see what will get back when we replay it. This obviously depends on the cartridge sensitivity. That issue is dealt with in the next section, but it might as well be said now that, in general, MM cartridge sensitivity varies over a limited range of about 7 dB, while MC sensitivity variation is much greater.

Since MM input stages do not normally have gain controls, it is important that they can accept the whole range of input levels that occur. A well-known paper by Tomlinson Holman [10] quotes a worst-case peak voltage from an MM cartridge of 135 mV at 1 kHz given by [11]. This is equivalent to 95 mVrms at 1 kHz. He says, 'this is a genuinely worst-case combination which is not expected to be approached typically in practice.'

Shure are a well-known manufacturer of MM cartridges, and their flagship V15 phonograph cartridge series, (the 15 in each model name referred to the cartridges' 15-degree tracking angle) for many years set the standard for low tracking force and high tracking ability. Its development necessitated much research into the maximum levels on vinyl. Many other workers also contributed in this field. The results are usually expressed in velocity (cm/s) as this eliminates the effect of cartridge sensitivity. I have boiled down the Shure velocity data into Table 8.1. I have included the acceleration of the stylus tip required for the various frequency/velocity pairs; this is not of direct use, but given that the maximum sustained acceleration the human body can withstand with is around 3 g, it surely makes you think. Since the highest MM cartridge sensitivity for normal use is 1.6 mV per cm/s (see next section), Table 8.1 tells us that we need to be able to handle an MM input of $1.6 \times 38 = 61$ mVrms. This is not far out of line with the 95 mVrms quoted by Holman, being only 3.8 dB lower.

TABLE 8.1 Maximum groove velocities from vinyl (after Shure)

	400 Hz	500 Hz	2 kHz	5 kHz	8 kHz	10 kHz	20 kHz
Velocity (cm/s)	26	30	38	35	30	26	10
Acceleration (g)	66.5	96	487	1120	1535	1665	1281

The website of Jim Lesurf [12] has many contemporary measurements of maximum groove velocities. The maximum quoted is 39.7 cm/s, which gives $1.6 \times 39.7 =$ 63.5 mVrms. Rooting through the literature, the Pressure Cooker discs by Sheffield Labs were recorded direct to disc and are said to contain velocities up to 40 cm/s, giving us $1.6 \times 40 = 64$ mVrms. It is reassuring that these maxima do not differ very much. On the other hand, the jazz record *Hey! Heard The Herd* by Woody Herman (Verve V/V6 8558, 1953) is said to peak at a velocity of 104 cm/s at 7.25 kHz, [13] but this seems out of line with all other data. If it is true, the input level from the most sensitive cartridge would be $1.6 \times 105 = 166$ mVrms.

So we may conclude that the greatest input level we are likely to encounter is 64 mVrms, though that 166 mVrms should perhaps not be entirely forgotten.

TABLE 8.2 Maximum input, overload margin, and nominal output for various MM preamp gains. All at 1 kHz

Gain dB (dB)	Gain times	Max input mVrms (mV)	Overload margin dB (dB)	5 mVrms would be raised to: (mVrms)
50	316	32	16	1580
45.5	188	53	21	942
40	100	100	26	500
35	56.2	178	31	281
30	31.6	316	36	158
25	17.8	562	41	89
20	10	1000	46	50

The maximum input a stage can accept before output clipping is set by its gain and supply rails. If we are using normal opamps powered from ±17 V rails, we can assume an output capability of 10 Vrms. Scaling this down by the gain in each case gives us Table 8.2 which also shows the output level from a nominal 5 Vrms input. You will see shortly why the odd gain value of +45.5 dB was used.

Clearly if we want to accept a 64 mVrms input the gain cannot much exceed +40 dB. In fact a gain of +43.8 dB will just give clipping for 64 mVrms in. If we want to accept the Woody Herman 166 mVrms, then the maximum permissible gain is +35.6 dB. I would suggest that a safety margin of at least 5 dB should be added, so we conclude that 30 dB (1 kHz) is an appropriate gain for an MM input stage; this will accept 316 mVrms from a cartridge before clipping. A recent review of a valve phono stage [14] described an MM input capability of 300 mVrms as 'extremely generous', and I suggest that an input capability of around this figure will render you immune to overload for ever, and be more than adequate for the highest quality equipment. The stage output with a nominal 5 mVrms input is only 158 mVrms, which is not enough to operate your average power amplifier, and so there will have to be another amplifying stage after it. This must have variable gain, or be preceded by a passive volume control, for otherwise it will clip before the first stage and reduce the overload margin.

While we must have a relatively low gain in the MM stage to give a good maximum signal capability, we do not want it to be too low, or the signal/noise ratio is likely to be degraded as the signal passes through later stages. It is one of the prime rules of audio that you should minimise the possibility of this by getting the signal up to a decent level as soon as possible, but it is common practice and very sensible for the MM output to go through a unity-gain subsonic filter before it receives any further amplification; this is because the subsonic stuff coming from the disc can be at disturbingly high levels.

In the history of preamplifiers MM input overload margin used to be a test of machismo – this is less true now as the changeover to MC cartridges with a wide range of output levels makes a single input overload figure much less meaningful.

Moving-magnet cartridge sensitivities

Having looked at the limitations on the signal levels put onto disc, we need to see what levels we will get back when we replay it. The level reaching the preamplifier is clearly proportional to the cartridge sensitivity. Due to their electromagnetic nature, MM cartridges respond to stylus velocity rather than displacement (the same applies to MC cartridges), so output voltage is usually specified at a velocity of 5 cm/sec. That convention is followed throughout this chapter.

A survey of 72 MM cartridges on the market in 2012 showed that they fall into two groups – what might be called normal hifi cartridges, (57 of them) and specialised cartridges for DJ use (15 of them). The DJ types have a significantly higher output than the normal cartridges – the Ortofon Q-Bert Concorde produces no less than 11 mV, the highest output I could find. It seems unlikely that the manufacturers are trying to optimise the signal-to-noise ratio in a DJ environment, so I imagine there is some sort of macho 'my cartridge has more output than yours' thing going on. Presumably DJ cartridges are also designed to be exceptionally mechanically robust. We will focus here on the normal cartridges, but to accommodate DJ types all you only really need to do is allow for 6 dB more input level to the preamplifier.

The outputs of the 57 normal cartridges at a velocity of 5 cm/sec are summarised in the histogram of Figure 8.2. The range is from 3.0 mV to 8.0 mV, with significant clumps around 4–5 mV and 6.5 mV. If we ignore the single 8.0 mV cartridge, the output range is

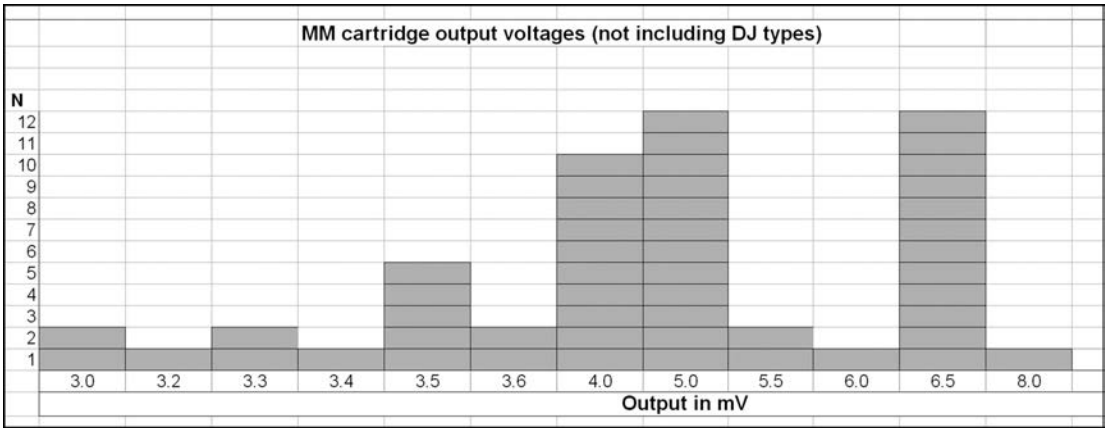

Figure 8.2: The output voltages for 57 MM cartridges, excluding specialised DJ types

restricted to 3.0 to 6.5 mV, which is only 6.7 dB. This is a very small range compared with the very wide one shown by MC cartridges, and makes the design of a purely MM input a simpler matter. There is no need to provide different amplifier sensitivities as a 6.7 dB range can be easily accommodated by adjustment of a volume control later in the audio path.

Overload margins and amplifier limitations

The safety factor between a nominal 5 mVrms input and the clipping point may be described as either the input headroom in mVrms, or the 'overload margin' which is the dB ratio between the nominal 5 mVrms input and the maximum input. Table 8.2 above shows that an MM stage with a +35 dB gain (1 kHz) gives an output of 280 mV, an input overload level of 178 mVrms and an overload margin of 31 dB, which might be called very good. A +30 dB (1 kHz) stage gives a nominal 158 mV out, an input overload level of 316 mVrms, and an overload margin of 36 dB, which is definitely excellent, giving 10 dB more headroom.

The maximum input capability of an MM stage is not always defined by simple frequency-independent clipping at its output. Things may be complicated by the stage output capability varying with frequency. An RIAA feedback network, particularly one designed with a relatively low impedance to reduce noise, presents a heavier load as frequency rises because the impedance of the capacitors falls. This heavy loading at HF was very often a major cause of distortion and headroom-limitation in discrete RIAA stages that had either common-collector or emitter-follower output topologies with highly asymmetrical drive capabilities; for example, an NPN emitter-follower is much better at sourcing current than sinking it. With conventional discrete designs, the 20 kHz output capability, and thus the overload margin, was often reduced by 6 dB or even more. Replacing the emitter resistor of an emitter-follower with a current-source much reduces the problem, and the very slight extra complication of using a push-pull Class-A output structure can bring it down to negligible proportions (for more details see Chapter 10 on discrete MM input design). Earlier opamps such as the TL072 also struggled to drive RIAA networks at HF, as well as giving a very poor noise performance. It was not until the advent of the 5532 opamp, with its excellent load-driving capabilities that the problem of driving low-impedance RIAA networks was solved; the noise performance was much better, too. However, if a low-impedance HF correction pole (more on this later) is being driven as well, there may still be some slight loss of output capability at 20 kHz.

Further headroom restrictions may occur when not all of the RIAA equalisation is implemented in one feedback loop. Putting the IEC Amendment roll-off after the preamplifier stage, as in Figure 8.5 below, means that very low frequencies are amplified by 3 dB more at 20 Hz than they otherwise would be, and this is then undone by the later roll-off. This sort of audio impropriety always carries a penalty in headroom as the signal will clip before it is attenuated, and the overload margin at 20 Hz is reduced by 3 dB. This effect reduces quickly as frequency increases, being 1.6 dB at 30 Hz and only 1 dB at 40 Hz. Whether this loss of overload margin is more important than providing an accurate IEC Amendment response is a judgement call, but in my experience it creates no trace of any problem in an MM stage with a gain of +30 dB (1 kHz). Passive-equalisation input architectures that put flat amplification before an RIAA stage suffer much more severely from this kind of headroom restriction, and it is quite common to encounter preamplifiers that claim to be high-end, with a high-end price-tag, but a very low-end overload margin of 20 to 22 dB. Bad show, chaps.

At any rate it is clear that we have to be careful not to unduly compromise the headroom at very low frequencies. We saw in the earlier section on spurious signals that Tomlinson Holman concluded that to accommodate the worst of the worst, a preamplifier should be able to accept not less than 35 mVrms in the 3–4 Hz region [7]. If the IEC Amendment is after the preamplifier stage, and C0 is made very large so it has no effect, the RIAA gain in the 2–5 Hz region has flattened out at 19.9 dB, implying that the equivalent overload level at 1 kHz will need to be 346 mVrms. The +30 dB (1 kHz) gain stage of Figure 8.5 has a 1 kHz overload level of 316 mVrms, which is only 0.8 dB below this rather extreme criterion; we are good to go. Using a +35 dB (1 kHz) gain stage instead would significantly reduce the safety margins.

At the other end of the audio spectrum, adding an HF correction pole after the preamplifier, to correct the RIAA response with low gains, also introduces a compromise in the overload margin, though generally a much smaller one. The 30 dB (1 kHz) stage in Figure 8.5 has a mid-band overload margin of 36 dB, which falls to +33 dB at 20 kHz. Only 0.4 dB of this is due to the amplify-then-attenuate action of the HF correction pole, the rest being due to the heavy capacitative loading on A1 of both the main RIAA feedback path and the pole-correcting RC network. This slight compromise could be eliminated by using an opamp structure with greater load-driving capabilities, so long as it retains the low noise of a 5534A.

An attempt has been made to show these extra preamp limitations on output level in Figure 8.1e above, and comparing 7.1d, it appears that in practice they are almost irrelevant because of the falloff in possible input levels at each end of the audio band.

To put all this into some sort of perspective, here are the 1 kHz overload margins for a few of my published designs. My first preamplifier, the Advanced Preamplifier [15], achieved +39 dB in 1976, partly by using all-discrete design and ± 24 V supply rails. A later discrete design in 1979 [16] gave a tour de force +47 dB, accepting over 1.1 Vrms at 1 kHz, but I must confess this was showing off a bit and involved some quite complicated discrete circuitry, including the push-pull Class-A output stages mentioned in Chapter 10. Later designs such as the Precision Preamplifier [17] and its linear descendant the Precision Preamplifier ’96 [18] accepted the limitations of opamp output voltage in exchange for much greater convenience in most other directions, and still have an excellent overload margin of 36 dB.

Equalisation and its discontents

Both moving-magnet and moving-coil cartridges operate by the relative motion of conductors and magnetic field, so the voltage produced is proportional to rate of change of flux. The cartridge is therefore sensitive to groove velocity rather than groove amplitude, and so its sensitivity is expressed as X mV per cm/sec. This velocity-sensitivity gives a frequency response rising steadily at 6 dB/octave across the whole audio band for a groove of constant amplitude. Therefore a maximal signal on the disc, as in Figure 8.1a, would give a cartridge output like Figure 8.1b, which is simply 8.1a tilted upwards at 6 dB/octave. From here on the acceleration limits are omitted for greater clarity.

The RIAA replay equalisation curve is shown in Figure 8.1c. It has three corners in its response curve, with frequencies at 50.05 Hz, 500.5 Hz, and 2.122 kHz, which are set by three time-constants of 3180 μs, 318 μs, and 75 μs. The RIAA curve was of USA origin but was adopted internationally with surprising speed, probably because everyone concerned was heartily sick of the ragbag of equalisation curves that existed previously (it became part of the IEC 98 standard, first published in 1964, and is now enshrined in IEC 60098, 'Analogue Audio Disk Records and Reproducing Equipment').

Note the flat shelf between 500 Hz and 2 kHz. It may occur to you that a constant downward slope across the audio band would have been simpler, required fewer precision components to accurately replicate, and would have saved us all a lot of trouble with the calculations. But such a response would require 60 dB more gain at 20 Hz than at 20 kHz, equivalent to 1000 times. The minimum open-loop gain at 20 Hz would have to be 70 dB (3000 times) to allow even a minimal 10 dB of feedback at that frequency, and implementing that with a simple two-transistor preamplifier stage would have been difficult if not impossible (must try it sometime). The 500 Hz–2 kHz shelf in the RIAA curve reduces the 20 Hz–20 kHz gain difference by 12 dB to only 48 dB, making a one-valve or two-transistor preamplifier stage practical. One has to conclude that the people who established the RIAA curve knew what they were doing.

When the RIAA equalisation of Figure 8.1c is applied to the cartridge output of Figure 8.1b, the result looks like Figure 8.1d, with the maximum amplitudes occurring around 1–2 kHz. This is in agreement with Holman's data [7].

Figure 8.1e shows some possible output level restrictions that may affect Figure 8.1d. If the IEC Amendment is implemented after the first stage, there is a possibility of overload at low frequencies which does not exist if The Amendment is implemented in the feedback loop by restricting C0. At the high end, the output may be limited by problems driving the RIAA feedback network which falls in impedance as frequency rises (there is more on this later).

The unloved IEC Amendment

Figure 8.1c shows in dotted lines an extra response corner at 20.02 Hz, corresponding to a time-constant of 7950 μs. This extra roll-off is called the 'IEC Amendment' and it was added to what was then IEC 98 in 1976. Its apparent intention was to reduce the subsonic output from the preamplifier, but its introduction is something of a mystery. It was certainly not asked for by either equipment manufacturers or their customers, and it was unpopular with both, with some manufacturers simply refusing to implement it. It still attracts negative comments today. The likeliest explanation seems to be that several noise reduction systems, for example dbx, were being introduced for use with vinyl at the time and their operation was badly affected by subsonic disturbances. None of these systems caught on.

On one hand it was pointed out that as an anti-rumble measure it was ineffective, as its slow first-order roll-off meant that the extra attenuation at 13 Hz, a typical cartridge-arm resonance frequency, was a feeble −5.3 dB; however at 4 Hz, a typical disc warp frequency, it did give a somewhat more useful −14.2 dB, reducing the unwanted frequencies to a quarter of their original amplitude. On the other hand there were loud complaints that the extra unwanted replay time constant caused significant frequency response errors at the low end of the audio band, namely −3.0 dB at 20 Hz and −1.0 dB at 40 Hz. Some of the more sophisticated equipment allows the Amendment to be switched in or out; a current example is the Audiolab 8000PPA phono preamplifier.

The 'Neumann pole'

The RIAA curve is only defined to 20 kHz, but by implication carries on down at 6 dB/octave forever. This implies a recording characteristic rising at 6 dB/octave forever, which could clearly endanger the cutting head if ultrasonic signals were allowed through. From 1995 a belief began to circulate that record lathes incorporated an extra unofficial pole at 3.18 μs (50.0 kHz) to limit HF gain. This would cause a loss of 0.17 dB at 10 kHz and 0.64 dB at 20 kHz, and would require compensation if an accurate replay response was to be obtained. The name of Neumann became attached to this concept simply because they are the best-known manufacturers of record lathes.

The main problem with this story is that it is not true. The most popular cutting amplifier is the Neumann SAL 74B which has no such pole. For protection against ultrasonics and RF it has instead a rather more effective second-order low-pass filter with a corner frequency of 49.9 kHz and a Q of 0.72 [19], giving a Butterworth (maximally flat) response rolling-off at 12 dB/octave. Combined with the RIAA equalisation this gives a 6 dB/octave roll-off above 50 kHz. The loss from this filter at 20 kHz is less than −0.1 dB, so there is little point in trying to compensate for it, particularly because other cutting amplifiers are unlikely to have identical filters.

Opamp MM disc input stages

Satisfactory discrete MM preamplifier circuitry is not that straightforward to design, and there is a lot to be said for using a good opamp, which, if well-chosen, will have more than enough open-loop gain to implement the RIAA bass boost without introducing detectable distortion at normal operating levels. The 5534/5532 opamps have input noise parameters that are well suited to moving-magnet (MM) cartridges. Having digested this chapter so far, we are in a position to summarise the requirements for a good RIAA preamplifier. These are:

1. Use a series feedback RIAA network, as shunt feedback is approximately 14 dB noisier.
2. Correct gain at 1 kHz. This sounds elementary, but getting the RIAA network right is not a negligible task.
3. Accuracy. My 1983 preamplifier was designed for ±0.2 dB accuracy from 20 Hz–20 kHz, the limit of the testgear I had at the time. This was tightened to ±0.05 dB without using rare parts in my 1996 preamplifier.
4. Use obtainable components. Resistors will often be from the E24 series, though E96 is much more available than it used to be. Capacitors will probably be from the E6 series, so intermediate values must be made by series or parallel combinations.
5. R0 must be as low as possible as its Johnson noise is effectively in series with the input signal. This is particularly important when the MM preamplifier is fed from a low impedance, which typically occurs when it is providing RIAA equalisation for the output of an MC preamplifier, rather accepting input direct from an MM cartridge with its high inductance.
6. The feedback RIAA network impedance to be driven must not be so low as to increase distortion or limit output swing, especially at HF.
7. The resistive path through the feedback arm should ideally have the same DC resistance as input bias resistor R18, to minimise offsets at A1 output. This is a bit of a minor point as the offset would have to be quite large to significantly affect the output voltage swing. Very often it is not possible to meet this constraint as well as other more important requirements.

Calculating the RIAA equalisation components

Calculating the values required for series feedback configuration is not straightforward. You absolutely cannot take Figure 8.3 and calculate the time-constants of R2, C2 and R3, C3 as if they were independent of each other; the answers will be wrong. Empirical approaches (cut-and-try) are possible if no great accuracy is required, but attempting to reach even ±0.2 dB by this route is tedious, frustrating, and generally bad for your mental health.

The definitive paper on this subject is by Stanley Lipshitz [20]. This heroic work covers both series and shunt configurations and much more besides, including the effects of low

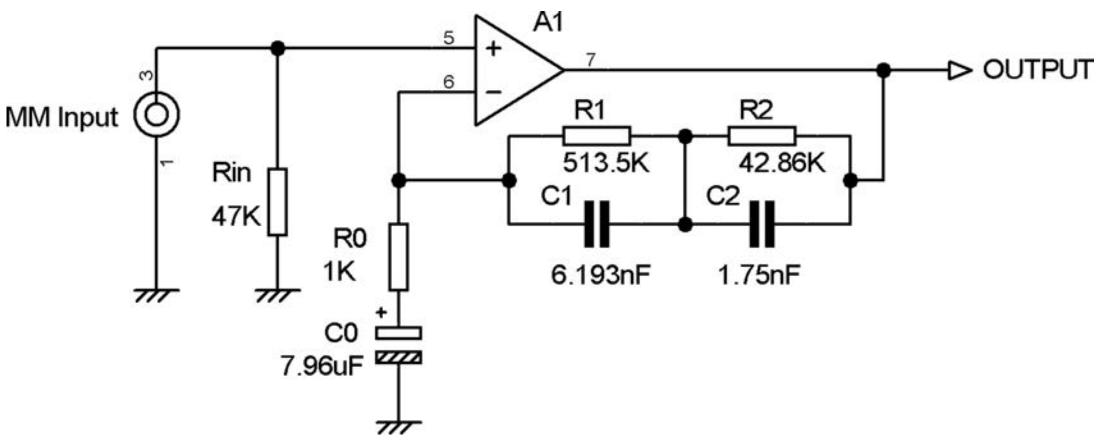

Figure 8.3: Series-feedback RIAA equalisation Configuration-A, IEC Amendment implemented by C0. Component values for 35.0 dB gain (1 kHz). Maximum input 178 mVrms (1 kHz). RIAA accuracy is within ±0.1 dB from 20 Hz to 20 kHz, without an HF correction pole

open-loop gain. It is relatively straightforward to build a spreadsheet using the Lipshitz equations that allows extremely accurate RIAA networks to be designed in a second or two; the greatest difficulty is that some of the equations are long and complicated – we're talking real turn-the-paper-sideways algebra here – and some very careful typing is required.

Exact RIAA equalisation cannot be achieved with preferred component values, and that extends to E24 resistors. If you see any single-stage RIAA preamp where the equalisation is achieved by two resistors and two capacitors in the same feedback loop, you can be sure it is not very accurate.

My spreadsheet model takes the desired gain at 1 kHz and the value of R0, which sets the overall impedance level of the RIAA network. In my preamplifier designs the IEC Amendment is definitely *not* implemented by restricting the value of C0; this component is made large enough to have no significant effect in the audio band, and the Amendment roll-off is realised in the next stage.

Implementing RIAA equalisation

It can be firmly stated from the start the best way to implement RIAA equalisation is the traditional series-feedback method. So-called passive (usually only semi-passive) RIAA configurations suffer from serious compromises on noise and headroom. For completeness they are dealt with towards the end of this chapter.

There are several different ways to arrange the resistors and capacitors in an RIAA network, all of which give identically exact equalisation when the correct component values are used. Figure 8.3 shows a series-feedback MM preamp built with what I call RIAA Configuration-A, which has the advantage that it makes the RIAA calculations somewhat easier, but otherwise is not the best; there will be much more on this topic later. Don't start building it until you've read the rest of this chapter; it gives an accurate RIAA response but is not otherwise optimised, as it attempts to represent a 'typical' design. We will optimise it later; we will lower the gain, and reduce the value of R0 to reduce its Johnson noise contribution, and the effect of opamp current noise flowing in it. Also note that details that are essential for practical use, like input DC-blocking capacitors, DC drain resistors, and EMC/cartridge-loading capacitors have been omitted to keep things simple; often the 47 kΩ input loading resistor is omitted as well. The addition of these components is fully described in the section on practical designs at the end of this chapter.

This stage is designed for a gain of 35.0 dB at 1 kHz, which means a maximum input of 178 mV at the same frequency. With a nominal 5 mVrms input at 1 kHz the output is 280 mV. The RIAA accuracy is within ± 0.1 dB from 20 Hz to 20 kHz, and the IEC Amendment is implemented by making C0 a mere 7.96 µF. You will note with apprehension that only one of the components, R0, is a standard value, and that is because it was used as the input to the RIAA design calculations that defined the overall RIAA network impedance. This is always the case for accurate RIAA networks. Here, even if we assume that capacitors of the exact value could be obtained, and we use the nearest E96 resistor values, systematic errors of up to 0.06 dB will be introduced. Not a long way adrift, it's true, but if we are aiming for an accuracy of ± 0.1 dB it's not a good start. If E24 resistors are the best available the errors grow to a maximum of 0.12 dB, and don't forget that we have not considered tolerances – we are assuming the values are exact. If we resort to the nearest E12 value, (which really shouldn't be necessary these days) then the errors exceed 0.7 dB at the HF end. And what about those capacitors?

The answer is of course that by using multiple components in parallel or series we can get pretty much what value we like, and it is perhaps surprising that this approach is not adopted more often. The reason is probably cost – a couple of extra resistors are no big deal but extra capacitors make more of an impact on the costing sheet. The use of multiple components also improves the accuracy of the total value, as described in Chapter 2 (there is more on this important topic later in this chapter).

The only drawback to the series-feedback RIAA configuration is what might be called the unity-gain problem. While the RIAA equalisation curve is not specified above 20 kHz, the implication is clear that it will go on falling indefinitely at 6 dB/octave. A series feedback stage cannot have a gain of less than unity, so at some point the curve will begin to level out and eventually become flat at unity gain; in other words there is a zero in the response. Figure 8.4 shows the various poles and zero frequencies of the circuit in Figure 8.3, with their associated time-constants. T3, T4 and T5 are the time-constants that define the basic RIAA

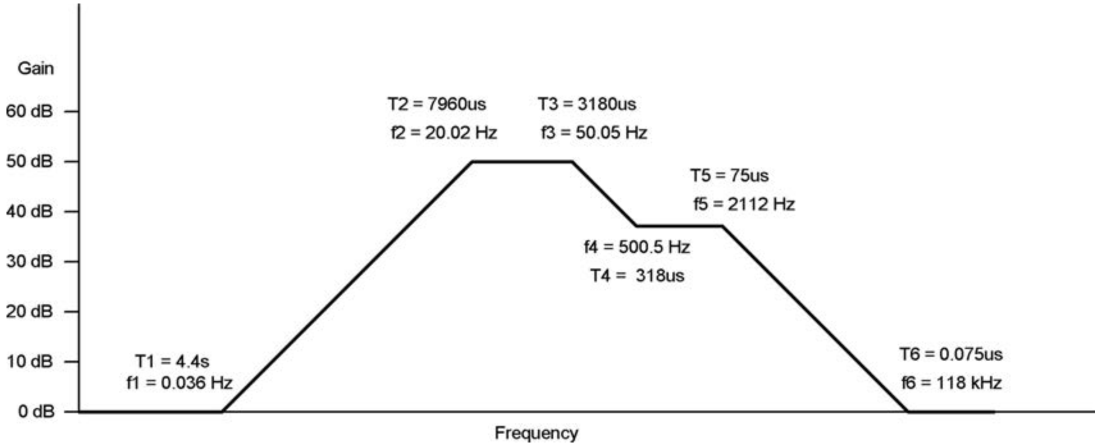

Figure 8.4: The practical response for series-feedback RIAA equalisation, including the IEC Amendment which gives an extra roll-off at 20.02 Hz

curve, while f3, f4 and f5 are the equivalent frequencies. This is the naming convention used by Stanley Lipshitz in his landmark paper, [20] and is used throughout this book. Likewise, T2 is the extra time-constant for the IEC Amendment, and T1 shows where its effect ceases at very low frequencies when the gain is approaching unity at the low frequency end due to C0. At the high end, the final zero is at frequency f6, with associated time-constant T6, and because the gain was chosen to be +35 dB at 1 kHz it is quite a long way from 20 kHz and has very little effect at this frequency, giving an excess gain of only 0.10 dB. This error quickly dies away to nothing as frequency falls below 20 kHz.

However, if the gain of the stage is set lower than +35 dB to maximise the input overload margin, the 6dB/octave fall tends to level out at unity early enough to cause significant errors in the audio band. Adding a HF correction pole (i.e. low-pass time constant) just after the input stage makes the simulated and measured frequency response exactly correct. It is not a question of bodging the response to make it roughly right. If the correction pole frequency is correctly chosen then its roll-off cancels *exactly* with the 'roll-up' of the final zero at f6.

An HF correction pole is demonstrated in Figure 8.5, where several important changes have been made compared with Figure 8.3. The overall impedance of the RIAA network has been reduced by making R0 220 Ω, to reduce Johnson noise from the resistors; we still end up with some very awkward values. The IEC Amendment is no longer implemented in this stage; if it was then the correct value of C0 would be 36.18 μF, and instead it has been made 220 μF so that its associated −3 dB roll-off does not occur until 3.29 Hz. Even this wide spacing introduces an unwanted 0.1 dB loss at 20 Hz, and perfectionists will want to use 470 μF here, which reduces the error to 0.06 dB. Most importantly, the gain has been reduced to +30 dB at 1 kHz to get more overload margin. With a nominal 5 mVrms input at 1 kHz

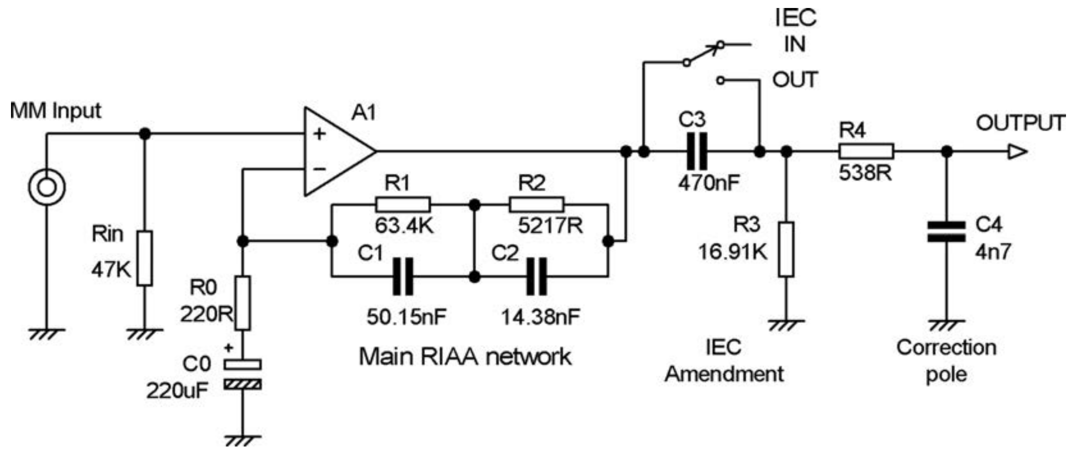

Figure 8.5: Series-feedback RIAA equalisation Configuration-A, redesigned for +30.0 dB gain (1 kHz) which allows a maximum input of 316 mVrms (1 kHz). R0 has been set to 220 Ω to reduce RIAA network impedance. The switchable IEC Amendment is implemented by C3, R3. HF correction pole R4,C4 is added to keep RIAA accuracy within ±0.1 dB, 20 Hz to 20 kHz.

the output will be 158 mV. The result is that the final zero f6 in Figure 8.3 is now at 66.4 kHz, much closer in, and it introduces an excess gain at 20 kHz of 0.38 dB, which is too much to ignore if you are aiming to make high-class gear. The HF correction pole R4, C4 is therefore added, which solves the problem completely. Since there are only two components, and no interaction with other parts of the circuit, we have complete freedom in choosing C4 so we use a standard E3 value and then get the pole frequency exactly right by using two resistors in series for R4–470 Ω and 68 Ω. Since these components are only doing a little fine tuning at the top of the frequency range, the tolerance requirements are somewhat relaxed compared with the main RIAA network. The design considerations are a) that the resistive section R4 should be as low as possible in value to minimise Johnson noise, and, on the other hand b) that the shunt capacitor C4 should not be large enough to load the opamp output excessively at 20 kHz. At this level of accuracy, even the finite gain open-loop gain of even a 5534 at HF has a slight effect, and the frequency of the HF pole has been trimmed to compensate for this.

Implementing the IEC Amendment

The unloved IEC Amendment was almost certainly intended to be implemented by restricting the value of the capacitor at the bottom of a series feedback arm, i.e. C0 in Figures 8.3 and 8.5. While electrolytic capacitors nowadays (2013) have relatively tight tolerances of ±20%, in the 1970's you would be more likely to encounter −20% +50%, the asymmetry reflecting the assumption that electrolytics would be used for non-critical coupling or decoupling purposes where too little capacitance might cause a problem, but more than expected would be fine. This wide tolerance meant that there could be significant errors in the LF response due to C0. Figure 8.6 shows the effect of a ±20% C0 tolerance on the RIAA response of a preamplifier similar to Figure 8.5, with a gain of +30 dB (1 kHz) and C0 = 36.13 μF. The gain will be +0.7 dB up at 20 Hz for a +20% C0, and −1.1 dB down at 20 Hz for a −20% C0. The effect of C0 is negligible above 100 Hz, but this is clearly not a good way to make accurate RIAA networks.

To get RIAA precision it is necessary to implement the IEC Amendment separately with a non-electrolytic capacitor, which can have a tolerance of ±1% if necessary. In several of my designs the IEC Amendment has been integrated into the response of the subsonic filter that immediately follows the RIAA preamplifier; this gives economy of components but means that it is not practicable to make it switchable in and out. Unless buffering is provided the series resistance in the HF correction network can interfere with the subsonic filter action, causing an early roll-off that degrades RIAA accuracy in the 20–100 Hz region.

The best solution is a passive CR high-pass network after the preamplifier stage. We make C0 large to minimise its effect, and add a separate 7950 μs time-constant after the preamplifier, as shown in Figure 8.5, where R3 and C3 give the required −3 dB roll-off at 20.02 Hz. Once again we can use a standard E3 capacitor value and 470 nF has been chosen here, and once again an unhelpful resistor value results; in this case 16.91 kΩ. However, with E24 values, this can be implemented exactly as 16 kΩ + 910 Ω (near-equal series or parallel resistor pairs would give more accuracy for a given tolerance). The switch as shown will not be entirely click-free because of the offset voltage at A1 output, but that is relatively unimportant as it will probably only be operated a few times in the life of the equipment.

When C0 is made large and the IEC Amendment is done later, we find C0 still has some effect. Since it is not infinite in value it will cause a roll-off of gain at some frequency. If we make C0 220 μF, which will be a handily compact component, there is an error of −0.128 dB at 20 Hz (assuming no IEC Amendment is used). If C0 is +20% high in value the error is reduced to −0.094 dB, and if it is −20% low the error is increased to −0.192 dB. Making C0 larger, such as 470 μF, reduces the basic error to a rather small −0.040 dB, and the variability due to its tolerance becomes negligible. 470 μF is a reasonable size at 6V3 rating, which is quite adequate for a component that is only exposed to opamp offset voltages. Going to 1000 μF or even 2200 μF starts to make significant demands on PCB area, and gains very little extra precision. This is summarised in Table 8.3. Most of the design examples in this chapter use C0 = 220 μF, but feel free to use 470 μF if you prefer; no other changes are required.

It is possible to compensate for the effect of C0 by tweaking the IEC Amendment. In Figure 8.5 the 220 μF value for C0 gives an error of −0.128 dB at 20 Hz. If R3 is changed

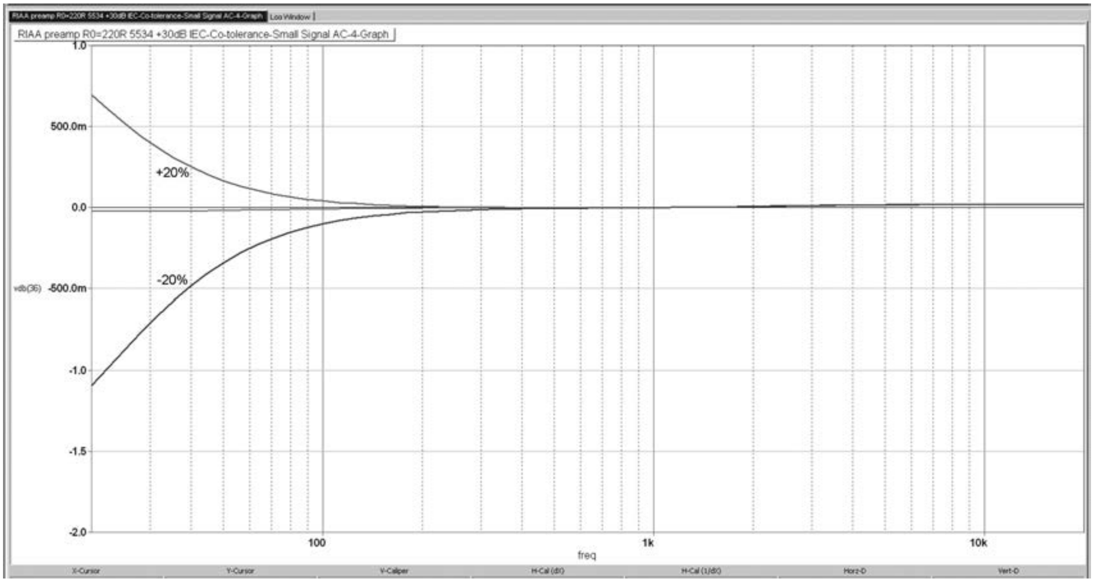

Figure 8.6: The effect of a ±20% tolerance for C0 when it is used to implement the IEC amendment

TABLE 8.3 Effect of C0 with ±20% tolerance on RIAA accuracy at 20 Hz. Preamp gain +30 dB, (1 kHz) R0 = 220R

Nominal C0 value µF	C0 nominal (dB)	C0 +20% (dB)	C0 −20% (dB)
100	−0.542	−0.385	−0.806
220	−0.128	−0.094	−0.192
470	−0.040	−0.032	−0.054
1000	−0.020	−0.019	−0.024
2200	−0.016	−0.015	−0.017

from 16.91 kΩ to 17.4 kΩ the overall response is made accurate to ±0.005 dB. The compensation is not mathematically exact – there is a + 0.005 dB hump around 20 Hz – but I suggest it is good enough for most of us. This process does not of course do anything to reduce the effects of the tolerance of C0, and is not usable if it is desired to make the IEC Amendment switchable in/out. If a subsonic filter is used, it is probably starting to take action at 20 Hz, and so small RIAA errors at this frequency are likely to be irrelevant.

The IEC network should come before the HF correction pole, as in Figure 8.5, so that R4 is not loaded by R3, which would cause a 0.3 dB loss; a small amount, perhaps, but you would have to recover it somewhere. Instead C3 is loaded by C4, but this has much less effect. The −0.3 dB figure assumes there is no significant external loading on the output at C4. Often the stage will be feeding the high-impedance input of a non-inverting gain stage, but if not some sort of buffering may be required so the two output networks behave as designed.

Another problem with the 'small C0' method of IEC Amendment is the non-linearity of electrolytic capacitors when they are asked to form part of a time-constant. This is described in detail in Chapter 2. Since the MM preamps of the Seventies tended to have poor linearity at LF anyway, because the need for bass boost meant a reduction in the LF negative feedback factor, introducing another potential source of distortion was not exactly an inspired move; on the other hand the signal levels are low. There is no doubt that even a simple second-order subsonic filter, switchable in and out, would be a better approach to controlling subsonic disturbances. If a Butterworth (maximally flat) alignment was used, with a −3 dB point at 20 Hz, this would only attenuate by 0.3 dB at 40 Hz, but would give a more useful −8.2 dB at 13 Hz and a thoroughly effective −28 dB at 4 Hz. Not all commentators are convinced that the more rapid LF phase changes that result are wholly inaudible, but they are; you cannot hear phase [21]. Subsonic filters are examined more closely at the end of this chapter.

RIAA equalisation by cartridge loading

It is possible to implement the LF-boost part of the RIAA characteristic by loading the cartridge inductance with a relatively low amplifier input resistance, giving a 6 dB/octave slope. As frequency increases the impedance of the inductance increases and the current into the input decreases.

This idea goes back a long way. The first use of it with a transistor amplifier I am aware of is in a 1961 preamplifier design by Tobey and Dinsdale [22], where the input stage was a single transistor with shunt feedback around it to implement the HF part of the RIAA curve; the cartridge was loaded with a 3.9 kΩ series input resistor rather than the standard 47 kΩ. There was another example in 1963 where a 6.8 kΩ input resistor was used with a two transistor shunt feedback amplifier. [23] The idea may well have been used in valve circuitry long before that. The notion has resurfaced many times since, most recently due to Bob Cordell in Jan Didden's *Linear Audio* [24].

Unfortunately there is a crippling snag: the LF equalisation now depends crucially on the cartridge inductance, so if you change your cartridge you have to redesign your preamplifier. You can, of course, make the loading variable, with a knob labelled 'cartridge inductance', but this assumes you actually know the cartridge inductance, and know it accurately. Any inaccuracy in dialling in the inductance is directly reflected in errors in the RIAA response. Most people would have to rely on the manufacturer's specification for inductance (a few do not specify it at all) and this is often quoted in suspiciously round figures. I also wonder how much the inductance varies between the two stereo channels. The normal range of MM cartridge inductance is from 400 mH to 800 mH, but you might come completely unstuck with unconventional cartridges like the moving-iron Grado Prestige series, which quotes an inductance of only 45 mH. The technique was criticised by Dinsdale in 1965, who acknowledged the problem of varying inductance, and also claimed that causing larger currents to flow through the cartridge degraded inter-channel crosstalk because of transformer action between the left and right coils [25].

The idea of loading the inductance by a resistance is not wholly worthless. It is used in conventional MM inputs, but in a less heavy-handed way. If the MM inductance is, say, 500 mH, in conjunction with the standard 47 kΩ loading resistor this gives a 6 dB/octave roll-off starting at 14.96 kHz. This is used by cartridge manufacturers to control HF resonances and flatten the top octaves of the frequency response. There is much more on MM cartridge inductance and its range of variation in Chapter 11 on MM-input noise, because it has a major effect on this area of performance.

It is worth noting that guitar pickups, which have substantial series inductance, are always operated into a high impedance of the order of 500 kΩ to 1 MΩ to avoid the loss of high frequencies. The capacitance of those long curly leads was often a problem, but now many guitarists use radio links.

RIAA series-feedback network configurations

There are four possible configurations described by Lipshitz in his classic paper [20]. These, with his component values, are shown in Figure 8.7; the same identifying letters have been used. They are all accurate to within ±0.1 dB when implemented with a 5534 opamp, but in the case of Figure 8.7A the error is getting close to −0.1 dB at 20 Hz due to the relatively high closed-loop gain (+46.4 dB at 1 kHz) and the finite open-loop gain of the 5534. All have RIAA networks at a relatively high impedance. They all have relatively high gain and therefore a low maximum input. The notation R0, C0, R1, C1, R2, C2 is as used by Lipshitz; C1 is always the larger of the two. In each case the IEC Amendment is implemented by the value of C0.

In recent years I have always used Configuration-A, mainly because long ago I wrote a design tool to implement the Lipshitz equations for it. I choose A simply because it was the easiest mathematical case. To repeat that for the other three configurations would be a significant amount of work, so the question arises, do any of the other three configurations have advantages that might make that work worthwhile?

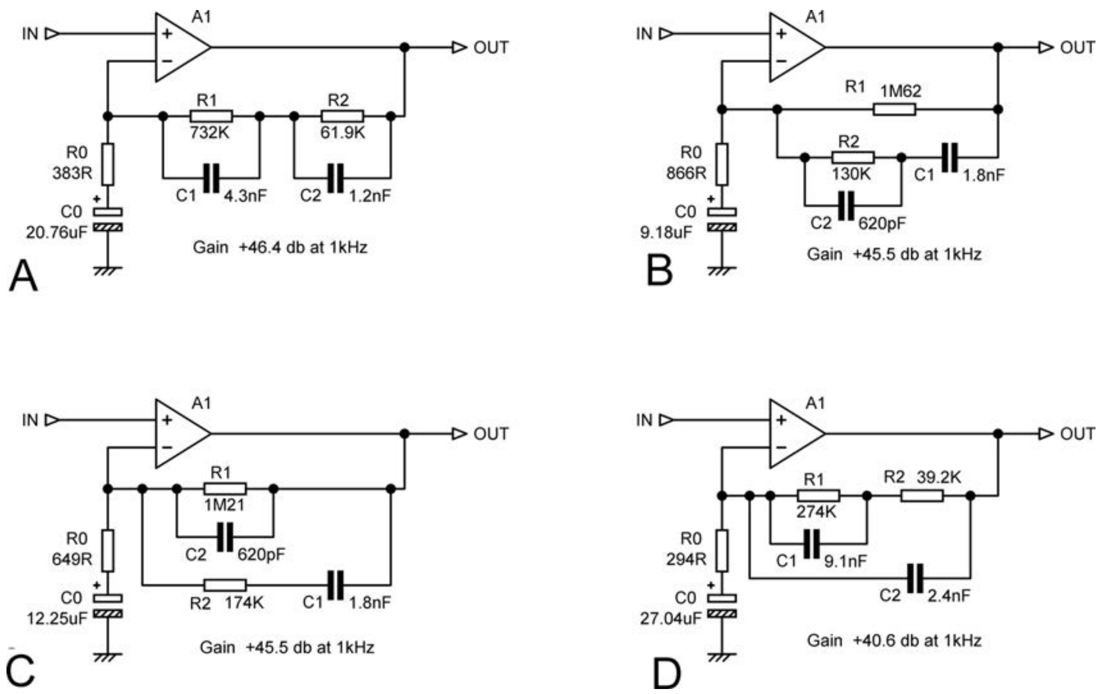

Figure 8.7: The four RIAA feedback configurations in the Lipshitz paper, identified by letter

Two things to examine come to mind:

Firstly, each configuration in Figure 8.7 contains two capacitors, a large C1 and a small C2, that set the RIAA response. If they are close-tolerance (to get accurate RIAA) and non-polyester (to prevent capacitor distortion) then they will be expensive, so if there is a configuration that makes the large capacitor smaller, even if it is at the expense of making the small capacitor bigger, it is well worth pursuing. The large capacitor C1 is probably the most expensive component in the RIAA MM amplifier by a large margin.

Secondly, the signal voltages across each capacitor are going to be different. If polyester capacitors must be used for cost reasons, then if there is a configuration that puts less voltage across a capacitor then that capacitor will generate less distortion. Capacitor distortion at least triples, and may quadruple, as the voltage across it doubles, so choosing the configuration that minimises the voltage is worthwhile.

RIAA configurations compared for capacitor cost

Looking at the capacitor sizing first, we need to put the four configurations A, B, C, and D into a form where they can be directly compared. Since in Figure 8.7 they are all working at different impedance levels, as shown by the differing values of R0, the first step is to scale all

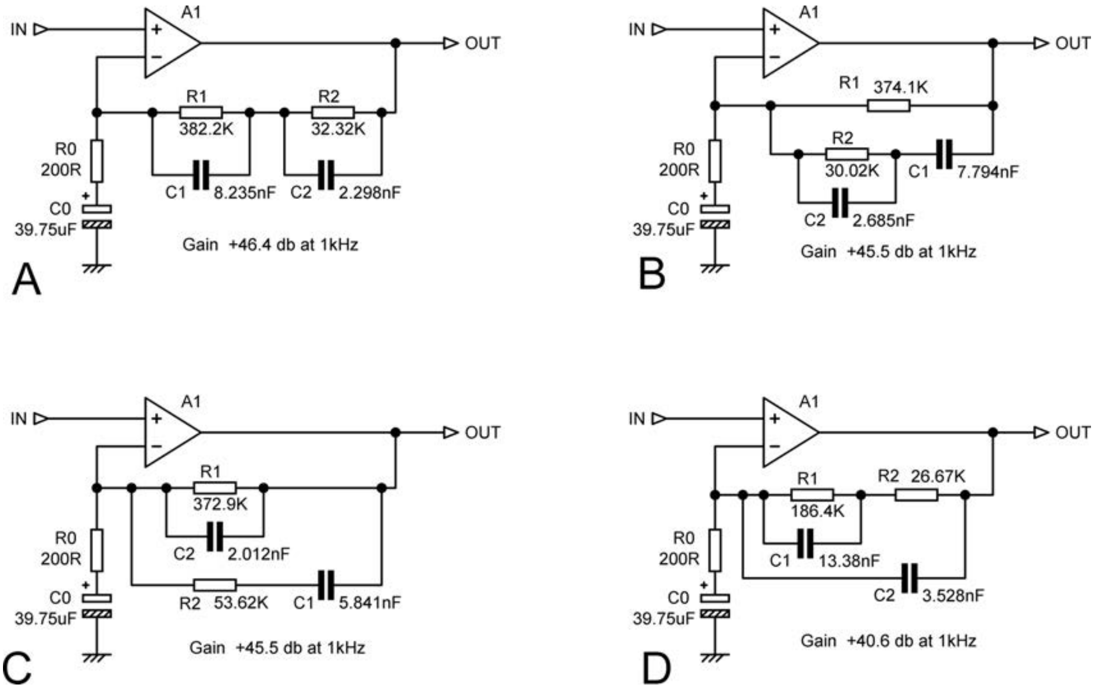

Figure 8.8: The four RIAA feedback configurations, with component values scaled so that R0 = 200 Ω in each case

TABLE 8.4 The values of C1 and C2 in Figure 8.8, with networks scaled so R0 = 200 Ω in each case

Configuration	Gain at 1 kHz (dB)	Large cap C1 (nF)	Small cap C2 (nF)	C1/C2 ratio
A	+46.4	8.235	2.298	3.583
B	+45.5	7.794	2.685	2.903
C	+45.5	5.841	2.012	2.903
D	+40.6	13.38	3.528	3.791

the RIAA component values to make R0 exactly 200 Ω, as in Figure 8.8. C0 then comes out as 39.75 μF in each case, and implements the IEC Amendment. The scaling does not affect the gain or the RIAA accuracy; this was checked by simulation for each configuration. The new capacitor values are summarised in Table 8.4

Fortunately the gains of A, B, and C are nearly the same, so we can compare the values for C1, and it looks as if A might actually be the worst case for capacitor size. To be certain about this we have to alter the gain of A to be exactly the same as B and C at +45.5 dB. Configuration-D has significantly bigger capacitors than A, B, and C because it has about half the gain but the same value of R0, so that gain also has to be altered. Changing the gain of an RIAA network is of course a non-trivial task, and we don't have the software tools for B, C, or D to do it quickly and with precision. We have to work out if it is worth writing one or more of those three tools.

Not having the tools, we can change the gain by simply scaling the RIAA network values of A and D, with R0 kept constant. We must accept that the results may not be very accurate but should be good enough for us to judge which configuration is superior. We need to reduce the gain of A by 0.899 dB, or a factor of 1.109 times; we therefore multiply the capacitors C1 and C2 by this factor, and divide the resistors R1 and R2 by it. For D, we want to increase the gain by 4.948 dB, or 1.767 times, so now we divide the capacitors C1 and C2 by this factor, and multiply the resistors R1 and R2 by it. This gives the values shown in Figure 8.9 and Table 8.5. The C1/C2 ratios are unchanged.

After this process, the Configuration-A, though less accurate than before, is still within a ±0.1 dB error band; a completely accurate version with the same gain was calculated directly from the Lipshitz equations and is shown in Figure 8.10A; note that the values of C1, C2 and R1, R2 are all slightly different, as you cannot change RIAA gain simply by scaling against R0 and get the exactly correct result. For Configuration-D, which

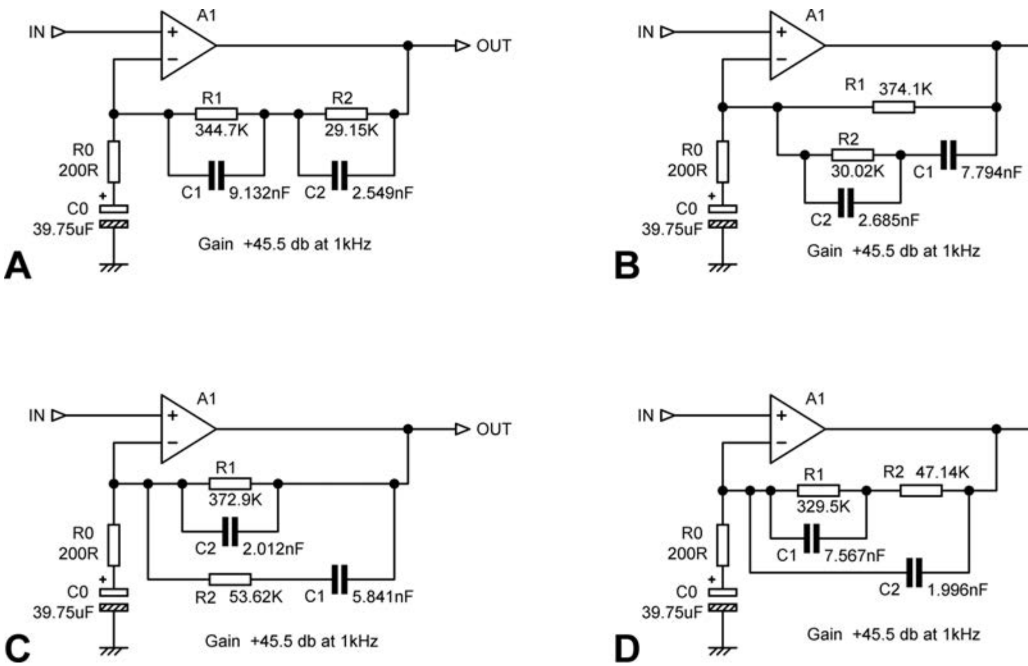

Figure 8.9: The RIAA feedback configurations, with component values scaled so that R0 = 200 Ω and the gain is +45.5 dB at 1 kHz in each case. Note that the RIAA response of A and D here is not wholly accurate. Maximum input in each case is only 53 mVrms (1 kHz) which is not generally adequate

TABLE 8.5 The values of C1 and C2 as in Figure 8.9, after scaling so R0 = 200 Ω and gain = +45.5 dB (1 kHz) for all configurations

Configuration	Gain at 1 kHz (dB)	Large cap C1 (nF)	Small cap C2 (nF)	C1/C2 ratio
A	+45.5	9.132	2.549	3.583
B	+45.5	7.794	2.685	2.903
C	+45.5	5.841	2.012	2.903
D	+45.5	7.567	1.996	3.791

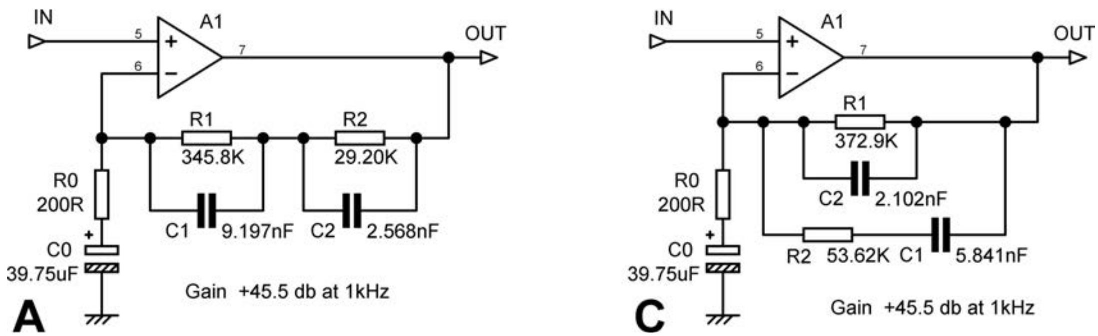

Figure 8.10: Configuration-A with values calculated from the Lipshitz equations to give accurate RIAA response. Configuration-C from Figure 8.8 shown for comparison; C1 in Configuration-A is much larger than C1 in Configuration-C, so the latter is superior. Gain +45.5 dB at 1 kHz for both

has undergone a greater gain change, the RIAA errors now exceed 0.1 dB (though not by much) at several frequencies. This is accurate enough to allow assessment of the configurations.

It is immediately obvious from Figures 8.9 and 8.10, and Table 8.5, that C1 in Configuration-C is only 64% the size of C1 in Configuration-A. I was afraid that this might be accompanied by an increase in C2 in Configuration-C, but this is also smaller at 81%. Unhappily it looks as if Configuration-A (which I have been using for years) makes the least efficient use of its capacitors, since they are effectively in series, reducing the effective value of both of them. Configurations B and D have intermediate values for C1, but of the two D has a significantly smaller C2. Configuration-C would appear to be the optimal solution in terms of capacitor size and hence cost. To design it accurately for gains other than +45.5 dB (1 kHz) meant building a software tool for it from the Lipshitz equations for Configuration-C. This I duly did, though, just as anticipated, it was somewhat more difficult than it had been for Configuration-A.

While Configuration-C in Figures 8.9 and 8.10 has come out as the most economical, our work here is not done. It will not have escaped you that a gain as high as +45.5 dB at 1 kHz is not going to give a great overload margin; it has only been used so far because it was the gain adopted in the Lipshitz paper. If we assume our opamp can provide 10 Vrms out, then the maximum input at 1 kHz is only 53 mVrms, which is mediocre at best. The gain of an MM input stage should not, in my opinion, much exceed 30 dB at 1 kHz (see the earlier example in Figure 8.5).

My Precision Preamplifier design [18] has an MM stage gain of +29 dB at 1 kHz, allowing a maximum input of 354 mVrms (1 kHz). The more recent Elektor Preamplifier 2012 [26] has an MM stage gain of +30 dB (1 kHz), allowing a maximum input of 316 mVrms; it is followed by a flat switched-gain stage which allows for the large range in MC cartridge sensitivity.

I used the new software tool for Configuration-C to design the MM input stage in Figure 8.11, which has a gain of +30 dB (1 kHz). This design has an RIAA response, including the IEC Amendment, that is accurate to within ±0.01 dB from 20 Hz to 20 kHz (it is assumed C0 is accurate). The relatively low gain means that an HF correction pole is required to maintain accuracy at the top of the audio band, and this is implemented by R3 and C3. Without this pole the response is 0.1dB high at 10 kHz, and 0.37 dB high at 20 kHz. R3 is a non-preferred value as we have used the E6 value of 2n2 for capacitor C3.

In Figure 8.11, and in the examples that follow, I have implemented the IEC Amendment by using the appropriate value for C0, rather than by adding an extra time-constant after the amplifier as in Figure 8.5. We noted above that using C0 is not the best method, but I have stuck with it here as it is instructive how the correct value of C0 changes as other alterations are made to the RIAA network.

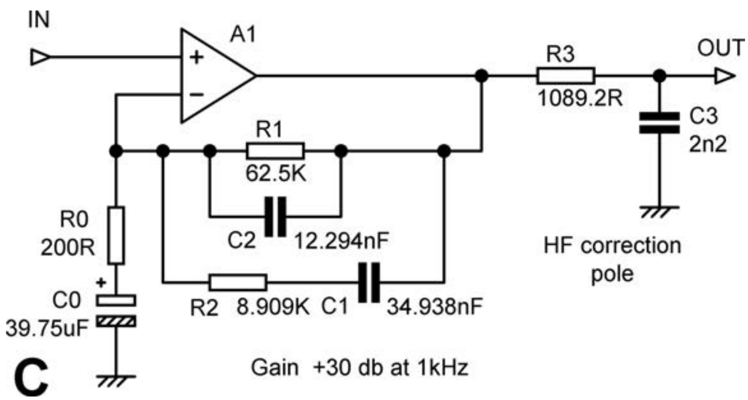

Figure 8.11: Configuration-C with values calculated from the Lipshitz equations to give +30.0 dB gain at 1 kHz, and an accurate RIAA response within ±0.01 dB; the lower gain now requires HF correction pole R3, C3 to maintain accuracy at the top of the audio band

RIAA network optimisation: C1 as a single E6 capacitor

Looking at Figure 8.11, a further stage of optimisation is possible after choosing the best RIAA configuration. There is nothing magical about the value of R0 at 200 Ω, (apart from the bare fact that it's an E24 value) it just needs to be suitably low for a good noise performance, so it can be manipulated to make at least one of the capacitor values more convenient, the larger one being the obvious candidate. Compared with the potential savings on expensive capacitors here, the cost of a non-preferred value for R0 is negligible. It is immediately clear that C1, at 34.9 nF, is close to 33 nF. If we twiddle the new software tool for Configuration-C so that C1 is exactly 33 nF, we get the arrangement in Figure 8.12. R0 has only increased by 6%, and so the effect on the noise performance will be quite negligible. All the values in the RIAA feedback network have likewise altered by about 6%, including C0, but the HF correction pole is unchanged; we would only need to alter it if we altered the gain. As for Figure 8.11, the RIAA accuracy of this version is well within ±0.01 dB from 20 Hz to 20 kHz when implemented with a 5534.

The circuit of Figure 8.12 has two preferred-value capacitors, but that is the most we can manage. All the other values are, as expected, thoroughly awkward. In Chapter 2 I described how to make up arbitrary resistor values by paralleling two or more resistors, and how the optimal way to do this is with resistors of as nearly equal values as you can manage. If the values are equal, then the tolerance errors partly cancel, and the accuracy of the combination is $\sqrt{2}$ times better than the individual resistors. The resistors are assumed to be E24, and the parallel pairs were selected using a specially written software tool. The three-part combination for C2, which I have assumed restricted to E6 values, was done by manual

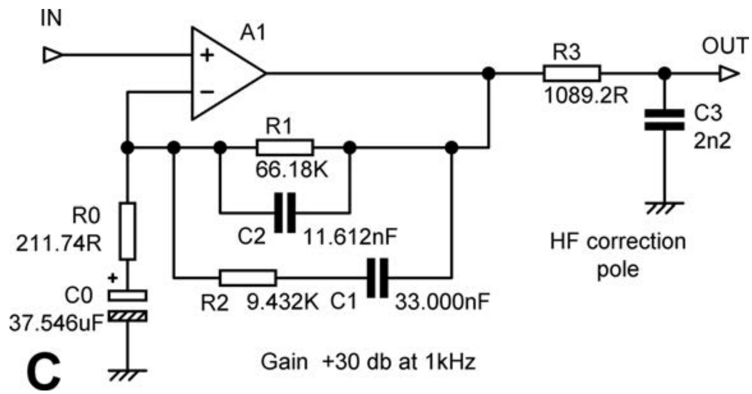

Figure 8.12: Configuration-C from Figure 8.10 with RO tweaked to make C1 exactly the E6 preferred value of 33.000 nF. Gain is still 30.0 dB at 1 kHz, and RIAA accuracy within ±0.01 dB. The HF correction pole R3, C3 is unchanged

TABLE 8.6 Approximation to the exact values in Figure 8.12 by using parallel components, giving Figure 8.13

Component	Desired value	Actual value	Parallel part A	Parallel part B	Parallel part C	Error
R0	211.74 Ω	211.03 Ω	360 Ω	510 Ω	–	−0.33 %
R1	66.18 kΩ	65.982 kΩ	91 kΩ	240 kΩ	–	−0.30 %
C1	33 nF	33 nF	33 nF	–	–	0 %
R2	9.432 kΩ	9.474 kΩ	18 kΩ	20 kΩ	–	+0.44 %
C2	11.612 nF	11.60 nF	4n7	4n7	2n2	−0.10 %
R3	1089.2 Ω	1090.9 Ω	2 kΩ	2.4 kΩ	–	+0.16 %

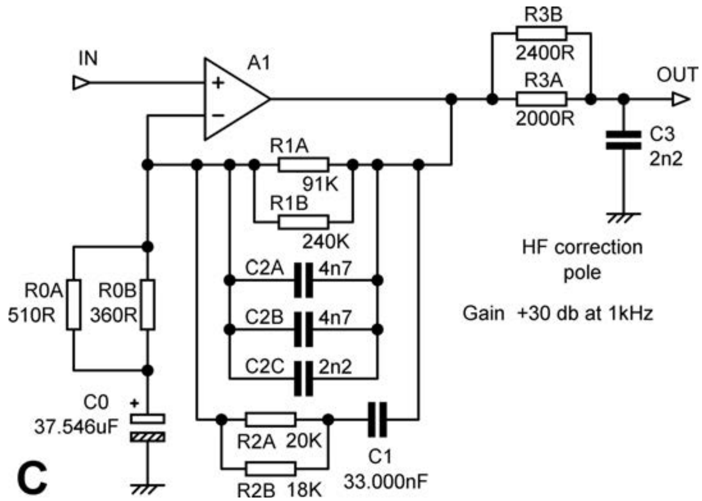

Figure 8.13: Configuration-C from Figure 8.14 with the resistors made up of optimal parallel pairs to achieve the correct value. C2 is now made up of three parts. Gain +30.05 dB at 1 kHz, RIAA accuracy is worsened but still within ±0.048 dB

bodging, though as you can see from Table 8.6, we have been rather lucky with how the values work out, with only three components getting us very close to the exact value we want. Figure 8.13 shows the resulting circuit.

The criterion used when selecting the parallel resistor pairs was that the error in the nominal value should be less than half of the component tolerance, assumed to be ±1%. R2 only just squeaks in, but its near-equal values will give almost all of the $\sqrt{2}$ improvement. Remember that in Table 8.6 we are dealing here with nominal values, and the % error in the nominal value shown in the rightmost column has nothing to do with the resistor tolerances.

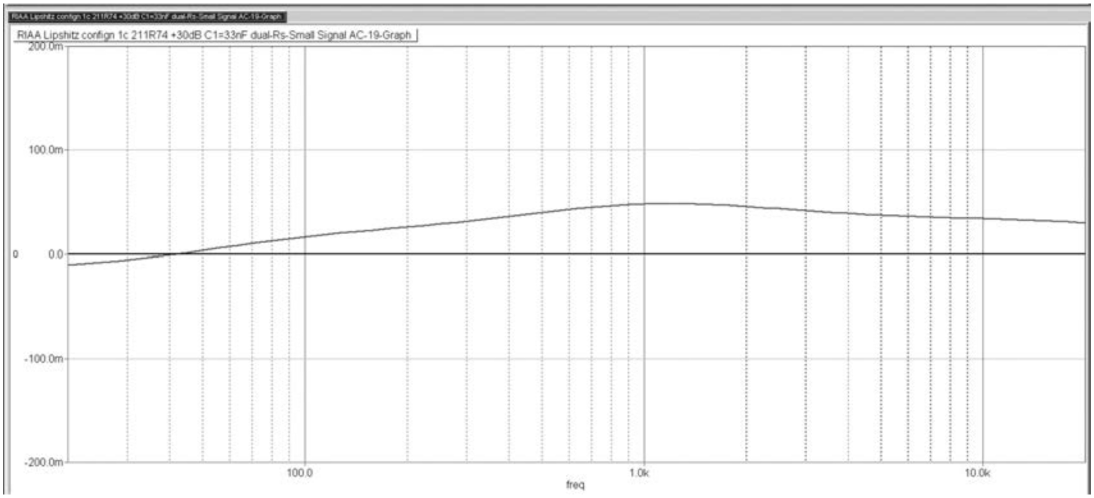

Figure 8.14: The RIAA accuracy of Figure 8.13. Gain is 30.05 dB at 1 kHz, and RIAA error reaches a maximum of +0.048 dB midband

No attempt has been made here to deal with the non-standard value for C0. In practice C0 will be a large value such as 220 µF so its wide tolerance will have no significant effect on RIAA accuracy. The IEC Amendment will be implemented (if at all) by a later time-constant using a non-electrolytic, as shown earlier in Figure 8.5.

Obviously the slight errors in nominal value seen in the rightmost column of Table 8.6 have some effect. Figure 8.14 shows that the gain at 1 kHz now peaks by +0.048 dB at 1 kHz, which is not the end of the world. By pure coincidence 1 kHz is actually where the RIAA accuracy is worst. At higher frequencies the error slowly declines to +0.031 dB at 20 kHz. Most of this deviation is caused by the +0.44% error in the nominal value of R2; greater accuracy could be got by using a three-resistor combination.

RIAA network optimisation: C1 as multiple 10 nF capacitors

We have just modified the RIAA network so that the major capacitor C1 is a single preferred value. The optimisation of the RIAA component values can be tackled in another way however; much depends on the initial assumptions about component availability. In many polystyrene capacitor ranges 10 nF is the highest value that can be obtained with a tolerance of 1%; in other cases the price goes up rather faster than proportionally above 10 nF. Paralleling several 10 nF polystyrene capacitors is *much* more cost-effective than using a single precision polypropylene part.

To use this method we need to redesign the circuit of Figure 8.12 so that C1 is either exactly 30 nF or exactly 40 nF (there is a practical design using Configuration-A with

5 × 10 nF = 50 nF at the end of this chapter, underlining the fact that Configuration-A makes less efficient use of its capacitance). The 40 nF version costs more than the 30 nF version but gives a total capacitance that is twice as accurate as one capacitor (because $\sqrt{4} = 2$), while the 30 nF version only improves accuracy by $\sqrt{3}$ (= 1.73) times. Using 40 nF gives somewhat lower general impedance for the RIAA network that may reduce noise very slightly. Figure 8.15 shows the result for C1 = 30 nF, and Figure 8.16 shows the result for C1 = 40 nF. Since the gain is unchanged the values for the HF correction pole R3, C3 are also unchanged in each case.

To turn Figure 8.15 into a practical circuit, the awkward component values are made up with optimally-selected parallel pairs. The results of this process for C1 = 30 nF are in Table 8.7 and Figure 8.16. In this case we have been unlucky with the value of C2, which needs to be trimmed with a 120 pF capacitor to meet the criterion that the error in the nominal value will not exceed half the component tolerance. This configuration has been built with 1% capacitors and measured, and it works exactly as it should. It gave a cost saving of about £2 on the product concerned.

The same process can be applied to the 4 × 10 nF version in Figure 8.17, giving the results in Table 8.8 and Figure 8.18.

This time we are much luckier with the value of C2; three 4n7 capacitors in parallel give almost exactly the required value. On the other hand we are very unlucky with R0, where 180 Ω in parallel with 6.2 kΩ is the most 'equal-value' solution that falls within our error criterion.

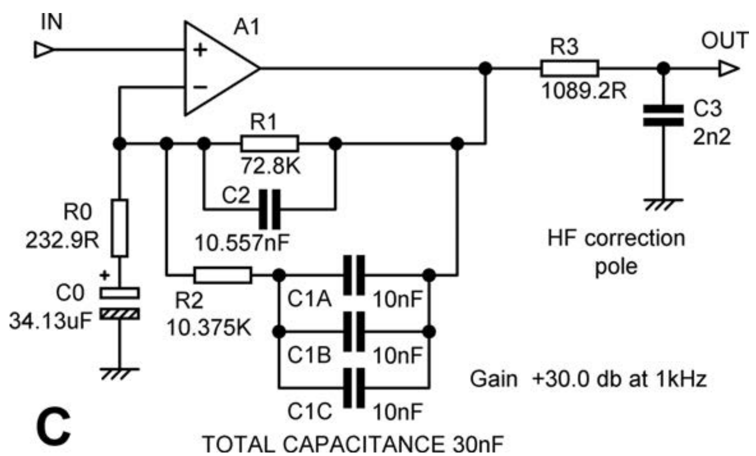

Figure 8.15: Configuration-C from Figure 8.11 redesigned so that C1 is 30 nF, made up with three paralleled 10 nF capacitors. Gain +30.0 dB at 1 kHz, RIAA accuracy is worsened but still within ±0.048 dB

TABLE 8.7 Approximation to the exact values in Figure 8.15 by using parallel components, giving Figure 8.16

Component	Desired value	Actual value	Parallel part A	Parallel part B	Parallel part C	Error
R0	232.9 Ω	233.3 Ω	430 Ω	510 Ω	–	−0.17%
R1	72.80 kΩ	72.97 kΩ	100 kΩ	270 kΩ	–	+0.24%
C1	30 nF	30 nF	10 nF	10 nF	10 nF	0%
R2	10.375 kΩ	10.359 kΩ	13 kΩ	51 kΩ	–	−0.15%
C2	10.557 nF	11.60 nF	4n7	4n7	1nF + 120 pF	−0.34%
R3	1089.2 Ω	1090.9 Ω	2 kΩ	2.4 kΩ	–	+0.16%

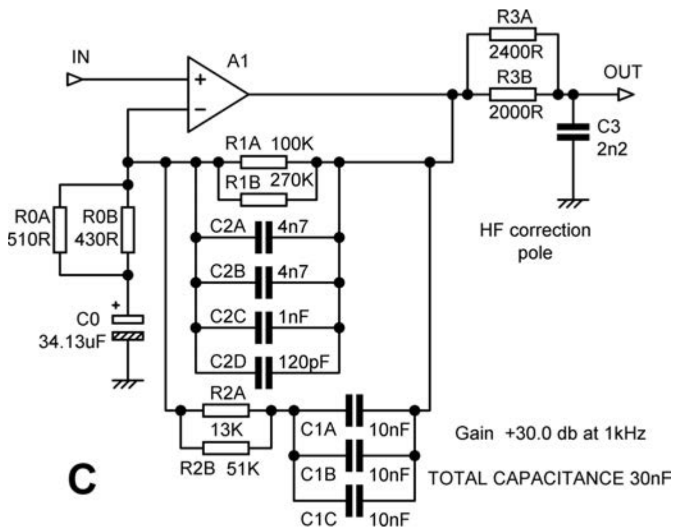

Figure 8.16: Configuration-C from Figure 8.15 with resistors made up of optimal parallel pairs. C2 is made up of four parts. Gain +30.0 dB at 1 kHz, RIAA accuracy is within ±0.01 dB

Both my Precision Preamplifier [18] and the more recent Elektor Preamplifier 2012 [26] have MM stage gains close to +30 dB (1 kHz), like the examples above, but both use Configuration-A, and five paralleled 10 nF capacitors are required.

Our investigations have shown that there are very real differences in how efficiently the various RIAA networks use their capacitors, and it looks clear that using Configuration-C rather than Configuration-A will cut the cost of the expensive capacitors C1 and C2 in an MM stage by 36% and 19% respectively, which I suggest is both a new result and well worth having. From there we went on to find that different constraints on capacitor availability lead to different optimal solutions for Configuration-C.

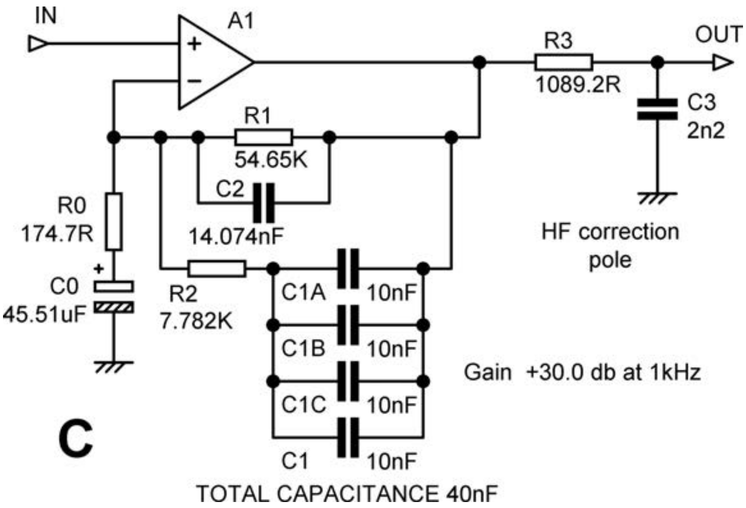

Figure 8.17: Configuration-C from Figure 8.11 redesigned so that C1 is 40 nF, made up with four paralleled 10 nF capacitors. Gain +30.0 dB at 1 kHz

TABLE 8.8 Approximation to the exact values in Figure 8.17 by using parallel components, giving Figure 8.18

Component	Desired value	Actual value	Parallel part A	Parallel part B	Parallel part C	Parallel part D	Error
R0	174.7 Ω	174.9 Ω	180 Ω	6.2 kΩ	–	–	+0.1%
R1	54.44 kΩ	54.54 kΩ	100 kΩ	120 kΩ	–	–	−0.41%
C1	40 nF	40 nF	10 nF	10 nF	10 nF	10 nF	0%
R2	7.821 kΩ	7.765 kΩ	12 kΩ	22 kΩ	–	–	−0.22%
C2	14.074 nF	14.1 nF	4n7	4n7	4n7	–	+0.18%
R3	1089.2 Ω	1090.9 Ω	2 kΩ	2.4 kΩ	–	–	+0.16%

It seems likely that further optimisation of the RIAA networks shown here is possible. For example, we noticed that changing R0 from 200 Ω to 211.74 Ω had a negligible effect on the noise performance; worse by only 0.02 dB. That is well below the limits of measurement, and we could ask what happens if we grit our teeth and accept a 0.1 dB noise deterioration? That is still at or below most measurement limits. It implies that R0 is 270 Ω, and the RIAA network impedance is therefore increased by 35%, so we could for example omit one of the 10 nF capacitors in Figure 6.18, with of course suitable adjustments to all the other circuit values, and save some more of our hard-earned money.

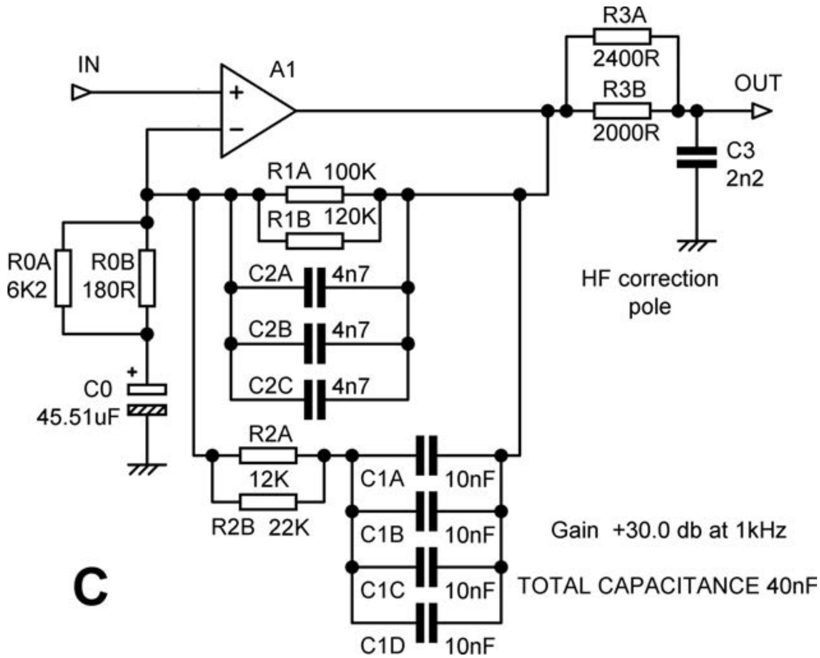

Figure 8.18: Configuration-C from Figure 8.17 with resistors made up of parallel pairs. C2 is made up of three parts. Gain +30.0 dB at 1 kHz, RIAA accuracy is within ±0.01 dB

I hope you will forgive me for not making public the software tools mentioned in this article. They are part of my stock-in-trade as a consultant engineer, and I have invested significant time in their development.

RIAA configurations compared for capacitor voltages

Turning to our second issue – the signal voltages across the capacitors in each configuration – a good deal of simulation tells us that there is very little to choose between the four configurations as regards the signal voltage across the capacitors. Not a helpful result or an interesting result, but sometimes things just are that way.

A related but different question is: if we assume a certain amount of non-linearity in one or both of the RIAA capacitors, are the configurations different in their sensitivity to that non-linearity? In other words, how much distortion will appear at the output? This question could be resolved in simulation, by using non-linear capacitor models constructed with Analog Behavioural Modelling, but it would be a lot of work, and since the emphasis of this book is on high quality, where we can presumably afford a polypropylene capacitor or two, I have put that one on the back-burner. Indeed, it may fall completely off the back of the cooker.

Equivalent RIAA configurations

You may be wondering if the four configurations in Figure 8.9 actually cover all possible single-stage series-feedback RIAA arrangements. The configurations can be differently arranged as if two components are in series, with nothing connected to their junction point, it does not matter in which order they occur. In Figure 8.19 the configurations A and A' are electrically identical. This looks pretty obvious, but it is perhaps a bit less so for B B', C C', and D D', in which the identical topology can be drawn in several different ways. Burkhard Vogel, in his monumental book on noise *The Sound of Silence,* describes Configuration-A as a Type-Eub network, Configuration-B' as a Fub-B network, and Configuration-C as a Fub-A network [27]. Configuration-D is not examined.

RIAA components

Many of the factors affecting the choice of components for the RIAA network, such as accuracy, linearity, and cost have already been dwelt on at length. Here are a few more points to ponder:

- Resistors should be metal film for good linearity, with two or more near-equal E24 values paralleled to obtain the non-preferred resistance values. See Chapter 2.
- For close-tolerance capacitors the best solution seems to be axial polystyrene types which are freely available at 1% tolerance up to 10 nF. Some paralleling is required, and in fact is highly desirable. This is because the sum of multiple capacitors is more accurate than a single component of the same tolerance, so long as the mean is well controlled, because the capacitances sum arithmetically but the random errors partially cancel. This is described in detail in Chapter 2. Both the preamp examples at the end of this chapter use multiple capacitors in this way.
- Polystyrene capacitors have two foils, and one of them will be on the outside of the component, and so vulnerable to capacitive crosstalk and hum pickup. It is desirable that the capacitor is orientated so that the outer foil is connected to the circuit node with the lowest impedance; very often this can be arranged to be the stage output, which is at a very low impedance and immune to capacitive pickup. Some capacitor manufacturers mark the outer foil; for example, the outer foil of polystyrene capacitors manufactured by LCR is indicated by the mitred corner on the packaging. Inexpensive polystyrene caps may not have consistent foil placement. Other types of capacitor, such as polypropylene, have similar considerations. With axial types the outer foil is likely to be marked by a line at one end, if indeed it is marked at all. The outer foil of a capacitor can be quickly identified with an oscilloscope. Ground one lead and put the probe on the other, and see how much hum the capacitor picks up from your fingers. Reverse the connections and repeat. When the outer foil is grounded, much less hum is picked up.

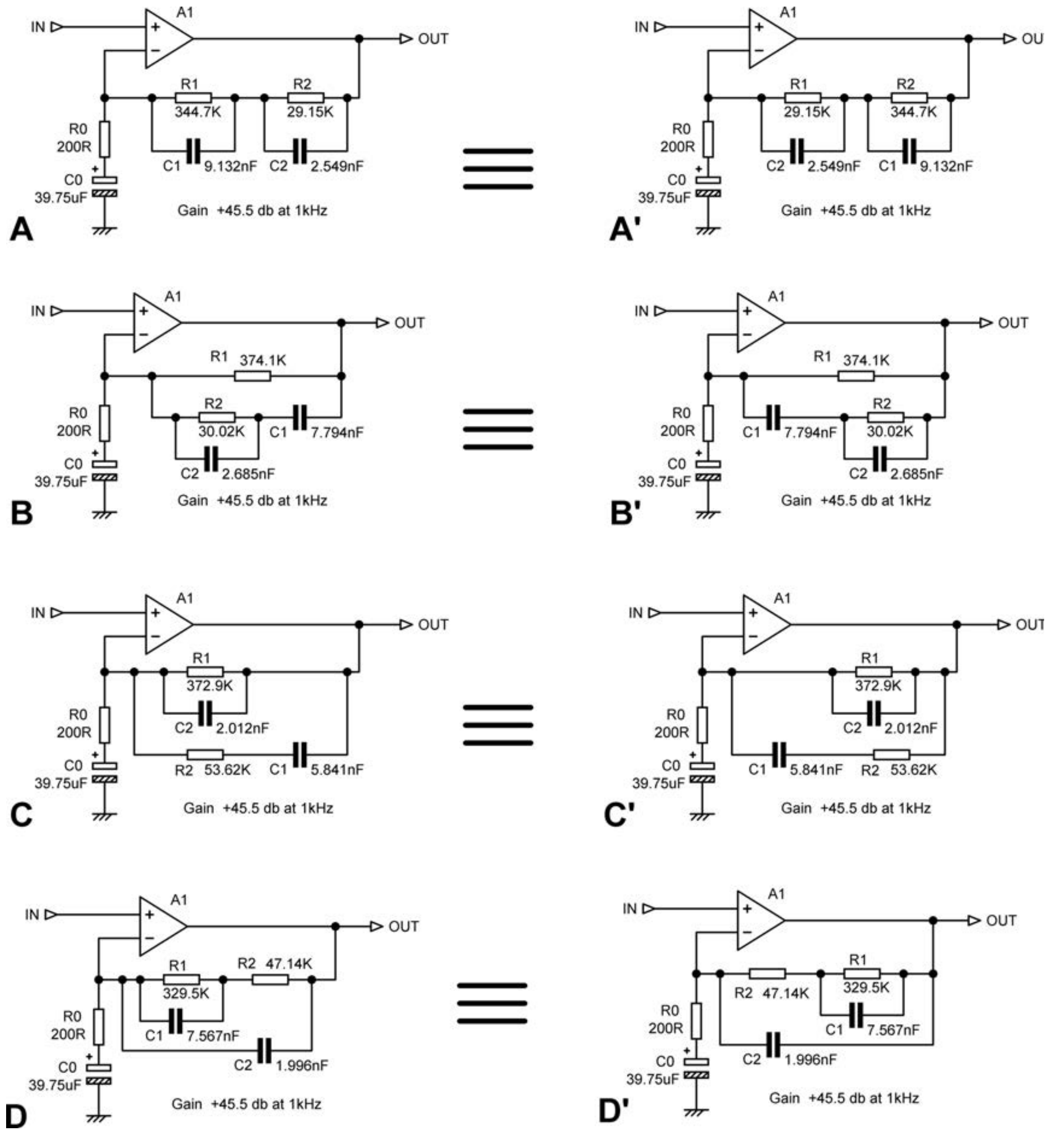

Figure 8.19: Equivalent RIAA feedback configurations

RIAA component sensitivity: Configuration-A

The 'component sensitivity' of a circuit defines how much its response varies as a result of component value tolerances. It has nothing to do with gain or signal levels. It is much affected by the way the circuit works – the higher the Q of an active filter, the greater its component sensitivity.

Components have tolerances on their value, and we need to assess what RIAA accuracy is possible without spending a fortune on precision parts; ±1% is the best tolerance readily available for metal-film resistors and polystyrene capacitors, so at first it appears anything better than ±0.1 dB accuracy is out of the question. This is not so, for the simple reason that across the audio band more than one component determines the response. Higher precision can be obtained by using multiple components, as we have noted before.

We saw above that Configuration-C was more economical than Configuration-A. We might wonder whether we pay for that in increased component sensitivity for Configuration-C. Sensitivity analysis can be done by involved mathematics, or more simply by making ±1% changes in the components of a preamp simulation. The results in Figures 8.20 and 8.21 include the IEC Amendment, implemented by C0. The effect of C0 tolerances on accuracy has already been examined in this chapter.

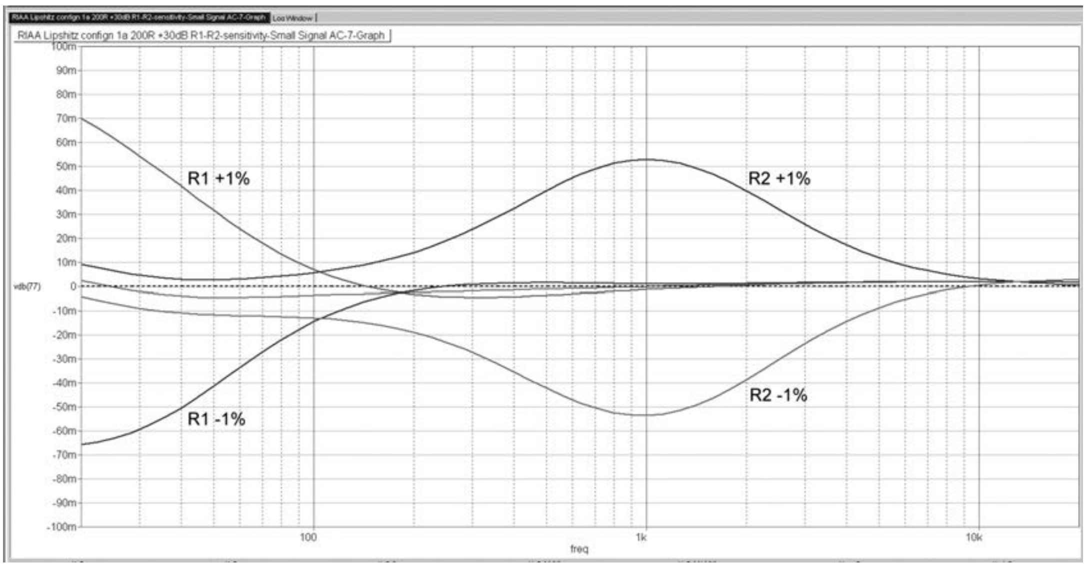

Figure 8.20: Configuration-A: the effect on RIAA accuracy of separate ±1% changes in R1 and R2. Maximum errors are 0.070 dB for R1, and 0.053 dB for R2

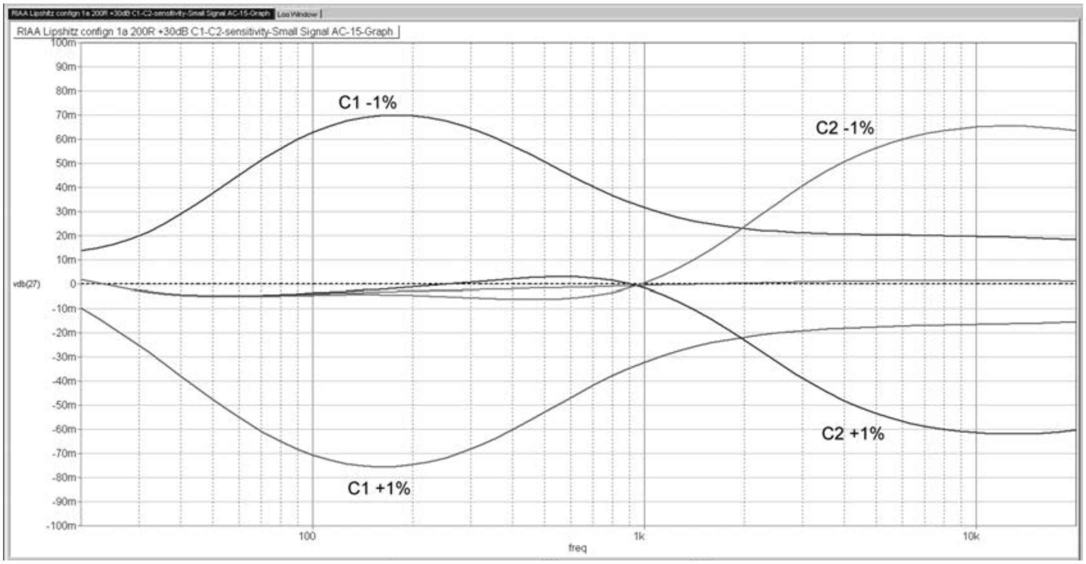

Figure 8.21: Configuration-A: the effect on RIAA accuracy of separate ±1% changes in C1 and C2. Maximum errors are 0.070 dB for C1, and 0.065 dB for C2

TABLE 8.9 Maximum errors for 1% deviation in components values for Configurations A and C

Configuration	R1 (dB)	R2 (dB)	C1 (dB)	C2 (dB)
A	0.070	0.053	0.070	0.065
C	0.075	0.04	0.052	0.082

RIAA component sensitivity: Configuration-C

Figures 8.22 and 8.23 show the same results but for Configuration-C. The component sensitivities for both configurations are summarised in Table 8.9, which shows that there is not much to choose between them. Configuration-A is better for C2, but Configuration-C is better for R2 and C1. There is no reason here to not use Configuration-C.

Open-loop gain and RIAA accuracy

There is no point in having a super-accurate RIAA network if the active element does not have enough open-loop gain to correctly render the response demanded. This was a major problem for two- and three-transistor discrete MM input stages, but one might have hoped that it would have disappeared with the advent of usable opamps. However, life is flawed, and gain problems

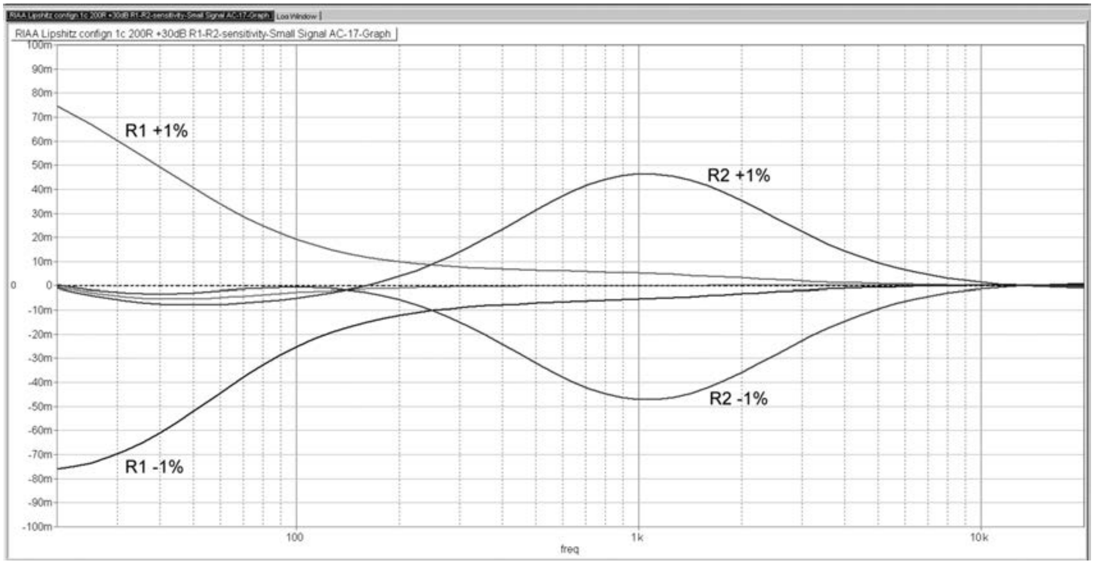

Figure 8.22: Configuration-C: the effect on RIAA accuracy of separate ±1% changes in R1 and R2. Maximum errors are 0.075 dB for R1, and 0.047 dB for R2. Similar to Configuration-A except that R2 has very little effect below 200 Hz

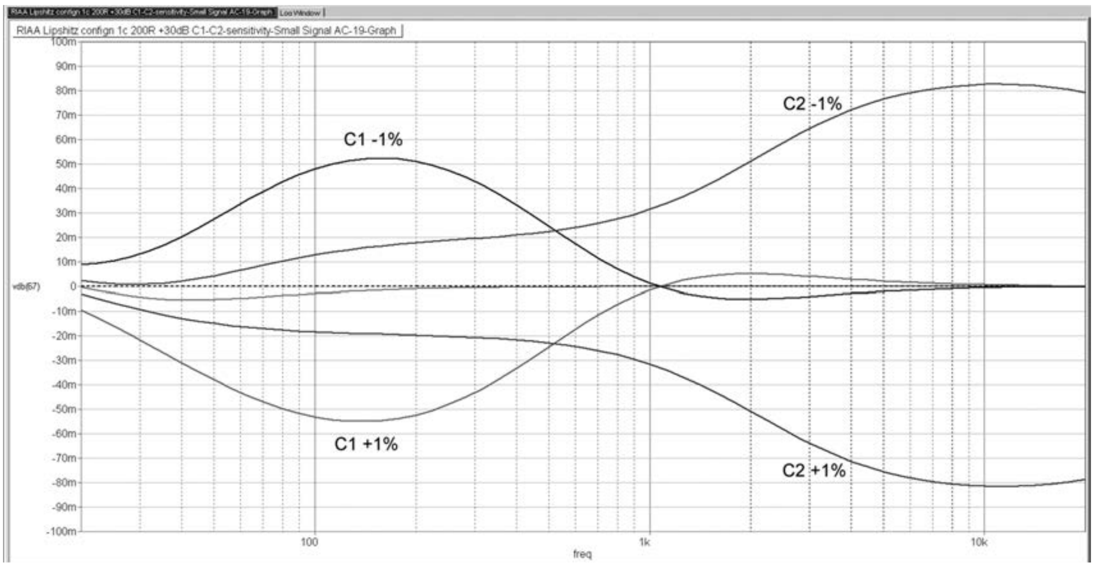

Figure 8.23: Configuration-C: the effect on RIAA accuracy of separate ±1% changes in C1 and C2. Maximum errors are 0.052 dB for C1, and 0.082 dB for C2. Very similar to Configuration-A except that C2 has significant effect below 1 kHz

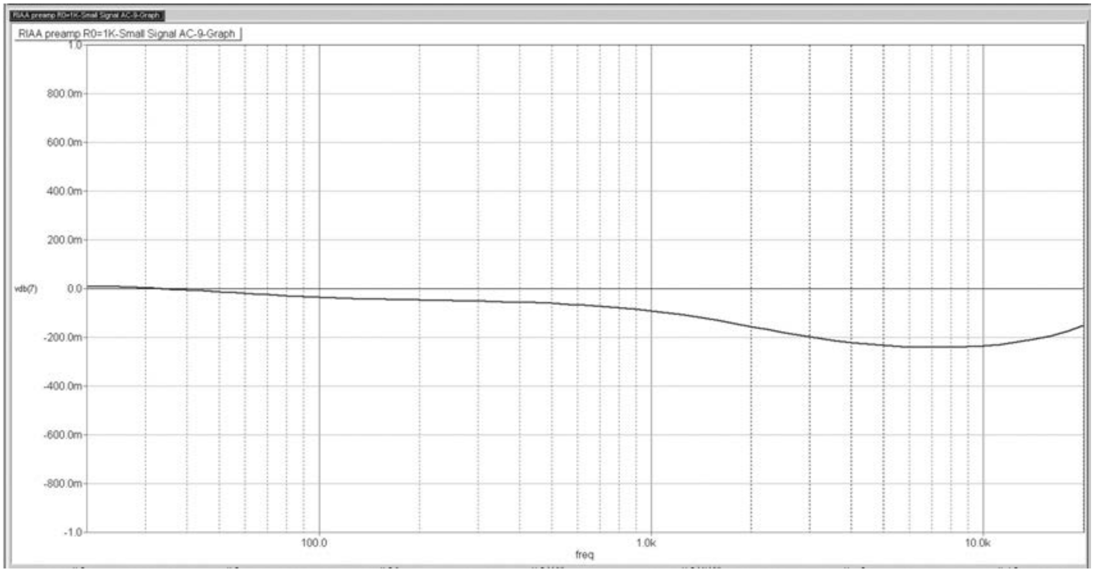

Figure 8.24: RIAA error using a TL072 in a +35 dB preamp; lack of open-loop gain causes a 0.2 dB dip between 3 and 10 kHz. Scale +1 dB

did not wholly vanish. The TL072 was at one time widely used for MM inputs because of its affordability, even though its JFET input devices are a poor match to MM cartridge impedances, and its distortion performance was not of the best. However, there was another lurking problem.

As we have seen earlier in this chapter, the appropriate gain (at 1 kHz) for an MM input is between +30 and +40 dB. The TL072 does not have enough open-loop gain to give an accurate response with a closed-loop gain of a +35 dB. Figure 8.24 shows the result of a simulation using Configuration-A RIAA with accurate values derived from the Lipshitz equations. There is a 0.2 dB dip between 3 and 10 kHz; the vertical scale is ±1 dB. In the simulations that follow, the IEC Amendment is not implemented.

Replacing the TL072 with a 5534 opamp, which has more open-loop gain, reduced the RIAA error to much less than 0.1 dB across the audio band. The TL072 has an LF gain of 200,000 times and a dominant pole at 20 Hz (all typical specs). The 5534 has a lower LF gain of 100,000 but its pole (uncompensated) is much higher at 1 kHz, so at any frequency above 40 Hz the 5534 has more open-loop gain to offer. At all frequencies above 1 kHz the 5534 has 30 dB more gain than the TL072. These parameters are subject to production variations. Figure 8.25 shows the RIAA accuracy using a 5534A in a +35 dB (1 kHz) gain preamp; note that the amplitude scale has changed from ±1 dB down to ±0.2 dB. The RIAA error is negligible (less than ±0.01 dB) above 100 Hz, but reaches a maximum of −0.025 dB at 30 Hz. The rising response error at the HF end (+0.12 dB at 20 kHz) is due to the gain

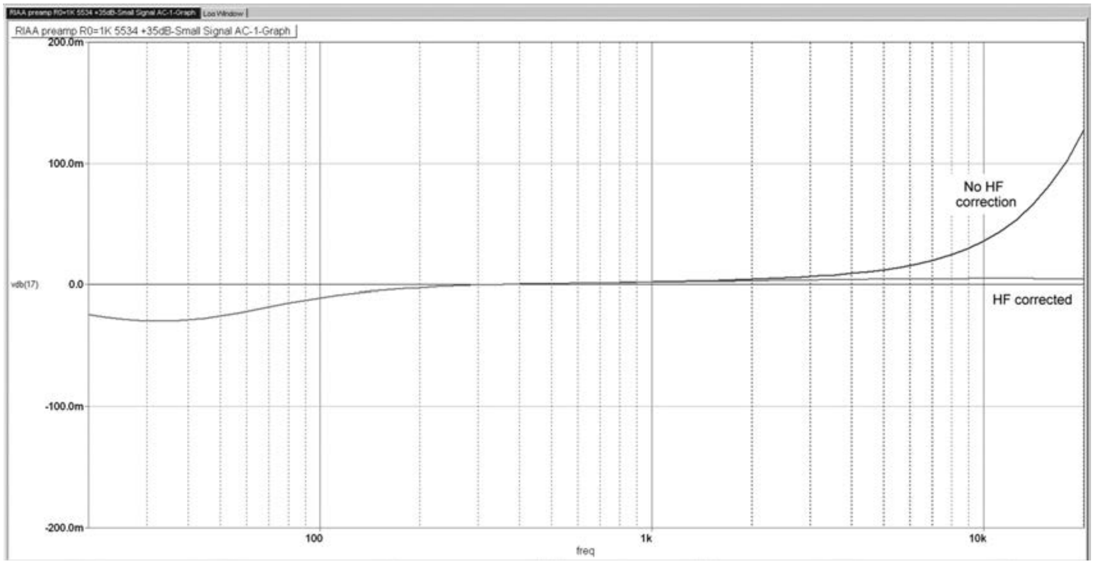

Figure 8.25: RIAA error using a 5534A in a +35 dB preamp, with and without HF correction pole. Scale ±0.2 dB

levelling off at f6, but this is here cancelled by an HF correction pole that reduces the HF error to less than ±0.01 dB.

Figure 8.26 shows the same results for a 5534 in a +30 dB (1 kHz) preamp. HF accuracy is better, but the LF error is not much reduced, being −0.020 dB at 30 Hz, indicating it may not be due to a lack of open-loop gain. This error can be reduced to −0.005 dB by adjusting the value of R1 upwards by trial and error, no other RIAA components being altered. One wonders if there might be a small systematic error buried in the Lipshitz equations, or more likely in my application of them. The uncorrected HF response has a greater error of 0.37 dB at 20 kHz, as the gain is levelling off towards unity at a lower f6 frequency, but this is once more fully sorted out by the HF correction pole.

The 5534 has quite a complicated internal structure with nested Miller loops for compensation. It is not obvious (to me, anyway) whether this has anything to do with the interaction of open-loop gain with RIAA accuracy. It seemed worthwhile to do a few more simulations to get a better idea of the situation. The SPICE opamp model is replaced with the simplest possible conceptual opamp. This has only two parameters – the LF open-loop gain and the dominant-pole frequency – and is modelled by a voltage-controlled-voltage source (VCVS) with a gain of 100,000 times, combined with a first-order RC filter that is −3 dB at 100 Hz.

We will start off with a +40 dB (1 kHz) preamp. The simulation results are shown in Figure 8.27; the maximum errors are −0.09 dB at 150 Hz and −0.16 dB at 20 kHz.

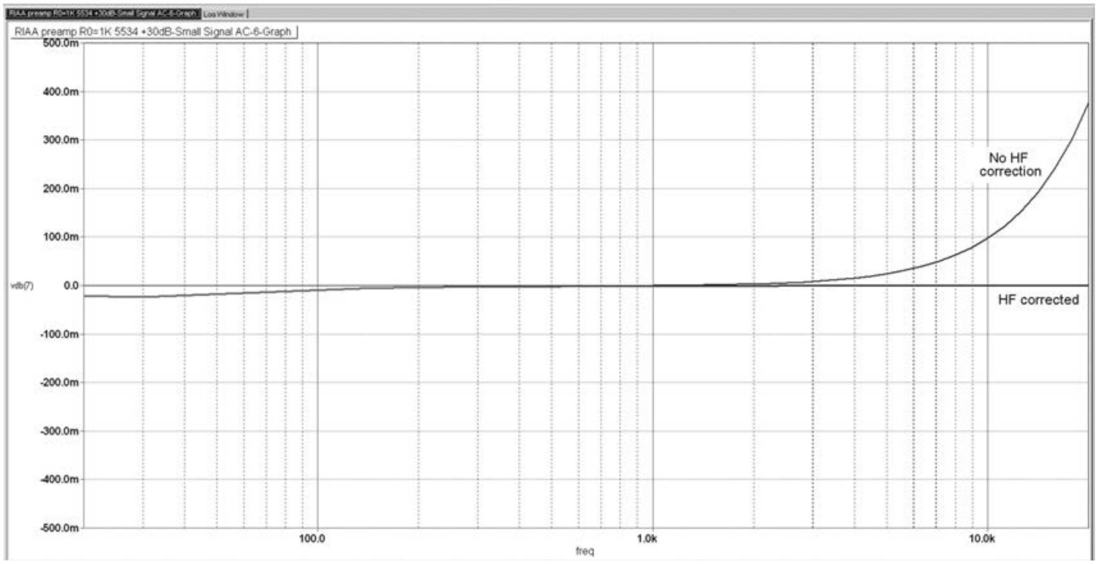

Figure 8.26: RIAA error using a 5534A in a +30 dB preamp, with and without HF correction pole. Scale ±0.5 dB

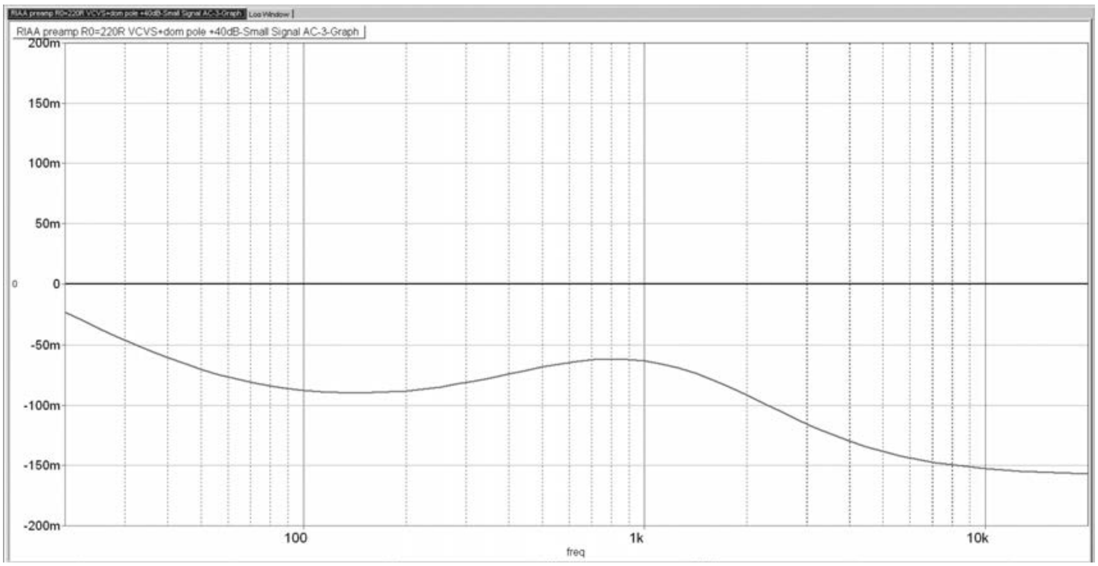

Figure 8.27: RIAA error with conceptual opamp. Closed-loop gain +40 dB at 1 kHz. Scale ±0.2 dB

Rinse and repeat with a +35 dB (1 kHz) preamp, and we get the Figure 8.28; the maximum errors are −0.03 dB at 40 Hz and −0.095 dB at 20 kHz. This is a useful improvement in accuracy at both LF and HF, and gives some idea of what basic opamp performance is required for a precise RIAA characteristic.

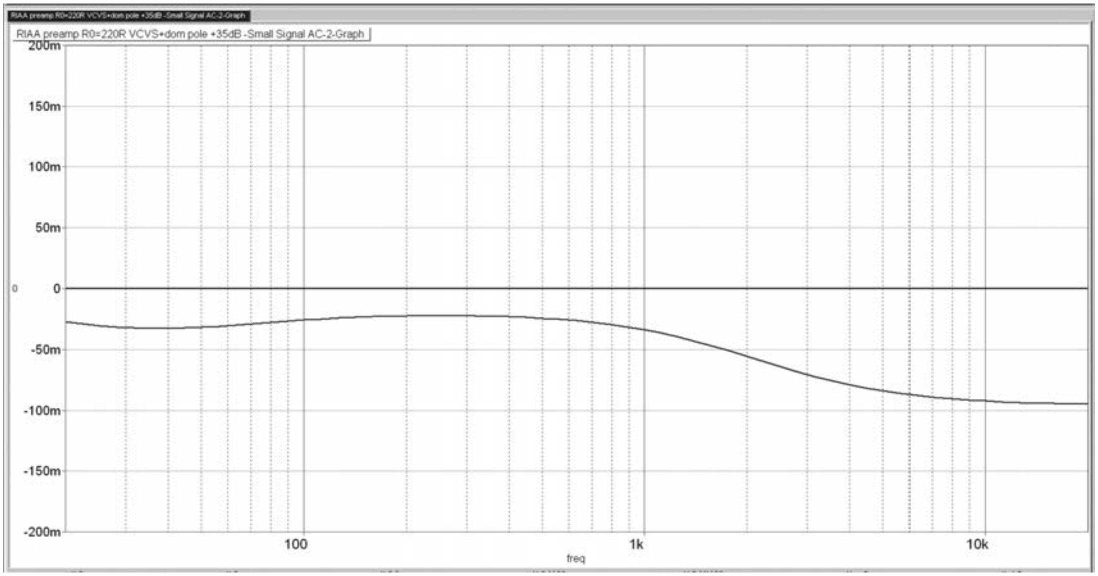

Figure 8.28: RIAA error with conceptual opamp. Closed-loop gain +35 dB at 1 kHz. Scale ±0.2 dB

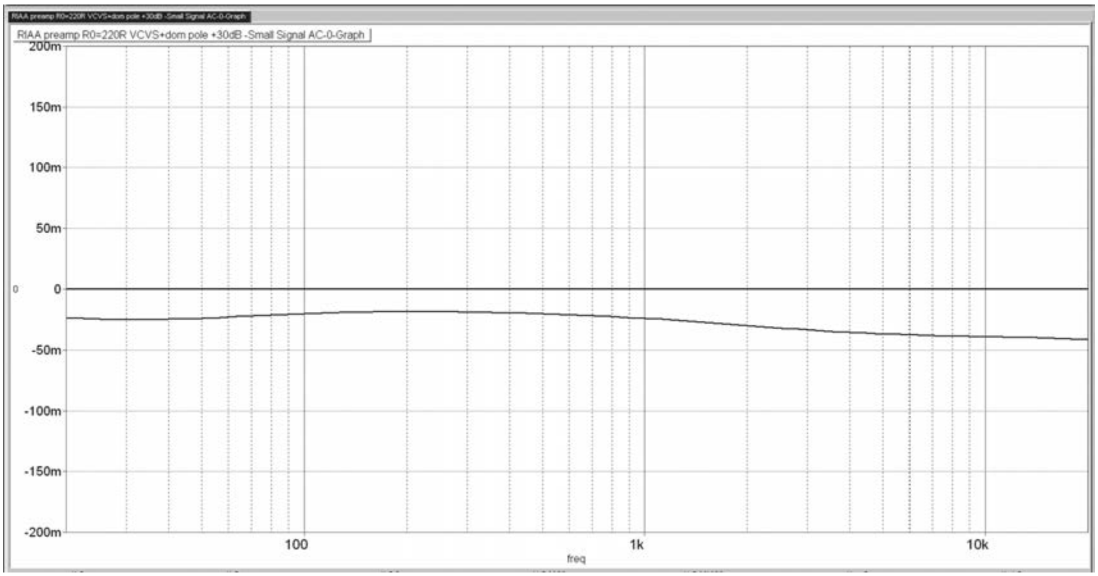

Figure 8.29: RIAA error with conceptual opamp. Closed-loop gain +30 dB at 1 kHz. Scale ±0.2 dB

Do it again with a +30 dB (1 kHz) preamp, and we get Figure 8.29; the maximum errors are now −0.025 dB at 30 Hz and −0.04 dB at 20 kHz. This is another handy improvement in accuracy at HF, but, as with the 5534 simulations, we note that there seems to be a small but persistent error at the LF end.

The error curves shown above have very gentle slopes. This is because the open-loop gain is falling at 6 dB/octave, but the demanded closed-loop gain is also falling at roughly this rate (not forgetting the plateau between 500 Hz and 2 kHz) so the error would be expected to be very roughly constant across the audio band.

To summarise, the open-loop gain of a 5534A is not adequate for a closed-loop gain of +40 dB at 1 kHz if you are aiming for an accurate RIAA response, and the +35 dB (1 kHz) situation is marginal. For +30 dB (1 kHz) the errors due to limited open-loop gain are negligible compared with the expected tolerances of the passive RIAA components. We have already seen that, if a wide range of cartridges and recording levels are to be accommodated, the minimum gain should be no more than +30 dB (1 kHz), so this works out quite nicely.

Switched-gain RIAA amplifiers

As noted above, it is not necessary to have a wide range of variable or stepped gain if we are only dealing with MM inputs, due to the limited spread of MM cartridge sensitivities – only about 7 dB. However, the sensitivity range of MC cartridges is very much greater (about 36 dB), so when the MM stage is being used as an RIAA equaliser, it would be very convenient if its gain could be easily changed. This would eliminate the need for a switched gain stage after the MM stage; such stages are dealt with later in this chapter. According to Peter Baxandall, at least two gain options are desirable [28].

However, as we have seen, the design of one-stage RIAA networks is not easy, and you might suspect that altering R0 away from the design point to change the gain is going to lead to some response errors. How right you are. Changing R0 introducing an LF RIAA error, plus an HF error as the HF correction pole is now wrong. Here are some examples, where the RIAA components are calculated for a gain of +30 dB, with R0 = 200 Ω, and then the gain increased by reducing R0:

- For +30 dB gain switched to +35 dB gain, (R0 = 112.47 Ω) the RIAA LF error is +0.07 dB 20 Hz–1 kHz, but the HF error is much bigger at −0.26 dB at 20 kHz.
- For +30 dB gain switched to +40 dB gain, (R0 = 63.245 Ω) the RIAA LF error is +0.10 dB 20 Hz–1 kHz, the HF error is at −0.335 dB at 20 kHz.

This includes the effect of finite open-loop gain when using a 5534 as the opamp.

Thus for real accuracy we need to switch not only R0 but also R1 in the RIAA feedback path, and R3 in the HF correction pole. If the RIAA error tolerance is ±0.1 dB, switching R1 could be omitted but two resistors still need to be switched. This assumes that the IEC Amendment is performed by a CR network after the MM stage, as described above; this will be unaffected by changes in R0. Otherwise, if the IEC Amendment is implemented by a small value of C0, you would need to switch that component as well, to avoid gross RIAA errors below 100 Hz. All in all, switched-gain RIAA amplifiers are not an attractive proposition. If you are catering for the full MC cartridge range then you will need five 5-dB steps, and you are switching an absolute minimum of two precision components if you are making any attempt at all at accuracy. That means lots of precise components and a 5-way 4-pole switch for stereo, which is an expensive and clumsy solution.

Shunt-feedback RIAA equalisation

The shunt-feedback equivalent of the basic RIAA stage is shown in Figure 8.30. It has occasionally been advocated because it avoids the unity-gain problem, but it has the crippling disadvantage that with a real cartridge load, with its substantial inductance, it is about 14 dB noisier than the series RIAA configuration [29]. A great deal of grievous twaddle has been talked about RIAA equalisation and transient response, in perverse attempts to render the shunt RIAA configuration acceptable despite its serious noise disadvantage. Since the input resistor R0 has to be 47 kΩ to load the cartridge correctly, the RIAA network has to operate at a correspondingly high impedance and will be noisy.

A series-feedback disc stage cannot make its gain fall below one, as described above, while the shunt-feedback version can; however an HF correction pole solves that problem completely. Shunt feedback eliminates any possibility of common-mode distortion, but then at the signal levels we are dealing with that is not a problem, at least with bipolar input opamps. A further disadvantage is that a shunt-feedback RIAA stage gives a phase inversion that can be highly inconvenient if you are concerned to preserve absolute phase.

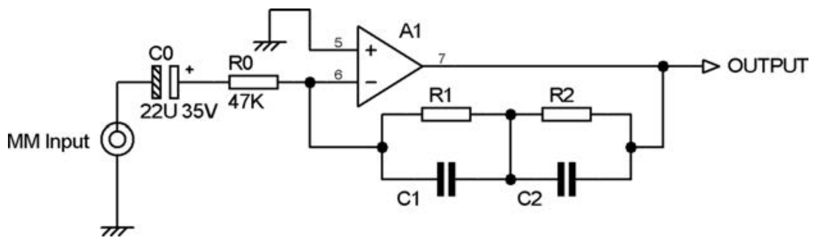

Figure 8.30: Shunt-feedback RIAA configuration. This is 14 dB noisier than the series-feedback version

Simulating inverse RIAA equalisation

SPICE simulation is well suited to the task of checking that the RIAA component values chosen are accurate. The best way to do this is to build an inverse-RIAA model to feed the RIAA preamplifier being simulated. This is much, much simpler than designing the preamplifier RIAA network because the time-constants can be completely decoupled from each other by using unity-gain buffers with zero-impedance outputs. The required response can be implemented in many ways, but my version is shown in Figure 8.31. The component values have nothing to do with practical circuitry and are chosen simply for ease of calculation.

The first network C1, R1 implements the 7960 μs time-constant of the notorious IEC Amendment. Since this first network is the inverse of a bass-roll-off, its output must continue to rise indefinitely at 6 dB/octave as frequency falls, and it is therefore implemented with a current-source, so that as the impedance of C1 rises, the output voltage at node 20 rises indefinitely. The apparently odd value of 1.011 A for the current source is in fact cunningly chosen to give a final output of 0 dBV at 1 kHz, which simplifies SPICE output plotting. The 10 GigaOhm resistor Rdummy is required as SPICE otherwise considers node 20 to be at an undefined DC level, and objects strongly. The voltage at node 20 controls the output of the VCVS (voltage-controlled voltage source) E1 which has its gain set to unity. It has zero output impedance and so acts as a mathematically perfect buffer. E is the conventional designator for a VCVS in SPICE.

E1 then drives the network R2, C3, R3, which implements the 3180 μs and 318 μs time-constants. E2 acts as another perfect buffer for the voltage at node 23, and drives R4, C3, R5, which implement the 75 μs time-constant. The very low value for R5 allows the output to go rising at 6 dB/octave to well beyond 20 kHz; the response does not level out until the T6 zero at 2.12 MHz is reached. If the IEC Amendment is not required, increase C1 to 10,000 μF so it has no effect in the audio band.

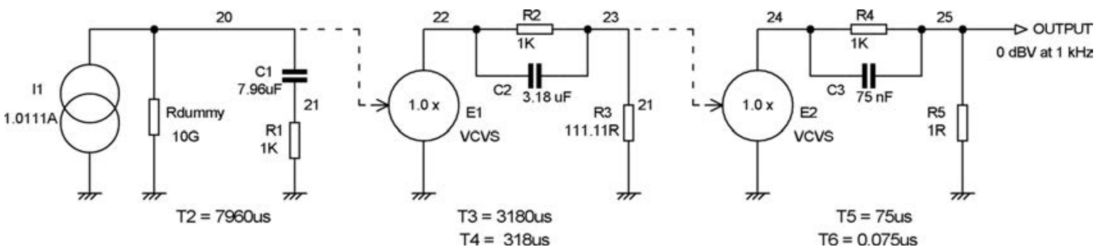

Figure 8.31: Inverse RIAA network for SPICE simulation

Physical inverse RIAA equalisation

Building a sufficiently accurate inverse RIAA network for precision measurements is not to be entered upon lightly or unadvisedly. The component values will need to have an accuracy a good deal better than 1%, and this makes sourcing components difficult and expensive. A much better alternative is to use a test system such as those by Audio Precision that allow an equalisation file to modify the generator output level during a frequency sweep.

Passive and semi-passive RIAA equalisation

For many years, series-feedback RIAA preamplifiers as described above were virtually universal, it being accepted by all that they gave the best noise, overload performance, and economy, especially of active components. However, human nature being what it is, some people will always want to do things the hard way, and this is exemplified by the fashion for passive (actually, semi-passive is more accurate) RIAA equalisation. The basic notion is to split the RIAA equalisation into separate stages, and I have a dark and abiding suspicion that this approach may be popular simply because it makes the design of accurate RIAA equalisation much easier, as all you have to do is calculate simple time-constants instead of grappling with foot-long equations. There is a price, and a heavy one: the overload and/or noise performance is inevitably compromised.

Clearly a completely passive RIAA stage is a daft idea because a lot of gain is required somewhere to get the 5 mV cartridge signal up to a usable amplitude. The nearest you can get to completely passive is the scheme shown in Figure 8.32a, where the amplification and the equalisation are wholly separate, with no frequency-dependent feedback used at all. R2, R3 and C1 implement T3 and T4, while C2 implements T5. There is no inconvenient T6 because the response carries on falling indefinitely with frequency. This network clearly gives its maximum gain at 20 Hz, and at 1 kHz it attenuates by about 20 dB. Therefore, if we want the modest +30 dB gain at 1 kHz used in the previous example, the A1 stage must have a gain of no less than 50 dB. A 5 mVrms 1 kHz input would therefore result in 1.58 V at the output of A1. This is only 16 dB below clipping, assuming we are using the usual sort of opamps, and an overload margin of 16 dB is much too small to be usable. It is obviously impossible to drive anything like a volume control or tone control stage from the passive network, so the buffer stage A2 is shown to emphasise that extra electronics is required with this approach.

The only way to improve the overload margin is to split the gain so that the A1 stage has perhaps 30 dB, while A2 after the passive RIAA network makes up the loss with 20 dB more gain. Sadly, this second stage of amplification must introduce extra noise, and there is always the point that you now have to put the signal through two amplifiers instead of one, so there is the potential for increased distortion.

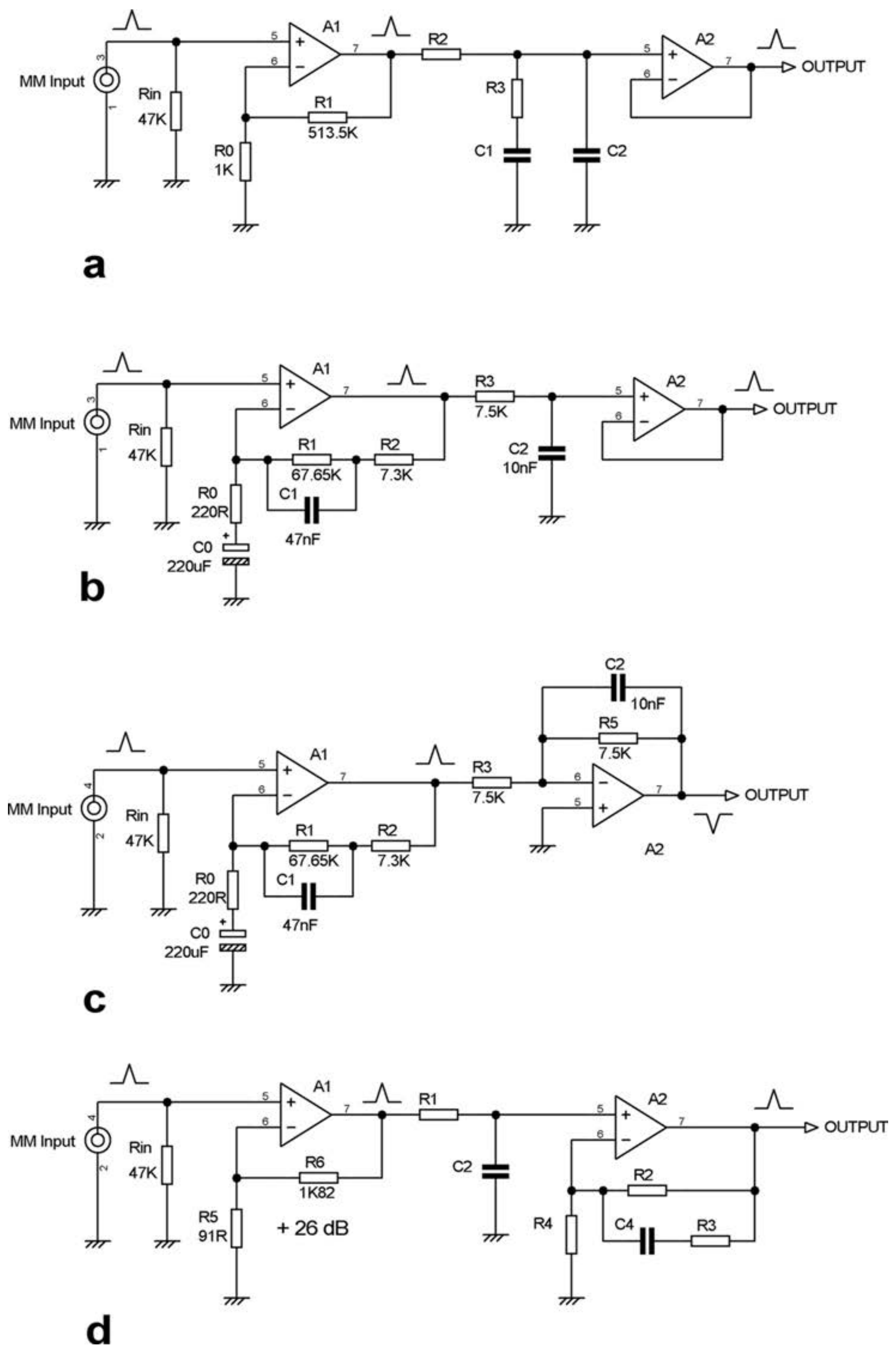

Figure 8.32: Passive and semi-passive RIAA configurations

The most popular architecture that separates the high and low RIAA sections is seen in Figure 8.32b. Here there is an active LF RIAA stage using feedback to implement T3 and T4 with R1, C1, R2, followed by R3, C2 which give a passive HF cut for T5. This is what I call an active-passive configuration. The values shown give an RIAA curve correct to within 0.04 dB from 20 Hz to 20 kHz. Note that because of the lack of time-constant interaction, we can choose standard values for both capacitors, but we are still left with awkward resistor values.

As before, amplification followed by attenuation means a headroom bottleneck, and this passive HF roll-off is no exception. Signals direct from disc have their highest amplitudes at high frequencies, so both these configurations give poor HF headroom, overload occurring at A1 output before passive HF cut can reduce the level. Figure 8.33 shows how the level at A1 output (Trace B) is higher at HF than the output signal (Trace A). The difference is Trace C, the headroom loss; from 1 dB at 1 kHz this rises to 14 dB at 10 kHz and continues to increase in the ultrasonic region. The passive circuit was driven from an inverse RIAA network, so a totally accurate disc stage would give a straight line just below the +30 dB mark.

A related problem in this semi-passive configuration is that the opamp A1 must handle a signal with much more HF content than the opamp in the single-stage series-feedback configuration, worsening any difficulties with slew-limiting and HF distortion. It uses two amplifier stages rather than one, and more precision components, because of the extra resistor. Another difficulty is that A1 is more likely to run out of open-loop gain or slew rate at HF, as the response

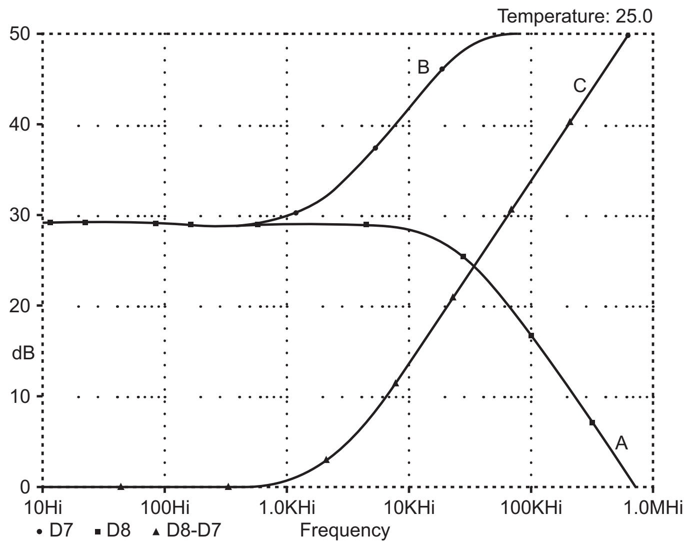

Figure 8.33: Headroom loss with passive RIAA equalisation. The signal level at A1 (Trace B) is greater than at A2 (Trace A) so clipping occurs there first. Trace C shows the headroom loss, which reaches 18 dB at 20 kHz

plateaus above 1 kHz rather than being steadily reduced by increasing negative feedback. Once again a buffer stage A2 is required to isolate the final time-constant from loading.

A third method of equalisation is shown in Figure 8.32c, where the T5 roll-off is done by feedback via R4, C2 rather than by passive attenuation. This is not really passive in any way, as the equalisation is done in two active stages, but it does share the feature of splitting up the time-constants for easier design. As with the previous circuit, A1 is running under unfavourable conditions as it has to handle a larger HF content than in the series-feedback version, and there is now an inconvenient phase reversal. The values shown give the same gain and RIAA accuracy as the previous circuit, though in this case the value of R3 can be scaled to change the gain.

There are many other alternative arrangements that can be used for passive or semi-passive equalisation. There could be a flat input stage followed by a passive HF cut and then another stage to give the LF boost, as in Figure 8.32d, which has even more headroom problems and uses yet more bits. I call this a passive-active configuration. In contrast the 'all-in-one-go' series feedback configuration avoids unnecessary headroom restrictions and has the minimum number of stages.

Passive RIAA is not an attractive option for general use, but comes into its own in the archival transcription of recordings, where there are dozens of different pre-RIAA equalisation schemes, and it must be possible to adjust the turnover frequencies f3, f4 and f5 independently. This is done most straight-forwardly by a fifth passive-passive equalisation configuration which is described in detail in Chapter 9.

As always, what you think is a recent trend has its roots in the past. An active-passive MM input stage was published in *Wireless World* in 1961 [30]. This had a two-transistor series-feedback amplifier which dealt with the LF equalisation, followed by a passive RC HF roll-off.

Peter Baxandall published a circuit in 1981 [28] with the configuration of Figure 8.32b that gave easy switched gain control and allowed the use of preferred values, with only two of them in the E24 series. Like all Peter's ideas it is well worth studying and is shown in Figure 8.34. The gains are +20, +30 and +40 dB, accurate to within ±0.3 dB; the switchable gain largely avoids the headroom problems of passive RIAA equalisation. The RIAA accuracy is within ±0.03 dB between 1 kHz and 20 kHz in each case, falling off to about −0.1 dB at 100 Hz. This is due to the way that R0 and C0 implement the IEC Amendment, giving f2 = 21.22 Hz rather than the correct 20.02 Hz; that is as close as you can get with a single 750 Ω E24 resistor for R0. It results in a response 0.34 dB too low at 20 Hz. The correct value for R0 is 795 Ω, so f2 could be made much more accurate by using the parallel pair 1 kΩ and 3.9 kΩ, which is only 0.1% too high. However, there is the tolerance of C0 to be considered, and when Peter was writing (1981) that would have been larger than we would expect today, so 750 Ω was close enough.

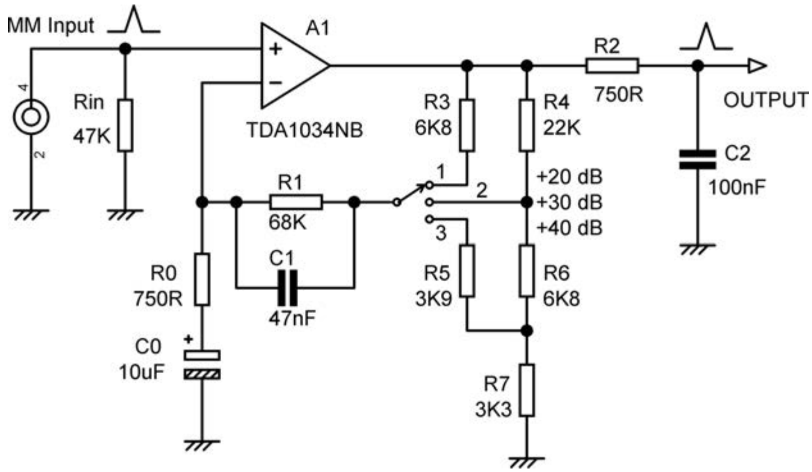

Figure 8.34: Active-passive RIAA stage with switched gain by Peter Baxandall

One problem with this circuit suggests itself. When the gain switch is between contacts, A1 has no feedback and will hit the rails. Very likely Peter was thinking of a make-before break switch. Another possible way of solving this is given in Figure 8.36 below, where feedback is maintained when the switch moves. The TDA1034B was an early version of the 5534, and capable of driving the relatively low impedance of the R2−C2 combination.

MM cartridge loading and frequency response

The standard loading for a moving-magnet cartridge is 47 kΩ in parallel with a certain amount of capacitance, the latter usually being specified by the maker. The resulting resonance with the cartridge inductance is deliberately used by manufacturers to extend the frequency response, so it is wise to think hard before trying to modify it. Load capacitance is normally in the range 50 to 200 pF. The capacitance is often the subject of experimentation by enthusiasts, and so switchable capacitors are often provided at the input of high-end preamplifiers which allow several values to be set up by combinations of switch positions. The exact effect of altering the capacitance depends on the inductance and resistance of the cartridge, but a typical result is shown in Figure 8.35 where increasing the load capacitance lowers the resonance peak frequency and makes it more prominent and less damped. It is important to remember that it is the total capacitance, including that of the connecting leads, which counts.

Because of the high inductance of an MM cartridge, adjusting the load resistance can also have significant effects on the frequency response, and some preamplifiers allow this too to be altered. The only objective way to assess the effects of these modifications is to measure the output when a special (and expensive) test disc is played.

When loading capacitance is used it should be as near to the input socket as possible so it can contribute to filtering out RF before it radiates inside the enclosure. However, its effectiveness for EMC purposes is likely to be much compromised if the capacitors are switched. Normal

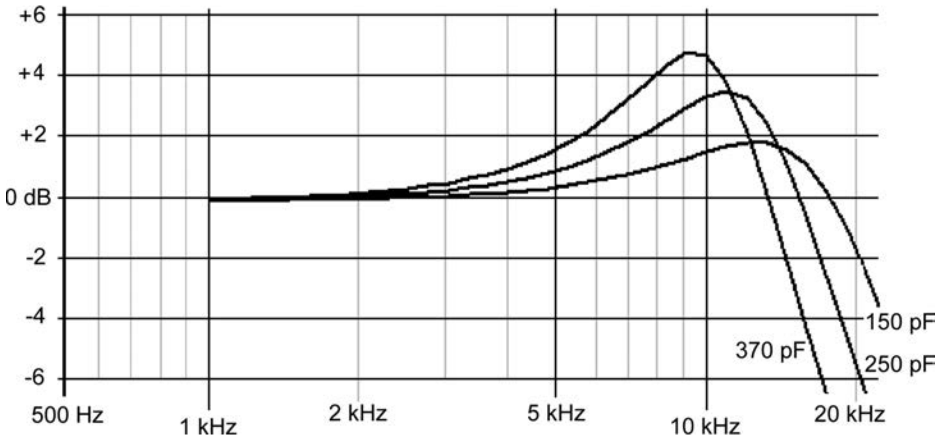

Figure 8.35: The typical effect of changing the loading capacitance on an MM cartridge

practice is that the smallest capacitor is permanently in circuit so it can be mounted right on the rear of the input socket. A continuously variable loading capacitance could be made with an old-style tuning capacitor (two-section for stereo); looking back they were marvels of mass-produced precision engineering. The maximum value in an old medium-wave radio is often a rather convenient 500 pF. This would look well cool but naturally takes up a lot of space, and the variable-bootstrapping of a fixed capacitor (see the variable-frequency tone control in Chapter 15) would be much more compact.

The exact nature of this resonance does not have a consensus in the hifi community. There is also the possibility of what is usually called the 'cantilever resonance' which is a mechanical resonance between the effective tip mass of the stylus and the compliance of the vinyl it is tracking, the latter making up the spring part of the classic mass-and-spring system. The effective tip mass of the stylus is contributed to by the mass of the diamond tip, the cantilever, and the generator element on the other end, which may be a piece of iron, a magnet, or coils; it usually ranges from 0.2 to 0.7 milligrams. There is also the question of the contribution of the cantilever compliance, and the possibility of a torsional resonance of the cantilever [31]. You are probably thinking by now that this is a mass of electromechanical compromises that should be left alone, and you are probably right.

Not everyone agrees. A scheme for cancelling the effects of the cantilever resonance with a sophisticated active filter was put forward by Steven van Raalte in *Linear Audio* Volume 3 [32]. A slightly earlier attempt, in 1953 [33], simply put a series LC circuit across the cartridge output.

MM cartridge–preamplifier interaction

One often hears that there can be problems due to interaction between the impedance of the cartridge and the negative-feedback network. Most commentators are extremely vague as to what this actually means, but according to Tomlinson Holman [34], the factual basis is that it used to be all too easy to design an RIAA stage, if you are using only two or three discrete transistors, in which the NFB factor is falling significantly with frequency in the upper reaches of the audio band, perhaps due to excessive dominant-pole compensation to achieve HF stability (on the other hand, the amount of feedback is increasing with frequency due to RIAA equalisation). Assuming a series-feedback configuration is being used, this means that the input impedance will fall with frequency, which is equivalent to having a capacitive input impedance. This interacts with the cartridge inductance and allegedly can cause a resonant peak in the frequency response, in the same way that cable capacitance or a deliberately added load capacitance can do.

For this reason a flat-response buffer stage between the cartridge and the first stage performing RIAA equalisation was sometimes advocated. One design including this feature was the Cambridge Audio P50, which used a Darlington emitter-follower as a buffer; with this approach there is an obvious danger of compromising the noise performance because there is no gain to lift the signal level above the noise floor of the next stage.

MM cartridge DC and AC coupling

Some uninformed commentators have said that there should be no DC blocking capacitor between the cartridge and the preamplifier. This is insane. Keep DC out of your cartridge. The signal currents are tiny (for MM cartridges 5 mV in 47 kΩ = 106 nA, while for MC ones 245 μV in 100 Ω = 2.45 μA – a good deal higher) and even a small DC bias current could interfere with linearity. I am not aware of any published work on how cartridge distortion is affected by DC bias currents, but I think it pretty clear they will not improve things and may make them very much worse. Large currents might partially demagnetise the magnet, be it moving or otherwise, ruining the cartridge. Even larger currents due to circuit faults might set fire to the coils, ruining the cartridge even more effectively. You may call a lack of blocking capacitors high-end, but I call it highly irresponsible.

If I had a £15,000 cartridge (and they do exist, by Koetsu and Clearaudio) I would probably put two blocking capacitors in series. Or three.

Switched-gain flat stages

In the earlier parts of this chapter we saw that the appropriate gain (at 1 kHz) for an MM input with pretensions to quality is between +30 and +40 dB, giving maximum inputs at 1 kHz of 316 and 100 mVrms respectively. Lower gains give an inconveniently low output signal, and a greater headroom loss at HF due to the need for a lower HF correction pole frequency. Higher gains give too low a maximum input.

The nominal output for 5 mVrms input (1 kHz) from a +30 dB stage is 158 mVrms, and from a +40 dB stage is 500 mVrms. Bearing in mind that the line signals between pieces of equipment are, in these digital days usually in the range 1−2 Vrms, it is obvious that both

158 mVrms and 500 mVrms are too low. If we put a fixed gain stage after the MM input stage, it will overload first and the maximum inputs just quoted are no longer valid. It is therefore desirable to make such a stage switchable in gain, to cope with differing conditions of cartridge sensitivity and recorded level. One of the gain options must be unity (0 dB) if the maximum MM inputs are to be preserved; having less than unity gain is pointless as the MM stage will clip first. It would of course be possible to have continuously variable gain controlled by a pot, but this brings in difficult issues of stereo level matching (these are described in Chapter 13 on volume controls). It is not, in my opinion, necessary to have finer control of the post-MM-input gain than 5 dB steps.

If we are dealing with just MM inputs, then not many gain options are required. If we assume a +30 dB (1 kHz) MM stage with its nominal 158 mVrms output, then we need 6.3 times or +16 dB of gain to raise that level to 1 Vrms. This suggests that gain options of 0 dB, +5 dB, +10 dB, and +15 dB are all that are needed, with the lower gains allowing for more sensitive cartridges and elevated recording levels.

However, it will be seen in Chapter 12 that MC cartridges have a much wider spread of sensitivities than the MM variety, and if the MM input stage followed by the flat switched-gain stage are going to be used to perform the RIAA equalisation after a flat +30 dB MC head amp, a further +20 dB gain option in the switched-gain stage is required to ensure that even the most insensitive MC cartridges can produce a full 1 Vrms nominal output.

The stage in Figure 8.36 is derived from my Elektor 2012 preamp [26] and gives those gain options. The AC negative feedback is tapped from the divider R51−R60, which is made up

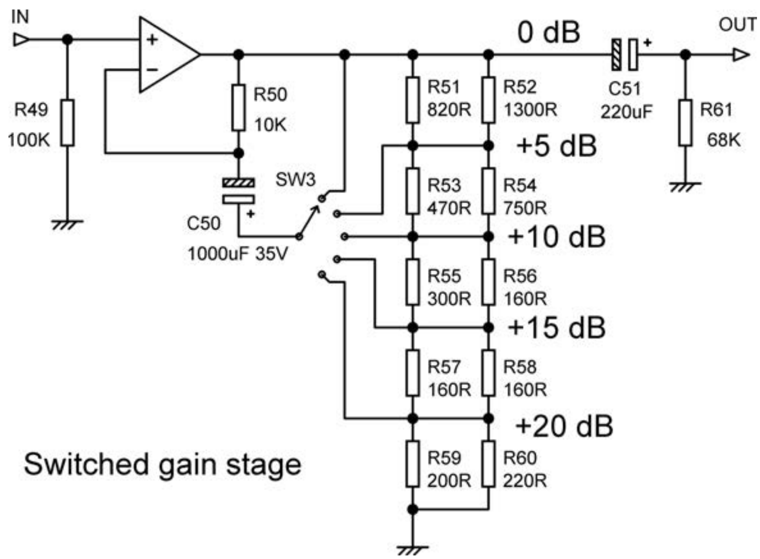

Figure 8.36: A flat gain stage with accurate switched gains of 0, +5, +10, +15, and +20 dB. Resistor pairs are used to get the exact gains wanted, and to reduce the effect of tolerances

of pairs of resistors to achieve the exact gain required and to reduce the effect of the resistor tolerances. The DC feedback for the opamp is always through R50 to prevent the opamp hitting the rails when switching the gain; the blocking capacitor C50 is more than large enough to prevent any frequency response irregularities in the audio band. Assuming the source impedance is reasonably low, an LM4562 will give better noise and distortion results than a 5532 section.

The correct setting for the gain switch can be worked out by considering cartridge sensitivity specs and recording levels, but the latter are usually unknown, so some form of level indicator is very useful when setting up. A bar-graph meter seems a bit over the top for a facility that will not be used very often, and a single LED indication makes more sense. For this reason the Log Law Level LED was developed, giving about as much level information as can be had from one LED. It is fully described in Chapter 23 on metering. It is desirable that any level-indication can be switched off as not everyone thinks that flashing lights add to the musical experience.

Subsonic filters

In the earlier parts of this chapter we have seen that the worst subsonic disturbances occur in the 2−4 Hz region, due to disk warps, and are about 8 dB less at 10 Hz. We have also seen that the IEC Amendment gives only 14 dB of attenuation at 4 Hz, and in any case is often omitted by the manufacturer or switched out by the user. It is therefore important to provide authoritative subsonic filtering. What needs to be settled is what order filter to use, because some people at least will be concerned about the audibility of LF phase shifts, and how far into the audio band the filter should intrude. There is nothing approaching a consensus on either point, so it can be a wise move to configure the subsonic filter so it can be switched out.

The third-order filter already described in this chapter also did the job of implementing the IEC Amendment, so it did not have one of the classic filter characteristics. All the filters described here do just the filtering job and it is assumed that the IEC Amendment is implemented elsewhere, if at all.

High-pass filters used for RIAA subsonic are typically of the second-order or third-order Butterworth (maximally flat) configuration, rolling off at rates of 12 dB/octave and 18 dB/octave respectively, as shown in Figure 8.37. Fourth order 24 dB/octave filters are much less common, presumably due to worries about the possible audibility of rapid phase changes at the very bottom of the audio spectrum. The Butterworth response is but one of many possible filter alignments; the Bessel response gives a slower roll-off, but aims for linear phase – i.e. a constant delay versus frequency – and so reproduces the

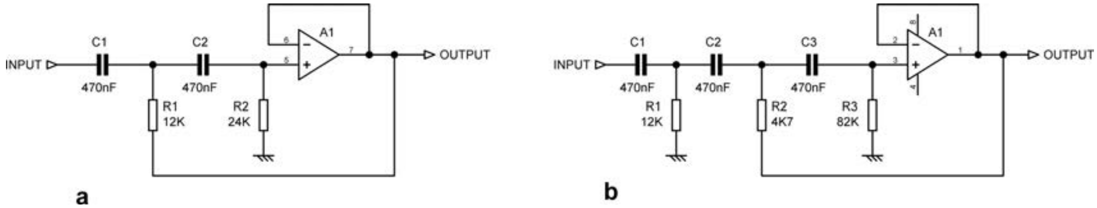

Figure 8.37: Subsonic filters: a) second-order, and b) third-order Butterworth high-pass filters, both 3 dB down at 20 Hz

shape of transients better. Other filter alignments such as Chebyshev give faster initial roll-offs than the Butterworth, but they do so at the expense of ripples in the passband or stopband gain which is not helpful if you are aiming for a ruler-flat response after RIAA equalisation.

A very handy filter configuration is the well-known Sallen and Key type; it has drawbacks when used as a low-pass handling high frequencies (the response comes back up at RF due to the non-zero opamp output impedance) but works very well for our purposes here. A second-order Sallen and Key is simple to design; the two series capacitors C1 and C2 are made equal and R2 is made twice the value of R1. Such a filter with a −3 dB point at 20 Hz is shown in Figure 8.37a. Other roll-off frequencies can be obtained simply by scaling the component values while keeping C1 equal to C2 and R2 twice R1. The response is 24.0 dB down at 5 Hz, by which time the 12 dB/octave slope is well-established, and we are well protected against disk warps. It is however only 12.3 dB down at 10 Hz, which gives little protection against arm-resonance problems. Above the −3 dB roll-off point the response is still −0.78 dB down at 30 Hz, which is intruding a bit into the sort of frequencies we want to keep. We have to conclude that a second-order filter really does not bifurcate the condiment, and the faster roll-off of a third-order filter is preferable.

Third-order filters are a little more complex. Some versions are made up of a second-order filter cascaded with a first-order roll-off, using two opamp sections. It can however be done with just one, as in Figure 8.37b, which is a third-order Butterworth filter also with a −3 dB point at 20 Hz. The resistor value ratios are now a less friendly 2.53:1.00:17.55, and the circuit shown uses the nearest E24 values to this – which by happy chance come out as E12 values. The frequency response is shown in Figure 8.38, where it can be seen to be 18.6 dB down at 10 Hz, which should keep out any arm-resonance frequencies. It is 36.0 dB down at 5 Hz so disk warp spurii won't have a chance. The 30 Hz response is now only down by an insignificant −0.37 dB, which demonstrates that how a third-order filter is much better than

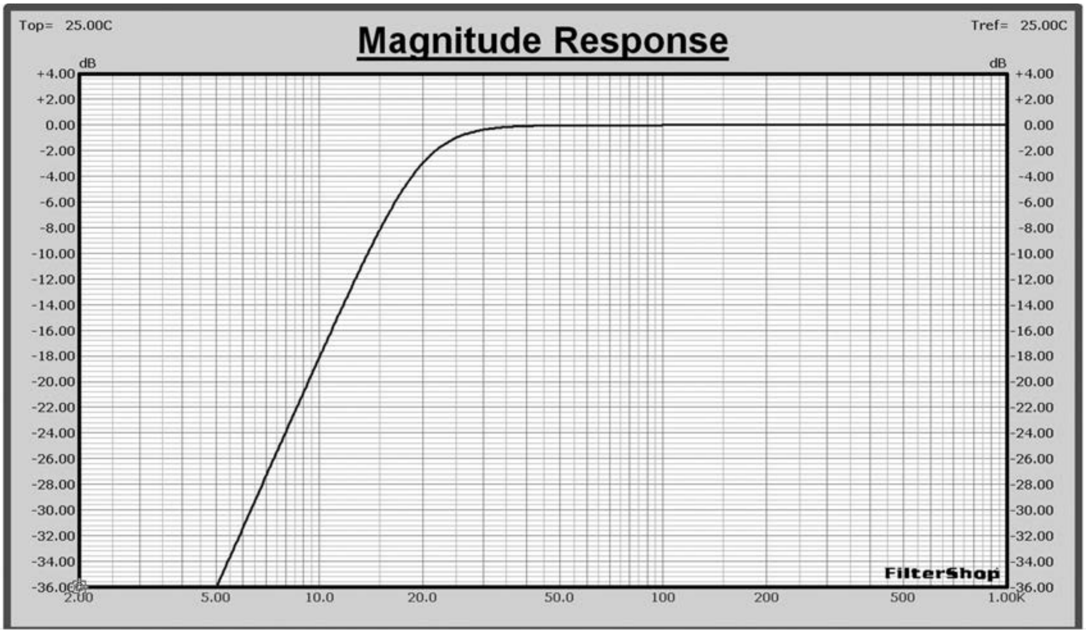

Figure 8.38: Frequency response of a third-order Butterworth subsonic filter, 3 dB down at 20 Hz.

a second-order filter for this application. As before, other roll-off frequencies can be had by scaling the component values while keeping the resistor ratios the same.

When dealing with frequency-dependant networks like filters you need to keep an eye on the input impedance, because it can drop to unexpectedly low values, putting excessive loading on the stage upstream and degrading its linearity. In a high-pass Sallen and Key filter, the input impedance is high at low frequencies but falls with increasing frequency. In the third-order version, it tends to the value of R1 in parallel with R3, which here is 10.6 kΩ. This should not worry the previous stage.

Because of the large capacitances, the noise generated by the resistors in a high-pass filter of this sort is usually well below the opamp noise. The capacitances do not, of course, generate any noise themselves. With the values used here, SPICE simulation shows that the resistors produce −125.0 dBu of noise at the output (22 kHz bandwidth, 25 °C). The use of the LM4562 will reduce voltage-follower CM distortion compared with the 5534/5532, but may be noisier in some cases due to the higher current noise of the LM4562.

Capacitor distortion in electrolytics is (or should be) by now a well-known phenomenon. It is perhaps less well known that non-electrolytics can also generate distortion in filters like these. This has nothing to do with Subjectivist musicality, but is all too real and measurable.

Details of the problem are given in Chapter 2, where it is concluded that only NP0 ceramic, polystyrene, and polypropylene capacitors can be regarded as free of this effect. The capacitor sizes needed for subsonic filters are large, if impedances and hence noise are to be kept low, which means it has to be polypropylene. Anything larger than 470 nF gets to be big and expensive, so that is the value used here. 220 nF polypropylene is substantially smaller and about half the price. There is more information on this, and on high-pass filters in general, in Chapter 6 on filters.

Since most of the low frequency disturbances from a disc are due to up-and-down motion, they are reproduced as two out-of-phase signals by a stereo pickup cartridge. It has often been suggested that severe rumble overlapping the audio band can be best dealt with by reducing the stereo signal to mono at low frequencies, cancelling the disturbances but leaving the bass, which is usually panned towards the middle, relatively unaffected. This is usually done by cross-feeding the outputs of two low-pass filters between the channels. Several circuits have been published to perform this; one example is reference [35].

Ultrasonic filters

Scratches and groove debris create clicks that have a large high-frequency content, some of it ultrasonic and liable to cause slew rate and intermodulation problems further down the audio chain. The transients from scratches can easily exceed the normal signal level. It is often considered desirable to filter this out as soon as possible (though of course some people are only satisfied with radio-transmitter frequency responses).

If an MM input stage is provided with an HF correction pole, in the form of an RC first-order roll-off after the opamp, this in itself provides some protection against ultrasonics as its attenuation continues to increase with frequency and it is inherently linear. The opamp ahead of it naturally does not benefit from this; while it might be desirable to put some ultrasonic filtering in front of the first active stage, in practice it is going to be very hard to do this without degrading the noise performance.

A true ultrasonic filter could be a passive LC design, but inductors are not much loved in audio. A more likely choice is a second or third-order active filter, probably opamp-based, but if Sallen and Key filters are used then a discrete emitter-follower is an option, and this should be free from the bandwidth and slew rate limitations of opamps. If an ultrasonic filter is incorporated it is usually second-order, very likely due to misplaced fears of perceptible phase effects at the top of the audio band. If a Sallen and Key filter with an opamp is used, be aware that the response does not keep going down forever but comes back up due to the non-zero output impedance of the opamp at high frequencies; the multiple-feedback (MFB) filter configuration is free from this problem. The design of suitable low-pass filters to remove ultrasonics is fully explained in Chapter 6 on filters.

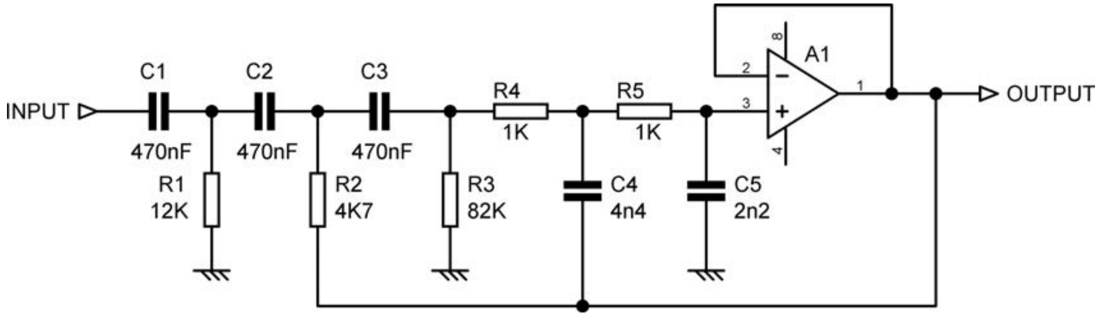

Figure 8.39: A third-order Butterworth subsonic filter combined with a second-order ultrasonic filter

The combination of a subsonic filter and an ultrasonic filter is sometimes called a bandwidth definition filter.

Combining subsonic and ultrasonic filters in one stage

An obstacle to the inclusion of an ultrasonic filter is the extra cost and power consumption of another filter stage. This difficulty can be resolved by combining it with a subsonic filter in the same stage. Combined filters also have the advantage that the signal now passes through one opamp rather than two, and can be extremely useful if you only have one opamp section left.

This cunning plan is workable only because the high-pass and low-pass turnover frequencies are widely different. Figure 8.39 shows the third-order Butterworth subsonic filter combined with a second-order 50 kHz Butterworth low-pass filter; the response of the combination is exactly the same as expected for each separately. The low-pass filter is cautiously designed to prevent significant loss in the audio band, and has a −3 dB point at 50 kHz, giving very close to 0.0 dB at 20 kHz. The response is −12.6 dB down at 100 kHz and −24.9 dB at 200 kHz. C1 is made up of two 2n2 capacitors in parallel.

Note that the mid band gain of the combined filter is −0.15 dB rather than exactly unity. The loss occurs because the series combination of C1, C2 and C3, together with C5, form a capacitive potential divider with this attenuation, and this is one reason why the turnover frequencies need to be widely separated for filter combining to work. If they were closer together then C1, C2, C3 would be smaller, C5 would be bigger, and the capacitive divider loss would be greater.

Scratch filters

In what might be called the First Age of Vinyl, a fully-equipped preamplifier would certainly have a switchable low-pass filter, usually called, with brutal frankness, the ‘scratch’ filter. It would have a roll-off slope of at least 12 dB/octave, faster than the 6 dB/octave maximum slope of the tone-control stage, and commencing roll-off at a much higher frequency. It was aimed at suppressing, or at any rate dulling and hopefully rendering acceptable, not just record surface noise and the inevitable ticks and clicks, but also HF distortion. This function is quite separate from that of an ultrasonic filter, and the turnover frequency is very much in the audio band, usually in the range 3–10 kHz.

The more highly-specified preamps would have two, three, or more alternative filter frequencies, and the really posh models had variable filter slope as well. The fact that the need was felt for some really quite sophisticated filtering to smooth the listening experience does rather emphasise the inherent vulnerability of a mechanical groove for delivering music. Now we seem to be in the Second Age of Vinyl, a reassessment of this once-abandoned bit of technology would seem to be timely.

Historically scratch filters were often passive LC configurations as that was much cheaper than putting in an extra valve to make an active filter. A modern scratch filter, if such a thing was desired, would almost certainly be an active filter. Some versatile scratch filters with variable slope are described in Chapter 9 on archival disc transcription; the design of low-pass active filters in general is explained in Chapter 5 on filters.

A practical MM amplifier: #1

The closely observed designs given here are intended to demonstrate the various techniques discussed in this chapter, and Chapter 11 on MM amplifier noise and distortion. The MM amplifier shown in Figure 8.40 is based on the MM section of the Signal Transfer MM/MC phono amplifier [36]. This practical design includes cartridge loading capacitor C1, input DC blocking capacitor C2 and DC drain R1 which stops mighty thumps being caused by charge left on C2 if the input is unplugged. R1 in parallel with R2 makes up the 47 kΩ resistive input load. I have used this circuit for many years and it has given complete satisfaction to many customers, though in the light of the latest knowledge it could be further optimised to economise on precision capacitors. It includes a typical subsonic filter which is designed with a slow initial roll-off that implements the IEC Amendment, so a separate network is not required. A 5534A is used at the input stage to get the best possible noise performance. A 5534A without external compensation has a minimum stable closed-loop gain of about three times; that is close to the gain at 20 kHz here, so a touch of extra compensation is required for stability. The capacitor used here is 4.7 pF, which experience shows is both definitely required and also gives completely reliable stability. This is tested by sweeping a large signal from 20 kHz downwards; single-frequency testing can miss this sort of problem.

The resistors have been made more accurate by combining two E24 values. In this case they are used in series and no attempt was made to try and get the values equal for

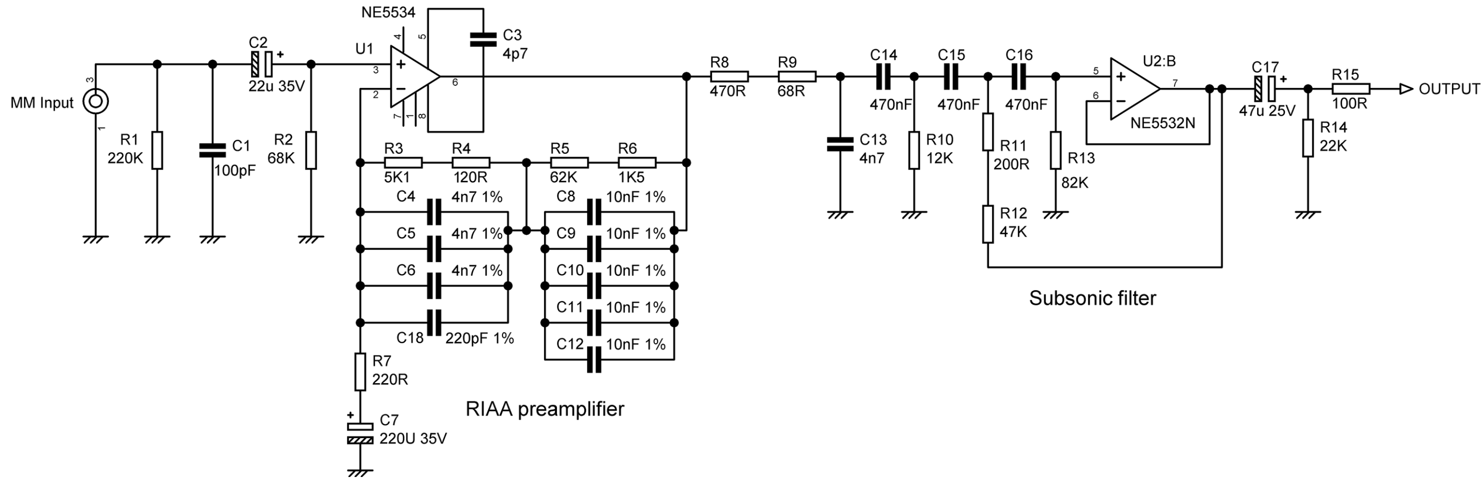

Figure 8.40: MM input with HF correction pole, and IEC amendment implemented by third-order subsonic filter. Based on Signal Transfer design. Gain +30 dB at 1 kHz

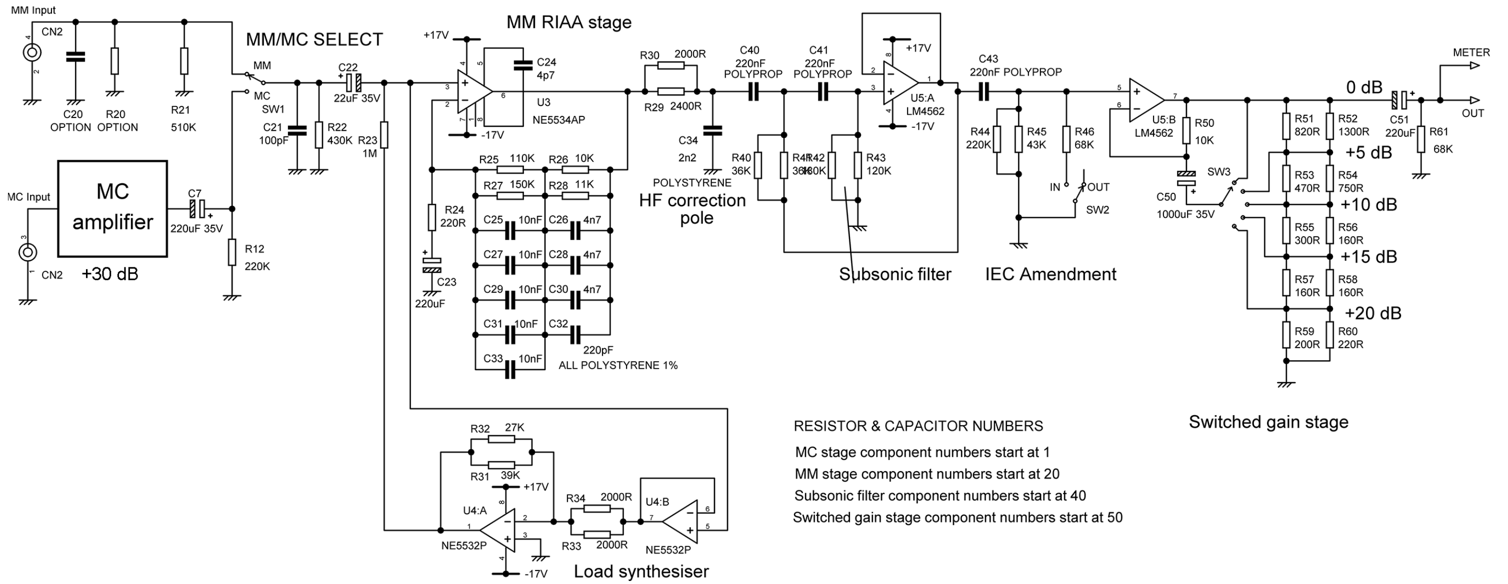

Figure 8.41: MM and MC input with HF correction pole, switchable IEC amendment third-order subsonic filter. Gain minimum +30 dB at 1 kHz

the maximum reduction of tolerance errors. That statistical work was done at a later date. The Configuration-A RIAA network capacitances are made up of multiple 1% polystyrene capacitors for improved accuracy. Thus for the five 10 nF capacitors that make up C1, the standard deviation (square root of variance) increases by the square root of five, while total capacitance has increased five times, and we have inexpensively built an otherwise costly 0.44% close-tolerance 50 nF capacitor. You will note that 5 × 10 nF capacitors are required, whereas a Configuration-C RIAA network can do the same job with 4 × 10 nF.

C2 is essentially composed of three 4n7 components and its tolerance is improved by $\sqrt{3}$, to 0.58%. Its final value is tweaked by the addition of C15. An HF correction pole R8, R9, C13 is fitted; here the resultant loss of HF headroom is only 0.5 dB at 20 kHz, which I think I can live with.

Immediately after the RIAA stage is the subsonic filter, a third-order Butterworth high-pass filter which also implements the IEC Amendment by using a value for R11 + R12 which is lower than that for maximal Butterworth flatness. The stage also buffers the HF correction pole R8, R9, C13 from later circuitry, and gives the capability to drive a 600 Ω load, if you can find one. A version of this design, using appropriate precision components, is manufactured by the Signal Transfer Company in bare PCB, kit, and fully built and tested formats [36].

A practical MM amplifier: #2

The MM/MC input system shown in Figure 8.41 is based on my recent Elektor preamp design [26], which was a no-holds-barred attempt at getting the best possible performance. The input is switchable between a phono socket for the connection of MM cartridges, and an MC head amplifier with a flat gain of +30 dB. A switched-gain stage allows every MC and MM cartridge on the market to be catered for.

The MM stage uses Configuration-A (the work on optimising RIAA networks not having been done at the time) and the gain is +30 dB at 1 kHz using 5 × 10 nF polystyrene capacitors to obtain the required value and to improve RIAA accuracy (because random errors in the capacitor values tend to cancel). Multiple RIAA resistors R22, R23 and R24, R25 are used to improve accuracy in the same way, and this time the paralleled values were optimised by being made as near equal as possible. The value of C12 is large at 220 µF as the IEC Amendment is not implemented in this stage, and tweaking the IEC Amendment to compensate for its less-than-infinite value (see earlier in this chapter) was not considered necessary. The +30 dB (1 kHz) gain of the MM stage requires that the RIAA characteristic is corrected at HF by R26, R27, and C22. Two near-equal resistors are again used to improve accuracy; C22 is polystyrene to avoid capacitor distortion.

A 5534A is used here for IC3 as it is quieter than half a 5532, and considerably quieter than an LM4562 with its higher current noise. The high inductance of an MM cartridge makes low current noise important.

A load-synthesis circuit IC4 is used to make an electronic version of the required 47 kΩ loading resistor from the 1M resistor R16. The Johnson noise of the resistor is however *not* emulated and so noise due to the rising impedance of the MM cartridge inductance is much reduced. R16 is made to appear as 47 kΩ by driving its bottom end in anti-phase to the signal at the top. IC4B prevents loading on the MM input while IC4A is the inverting stage. Paired resistors R19, R20 and R17, R18 are used to improve the gain accuracy of the inverting stage, and therefore the accuracy of the synthesised impedance. There is much more on this technique in Chapter 11.

The subsonic filter is a two-stage third-order Butterworth high-pass filter that is −3 dB at 20 Hz, with paired resistors again used to improve accuracy. It is a two-stage filter in that it consists of an under-damped second-order high-pass stage (C40, C41, R40−R43 etc.) combined with a first-order stage (C43, R44, R45) to give the maximally-flat Butterworth response. My previous preamp designs have used a single-stage third-order Butterworth, but I have found the two-stage configuration is preferred when seeking the best possible distortion performance [37]. An LM4562 is used in the filter as it significantly reduces distortion and its higher current noise makes a negligible contribution to the overall noise performance.

The IEC Amendment is an extra LF roll-off that was added to the RIAA spec at a later date. Most people regard it as unwelcome, so it is sometimes just omitted. Here it can be switched in by placing an extra resistance R46 across the subsonic filter resistances R44, R45. This is something of an approximation, but saves an opamp stage and is accurate to ± 0.1 dB down to 29 Hz. Below this the subsonic filter roll-off begins and the Amendment accuracy is irrelevant.

The switched-gain stage comes last in the signal path. This stage allows every MC and MM cartridge currently on the market, including the very low output MC models made by Audio Note, to receive the amount of gain required for optimal noise and headroom. The gain stage is fully described earlier in this chapter; gain is varied in 5 dB steps by a switch SW3 which selects the desired tap on the negative-feedback divider R51−R60. Each divider step is made with two resistors, with their values as close as possible, to get the exact value and improve accuracy. R35 provides continuity of DC feedback when the switch is altered.

Setting the correct gain can be done by reference to cartridge sensitivities and so on, but it is far more convenient to have some sort of level indication for guidance, hence the meter output shown in Figure 8.41. The Log Law Level LED (LLLL) is my attempt to get as much useful information as possible from a single LED; it is to the best of my knowledge, a new idea, and is fully described in Chapter 23 on metering.

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CHAPTER 9

Moving-magnet inputs: archival and non-standard equalisation

Archival transcription

This chapter deals with the differing requirements for reproducing discs, or indeed cylinders, that are not made to the microgroove standard. The subject of archival transcription is a large and complex one, and here I concentrate on the electronics. For disc replay there are many factors relating to stylus size and shape, tracking weight, etc. which we cannot explore here.

The major electronic problem is replay equalisation. Forget about having one universally-accepted equalisation, like the RIAA characteristic for microgroove discs. As you have seen in earlier chapters, implementing the RIAA both accurately and economically is quite a challenge. Early discs used a wide variety of equalisations and so an archival phono preamplifier must be able to provide all these in an accurate manner, though there is no need for great economy in specialised equipment that will only be made in very small numbers.

Although the media involved is always monaural, a stereo cartridge is normally used for transcription, and archival preamplifiers are likewise stereo, with a facility for summing varying proportions of the two channels to create a final mono output. This is because the two walls of a mono groove are unlikely to be identical and the best results may be obtained by summing them unequally.

The preservation and distribution of archival material in analogue formats requires that it be replayed with the highest quality attainable and converted to digital. The most comprehensive book I am aware of on the general subject is the *Manual Of Analogue Sound Restoration Techniques* [1] by Peter Copeland [2].

Coarse groove discs

The great majority of the material for archival transcription is represented by coarse groove discs, often referred to as '78s' because they were designed to rotate at 78 rpm rather than the 33 rpm of microgroove discs. They were produced from 1898 until the mid-fifties,

TABLE 9.1 Approximate groove dimensions for coarse and microgroove discs

	Coarse groove		Microgroove	
	Micron (μm)	Mil (inch/1000)	Micron (μm)	Mil (inch/1000)
Groove width	150	6	50	2
Groove depth	75	3	25	1
Groove bottom radius	38	1.5	8	0.3
Flat inter-groove spacing	100	4	38	2.5
Stylus tip radius	63	2.5	12.5–23	0.5–1

and were composed of mineral powders in an organic binder ('shellac') which gave much higher surface noise than the vinyl used for microgroove records. Playing time was limited to three minutes, later extended to six. Table 9.1 summarises the radical changes in groove dimensions required to get 22 minutes playing time from microgroove discs – completely different styli are required.

Wax cylinders

As I have made clear, I think vinyl is an obsolete technology. However, if you're going to be retro, I say no half-measures. The first medium for recording and playback was tin-foil on a cylinder, introduced by Edison in 1877, but the results were poor even by the standards of the day. Wax cylinders proved much better. A standard cylinder system was agreed by Edison Records, Columbia Phonograph, others in the late 1880s. These cylinders were 4 inches (10 cm) long, 2¼ inches in diameter, and played for about two minutes. The grooves made 100 turns per lateral inch.

Disc records appeared in 1901, first 10-inch then, later, 12-inch. These played for about three and four minutes respectively, but had poor quality compared with contemporary cylinders. To meet the competition on playing time, Edison introduced the Amberol cylinder in 1909, which increased the maximum playing time to 4.5 minutes, turning at 160 rpm, by increasing the groove density to 200 turns per lateral inch. Blue Amberol cylinders came later, with a celluloid playing surface on a core of plaster of Paris. Despite these developments, discs decisively defeated cylinders in what must have been the first audio format war, though Edison continued to produce new cylinders until October 1929.

Wax cylinders have certain advantages. The format is inherently linear-tracking, which eliminates all the problems of angular alignment and stylus side-force. A worm gear is used to move the stylus in alignment with the grooves on the cylinder, whereas disc replay uses the grooves to pull the stylus across the playing surface, creating a side-force and increasing groove wear. The speed of the stylus in the groove is constant, which eliminates the end-of-side distortion on the inner grooves of discs. Around 1900, it was acknowledged that cylinders in general had significantly better audio quality than discs, but by 1910 disc technology had improved and the difference disappeared.

Probably the greatest disadvantage of cylinders compared with discs is the space they take up. The volume on the inside of the cylinder has no use beyond supporting the player surface. Discs, however, can be stacked in a compact pile with no wasted space.

Various one-off cylinder replay machines have been built for the transcription of old recordings. These include the Archéophone [3] designed by Henri Chamoux, in 1998, and a cylinder player built by BBC engineers in the early 1990s. The latter used a linear-tracking arm from a contemporary turntable and an Ortofon cartridge. It could use a wide range of different equalisations and was used to transfer archival content to DAT tapes. Cylinders can also be 'replayed' by optical scanning, which has the great advantage that it can cause no groove damage.

It is impossible to give a single correct equalisation characteristic for the replay of cylinders. The recordings were made acoustically, without any standard electrical equalisation being applied as would later be done as part of the RIAA standard, and each case must be decided on its merits. Variable equalisation preamplifiers exist, but are rare and expensive. There is information on how to build them later in this chapter.

It is, as you might imagine, now extremely unusual for music to be released in cylinder format. A notable (and probably unique) exception being the release of the track 'Sewer' in 2010 by the British steampunk band The Men That Will Not Be Blamed For Nothing [4]. This was a very limited edition indeed; only 40 cylinders were produced and only 30 were put on sale.

Non-standard replay equalisation

The familiar RIAA curve for microgroove records has three basic corner frequencies, f3, f4 and f5, plus f2 if you count the IEC amendment. This is the Lipshitz numbering convention. Early replay equalisation was simpler than this, with only f4 specified as the 'bass turnover frequency' below which the signal received bass boost at 6 dB/octave.

The RIAA curve is usually specified in terms of its time-constants. These have the Lipshitz names of T3, T4 and T5, corresponding to the turnover frequencies f3, f4, and f5, and their values are 3180 µs (f3 = 50.05 Hz), 318 µs (f4 = 500.5 Hz) and 75 µs (f5 = 2112 Hz). The time-constant T2 is the IEC amendment of 7960 µs (f2 = 20.02 Hz) but as we saw in Chapter 8, this merely acts as a feeble substitute for a subsonic filter, and is of no interest here as it is only intended to be used with microgroove (LP) discs. Any preamp intended for archival transcription is almost certain to have a proper subsonic filter, often with a variable cut-off frequency. Non-standard equalisation is more often defined in terms of the frequencies rather than the time-constants. The two are easily interconverted; see Equations 9.1 and 9.2.

$$f = \frac{1}{2\pi T} \qquad \text{(Equation 9.1)}$$

$$T = \frac{1}{2\pi f} \qquad \text{(Equation 9.2)}$$

All cartridges (except for very rare strain-gauge types) respond to the velocity of the stylus, and not its displacement. If a recording is made to suit this, then the groove amplitude will continuously decrease from LF to HF; this is called constant-velocity recording. This causes two problems: excessively large groove amplitude at LF which limits playing time and at some point becomes untrackable, and excessively small groove amplitude at HF so surface noise becomes a problem. Early recording equalisation, from the mid-1920s, tackled the first problem by introducing a 6 dB/octave LF roll-off when recording, starting at a set frequency, f4, and giving constant-amplitude recording below that. An example is curve A in Figure 9.1, which corresponds to the Radiofunken characteristic in Table 9.2. In the literature, f4 is often called the 'bass turnover frequency'. To undo this at replay, complementary LF boost had to be provided, starting at the same frequency. This boost had to be curtailed before sub-sonic frequencies were reached, or record warps and turntable rumble would be grotesquely exaggerated. Therefore replay

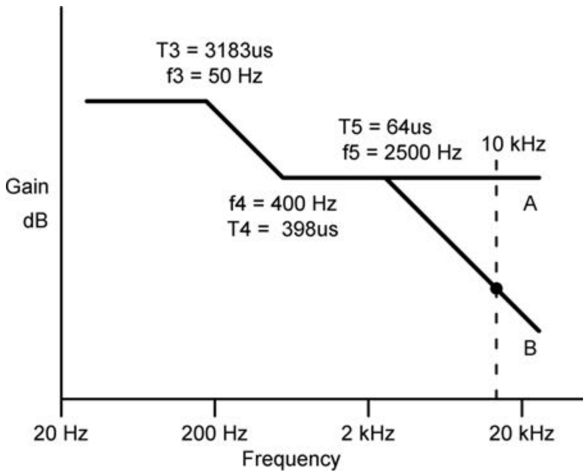

Figure 9.1: Early equalisation systems used curves like A, later ones like B added de-emphasis to attenuate surface noise. f5 is sometimes specified in terms of the attenuation at 10 kHz

TABLE 9.2 Generally agreed replay equalisation frequencies (without de-emphasis)

Manufacturer or organisation	f4 (Hz)
Acoustic gramophone	0
AES standard	400
Brunswick	500
Columbia (UK)	250
Decca 78	150
Early 78s (mid-1930s)	500
EMI (1931)	250
HMV (1931)	250
NAB standard	500
New Records	750
Oriole	Inconsistent
Parlophone	500
Pathe	Inconsistent
Radiofunken	400

equalisation always had another frequency, f3, where the gain flattened out; I have used 50 Hz in Figure 9.1. Since f3 was not part of any official recording characteristic until about 1940, it seems to have been left to the judgement of those designing replay amplifiers.

Later, from the mid 1930s, the second problem of excessively small groove amplitude at HF was addressed by increasing the HF content recorded. This is called pre-emphasis and relies on the fact that an audio signal has relatively low levels at HF. At replay, HF cut (de-emphasis) is applied to undo the pre-emphasis, and this attenuates the surface noise (see curve B in Figure 9.1, which corresponds to the Decca 1934 characteristic in Table 9.3). In the literature, f5 is often called the 'treble transition frequency'. The HF cut curve also dropped at 6 dB/octave, for ease of implementation, except in the case of Decca's FFRR (full frequency range recording) records which fell at a 3 dB/octave slope. I have found no information on how this was supposed to be implemented – presumably it was approximated by using extra overlapping time-constants. See my crossover book [5] for how to do this with varying degrees of accuracy.

There were dozens of different recording characteristics used and choosing the corresponding replay characteristic is a significant problem in itself. The history of even one record label can be complicated with different characteristics used in different years and, according to Gary Galo, sometimes even on different sides of the same disc. The histories of the characteristics

TABLE 9.3 Generally agreed replay equalisation frequencies (with de-emphasis)

Manufacturer	f4 (Hz)	f5 (Hz)	10 kHz attenuation (dB)
Capitol (1942)	400	2500	−12
Columbia (1925)	200 (250)	5500 (5200)	−7 (−8.5)
Columbia (1938)	300 (250)	1590	−16
Decca (1934)	400	2500	−12
Decca FFRR (1949)	250	3000*	−5
London FFRR (1949)	250	3000*	−5
Mercury	400	2500	−12
MGM	500	2500	−12
Victor (1925)	200–500	5500 (5200)	−7 (−8.5)
Victor (1938–47)	500	5500 (5200)	−7 (−8.5)
Victor (1947–52)	500	2120	−12

* 3 dB/octave slope above f5

are not universally agreed and are subject to some debate, much of which can be found on the internet. In Tables 9.2 and 9.3 I have done my best to give some generally accepted examples, including the extremes, so we can see what range of frequencies is required in an archival preamp. It is far from comprehensive, and much greater depth of information can be found in Peter Copeland's book [1] (other useful references are [6] and [7]). In most of the literature, the frequencies f4 and f5 are used rather than time-constants. Sometime f5 is not quoted directly, but in terms of the attenuation at 10 kHz with respect to 1 kHz. In difficult cases the only thing to do is to judge by ear the correct characteristic to use, and this has implications that will surface later.

The entry for 'acoustic gramophone' in Table 9.2 indicates that no equalisation at all was used, giving a straight constant-velocity characteristic. This was suitable for purely mechanical reproducers such as the Orthophonic Victrola with its folded exponential horn. Note the entries marked 'inconsistent'.

Tables 9.2 and 9.3 show that f4 must be variable at least between 150 and 750 Hz, and Table 9.3 shows that f5 must be at least variable between 1590 and 5500 Hz. Having decided on f3, f4 and f5 you have to implement them. In Chapter 8 I stated firmly that RIAA equalisation should be done in one active series-feedback stage, and demonstrated that all other approaches involving partly-passive RIAA introduced serious compromises in noise or headroom, or both. The downside is that the active one-stage method is much more difficult to design because of the interacting values in the RIAA feedback network.

For archival transcription the priorities are different. If f3, f4 and f5 are combined in one network, changing one of them requires recalculating the whole network and then changing every component in it. This is obviously impractical, and the only answer is to use some form of partly-passive configuration that allows the time-constants to be set independently. Increased electronic noise is not a problem because it will be well below the surface noise of ancient media. Headroom is likewise unlikely to be an issue because of the generally low recording levels, and the ease of including a flat variable-gain stage in a partly-passive configuration.

Figure 9.2 shows a typical passive-passive equalisation circuit (so-called because both equalisation stages are passive and the amplifiers are separate). It follows the circuitry discussed in Chapter 8 by having a gain of +30 dB at 1 kHz, so 5 mV in gives 158 mV out. The values of the equalisation components are taken from Gary Galow's *Linear Audio* article [8] for reasons that will become clear; they give an RIAA response accurate to ±0.02 dB. As noted, the IEC amendment f2 is unlikely to be used, but if required it can be implemented by making C0 equal to 79.05 μF as shown. If f2 is not required, C0 should be 220 μF or 470 μF. In this circuit f3 is set by the time-constant T3, the product of (R2 + R3) and C1. Likewise f4 is set by the time-constant T4, the product of R3 and C1. The treble turnover frequency f5 is set simply by R6 and C2; the output must not be significantly loaded, and in many cases a buffer stage will be needed. Here it is very clear that 7.5 kΩ and 10 nF give a T5 of 75 μs, so f5 is 2112 Hz. If f5 is not used then R6 and C2 are simply omitted. It is normal practice to make the time constants switchable. Making them fully variable by using ganged pots would degrade the matching of the stereo channels.

Sometimes you have to judge by ear the 'correct' characteristic to use, or at least one that gives satisfactory results. When doing this, it is very convenient if the gain stays the same at 1 kHz while the frequency extremes are altered. This does not occur with the simple

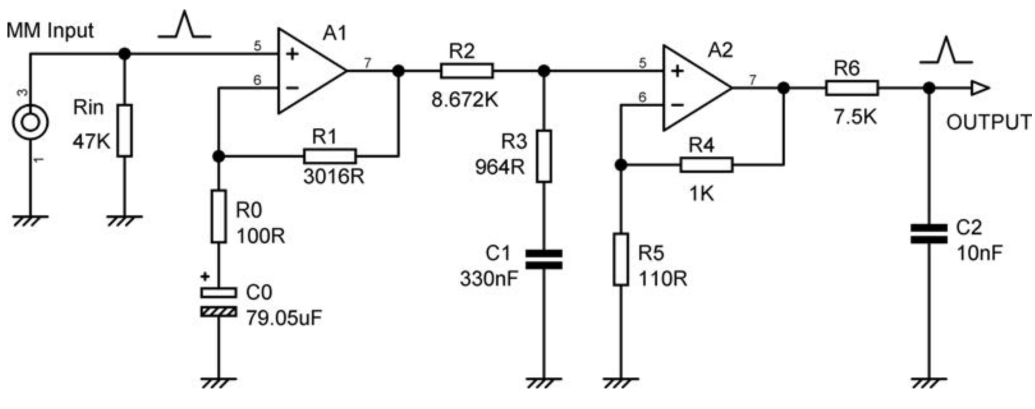

Figure 9.2: Passive-passive equalisation configuration

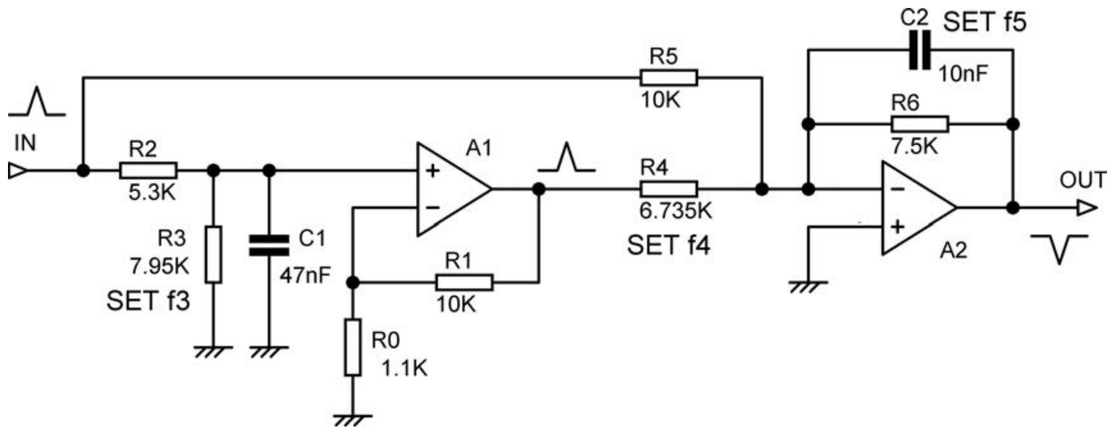

Figure 9.3: Partly-passive equalisation with constant level at 1 kHz

circuit of Figure 9.2, and great credit goes to Gary Galow and Mike Shields for devising the circuit in Figure 9.3 [8]. The values here give the standard RIAA curve, to allow easy checking of the circuit operation; when simulated using 5534s it is accurate to ±0.02 dB. The circuit works by summing the path through R5, which sets the 1 kHz gain, with the path through R4, which sets f3 and f4. Once again it is easy to see that R6 = 7.5 kΩ and C2 = 10 nF gives a T5 of 75 us and thus an f5 of 2112 Hz. The equations linking f3 and f4 with the related component values are more complicated but are fully explained in [8].

This circuit is purely an equaliser and not an amplifier; it has unity gain at 1 kHz. Be aware that adding a C0 in series with R0 will *not* give an accurate IEC amendment f2, because of the two paths.

Scratch filters

In the First Age of Vinyl (say 1948 to 1983 if we restrict ourselves to microgroove records), hifi preamplifiers commonly included a switchable low-pass filter, called, with no attempt at circumlocution, the scratch filter. The roll-off slope would be 12 or 18 dB/octave, commencing roll-off in the range 3–10 kHz. It was intended for the suppression, not just of ticks and clicks from scratched discs, but also the smoothing out of record surface noise and the dulling of the inevitable HF distortion. Upmarket preamplifiers would have two or more switched filter frequencies, and switchable or variable filter slope as well. The need for such versatile filtering is clear when transcribing archival discs in less-than-perfect condition. Note that a scratch filter is quite distinct from an ultrasonic filter (see Chapter 8), the latter only operating outside the audio band.

Variable-slope scratch filters: LC solutions

Making a variable-slope filter is not that straightforward, because the natural slopes you get with resistors and capacitors are 6 dB per octave. Using active filters gives access to final slopes with 12, 18, 24 or more dB per octave, but intermediate slopes are usually only found in the transitions between flat and the ultimate roll-off slope.

A popular way of achieving variable-slope back in the valve era involved an LC filter bypassed by a variable resistance. A classic example of this approach, published in *Wireless World* in 1956 [9], is shown in Figure 9.4. If RV1 is absent, there is a deep notch centred on 10 kHz, but adding it abolishes the notch and gives the response shown in the bottom trace of Figure 9.5.

Adjusting RV1 to steadily reduce the resistance across L1, C2 gives a set of responses that have more or less the same turnover frequency (−3 dB at about 3 kHz) but reducing slope. If we measure the average slopes across the octave 5–10 kHz, we get Table 9.4, which shows a

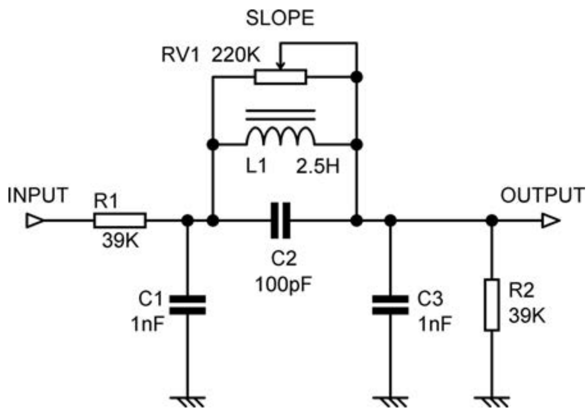

Figure 9.4: Historical variable-slope LC low-pass filter based on notch filter with bypass resistance RV1: 1956

TABLE 9.4 LC filter slope in dB/octave over 5–10 kHz

Bypass resistance	Slope (dB/octave)
220 kΩ	17.6
100 kΩ	11.7
47 kΩ	7.8
22 kΩ	5.7
10 kΩ	4.8
0 Ω	4.5

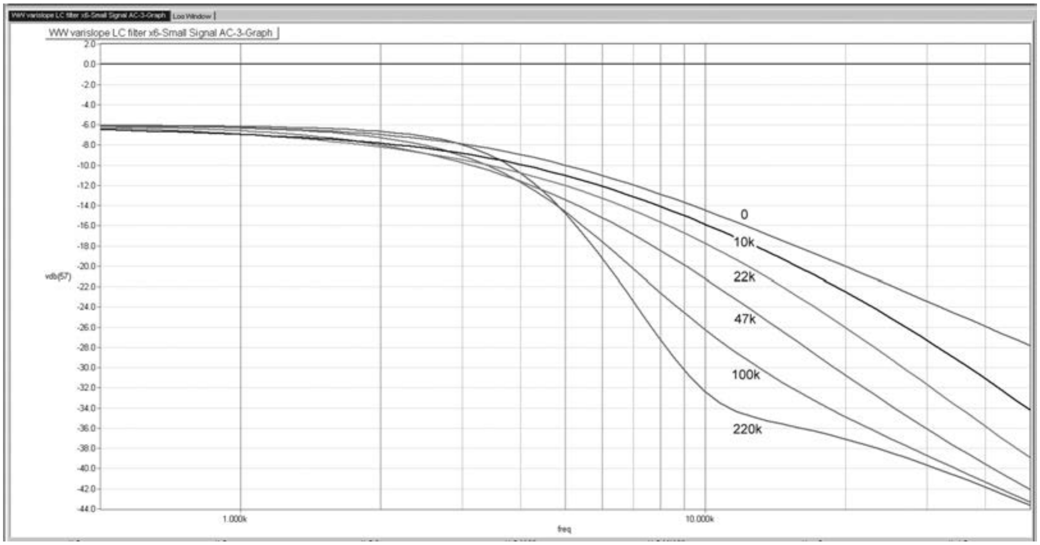

Figure 9.5: Variable-slope LC low-pass filter response for varying bypass resistance

handy variation in slope from 4.5 to almost 18 dB/octave. As revealed in Figure 9.5, the filter has 6 dB loss in the passband, which is not ideal.

The full published circuit included switching of all three capacitors to give nominal turnover frequencies of 5, 7, and 10 kHz, which of course means a rather clumsy 6-way switch for stereo use. It had a 1 H inductor, and even that reduced size required 1100 turns to be wound on a Ferroxcube core, so its construction required some degree of commitment. RV1 was a log-law component. This form of filter resurfaced in a transistor preamplifier design by Reg Williamson in 1967 [10]. In contrast, the well-known Leak Varislope preamplifiers used purely RC filtering, and only offered two switched slope settings.

Variable-slope scratch filters: active solutions

The prospect of winding 1100 turns on a former to make an inductor is not at all appealing, and would turn anybody's mind to RC active filters. This is quite apart from the well-known inductor drawbacks of weight, cost, non-linearity, and susceptibility to hum fields. In 1970 John Linsley-Hood published an RC variable-slope filter design [11], but its operation was quite unacceptable, having irregularities in the passband down to 100 Hz, and +3 dB of internal peaking that eroded headroom. In 1990, Reg Williamson most ingeniously converted the LC filter to a more practical form by replacing the floating inductor with two gyrators [12]; the downside was that it required five opamp sections.

When I needed a varislope filter in 1978 I decided to have a go at designing my own version. The first attempt was adding a bypass resistance to a standard second-order Butterworth

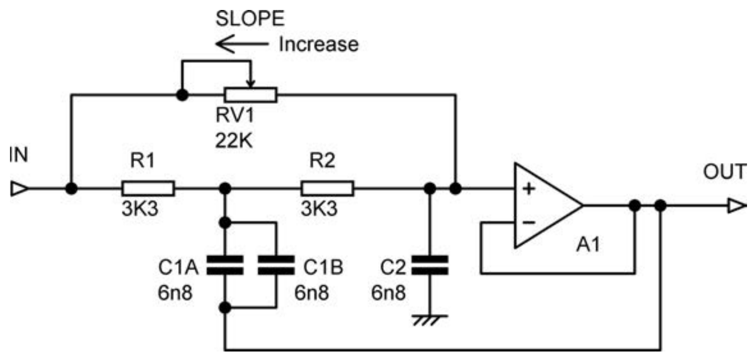

Figure 9.6: Variable-slope active RC second-order low-pass filter with bypass resistance RV1 added

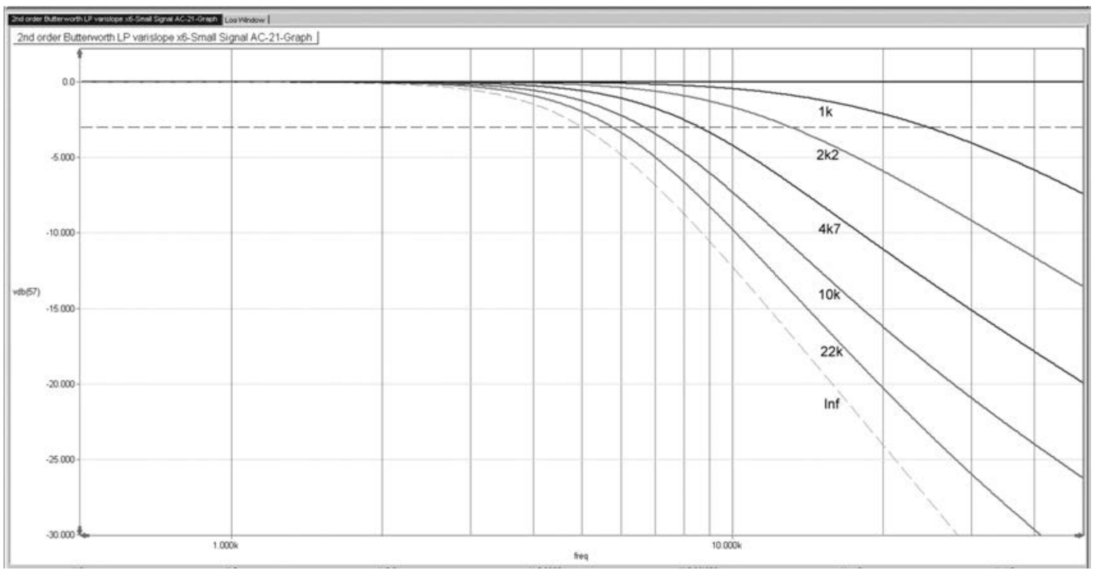

Figure 9.7: Variable-slope active RC second-order low-pass filter response for varying bypass resistance

Sallen and Key filter, as shown in Figure 9.6, with the response results in Figure 9.7, where the dashed bottom trace shows the pure second-order Butterworth characteristic obtained when the bypass resistance is entirely absent. The slopes are indeed varying smoothly, and are summarised in Table 9.5. Naturally the maximum slope is somewhat less than 12 dB/octave, since we started out with a second-order filter. The turnover frequency varies with the slope, though this is not necessarily a disadvantage. We are trying to make tolerable the reproduction from a very imperfect medium, not design a laboratory instrument. Here, increasing the bypass resistance gives 'more filtering' in two ways because the turnover frequency is reduced as the slope is increased.

TABLE 9.5 RC second-order filter slope in dB/octave over 5–10 kHz

Bypass resistance	Slope (dB/octave)
None	11.8
22 kΩ	10.5
10 kΩ	8.9
4k7	6.9
2k2	4.2
1 kΩ	1.7

In practice, adding a 1 kΩ end-stop resistor in series with the slope pot would probably be a good idea, as this will prevent C2 being connected directly to the output of the previous stage, which may object by going unstable. It is assumed there is a way of switching out the filter stage completely.

While the second-order low-pass filter with bypass resistance delivers variable slopes quite well, it is doubtful if a maximum slope of 12 dB/octave is really enough; the historical LC filter gave a maximum of almost 18 dB/octave.

I therefore took another swing at the problem by starting with a third-order Butterworth Sallen and Key filter, as in Figure 9.8. I have not converted the exact capacitor values to combinations of preferred values. Connecting a bypass resistance between the input and C3 gives no useful result, but wiring it between C1 and C3 gives the response in Figure 9.9. There is something like a constant turnover frequency, but this does not align with the dashed line at −3 dB; rather it occurs around −7 dB. A variable-slope characteristic is obtained with slopes of between 5 and 15 dB/octave, as summarised in Table 9.6.

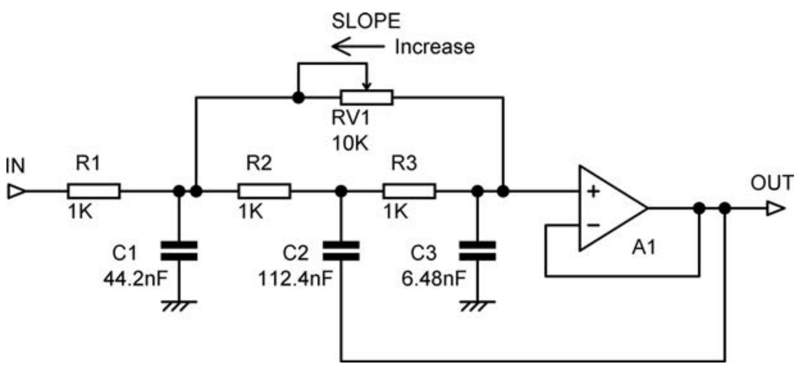

Figure 9.8: Variable-slope active RC third-order low-pass filter with bypass resistance RV1 added

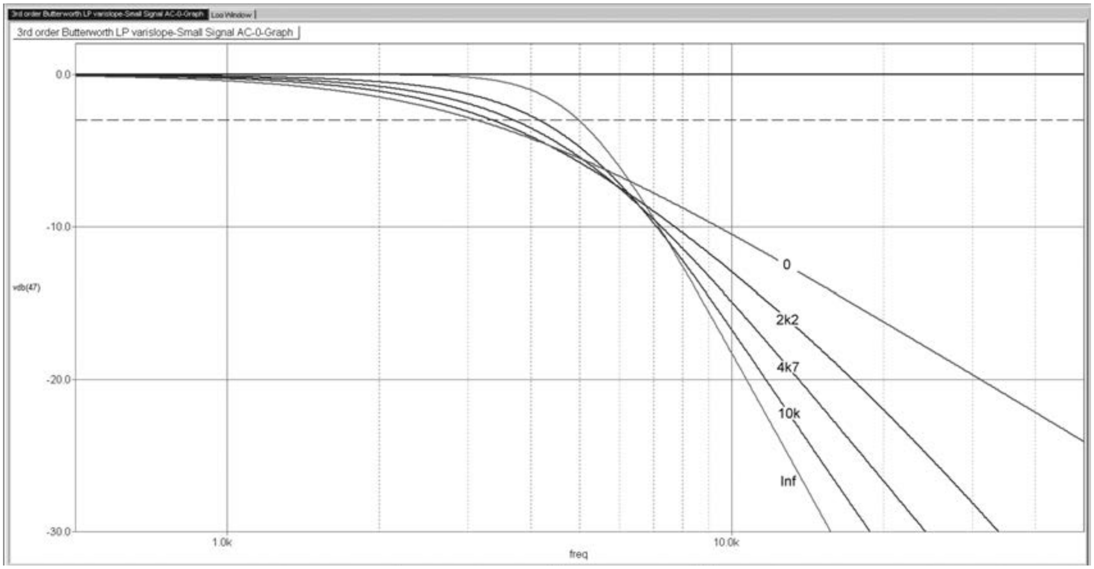

Figure 9.9: Variable-slope active RC third-order low-pass filter response for varying bypass resistance

TABLE 9.6 RC third-order filter slope in dB/octave over 5–10 kHz

Bypass resistance	Slope (dB/octave)
None	15.2
10 kΩ	12.0
4k7	9.4
2k2	7.1
0 Ω	5.0

There is now no need to add an end-stop resistor in series with the slope pot, as C3 is never connected directly to the input. I am not claiming that these bypass filters are the last word on the subject, but they are certainly more economical of parts than the Williamson gyrator filter. To the best of my knowledge this is the first time this technique has been published.

Variable-slope scratch filters: the Hamill filter

If slightly different design criteria are used, a different form of variable scratch filter results. In a 1981 article in *Wireless World*, [13] David Hamill concluded that low-pass filters with rapid roll-offs around the cutoff frequency introduced colouration, but this could be avoided if the area around cutoff had a slow roll-off to prevent ringing on transients. The Gaussian filter characteristic is optimised for its time response, giving no overshoot and minimum rise and fall times on edges, but it gives a slow roll-off when near the cutoff frequency. Hamill stated that making the roll-off Gaussian for the first 10 dB or so is enough to prevent ringing, but as the filter cutoff frequency increases the roll-off can be made faster as the ear becomes less sensitive to colouration. His design therefore had a variable slope around the cutoff frequency, transitioning to a fixed 18 dB/octave ultimate slope at higher frequencies, unlike the filters described earlier which have variable ultimate slopes.

One slight problem with Gaussian filters is that they are impossible to construct. The slow roll-off that gives the good time response is obtained by cascading a series of first-order stages of differing and carefully chosen frequencies; the Gaussian roll-off slope steadily increases with frequency and is infinite at infinite frequency. You therefore need an infinite number of first-order stages, which makes construction rather difficult. It is however entirely practical to build an *approximation* to a Gaussian filter, by keeping the slow initial roll-off, but then smoothly splicing this to a constant, and therefore more practical, filter slope such as the 18 dB/octave of a third-order Butterworth or Bessel filter. All real 'Gaussian' filters are therefore actually transitional filters.

The original published design included a fixed-frequency third-order subsonic filter, but its presence or absence does not affect the operation of the scratch filter, and it is omitted here for clarity. The filter very cleverly has only one control to set cutoff frequency and slope around cutoff. As the control is turned the cutoff frequency increases, and at the same time the slope around the cutoff increase, the filter characteristic moving from a Bessel approximation to a Gaussian filter, through Butterworth, to a Chebyshev response with 0.5 dB passband ripple. This single 'ripple' is a very shallow peak around 15 kHz and is highly unlikely to be perceptible. The control is a single pot making stereo operation with a dual-gang pot simple. I think we should call this a Hamill filter.

My interpretation of this filter is shown in Figure 9.10. The basic configuration is that of a two-stage third-order filter, but with capacitors C2 and C4 driven by a scaled version of the output voltage; the scaling factor is set by RV1, with R6 bending the control law so it approximates to logarithmic, i.e. linear in octaves. The circuit impedances have been reduced by a factor of 4.25 to improve the noise performance with modern opamps (the original design used discrete transistors), the precise factor being chosen to make the largest capacitor C3 exactly 20 nF, so it can be made from two 10 nF polystyrene capacitors C3A, C3B in parallel. This luckily gave a value of almost exactly 1.2 kΩ for R1, R2, and R3. The other capacitors C1, C2 and C4 inevitably have awkward values; they are made up assuming only E6 series components are available. As luck would have it, C1 and C4 come out very nicely, their combined nominal values being within 0.1% of the target, though three parallel capacitors C4A, C4B, and C4C are required to achieve this for C4. C2 is a bit less favourable, coming out 1% high, but this seems to have very little effect on

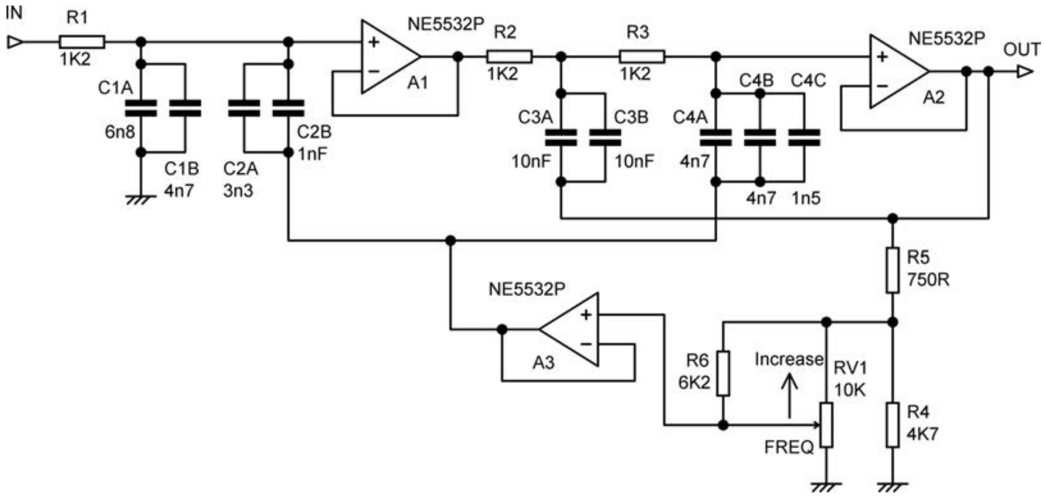

Figure 9.10: Hamill filter variable-slope active RC third-order low-pass filter response

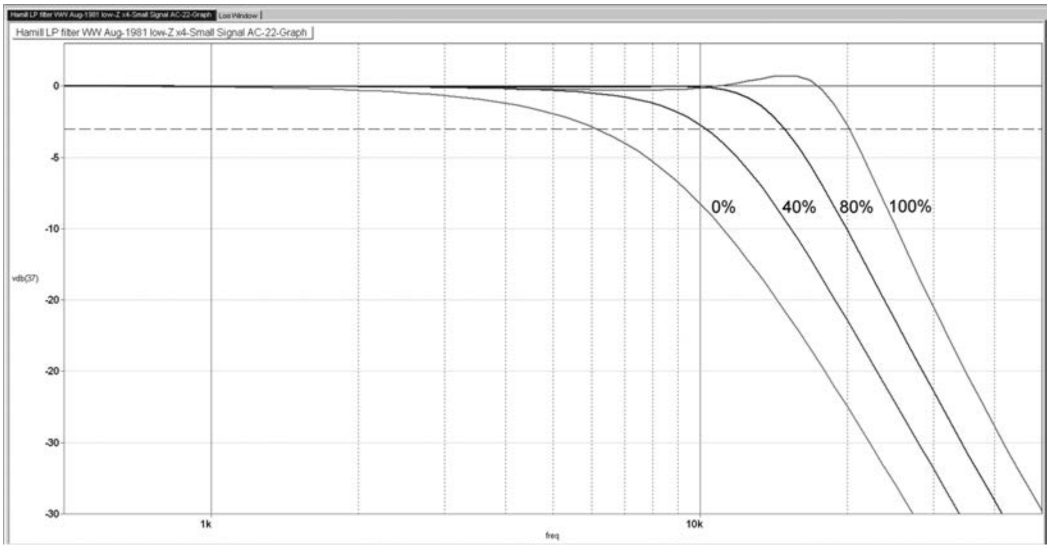

Figure 9.11: Hamill filter with variable-slope around cutoff, merging into a third-order 18 dB/ octave roll-off. Four control settings shown

the responses, which as far as the eye can judge are identical to those published in *Wireless World*.

The frequency responses for four control settings as the pot wiper is moved upwards are shown in Figure 9.11, and it can be seen that the Hamill filter does its work most effectively and ingeniously. It should prove useful for archival transcription work.

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CHAPTER 10

Moving-magnet inputs: discrete circuitry

Discrete MM input stages

Discrete moving-magnet (MM) input amplifiers were almost universal until the early 1970s. For a long time opamps had, quite deservedly, a poor reputation for noise when used in this application. When the first bipolar transistor MM inputs were designed, active components were still expensive, and adding another transistor to a circuit was not something to be done lightly. The circuitry from that era therefore looks to us very much cut-to-the-bone, and before disrespecting it we need to remember that it was designed under very different economic constraints.

A major problem with early discrete MM amplifiers was a simple lack of open-loop gain to give an accurate RIAA response network in the low frequency region, even if the RIAA network was accurate, which it rarely was. Another problem was that an RIAA feedback network, particularly one designed for low closed-loop gain, and/or a relatively low RIAA network impedance to reduce noise, presents a heavy load at high frequencies because the impedance of the capacitors becomes low. Heavy loading at HF was commonly a major cause of increased distortion and headroom-limitation in discrete RIAA stages that had either common-collector or emitter-follower output topologies with asymmetrical clipping behaviour; an NPN emitter-follower is much better at sourcing current than sinking it. The 20 kHz output capability, and thus the overload margin, was often brought down by 6 dB or even more. Replacing the emitter resistor of an emitter-follower with a current-source gives a much better HF output current capability, and this can be further doubled for the same quiescent dissipation by using a simple push-pull Class-A output structure.

One-transistor MM input stages

I will say at once that the one-transistor MM input stage is purely a historical curiosity. Its performance cannot be expected to be anything other than dreadful by today's standards. Nevertheless, the idea is worth looking at. Figure 10.1 shows a one-transistor MM input stage designed by Jack Dinsdale (of whom more later) in 1961 [1]. At that time transistors were *very* expensive, and using two in a single stage would have been thought highly extravagant.

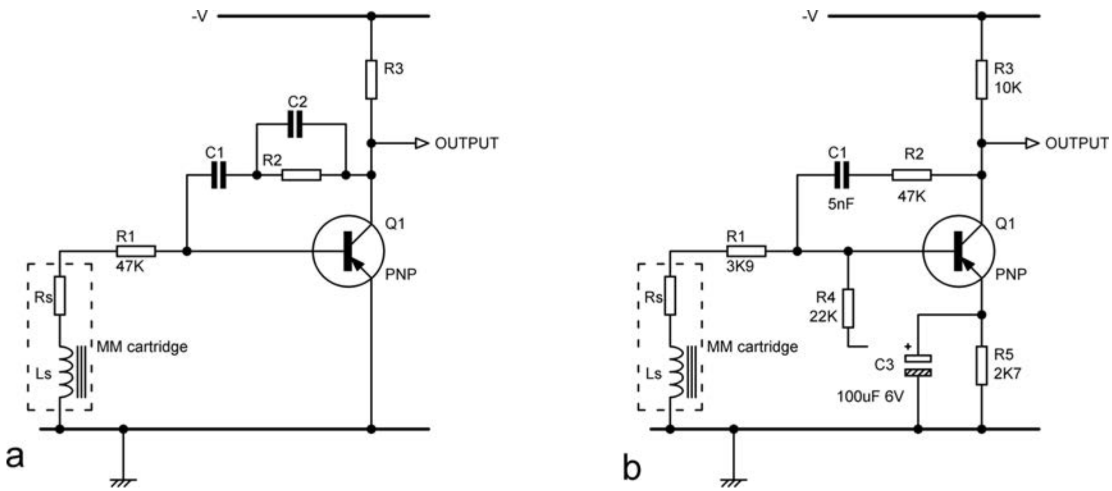

Figure 10.1: a) Basic one-transistor shunt-feedback MM amplifier, b) using the cartridge inductance to perform the LF part of the RIAA equalisation (Dinsdale 1961)

If you have but one transistor to play with, there are only three possible configurations; common-collector, (i.e. emitter-follower) common-base, and common-emitter. The first gives no voltage gain, and the second has a low input impedance that looks unpromising. That leaves the common-emitter configuration, as in both halves of Figure 10.1. Since it inherently inverts, the only possibility is shunt-feedback, which, as we saw in Chapter 8, is inherently much noisier than its series-feedback equivalent. The standard approach at the time, derived from valve designs, was that in Figure 10.1a, which had an input resistor R1 of 47 kΩ to give the correct cartridge loading, with the RIAA equalisation performed by the negative feedback network C1, R2, C2 in conjunction with the impedance of the collector load R3.

This arrangement is bound to give a poor noise performance. Dinsdale's solution [1], in Figure 10.1b, was to make the input impedance low and implement the LF part of the RIAA equalisation by the interaction of the cartridge inductance with it, giving a 6 dB/octave slope. As frequency falls the impedance of the inductance falls and the current into the input increases. He described this method as more 'efficient', which presumably means a greater transfer of energy through R1 and hence a better signal/noise ratio. Components R5 and C3 set the DC conditions, with base bias provided through R4.

The idea of loading the MM cartridge inductance with a low input resistance to achieve the LF boost section of the RIAA equalisation has come back to haunt us many times since then. The terrible snag is that since the LF equalisation is set by the cartridge inductance, changing the cartridge type almost certainly means you have to change the loading resistor too. You can, of course, add a control marked 'cartridge inductance', but this assumes you actually know the cartridge inductance, and know it precisely. Inaccuracy in the inductance setting will give errors in the RIAA response. You will have to rely on the manufacturer's technical specification for the value of the inductance unless you plan to measure it yourself.

Two-transistor MM input stages

Figure 10.2 shows a typical two-transistor MM input amplifier from the late 60s. The configuration is generally considered to have been introduced by Jack Dinsdale in 1965, in a classic preamplifier design [2] that was one of the first to deal effectively with the new RIAA equalisation requirements for microgroove records. It is a two-stage series-feedback amplifier composed of two common-emitter stages. R3 and C1 make up an RF filter; note R3 is 3k3. This is considerably greater than the DC resistance of most MM cartridges and looks like it would introduce unnecessary Johnson noise and turn the current noise of Q1 into voltage noise. The RIAA network is R6, R7, C4, C5, and it has a high impedance to reduce loading on the stage output. Since R6 has the high value of 1M8, the RIAA network cannot be used for the DC feedback that is required to set the quiescent conditions. There is a separate DC feedback network comprising R1, R4, R5, R10, and C3 which establishes the appropriate voltage across R10. C3 keeps signal frequencies out of this path. The RIAA network is in Configuration-A (see Chapter 8). No attempt is made to implement the IEC Amendment, as it was not introduced until 1976.

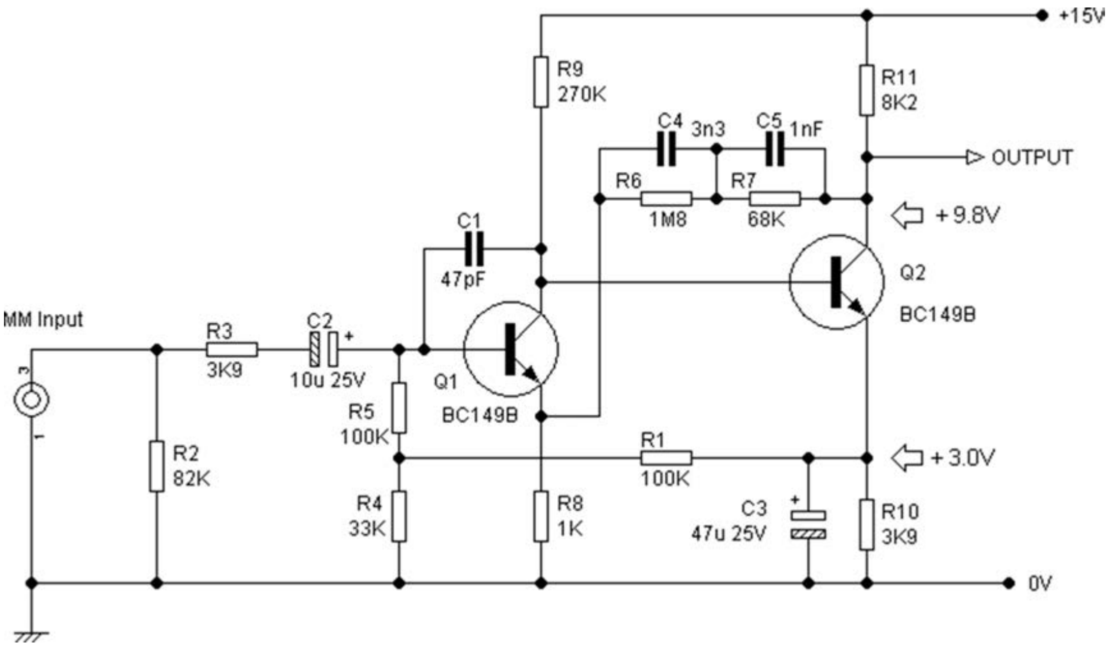

Figure 10.2: A typical two-transistor MM amplifier as commonly used in the 1960's and early 70's. Gain +39 dB at 1 kHz

Because of its simplicity, this stage inevitably contains compromises. The second collector resistor R11 needs to be high in value to maximise open-loop gain, but low to adequately drive the RIAA network and any external loading.

An MM preamp has to deliver a maximum low frequency boost of nearly 20 dB, on top of the gain required to get the desired output level at 1 kHz. If the cartridge output is taken as 5 mVrms at 1 kHz, and the amplifier output is 150 mVrms (which is about as low as you could hope to get away with then if you were sending this signal to the outside world) then a total closed-loop gain of 20 + 34 = 54 dB is required at low frequencies. The open-loop gain obviously needs to be considerably higher than this, for a decent feedback factor is required not only to reduce distortion, but also to ensure that the RIAA equalisation is accurately rendered by the feedback network. By 1970 it had become clear that the two-transistor configuration was really not up to the job, and more sophisticated circuits using three transistors or more were developed, aided by falling semiconductor costs. While the two-transistor MM preamplifier must now be regarded as of purely historical interest, it is highly instructive to see just what can be done with it by modification.

The circuit shown in Figure 10.2 was deliberately chosen as representative of contemporary practice in its era, and it has not been modified or optimised in any way. It is closely based on a small RIAA preamplifier PCB called the 'Lenco VV7', which was intended for upgrading systems to use MM cartridges where the amplifier had only a ceramic pick-up input (see Figure 10.3). It was a Swiss product distributed in Britain by Goldring in the early 1970's. It had an integral mains PSU (see the tiny transformer on the left) with half-wave rectification and RC smoothing. What the proximity of that transformer did to the hum levels I do not know, but it looks awfully close to the preamp, which is in the screening can to the right. You will note that the single-rail supply is, by modern standards, low at +15 V; opamp-based preamplifiers today normally run from ±15 V or ±17 V, giving them a 6 dB headroom advantage at once. The gain is +39 dB at 1 kHz

I built up Figure 10.2 with BC184 transistors, using an external DC supply. I found that the first-stage (Q1) collector current was 42 μA, and the second-stage (Q2) collector current was 0.63 mA. On measuring it I was not exactly surprised that the performance was mediocre. There was a high level of hum at the output: −66 dBu at 50 Hz. Careful screening of the

Figure 10.3: The Lenco phono preamplifier

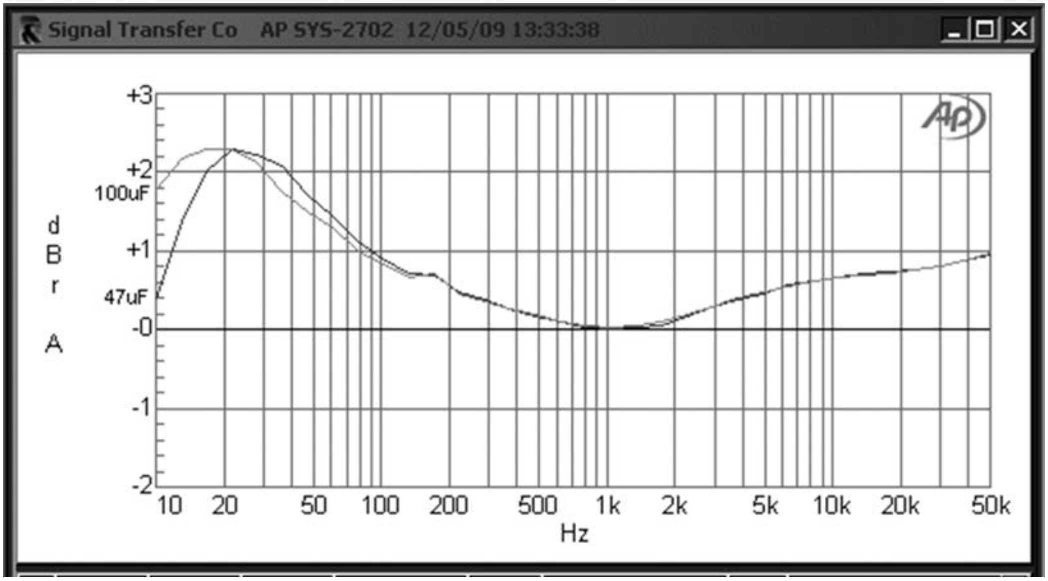

Figure 10.4: The RIAA errors are pretty gross by today's standards. Things are improved somewhat by increasing C3 to 100 μF

whole circuit only reduced this to −68 dBu, so electrostatic pickup was clearly not the only, or even the major, problem.

The RIAA equalisation accuracy, shown in Figure 10.4, is not good, which is only to be expected when you look at the standard component values in the RIAA network. Accurate RIAA networks cannot have more than one preferred value. The errors reach +2.3 dB at 20 Hz and +0.7 dB at 20 kHz; the IEC amendment is not implemented; it would have given an extra attenuation of −3 dB at 20 Hz and −1 dB at 40 Hz. The roll-off below 20 Hz is caused by C3. Increasing it from 47 μF to 100 μF much reduces the roll-off, and slightly improves RIAA accuracy between 20 and 200 Hz.

The preamplifier was being powered from a perfectly respectable bench PSU, but it still seemed possible that hum was getting in from the supply rail, as there is absolutely no filtering in the supply to Q1 collector. Inserting a 1 kΩ−22 μF RC filter in the supply to R9 dropped the noise output from −68 to −73.4 dBu (this figure is the average of six readings, to reduce the tendency of a noise reading to jump about when there is significant low-frequency content. Measurement bandwidth is always 22 Hz−22 kHz unless otherwise stated). A bandpass sweep of the noise output showed that there was now very little extra 50 Hz or 100 Hz content. The RC filter gives an attenuation of −16.9 dB at 50 Hz and −22.8 dB at 100 Hz. Increasing the filter capacitance to 100 μF however did give a slight improvement, so this was adopted; the attenuation at 50 Hz is now −29.9 dB.

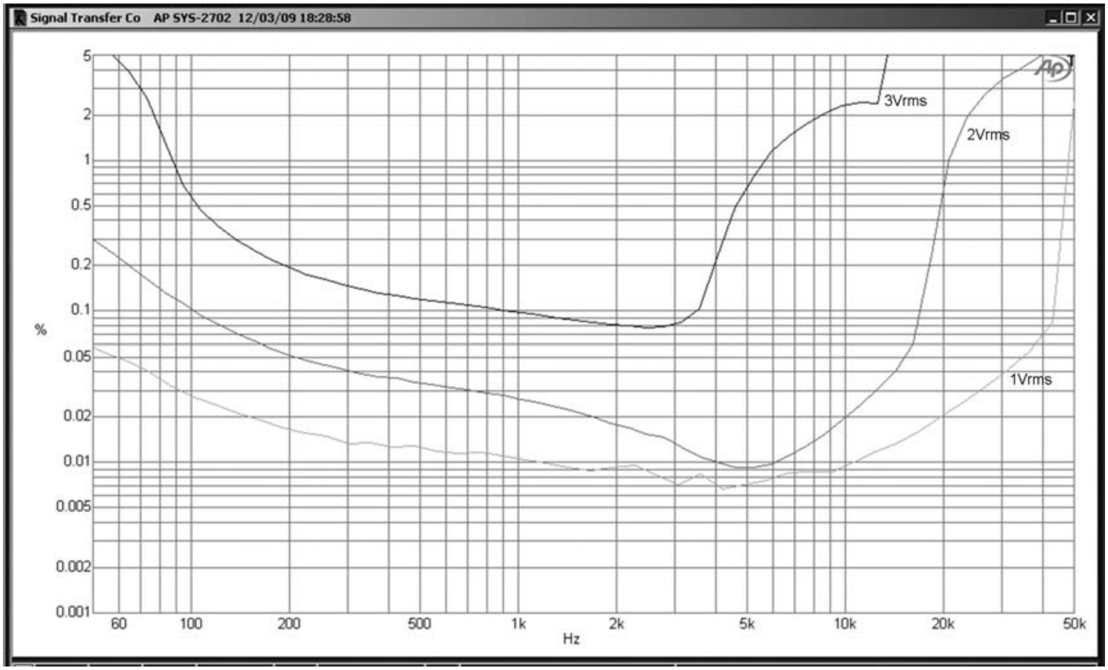

Figure 10.5: Two-transistor MM amplifier THD with a +15 V rail, at 1, 2, and 3 Vrms out. Bandwidth 100 Hz–80 kHz

Maximum output with a +15 V supply rail was 3.4 Vrms at 1 kHz (1% THD), and it is noticeable that clipping is not symmetrical, occurring first on the positive peaks. When this clipping does occur, there is a shift in the DC conditions of the circuit due to the way the biasing works through the filtering action of C3.

The THD at 1 Vrms out (1 kHz) was 0.010%, which by modern standards is a lot for such a low level. Figure 10.5 shows the distortion performance with a +15 V rail, at 1, 2, and 3 Vrms out. The input signal was inverse-RIAA equalised so that the output level remains constant with frequency. It was necessary to use the 100 Hz filter on the AP to get consistent results, despite having got rid of the 50 Hz problem with the RC filter, as there is still a large LF noise component due to the RIAA LF boost.

You can see that for 1 Vrms, the mid-band distortion is around 0.01%, but there is a steady rise below 1 kHz. This is caused by the falling negative feedback factor as the RIAA curve demands more gain at lower frequencies. The other area of concern is at high frequencies; at 1 Vrms nothing too bad happens in the audio band, though THD has reached 0.02% at 20 kHz.

At the higher output level of 2 Vrms, the mid-band THD is tripled. The output stage starts to clip around 15 kHz, as Q2 can no longer drive the RIAA network, which has a falling impedance at high frequencies. Things are pretty gross at 3 Vrms out, with THD around 0.1% mid-band, and HF clipping starting at 4 kHz.

Clearly this historical RIAA preamp could use a bit of improvement, starting with linearity and headroom. Let's see what can be done with it; the process will reveal a lot about how discrete circuitry works.

Two-transistors: increasing supply voltage to +24 V

A pretty sure bet for improving both the linearity and headroom of a discrete amplifier is simply to increase the supply voltage. We will start by turning it up to +24 V, a voltage that can conveniently be obtained from a 7824 IC regulator. Figure 10.6 shows the results: distortion is somewhat reduced overall and the HF overload problem has been pushed to slightly higher frequencies, but the effect is not as dramatic as we might have hoped. The maximum output has only increased to 3.8 Vrms at 1 kHz, (1% THD) which is not much of a return for increasing the supply voltage by 60%.

Casting a suspicious eye over the circuit, it's clear that it is still clipping asymmetrically. There is +18.4 V on Q2 collector, whereas for a symmetrical output swing we would expect something more like +12 V. Improving the bias conditions by changing R10 from

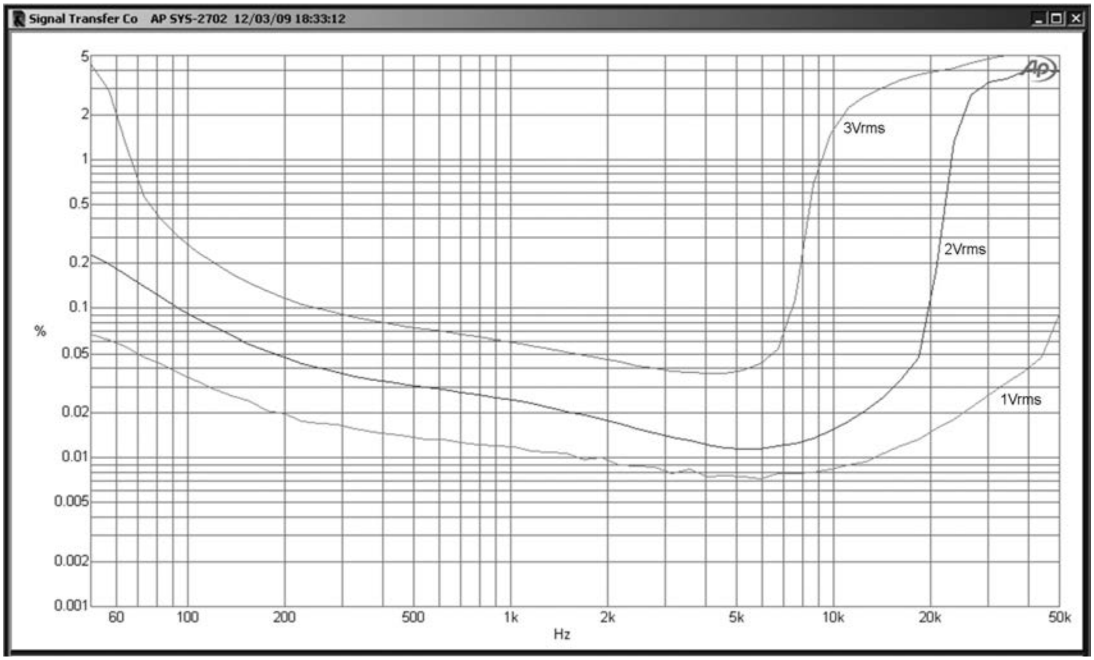

Figure 10.6: Two-transistor MM amplifier THD with a +24 V rail, at 1, 2, and 3 Vrms out. Bandwidth 100 Hz–80 kHz

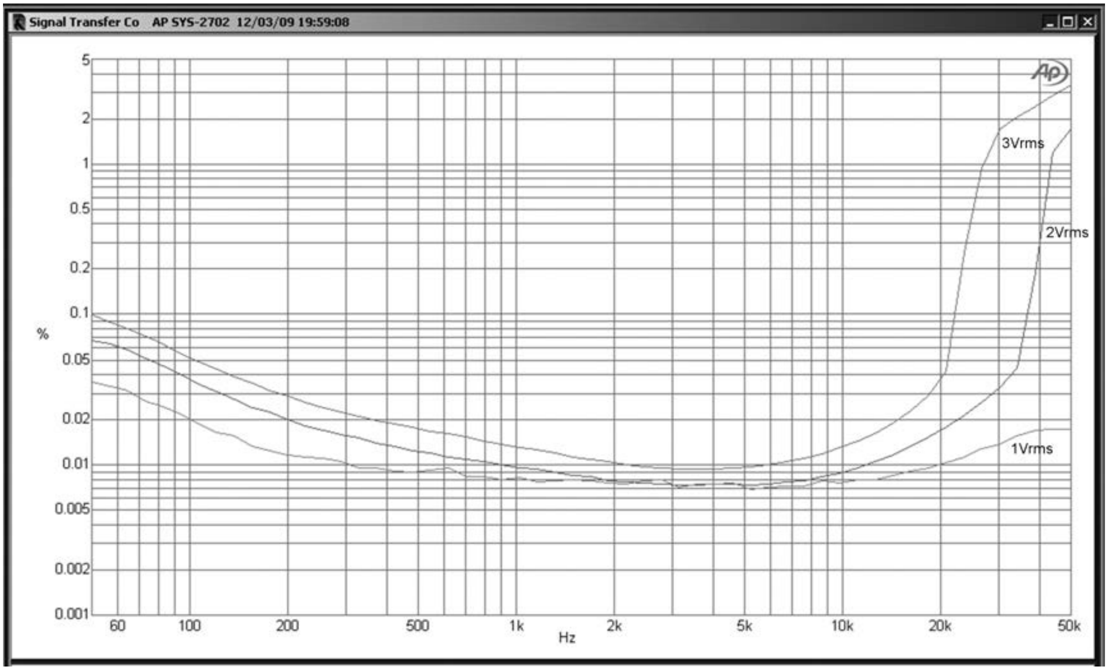

Figure 10.7: Two-transistor MM amplifier THD with a +24 V rail, at 1, 2, and 3 Vrms out, after rebiasing. Bandwidth 100 Hz–80 kHz

3k9 to 2k4 reduces Q2 collector volts to +15 V, and gives much more output voltage swing capability, as well as increasing the standing current in Q2, which improves load-driving capability. The maximum output is now 6.0 Vrms at 1 kHz, an improvement of +4 dB. While this does not give exact symmetry of clipping, it does seem to be close to optimal biasing for linearity. A good indication of this is that the distortion residual at 1 kHz is third-harmonic, which suggests that some cancellation of second-harmonic distortion is going on.

The distortion performance is transformed −3 Vrms out (1 kHz) gave 0.06% in Figure 10.6. After re-biasing it has fallen to 0.014%, as in Figure 10.7. The HF overload effect has also been pushed out to above 20 kHz, even for the 3 Vrms case. Not bad for modifications that essentially cost nothing.

The distortion improvement at lower output voltages in the mid-band is barely visible even with 100 Hz AP filtering because of the high noise output from a circuit with +39 dB of gain at 1 kHz.

The modified circuit, with the added RC filter for the first stage, supply increased to +24 V, and biasing adjusted by changing R10, is shown in Figure 10.8. The I_c of Q1 is now 75 μA, and the I_c of Q2 is 1.1 mA.

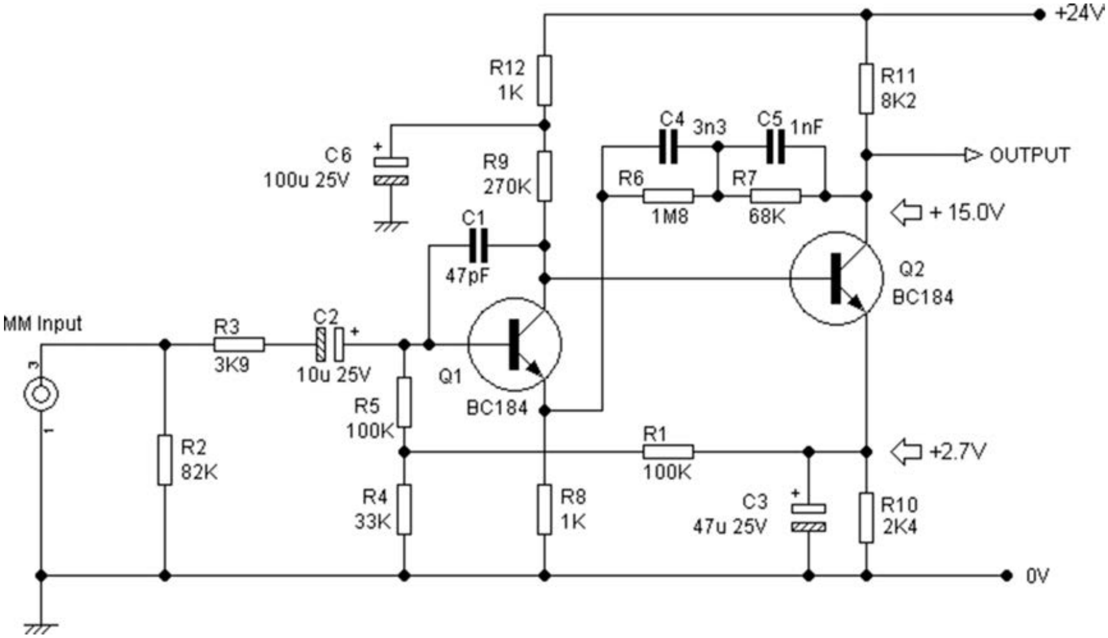

Figure 10.8: The two-transistor MM amplifier using a +24 V rail, with the RC filter R6, C12 added to the collector of Q1, and after rebiasing by altering R10

Two-transistors: increasing supply voltage to +30 V

Since increasing the supply to +24 V gave considerable benefits (after re-biasing), we will increase it further to +30 V. This increases the maximum output to 6.8 Vrms at 1 kHz (1% THD), which is 6 dB up on the original circuit. R10 has been changed again to 2k2 to optimise the biasing. The results are seen in Figure 10.9. The THD for the 1 Vrms case is now completely submerged in low-frequency noise, so I used the 400 Hz AP filter, which shows the THD at 1 Vrms (1 kHz) is about 0.0055%. This number however still contains a significant amount of noise.

HF overload behaviour has improved again, but HF distortion is somewhat worse, at all three output levels. The LF distortion is notably improved, being more than halved.

As a side issue, we might consider how to generate the supply rail required. The 7824 IC regulator will accept a maximum input of 40 V, so it is feasible to use that with the ADJ pin elevated by some means, as described in Chapter 25. For voltages above +30 V this does not leave enough regulator headroom, and we might need to use the TL783 high-voltage regulator. This is a favourite device for generating +48 V supplies for microphone phantom power, and can definitely be relied on up to this voltage.

Two-transistors: gain distribution

At this point I began wondering what else could be done to reduce the distortion. There are practical limits to raising the supply voltage: power dissipation increases and there is

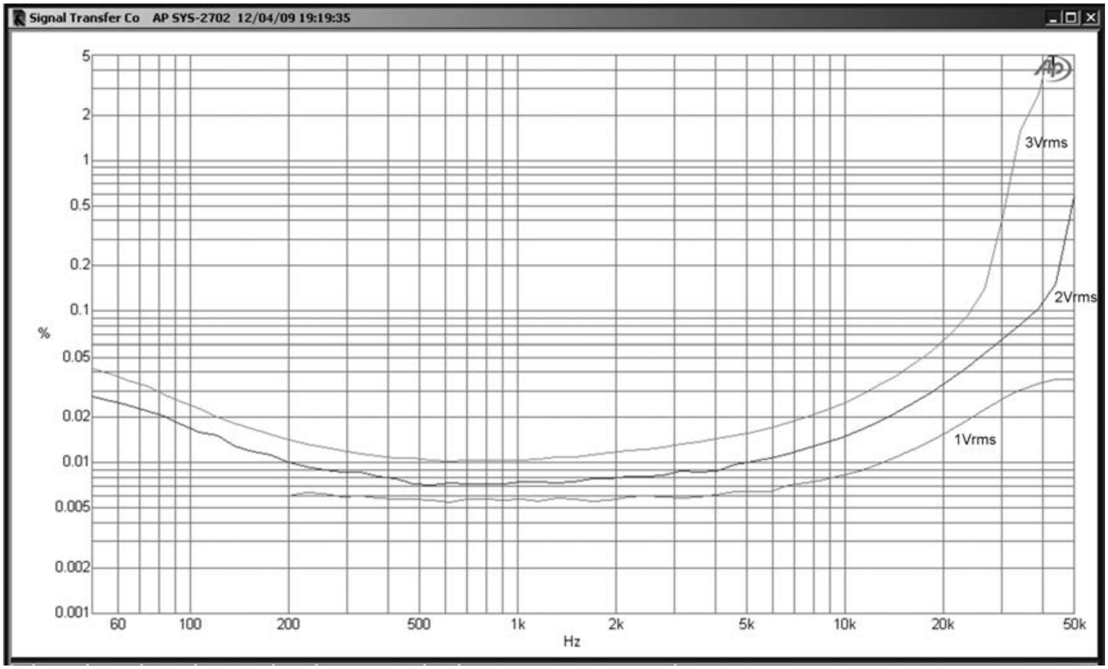

Figure 10.9: THD with a +30 V rail, at 1, 2, and 3 Vrms out, rebiased again. Bandwidth 100 Hz–80 kHz for 3 and 2 Vrms, 400 Hz–80 kHz for 1 Vrms

a danger that the circuit could generate turn-on or turn-off transients that would damage stages downstream.

At this point it's worth considering what the sources of non-linearity are. The two transistors, obviously, but the RIAA capacitors could also be contributing, as I used ordinary polyester types, and if you've read the chapter on components, you will know that these are not wholly linear, and in this application there is a significant signal voltage across them. However the distortion generated by polyester caps is typically of the order of 0.001% at 10 Vrms, and the signal levels we are using here are much lower than that, so the capacitor contribution is almost certainly negligible. At the time of writing I haven't got round to proving the point by substituting polypropylene capacitors, which *are* linear.

The configuration is made up of two cascaded voltage amplifiers, and it seemed to be a good idea to find out how the open-loop gain is distributed between them. The high value of the Q1 collector load suggests that it is intended to give a high voltage gain.

Measurement showed that the signal on Q1 collector was −49 dB with reference to that on the output at Q2 collector, at 1 kHz. This was confirmed by SPICE simulation, which gave −45 dB on Q1 collector between 100 Hz and 10 kHz. Note however that the emitter of Q2 is connected to AC ground via C3, which suggests that the second stage has a low input impedance and perhaps the first stage is working as a transconductance stage, feeding a current into the base of Q1 rather than a voltage, if you see what I mean. SPICE gives the error voltage, i.e. that between the base and emitter of Q1, as 38 dB below the output voltage, so the voltage on Q1 collector is less than that going into the first stage, and this indicates that Q1 is indeed feeding a current to Q2. This is an important finding as it means that the open-loop gain, and hence the feedback factor, cannot be increased by bootstrapping the collector load of Q1, which was the idea I had at the back of my mind all along.

The low-impedance at Q2 base will be further reduced by Miller feedback through the Cbc of Q2, though how significant that is uncertain at present. Since Cbc is a function of collector voltage this is another potential source of non-linearity.

Two-transistors: dual supply rails

The question arises as to how easy it would be to convert this stage to run off dual supply rails, i.e. ±V and 0 V. The answer appears to be not easy at all, because the input transistor Q1 that performs the input-NFB subtraction is sitting very near the bottom rail.

Two-transistors: the historical Dinsdale MM circuit

The first two-transistor MM stage is generally accepted to have been put forward by J. Dinsdale, in an article ‘Transistor High-Quality Audio Amplifier’, in *Wireless World* for January 1965. This article may be 45 years old but it is still worth reading if you can get hold of it, not least because it discusses how to make an MM preamplifier using just *one* transistor (see the start of this chapter).

You will note at once from Figure 10.10 that the circuit is upside-down to modern eyes, with a negative supply rail at the top. This was common in circuits of the era, and stemmed from the fact that most germanium transistors were PNP, so if you drew the emitter at the bottom (which is where people were used to drawing valve cathodes) then inevitably you end up with a negative supply rail at the top. When silicon transistors came in, they were more commonly NPN, so a sigh of relief went up all round as we reverted to the more logical approach of having the most positive rail at the top.

I have not so far tried building this circuit, due to the difficulty of obtaining the transistors. The OC44 was a PNP germanium transistor made by Mullard. Remarkably, it is still in much demand as it is held to give a unique sound in vintage-style fuzz boxes [3].

Comparing this circuit with Figure 10.2 you can see that there are the characteristic two separate feedback loops, with DC feedback through R13, R14 and R7, and AC feedback to Q1 emitter via the RIAA network R15, C8, R16 and C7. The DC path through this network

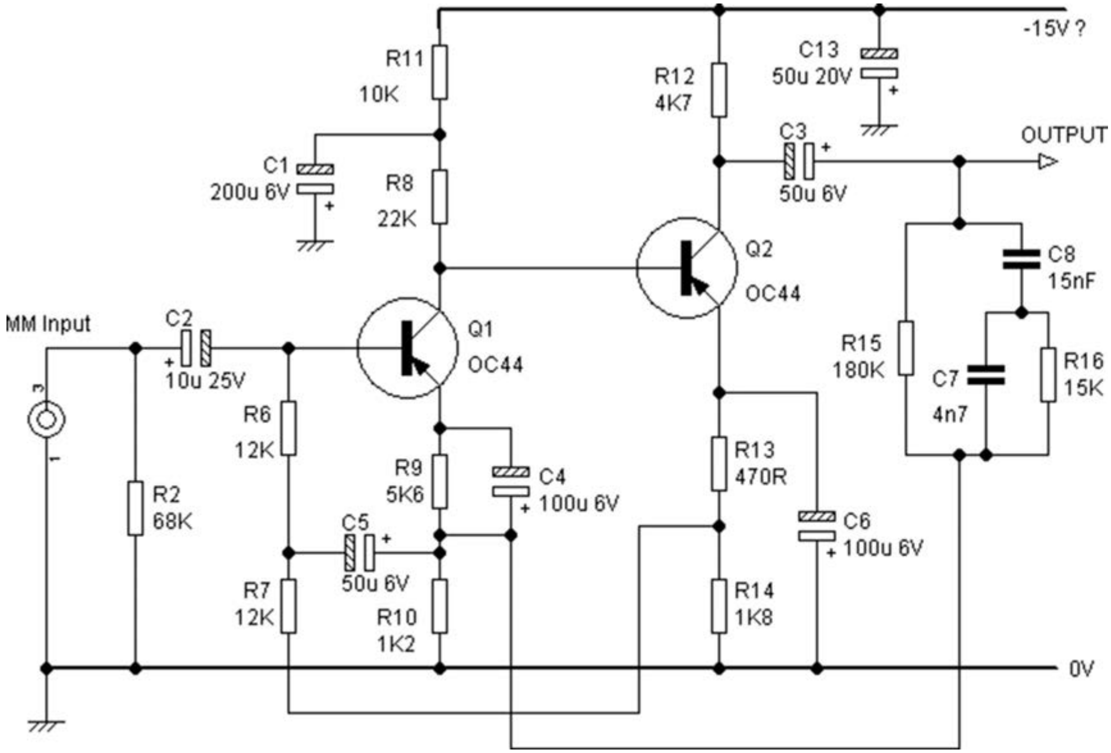

Figure 10.10: The circuit of the original Dinsdale MM stage: 1965

is blocked by C3. The RIAA network is in Configuration-B (see Chapter 8). There is no IEC Amendment.

C3 bootstraps R6 to raise the input impedance high enough to give 47 kΩ in conjunction with loading resistor R2; disconcertingly Dinsdale refers to this as 'feedback' in his article, which it is not.

The supply rail voltage is not precisely known, as in the complete preamplifier circuit the MM stage is fed through a network of RC filters that leave the final voltage in some doubt. It clearly is not greater than −20 V, judging by the rating of C13, and it seems pretty safe to assume it was around −15 V. However the OC44 had a collector breakdown voltage of only 15 V, so the rail might have been lower – perhaps −12 V.

This circuit uses more parts than the two-transistor circuit of Figure 10.2, largely as a result of the different DC bias arrangements. Apart from the transistors, it has 12 resistors and five electrolytic capacitors. Figure 10.2 has (ignoring EMC filtering) 10 resistors and two electrolytic capacitors, and is the more economical solution.

Three-transistor MM input stages

Adding an extra transistor improves the possible performance remarkably. The most common three-transistor configuration was introduced by Arthur Bailey in 1966 [4], though his version had a rather awkward level-shift between the first and second transistors. Figure 10.11 shows a much improved version from the early 70s, designed by H. P. Walker [5], and later enhanced by my own good self [6]. It consists of two voltage-amplifier stages as before, but an emitter-follower Q3 is added to buffer the collector of the second transistor from the load of the RIAA network and any external load. This means that the second transistor collector can be operated at a much higher impedance, generating more open-loop gain.

The original Walker design had a simple 22 kΩ resistor as a collector load for Q2; when I was using this configuration in [6] I split this into two 12 kΩ resistors, bootstrapping their central point from the emitter of Q3 as shown in Figure 10.11; this further increased the open-loop gain and reduced the stage distortion by a factor of three. Dominant-pole compensation is applied to Q2 by C1. Once again, the RIAA network has a high impedance and a separate path for DC feedback must be provided by R1; there is in fact no DC feedback at all through the RIAA network as it is connected to the outside of C9. R3 and C8 are the input RF filter. The supply to the first stage is heavily filtered by R8 and C5.

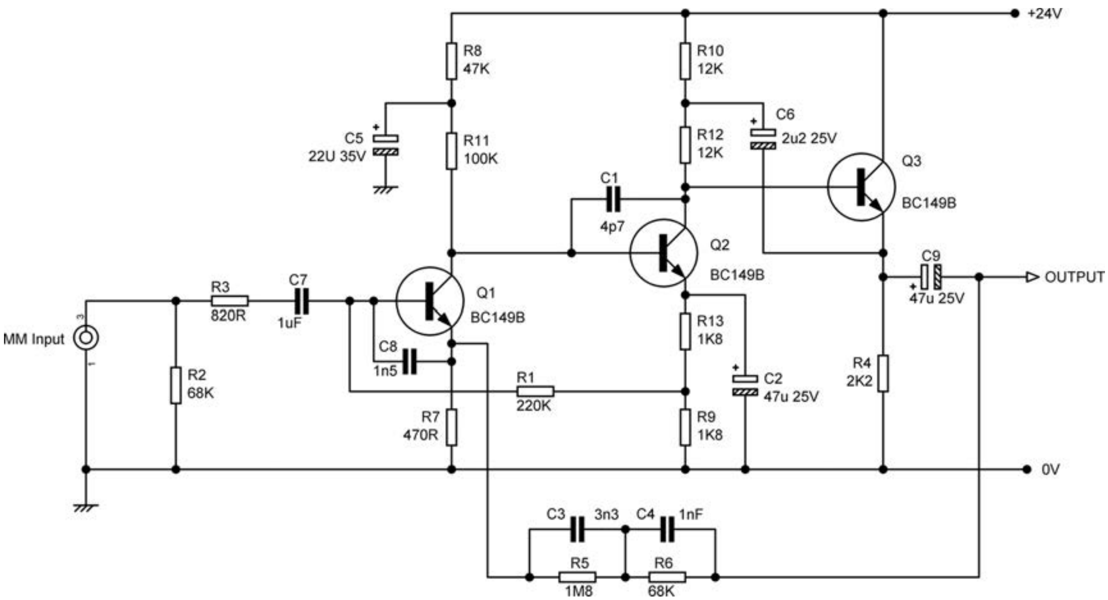

Figure 10.11: A typical three-transistor MM amplifier as commonly used in the 1960s and early 70s. The original design was by H. P. Walker, the bootstrapping of Q2 collector was added by me

The RIAA network is in Configuration-A. There is no IEC Amendment, as it was not introduced until 1976.

The added emitter follower Q3, running at a much higher collector current than Q2 (5 mA versus 500 µA) much increases the output voltage swing into a load and so improves the input overload margin. However, a simple emitter follower output stage has an asymmetrical output current capability, and so this is less effective at high frequencies where the impedance of the RIAA network is falling. Low-gain versions of this circuit may have the overload margin compromised by several dB at 20 kHz. This can be overcome by making the output stage more sophisticated – replacing the emitter resistor R4 with a current-source greatly improves matters, and using a push-pull Class-A output doubles the output current capability again (see Chapter 3 on discrete design for more details).

While this stage is a great improvement on the two-transistor configuration, it also is not well-adapted to dual supply rails, for the same reason; Q1 is still referenced to the bottom rail.

Figure 10.12 demonstrates another way to use three transistors in an RIAA amplifier; this configuration consists of a voltage amplifier stage, an emitter-follower, and then another

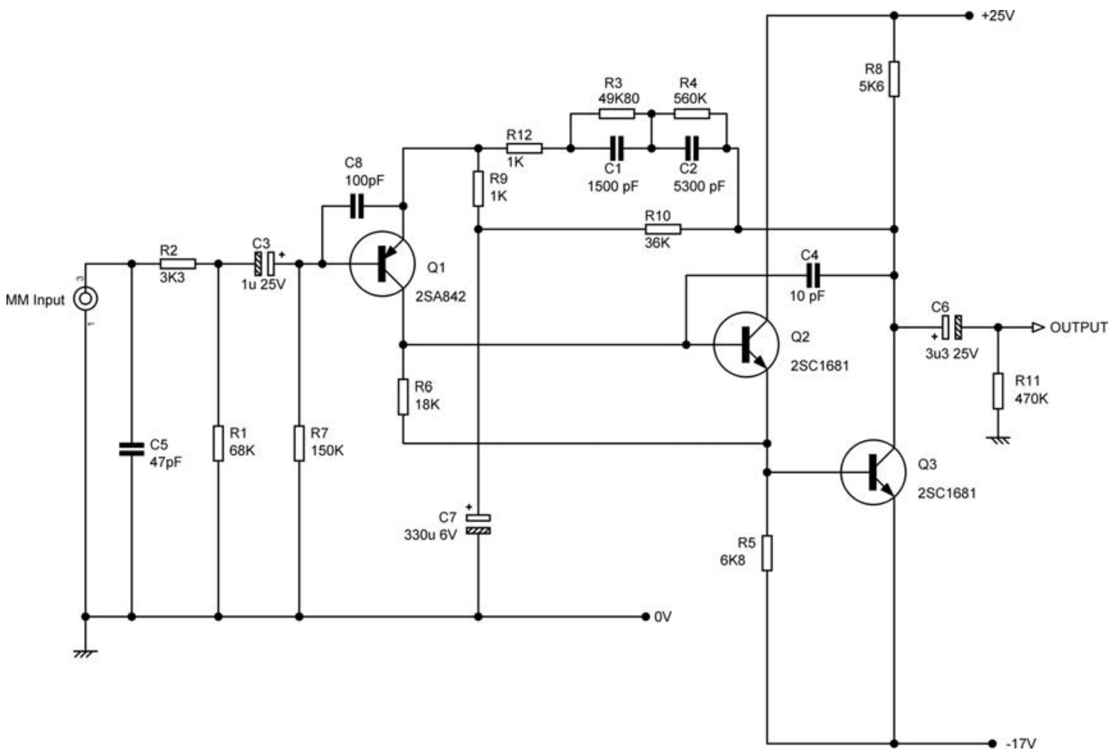

Figure 10.12: Another three-transistor MM amplifier configuration with a quite different structure. This is a simplified version of a circuit used by Pioneer and Sonab

voltage amplifier stage. The RIAA network is in Configuration-A. The first transistor is now a PNP type, its collector bootstrapped for increased gain by connecting the lower end of collector load R6 to the emitter-follower Q2. R2 and C8 make up an RF filter, and once more there is a high value of series resistance. Q2 drives the output transistor Q3, which is now a common-emitter voltage amplifier with collector load R8. Note the asymmetrical supply voltages; the positive rail is 8 V greater than the negative rail, and this is almost certainly intended to increase the positive-swing capabilities of the Q3 stage. Once again note the high impedances in the RIAA network to reduce the loading on the output, and the consequent need for a separate DC feedback path via R10, the AC content being filtered out by C7. Versions of this configuration were used by Pioneer and Sonab.

Four-transistor MM input stages

All of the amplifiers described so far have conventional output stages and so have trouble driving their RIAA network at high frequencies. Going from three transistors to four gives much greater freedom of design, and allows us to use a markedly more capable output stage. The design in Figure 10.13 is closely based on the MM input stage of my High-Performance

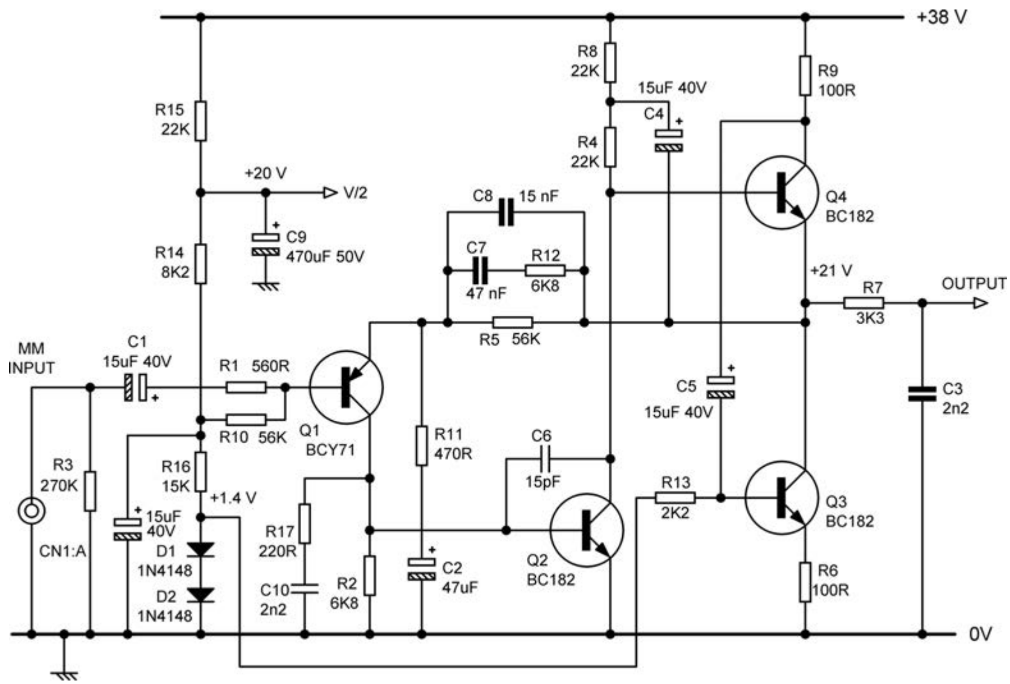

Figure 10.13: Four-transistor MM amplifier configuration with push-pull Class-A structure. Based on the MM stage of my 'High-Performance Preamplifier' published in *Wireless World* in 1979

Preamplifier published in *Wireless World* in 1979 [7]. By my own internal labelling system this was the MRP4. A slightly modified version (the MRP6) was used a year later in a preamplifier design for a consultancy client.

The MRP4 was the first 'conventional' preamplifier design – insofar as my designs are ever conventional – that I published. It was my own reaction to the considerable complexity of the Advanced Preamplifier of 1976 [6], with its multiple discrete op-amps. Back then the available IC opamps were looked at with entirely justified suspicion; they were relatively noisy and prone to crossover distortion in their output stages. Crossover might be inescapable in a power amplifier, but it was definitely not wanted in a preamplifier. Thus discrete Class-A circuitry was used throughout the preamplifier (the 5534 opamp was just becoming available at the time, but was horribly expensive).

This design was intended to display a very large overload margin, and so its gain at 1 kHz was only 10 times (+20 dB) and the output was only 50 mVrms nominal. Obviously, further downstream amplification (of variable gain) was required to get the signal up to a level that could be applied to a power amplifier. The configuration is much the same as many power amplifiers of the day, and if the singleton input transistor was replaced by a differential pair, it would look very much like the 'model amplifiers' used so extensively in my book *Audio Power Amplifier Design* [8]. I used a single-transistor input because I was worried about the noise contribution from the second transistor in a differential pair. That this might create a relatively large amount of second-harmonic distortion was considered of less importance, given the 50 mVrms nominal output of the stage. The RIAA network is an example of Configuration-C, which is preferred because of its lower capacitor values for the same gain and impedance (as described in Chapter 8). It has capacitor values similar to those used in the +30 dB gain opamp MM inputs in that chapter, but R0 (which in this case is actually labelled R11) is larger so the gain is 10 dB less. The stage could accept an input of 1.1 Vrms at 1 kHz, an overload margin of no less than 47 dB, and 3.8 Vrms at 10 kHz. The low gain made an HF correction pole (R7, C3) essential. No attempt was made to implement the IEC Amendment.

The whole preamplifier ran off a single +38 V rail that was not regulated; instead it had a post-reservoir RC filter that reduced the supply ripple to about 50 mVrms. This low-cost approach was combined with heavy RC decoupling of the bias network for each stage, and this was very effective at getting low hum figures.

The biasing network R14, R15, R16, D1, D2 provides three voltages. The +20 V 'V/2' bias rail was used throughout the rest of the preamplifier; it is heavily filtered by the large value of C9. The bias voltage at the top of R16 was lower to allow for the voltage drop of Q1 collector current through R5. This voltage-shift is an inherent problem with singleton inputs. D1, D2 provide the bias for the push-pull output stage, and were shared between the left and right inputs without any crosstalk problems.

The input impedance is defined by resistors R10 +R1, in parallel with DC drain R3; this comes to 46.8 kΩ. You will note that all the resistors in the stage (including those in the RIAA network) are E12, because E24 value resistors were specialised and expensive parts back in those days, and rarely if ever used in audio work.

The collector current of Q1 is set by the V_{be} of Q2 maintained across R2, and is here 53 μA. The signal is passed as a current from Q1 to the VAS Q2, on whose collector the full output swing is developed. Q2 collector load was a bootstrapped resistor rather than a current source as this still gave an economic advantage at the time. This is then buffered by the push-pull emitter-follower Q4 with its driven current source Q3. The operation of the push-pull output structure is fully described in Chapter 3 on discrete design.

Stability is an issue here because the amplifier is working with a closed-loop gain close to unity at HF. The Miller dominant-pole capacitor, C6, is made as small as possible to maximise the slew rate, and stability is assisted by the lead-lag network R17, C10 across Q1 collector resistor. This stage was followed by a third-order subsonic filter using a current-source emitter-follower, and the only THD figure I have to hand is for both stages together. THD was below 0.004% at 6 Vrms out, from 1 kHz to 10 kHz, the input signal being inverse-RIAA equalised to give constant output with frequency.

This design could be relatively easily converted to run off dual supply rails, as the first transistor is not referenced to the bottom rail. However, the voltage drop through R5 is a problem, for if Q1 is biased from 0 V, the output standing voltage will be several volts positive.

More complex discrete-transistor MM input stages

The record for the highest supply voltage to an RIAA stage was set in 1974 by the Technics SU9600, which employed ±24 V rails and a third rail at a staggering +136 V. This gives a whole new meaning to the phrase ‘third-rail electrification’. To the best of my knowledge the record still stands. The general configuration is shown in Figure 10.14; seven transistors are used, in three cascaded differential voltage amplifiers, followed by an emitter-follower output buffer. The final voltage amplifier and the output stage work on asymmetrical supplies, running between +136 V and −24 V. The output sits at +56 V to allow a symmetrical output swing, which accounts for the DC blocking capacitor C11. Note that several resistors around the output stage are high-wattage types. RV1 allowed gain adjustment, while RV2 and RV3 were for setting the DC conditions. My information is that the maximum input was 900 mV (frequency unstated, but presumably at 1 kHz), the THD was 0.08% (frequency and level unstated), the S/N ratio was 73 dB with reference to 2 mV, and the RIAA accuracy was ± 0.3 dB.

The output device dissipation is of course enormous for a preamp stage, and the use of a constant-current source, or better still a push-pull Class-A output stage, would have allowed

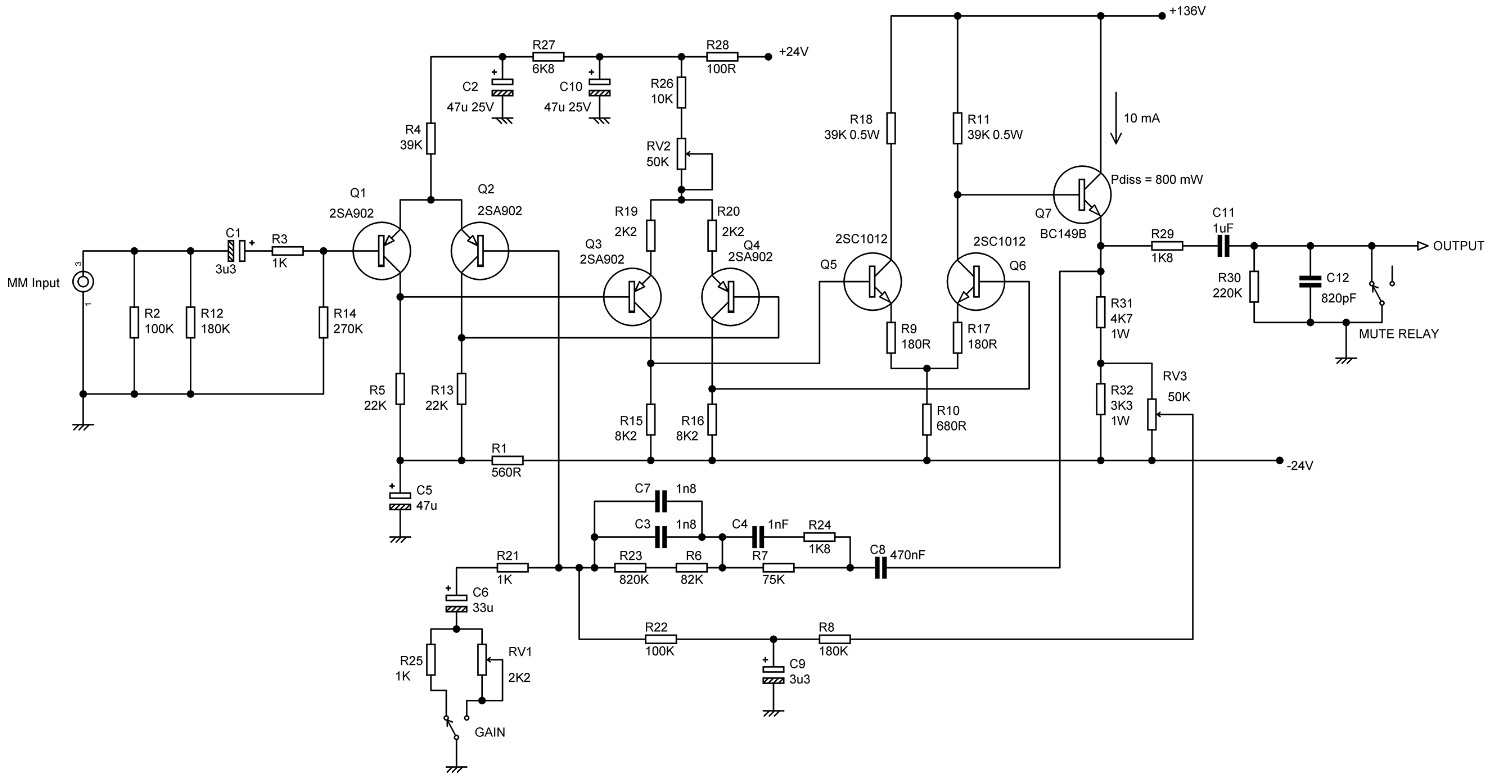

Figure 10.14: A simplified schematic of the Technics SU9600 RIAA stage, with its +136 V supply rail

this to be much reduced; one can only speculate as to why those techniques were not used. There would have been some fearsome transients at the output on switch-on, and it is notable that an output muting relay was required, probably not so much for reducing audible noise as to give the later stages in the preamplifier a chance of survival.

In the original circuit small capacitors were freely sprinkled over the diagram, leading me to suspect that HF stability was a serious issue during development.

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CHAPTER 11

Moving-magnet inputs: noise and distortion

Noise in MM RIAA preamplifiers

The subject of noise in moving-magnet (MM) RIAA preamplifiers is an involved business. An MM cartridge is a combination of resistance and a significant amount of inductance, with neither parameter having standard values, and this is combined with the complications of RIAA equalisation [1]. Burkhard Vogel has written a monumental 740-page book solely on RIAA amp noise [2], but even this does not exhaust the subject.

The first priority is to find out the physical limits that set how low the noise can be. The best possible equivalent input noise (EIN) for a purely resistive source, such as a 200 Ω microphone, is easily calculated to be −129.6 dBu with a noiseless amplifier at the usual temperature and bandwidth, but the same calculation for a moving-magnet input is much harder. Real amplifiers have their own noise, and the amount by which the source-amplifier combination is noisier than the source alone is the noise figure (NF) and we want to get this as low as possible. Noise figures are rarely, if ever, used in audio specifications, probably because they are very revealing; an NF of 20 dB usually indicates that someone doesn't know what they're about. Manufacturers seem to have no interest at all in quoting MM noise specs in a way that would allow easy comparison.

Most of the complications in calculating theoretical noise occur only when an MM cartridge is driving an RIAA preamp directly. When an MC cartridge is in use, the RIAA stage will be driven from an MC headamp, the output impedance of which should be very low, and this makes the noise situation much simpler (see later in this chapter).

The electronic noise will be much lower than the noise generated by the stylus sliding along a groove; hard data on this is in short supply but it is likely that the groove noise (sometimes called surface noise) will be 20 dB or more above the electronic noise [3]. Groove noise increases as the record wears with playing.

This presents a philosophical conundrum: is it not a waste of time to strive for low electronic noise when the groove noise is much greater and, since they will sum with rms-addition, the contribution of the electronic noise is negligible? If obtaining a good electronic noise performance was difficult and expensive this argument would have more force, but it is simply not so. This chapter will show how to get within a couple of dB of the lowest noise physically possible using cheap opamps and a little ingenuity. We will therefore do that, and sit back and see if record enthusiasts can improve their vinyl formulations to match. I think it will be a long wait.

A-weighting is not used in this chapter (or any other) except where explicitly stated.

Cartridge impedances

The impedance of the cartridge strongly influences the noise performance of an MM RIAA stage. Manufacturers do not always supply this data, and so I have had to make the best of what is available. Some of the cartridges listed in Table 11.1 are vintage while some are

TABLE 11.1 Some moving-magnet cartridge impedances, both current and historical

Type	Resistance (Ω)	Inductance (mH)
Audio-Technica AT440	Not stated	490
Audio-Technica AT15SS	500	720
CS1 ‘Carl Cox’	430	400
Glanz MFG-31E	900	110
Goldring 1006	660	570
Goldring 1042	660	570
Goldring 2044	Not stated	720
Goldring 2100	550	550
Goldring 2200, 2300	550	680
Goldring 2400, 2500	550	720
Grado Prestige Green 1, Black 1	475	45
Grado Prestige Red 1, Blue 1	475	45
Grado Reference Sonata 1, Platinum 1	475	45
Ortofon 2M Red, Blue	Not stated	700
Ortofon 2M Bronze, Black	Not stated	630
Shure ME75-ED Type 2	610	470
Shure ME95-ED	1500	650
Shure V15V MR	815	330
Shure V15V IV	1380	500
Shure V15V III	1350	500
Shure M44G	650	650
Stanton 5000 AL-II	535	400

up-to-date, the collection covering from about 1972 to 2013. Resistance ranges from 430 Ω to 1500 Ω, and inductance generally from 330 to 720 mH, apart from the Grado series which are more moving-iron than MM in operation, but given their 5 mV per cm/s output they are going to be used with an MM input. Moving-iron pickups go back a long way; see an article by Francis in *Wireless World* for 1947 [4], where a 1:100 step-up transformer was used to get the signal up to a suitable level to apply to a valve preamplifier. The Shure V15V values have been confirmed by Burkhard Vogel [5].

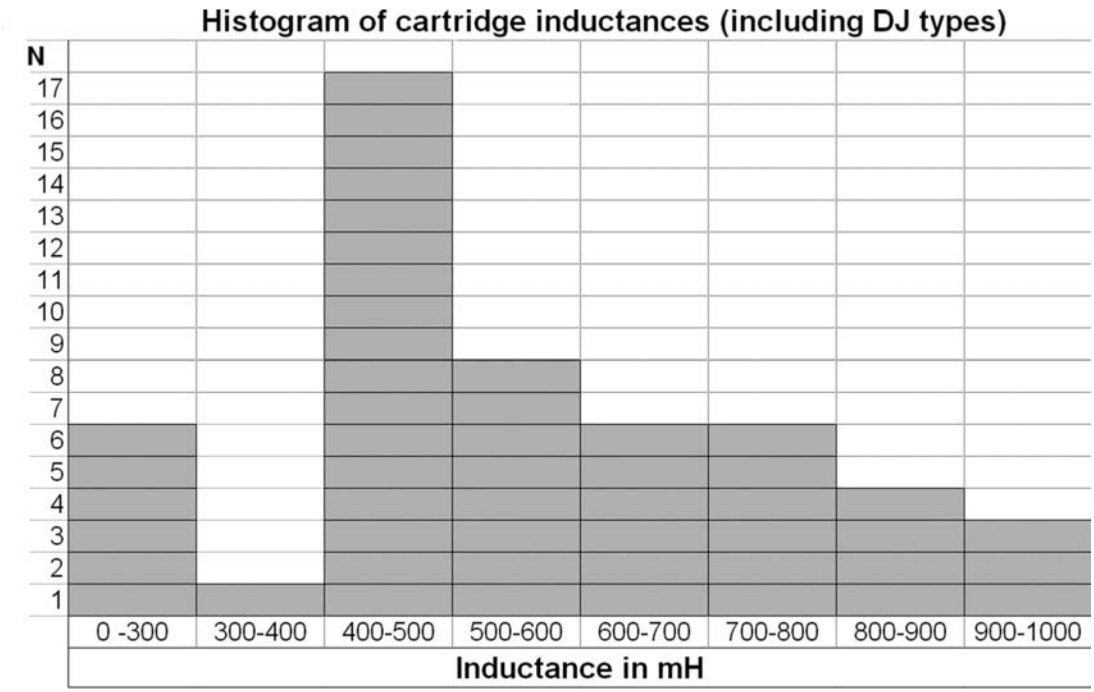

Figure 11.1: MM cartridge inductance, including DJ cartridges

The inductance of the cartridge is an important element in determining the noise performance, as will shortly be made clear. Figure 11.1 shows the inductance of 51 cartridges, covering both historical and contemporary models. The six types in the '0−300 mH' column are the Grado series. The seven types in the rightmost two columns are DJ types with higher output and inductance than normal hifi cartridges (see Chapter 8).

Noise modelling of RIAA preamplifiers

The basic noise situation for a series-feedback RIAA stage using an opamp is shown in Figure 11.2. The cartridge is modelled as a resistance R_{gen} in series with a significant inductance L_{gen}, and is loaded by the standard 47 kΩ resistor R_{in}; this innocent-looking

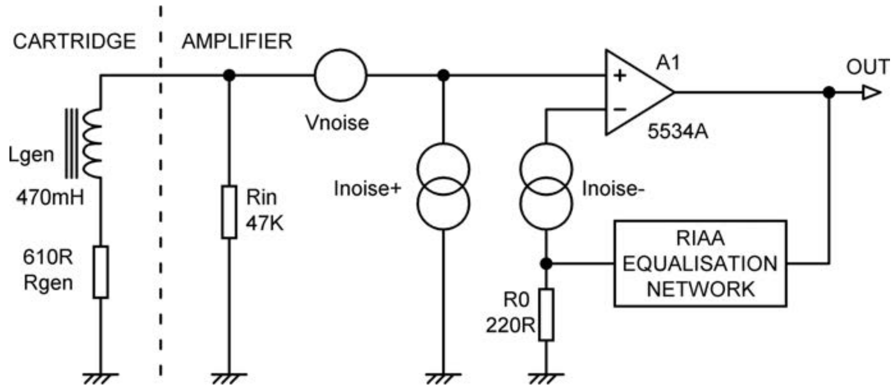

Figure 11.2: A moving-magnet input simplified for noise calculations, with typical cartridge parameter values (Shure ME75-ED2)

component causes more mischief than you might think. The amplifier A1 is treated as noiseless, its voltage noise being represented by the voltage generator V_{noise}, and the current noise of each input being represented by the current generators $I_{noise}+$ and $I_{noise}-$, which are uncorrelated. It does not matter to which side of V_{noise} $I_{noise}+$ is connected because V_{noise} has no internal resistance.

The contributions to the noise at the input of A1 are:

1. The Johnson noise of the cartridge resistance R_{gen}. This sets the ultimate limit to the signal/noise ratio. The proportion of noise from R_{gen} that reaches the amplifier input falls with frequency as the impedance of L_{gen} increases. Here the fraction reaching the amplifier falls from 0.99 to 0.48 from 36 Hz to 17.4 kHz. A complication that is not visible in the diagram is that the effective value of R_{gen} is not simply the resistance of the coils. It increases in value with frequency (while still remaining resistive − we are not talking about inductance here) as a consequence of hysteresis and eddy current magnetic losses in the iron on which the coils are wound, and possibly skin effect [6]. These losses are sometimes modelled by a frequency-dependent resistance placed across L_{gen}, as by Hallgren [7], or by a fixed resistance across part of a tapped L_{gen} [8]. According to Gevel [9] the losses have little effect on noise issues, and they are not modelled here, not least because of a sad lack of data.
2. The Johnson noise of the 47 kΩ input load R_{in}. Some of the Johnson noise generated by R_{in} is shunted away from the amplifier input by the cartridge, the amount decreasing with frequency due to the inductance L_{gen}. Here the fraction reaching the amplifier rises from 0.013 to 0.52 from 36 Hz to 17.4 kHz.
3. The opamp voltage noise V_{noise}. This contribution is unaffected by other components.

4. The noise voltage generated by I_{noise}+ flowing through the parallel combination of the cartridge impedance and R_{in}. This impedance increases with frequency due to L_{gen}. Here it increases from 619 Ω at 36 Hz to 24.5 kΩ at 17.4 kHz; the increase at the top end is moderated by the shunting effect of R_{in}. This increase has a major effect on the noise behaviour. For the lowest noise you must design for a higher impedance than you might think, and Gevel [9] quotes 12 kΩ as a suitable value for noise optimisation; this assumes A-weighting, inclusion of the IEC amendment, and cartridge parameters of 1000 Ω and 494 mH.
5. The Johnson noise of R0. For the values shown, and with A1 assumed to be 5534A, ignoring the Johnson noise of R0 improves the noise performance by only 0.35 dB. The other resistors in the RIAA feedback network are ignored, as R0 has a much lower value, but the RIAA frequency response must of course be modelled.
6. The noise voltage generated by I_{noise}- flowing through R0. For normal values of R0, say up to 1000 Ω, this contribution is negligible, affecting the total noise output by less than 0.01 dB.

Contributions 1, 2 and 4 are significantly affected by the rising impedance of the cartridge inductance L_{gen} with frequency. On top of this complicated frequency-dependent behaviour is overlaid the effect of the RIAA equalisation. This would reduce the level of white noise by 4.2 dB, but we are not dealing with white noise − the HF part of the spectrum has been accentuated by the effects of L_{gen}, and with the cartridge parameters given RIAA equalisation actually reduces the noise amplitude by 10.4 dB.

Clearly this model has some quite complex behaviour. It could be analysed mathematically, using a package such as MathCAD, or it could be simulated by SPICE. The solution I chose is a spreadsheet mathematical model of the cartridge input. The basic method is described by Sherwin [10]. The audio spectrum is divided into a number of octave bands so RIAA equalisation factors can be applied, and V_{noise}, I_{noise} and R_{gen} can be varied with frequency if desired. I extended the Sherwin scheme by using ten octave bands covering 22 Hz to 22 kHz; ten bands are enough to make the process accurate. An advantage of the spreadsheet method is that it is very simple to turn off various noise contributions so you can experiment with noiseless amplifiers or other flights from physical reality. For example, the noise generated by the 47 kΩ resistor R_{in} is modelled separately from its loading effects so they can be switched off independently (see load synthesis later in this chapter). It is also possible to switch off the bottom four octave bands to make the results comparable with real cartridge measurements that require a steep 400 Hz high-pass filter to remove the hum. A-weighting can also be switched on and off. The RIAA IEC Amendment can be switched on and off too, but since it only has an effect on very low frequencies the effect on the noise is negligible. The results match well with my 5534, 5532 and TL072 measurements, and experience shows the model is a usable tool. While it is no substitute for careful measurements, it gives a good physical insight and allows noise comparisons at the LF end where hum is very difficult to exclude completely. I call this model MAGNOISE2; it is an enhancement of the earlier MAGNOISE with improved accuracy. For this reason there are some small differences between the figures quoted here and in the first edition of this book, but the principles and the conclusions drawn remain the same.

Table 11.2 shows some interesting cases: output noise, EIN, and signal-to-noise ratio for a 5 mVrms input at 1 kHz are calculated for gain of +30.0 dB at 1 kHz. The IEC Amendment is included. The cartridge parameters were set to 610 Ω + 470 mH, the measured values for the Shure M75ED 2. Bandwidth is 22 Hz−22 kHz, no A-weighting is used, and 1/*f* noise was not considered. Be aware that the 5534A is a low-noise version of the 5534, with a typical

TABLE 11.2 RIAA noise results from the MAGNOISE2 spreadsheet model under differing conditions, in order of quietness. Cases 0 to 3 assume a noiseless amplifier and are purely theoretical

Case	Amplifier type	en (nV/√Hz)	in (pA/√Hz)	R_{in} (Ω)	R_0 (Ω)	Noise output (dBu)	S/N ref. 5mV input (dB)	EIN (dBu)	NF ref. Case 2 (dB)	Ref. Case 6a (dB)
0	Noiseless amp, no R_{gen}	0	0	1000M	0	−136.8	−123.1	−166.8	−41.2	−44.3
1a	Noiseless amp	0	0	1000M	0	−98.8	−85.0	−128.8	−3.2	−6.3
1b	Noiseless amp	0	0	10M	0	−98.7	−84.9	−128.7	−3.1	−6.2
1c	Noiseless amp	0	0	1M	0	−98.2	−84.4	−128.2	−2.6	−5.7
2	Noiseless amp	0	0	47k	0	−95.6	−81.8	−125.6	0 ref	−3.1
3	Noiseless amp	0	0	47k	220	−94.9	−81.1	−124.9	0.7	−2.4
4	J310 FET, Id = 10 mA	2	0	47k	220	−94.2	−80.4	−124.2	1.4	−1.7
5a	2SB737 BJT, I_c = 70 μA	1.75	0.39	47k	220	−93.6	−79.8	−123.6	2.0	−1.1
5b	2SB737 BJT, I_c = 100 μA	1.47	0.46	47k	220	−93.4	−79.6	−123.4	2.2	−0.9
5c	2SB737 BJT, I_c = 200 μA	1.04	0.65	47k	220	−92.7	−78.9	−122.7	2.9	+0.2
6a	5534A	3.5	0.4	47k	220	−92.5	−78.7	−122.5	3.1	0 ref
6b	5534A	3.5	0.4	47k	470	−92.1	−78.3	−122.1	3.5	+0.4
6c	5534A	3.5	0.4	47k	1000	−91.4	−77.6	−121.4	4.2	+1.1
7	5532A	5	0.7	47k	220	−90.5	−76.5	−120.5	5.1	+2.0
8	OPA2134	8	0.003	47k	220	−89.3	−75.5	−119.3	6.3	+3.2
9	LM4562	2.7	1.6	47k	220	−87.9	−74.1	−117.9	7.7	+4.6
10	OP275	6	1.5	47k	220	−87.3	−73.5	−117.3	8.3	+5.2
11	TL072	18	0.01	47k	220	−83.4	−69.6	−113.4	12.2	+9.1
12	LM741	20	0.7 ?	47k	220	−82.4	−68.6	−112.4	13.2	+10.1

voltage noise density of 3.5 rather than 4 $nV\sqrt{Hz}$, and a typical current noise density of 0.4 rather than 0.6 $pA/\sqrt{Hz}$. There is also an A-version of the 5532, but curiously the data sheets show no noise advantage. The voltage noise and current noise densities used here are the manufacturer's 'typical' figures. I am not aware of any data on how much they vary around the quoted values in practice.

Firstly let us see how quiet the circuit of Figure 11.2 would be if we had miraculously noise-free electronics.

Case 0: We will begin with a completely theoretical situation with no amplifier noise, and an MM cartridge with no resistance R_{gen}. L_{gen} is 470 mH. R_{in} is set to 1000 MΩ; the significance of that will be seen shortly. The noise out is a subterranean and completely unrealistic −136.8 dBu, and that is *after* +30 dB of amplification. This noise comes wholly from R_{in}, and can be reduced without limit if R_{in} is increased without limit. Thus if R_{in} is set to 1000 GΩ the noise out is −166.8 dBu.

You may ask why the noise is going up as the resistance goes down, whereas it is usually the other way around. This is because of the high cartridge inductance, which means the Johnson noise of R_{in} acts as a current rather than a voltage, and this goes up as resistance goes down.

Case 1: We now switch on the Johnson noise from R_{gen} (610 Ω). We will continue to completely ignore the cartridge loading requirements and leave R_{in} at 1000 MΩ, at which value it now has no effect on noise. The output noise with these particular cartridge parameters is then −98.8 dBu (Case 1a). This is the quietest possible condition, (if you can come up with a noiseless amplifier) but you will note that right from the start the signal/noise ratio of 85 dB compares badly with the 96 dB of a CD, a situation that merits some thought. And there is, of course, no groove noise on CDs. All of this noise comes from R_{gen}, the resistive component of the cartridge impedance. The only way to improve on this would be to select a cartridge with a lower R_{gen} but the same sensitivity, or start pumping liquid nitrogen down the tone-arm (as an aside, if you *did* cool your cartridge with liquid nitrogen at −196 °C, the Johnson noise from R_{gen} would only be reduced by 5.8 dB, and if you are using a 5534A in the preamplifier, as in Case 6a below, the overall improvement would only be 0.75 dB. And, of course, the compliant materials would go solid and the cartridge wouldn't work at all. Hold the cryostats!).

With lower, but still high, values of R_{in} the noise increases; with R_{in} set to 10 MΩ (Case 1b) the EIN is −128.7 dBu, a bare 0.1 dB worse. With R_{in} set to 1 MΩ (Case 1c) the EIN is now −128.2 dBu, 0.8 dB worse than the best possible condition (Case 1a).

Case 2: It is however a fact of life that MM cartridges need to be properly loaded, and when we set R_{in} to its correct value of 47 kΩ things deteriorate sharply, the EIN rising by 3.2 dB (compared with Case 1a) to −125.6 dBu. That 47 kΩ resistor is not innocent at all. This case still assumes a noiseless amplifier, and appears to be the appropriate noise reference for design, so the noise figure is 0 dB (however, see the section on load synthesis later in this chapter, which shows how the effects of noise from R_{in} can be reduced by some non-obvious methods). Cases 1a,b and c, therefore, have negative noise figures but this has little meaning.

Case 3: We leave the amplifier noise switched off, but add in the Johnson noise from R0 and the effect of I_{noise} to see if the value of 220 Ω is appropriate. The noise only worsens by 0.7 dB, so it looks like R0 is not the first thing to worry about. Its contribution is included in all the cases that follow. The noise figure is now 0.7 dB.

We will now take a deep breath and switch on the amplifier noise.

Case 4: Here we use a single J310 FET, a device often recommended for this application [9]. With I_d set to 10 mA, the voltage noise is about 2 $nV\sqrt{Hz}$; the current noise is negligible, which is why it is overall slightly quieter than the 2SB737 despite having more voltage noise.

Case 5: In these cases a single discrete bipolar transistor is used as an input device, not a differential pair. This can give superior noise results to an opamp. The transistor may be part of a fully discrete RIAA stage, or the front end to an opamp. If we turn a blind eye to supply difficulties and use the remarkable 2SB737 transistor (with R_b only 2 Ω typical) then some interesting results are possible. We can decide the collector current of the device, so we can to some extent trade off voltage noise against current noise, as described in Chapter 1. We know that current noise is important with an MM input, and so we will start off with quite a low I_c of 200 μA, which gives Case 5c in Table 11.2. The result is very slightly worse than the 5534A (Case 6a). Undiscouraged, we drop I_c to 100 μA (Case 5b) and voltage noise increases but current noise decreases, the net result being that things are now 0.9 dB quieter than the 5534A. If we reduce I_c again to 70 μA (Case 5a) we gain another 0.2 dB, and we have an EIN of −123.6 and a noise figure of only 2.0 dB. Voltage noise is now increasing fast and there is virtually nothing to be gained by reducing the collector current further.

We therefore must conclude that even an exceptionally good single discrete BJT with appropriate support circuitry will only gain us a 1.1 dB noise advantage over the 5534A, while the J310 FET gives only a 1.7 dB advantage, and it is questionable if the extra complication is worth it. You are probably wondering why going from a single transistor to an opamp does not introduce a 3 dB noise penalty because the opamp has a differential input with two transistors. The answer is that the second opamp transistor is connected to the NFB network and sees much more favourable noise conditions; a low and resistive source impedance in the shape of R0.

Case 6: Here we have a 5534A as the amplifying element, and using the typical 1 kHz specs for the A-suffix part, we get an EIN of −122.5 dBu and an NF of 3.1 dB (Case 6a with R0 = 220 Ω). Using thoroughly standard technology, and one of the cheapest opamps about, we are within three decibels of perfection; the only downside is that the opportunities for showing off some virtuoso circuit design with discrete transistors appear limited. Case 6a is useful as a standard for comparison with other cases, as in the rightmost column of Table 11.2.

Firstly, how does the value of R0 affect noise? In Case 6b, R0 is increased to 470 Ω, and the noise is only 0.4 dB worse; if you can live with that, the increase in the impedance of the RIAA feedback network allows significant savings in expensive precision capacitors. In Case 6c, R0 is raised further to 1000 Ω, and noise is now 1.1 dB worse than the 220 Ω case. Reducing R0 from 220 Ω to 100 Ω is do-able at some cost in capacitors, but only reduces the noise output by 0.2 dBu. In Chapter 8, the value of R0 can be manipulated to get convenient capacitor values in the RIAA network, because it has only a weak effect on the noise performance.

Secondly, we have seen that the presence of L_{gen} has a big effect on the noise contributions. In Case 6a, if we reduce L_{gen} to zero the noise out drops from −92.5 to −94.7 dBu. Halving it gives −93.8 dBu. Minimum cartridge inductance is a good thing.

Thirdly, what about R_{gen}? With the original value of L_{gen}, setting R_{gen} to zero only reduces the noise from −92.5 to −93.5 dBu; the cartridge inductance has more effect than its resistance.

Case 7: It is well-known that the single 5534A has somewhat better noise specs than the dual 5532A, with both e_n and i_n being significantly lower, but does this translate into a significant noise advantage in the RIAA application? Case 7 shows that on plugging in a 5532A the noise output increases by 2.0 dB, the EIN increasing to −120.5 dBu. The NF is now 5.1 dB, which looks a bit less satisfactory. If you want good performance then the inconvenience of a single package and an external compensation capacitor are well worth putting up with. If your circuit design ends up with an odd number of half-5532s per channel, a 5534A can be placed in the MM stage, where its lower noise is best used.

Case 8: Ho-hum, I hear you murmur, it all comes back to the 5534A, doesn't it? What about all the other opamps on the market? In Case 8 we take the FET-input OPA2134, which is a good opamp when DC accuracy and low bias currents are required; we find the e_n is much higher at 8 nV$\sqrt{\text{Hz}}$, but i_n is very low indeed at 3 fA$\sqrt{\text{Hz}}$. It looks like we might be in with a chance, but the greater voltage noise does more harm than the lower current noise does good and the EIN goes up to −119.3 dBu. The OPA2134 is therefore 3.2 dB noisier than the 5534A and 2.5 dB noisier than the 5532A, and it is not cheap. The noise figure is now 6.3 dB, which to a practised eye would show that something had gone amiss in the design process.

Case 9: The LM4562 BJT-input opamp gives significant noise improvements over the 5534/5532 when used in low-impedance circuitry, because its e_n is lower at 2.7 nV$\sqrt{\text{Hz}}$. However, the impedances we are dealing with here are not low, and the i_n, at 1.6 pA$\sqrt{\text{Hz}}$, is four times that of the 5534A, leading us to think it will not do well here. We are sadly correct, with EIN deteriorating to −117.9 dBu and the noise figure an unimpressive 7.7 dB. The LM4562 is almost 5 dB noisier than the 5534A and at the time of writing is a lot more expensive. Measurements confirm a 5 dB disadvantage.

Case 10: The OP275 has both BJT and FET input devices. Regrettably this appears to give both high voltage noise and high current noise, resulting in a discouraging EIN of −117.3 dBu. It is expensive.

Case 11: The TL072 with its FET input has very high voltage noise at 18 $nV\sqrt{Hz}$ but low current noise. We can expect a poor performance. We duly get it, with EIN rising to −113.4 dBu and a very indifferent noise figure of 12.2 dB. The TL072 is 9.1 dB noisier than a 5534A, and 8.4 dB noisier than a 5532A. The latter figure is confirmed (within experimental error, anyway) by the data listed in the section below on noise measurements. There is now no reason to use a TL072 in an MM preamp; it must be one of the worst you could pick.

Case 12: Just for historical interest I tried out the LM741. The voltage noise measures about 20 $nV\sqrt{Hz}$. I have no figures for the current noise, but I think it's safe to assume it won't be better than a 5532, so I have used 0.7 $pA\sqrt{Hz}$. Predictably the noise is the highest yet, with an EIN of −112.4 dBu, but it is a matter for some thought that despite using an ancient part it is only 10 dB worse than the 5534A.

You may be wondering what has happened to other well-known opamps, particularly the OP-27 and the LT1028. Both are sometimes recommended for audio use, because of their low e_ns, but this ignores a serious problem. The OP-27 has a low e_n and i_n, and in fact gives a calculated EIN of −123.0 dB, which beats the 5532A noise, but when you measure it in real life it is actually several dB noisier; I have confirmed this several times. This is due to extra noise generated by bias-current cancellation circuitry. Correlated noise currents are fed into both inputs, and will only cancel if both inputs see the same impedance. In this MM-RIAA application the impedances are wildly different, and the result is much increased noise. This problem with the OP-27 was originally pointed out to me by Marcel van de Gevel [9].

The LT1028 gives a poor performance in MM applications because while it has an appealingly low e_n of 0.85 $nV\sqrt{Hz}$, its i_n is high at 1 $pA\sqrt{Hz}$ as a result of running big input BJTs at high collector currents. The EIN is by calculation −120.9 dBu, making it a shade quieter than the 5532A. But the LT1028 also has bias-current cancellation circuitry, and the data sheet explicitly states, 'the cancellation circuitry injects two correlated current noise components into the two inputs.' According to Gevel [9] the effective voltage noise in an MM application is about 39 $nV\sqrt{Hz}$, so even the 741 would be a better choice.

Noise and A-weighting

The frequency response of human hearing is not flat, especially at lower listening levels (see Chapter 13 for more on this). Some commentators therefore feel it is appropriate to use psychoacoustic weighting when studying noise levels. This is almost invariably ANSI A-weighting despite the fact that it is generally considered inaccurate, as it undervalues low frequencies. ANSI B-, C-, and D-weightings also exist but are not used in audio. The ITU-R 468 weighting (CCIR-468) is much better but only rarely used. I prefer unweighted measurements as you are one step closer to the original data.

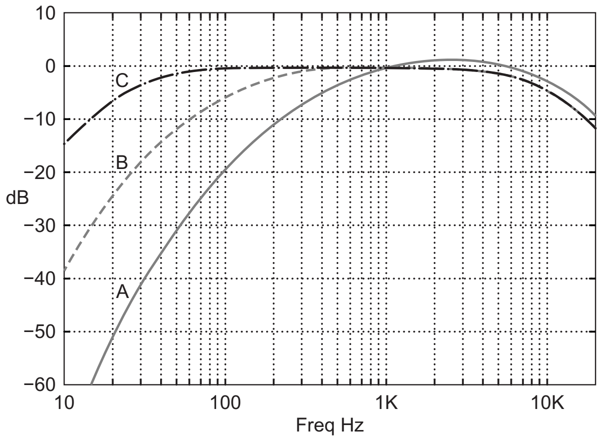

Figure 11.3: ANSI weighting curves A, B, and C. Only A-weighting is used in audio

TABLE 11.3 The effect of A-weighting on the calculated noise performances of various amplifiers

Case	Amplifier type	Unweighted noise out (dBu)	A-Weighted noise out (dBu)	A-Weighting difference (dB)
5a	2SB737 70 μA	−93.6	−96.3	2.7
5b	2SB737 100 μA	−93.4	−96.0	2.6
5c	2SB737 200 μA	−92.7	−95.1	2.4
6a	5534A	−92.5	−95.4	2.9
7	5532A	−90.5	−93.3	2.8
8	OPA2134	−89.3	−92.8	3.5
9	LM4562	−87.9	−89.9	2.0
11	TL072	−83.4	−87.1	3.7

The A-weighting curve is shown in Figure 11.3. It approximately follows the Fletcher-Munson 40-phon line. It passes through 0 dB at 1 kHz, and has a maximum gain of 1.3 dB at 2.5 kHz. The low-frequency roll-off steadily steepens as frequency falls, while the high-frequency roll-off is at 12 dB/octave (−3 dB at 12 kHz).

A-weighting reduces the level of white noise by 4.4 dB, but as noted earlier we are not dealing with white noise as the spectrum is altered by the effects of L_{gen} and the RIAA equalisation.

You may therefore be wondering how the unweighted results described above will be affected by the application of A-weighting. Will the order of merit in Table 11.2 be upset?

Table 11.3 shows the effects of A-weighting on selected cases from Table 11.2. The noise level drops by between 2 and 4 dB, depending on the magnitude of voltage noise compared to current noise. A-weighting does not introduce any revolutionary changes into the order of merit of the various amplifiers in Table 11.2.

RIAA noise measurements

In the past, many people who should have known better have recommended that MM input noise should be measured with a 1 kΩ load, presumably thinking that this emulates the resistance R_{gen}, which is the only parameter in the cartridge actually generating noise − the inductance is of course noiseless. This overlooks the massive effect that the inductance has in making the impedance seen at the preamp input rise very strongly with frequency, so that at higher frequencies most of the input noise actually comes from the 47 kΩ loading resistance. I am grateful to Marcel van de Gevel for drawing my attention to some of the deeper implications of this point [11].

The importance of using a real cartridge load is demonstrated in Table 11.4, where the noise performance of a TL072 and a 5532 are compared. The TL072 result is 0.8 dB too low, and 5532 result 4.9 dB too low − a hefty error. In general results with the 1 kΩ resistor will always be too low, by a variable amount. In this case you still get the right overall answer − i.e. you should use a 5532 for least noise − but the dB difference between the two has been exaggerated by almost a factor of two, by undervaluing the 5532 current noise.

The 1 kΩ recommendation was perhaps made because the obvious measurement method of loading the input with a MM cartridge has serious difficulties with hum from the ambient magnetic fields. To get useful results it is essential to enclose the cartridge completely in a grounded mu-metal can − I use one from a redundant microphone transformer and it works very well. I suppose the ideal load would be a toroidal inductor, but it would be an expensive custom part. It is also necessary to use complete electrostatic screening of the amplifier itself. If it has a 22 μF input coupling capacitor and the input is short-circuited, the impedance downstream of the capacitor is 145 Ω at 50 Hz, which is enough to make it susceptible to electrostatic hum pickup.

TABLE 11.4 Measured noise performance of 5532 and TL072 with two different source impedances

Zsource	TL072	5532 (dBu)	5532 benefit (dB)	5532 EIN (dBu)
1 kΩ resistor	−88.0	−97.2	+9.8	−126.7
Shure M75ED 2	−87.2	−92.3	+5.1	−121.8

These tests were done with an amplifier gain of +29.55 dB at 1 kHz. Bandwidth was 400−22 kHz to remove hum, rms sensing, no weighting, cartridge parameters were 610 Ω + 470 mH.

RIAA amps driven from MC head amp

All the discussion above deals with an RIAA preamplifier driven by an MM cartridge. As we noted at the start, the MM RIAA stage may also be driven from an MC head amp. The noise conditions for the RIAA amplifier are quite different, as it is now fed from a very low impedance, plus probably a series resistor in series with the MC amp output to give stability against stray capacitances. My current MC head amp design (see Chapter 12) has an EIN of −141.5 dBu with a 3.3 Ω input source resistance. Its gain is +30 dB, so the output noise is −111.5 dBu. The series output resistor is 47 Ω; its Johnson noise at −135 dBu is negligible, and likewise the effect of the RIAA stage input noise flowing in it. The noise at the output of the RIAA stage is then −85.7 dBu, which is higher than any of the figures in Table 11.2 except those for TL072 and LM741. In this situation the value of R0 is relatively unimportant.

Cartridge load synthesis for lower noise

Going back to Table 11.2 above, you will recall that when we were examining the situation with the amplifier and feedback network noise switched off, adding in the Johnson noise from the 47 kΩ loading resistor R_{in} caused the output noise to rise by 3.2 dB. In real conditions with amplifier noise included the effect is obviously less dramatic, but it is still significant. For the 5534A (Case 6a), the removal of the noise from R_{in} (but *not* the loading effect of R_{in}) reduces the noise output by 1.5 dB. Table 11.5 summarises the results for various amplifier options; the amplifier noise is unaffected, so the noisier the technology used, the less the improvement.

TABLE 11.5 The noise advantages gained by load synthesis with R_{in} = 1 MΩ

Case	Amplifier type	R_{in} = 47 KΩ EIN (dBu)	R_{in} = 47 KΩ NF (dB)	R_{in} = 1M EIN (dBu)	R_{in} = 1M NF (dB)	Advantage (dB)
5a	2SB737 70uA	−123.6	2.0	−125.3	0.3	1.7
5b	2SB737 100uA	−123.4	2.2	−125.1	0.5	1.7
5c	2SB737 200uA	−122.7	2.9	−124.1	1.5	1.4
6a	5534A	−122.5	3.1	−123.8	1.8	1.3
7	5532A	−120.5	5.1	−121.3	4.6	0.5
8	OPA2134	−119.3	6.3	−119.9	5.7	0.6
9	LM4562	−117.9	7.7	−118.3	7.3	0.4
11	TL072	−113.4	12.2	−113.5	12.1	0.1

From MAGNOISE2. NF is reference from Case 2 in Table 11.2

This may appear to be utterly academic, because the cartridge must be loaded with 47 kΩ to get the correct response. This is true, *but it does not have to be loaded with a physical 47 kΩ resistor*. An electronic circuit that has the V/I characteristics of a 47 kΩ resistor, but lower noise, will do the job very well. Such a circuit may seem like a tall order − it will, after all, be connected at the very input, where noise is critical, but unusually the task is not as difficult as it seems.

Figure 11.4a shows the basic principle. The 47 kΩ R_{in} is replaced with a 1 MΩ resistor whose bottom end is driven with a voltage that is phase-inverted and 20.27 times that at the top end. If we conceptually split the 1 MΩ resistor into two parts of 47 kΩ and 953 kΩ, a little light mathematics shows that with −20.27 times V_{in} at the output of A2, the voltage at the 47 kΩ−953 kΩ junction A is zero, and so as far as the cartridge is concerned it is looking at a 47 kΩ resistance to ground. However, the physical component is 1 MΩ, and the Johnson current noise it produces is less than that from a 47 kΩ (Johnson current noise is just the usual Johnson voltage noise applied through the resistance in question). The point here is that the apparent resistor value has increased by 21.27 times, but the Johnson noise has only increased by 4.61 times, because of the square-root in the Johnson equation; thus the current noise injected by R_{in} is also reduced by 4.61 times. The noise reduction gained with a 5534A (Case 6a) is 1.3 dB, which is very close to the 1.5 dB improvement obtained by switching off the R_{in} noise completely. If a resistor larger than 1 MΩ is used, slightly more noise reduction can be obtained, but that would need more gain in A2 and we would soon reach the point where it would clip before A1, restricting headroom. In this case we get a very good noise figure of 1.8 dB, though the lowest noise output comes from the 2SB737 at 70 μA.

The implementation made known by Gevel [9] is shown in Figure 11.4b. This ingenious circuit uses the current flowing through the feedback resistor R0 to drive a shunt feedback

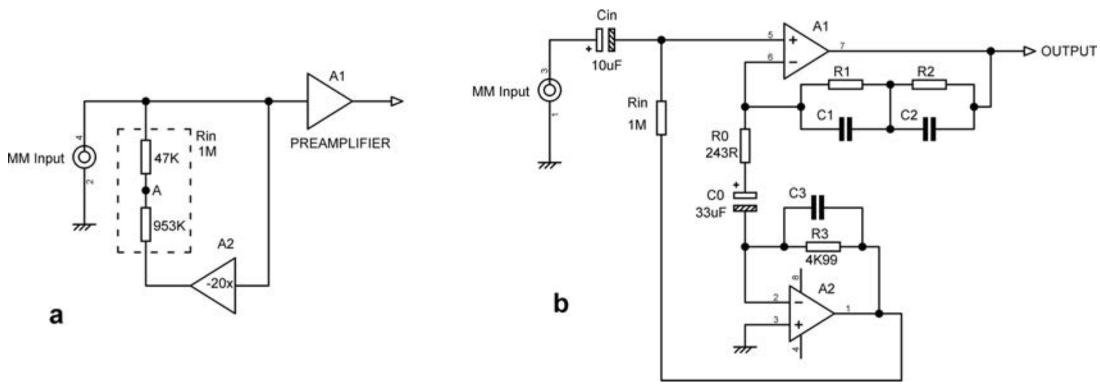

Figure 11.4: Electronic load synthesis: a) the basic principle, b) the Gevel circuit

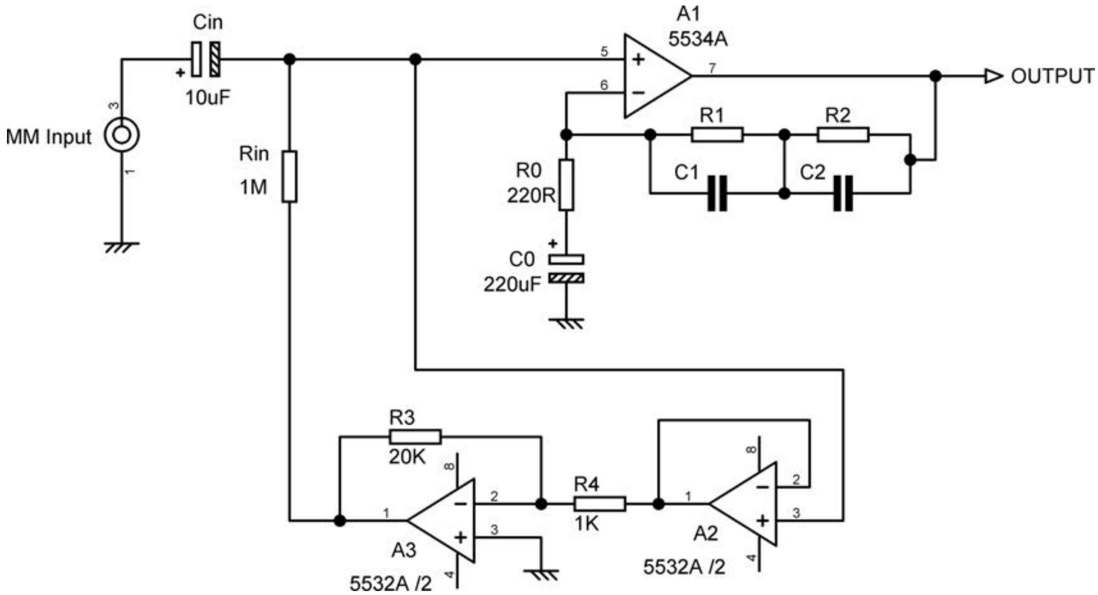

Figure 11.5: Electronic load synthesis: the Self circuit

stage around A2. With suitable scaling of R3 (note that here it has an E96 value) the output voltage of A2 is at the right level and correctly phase-inverted. When I first saw this circuit I had reservations about connecting R0 to a virtual ground rather than a real one, and thought that extra noise from A2 might find its way back up R0 into the main path (I hasten to add that these fears may be quite unjustified, and I have not found time so far to put them to a practical test). The inverting signal given by this circuit is amplified by 20.5 times rather than 20.27 but this has a negligible effect on the amount of noise-reduction.

Because of these reservations, I tried out my version of load-synthesis as shown in Figure 11.5. This uses the basic circuit of Figure 11.4a; it is important that the inverting stage A3 does not load the input with its 1 kΩ input resistor R4, so a unity-gain buffer A2 is added. The inverting signal is amplified by 20 times, not 20.27, but once again this has negligible effect on the noise-reduction.

In practical measurements with a 5534A as amplifier A1, I found that the noise improvement with a real cartridge load (Shure M75ED 2, cartridge parameters 610 Ω + 470 mH) was indeed 1.3 dB, just as predicted, which is as nice a matching of theory and reality as you are likely to encounter in this world. There were no HF stability problems. Whether the 1.3 dB is worth the extra electronics is a good question; I say it's worth having.

This technique has been called 'electronic cooling', presumably because it could be regarded as analogous to dipping the loading resistance in liquid nitrogen or whatever to reduce

Johnson noise. I must admit I don't like the term as it could be understood to mean that thermoelectric elements have been used to cool down the input stage, a technique I do not think has been used in hifi yet. I prefer to call it 'electronic loading', 'active input impedance' or 'load synthesis', the last being perhaps the most explicit. It would also be useful for tape head preamplifiers, but they are a bit of a minority interest these days.

The history of load synthesis

Yet again we encounter a technique that has a longer history than you might expect. It was referred to by Tomlinson Holman in his famous 1975 paper 'New Factors In Phonograph Preamplifier Design' [12] (which is still well worth reading) as a recent innovation:

'Two noise reduction techniques have appeared in recent designs. One is to use quite low impedances in the RIAA feedback. . . . The second involves the use of a synthesized input impedance through the use of an extra feedback loop which bootstraps the cartridge termination resistor to reduce its noise contribution. One commercial embodiment of the bootstrap method produced a signal-to-noise ratio of 85 dB re 10 mV, 1 kHz input, ANSI 'A' weighted with a cartridge input.' For comparison, assuming the reference level is 10 mVrms, the 5534A in Case 6a gives an *unweighted* S/N ratio of 84.7 dB without load synthesis, and 86.0 dB with it.

Bootstrapping is not really the right term; load synthesis is more 'anti-bootstrapping' in that it makes a large impedance look like a small one by applying a voltage in anti-phase, whereas conventional bootstrapping makes a small impedance look like a large one by applying a voltage in phase. I have no idea which 'commercial embodiment' he was talking about, and I would be glad to hear any suggestions.

The technique was analysed by Hoeffelman and Meys in the JAES in 1978 [13]. So far as I can find, there are no patents on this subject.

Distortion in MM RIAA amplifiers

An RIAA stage with a gain of +30 dB at 1 kHz will have a gain of about +50 dB at 20 Hz; that is high for a single opamp stage. The closed-loop gain at HF is about +10 dB so the feedback factor can be maintained there without problems; fortunately, the RIAA curve roughly follows the 6 dB/octave fall in open-loop gain. The CM voltage is very small and should not cause distortion.

Figure 11.6 shows the distortion at 5 V, 7 V, and 9 Vrms. The distortion at LF is mainly second harmonic. It was checked that this was not coming from C0; increasing it from 220 µF to 1000/6V3 gave no improvement. The sudden increase in distortion at about 27 kHz for the

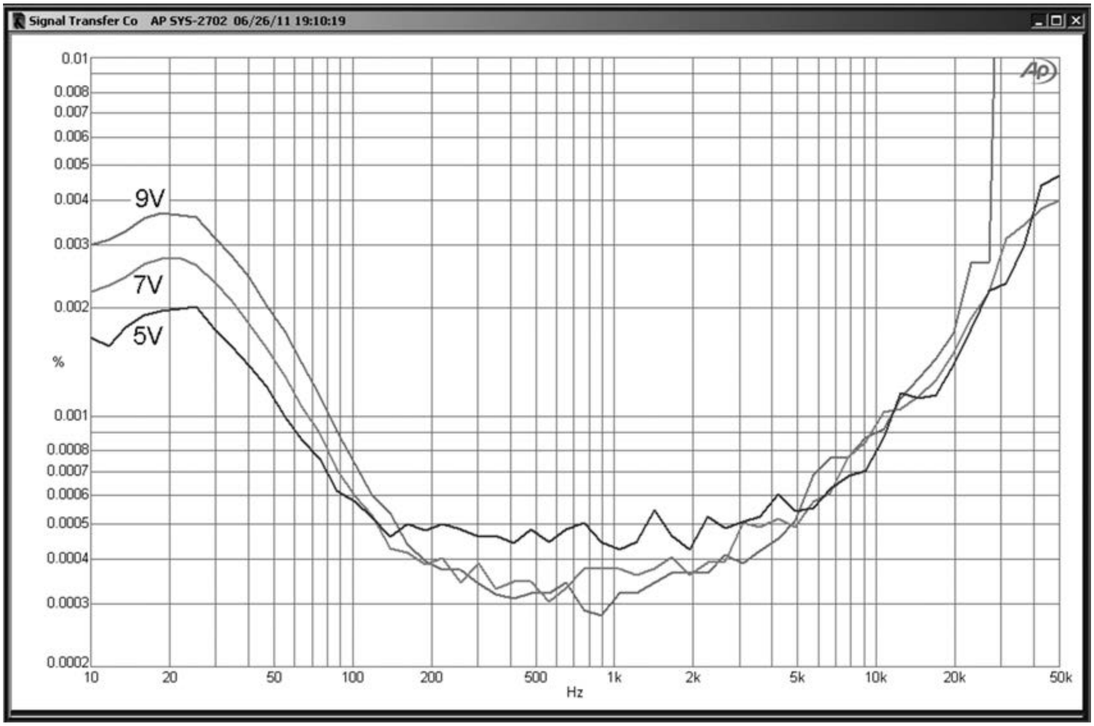

Figure 11.6: +30 dB (1 kHz) 5534A RIAA preamp THD at 5, 7, and 9 Vrms out. No IEC amendment, no external load except correction pole

9 Vrms case occurs when the current drawn by the RIAA network exceeds the capability of the 5534A. Note that 9 Vrms is a whole 36 dB above the nominal operating level of 150 mVrms.

The distortion will be aggravated by external loading. Here we have just the HF correction pole, which only places an extra load on the opamp at HF. If it is removed the distortion at low and middle frequencies is completely unchanged, but the sudden increase in distortion for the 9 Vrms case now occurs at the higher frequency of 36 kHz.

It is clear that the 5534A is not distortion-free in this application. The LM4562 has superior linearity and load-driving capabilities in general, though we know that the noise performance will be inferior due to its higher current noise. The results in Figure 11.7 are convincing.

The rise in distortion at LF has been completely eliminated, but disappointingly the HF distortion is barely improved at all. It is disconcerting that the opamp output is no more effective at driving the RIAA feedback network, given the excellent drive capabilities of the LM4562 into resistive loads. I suspect the reason is that the output stage uses VI limiting for overload protection, as opposed to the simple current-limiting used in the 5532, and this makes it more likely to come into action when driving a highly reactive load like an RIAA

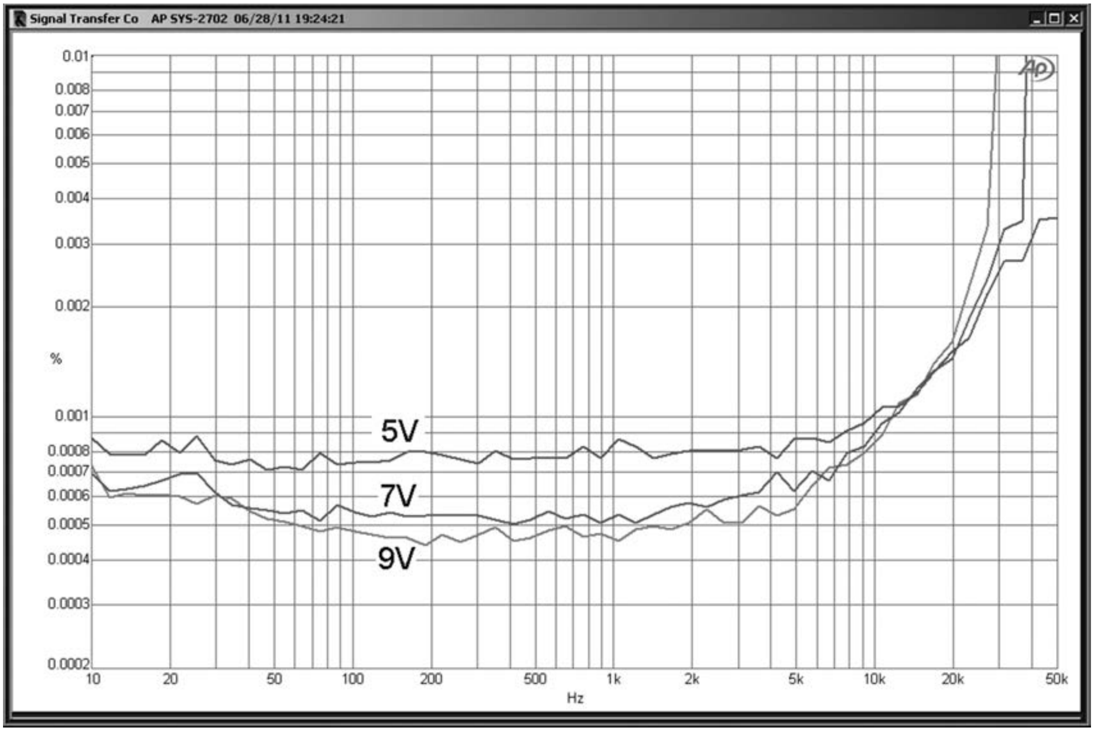

Figure 11.7: As for Figure 11.6 but using one LM4562 section

network, which above 1 kHz looks pretty much like a capacitance in series with a low-value resistor (R0) connected to ground. The use of a 'helper' opamp to assist in driving the RIAA network would be likely to reduce the distortion (see Chapter 1). Figure 11.7 confirms that the LF distortion is wholly from the 5534 opamp and is nothing to do with capacitor distortion.

Conclusions

While MM RIAA cartridge noise is a complicated business, the clear result is that the 5534A is the cheapest and easiest way to get within 3 dB of the theoretical best noise performance. This is because it not only has low noise in general, but also a favourable balance between its voltage noise and current noise for an MM source. It is a happy chance that it is inexpensive. To get within 2 dB of perfection, another 5532 can be added to implement load-synthesis.

Money can be saved by using a 5532 but you are then 5 dB from theoretical best noise, and running two channels through the same package is very likely to compromise crosstalk.

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CHAPTER 12

Moving-coil head amplifiers

Moving-coil cartridges are generally accepted to have a better tracking performance than moving-magnet cartridges because the moving element is a set of lightweight coils rather than a magnet which is inevitably made of a relatively dense alloy. Because the coils must be light, they consist of relatively few turns and the output voltage is very low, typically in the range 100 to 550 μVrms at a velocity of 5 cm/sec, compared with 5 mVrms from the average moving-magnet cartridge. Fortunately this low output comes from a very low impedance, which, by various technical means, allows an acceptable signal-to-noise performance to be obtained.

Moving-coil cartridge characteristics

There is much greater variation in impedance and output across the available range of moving-coil cartridges than for moving-magnet cartridges. The range quoted above is already much wider, but including the extremes currently on the market (2009) the output range is from 40 to 2500 μV, a remarkably wide span of 62 times or 36 dB. This is illustrated in Figure 12.1, which shows the results of a survey of 85 different MC cartridges (note that the ranges for the columns are wider at the right side of the diagram). When I first became involved in designing MC amplifiers in 1986, I compiled a similar chart [1] and it is interesting to note that the same features occurred – there are two separate clusters around 100–300 μV and 500–700 μV, and the lowest output is still 40 μV from an Audio Note cartridge (the loLtd model). The highest output of 2.5 mV comes from the Benz Micro H2, and this is only 6 dB below an MM cartridge.

Assuming that a conventional MM input stage is being used to raise 5 mV to nominal internal level and perform the RIAA equalisation, the Audio Note cartridge requires a gain of 125 times or +42 dB. The cartridge cluster around 200 μV output needs 25 times or +28 dB, while the 500 μV cluster requires 10 times or +20 dB. If an amplifier is to cover the whole range of MC cartridges available, some form of gain switching is highly desirable.

Cartridge impedances also vary markedly, over a range from 1 Ω (Audio Note loLtd) to 160 Ω (Denon DL-110 and DL160), with impedance increasing with output level, as you would expect – there are more turns of wire in the coils. The inductance of MC cartridges

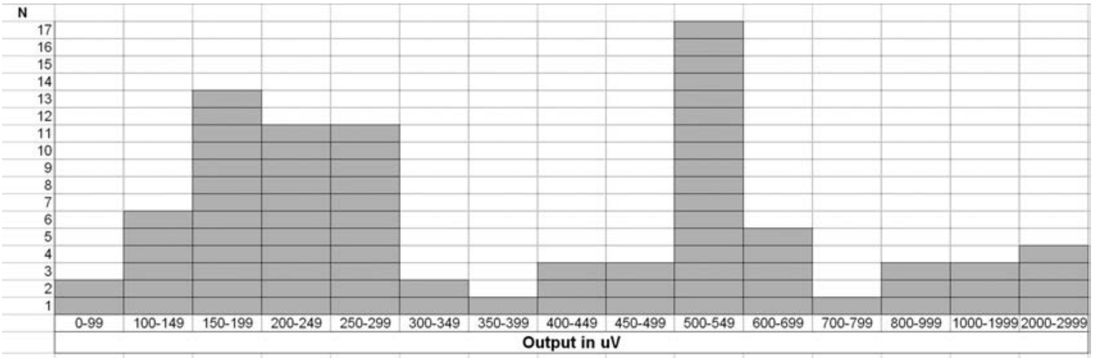

Figure 12.1: The output levels for 85 moving-coil cartridges at 5 cm/sec (2009)

TABLE 12.1 MC cartridge manufacturers whose product data was used to compile Figure 12.1

Audio Note	Immutable Music
Benz Micro	Koetsu
Cardas	Lyra
Clear Audio	Miyabi
Denon	Ortofon
Dynavector	Shelter
Goldring	Sumiko
Grado	van den Hul

is very low and their source impedance is normally treated as purely resistive. The recommended load impedances are also resistive (unlike the R–C combinations often used with MM cartridges) and are usually quoted as a minimum load resistance. Once more the variation is wide, from 3 Ω (Audio Note loLtd again) to 47 kΩ (Denon DL-110 and DL160 again) but a 100 Ω input load will be high enough for most of the cartridges surveyed, and 500 Ω will work for almost all of them. The Audio Note loLtd cartridge is unusual in another way – its magnetic field is produced not by permanent magnets but a DC-powered electromagnet, which presumably requires a very pure supply indeed. The manufacturers whose cartridges were included in the survey are listed in Table 12.1.

The limits on MC noise performance

Because MC cartridges can be modelled for noise purposes simply as their coil resistance, it is straightforward to calculate the best signal/noise ratio possible. Even if we assume a noiseless amplifier, the Johnson noise from the coil resistance sets an inescapable limit;

comparing this with the cartridge output gives the maximum signal/noise ratio. This was done for all the cartridges used to compile Figure 12.1, using the manufacturer's specs and the answers varied from 63.9 to 90.8 dB, which is a pretty big spread (this does not include RIAA equalisation).

In practice things will be worse. Even if we carry on assuming a noiseless amplifier, there is resistance in the tone-arm wiring, which has to be very thin for flexibility, and a bit more in the cable connecting the turntable to the preamp. Calculating the same figures for MM cartridges is a good deal more complicated because of the significant cartridge inductance (see Chapter 11).

Amplification strategies

There are two ways to achieve the high gains required for these low-output cartridges. In the most common method, a standard MM input stage, with RIAA equalisation, can be switched to either accept an MC input directly or the output of a specialised MC input stage which gives the extra gain needed; this may be either a step-up transformer or an amplifier configured to work well with very low source resistances. The large amount of gain required is split between the two stages, which makes it easier to achieve. Alternatively, a single stage can be used with switched gain; but this is not too hot an idea as:

1. Switchable gain makes accurate RIAA equalisation much harder.
2. For good noise performance, the input device operating current needs to be low for MM use (where it sees a high impedance) and high for MC use (where it sees a very low impedance). Making this operating current switchable would be a complicated business.
3. Achieving the very high gain required for MC operation together with low distortion and adequate bandwidth will be a challenge. It is unlikely to be possible with a single opamp, and so there is little likelihood of any saving on parts.

Moving-coil transformers

If you have a very low output voltage and very low impedance, an obvious way to deal with this is by using a step-up transformer to raise the voltage to the level where it can be appropriately applied to a moving-magnet amplifier stage, such as those discussed in Chapter 8. Such a transformer has most of the usual disadvantages such as frequency response problems and cost, though for hifi use the weight is not a difficulty, and non-linearity should not be an issue because of the very low signal levels.

In this application the cost is increased by the need for very high immunity to hum fields. While it is relatively straightforward to make transformers that have high immunity to external magnetic fields, particularly if they are toroidal in construction, it is not cheap, because the mu-metal cans that are required for the sort of immunity necessary are difficult to manufacture. The root of the problem is that the signal being handled is so very small. The transformer is usually working in a modern house which can have surprisingly large hum fields, and generally presents a hostile environment for very-low-signal transformers, so very good immunity indeed is required; some manufacturers use two nested screening cans, separately grounded, to achieve this, as shown in Figure 12.2. An inter-winding electrostatic screen is usually fitted. A stereo MC input naturally requires *two* of these costly transformers.

MC transformers are designed with low primary winding resistances, typically 2 or 3 Ω, to minimise the Johnson noise contribution from the transformer. Some transformers have windings made of silver rather than copper wire, but the conductivity of silver is only 5% higher than that of copper, and the increase in cost is startling. At the time of writing one brand of silver-wound transformer costs more than £1200 – each, not for a pair.

Because of the great variation in cartridge output levels and impedances some manufacturers (e.g. Jensen) offer transformers with two or three primary windings, which can be connected in series, parallel, or series-parallel to accommodate a wide variety of cartridges.

A transformer secondary must be correctly loaded to give the flattest possible frequency response, and this usually means that a Zobel R-C network must be connected across it, as explained in Chapter 18 on line inputs. This is Rd and Cd in Figure 12.2, where they have typical values. The values required depend not only on the transformer design but

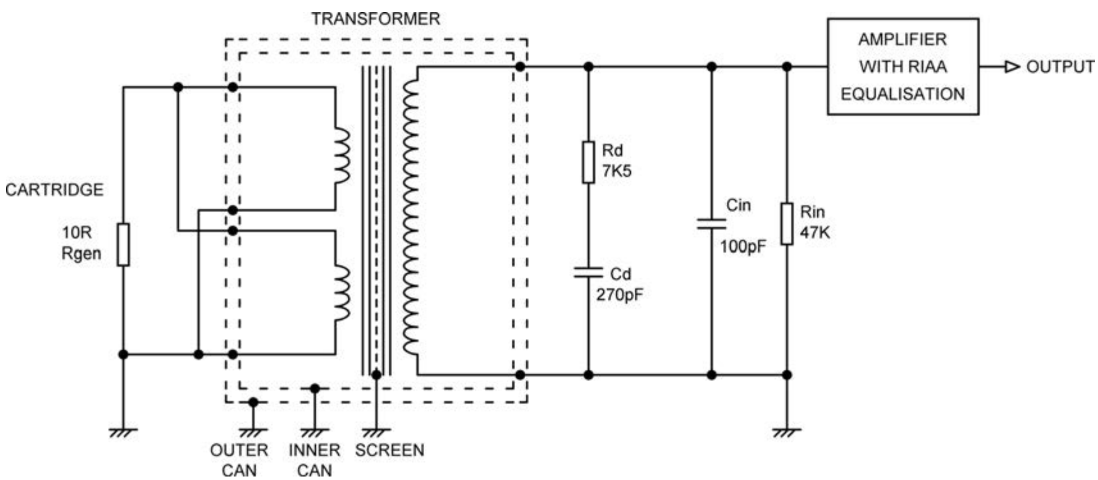

Figure 12.2: A typical MC step-up transformer circuit with twin primaries wired in parallel, dual screening cans, and Zobel network Rd, Cd across the secondary. Rin, Cin are the usual MM input loading components

also somewhat on the cartridge impedance, and some manufacturers such as Jensen are praiseworthily thorough in giving secondary loading recommendations for a wide range of cartridge impedances.

The very wide variation in cartridge outputs means that the step-up ratio of the transformer must be matched to the cartridge to get an output around 5 mV that is suitable for an MM input amplifier. For example, Jensen offer basic step-up ratios from 1:8 to 1:37. The maximum ratio is limited not only by transformer design issues but by the fact that the loading on the secondary is, as with all transformers, transferred to the primary divided by the square of the turns ratio. A 1:37 transformer connected to the 47 kΩ input impedance of an MM stage will have an impedance looking into the primary of only 34 Ω; such a transformer would, however, only be used with a very low-impedance low-output cartridge, which would be quite happy with such a loading. It is of course possible to arrange the input switching so the 47 kΩ input load is on the MM socket side of the MC/MM switch; the MM amplifier can then have a substantially higher input impedance.

Moving-coil input amplifiers

The high cost of transformers means that there is a strong incentive to come up with an electronic solution to the amplification problem. The only thing that makes it possible to achieve a reasonable signal-to-noise ratio is that the very small signal comes from a very low source impedance.

MC head-amplifiers come in many forms, but almost all in use today can be classified into one of the topologies shown in Figure 12.3, all of which use series-feedback. The configuration in Figure 12.3a is a complementary-feedback pair using a single input transistor chosen to have a low base series resistance R_b. The feedback network must also have a low impedance to prevent its Johnson noise from dominating the overall noise output, and this puts a heavy load on the second transistor. Typically a gain of 47 times will be implemented with an upper feedback resistor of 100 Ω and a lower resistor of 2 Ω, a total load on the amplifier output of 102 Ω. The combination of limited open-loop gain and the heavy load of the feedback network means that both linearity and maximum output level tend to be uninspiring, and the distortion performance is only acceptable because the signals are so small. An amplifier of this type is analysed in reference [2].

Figure 12.3b shows a classic configuration where multiple transistors are operated in parallel so that their gains add but their uncorrelated noise partly cancels. Two transistors gives a 3 dB improvement, four transistors 6 dB, and so on. The gain block A is traditionally one or two discrete devices, which again have difficulty in driving the low-impedance feedback network. Attention is usually paid to ensuring proper current-sharing between the input devices. This can be done by adding low-value emitter resistors to swamp V_{be} variations; they are

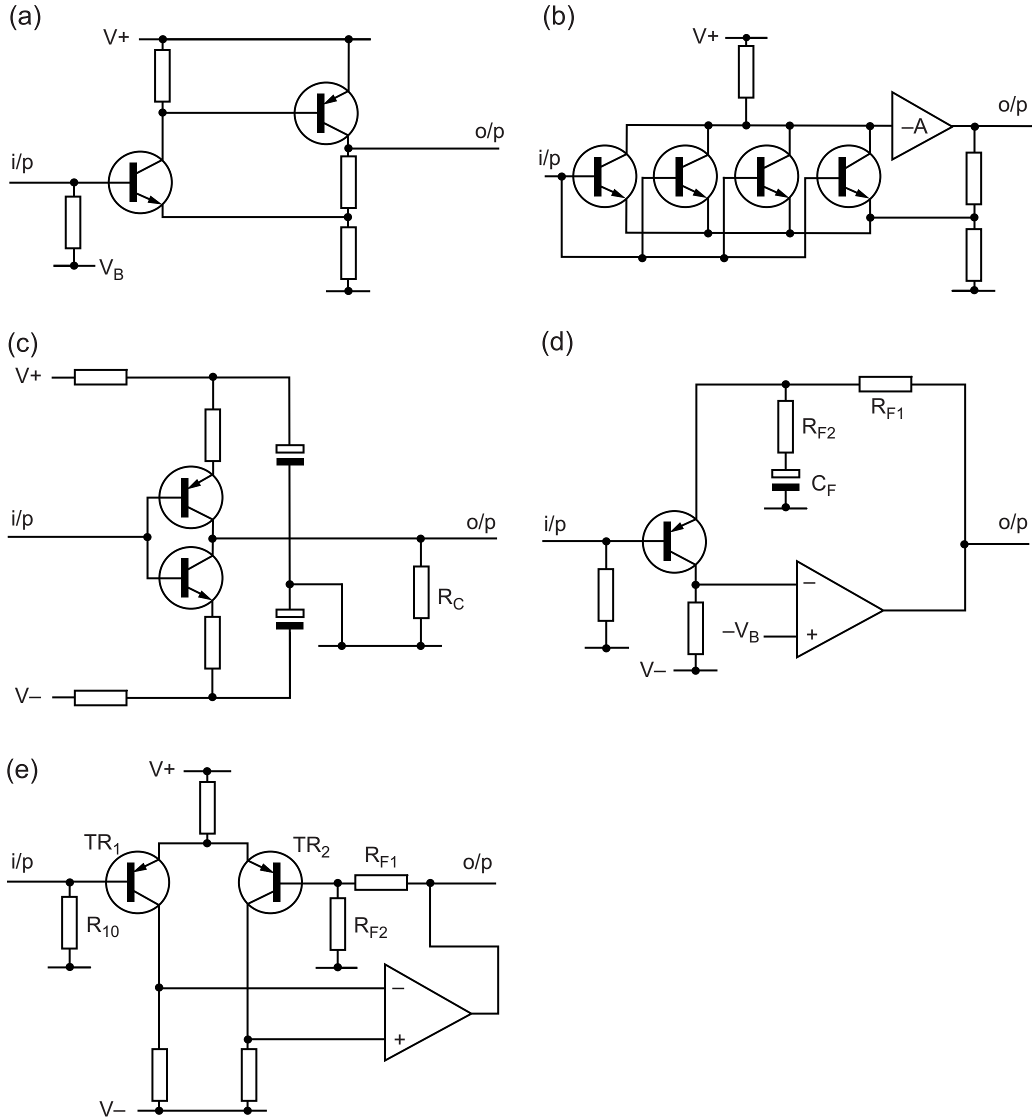

Figure 12.3: The popular MC amplifier configurations

effectively in series with the input path, and therefore degrade the noise performance unless each resistor is individually decoupled with a large electrolytic. Alternatively, each transistor can be given its own DC feedback loop to set up its collector current. For examples of this kind of circuitry see reference [3].

Figure 12.3c shows the series-pair configuration. This simple arrangement uses two complementary input transistors to achieve a 3 dB noise improvement without current-sharing problems as essentially the same collector current goes through each device. The collector signal currents are summed in R_c, which must be reasonably low in value to absorb collector-current imbalances. There is no feedback, so linearity is poor. The biasing arrangements are not shown.

Figure 12.3d is an enhancement of Figure 12.3a, with the input transistor inverted in polarity and the spadework of providing open-loop gain and output drive capability entrusted to an opamp. The much increased feedback gives excellent linearity, and less than 0.002% THD at full output may be confidently expected. However, problems remain. Rf_2 must be very low in value, as it is effectively in series with the input and will degrade the noise performance accordingly. If Rf_2 is 10 Ω (which is on the high side) C_f must be very large; for example, 1000 μF to limit the LF roll-off to −1 dB at 30 Hz. Adopting a quieter 3.3 Ω for the Rf_2 position gives significantly lower noise but demands 4700 μF to give −3 dB at 10 Hz; this is not elegant and leads to doubts as to whether the ESR of the capacitor will cause trouble. C_f is essential to reduce the gain to unity at DC because there is +0.6 V on the input device emitter, and we don't want to amplify that by fifty times.

The +0.6 V offset can be eliminated by the use of a differential pair, as in Figure 12.3e. This cancels out the V_{be} of the input transistor TR_1, at the cost of some degradation in noise performance. The pious hope is that the DC offset is so much smaller that if C_f is omitted, and the offset is amplified by the full AC gain, the output voltage swing will not be seriously reduced. The noise degradation incurred by using a differential pair was measured at about 2.8 dB. Another objection to this circuit is that the offset at the output is still non-negligible, about 1 V, mostly due to the base bias current flowing through R10. A DC-blocking capacitor on the output is essential.

An effective MC amplifier configuration

Finding none of these configurations satisfactory, I evolved the configuration shown as a block diagram in Figure 12.4. There is no C_f in the feedback loop, and indeed no overall DC feedback at all. The input transistor and the opamp each have their own DC feedback systems. The transistor relies on simple shunt negative feedback via DC loop 1; the opamp has its output held precisely to a DC level of 0 V by the integrator A2 which acts as DC loop 2. This senses the mean output level, and sets up a voltage on the non-inverting input of A1 that is very close to that at Q_1 collector, such that the output stays firmly at zero. The time-constant is made large enough to ensure that an ample amount of open-loop gain exists at the lowest audio frequencies. Too short a time-constant will give a rapid rise in distortion as frequency falls. Any changes in the direct voltage on Q_1 collector are completely uncoupled from the output, but AC feedback passes through Rf_1 as usual and ensures that the overall linearity is near-perfect, as is often the case with transistor opamp hybrid circuits. Due to

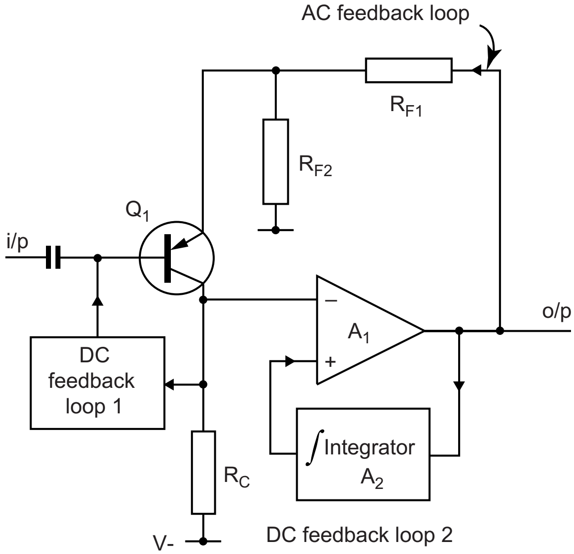

Figure 12.4: Block diagram of the MC preamplifier, showing the two DC feedback loops

the high open-loop gain of A the AC signal on Q_1 collector is very small and so shunt AC feedback through DC loop 1 does not significantly reduce the input impedance of the overall amplifier, which is about 8 kΩ.

As we have seen, MC cartridges vary greatly in their output and different amplifier gain settings are highly desirable. Usually it would be simple enough to alter Rf_1 or Rf_2, but here it is not quite so simple. The resistance Rf_2 is not amenable to alteration, as it is kept to the low value of 3.3 Ω by noise considerations, while Rf_1 must be kept up to a reasonable value so that it can be driven to a full voltage swing by an opamp output. This means a minimum of 500 Ω for the 5534/2. It is intriguing that amplifiers whose output is measured in milliVolts are required to handle so much current.

These two values fix a minimum closed-loop gain of about 44 dB, which is much too high for all but the most insensitive cartridges. My solution was to use a ladder output attenuator to reduce the overall gain; this would be anathema in a conventional signal path, because of the loss of headroom involved, but since even an output of 300 mVrms would be enough to overload virtually all MM amplifiers, we can afford to be prodigal with it. If the gain of the head amplifier is set to be a convenient 200 times (+46 dB) then adding output attenuation to reduce the overall gain to a more useful +20 dB still allows a maximum output of 480 mVrms. Lesser degrees of attenuation to give intermediate gains allow greater outputs, and these are summarized in Table 12.2. For testing, an Ortofon MC10 was used with +26 dB of gain, giving similar output levels to MM cartridges. This highly successful cartridge was in production for 30 years and has only recently been superseded; its impedance is 3.3 Ω.

TABLE 12.2 Gain options and maximum outputs

Gain	Gain (dB)	Max output (rms)
10 ×	+20	480 mV
20 ×	+26	960 mV
50 ×	+34	2.4 V
100 ×	+40	4.6 V
200 ×	+46	10 V

A final constraint on the attenuator is the need for low output impedances so the succeeding MM input stage can give a good noise performance. The MM input should have been optimized to give its best Noise Figure with relatively high source impedances, but a low source impedance will still reduce its actual noise output. This means that an output attenuator will need low resistor values, imposing yet more loading on the unfortunate opamp. This problem was solved by making the attenuator ladder an integral part of the AC feedback loop, as shown in Figure 12.5. This is practicable because it is known that the input impedance of the following MM stage will be too high at 47 kΩ to cause significant gain variations.

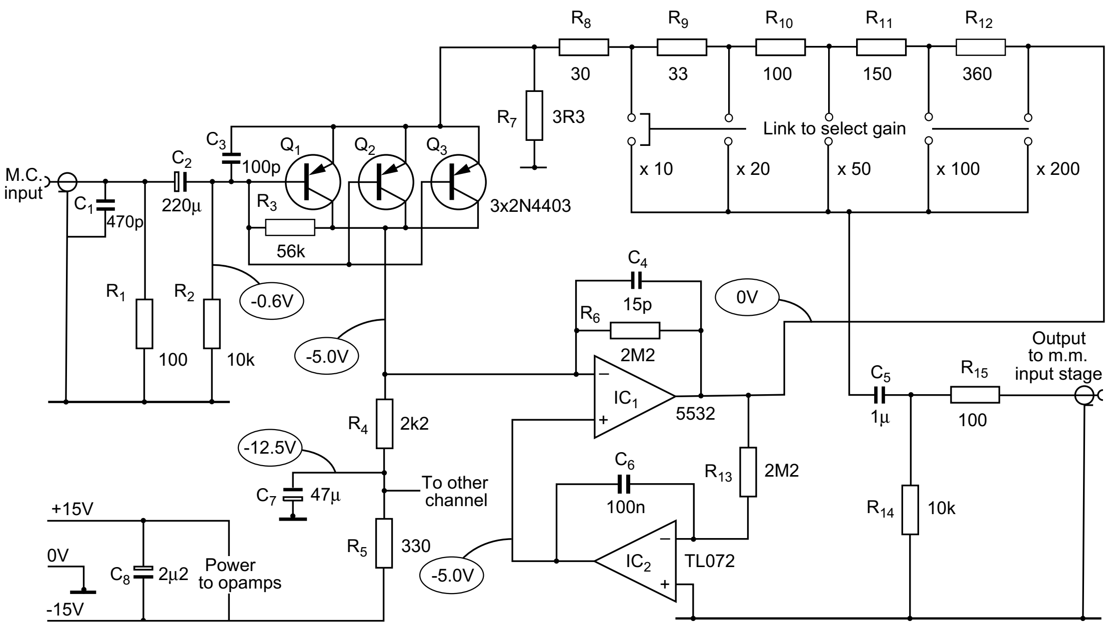

Figure 12.5: Circuit diagram of the MC preamplifier

The complete circuit

This is shown in Figure 12.5, and closely follows Figure 12.4, though you will note that the input devices have suddenly multiplied themselves by three. Capacitor C1 is soldered on the back of the MC input phono sockets and is intended for EMC immunity rather than cartridge response modification. If the need for more capacitive or resistive loading is felt, then extra components may be freely connected in parallel with R1. If R1 is raised in value, then load resistances of 5 kΩ or more are possible, as the impedance looking into C_2 is about 8 kΩ. Capacitor C2 is large to give the input devices the full benefit of the low source impedance, and its value should not be altered. Resistors R2, R3 make up DC loop 1, setting the operating conditions of Q1, Q2, Q3 while R_4 is the collector load, decoupled from the supply rail by C9 and R5, which are shared between two stereo channels. Opamp IC1 is a half of a 5532, providing most of the AC open-loop gain, and is stabilized at HF by C4. R6 has no real effect on normal operation, but is included to give IC1 a modicum of DC negative feedback and hence tidy behaviour at power-up, which would otherwise be slow due to the charging time of C2. IC2, half of a TL072, is the integrator that forms DC loop 2, its time-constant carefully chosen to give ample open-loop gain from IC1 at low frequencies, while avoiding peaking in the LF response that could occur due to the second time-constant of C2.

The ladder resistors R8–R12 make up the combined feedback network and output attenuator, the gain being selected by a push-on link in the prototype. A rotary switch could be used instead, but this should *not* be operated with the system volume up as this will cause loud clicks, due to the emitter current (about 4 mA) of Q1–Q3 flowing through R7, which causes voltage drops down the divider chain. Note that the current through R7 flows down the ground connection back to the PSU. Output resistor R15 ensures stability when driving screened cables, and C5 is included to eliminate any trace of DC offset from the output.

The power supply rails do not need to be especially quiet and a normal opamp supply is quite adequate.

Performance

The input transistor originally chosen was the 2N4403, a type that was acknowledged as superior for this kind of application for some years, due to its low R_b of about 40 Ω. A single device used in the circuit of Figure 12.5 gives an EIN of −138 dB with a 4 mA collector current and a 3.3 Ω source resistance. The Johnson noise from 3.3 Ω is −147.4 dBu, so we have a noise figure of 9.4 dB. It was then consistently found that putting devices in parallel without any current-sharing precautions whatever always resulted in a significant improvement in noise performance. On average, adding a second transistor reduced noise by 1.2 dB, and adding a third reduced it by another 0.5 dB, giving an EIN of −139.7 dBu and an NF of 7.7 dB. Beyond this, further multiplication was judged unprofitable, so a triple-device input was settled on. The current-sharing under these conditions was checked by measuring the voltage across 100 Ω resistors temporarily inserted in the collector paths. With 3.4 mA as the total current for the array it was found after much device-swapping that the worst case of imbalance was 0.97 mA in one transistor and 1.26 mA in another. The transistors were not all from the same batch. It appears that, for this device at least, matching is good enough to make simple paralleling practical.

A superior device for low source impedances was the purpose-designed 2SB737, with a stunningly low R_b of 2 Ω. Three of them improved the EIN to −141.0 dBu and the NF to 6.4 dB, albeit at significant cost. Sadly it is now obsolete (why, for heaven's sake?) but can still be obtained from specialised suppliers such as the Signal Transfer Company [4].

You will have spotted that R7, at 3.3 Ω, generates as much noise as the source impedance; this only degrades the noise figure by 1.4 dB, rather than 3 dB, as most of the noise comes from the transistors.

It would be instructive to compare this design with other MC preamplifiers, but it is not at all easy as their noise performance is specified in so many different ways it is virtually impossible to reduce them all to a similar form, particularly without knowing the spectral distribution of the noise (this chapter has dealt until now with unweighted noise referred to the input, over a 400 Hz–20 kHz bandwidth, and with RIAA equalisation *not* taken into account). Nonetheless, I suggest that this design is quieter than most, being within almost 6 dB of the theoretical minimum, with clearly limited scope for improvement. Burkhard Vogel has written an excellent article on the calculation and comparison of MC signal-to-noise ratios [5].

The performance is summarised in Table 12.3. Careful grounding is needed if the noise and crosstalk performance quoted is to be obtained.

When connected to a RIAA-equalised MM stage as described in Chapter 7, the noise output from the MM stage is −93.9 dBu at 10 times MC gain, and −85.8 dBu at 50 times. In the 10 times case, the MC noise is actually 1.7 dB lower than for MM mode.

TABLE 12.3 MC head amp performance figures

Input overload level	48 mV rms
Equivalent input noise	−141.0 dBu, unweighted, without RIAA equalisation (3.3 Ω source res)
Noise figure	6.4 dB (3.3 Ω source res)
THD	Less than 0.002% at 7 Vrms out (maximum gain) at 1 kHz Less than 0.004% 40 Hz–20 kHz
Frequency response	+0, −2 dB, 20 Hz–20 kHz
Crosstalk	Less than −90 dB, 1 kHz–20 kHz (layout dependent)
Power consumption	20 mA at ±15 V, for two channels

References

[1] Self, D. 'Design of Moving-Coil Head Amplifiers', *Electronics and Wireless World* (December 1987), p. 1206.

[2] Nordholt and Van Vierzen. 'Ultra Low Noise Preamp For Moving-Coil Phono Cartridges', *JAES* (April 1980), pp. 219–223.

[3] Barleycorn, J. (a.k.a. S. Curtis) *HiFi For Pleasure* (August 1978), pp. 105–106.

[4] www.signaltransfer.freeuk.com/

[5] Burkhard Vogel "The Sound of Silence" (Calculating MC preamp noise) *Electronics World* (Oct 2006) p. 28.

CHAPTER 13

Volume controls

Volume controls

A volume control is the most essential knob on a preamplifier – in fact the unhappily named ‘passive preamplifiers’ usually consist of nothing else but a volume control and an input selector switch. Volume controls in one guise or another are also freely distributed on the control surfaces of mixing consoles, examples being the auxiliary sends and the faders.

A volume control for a hifi preamplifier needs to cover at least a 50 dB range, with a reasonable approach to a logarithmic (i.e. linear-in dB) law, and have a channel balance better than ±1 dB over this range if noticeable stereo image shift is to be avoided when the volume is altered.

The simplest volume control is a potentiometer. These components, which are invariably called ‘pots’ in practice, come with various control laws, such as linear, logarithmic, anti-logarithmic, and so on. The control law is still sometimes called the ‘taper’ which is a historical reference to when the resistance element was actually physically tapered, so the rate of change of resistance from track-end to wiper could be different at different angular settings. Pots are no longer made this way, but the term has stuck around. An ‘audio-taper’ pot usually refers to a logarithmic type intended as a volume control.

All simple volume controls have the highest output impedance at the wiper at the −6 dB setting. For a linear pot this is when the control is rotated halfway towards the maximum, at the twelve o’clock position. For a log pot it will be at a higher setting, around three o’clock. The maximum impedance is significant because it usually sets the worst-case noise performance of the following amplification stage. The resistance value of a volume control should be as low as possible, given the loading/distortion characteristics of the stage driving it. This is sometimes called ‘low-impedance design’. Lower resistances mean:

1. Less Johnson noise from the pot track resistance
2. Less noise from the current-noise component of the following stage
3. Less likelihood of capacitive crosstalk from neighbouring circuitry
4. Less likelihood of hum and noise pickup.

Volume control laws

What constitutes the optimal volume control law? One answer is a strictly logarithmic or linear-in dB law, but this is in fact somewhat less than ideal, as an excessive amount of the pot rotation is used for very high and very low volume settings that are rarely used. It is therefore usual to have a law that is flatter in its central section, but falls off with increasing rapidity towards the low volume end – see the fader law later in this chapter. Sometimes the law steepens at the high-volume end as well but this is somewhat less common.

A linear pot is a simple thing – the output is proportional to the angular control setting, and this is usually pretty accurate, depending only on the integrity of the mechanical construction. Linear pots are given the code letter 'B'. See Table 13.1 for more code letters.

Log controls are rather less satisfactory. A typical log pot is not a precision attenuator with a fixed number of dB attenuation for each 10 degrees of shaft rotation. It is instead made up of two or three linear slopes that roughly approximate a logarithmic law, produced by superimposing two or three sections of track made with different resistivity material, the overlap usually being towards the bottom end of the control setting, such as 20% of full rotation (see Figure 13.1). These pots are usually given the code letter 'A'.

Anti-logarithmic pots are the same only constructed backwards, so that the slope change is at the top end of the control setting; these are typically used as gain controls for amplifying stages rather than as volume controls. These pots are usually given the code letter 'C'.

There is a more extreme version of the anti-logarithmic law where the slope change occurs at 10% of rotation instead of 20%. These are useful where you want to control the gain of an amplifying stage over a wide range, and still have something like a linear-in-dB control law. Typically they are used to set the gain of microphone input amplifiers, which can have a gain range of 50 dB or more. These pots are given the code 'RD', which stands for Reverse-D-law; I don't think I have ever come across a non-reverse D-law. Some typical laws are shown in Figure 13.1.

TABLE 13.1 Pot law identification letters

ALPS Code letter	Pot characteristic
A	Logarithmic
B	Linear
C	Anti-logarithmic
RD	Reverse-log

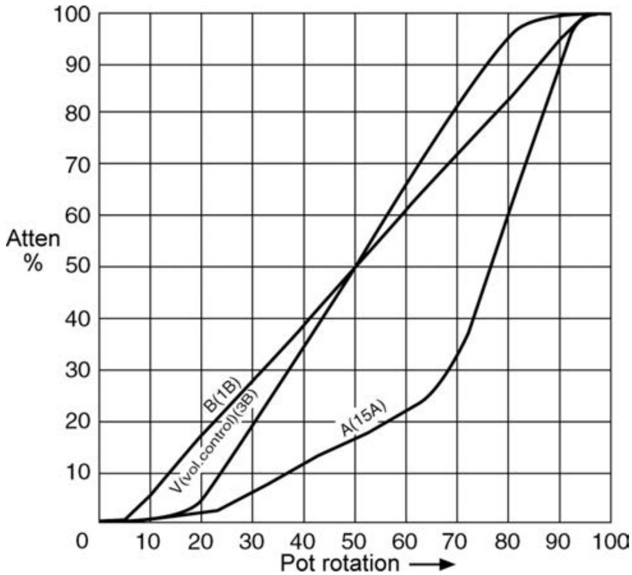

Figure 13.1: The control laws of typical linear and log pots. The log pot law A(15A) has three slope segments

Please note that the code letters are not adhered to quite as consistently across the world as one might hope. The codes given in Table 13.1 are those used by ALPS, one of the major pot makers, but other people use quite different allocations; for example, Radiohm, another major manufacturer, calls linear pots A and log pots B, but they agree that anti-log pots should be called C. Radiohm have several other laws called F, T, S, and X; for example, S is a symmetrical law apparently intended for use in balance controls. It clearly pays to check very carefully what system the manufacturer uses when you're ordering parts.

The closeness of approach to an ideal logarithmic law is not really the most important characteristic of a volume control – spreading out the most-used middle region is more useful. Of much greater importance is the matching between the two halves of a stereo volume control. It is common for the channel balance of log pots to deteriorate quite markedly at low volume settings, causing the stereo image to shift as the volume is altered. You may take it from me that customers really do complain about this, and so a good deal of ingenuity has been applied in attempts to extract good performance from indifferent components.

An important point in the design of volume controls is that their offness – the amount of signal that gets through when the control is at its minimum – is not very critical. This is in glaring contrast to a level control such as an auxiliary send on a mixer channel (see Chapter 22), where the maximum possible offness is very important indeed. A standard log pot will usually have an offness in the order of −90 dB with respect to fully up, and this is quite enough to render even a powerful hifi system effectively silent.

Loaded linear pots

Since ordinary log pots are not very accurate, many other ways of getting a log law have been tried. Trace 1 in Figure 13.3 (for a linear pot with no loading) makes it clear that the use of an unmodified linear law for volume control really is not viable; the attenuation is all cramped up at the bottom end. A rough approximation to a logarithmic law can be made by loading the wiper of a linear pot with a fixed resistor R1 to ground, as shown in Figure 13.2.

Adding a loading resistor much improves the law, but the drawback is that this technique really only works for a limited range of attenuation – in fact it only works well if you are looking for a control that varies from around 0 to −20 dB. It is therefore suitable for power amp level controls and aux master gain controls (see Chapter 22 for details of the latter), but is unlikely to be useful for a preamplifier gain control which needs a much wider logarithmic range. Figure 13.3 shows how the law varies as the value of the loading resistor is changed,

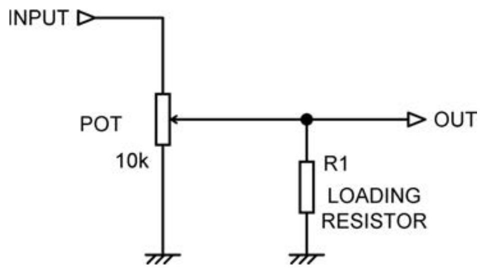

Figure 13.2: Resistive loading of a linear pot to approximate a logarithmic law

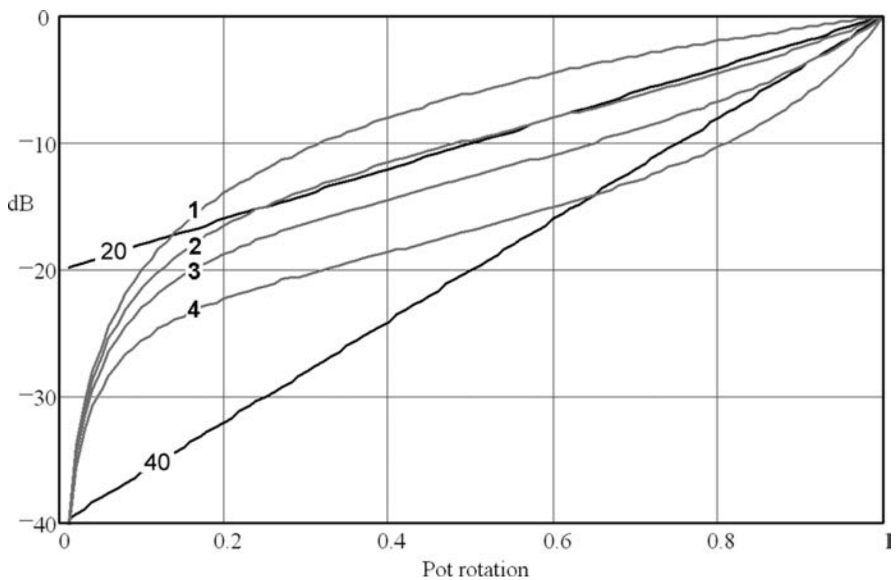

Figure 13.3: Resistive loading of a linear pot: the control laws plotted

and it is pretty clear that whatever its value, the slope of the control law around the middle range is only suitable for emulating the ideal log law labelled '20'. The value of the loading resistor for each trace number is given in Table 13.2.

Figure 13.3 shows that with the optimal loading value (Trace 2), the error in emulating a 0 to −20 dB log range is very small, lying within ±0.5 dB over the range 0 to −16 dB; below this the error grows rapidly. This error for Trace 2 only is shown in Figure 13.4.

Obtaining an accurate law naturally relies on having the right ratio between the pot track resistance and the loading resistor. The resistance of pot tracks is not controlled as closely as fixed resistors, their tolerance usually being specified as ±20%, so this presents a significant

TABLE 13.2 The loading resistor values used with a 10 kΩ pot in Figure 13.3

Trace number	Loading resistor R1 value
1	None
2	4k7
3	2k2
4	1 kΩ

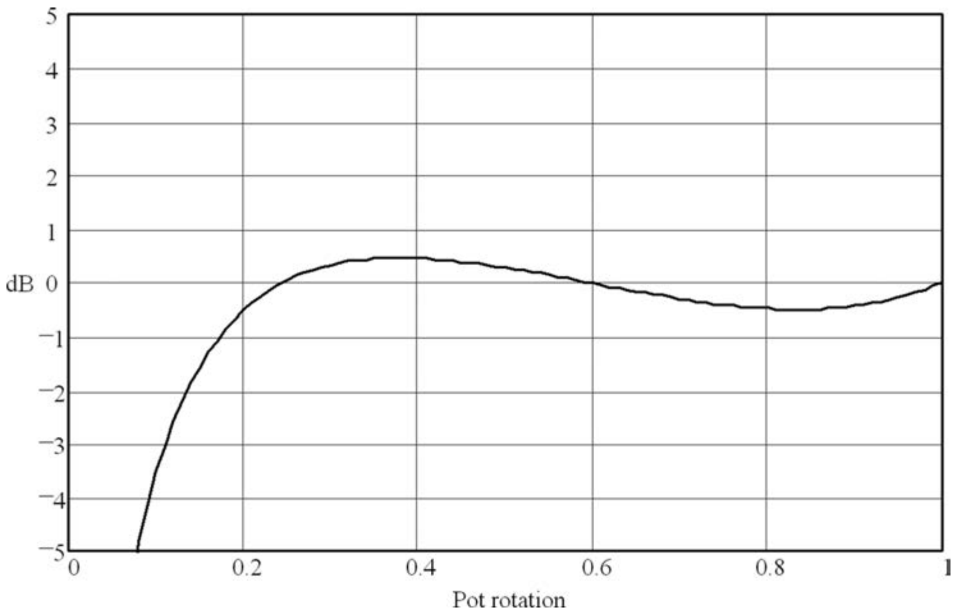

Figure 13.4: Loading of a linear pot: the deviation from an ideal 20 dB log law of Trace 2 in Figure 13.3

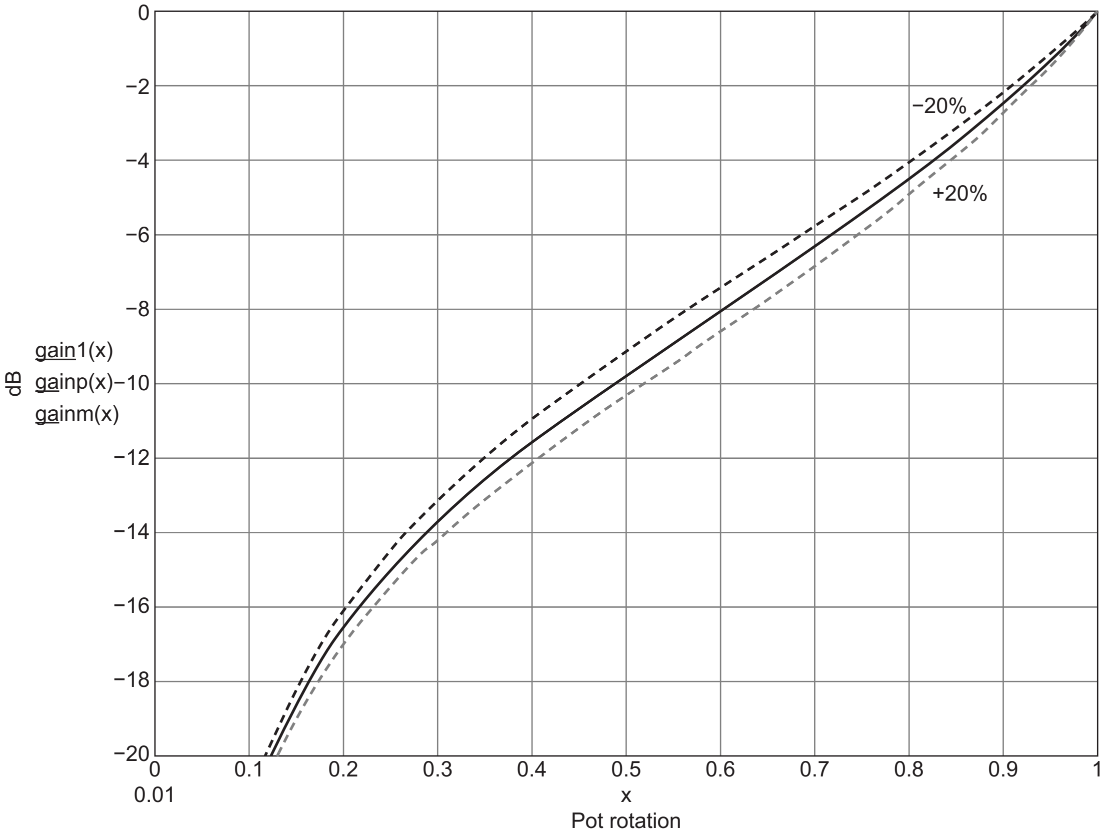

Figure 13.5: The effect of a ±20% pot track tolerance, with 4k7 resistive loading

balance-shift problem. The only solution would seem to be making the loading resistor trimmable, and this approach has been used in a master volume control by at least one mixing console manufacturer.

Figure 13.5 shows the effect on the control law of a ±20% pot track tolerance, with a loading resistor of 4k7 that is assumed to be accurate (trace 2 in Figure 13.3). With the pot track 20% low in value, the loading resistor has less effect and we get the dotted line above the solid (nominal) line. If it is 20% high the loading resistor has more effect, giving the lower dotted line. The error around the middle of control rotation is about ±0.7 dB, which is enough to give obvious balance errors in a stereo volume control. If you are unlucky enough to have the two tracks 20% out in opposite directions, the error will be 1.4 dB.

Another snag to this approach is that when the control is fully up, the loading resistor is placed directly across the input to the volume control, reducing its impedance drastically and possibly causing unhappy loading effects on the stage upstream. However the main problem is that the law is only good for a 0 to −20 dB range, inadequate for a volume control on a preamplifier.

The following stage must have a high enough impedance to not significantly affect the volume control law; this obviously also applies to plain logarithmic pots, and to all the passive controls described here.

Dual-action volume controls

In the previous section on loaded linear pots, we have seen that the control law is only acceptably logarithmic over a limited range – too limited for effective use as a volume control in a preamplifier, or as a fader or send control in a mixer. One way to fix this is the cascading of pots, so that their attenuation laws sum in dB. This approach does of course require the use of a four-gang pot for stereo, which may be objectionable because of increased cost, possible problems in sourcing, and a worsened volume-control feel. Nonetheless the technique can be useful, so we will give it a quick look.

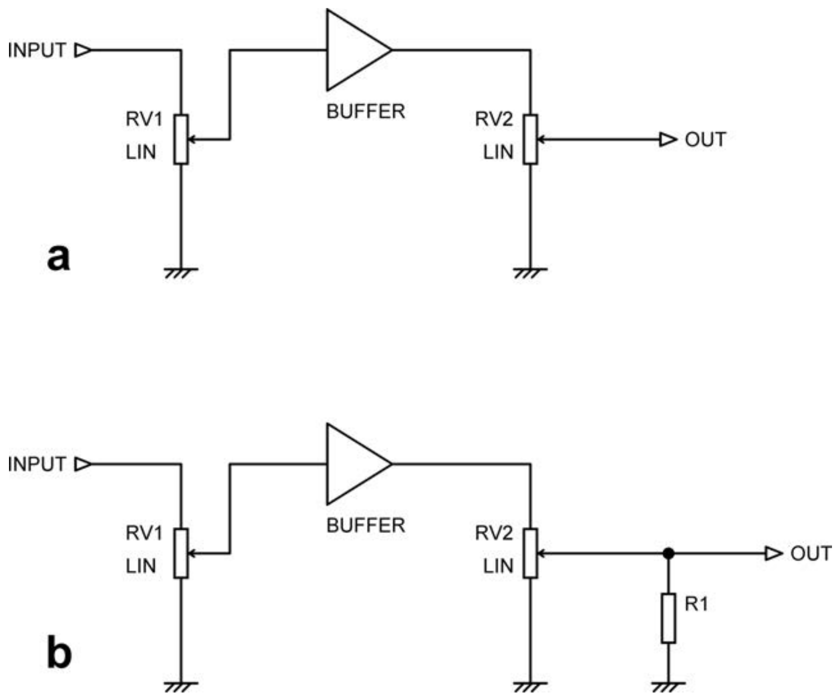

Figure 13.6: Dual-action volume controls: a) shows two linear pots cascaded, and b) is a linear pot cascaded with a loaded-linear pot

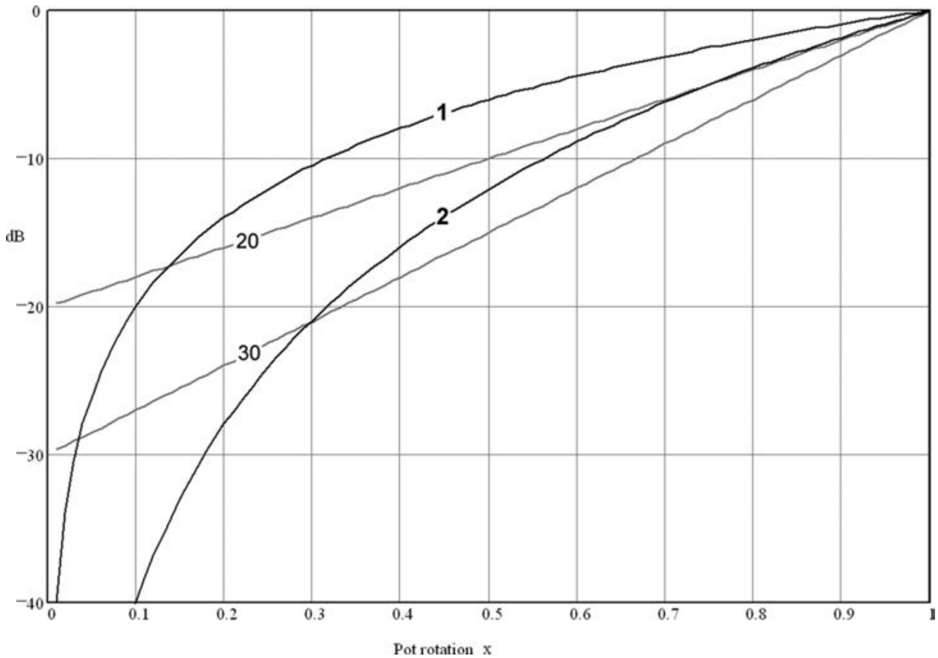

Figure 13.7: The dual-action volume control: Trace 1 is a linear pot, while Trace 2 is a square-law obtained by cascading two ganged linear pots with buffering between them

It is assumed there is no interaction between the two pots, so the second pot does not directly load the wiper of the first. This implies a buffer stage between them, as shown in Figure 13.6. There is no need for this to be a unity-gain stage, and in fact several stages can be interposed between the two pots. This gives what is usually called a *distributed gain control*, which can be configured to give a better noise/headroom compromise than a single volume control.

Figure 13.7 shows a linear law (Trace 1) and the square-law (Trace 2) made by cascading two linear pots. 20 dB and 30 dB ideal log lines are shown for comparison, and it is pretty clear that while the square-law is much more usable than the linear law, it is still a long way from perfect. It does however, have the useful property that, assuming the wiper is very lightly loaded by the next stage, the gain is dependant only upon control rotation and not the ratio between fixed 1% resistors and a ±20% pot track resistance. Figure 13.8 shows the deviation of the square-law from the 30 dB line; the error peaks at just over +3 dB before it plunges negative.

A *much* better attempt at a log law can be made by cascading an unloaded linear pot with a loaded-linear pot; the resulting law is shown in Figure 13.9. Trace 1 is the law of the linear pot alone, and Trace 2 is the law of a loaded linear 10 kΩ pot with a 2.0 kΩ loading resistor R1 from wiper to ground, alone. The combination of the two is Trace 3, and it can be seen that this gives a very good fit to a 40 dB ideal log line, and good control over a range of at least 35 dB.

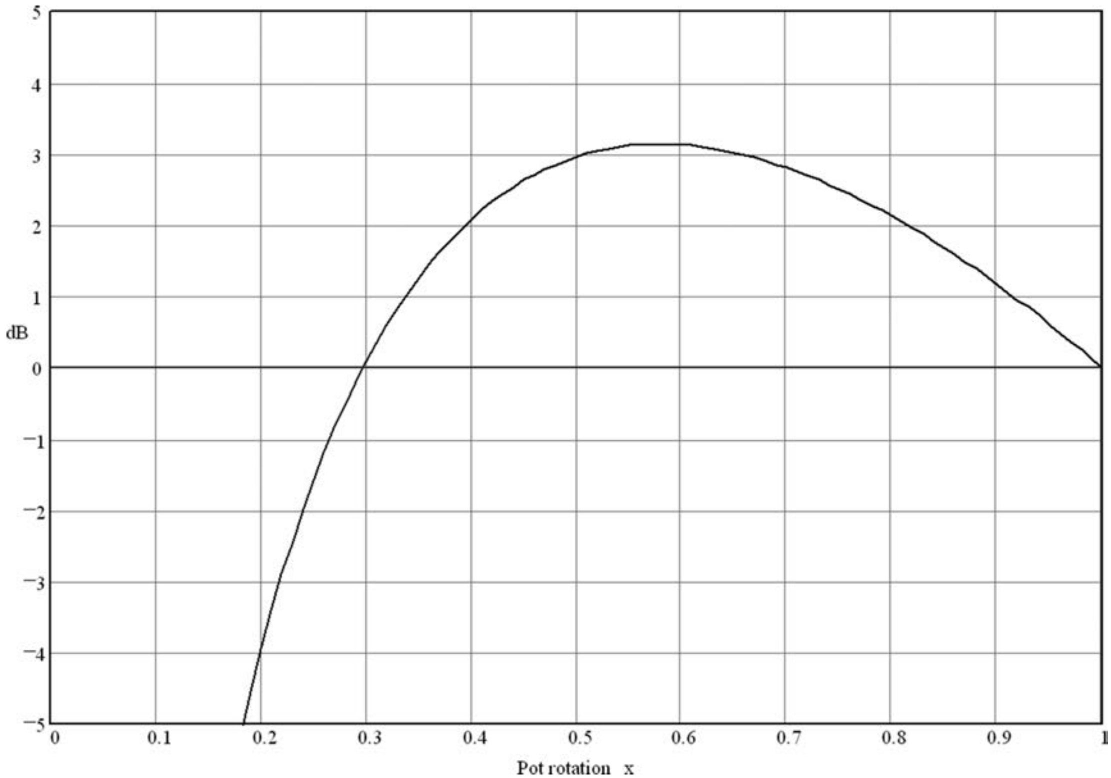

Figure 13.8: Dual-action volume control: the deviation of the square-law from the ideal 30 dB log line

Figure 13.10 shows the deviation of the combined law from the 40 dB line; the law error now peaks at just over ±1 dB. Unfortunately, adding the loading resistor means that once more the gain is dependent on the ratio between a fixed resistor and the pot track resistance.

Passive volume controls of various types can of course also be cascaded with an active volume control stage, and this can be a good way to obtain a desired control law. This is dealt with later in this chapter.

Tapped volume controls

The control law of a linear pot can be radically altered if it has a centre-tap (see Chapter 22 for the use of tapped pots for LCR panning). This can be connected to a potential divider that has a low impedance compared with the pot track resistance, and the attenuation at the tap point altered independently of other parameters. Figure 13.11 shows both unloaded and loaded versions of the arrangement.

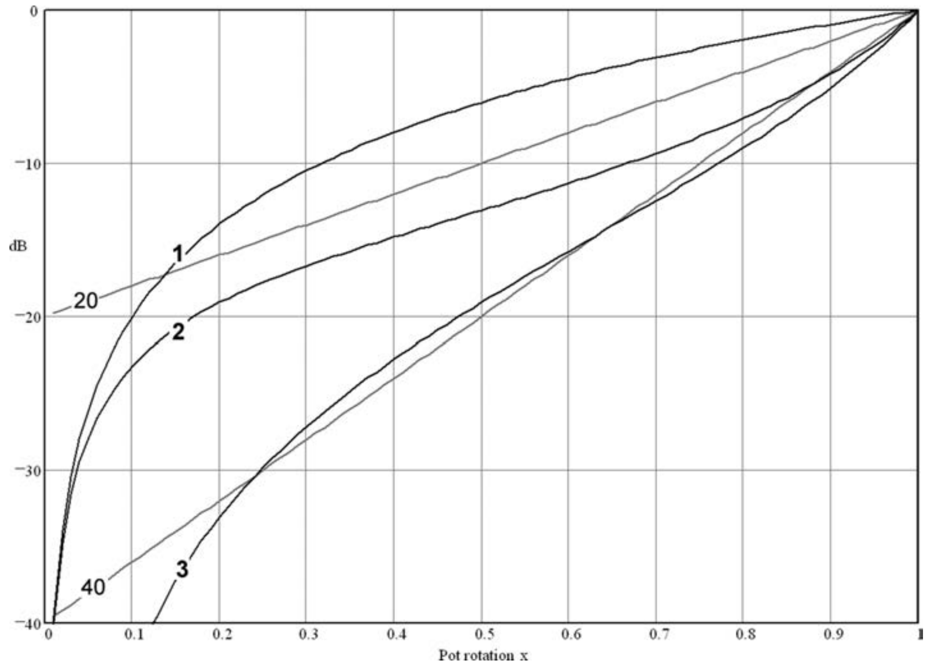

Figure 13.9: Dual-action volume control with improved law: linear pot cascaded with loaded-linear pot. Trace 3 is the combination of Traces 1 and 2 and closely fits the 40 dB log line

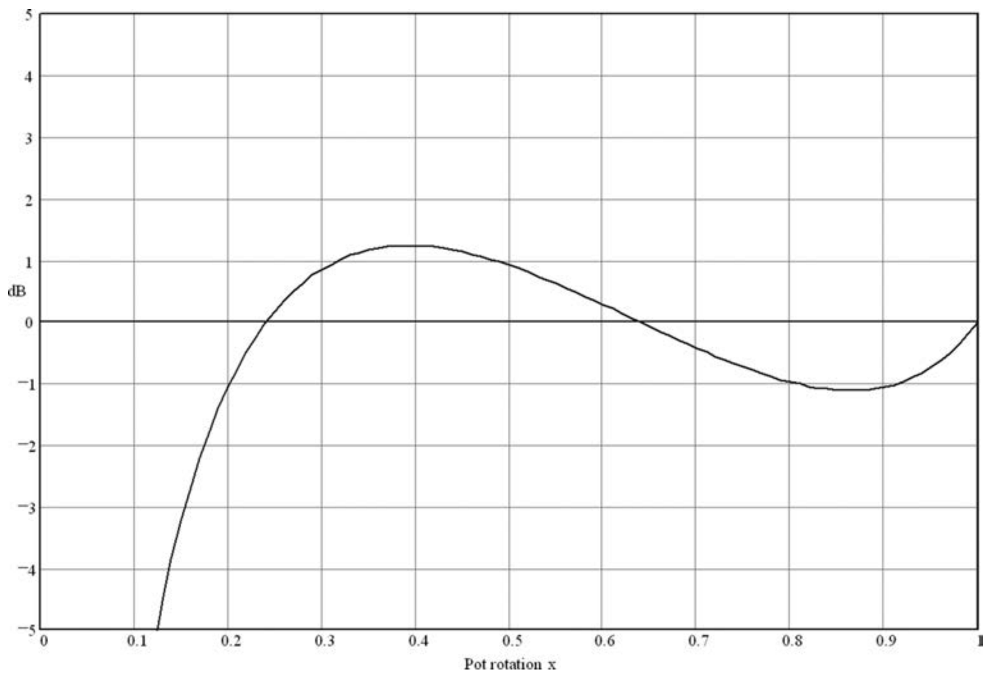

Figure 13.10: Dual-action volume control with improved law: deviation of the control law from the ideal 40 dB log line. Much better than Figure 13.8

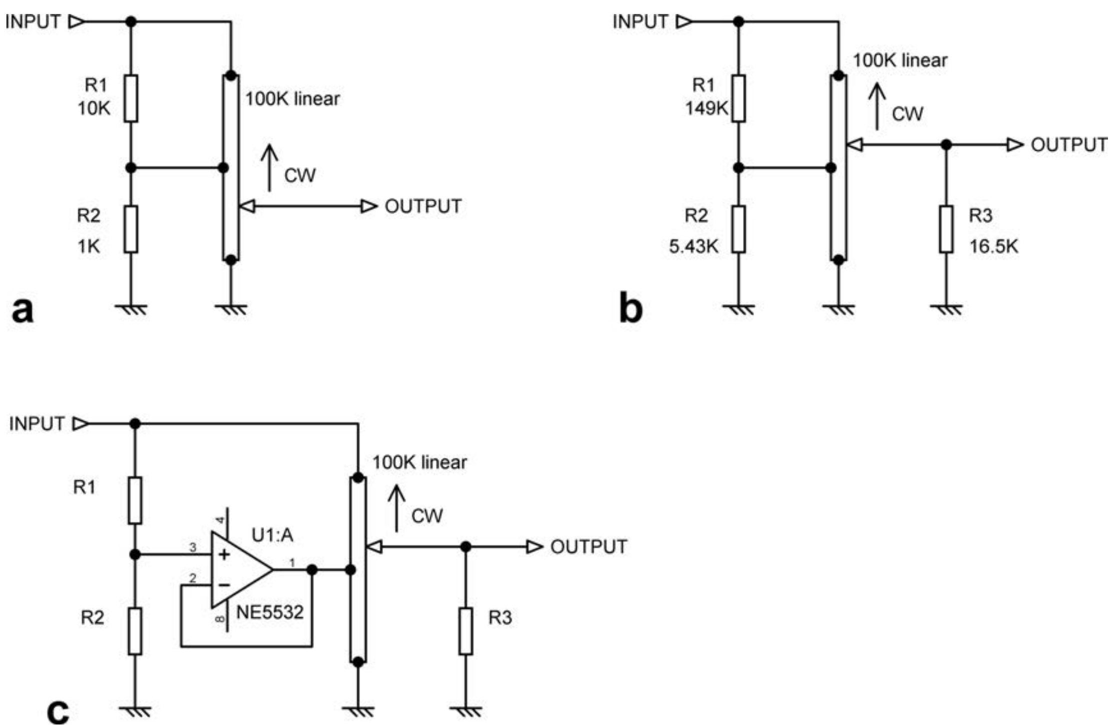

Figure 13.11: Tapped volume control: a) unloaded, b) loaded, c) with active control of tap voltage

The unloaded version shown in Figure 13.11a is arranged so that the attenuation at the tap is close to −20 dB. This gives the law shown in Figure 13.12; it approximates to a 33 dB log line, but there is an abrupt change of slope as the pot wiper crosses the tapping point.

Note that the track resistance has to be a good deal higher than that of the fixed resistors, so that they control the level at the tap. This means that, with the values shown, the source impedance at the wiper can be as high as 12.5 kΩ when it is halfway between the tap and one end, and this may degrade the noise performance, particularly if the following stage has significant current noise. A normal 10 kΩ pot, of whatever law, has a maximum output impedance of only 2.5 kΩ. To match this figure the values shown would have to be scaled down by a factor of 5, so that R1 = 2k, R2 = 200 Ω, and the track resistance is a more normal 10 kΩ. This scaling is quite practical, the load on the previous stage now being 1.62 kΩ.

The error is shown in Figure 13.13. This version of a tapped linear volume control covers a range of about 30 dB, and almost keeps the errors within ±3 dB; but that abrupt change of slope at the tap point is somewhat less than ideal.

A much better approach to a log law is possible if a loading resistor is added to the wiper, as with the loaded linear pot already examined. See Figure 13.11b for the circuit arrangement and Figure 13.14 for the control law. If the resistors are correctly chosen, using the ratios given in Figure 13.11b, the law can be arranged to have no change of slope at the tapping,

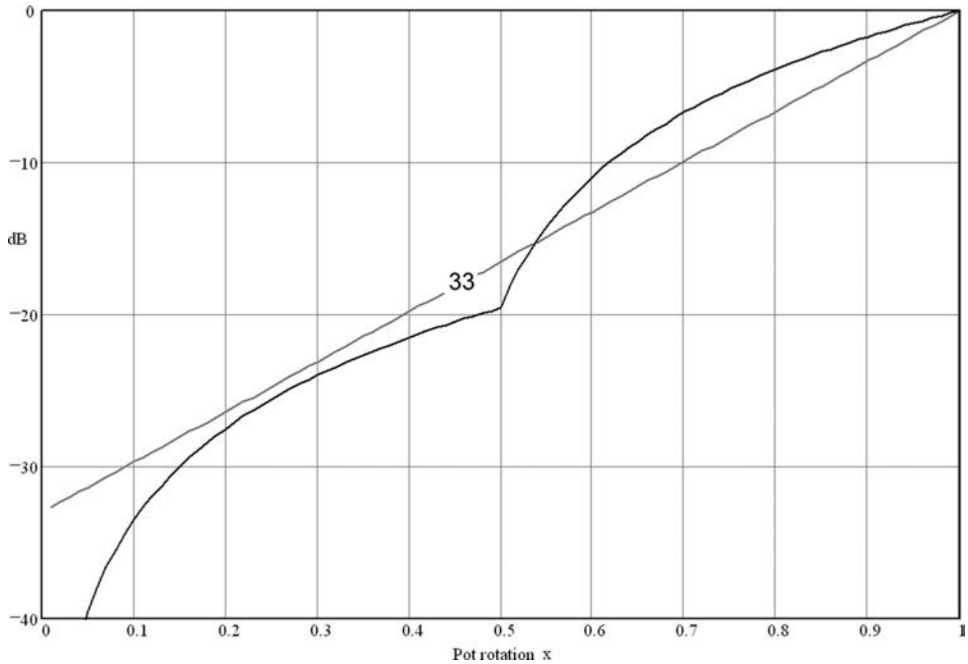

Figure 13.12: The law of an unloaded tapped volume control with −20 dB at the tap

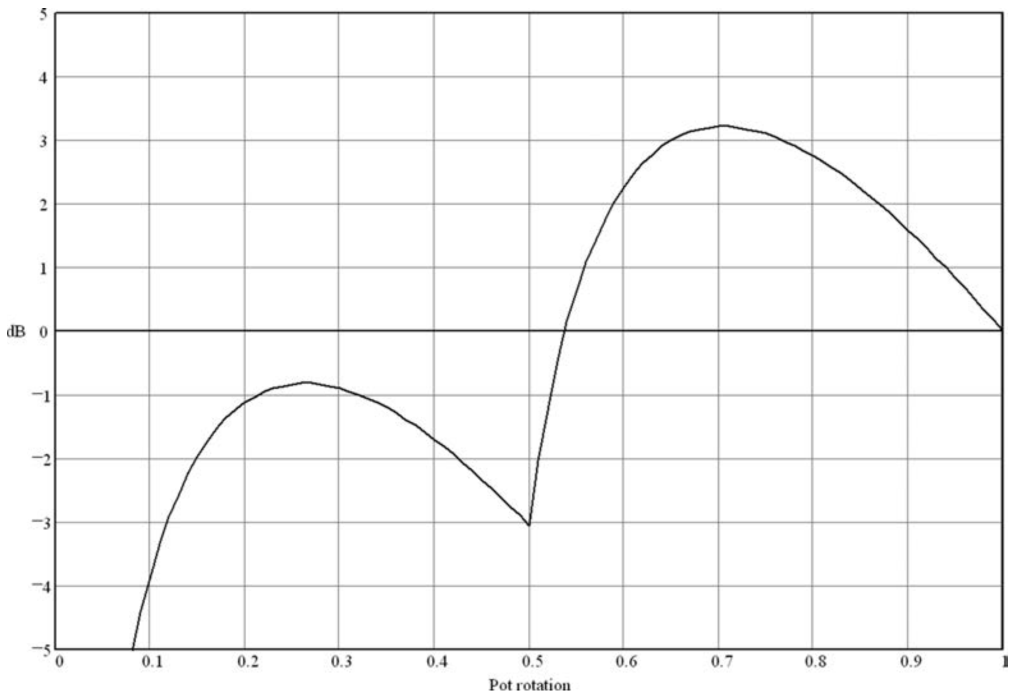

Figure 13.13: The deviation of the unloaded tapped volume control law from an ideal 33 dB log line

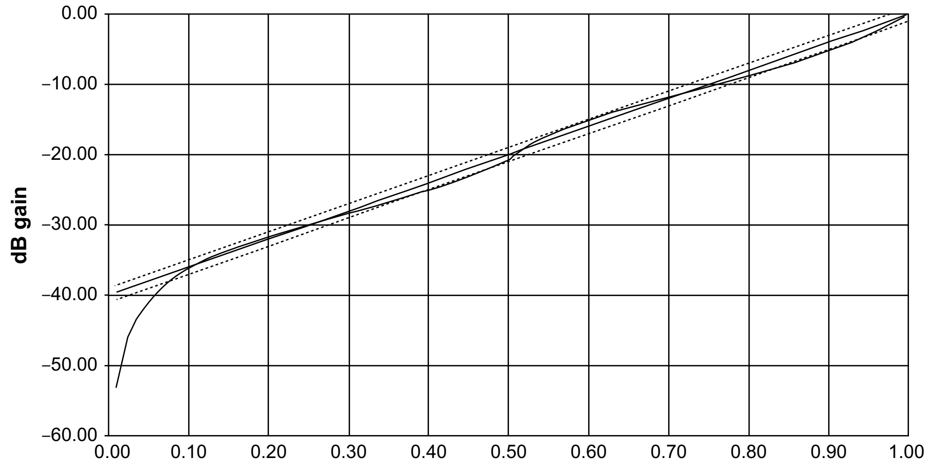

Figure 13.14: The law of a loaded and tapped volume control with −20 dB at the tap. The dotted lines show the width of a ±1 dB error band around the 40 dB ideal line

and the deviation from a 40 dB ideal line will be less than 1 dB over the range from 0 to −38 dB. The exact values of the resistors are given rather than the nearest preferred values. It is clear that this is a very effective way of giving an accurate law over a wide range, and it is widely used in making high-quality slide faders, using multiple taps and a conductive plastic track. Note that the accuracy still depends on good control of the track resistance value compared with the fixed resistors.

If the value of the resistors connected to the tap is such that the loading effect on the previous stage becomes excessive, one possible solution is shown in Figure 13.11c, where the R1 and R2 are kept reasonably high in value, and a unity-gain opamp buffer now holds the tap point in a vice-like grip due to its low output impedance. The disadvantage is that the noise of the opamp, and of R1 in parallel with R2, is fed directly through at full level when the wiper is near the tap. Often this will be at a lower level than the noise from the rest of the circuitry, but it is a point to watch. Note that because of the low-impedance drive from the opamp, the values of the fixed resistors will need to be altered from those of Figure 13.11b to get the best control law.

If the best possible noise performance is required, then it is better to increase the drive capability of the previous stage and keep the resistors connected to the tap low in value; if this is done by paralleling opamps then the noise performance will be improved rather than degraded.

Slide faders

So far as the design of the adjacent circuitry is concerned, a fader can normally be regarded as simply a slide-operated logarithmic potentiometer. Inexpensive faders are usually made using the same two-slope carbon-film construction as are rotary log volume controls, but

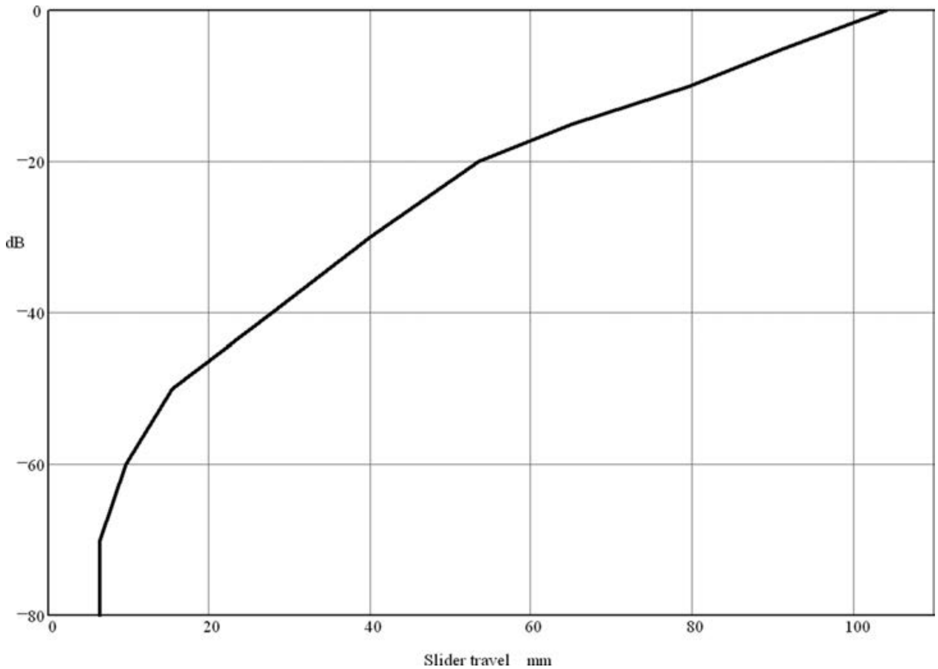

Figure 13.15: A typical control law for a 104 mm fader, showing how the attenuation is spread out over the upper part of the travel

the more expensive and sophisticated types use a conductive-plastic track with multiple taps connected to a resistor ladder, as described in the previous section. This allows much better control over the fader law.

High-quality faders typically have a conductive-plastic track, contacted by multiple gold-plate metal fingers to reduce noise during movement. A typical law for a 104 mm travel fader is shown in Figure 13.15. Note that a fader does not attempt to implement a linear-in-dB log scale; the attenuation is spread out over the top part of the travel and much compressed at the bottom. This puts the greatest ease of control in the range of most frequent use; there is very little point in giving a fader precise control over signals at −60 dB.

Figure 13.16a shows the straightforward construction used in smaller and less expensive mixers. There is some end resistance Ref at the bottom of the resistive track which compromises the offness, and it is further compromised by the voltage-drop down the resistance R_g of the ground wire that connects the fader to the channel PCB. To fix this a so-called 'infinity-off' feature is incorporated into the more sophisticated faders.

Figure 13.16b shows an infinity-off fader. When the slider is pulled down to the bottom of its travel, it leaves the resistive track and lands on an end-section with a separate connection back to the channel module ground. No signal passes down this ground wire and so there is no voltage-drop along its length. This arrangement gives extremely good maximum

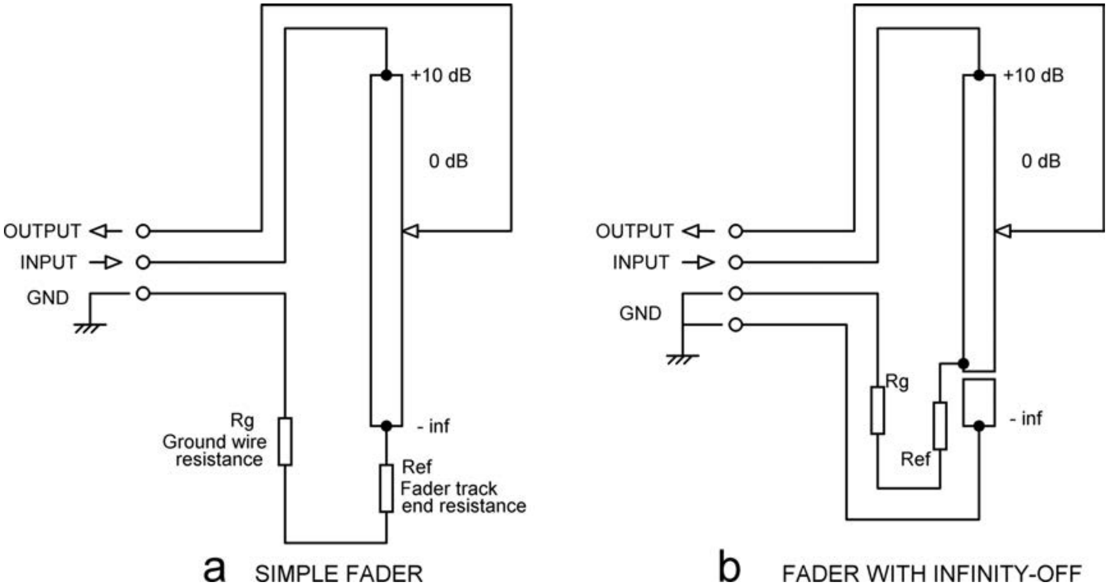

Figure 13.16: A simple fader a) and a more sophisticated version at b) with an 'infinity off' section to maximise offness

attenuation, orders of magnitude better than the simple fader, though of course 'infinity' is always a tricky thing to claim.

Faders are sometimes fitted with fader-start switches; these are microswitches which are actuated when the slider moves from the 'off' position. Traditionally these started tape cartridge machines; now they may be used to trigger digital replay.

Active volume controls

Active volume controls have many advantages. As explained in the section on preamplifier architectures, the use of an active volume control removes the dilemma concerning how much gain to put in front of the volume control and how much to put after it. An active gain control must fulfil the following requirements:

1. The gain must be smoothly variable from a maximum, usually well above unity, down to effectively zero, which in the case of a volume control means at least −70 dB. This at once rules out series-feedback amplifier configurations as these cannot give a gain of less than one, unless combined with a following passive control. Since the use of shunt feedback implies a phase-inversion, this can cause problems with the preservation of absolute polarity.
2. The control law relating shaft rotation and gain should be a reasonable approximation to a logarithmic law. It does not need to be strictly linear-in-decibels over its whole range; this would give too much space to the high-attenuation end, say around −60 dB, and it is better

to spread out the middle range of −20 to −50 dB, where the control will normally be used. These figures are naturally approximate as they depend on the gain of the power amplifier, speaker sensitivity, and so on. A major benefit of active gain controls is that they give much more flexible opportunities for modifying the law of a linear pot than does the simple addition of a loading resistor, which was examined and found somewhat wanting earlier in this chapter.

3. The opportunity to improve channel balance over the mediocre performance given by the average log pot should be firmly grasped. Most active gain controls use linear pots and arrange the circuitry so that these give a quasi-logarithmic law. This approach can be configured to remove channel imbalances due to the uncertainties of dual-slope log pots.

4. The noise gain of each amplifier involved should be as low as possible.

5. As for passive volume controls, the circuit resistance values should be as low as practicable to minimise Johnson noise and capacitive crosstalk.

Figure 13.17 shows a collection of possible active volume configurations, together with their gain equations. Each amplifier block represents an inverting stage with a large gain −A, i.e. enough to give plenty of negative feedback at all gain settings. It can be regarded as an opamp with its non-inverting input grounded. Figure 13.17a simply uses the series resistance of a log pot to set the gain. While you get the noise/headroom benefits of an active volume control, the retention of a log pot with its two slopes and resulting extra tolerances means that the channel balance is no better than that of an ordinary passive volume control using a log pot. It may in fact be worse, for the passive volume control is truly a potentiometer, and if it is lightly loaded differences in track resistance due to process variations should at least partially cancel, and one can at least rely on the gain being exactly 0 dB at full volume. Here, however, the pot is actually acting as a variable resistance, so variations in its track resistance compared with the fixed R1 will cause imbalance; the left and right gain will not even be the same with the control fully up. Given that pot track resistances are usually subject to a ±20% tolerance, it would be possible for the left and right channel gains to be 4 dB different at full volume. This configuration is not recommended.

Figure 13.17b improves on Figure 13.17a by using a linear pot and attempting to make it quasi-logarithmic by putting the pot into both the input and feedback arms around the amplifier. It is assumed that a maximum gain of 20 dB is required; it is unlikely that a preamplifier design will require more than that. The result is the law shown in Figure 13.18, which can be seen to approximate fairly closely to a linear-in decibels line with a range of −24 to +20 = 44 dB. This is a result of the essentially square-law operation of the circuit, in which the numerator of the gain equation increases as the denominator increases. This is in contrast to the loaded linear pot case described earlier, which approximates to a 20 dB line.

The deviation of the control law from the 44 dB line is plotted in Figure 13.19, where it can be seen that between control rotations of 0.1 and 1, and a gain range of almost 40 dB, the maximum error is ±2.5 dB. This sort of deviation from an ideal law is not very noticeable in

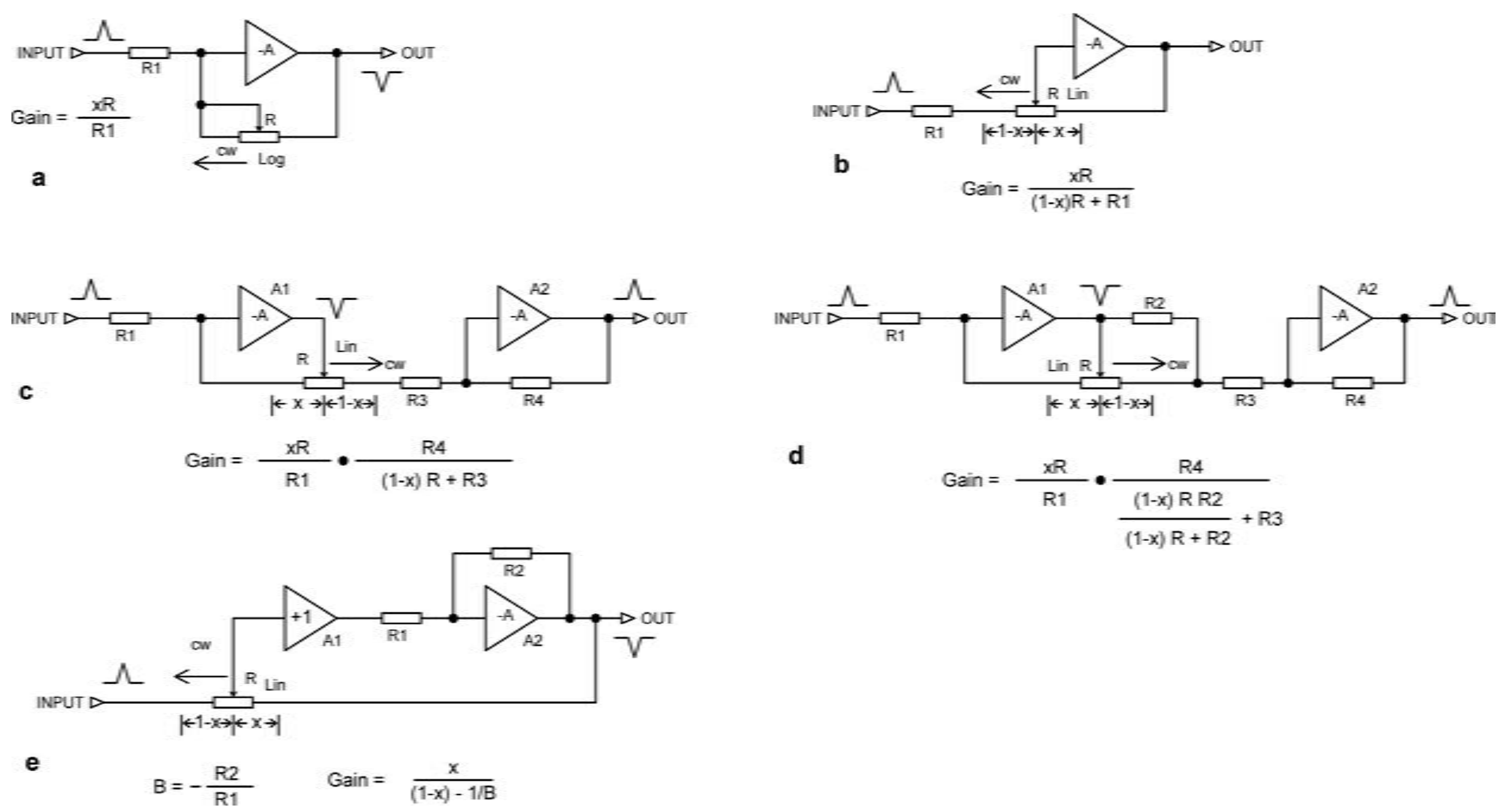

Figure 13.17: Active volume control configurations

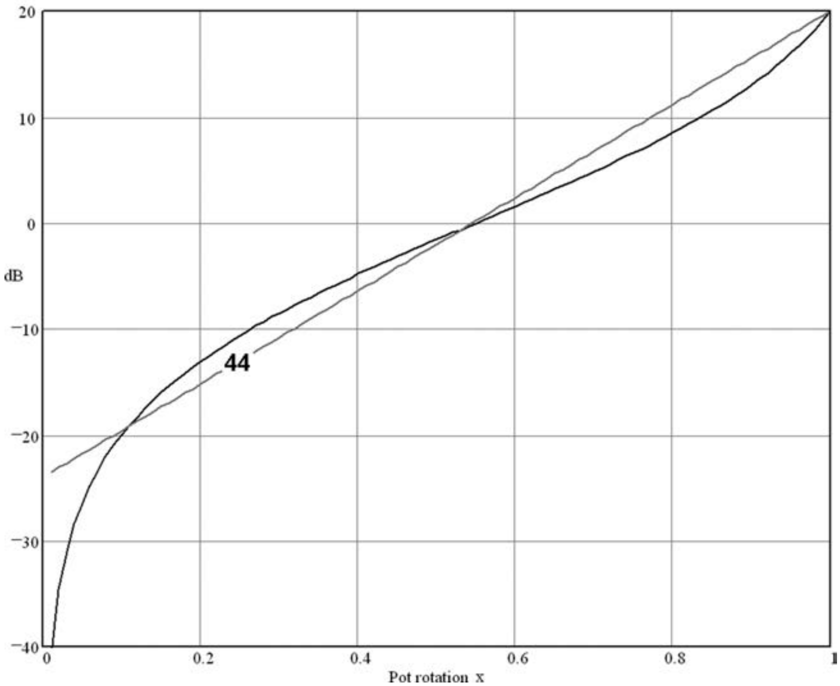

Figure 13.18: The control law of the active volume control in Figure 13.17b

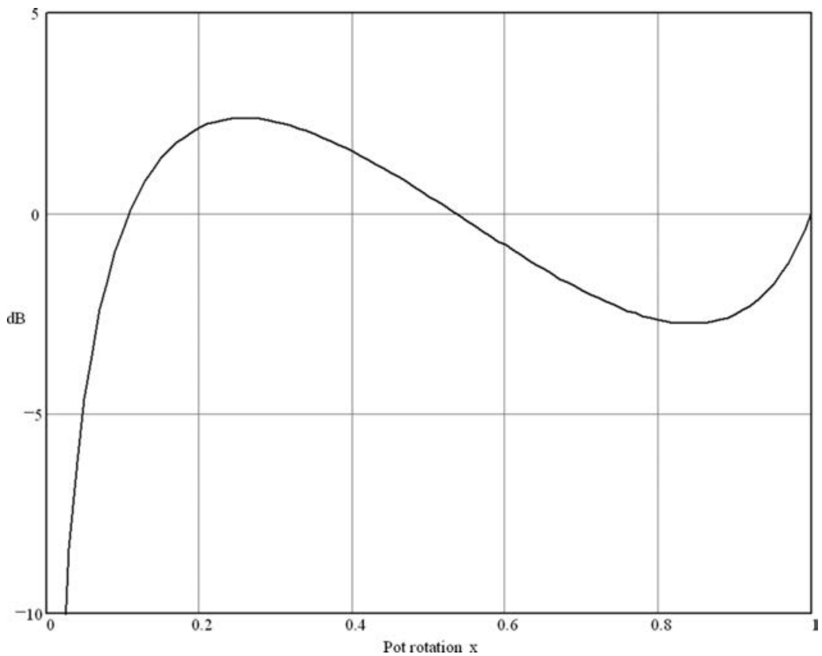

Figure 13.19: The deviation of the control law in Figure 13.18 from an ideal 44 dB logarithmic line

practice. A more serious issue is the way the gain heads rapidly south at rotations less than 0.1, with the result that volume drops rapidly towards the bottom of the travel, making it more difficult to set low volumes to be where you want them. Variations in track resistance tend to cancel out for middle volume settings, but at full volume the gain is once more proportional to the track resistance and therefore subject to large tolerances.

The configuration in Figure 13.17c also puts the pot into an input arm and a feedback arm, but in this case in separate amplifiers; the feedback arm of A1 and the input arm of A2. It requires two amplifier stages, but as a result the output signal is in the correct phase. When configured with R = 10 kΩ, R1 = 10 kΩ, R3 = 1 kΩ, and R4 =10 kΩ, it gives exactly the same law as Figure 13.17b, with the same maximum error of ±2.5 dB. It therefore may seem as pointless extra complication, but in fact the extra resistors involved give a greater degree of design freedom.

In some cases a linear-in decibels line with a range of 44 dB, which is given by the active gain stages already looked at, is considered too rapid; less steep laws can be obtained from a modified version of Figure 13.17c, by adding another resistor R2 to give the arrangement in Figure 13.17d. This configuration was used in the famous Cambridge Audio P50 integrated amplifier, introduced in 1970. When R2 is very high, the law approximates to that of Figure 13.17c. With R2 reduced to 4 kΩ, the law is modified to Trace 2 in Figure 13.20; the law is shifted up, but in fact the slope is not much altered, and is not a good approximation to

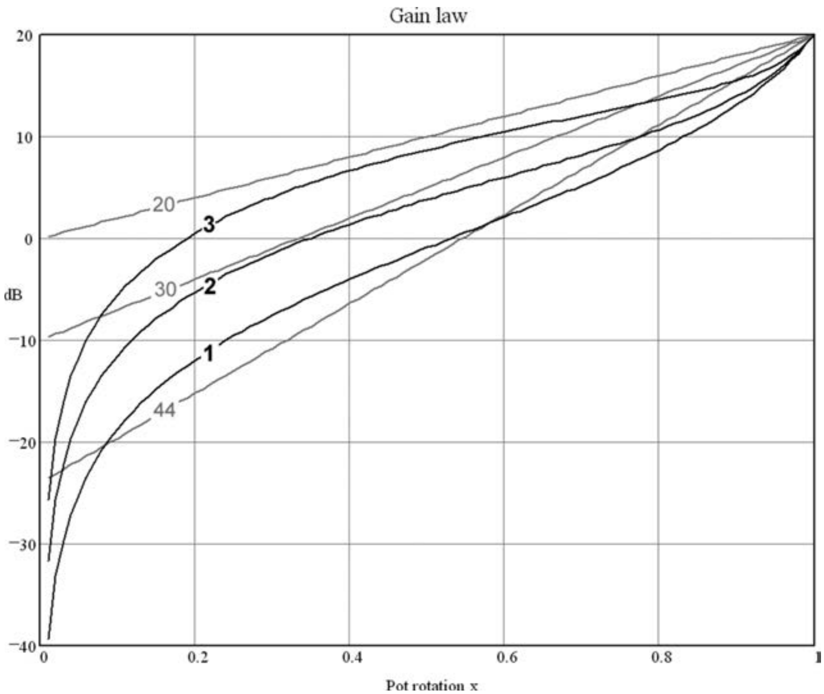

Figure 13.20: The control law of the active volume control in Figure 13.17d

the ideal 30 dB log line, labelled '30'. When R2 is reduced to 1 kΩ, the law is as Trace 3 in Figure 13.17d, and is a reasonable fit to the ideal 20 dB log line, labelled '20'. Unfortunately varying R2 can do nothing to help the way that all the laws fall off a cliff below a control rotation of 0.1, and in addition the problem remains that the gain is determined by the ratio between fixed resistors, for which a tolerance of 1% is normal, and the pot track resistance, with its ±20% tolerance. For this reason, none of the active gain controls considered so far are going to help with channel balance problems.

The Baxandall active volume control

The active volume control configuration in Figure 13.17e is due to Peter Baxandall. Like so many of the innovations conceived by that great man, it authoritatively solves the problem it addresses [1]. Figure 13.21 shows the law obtained with a maximum gain of +20 dB; the best-fit ideal log line is now 43 dB. There is still a rapid fall-off at low control settings.

You will note that there are no resistor or track resistance values in the gain equation in Figure 13.17e; the gain is only a function of the pot rotation and the maximum gain set up by R1, R2. As a result quite ordinary dual linear pots can give very good channel matching.

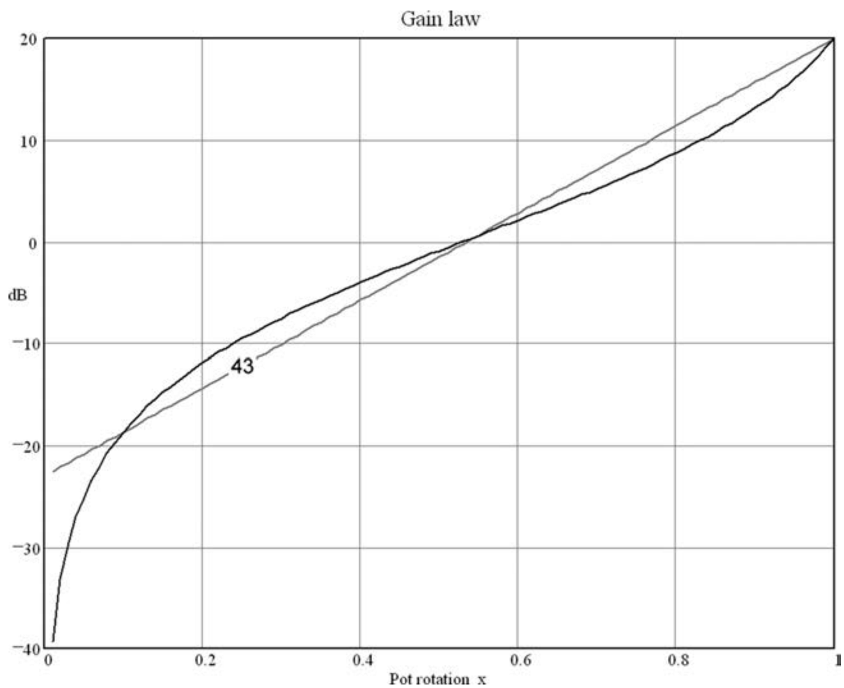

Figure 13.21: The control law of the +20 dB Baxandall active volume control in Figure 13.17e

When I tried a number of RadioOhm 20 mm diameter linear pots, the balance was almost always within 0.3 dB over a 46 dB gain range, with occasional excursions to an error of 0.6 dB.

However, the one problem that the Baxandall configuration cannot solve is channel imbalance due to mechanical deviation between the wiper positions. I have only once found that the Baxandall configuration did not greatly improve channel balance; in that case the linear pots I tried, which came from the same Chinese source as the log pots that were provoking customer irritation, had such poor mechanical alignment that the balance improvement obtained was small, and not worth the extra circuitry.

Note that all the active gain configurations require a low-impedance drive if they are to give the designed gain range; don't try feeding them from, say, the wiper of a balance control pot. The Baxandall configuration inherently gives a phase inversion that can be highly inconvenient if you are concerned to preserve absolute phase, but this can be undone by an inverting tone-control stage, such as the, er, Baxandall type.

An important point is that while at a first glance the Baxandall configuration looks like a conventional shunt feedback control, its action is modified by the limited gain set by R1 and R2. This means that the input impedance of the stage falls as the volume setting is increased, but does *not* drop to zero. With the values shown in Figure 13.23, input impedance falls steadily from a maximum of 10 kΩ at zero gain, to a minimum of 1.27 kΩ at maximum gain. If the preceding stage is based on a 5532 it will have no trouble driving this. Another consequence of the gain of the A2 stage is that the signals handled by the buffer A1 are never very large. This means that R1, and consequently R2, can be kept low in value to reduce noise without placing an excessive load on the buffer.

All the active volume controls examined here, including the otherwise superior Baxandall configuration, give a gain law that falls very rapidly in the bottom tenth of control rotation. It is not easy to see that there is any cure for this apart from using switched resistors instead of pots; any law can then be achieved.

The Baxandall volume control law

While the Baxandall configuration has several advantages, it is not perfect. One disadvantage is that the gain/rotation law is determined solely by the maximum gain, and it is not possible to bend it about by adding resistors without losing the freedom from pot-value-dependence.

Figure 13.22 shows the control laws for different maximum gains. Pot rotation is described here as Marks from Mk 0 to Mk 10 for full rotation. No provision is made for rock bands seeking controls going up to Mk 11 [2]. Changing the maximum gain has a much smaller effect on the gain at the middle setting (Mk 5).

Very often a maximum gain of +10 dB is required in preamplifier design, giving us −4 dB with the volume control central. In this case the control law is rather flat, with only 14 dB

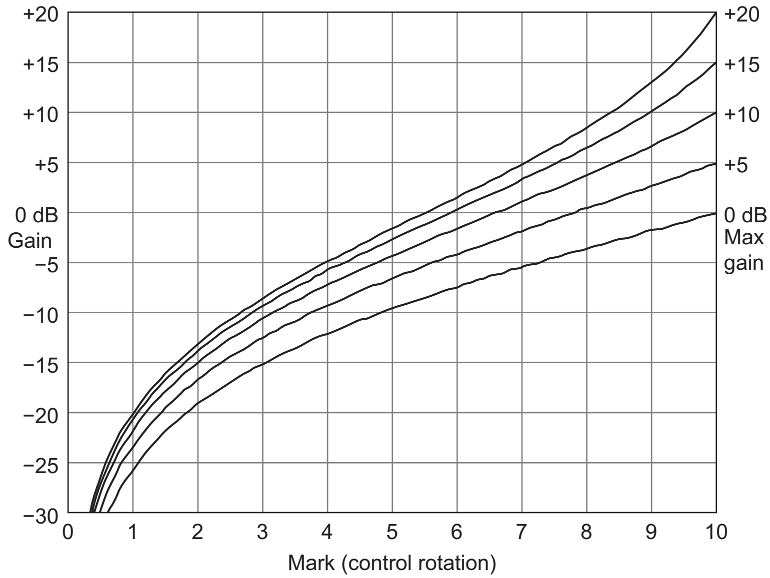

Figure 13.22: The gain law of a Baxandall volume control stage depends only on the maximum gain; plotted here for maximum gains of 0, +5, +10, +15 and +20 dB

change of gain in the top half of control rotation. The +10 dB law approximates closely to a linear-in-decibels line with a range of −18 to +10 = 28 dB. This has a shallower slope than the 43 dB log line shown in Figure 13.21, and is not ideal for a volume control law. Things can be much improved by combining it with a linear law; more on this later.

A practical Baxandall active volume stage

I have designed several preamplifiers using a Baxandall active volume control [3], [4], [5], [6]. The practical circuitry I employed for [4] is shown in Figure 13.23; this includes DC-blocking arrangements to deal with the significant bias currents of the 5532 opamp. The maximum gain is set to +17 dB by the ratio of R1, R2, to amplify a 150 mV line input to 1 V with a small safety margin.

This active volume-control stage gives the usual advantages of lower noise at gain settings below maximum, and excellent channel balance that depends solely on the mechanical alignment of the dual linear pot – all mismatches of its electrical characteristics are cancelled out. Note that in the first two preamplifier designs referenced here, all the pots were identical at 10 kΩ linear, apart from the question of centre-detents, which are desirable only on the balance, and treble and bass boost/cut controls.

The values given here are as used in the Precision Preamplifier '96 [4]. Compared with [3], noise has been reduced slightly by an impedance reduction on the gain-definition

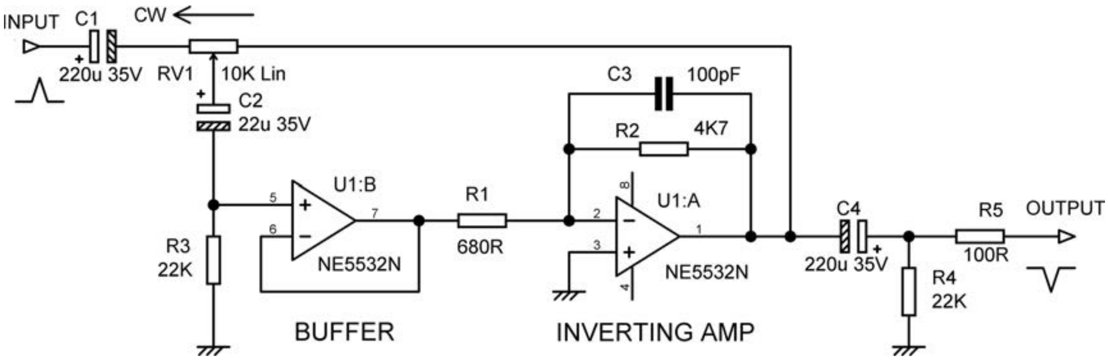

Figure 13.23: A practical Baxandall active volume control with DC-blocking, as used in the Precision Preamplifier '96 Maximum gain is +17dB

network R1, R2. The limit on this is the ability of buffer U1:B to drive R1, which has a virtual-earth at its other end. C3 ensures HF stability if there are excess phase-shifts due to stray capacitance. C1 prevents any DC voltages from circuitry upstream from reaching the volume-control stage. The input bias current of U1:B will produce a small voltage drop across R3, and C2 prevents this from reaching the control pot. Since two terminals of the pot are DC-blocked, it is now permissible to connect the third terminal to the output of U1:A as no current can flow through the control. The offset voltage at the output of U1:B will be amplified by U1:A, but should still be much too small to have any significant effect on available voltage swing, and it is prevented from leaving this stage by DC-blocking capacitor C4. R4 is a drain resistor to prevent voltage building up on the output due to leakage through C4, and R5 ensures stability by isolating the stage from load capacitance downstream, such as might be caused by the use of screened cable. Note that R4 is connected before R5, to prevent any loss of gain. The amount of loss is of course very small in one stage, but it can build up irritatingly in a large system if this precaution is not observed. Table 13.3 gives the noise performance.

The figure at full volume may look a bit mediocre, but results from the use of +17 dB of gain; at normal volume settings the noise output is well below −100 dBu.

TABLE 13.3 Noise performance of active gain control at various volume settings

Setting	Noise out
Zero gain	−114.5 dBu
Unity gain	−107.4 dBu
Full volume	−90.2 dBu

Low-noise Baxandall active volume stages

One of the themes of this book is the use of multiple opamps to reduce noise, exploiting the fact that noise from uncorrelated sources partially cancels. The Baxandall volume stage lends itself very well to this technique, which is demonstrated in Figure 13.24. The first version at a) improves on the basic design in Figure 13.23 by using two inverting amplifiers instead of one, and reducing the value of the pot from 10 kΩ to 5 kΩ. Bear in mind that the latter change reduces the input impedance proportionally, and it can fall to low values (1.2 kΩ in this case) with the volume control at maximum. The outputs of the two inverting amplifiers are averaged by the 10 Ω resistors R4 and R7, and the noise from them partially cancels. The noise from the unity-gain buffer U1:A is not reduced because it is reproduced identically by the two inverting amplifiers. C1 and R1 prevent the input bias current of the buffer from flowing through the pot wiper and causing rustling noises. This volume control stage was used in the variable-frequency-tone-control preamplifier published in Jan Didden's *Linear Audio*, Volume 5 [5].

Note that in Figure 13.24a there is a spare opamp section floating in space. If this is not required elsewhere in the signal path, pressing it into use as another inverting amplifier will usually give more noise reduction than doubling-up the buffer.

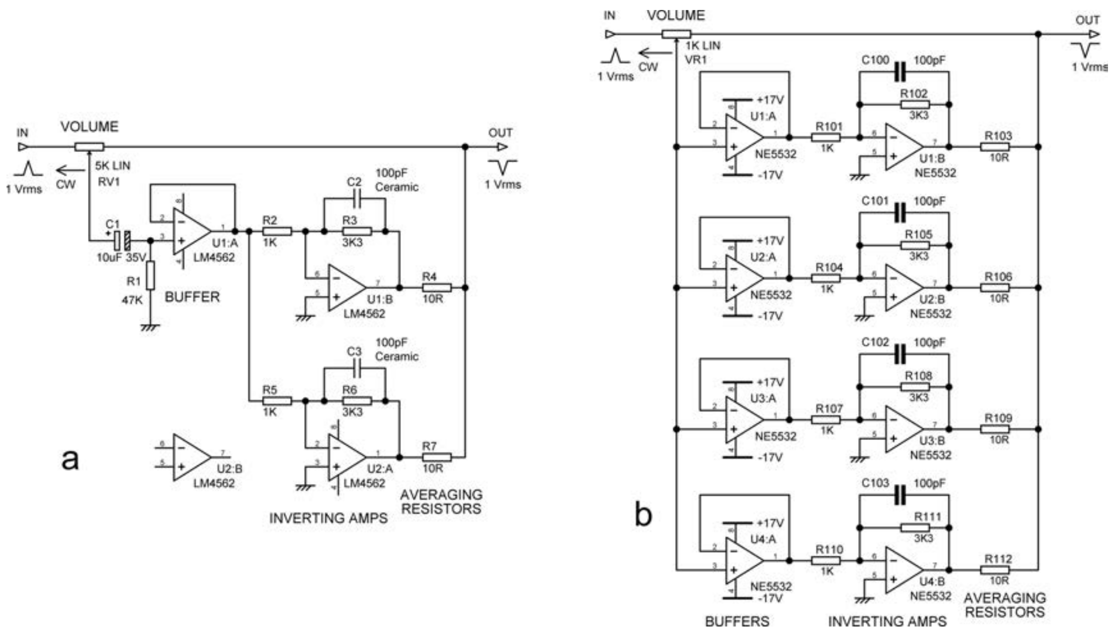

Figure 13.24: Baxandall active volume controls using multiple opamps to reduce noise: a) single buffer, dual inverting amps and 5 kΩ pot, b) quad buffer, quad inverting amps and 1 kΩ pot. Maximum gain +10 dB in both cases

TABLE 13.4 Measured noise output from the volume control stages in Figure 13.24

Volume setting (mark)	Single-amp control (+10 dB max) (dBu)	Dual-amp control (Fig. 13.24a) (dBu)	Quad-amp control (Fig. 13.24b) (dBu)
10	−107.3 dBu	−109.0 dBu	−112.1 dBu
9	−109.4 dBu	−110.3 dBu	−114.1 dBu
8	−111.1 dBu	−112.4 dBu	−115.7 dBu
7	−112.7 dBu	−113.7 dBu	−117.1 dBu
6	−113.8 dBu	−114.7 dBu	−118.1 dBu
5	−115.0 dBu	−116.9 dBu	−119.4 dBu
4	−116.1 dBu	−115.7 dBu	−120.8 dBu
3	−117.1 dBu	−118.1 dBu	−122.1 dBu
2	−117.9 dBu	−118.4 dBu	−123.0 dBu
1	−118.8 dBu	−119.8 dBu	−124.2 dBu
0	−119.8 dBu	−121.4 dBu	−126.1 dBu

Note: Readings corrected by subtracting −119.2 dBu testgear noise. Bandwidth 22 Hz−22 kHz, rms sensing, unweighted.

Figure 13.24b shows a more sophisticated design for still lower noise, using four inverting amplifiers and four separate buffers. There is nothing to be gained by averaging the buffer outputs before applying the signal to the inverting amplifiers, as the partial cancellation of buffer noise is carried out later by the 10 Ω resistors R103, etc. just as for the inverting amplifier noise. Four inverting amplifiers have enough drive capability to make it feasible to reduce the volume pot right down to 1 kΩ for lower noise without any distortion problems. DC-blocking components were not deemed to be required at the buffer inputs, because the bias currents are flowing through low resistances. This stage was used in my Elektor Preamplifier 2012 design [6], and, as for all the stages in this preamplifier, a really serious attempt was made to make the noise as low as was reasonably practical. There are no great technical difficulties in using an even lower pot value, such as 500 Ω, but there are sourcing problems with dual gang pots of less than 1 kΩ.

The measured noise performance for a single–inverting amplifier stage (Figure 13.23 with maximum gain reduced to +10 dB) and the stages in Figure 13.24 are summarised in Table 13.4. While the reduction in noise on adding amplifiers is not perhaps very dramatic, it is as bullet-proof as any electronic procedure can be.

The Baxandall volume control: loading effects

With circuits like the Baxandall volume stage that are not wholly obvious in their operation, it pays to keep a wary eye on all the loading conditions. We will take the dual amplifier

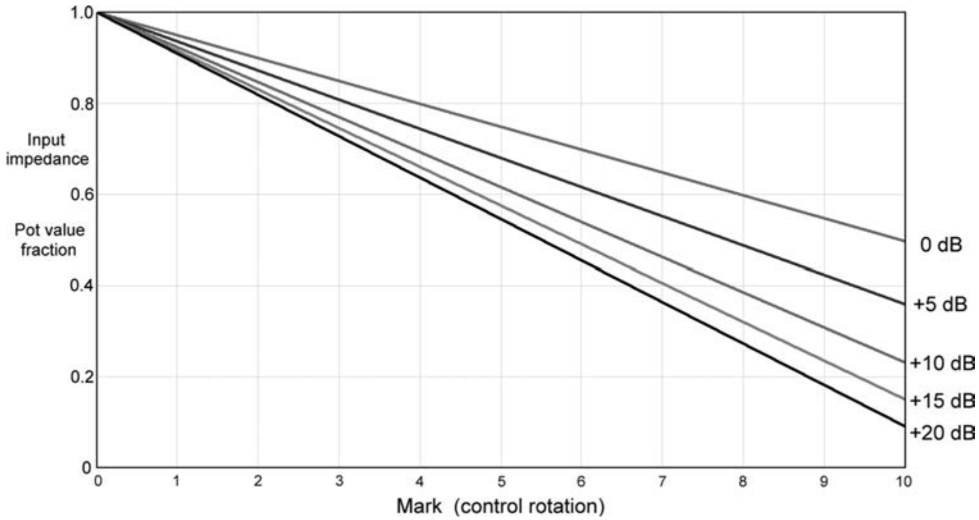

Figure 13.25: The input impedance of the volume control stage as a proportion of the pot track resistance falls more rapidly when the stage is configured for higher maximum gains

volume control stage Figure 13.24a as an example. There are three loading conditions to consider:

Firstly, the input impedance of the stage. This varies from the whole pot track resistance at Mk 0, to a fraction of this at Mk10, that fraction being determined by the maximum gain of the stage. It falls proportionally with control rotation as the volume setting is increased. Figure 13.25 shows how the minimum input impedance becomes a smaller proportion of the track resistance as the maximum gain increases. With a maximum gain of +10 dB the minimum input impedance is 0.23 times the track resistance, which for a 5 kΩ pot gives 1.2 kΩ. If the preceding stage is based on a 5532 or an LM4562 it will have no trouble at all in driving this load.

Secondly, the loading on the buffer stage U1:A. A consequence of the gain of the two inverting amplifiers U1:B, U2:A stage is that the signals handled by the unity-gain buffer U1:A are never very large; less than 3 Vrms if output clipping is avoided. This means that R2, R3 and R5, R6 can all be kept low in value to reduce noise without placing an excessive load on unity-gain buffer U1:A, which would cause increased distortion at high levels.

Thirdly, the loading on the inverting stages U1:B, U2:A. At Mk 10 the loading is a substantial fraction of the value of the pot, but it gets heavier as volume is reduced, as demonstrated in the rightmost column of Table 13.5. We note thankfully that the loading stays at a reasonable level over the mid-volume settings. Only when we get down to a setting of Mk 1 does the load get down to a slightly worrying 383 Ω; however, at this setting the attenuation is −21.6 dB, so even a maximum input of 10 Vrms would only give an output of 830 mV. We also have two opamps

TABLE 13.5 Volume stage gain, noise, input impedance, and gain stage loading versus control setting

Control position (mark)	Gain (dB)	Noise output (dBu)	Input impedance Ω	Opamp load Ω
10	+10.37	−109.0	1162	3900
9	+6.98	−110.3	1547	3456
8	+4.03	−111.4	1929	3069
7	+1.29	−112.7	–	2687
6	−1.38	−113.7	–	2305
5	−4.11	−114.9	3083	1918
4	−7.07	−115.7	–	1534
3	−10.48	−117.1	–	1151
2	−14.83	−118.4	–	767
1	−21.61	−119.8	–	383
0	−infinity	−121.4	5000	–

Note: Corrected for AP noise at −119.2 dBu. Measurement bandwidth 22 Hz−22 kHz, rms sensing, unweighted.

in parallel to drive the load, so the opamp output currents are actually quite small. Note that the loading considered here is only that of the pot on the inverting stages. The inverting opamps also have to drive their own feedback resistors R3 and R6, which are effectively grounded at the other end, and this should be taken into account when working out the total loading. U1:B, U2:A also have to drive whatever load is connected to the volume control stage output; it is likely to be the final stage in the preamplifier and may be connected directly to the outside world.

Table 13.5 shows the gain and the noise output at the various control settings.

I haven't bothered to fill in all the entries for input impedance, as it simply changes proportionally with control setting as seen in Figure 13.25, ranging from the minimum of 1162 Ω to 5 kΩ, the resistance of the pot track.

Going back to the loading on the inverting opamps at low volume, we have 383 Ω at Mk 1, which is 766 Ω per opamp and no cause for alarm. However, I have heard doubts expressed about a possible rise in distortion at very low volume settings below this, because the inverting amplifiers U1:B, U2:A then see even lower load impedances. The impedances may be low, but the current to be absorbed by the inverting stages is actually very limited because almost the whole of the pot track is in series with the input at low settings. To prove there is not a problem here, I set the volume to Mk 1 and pumped 20 Vrms in, getting 1.6 Vrms out. The THD residual was indistinguishable from the GenMon output of the AP SYS-2702. In use the input cannot exceed 10 Vrms as it comes from an opamp.

To push things further, I set the volume to Mk 0.2, (i.e. only 2% off the endstop) and shoved 20 Vrms in to get only 300 mVrms out. The THD+N residual was 0.0007%, composed entirely of noise with no trace of distortion. I then replaced the 4562s in the U1:B, U2:A positions with Texas 5532s (often considered the worst make for distortion) and the results were just the same except the noise level was a bit higher giving a THD+N of 0.0008%. There is not a problem here.

Baxandall active volume stage plus passive control

One of the few disadvantages of the Baxandall volume control is that, unlike passive volume controls, the noise output is not absolutely zero when the volume is set to zero. Whatever the configuration, at least the voltage noise of the opamps will appear at the output. This limitation applies to all active volume controls. The only other real drawback of the Baxandall control is the control law, which as we saw earlier, is rather too flat over its upper and middle ranges when configured for the popular maximum gain of +10 dB. The law cannot be bent around directly without introducing pot-dependence.

Both of these problems can be solved by putting a passive linear volume control after the active volume stage, with the two pots ganged together. This obviously means a four-gang control but the benefits are worth it. The noise level is now really zero at Mk 0 because the wiper of the linear pot is connected to ground.

Furthermore, the combination of the linear law with that of the Baxandall control gives a steeper characteristic which approximates to a linear-in-decibels line with a range of −32 to +10 = 42 dB, which is much more usable than the 28 dB line of the Baxandall control alone (see Figure 13.26).

This arrangement is also free from pot-dependence, so the gain depends only on the angular setting of the ganged control. However, this will no longer hold if the wiper of the linear pot is significantly loaded. In practice a high-input-impedance buffer stage will be required, and this compromises the 'zero-noise at zero volume' property. On the other hand, a unity-gain buffer opamp is working in good conditions for low noise, having no feedback resistors to introduce Johnson noise or convert opamp current noise to voltage noise. The source impedance for the buffer is usually non-zero, as it is fed from the wiper of the linear pot, but this can be made low in value as the multiple inverting amplifiers in the Baxandall stage give plenty of drive capability. The linear pot does not have to be directly after the active volume control; intermediate stages could be inserted if required, giving a distributed volume control.

A practical version of this arrangement is shown in Figure 13.27, where both pots are 2 kΩ. Making them different in value would introduce some unwelcome complications in component sourcing, but such a custom component could be specified if there was a powerful reason to do so (which, as far as I can see, there isn't).

In this case a 3-path Baxandall control is used, which has in itself an intermediate noise performance between the 2-path and 4-path versions in Table 13.5 above. The only downside to this technique is the expense of the four-gang pot.

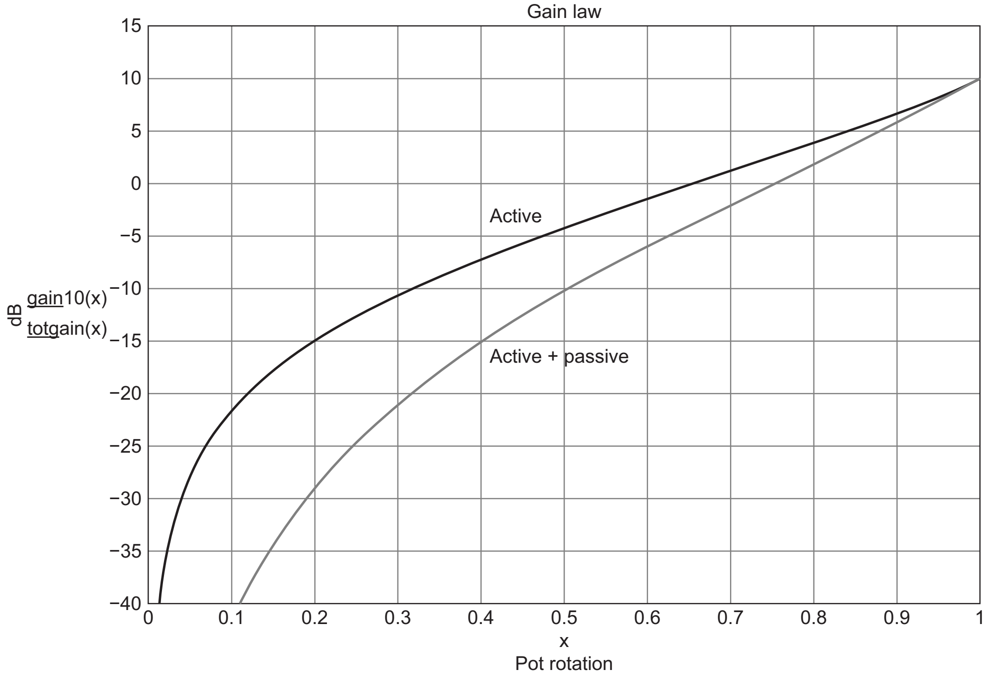

Figure 13.26: Baxandall active volume and active + passive control laws

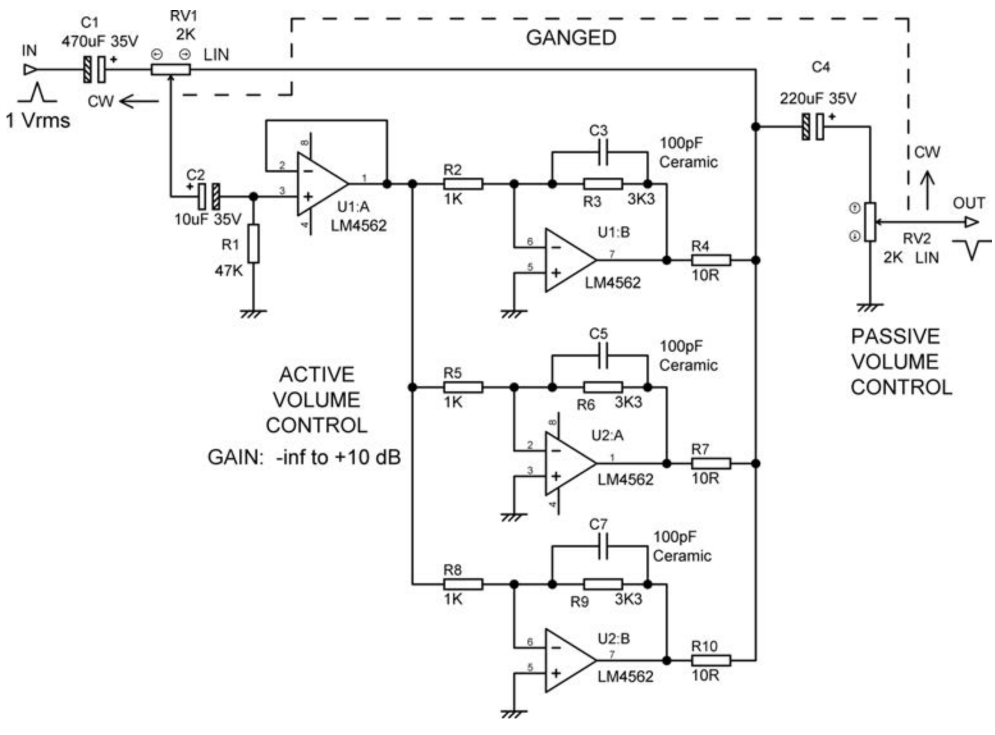

Figure 13.27: Baxandall active volume controls followed by linear passive attenuator

Potentiometers and DC

As noted in the previous section, it is never a good idea to have DC flowing through any part of a potentiometer. If there is a DC voltage between the ends of the pot track there will be rustling noises generated as the wiper moves up and down over minor irregularities in track resistance.

Feeding a bias current through a wiper to the next stage tends to create more serious noise because the variations in wiper contact resistance are greater. This tends to get worse as the track surface becomes worn. This practice is often acceptable for FET input opamps like the TL072 but it is definitely not a good idea for bipolar opamps such as the 5532 because the bias current is much greater and so therefore is the noise on wiper movement. AC coupling is essential when using bipolar opamps. If you are using electrolytic capacitors then make sure that the coupling time-constant is long enough for capacitor distortion to be avoided (see Chapter 2).

Motorised potentiometers

Motorised pots are simply ordinary pots driven by an attached electric motor, heavily geared down and connected to the control shaft through a silicone slipping clutch. This clutch allows manual adjustment of the volume when the motor is off, and prevents the motor stalling when the pot hits the end of its travel; limit switches are not normally used. Motorised pots are now considerably cheaper than they used to be, due to the manufacture of components in China that represent a very sincere homage to designs by ALPS and others, and they appear in lower to middle-range integrated amplifiers. Motorisation can be added to any control that uses a rotary pot.

In many ways they are the ideal way to implement a remote-controlled volume function. There is no variable gain electronics to add noise and distortion, manual control is always available if you can't find the remote (so-called because it is never to hand) and the volume setting is inherently non-volatile, as the knob stays where it was left when you switch off.

A disadvantage is that the 'feel' of a motorised pot is pretty certain to be worse than a normal control, because of the need for the slipping clutch between the control shaft and the motor; a large diameter weighted knob helps with this. The channel balance is, of course, no better than if the same pot was used as a manual control.

Since the motor has to be able to run in either direction, and it is simplest to run it from a single supply, an H-bridge configuration (as shown in Figure 13.28) is used to drive it. Normally all four transistors are off. To run the motor in one direction Q1 and Q4 are turned on; to run in the other direction Q2 and Q3 are turned on. The H-bridge and associated logic to interface with a microcontroller can be obtained in convenient ICs such as the BA6218

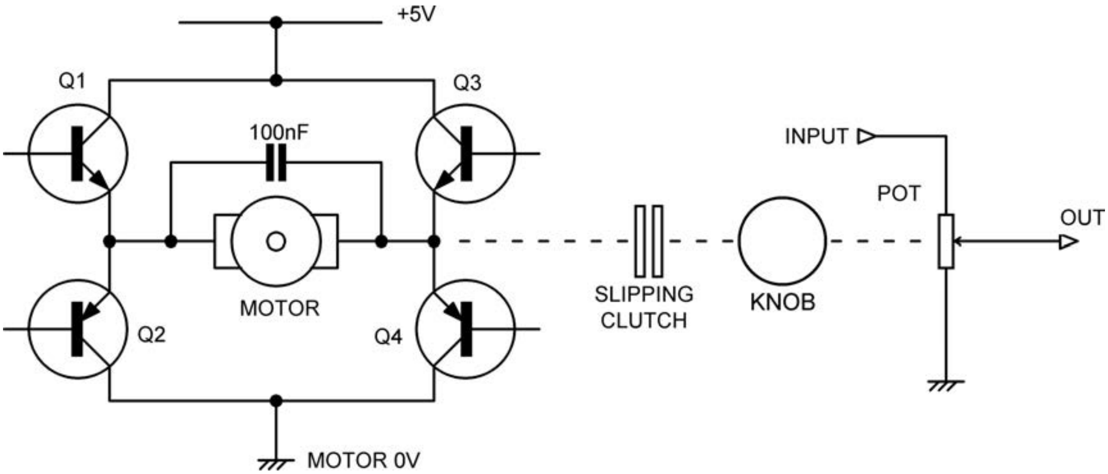

Figure 13.28: Control circuitry for a motorised volume control

by Rohm. This IC can supply an output current of up to 700 mA. Two logic inputs allow four output modes: forward, reverse, idle (all H-bridge transistors off) and dynamic braking (motor shorted via ground). The logic section prevents input combinations that would turn on all four devices in the H-bridge and create (briefly) electronic mayhem.

The absence of limit switches means that if continuous rotation is commanded (for example, by sitting on the remote) there is the potential for the motor to overheat. It is a wise precaution to write the software so that the motor is never energised for longer than it takes to get from one extreme of rotation to the other, plus a suitable safety margin.

It would appear that there might be problems with electrical noise from the motor getting into the audio circuitry, but I have not myself found this to be a problem. In the usual version the motor is screened with a layer of what appears to be GOSS (grain-oriented silicon steel) to keep magnetic effects under control, and the motor terminals are a long way from the audio terminals. A 100 nF capacitor across the motor terminals, and as close to them as practicable, is always a good idea. The motor should be driven from a separate non-audio supply unless you're really looking for trouble; motorised controls with 5 V motors are popular as they can run off the same +5 V rail as a housekeeping microcontroller. I have never had problems with motor noise interfering with a microcontroller.

Linear faders, as used on mixing consoles, are also sometimes motorised, not for remote control, but to allow previously stored fader movements to be played back in an automatic mixdown system. A linear servo track next to the audio track allows accurate positioning of the fader. Motorisation has usually been done by adding a small electric motor to one end of the fader, and moving the control knob through a mechanism of string and pulleys that is strongly reminiscent of an old radio dial. Such arrangements have not always been as reliable as one might have hoped.

Stepped volume controls

The great feature of potentiometer-based volume controls is that they have effectively infinite resolution, so you can set exactly the level you require. If, however, you are prepared to forego this and accept a volume control that changes in steps, a good number of new possibilities open up and, in return, promise much greater law accuracy and channel balance. The technologies available include rotary-switched resistive attenuators, relay-switched resistive attenuators, switched multi-tap transformers, and specialised volume-control ICs. These options are examined below.

The obvious question is: how small a step is needed to give satisfactory control? If the steps are made small enough, say less than 0.2 dB, they are imperceptible (always providing there are no switching transients) but there are powerful economic reasons for not using more steps than necessary. 2 dB steps are widely considered acceptable for in-car entertainment (implemented by an IC) but my view is that serious hifi requires 1 dB steps.

Switched attenuator volume controls

For high-end products where the imperfections of a ganged-potentiometer volume control are not acceptable, much superior accuracy can be achieved by using switched attenuators to control level. It is well-known that the BBC for many years used rotary faders that were stud-switched attenuators working in 2 dB steps; some of these were still in use in 1961.

The normal practice is to have a large rotary switch that selects a tap on a resistor ladder; since the ladder can be made of 1% tolerance components the channel matching is much better than that of the common dual-gang pot. A stereo control naturally requires two resistor ladders and a two-pole switch. The snag is of course the much greater cost; this depends to a large extent on how many control steps are used. The resistor ladders are not too costly, unless exotic super-precision parts are used, but two-pole switches with many ways are neither cheap nor easy to obtain.

At the time of writing, one commercial preamp offers twelve 5 dB steps; the component cost has been kept down by using separate switches for left and right channels, which is not exactly a triumph of ergonomics. Another commercial product has an 11-position ganged switched attenuator. In my opinion neither offers anything like enough steps.

The largest switches readily available are made up from 1 pole 12-way wafers. The most common version has a break-before-make action, but this causes clicky transients due to the interruption of the audio waveform. Make-before-break versions can usually be obtained and these are much more satisfactory as the level changes as the switch is rotated are very much smaller and the transients correspondingly less obtrusive. Make sure no circuitry gets overloaded when the switch is bridging two contacts.

Moving beyond 12-way, a relatively popular component for this sort of thing is a 24-position switch from the ELMA 04 range. A bit of care is needed in selecting the part required as one version has 10 μm of silver on the contacts with a protective layer of only 0.2 μm thick gold. This very thin layer is for protection during storage and transport only and in use will wear off quite quickly, exposing the silver, which is then subject to tarnishing and the production of non-conductive silver sulphide; this version should be avoided unless used in a sealed environment. Other versions of these switches have thick 3 μm gold on the contacts which is much more satisfactory. They can be obtained with one, two or four 24-way wafers, but they are not cheap. At the time of writing the first edition of this book (January 2009), two-wafer versions were being advertised by third parties at about $130 each, which meant that in almost all cases the volume switch would cost a lot more than the rest of the preamplifier put together.

If 24 positions are not enough (which in my view is the case) 48-way switches are available from ELMA and Seiden, though in practice they have to be operated as 47-way because the mechanical stop takes up one position. This stop is essential, as a sudden transition from silence to maximum volume is rarely a good idea. 58-way switches are also available from Seiden.

A switched attenuator can be made very low-impedance to minimise its own Johnson noise and the effect of the current noise of the following stage. The limiting factors are that the attenuator input must present a load that can be driven with low distortion from the preceding stage, and that the resistor values at the bottom of the ladder do not become too small for convenience. An impedance of around 1000 Ω from top to bottom is a reasonable choice. This means that the highest output impedance, (which is at the −6 dB setting, a very high setting for a volume control) will be only 250 Ω. This has a Johnson noise of only −128.2 dBu (22 kHz bandwidth, 25 °C), and is unlikely to contribute much to the noise output of a system. The choice of around 1000 Ω does assume that your chosen range of resistors goes down to 1 Ω rather than stopping at 10 Ω, though if necessary the lower values can of course be obtained by paralleling resistors.

Assuming you want to use a 12-way switch, a possible approach is to use 12 steps of 4 dB each, covering the range 0 to −44 dB. My view is that such steps are too large, and 2 dB is much more usable, but you may not want to spend a fortune on a rotary switch with more ways. In fact, 1 dB steps are really required so you can always get exactly the volume you want, but implementing this with a single rotary switch is going to be very difficult, as it

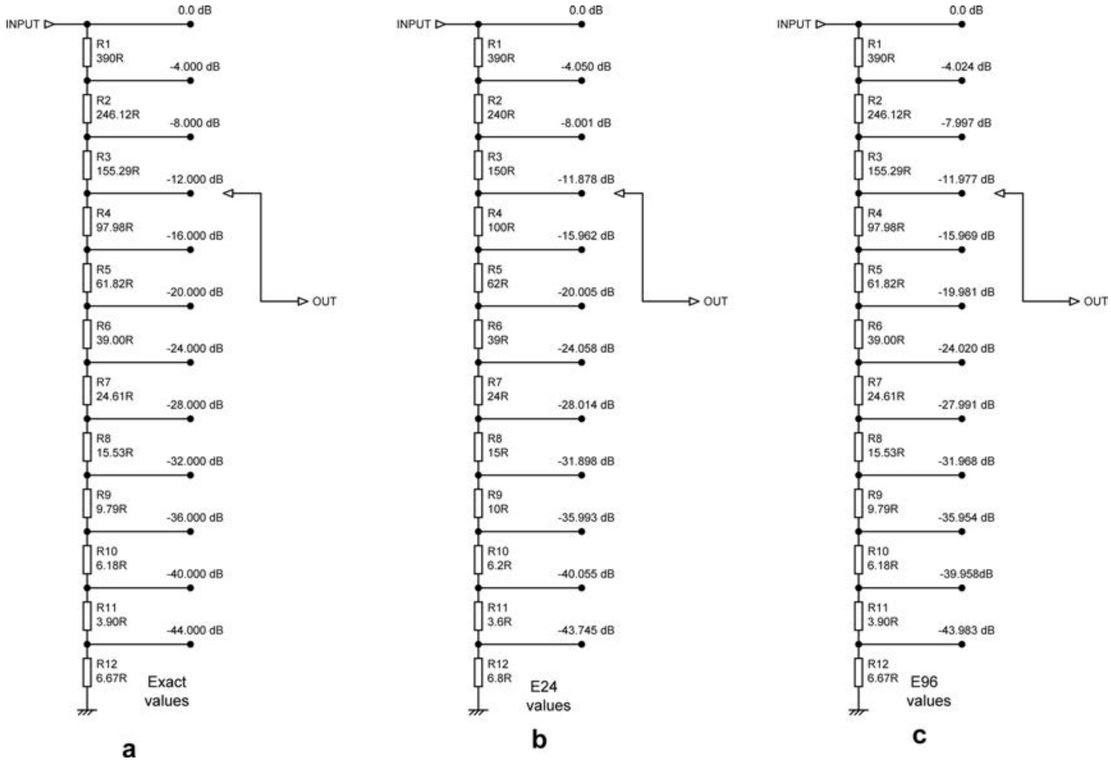

Figure 13.29: A 12-way switched attenuator volume control: a) gives the theoretical resistor values, b) is the best accuracy with E24 resistor values, and c) shows how the step errors are much reduced by using E96 values

implies something like a 60- or 70-way switch. My current preamplifier just happens to be a relay-switched Cambridge Audio 840E with some 90-odd 1 dB steps.

There is no reason why the steps have to be equal in size, and it could be argued that the steps should be smaller around the centre of the range and larger at the top and bottom, to give more resolution where the volume control is most likely to be used. It does not really matter if the steps are not exactly equal – the vital thing is that they should be identical for the left and right channels to avoid image shift.

Figure 13.29a shows a typical switched attenuator of this sort, with theoretically exact resistor values; the size of each step is 4.00000 dB, accurate to five decimal places. In view of resistor tolerances, such accuracy may appear pretty pointless, apart from the warm feeling it produces, but it costs nothing to start off with more accuracy than you need. The total resistance of the ladder is 1057 Ω, which is a very reasonable load on a preceding stage, so long as it is implemented with a 5532, LM4562, or similar.

A lot depends on what range of resistor values you have access to. When I began to design preamplifiers, E12 resistors (12 values per decade) were the norm, and E24 resistors (24 values per decade) were rather rare and expensive. This is no longer the case, and E24 is freely available. Figure 13.29b shows the same attenuator with the nearest E24 value used for each resistor. The attenuation at each tap is still very accurate, the error never exceeding 0.12 dB except for the last tap which is 0.25 dB high; this could be cured by making R12 a parallel combination of 6.8 Ω and 330 Ω and making R11 3.9 Ω, which reduces the last-tap error to 0.08 dB. To reduce the effect of tolerances R12 would be better made from two near-equal components; 12 Ω in parallel with 15 Ω is only 0.060% high in value.

The next most prolific resistor range is E96, with no less than 96 values in a decade – nobody seems interested in making an E48 range as such, though it is of course just a sub-set of E96. Using the nearest E96 value in the attenuator, we get Figure 13.29c, where the taps are accurate to within 0.04 dB; remember that this assumes that each resistor is exactly the value it should be and does not incorporate any tolerances. The improvement in accuracy is not enormous, and if you can get a tighter tolerance in the E24 range than E96, E24 is the preferred option.

There is also an E192 resistor range, but it is rather rare, and there seems to be no pressing need to use so many different values in volume control attenuators.

Almost all the work in the design of a switched attenuator is the calculation of the resistor values in the ladder. Putting in likely component values and attempting to tweak them by hand is a most unpromising approach, because all the values interact and you will boil off your sanity. A systematic spreadsheet approach is the only way. This is probably the simplest:

1. Decide the approximate total resistance of the resistor ladder.
2. Decide the step size, say 4 dB.
3. Take a two resistor potential divider, where the sum of the resistors equals the desired total resistance, and choose exact values for a 4 dB attenuation; use the goal-seek tool to get the exact value for the bottom resistor (an exact E24 value at the top of the ladder is not essential, but it is convenient).
4. Now split the bottom resistor into two, so that another 4 dB attenuator results. Check that the attenuation really is correct, because an error will propagate through the rest of the process, and you will have to go back and do it all again from that point. Repeat this step until you have enough resistors in the ladder for the number of taps required.
5. When the table of resistor values is complete, pick the nearest value for each resistor from the E-series you are using. Alternatively select a parallel pair to get closer to the desired value, keeping the two resistors as near-equal as possible.

You will then have constructed something like the spreadsheet shown in Table 13.6, which gives the E24 resistor values shown in Figure 13.29b. The chosen value at the top of the ladder is 390 Ω. The spreadsheet has been set up to give more information than just the resistor value and the tap attenuation; it gives the voltage at each tap for a given input (10 V in this case) in column 5, the output impedance of each tap in column 6, the step size in dB

TABLE 13.6 A spreadsheet that gives all the relevant information about a divider ladder for a switched attenuator. Here a 12-way switch gives 4 dB steps, using single E24 resistors

1	2	3	4	5	6	7	8	9	10
Nearest E24 value					Divider current =	0.0096	Amps rms		
		INPUT =	10	Vrms	Total res =	1046.6			
		dB ref =	0						
					Tap Z		Step	Tap error	Tap No.
	(Ω)	ratio	(dB)	(O/P V)	(Ω)	Power (mW)	(dB)	(dB)	1
R1	390					35.6			
		0.62736	−4.050	6.2736	244.672		4.0496	−0.0496	2
R2	240					21.9			
		0.39805	−8.001	3.9805	250.77		3.952	−0.0012	3
R3	150					13.7			
		0.25473	−11.878	2.5473	198.69		3.877	0.1216	4
R4	100					9.1			
		0.15918	−15.962	1.5918	140.08		4.084	0.0379	5
R5	62					5.7			
		0.09994	−20.005	0.9994	94.15		4.043	−0.0050	6
R6	39					3.6			
		0.06268	−24.058	0.6268	61.49		4.053	−0.0575	7
R7	24					2.2			
		0.03975	−28.014	0.3975	39.95		3.956	−0.0137	8
R8	15					1.4			
		0.02542	−31.898	0.2542	25.92		3.884	0.1020	9
R9	10					0.9			
		0.01586	−35.993	0.1586	16.34		4.095	0.0065	10
R10	6.2					0.6			
		0.00994	−40.055	0.0994	10.30		4.061	−0.0549	11
R11	3.6					0.3			
		0.00650	−43.745	0.0650	6.76		3.690	0.2546	12
R12	6.8					0.6			

in column 8, and the absolute error for each tap in column 9. It also gives the total resistance of the ladder, the current through it for the specified input voltage, and the resulting power dissipation in each resistor in column 7. The last parameter is unlikely to be a major concern in an audio attenuator, but if you're working with very low impedances to minimise noise, it is worth keeping an eye on.

Before the design work begins you must consider the stages before and after the attenuator. It is strongly recommended that the attenuator input is driven from a very low impedance such as the output of an opamp with plenty of negative feedback, so the source impedance can be effectively considered as zero and does not enter into the calculations. The loading on the output of the attenuator is more of a problem. You can either take loading on the output into account in the calculations, in which case the shunting effect of the load must be incorporated into step 3 above, or else make the input impedance of the next stage so high that it has a negligible effect.

As an example, the E24 network in Figure 13.29b was calculated with no allowance for loading on the output. Its highest output impedance is 253 Ω at Tap 3, so this is the worst case for both loading-sensitivity and noise. If a load of 100 kΩ is added, the level at this tap is only pulled down by 0.022 dB. A 100 kΩ input impedance for a following stage is easy to arrange, so the extra computation required in allowing for loading is probably not worthwhile unless for some very good reason the loading is much heavier. The Johnson noise of 253 Ω is still only −128.2 dBu.

If you feel you can afford 24-way switches, then there is rather more flexibility in design. You could cover from 0 to −46 dB in 2 dB steps, 0 to −57.5 dB in 2.5 dB steps, or 0 to −69 dB in 3 dB steps. There are infinite possibilities for adopting varying step sizes.

Table 13.7 shows the resistor values for 0 to −46 dB in 2 dB steps, Table 13.8 gives those for 0 to −57.5 dB in 2.5 dB steps, and Table 13.9 gives the values for 0 to −69 dB in 3 dB steps. All three tables give the nearest E24 and E96 values, and the resulting errors. Note the first two versions start off with a 220 Ω resistor at the top, but on moving from 2 dB to 2.5 dB steps, the resistors towards the bottom naturally get smaller to give the greater attenuation required, and the total ladder resistance falls from 1060 Ω to 866 Ω. To prevent the resistor values becoming inconveniently small and the total resistance too low, the 3 dB step version starts off with a higher value resistor of 430 Ω at the top; this increases the total resistance of the ladder to 1470 Ω, and raises the maximum output impedance to 368 Ω at Tap 3. The Johnson noise of 368 Ω is naturally higher at −126.5 dBu, but this is still very low compared with the likely noise from amplifier stages downstream.

The exact values on the left hand of each table can be scaled to give the total divider resistance required, but it will then be necessary to select the nearest E24 or E96 value

TABLE 13.7 Resistor values and accuracy for a 2 dB step switched attenuator with a 24-way switch

		Exact		Using E24 values			Using E96 values		
Tap		(Ω)	Step (dB)	(Ω)	Step (dB)	Error (dB)	(Ω)	Step (dB)	Error (dB)
1	R1	220.0000	0.0000	220	0.0000	0.0000	220	0.0000	0.0000
2	R2	174.7490	−2.0000	180	−2.0435	−0.0435	174	−2.0208	−0.0208
3	R3	138.8080	−4.0000	130	−4.1427	−0.1427	137	−4.0114	−0.0114
4	R4	110.2591	−6.0000	110	−6.0660	−0.0660	110	−5.9942	0.0058
5	R5	87.5819	−8.0000	82	−8.1104	−0.1104	86.6	−7.9967	0.0033
6	R6	69.5688	−10.0000	68	−10.0251	−0.0251	69.8	−9.9799	0.0201
7	R7	55.2605	−12.0000	56	−12.0124	−0.0124	54.9	−11.9914	0.0086
8	R8	43.8950	−14.0000	43	−14.0788	−0.0788	44.2	−13.9838	0.0162
9	R9	34.8670	−16.0000	33	−16.0850	−0.0850	34.8	−16.0046	−0.0046
10	R10	27.6958	−18.0000	27	−18.0166	−0.0166	27.4	−18.0107	−0.0107
11	R11	22.0000	−20.0000	22	−19.9959	0.0041	22.0	−19.9987	0.0013
12	R12	17.4749	−22.0000	18	−22.0272	−0.0272	17.4	−22.0077	−0.0077
13	R13	13.8808	−24.0000	13	−24.1361	−0.1361	13.7	−24.0093	−0.0093
14	R14	11.0259	−26.0000	11	−26.0577	−0.0577	11.0	−25.9916	0.0084
15	R15	8.7582	−28.0000	8.2	−28.1000	−0.1000	8.66	−27.9934	0.0066
16	R16	6.9569	−30.0001	6.8	−30.0120	−0.0120	6.98	−29.9758	0.0242
17	R17	5.5260	−32.0001	5.6	−31.9960	0.0040	5.49	−31.9863	0.0137
18	R18	4.3895	−34.0001	4.3	−34.0580	−0.0580	4.42	−33.9773	0.0227
19	R19	3.4867	−36.0001	3.3	−36.0588	−0.0588	3.48	−35.9964	0.0036
20	R20	2.7696	−38.0001	2.7	−37.9839	0.0161	2.74	−38.0004	−0.0004
21	R21	2.1999	−40.0001	2.2	−39.9548	0.0452	2.20	−39.9857	0.0143
22	R22	1.7475	−42.0000	1.8	−41.9753	0.0247	1.74	−41.9913	0.0087
23	R23	1.3881	−44.0000	1.5	−44.0701	−0.0701	1.37	−43.9886	0.0114
24	R24	5.3609	−46.0001	5.1	−46.3095	−0.3095	5.36	−45.9657	0.0343
				Total res = 1055 Ω			Total res = 1065 Ω		
				Avg abs error = 0.0654 dB			Avg abs error = 0.0117 dB		

manually. If you are increasing all the resistances by a factor of 10, then the same E24 or E96 values with a zero added can be used. The average of the absolute error for all 24 steps is shown at the bottom of each section; note that simply taking the average error would give a misleadingly optimistic result because the errors are of random sign and would partially cancel.

TABLE 13.8 Resistor values and accuracy for a 2.5 dB step switched attenuator with a 24-way switch

		Exact		Using E24 values			Using E96 values		
Tap		(Ω)	Step (dB)	(Ω)	Step (dB)	Error (dB)	(Ω)	Step (dB)	Error (dB)
1	R1	220.0000	0.0000	220	0.0000	0.0000	220	0.0000	0.0000
2	R2	164.9809	−2.4999	160	−2.5471	−0.0471	165	−2.4997	0.0003
3	R3	123.7182	−4.9999	120	−5.0208	−0.0208	124	−4.9907	0.0093
4	R4	92.7756	−7.4999	91	−7.4864	0.0136	93.1	−7.4901	0.0099
5	R5	69.5719	−10.0000	68	−9.9726	0.0274	69.8	−9.9928	0.0072
6	R6	52.1715	−12.5000	51	−12.4441	0.0559	52.3	−12.4958	0.0042
7	R7	39.1231	−15.0000	39	−14.9066	0.0934	39.2	−14.9974	0.0026
8	R8	29.3382	−17.5000	30	−17.4130	0.0870	29.4	−17.4982	0.0018
9	R9	22.0005	−20.0000	22	−19.9969	0.0031	22	−19.9995	0.0005
10	R10	16.4981	−22.5000	16	−22.5429	−0.0429	16.5	−22.4952	0.0048
11	R11	12.3718	−25.0000	12	−25.0153	−0.0153	12.4	−24.9899	0.0101
12	R12	9.2775	−27.5000	9.1	−27.4790	0.0210	9.31	−27.4890	0.0110
13	R13	6.9572	−30.0000	6.8	−29.9627	0.0373	6.98	−29.9914	0.0086
14	R14	5.2172	−32.5000	5.1	−32.4310	0.0690	5.23	−32.4939	0.0061
15	R15	3.9123	−35.0000	3.9	−34.8893	0.1107	3.92	−34.9949	0.0051
16	R16	2.9338	−37.5000	3.0	−37.3900	0.1100	2.94	−37.4948	0.0052
17	R17	2.2000	−40.0000	2.2	−39.9658	0.0342	2.2	−39.9950	0.0050
18	R18	1.6498	−42.5000	1.6	−42.5014	−0.0014	1.65	−42.4892	0.0108
19	R19	1.2372	−45.0000	1.2	−44.9601	0.0399	1.24	−44.9818	0.0182
20	R20	0.9277	−47.5001	0.91	−47.4057	0.0943	0.931	−47.4783	0.0217
21	R21	0.6957	−49.9999	0.68	−49.8653	0.1347	0.698	−49.9771	0.0229
22	R22	0.5217	−52.4999	0.51	−52.3019	0.1981	0.523	−52.4748	0.0252
23	R23	0.3912	−54.9997	0.39	−54.7183	0.2817	0.392	−54.9694	0.0306
24	R24	1.1731	−57.4997	1.20	−57.1626	0.3374	1.18	−57.4608	0.0392
				Total res = 866 Ω			Total res = 881 Ω		
				Avg abs error = 0.0816 dB			Avg abs error = 0.0113 dB		

If the available switches do not give enough steps, a possible solution is to use two rotary switches; one for coarse volume control, and the other for vernier control, the latter perhaps in 1 dB steps or less. While straightforward to design, this is not exactly user-friendly. A buffer stage is usually desirable to prevent the second attenuator from loading the first one; the second attenuator may present a constant load, but it is a heavy one. With no buffer, the

TABLE 13.9 Resistor values and accuracy for a 3 dB step switched attenuator with a 24-way switch

		Exact		Using E24 values			Using E96 values		
Tap		(Ω)	Step (dB)	(Ω)	Step (dB)	Error (dB)	(Ω)	Step (dB)	Error (dB)
1	R1	430.0000	0.0000	430	0.0000	0.0000	430	0.0000	0.0000
2	R2	304.4210	−3.0000	300	−2.9981	0.0019	301	−3.0066	−0.0066
3	R3	215.5135	−6.0000	220	−5.9437	0.0563	215	−5.9727	0.0273
4	R4	152.5719	−9.0000	150	−8.9930	0.0070	154	−8.9583	0.0417
5	R5	108.0126	−12.0000	110	−11.9281	0.0719	107	−11.9800	0.0200
6	R6	76.4671	−15.0000	75	−14.9622	0.0378	76.8	−14.9437	0.0563
7	R7	54.1346	−18.0000	56	−17.8769	0.1231	54.9	−17.9414	0.0586
8	R8	38.3243	−21.0000	39	−20.9466	0.0534	38.3	−20.9732	0.0268
9	R9	27.1315	−24.0000	27	−23.9857	0.0143	27.4	−23.9652	0.0348
10	R10	19.2077	−27.0000	20	−26.9606	0.0394	19.1	−26.9917	0.0083
11	R11	13.5980	−30.0000	13	−30.0906	−0.0906	13.7	−29.9736	0.0264
12	R12	9.6266	−33.0000	10	−32.9681	0.0319	9.53	−32.9950	0.0050
13	R13	6.8151	−36.0000	6.8	−36.0924	−0.0924	6.81	−35.9624	0.0376
14	R14	4.8248	−38.9999	4.7	−39.1209	−0.1209	4.87	−38.9497	0.0503
15	R15	3.4157	−42.0000	3.3	−42.0755	−0.0755	3.40	−41.9678	0.0322
16	R16	2.4181	−45.0000	2.4	−44.9831	0.0169	2.43	−44.9436	0.0564
17	R17	1.7119	−48.0000	1.8	−47.9476	0.0524	1.74	−47.9433	0.0567
18	R18	1.2119	−51.0001	1.2	−51.1090	−0.1090	1.21	−50.9841	0.0159
19	R19	0.8580	−54.0001	0.82	−54.1167	−0.1167	0.866	−53.9772	0.0228
20	R20	0.6074	−57.0002	0.62	−57.0034	−0.0034	0.604	−57.0068	−0.0068
21	R21	0.4300	−60.0003	0.43	−60.0776	−0.0776	0.432	−59.9949	0.0051
22	R22	0.3044	−63.0003	0.30	−63.1079	−0.1079	0.301	−63.0148	−0.0148
23	R23	0.2155	−66.0003	0.22	−66.0982	−0.0982	0.215	−65.9860	0.0140
24	R24	0.5224	−69.0003	0.51	−69.2132	−0.2132	0.523	−68.9771	0.0229
				Total res = 1473 Ω			Total res = 1470 Ω		
				Avg abs error = 0.0701 dB			Avg abs error = 0.0281 dB		

total loading of the two attenuators could also present an excessive load to the stage before the first attenuator.

A rotary switch does not of course have the smooth feel of a good potentiometer. To mitigate this, a switched volume control may need a large-diameter weighted knob, possibly with some sort of silicone damping. Sharp detents and a small knob do not a good volume control make.

Relay-switched volume controls

If you need more steps than a switch can provide, it is relatively simple to come up with a system that emulates a rotary switch with as many steps as desired, using relays, a microcontroller and an inexpensive shaft encoder; you will need a relay for each step and that will be an awful lot of relays. Another approach is to use relays in a ladder attenuator, which greatly reduces the relay count. As soon as a microcontroller is introduced, then there is of course the possibility of infra-red remote control.

The design of a ladder attenuator relay volume control is not all as simple as it may appear. If you try to use a binary system, perhaps based on an R–2R network, you will quickly find that horrible transients erupt on moving from, say, 011111 to 100000. This is because relays have an operating time that is both long by perceptual standards and somewhat unpredictable. It is necessary to use logarithmic resistor networks with some quite subtle relay timing.

Currently the highest expression of this technology is the Cambridge Audio 840A Integrated amplifier and 840E preamplifier. The latter has second generation dual-path relay volume control technology and uses only 12 relays to give 1 dB steps over most of the range. A rather unexpected feature of relay volume controls is that having purchased a very sophisticated preamplifier, where the precision relay control is the major unique feature, some customers then object to the sound of the relays clicking inside the box when they adjust the volume setting. You just can't please some people.

Transformer-tap volume controls

There are other ways of controlling volume than with resistors. At the time of writing (2013) there is at least one passive preamplifier on the market that controls volume by changing the taps on the secondary of a transformer. The unit I have in mind has only 12 volume steps, apparently each of 4 dB. These are very coarse steps for a volume control – unacceptably so, I would have thought; a 48 dB range also seems too small. 24-way switches are available (see the section on switched attenuator volume controls above) but they are expensive. There are several potential problems with this approach – transformers are well-known to fall much further short of being an ideal component than most electronic parts do. They can introduce frequency response irregularities, LF distortion, and hum. They are relatively heavy and expensive, and the need for a large number of taps on the secondary puts the price up further. The multi-way switch to select the desired tap must also be paid for.

There are however some advantages. The output impedance of a resistive potential divider varies according to its setting, being at maximum at −6 dB. Assuming it is fed from a low impedance, a transformer volume control has a low impedance at every tap, greater than zero only by the resistance of the windings, and handily lower than even a low-resistance pot. This gives lower Johnson noise and minimises the effect of the current noise of the following stage. The much-respected Sowter transformer company make a number of different volume-control transformers, of which the most representative is probably the 9335 model. This is basically a 1:1 transformer with taps on the secondary that give attenuation from 0 to −50 dB in 26 steps of 2 dB each. The DC resistance of both primary and secondary is 310 Ω. The total resistance of the windings is therefore 620 Ω, which is much lower than the value of the volume pots normally used. Note, however, that it is not much lower than the 1 kΩ pots I used in the Elektor preamplifier [6]. At the time of writing 9335s cost £132 each. You need two for stereo.

Integrated circuit volume controls

Specialised volume-control ICs based on switched networks have been around for a long time, but have only recently reached a level of development where they can be used in high-quality audio. Early versions had problems with excessively large volume steps, poor distortion performance, and nasty glitching on step changing. The best contemporary ICs are greatly improved; a modern example is the PGA2310 from Burr-Brown (Texas Instruments) which offers two independent channels with a gain range from +31.5 to −95.5 dB in 0.5 dB steps, and is spec'd at 0.0004% THD at 1 kHz for a 3 Vrms input level. The IC includes a zero crossing detection function intended to give noise-free level transitions. The concept is to change gain settings on a zero crossing of the input signal, and so minimise audible glitches. Gain control is by means of a three-wire serial interface, which is normally driven by a rotary encoder and a microcontroller such as one of the PIC series. The Yamaha YAC523 has a similar specification but incorporates seven gain-controlled channels for AV applications. Cirrus Logic make the CS3308 and CS3318, both with 8 channels.

Loudness controls

It is an awkward but inescapable feature of human hearing that its frequency response varies with the level of the sound. As the level drops the perceived loudness of high and low frequencies falls faster than the middle frequencies. This is very clearly expressed by the well-known Fletcher and Munson curves [7] shown in Figure 13.30. Their work was repeated by Robinson and Dadson [8] whose results are sometimes considered to be more accurate, and were the basis for the ISO equal-loudness curves. A great deal of information on these curves, the differences between them, and on the whole subject of loudness controls, can be found in JAES papers by Newcomb and Young [9], and Holman and Kampmann [10].

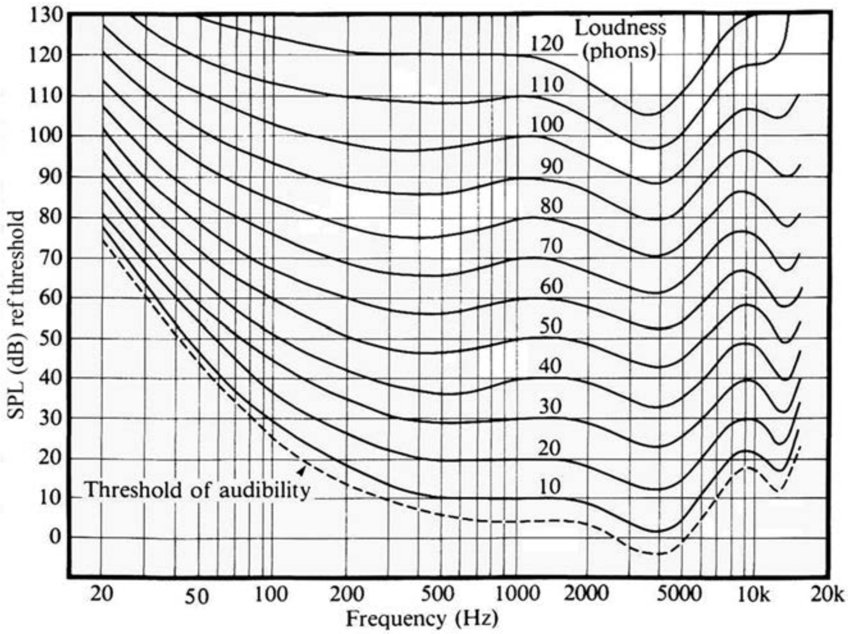

Figure 13.30: The Fletcher and Munson curves show the sound pressure level required to give a certain subjective loudness, which is measured in phons

The graph shows what sound pressure level above the threshold of hearing is needed to get a constant result in phons, the unit of subjective loudness. The first disconcerting feature is that none of the curves are remotely flat – at 20 Hz the sensitivity of the ear is always much lower than at 1 kHz, no matter what the level. A vital point is that the slopes at the bass end increase as the level falls, as shown in Column 2 of Table 13.10, which gives the average slope in dB per octave between 20 and 100 Hz. At 20 phons the slope is greater than that of a third order filter, while at 100 phons it is less than that of a second-order filter.

This increase of slope means that if you have satisfactory reproduction at a given sound level, turning down the volume gives a subjective loss of bass, demonstrated by the increased SPLs required to get the same number of phons at 20 Hz in Column 3. There are also some complicated but much smaller variations in the treble (see Column 4). This effect used to be called 'scale distortion' in the 1950s, which is a less than satisfactory term, not least because it sounds as if it would be of interest only to aquarists. At the time there was intense debate as to whether attempting to compensate for it – by making the frequency response vary with the volume control setting, turning it into what became called a 'loudness control' – was the right thing to do or not. A switch to make the volume control into a loudness control was a common feature in the 1970s, though there was still much debate as to whether it was a useful approach, with hifi purists roundly condemning the notion. Loudness controls are

TABLE 13.10 Characteristics of the Fletcher and Munson curves

1 Loudness (phons)	2 LF Slope (dB/octave)	3 SPL at 20 Hz ref 1 kHz (dB)	4 SPL at 10 kHz ref 1 kHz (dB)
100	10.7	+28	+6
90	11.6	+30	+5
80	12.5	+34	+5
70	13.4	+37	+5
60	14.7	+42	+5
50	15.1	+46	+8
40	16.8	+50	+9
30	17.7	+57	+10
20	18.6	+61	+10
10	20.3	+68	+9

currently out of fashion; if you're the kind of person that won't tolerate tone-controls you probably regard loudness controls as totally anathema. However, the wheel of fashion turns, and it is well worth looking at how they work.

Note that we are *not* trying to make the Fletcher and Munson curves flat. If we take the 100 phon level as an example, we would need some 42 dB of boost at 20 Hz, which is clearly not right. What we are trying to do is reduce the *differences* in perceived bass level with volume changes. Let us assume we have a 'flat' subjective response at a level of 100 phons, which since it roughly corresponds to 100 dB SPL which is pretty loud, typically described as 'a jackhammer at 1 metre'. Figure 13.31 shows the correction response we would need to bring the lower levels into line with the 100 phons curve. This was obtained by subtracting the 100 phon data from that for the other levels. The curves look a little wobbly as they cannot be read from the graph in Figure 13.30 with accuracy much better than 1 dB.

Clearly the correction shown above 1 kHz is neither large nor consistent. Holman and Kampmann state clearly that above 400 Hz correction is neither 'necessary or desirable' [11]. Correction below 1 kHz, however, is a different matter. Fortunately it consists of monotonic curves, though they do have varying slopes. For the higher levels, the hinge point between sloping and flat is around 600 Hz, increasing as the level falls to about 1 kHz. The slope, averaged between 20 Hz and 1 kHz, varies between 0.5 dB/octave at 90 phons to 7.3 dB/octave at 10 phons. If we assume that 10 phons is too low a level to worry about, we find the 20-phons correction curve has a slope of only 5.8 dB, which promises that we could implement a compensation network with simple first-order circuitry.

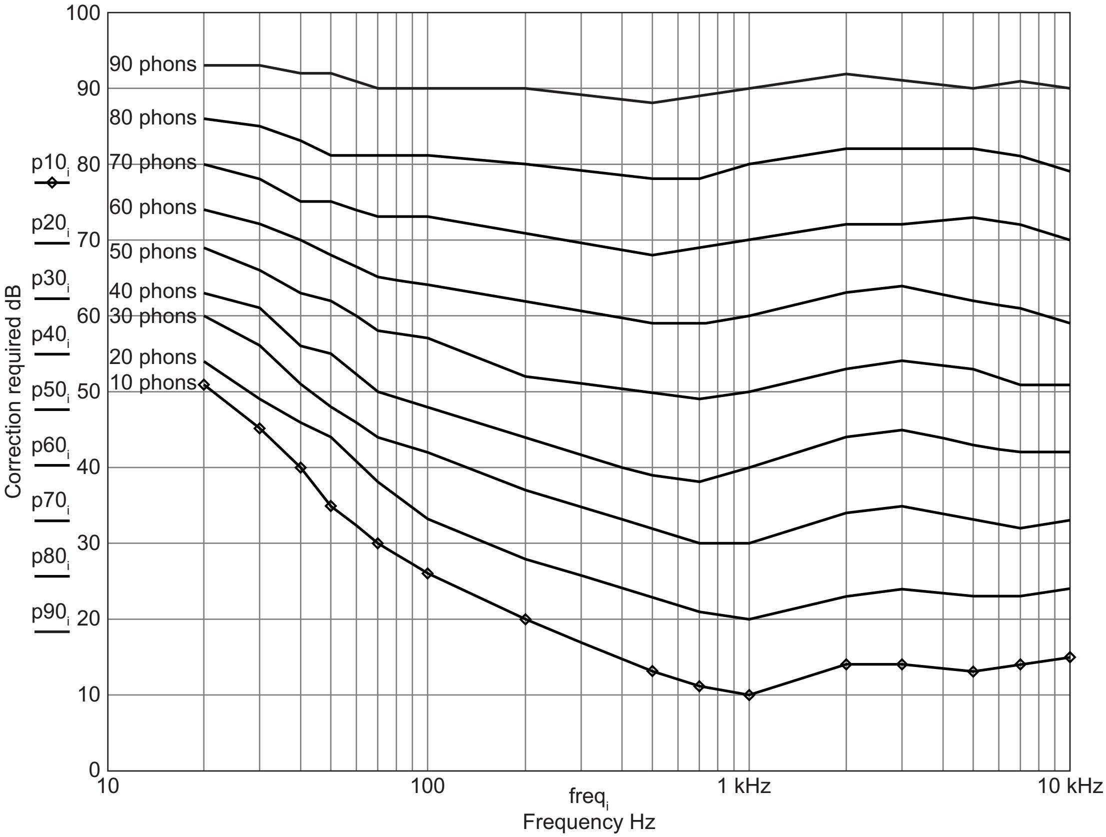

Figure 13.31: Correction curves to make the 10 to 90 phon Fletcher and Munson curves have the same response as the 100-phon curve

The Fletcher and Munson curves depend on the absolute level at the ear, and this obviously depends on amplifier gain, loudspeaker efficiency, room acoustics and many other factors. Unless you set up your system for a defined SPL which you regard as 'flat', loudness compensation is going to be a rather approximate business, and this is one argument frequently advanced against it.

Switchable loudness facilities were fitted to a large number of Japanese amplifiers in the 1970s and 80s, but European and USA manufacturers rarely used them. The circuitry used was virtually standard, with three typical versions shown in Figure 13.32. With the loudness switch set to 'OUT' the capacitors do nothing and we have a linear pot loaded down at its tap by R1, to approximate a logarithmic volume law (I assume that the pot must be linear; Japanese service manuals seem remarkably coy about the law of the pot, and only in one case have I found it confirmed as linear). When the loudness switch is set to 'IN', C1 gives HF boost and C2 gives LF boost by reducing the loading effect of R1.

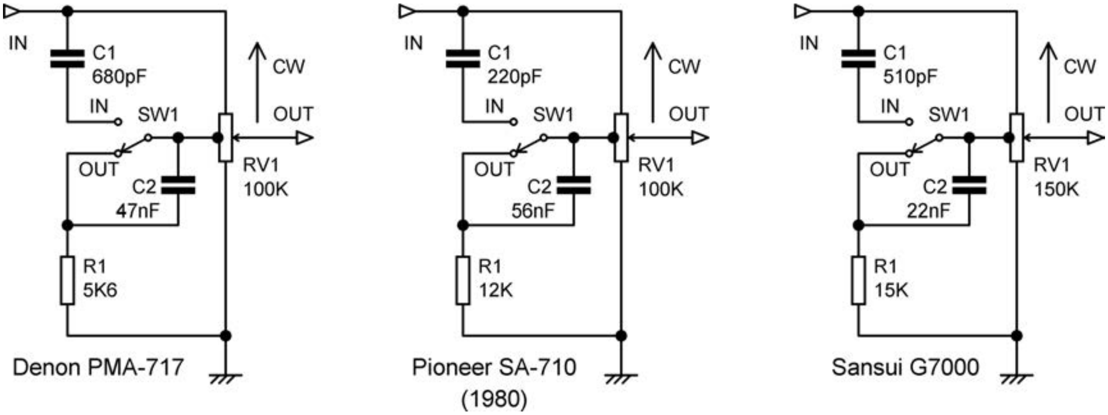

Figure 13.32: Three typical switchable loudness controls using a centre-tapped linear pot

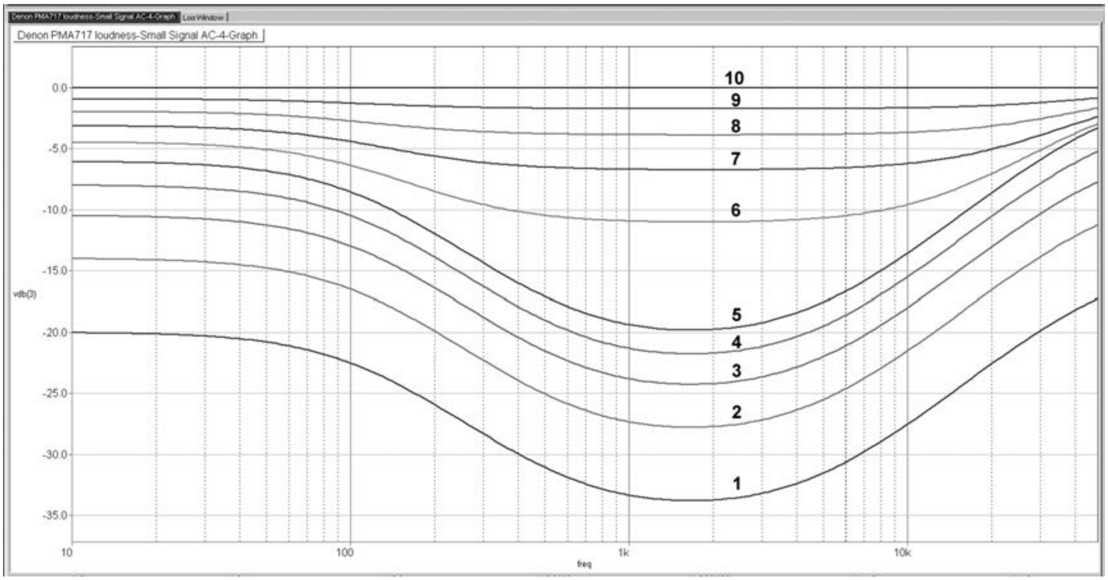

Figure 13.33: Frequency response of the Denon loudness control in Figure 13.32a. Complete pot rotation is from Marks 0–10, curve 0 not shown

The simulated frequency response for the Denon circuit is shown in Figure 13.33. There are three notable features: first, there is a lot of boost applied at HF, despite the fact that, as we saw above, no HF correction is actually wanted. Second, the correction curves only change in shape over the top half of the control rotation, the response being fixed below that. Third, there is a rather dramatic change in mid-band level between Marks 5 and 6, as the pot wiper approaches the centre-tap.

The almost universal inclusion of HF boost in these loudness circuits is a bit of a mystery. I can only assume that people took a quick look at the Fletcher and Munson curves, and saw increases at the HF end, but did not notice that the curves are essentially equally spaced, not varying in shape with level. I wonder if this unwanted HF boost was one of the reasons why loudness controls were so poorly regarded; in their heyday, the primary source was vinyl, with bad distortion characteristics at HF. It seems likely that the HF boost would have accentuated this.

A few other loudness schemes were used. The Sansui A60 and A80 amplifiers had separate loudness and volume pots. Newcomb and Young [9] put forward a very ingenious active control that compensated at LF only.

These loudness controls must represent the widest ever use of centre-tapped pots, which are otherwise rarely encountered. Pioneer used a centre-tapped pot with a simple loading resistor to ground connected to the tap, to emulate a logarithmic law; there was no loudness facility (see the Pioneer A-88X and the A-77X (1985)).

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CHAPTER 14

Balance controls

The ideal balance law

The job of a balance control on a preamplifier is to adjust the relative gain of the left and right channels, and so alter the position of the stereo image to the left or right, and do nothing else at all. As you would expect, increasing the left gain moves things to the left, and vice versa. The ideal law is therefore that of a stereo panpot in a mixing console. Two uncorrelated sound sources (and any signal will be uncorrelated once it has left the loudspeakers and bounced around the room a bit) add just like white noise sources, and so give a combined signal that is 3 dB higher, and not 6 dB higher. The sine/cosine panpot law in Figure 14.1 is therefore appropriate for the balance control of a stereo signal, because the level of the combined signals will not alter as the image is moved across the stereo stage from one side to the other (see panpots in Chapter 22). It is a constant-volume balance control.

I am aware I said, in the First Edition of this book, that the panpot law was *not* the ideal balance law, so let me clarify that. The sine/cosine law is functionally ideal in terms of its operation, but is not optimal in terms of electrical performance. Since the gain is −3 dB at the central position, there will be some compromise. If the 3 dB loss is made up by gain before the balance control, the headroom is reduced by 3 dB. If the loss is made up with 3 dB of gain after the balance control, the noise performance is likely to suffer. The only way to avoid this is to use an active balance control, which alters the gain of a stage rather than passively introducing attenuation.

A panpot in a mixing console reduces one of the signals to zero at each extreme of the pot rotation. There is no need to do this in a preamplifier. A balance control does not have to make radical changes to signal level to do its job. Introducing a channel gain imbalance of 10 dB is quite enough to shift the sound image completely to one side, so it appears to be coming from one loudspeaker only, and there is nothing at all to be gained by having the ability to fade out one channel completely.

This means that the ideal balance control is not a mixer panpot as such, but a panpot effectively limited in its rotation so that neither endstop is reached and the signal is never reduced to zero. This might be called a truncated sine/cosine law. An example is shown in

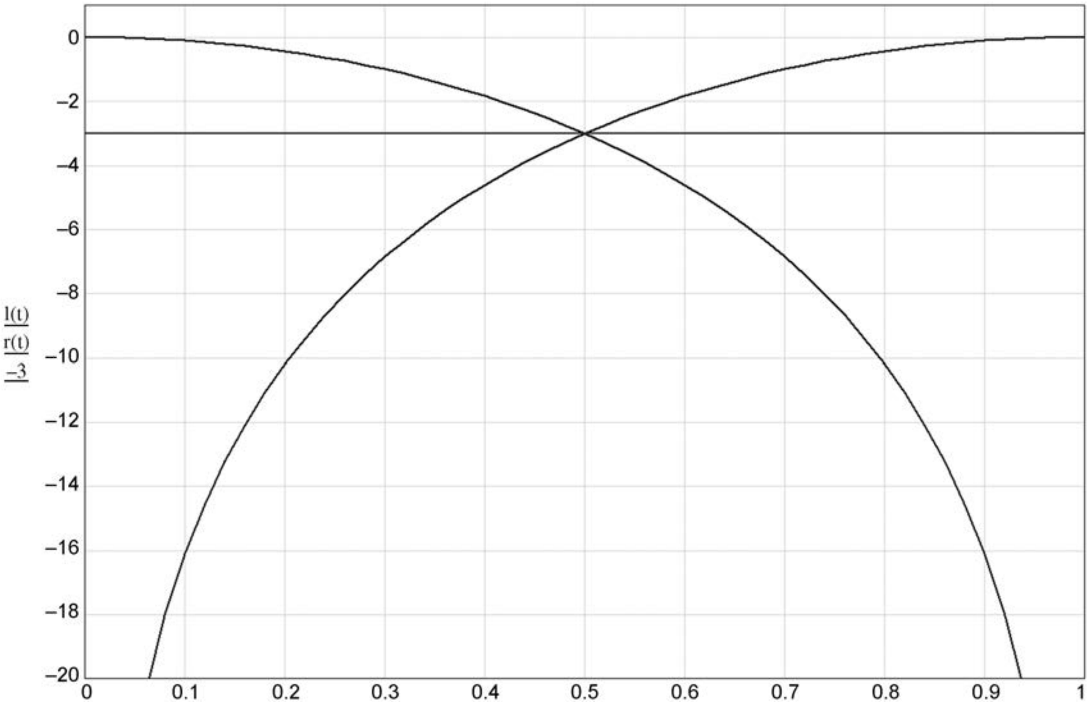

Figure 14.1: The sine/cosine law for a constant-volume stereo balance control

Figure 14.2, where the 20% of the control travel at each end is not used, and the central 60% spread out to give more precise control. As a result the maximum attenuation, with the balance control hard over, is −10.4 dB, while the minimum attenuation is −0.41 dB. The attenuation at the centre is unchanged at −3.0 dB; a central detent on a balance control pot is highly desirable.

In the case of a switched balance control, this is equivalent to using a 23-way switch (there must be an odd number to give a central position) and treating it as a 37-way switch, with the extreme seven positions at each end inaccessible. This gives a greater number of steps over the range in which the control is actually used.

Since the balance control is usually a set-and-forget function that does not require readjustment unless the listening room is rearranged, it is not often considered when controls are being motorised. In fact, getting the balance exactly right by leaping up and down between sofa and preamplifier is rather more tiresome than manual adjustment of volume.

If we look at the issue purely from the point of electrical performance rather than functionality, and ignore the desirability of a constant-volume balance control, the ideal balance law would have no attenuation when set centrally, and when moved to left or right

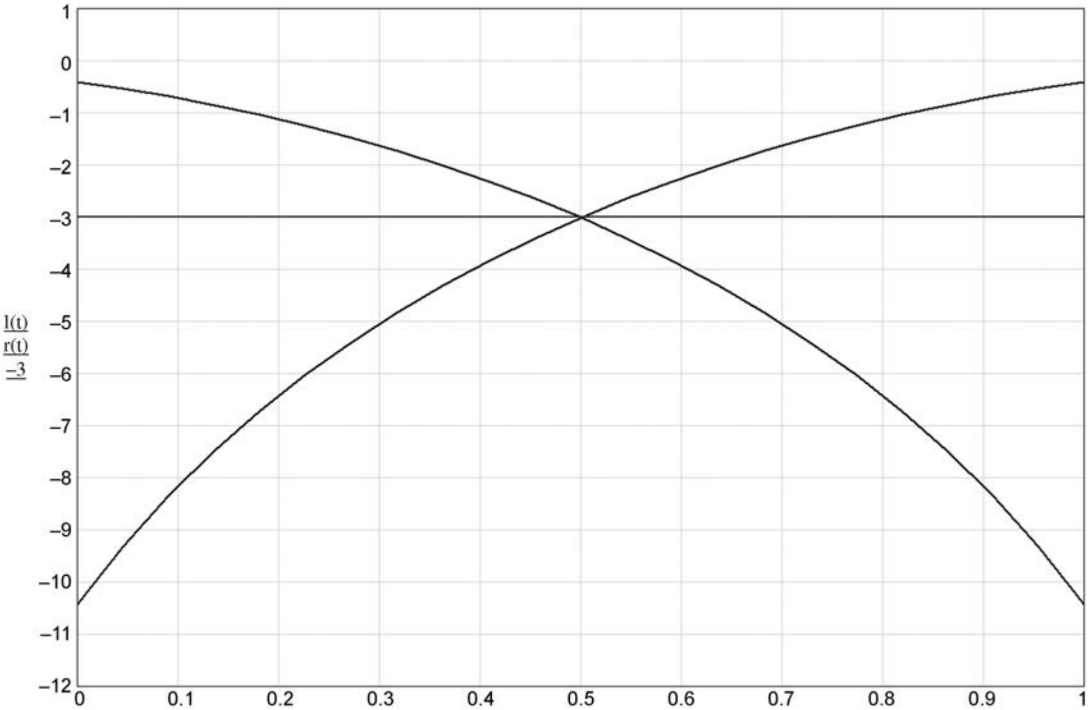

Figure 14.2: A truncated sine/cosine law for a constant-volume stereo balance control

will attenuate only one channel without affecting the gain of the other. This can be done with special balance pots that are available from several manufacturers. If there *is* attenuation at the central position then it needs to be made up by extra amplification either before or after the balance control, and this means that either the overload margin or the noise performance will be compromised to some degree. As noted earlier, the only way to avoid this is to use an active balance control, which alters gain rather than attenuation.

Balance controls: passive

Figure 14.3 shows the various forms of passive balance control; only the right channel is shown in the first three cases. Anti-clockwise movement of the control is required to introduce attenuation into the right channel and shift the sound image to the left. In Figure 14.3a a simple pot is used. If this has a linear track it will give a 6 dB loss when set centrally, and when the two uncorrelated stereo channels are rms-summed, there will be a 3 dB drop in the centre as the control is moved from hard left to hard right. See Figure 14.4, where no truncation of the balance law is used. This is a long way from being an ideal constant-volume balance control. An extra 6 dB of gain has to be built into the system to

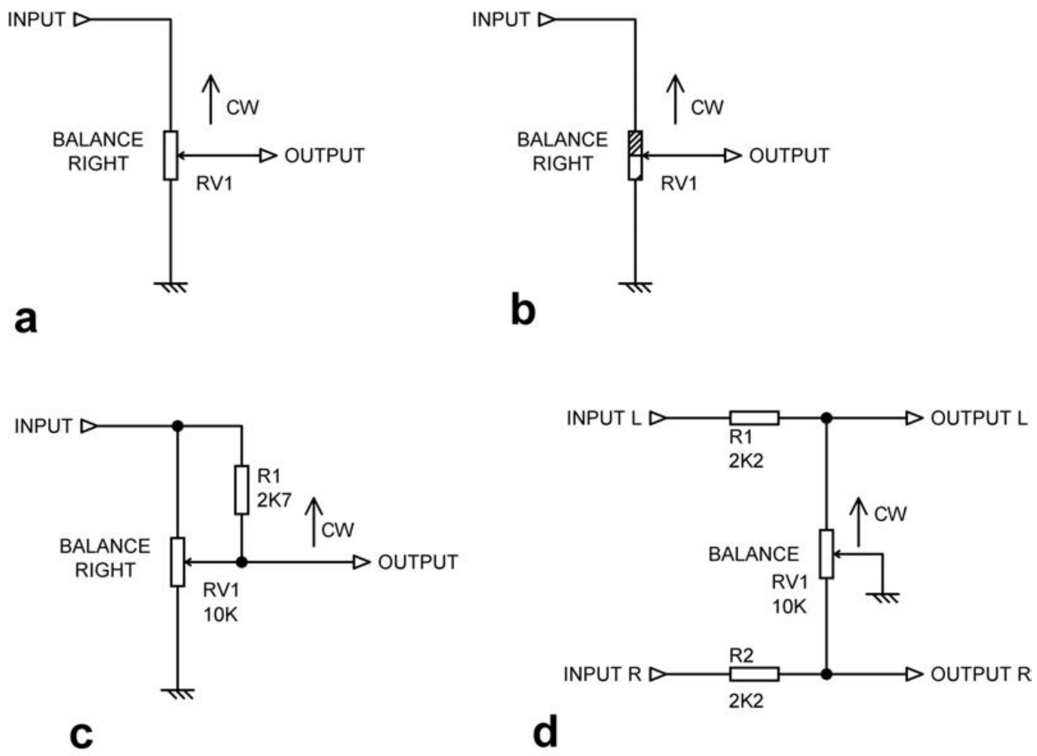

Figure 14.3: Passive balance controls

counteract the central loss, and the noise/headroom compromise is undesirable. If a log law is used for one channel, and an anti-log law for the other, the central loss can be made smaller, say 3 dB or less. Such dual-slope pots are not noted for their accuracy, so the uncertainties of log-law tolerancing mean it will not be possible to guarantee that the channel gains are identical when the control is centralised.

Special balance pots, as shown at Figure 14.3b, have half of each track, including the central position, made of a low resistance material, so that neither channel is attenuated at the central setting. On moving the control clockwise the right channel wiper stays on the low-resistance section and gain stays at 0 dB, while the left wiper moves onto the normal log/anti-log section and the signal is attenuated. Since there is no central attenuation the combined signal level falls off at each side; this is not a constant-volume balance control.

The linear control in Figure 14.3a can be much improved by the addition of a pull-up resistor, as in Figure 14.3c, which reduces the attenuation at the central position. With a 10 kΩ pot, a 3.6 kΩ pull-up resistor gives a central drop of very close to −3.0 dB, making the balance law approximately constant-volume.

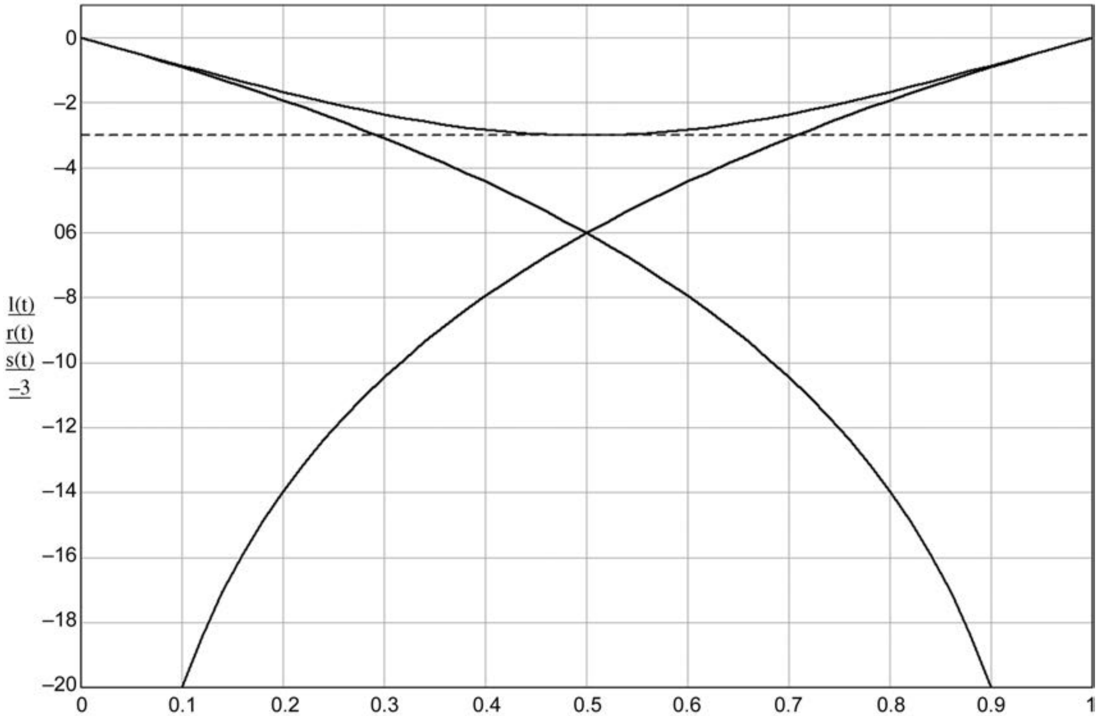

Figure 14.4: The −3 dB central drop in overall volume when a linear balance law is used

The result is seen in Figure 14.5; we have the right combined volume at the centre, but the volume is up to 0.35 dB low elsewhere. This is not likely to be perceptible under any circumstances, and we have very simply created what is for all but the most demanding applications a constant-volume balance control.

We can make things even better by reducing the pull-up resistor to 3.0 kΩ, which spreads the volume error out over both sides of the 0 dB line, reducing the worst case volume error to 0.25 dB, as in Figure 14.6. You might think this improvement is mere pedantry, but if it is possible to make things work slightly better simply by changing a resistor value, I can think of no reason not to do it.

More information on what is essentially a panpot configuration can be found in Chapter 22, though there the emphasis is more on constant volume when panning a mono signal to stereo.

The economical method in Figure 14.3d was once popular as only a single pot section is required. Unfortunately it has the unavoidable drawback that the relatively high resistance between wiper and track causes serious degradation of the interchannel crosstalk performance. If the resistance values are reduced to lower Johnson noise, the track-wiper resistance is

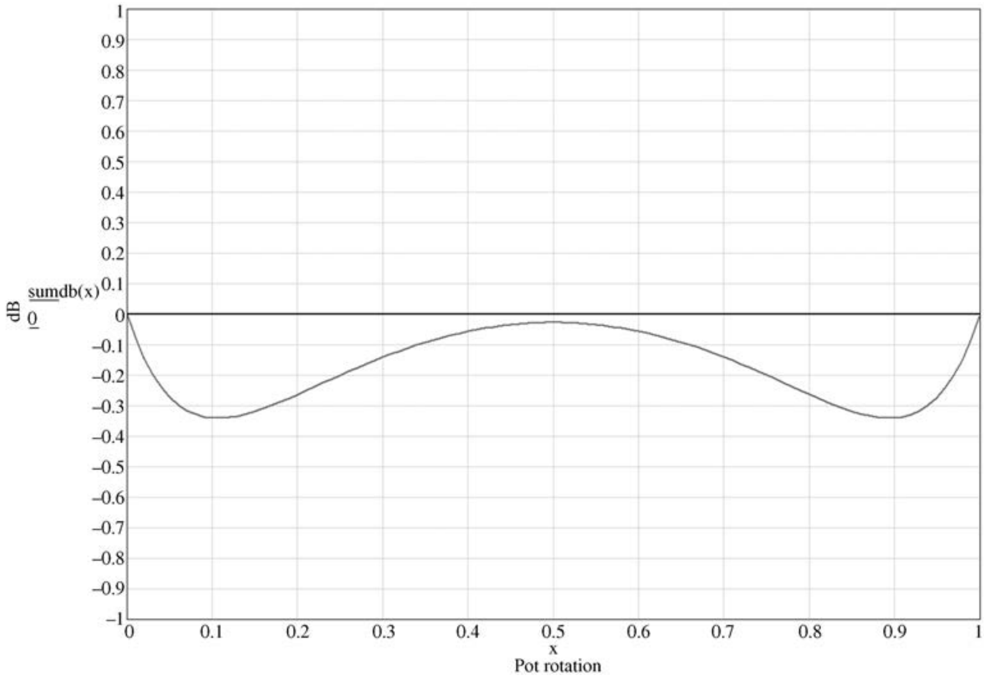

Figure 14.5: Not quite constant-volume when a linear pot with a pullup resistor is used to approximate the sine/cosine law maximum error 0.35 dB

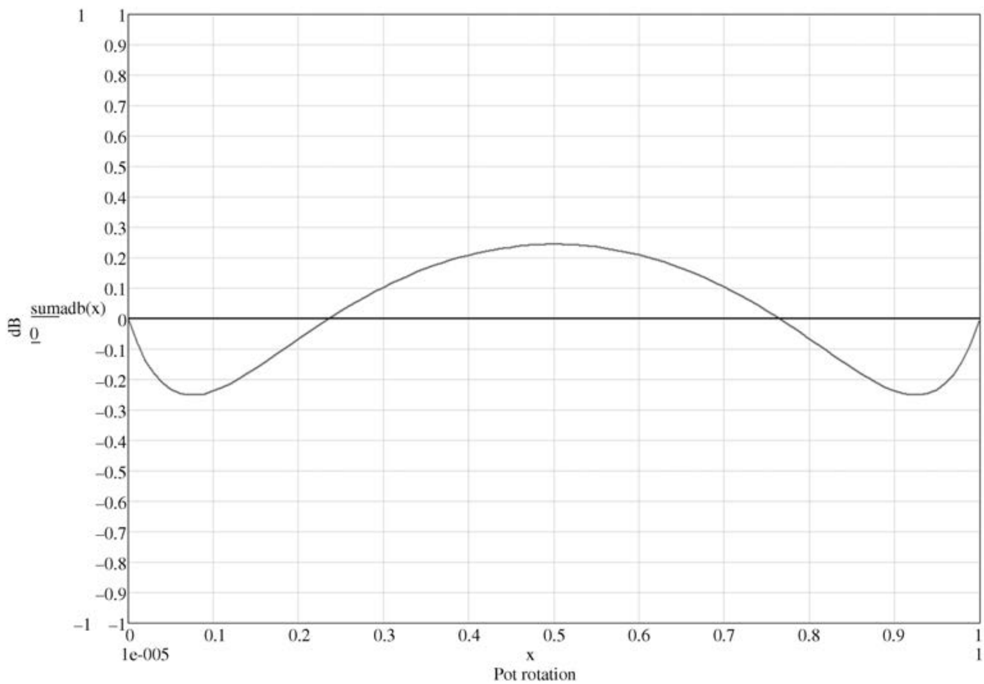

Figure 14.6: Reducing the pullup resistor value from 3.6 kΩ to 3.0 kΩ reduces the maximum deviation from constant-volume to 0.25 dB

unlikely to decrease proportionally and the crosstalk will be worsened. With the values shown, the loss for each channel with the control central is −3.2 dB. This loss can be reduced by decreasing the value of R1, R2 with respect to the pot, but this puts a correspondingly heavier load on the preceding stages when the control is well away from central.

Some preamplifier designs have attempted to evade the whole balance control problem by having separate but concentric volume knobs for left and right channels. The difficulty here is that almost all the time only the volume will require adjustment, and the balance function will be rarely used; it is therefore highly desirable that the left and right knobs are linked together in some sort of high-friction way so that the two normally move together. This introduces some awkward mechanical complications.

Balance controls: active

An active balance control is configured so that it makes a small adjustment to the gain of each channel rather than introducing attenuation, so that any noise/headroom compromise can be avoided altogether. Since all active preamplifiers have at least one gain stage, the extra complication is likely to be minimal. It is not elegant to add an extra active stage just to implement the balance function.

Figure 14.7a shows an active balance control that requires only a single-gang pot. However, it suffers from the same serious disadvantage as the passive version in Figure 14.3d; the wiper connection acts as a common impedance in the two channels and causes crosstalk. This kind of balance control cannot completely fade out one channel as it is not possible to reduce the stage gain below unity; in fact even unity cannot be achieved with this configuration because whatever the setting, there is some resistance to ground making up the lower feedback arm.

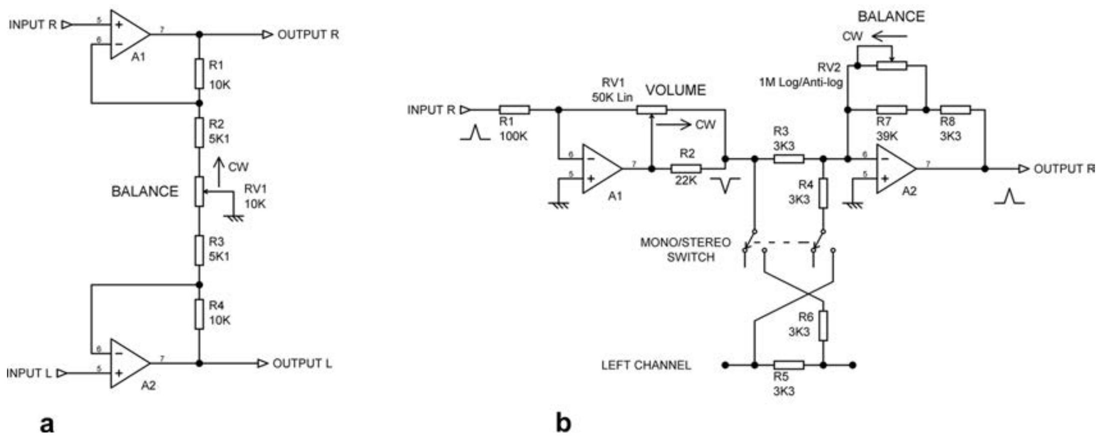

Figure 14.7: Active balance controls

With the values shown the gain for each channel with the control central is +6.0 dB. With the control fully clockwise the gain increases to +9.4 dB, and decreases to +4.4 dB with it fully anti-clockwise; the range is deliberately quite restricted. It is a characteristic of this arrangement that the gain increase on one channel is greater than the decrease on the other.

Combining balance controls with other stages

The balance control function is a very simple one; just control attenuation or gain over a rather limited range. I have always felt it is inelegant and uneconomic to design into a system a stage that does that and nothing else.

My earliest attempts in this direction were to combine the balance control with a Baxandall tone control, the balance pot varying the amount of output signal fed to the NFB side of the Baxandall network, with provision for minimising the source impedance variations. This worked quite well but did not give laboratory-instrument accuracy for the tone-control curves.

More recently I have found it expedient to combine the balance control with a balanced (differential) line input amplifier. It is a pity that there are two different uses of the word ‘balanced’ in the same circuit block, but there’s not much that can be done about it except to call it a ‘differential input’ which is less common in audio usage. There is more on this combination later in this chapter, and in Chapter 18 on line inputs.

Figure 14.7b shows an active balance control combined with an active gain control and mono/stereo switching. This configuration was used in the original Cambridge Audio P-series amplifiers; in that application A1 was a simple two-transistor inverting stage and A2 an even simpler single transistor. The left-hand section of volume control RV1 is the feedback resistance for A1, while the right hand section forms part of the input resistance to shunt stage A2, both changing to give a quasi-logarithmic law when the control is altered. The balance control RV2 is a variable resistance in the shunt feedback network of A2. The mono/stereo switch feeds the virtual-earth node of A2 with both channels via R3, R4 when in mono mode. The circuit has two inverting stages and so handily maintains the absolute phase of the output.

Switched balance controls

A balance control implemented with a pot is subject to inaccuracy due to the loose tolerances of the pot track (20%) compared with the fixed resistors around it. This does not apply to the arrangements in Figure 14.3a and Figure 14.3b because here the pot is acting as a pure potentiometer, with its output determined only by the wiper position (this assumes that the next stage puts negligible loading on the pot). If Figure 14.3a has a linear track to avoid the inaccuracies of log and anti-log tracks, there is an excessive 6 dB loss when set centrally. The accuracy of the control in Figure 14.3b depends on whether the resistive (non-metal) part of the track is linear or log.

A switched balance control, like a switched volume control, offers more than enough accuracy for even very precise audio work, and allows complete freedom in selecting the control law. As for volume controls, the drawbacks are much increased cost and a limited number of control steps. The latter is less of an issue for a balance control as it can have far fewer steps and still give all the control resolution for image position that is required. As with switched volume controls, the cost of the switch increases steeply with the number of steps. I suggest 24 steps are enough for a truly world-class balance control, and such switches are readily available. One of the switch positions must correspond to the central setting, so we actually need an odd number of switch positions. The 24-way switch is therefore mechanically stopped down to 23 positions, with position 12 being the centre. The switch should be make-before-break to minimise glitching.

Table 14.1 shows a truncated-sine law, as described above; the maximum attenuation at each extreme is −10 dB. This is shown implemented in Figure 14.8.

TABLE 14.1 Gain at each switch position for a truncated-sine law balance control

Switch position	Sine law (×)	Gain (dB)	RMS summation (dB)
23	0.9537	−0.412	0.000
22	0.9397	−0.540	0.000
21	0.9239	−0.688	0.000
20	0.9063	−0.854	0.000
19	0.8870	−1.041	0.000
18	0.8660	−1.249	0.000
17	0.8434	−1.479	0.000
16	0.8192	−1.733	0.000
15	0.7934	−2.011	0.000
14	0.7660	−2.315	0.000
13	0.7373	−2.647	0.000
12	0.7071	−3.010	0.000
11	0.6756	−3.406	0.000
10	0.6428	−3.839	0.000
9	0.6088	−4.311	0.000
8	0.5736	−4.828	0.000
7	0.5373	−5.396	0.000
6	0.5000	−6.021	0.000
5	0.4617	−6.712	0.000
4	0.4226	−7.481	0.000
3	0.3827	−8.343	0.000
2	0.3420	−9.319	0.000
1	0.3007	−10.437	0.000

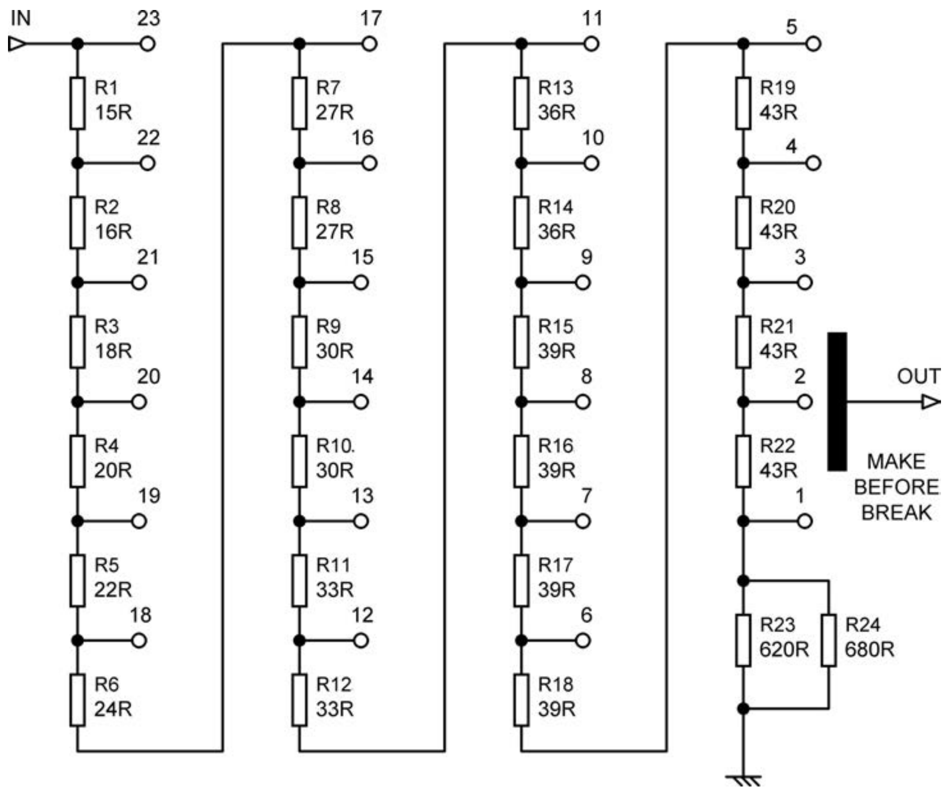

Figure 14.8: A 23-step passive constant-volume switched balance control, with make-before-break wiper contact

Mono-stereo switches

It was once commonplace for preamplifiers to have mono-stereo switches, which allowed a mono source to be played over both channels of a stereo amplifier system. Some of these were configured so that the two channels were simply joined together somewhere in the middle of the preamp stages, which was not very satisfactory unless the unused input was terminated in a low impedance to minimise noise. More sophisticated versions allowed either the left or the right input only to be routed to both outputs.

Since all modern sources are at least stereo, mono-stereo switches are now rarely if ever fitted.

Width controls

Another facility which has always been rare in preamps but is now almost unheard-of is the width control. Summing a small proportion of each channel into the other reduces the width of the sound image, and this was sometimes advocated as a small width reduction would make the image less associated with the loudspeakers, and so give a stronger illusion of acoustic reality. This of course

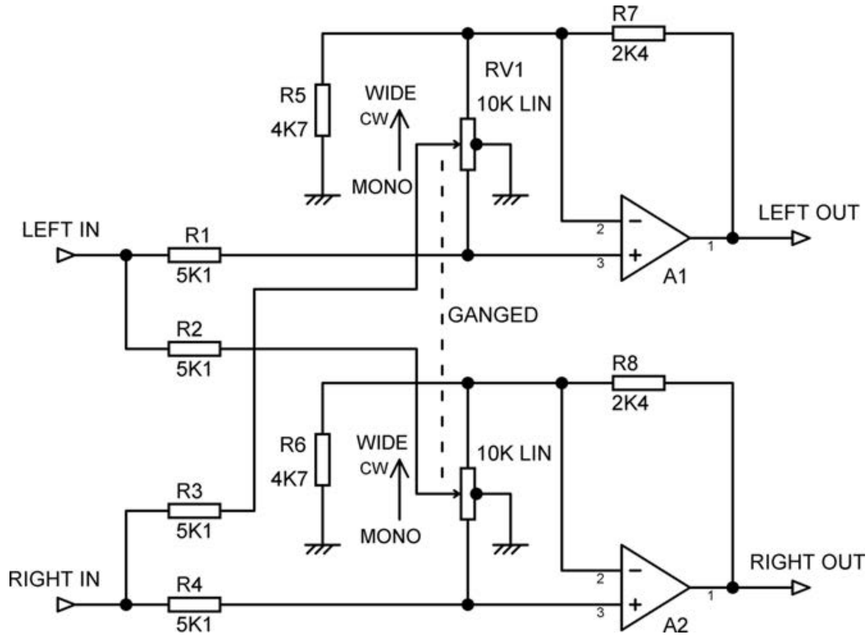

Figure 14.9: Stereo width control, with variation from mono through normal to extra-wide

runs directly counter to more contemporary views that very high levels of interchannel isolation are required to give a good stereo image. The latter is flat-out untrue; it was established long ago by the BBC in extensive testing before the introduction of stereo broadcasting, that a stereo separation of 20 to 25 dB is enough to give the impression of full image width.

By cross-feeding antiphase signals, the width of a stereo image can be increased. A famous circuit published by Mullard back in 1972 [1] gave continuous variation between mono, normal, and enhanced-width stereo. It was stated that antiphase crossfeed of greater than 24% should not be used as it would cause the sound image to come apart into two halves. Figure 14.9 shows an up-to-date version that I have used in many mixing console designs, usually on stereo input modules.

With the control central, the stereo width is unaltered, as the cross-feed through R2 and R3 is grounded via the centre-taps of the pot tracks. If there were no centre-taps we would be relying on the limited mechanical accuracy of the dual pot to minimise the crossfeed, and this would not work well. With the pot fully anti-clockwise 100% of each channel is summed with the other, giving mono. With the pot fully clockwise a fraction of each channel is summed with the other in anti-phase, increasing stereo width.

Reference

[1] Rose, M. J. (ed.). *Transistor Audio and Radio Circuits* 2nd edn (Mullard, 1972), p. 180.

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CHAPTER 15

Tone controls and equalisers

Introduction

Facilities that alter the shape of the frequency response are called tone controls when they are incorporated in hifi systems, and equalisation (or EQ) in mixing consoles.

Tone controls have suffered at the hands of fashion for some years now. It has been claimed that tone-controls cause an audible deterioration even when set to the flat position. This is usually blamed on 'phase-shift'. For a long time tone controls on a preamp damaged its chances of street (or rather sitting-room) credibility, for no good reason. A tone-control set to 'flat' – assuming it really is flat – cannot possibly contribute any extra phase-shift unless you have accidentally built in an allpass filter, which would require truly surreal levels of incompetence. Even if you managed to do it, it would still be inaudible except possibly on artificial test signals such as isolated clicks. This is well-known; most loudspeaker crossovers have an all-pass phase response, and this is considered entirely acceptable. A tone-control set to flat really is inaudible.

My view is that hifi tone controls are absolutely indispensable for correcting room acoustics, loudspeaker shortcomings, or the tonal balance of the source material, and that a lot of people are suffering sub-optimal sound as a result of this fashion. It is commonplace for audio critics to suggest that frequency-response inadequacies should be corrected by changing loudspeakers; this is an extraordinarily expensive way of avoiding tone-controls.

The equalisation sections of mixers have a rather different function, being creative rather than corrective (from now on I am going to just call it EQ). The aim is to produce a particular sound, and to this end mixer EQ is much more sophisticated than that found on most hifi preamplifiers. There will be middle controls as well as bass and treble (which in the mixing world are more often called LF and HF) and these introduce a peak or dip into the middle range of the audio band. On more complex consoles the middle frequencies are infinitely variable, and the most advanced examples have variable Q as well, to control the width of the peak or dip introduced. No one has so far suggested that mixing consoles should be built without EQ.

It is not necessary to litter these pages with equations to determine centre frequencies and so on. In each case, altering the range of frequency controlled can be done very simply by scaling the capacitor values given. If a stage gives a peaking cut/boost at 1 kHz, but you want 2.5 kHz, then simply reduce the values of all the capacitors by a factor of 2.5 times. Scaling the associated resistors instead would give the same frequency response, but may also affect noise performance, because if the resistor values are raised, Johnson noise will increase. Distortion will increase if reducing the resistor values places excessive loading on the opamps used. Another consideration is that potentiometers come in a very limited number of values, usually multiples of one, two, and five. Changing the capacitors is simpler and much more likely to be trouble-free.

Passive tone controls

For many years all tone controls were passive, simply providing frequency-selective attenuation. The famous *Radio Designer's Handbook* [1] shows that they came in a bewildering variety of forms; take a look at the chapter on 'Tone compensation and tone control' which is 42 pages long with 90 references, but does not include the Baxandall tone control. That is hidden away in an appendix at the back of the book (the final edition was published in 1953 but modern reprints are available). Some of the circuits are incredibly complex, requiring multi-section switches and tapped inductors to give quite limited tone control possibilities. Figure 15.1 shows one of the simpler arrangements [2] which is probably the best-known passive tone control configuration. It was described by Sterling [3]. though I have no idea if he originally invented it. The arrangement gives about ±18 dB of treble and bass boost and cut, the curves looking something like those of a Baxandall control but with less symmetry.

Such circuitry has several disadvantages. When set to flat it gives a loss in each network of 20.8 dB, which means a serious compromise in either noise (if the make-up gain is after the

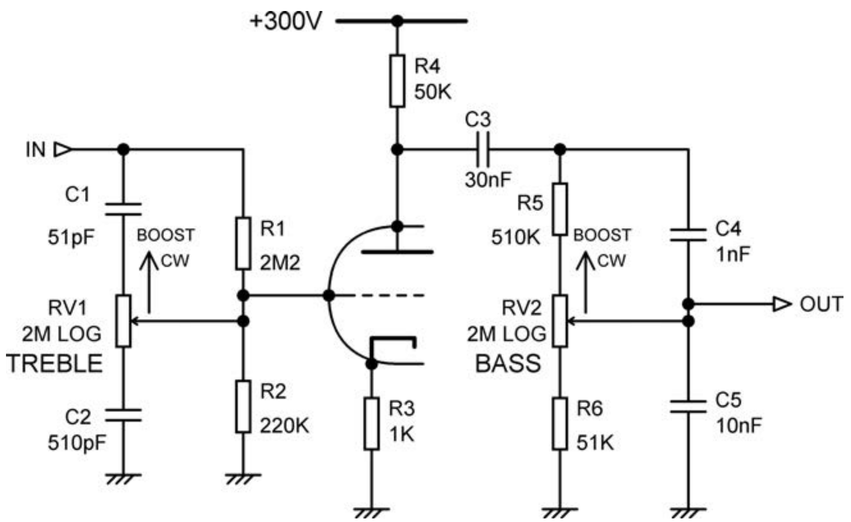

Figure 15.1: A pre-Baxandall passive tone control, with severe limitations

tone-control network) or headroom (if the make-up gain is before the tone-control network). In the days of valves, when these networks were popular, headroom may have been less of a problem, but given the generally poor linearity of valve circuitry, the increased levels probably gave rise to significantly more distortion.

Another problem is that if linear pots are used, the flat position corresponds to one-tenth of the rotation. It is therefore necessary to use log pots to get the flat setting to somewhere near the centre of control travel, and their large tolerances in law and value mean that the flat position is actually rather variable, and is unlikely to be the same for the two channels of stereo.

This circuit counts as a passive tone control because the valve in the middle of it is simply providing make-up gain for the treble network and is not in a frequency dependent feedback loop. In the published circuit there was another identical valve stage immediately after the bass network to make up the losses therein. There are some other interesting points about this valve-based circuit; it runs at a much higher impedance than solid-state versions, using 2 MΩ pots rather than 10 kΩ, and it uses a single supply rail at an intimidating 300 Volts. Circuitry running at such high impedances is very susceptible to capacitive hum pickup. There will also be a lot of Johnson noise from the high-value resistors.

Baxandall tone controls

The Baxandall bass-and-treble tone control swept all other versions before it. The original design, famously published in *Wireless World* in 1952 [4] was, in fact, rather more complex than the simplified version which has become universal. Naturally at that date it used a valve as the active component. The original schematic is rarely if ever displayed, so there it is in Figure 15.2. THD was quoted as less than 0.1% at 4 Vrms out up to 5 kHz.

The 1952 circuit specified a centre-tapped treble pot to give minimum interaction between the two controls, but such specialised components are almost as unwelcome to manufacturers as they are to home constructors, and the form of the circuit that became popular is shown in Figure 15.3, some control interaction being regarded as acceptable. The advantages of this circuit are its simplicity and its feedback operation, the latter meaning that there are no awkward compromises between noise and headroom, as there are in the passive circuit above. It also gives symmetrical cut and boost curves and is easily controlled.

It is not commonly realised that the Baxandall tone control comes in several versions. Either one or two capacitors can be used to define the bass time-constants, and the two arrangements give rather different results at the bass end. The same applies to the treble control. The original Baxandall design used two capacitors for bass and one for treble.

In the descriptions that follow, I have used the term ‘break frequency’ to indicate where the tone control begins to take action. I have defined this as the frequency where the response is ±1 dB away from flat with maximum cut or boost applied.

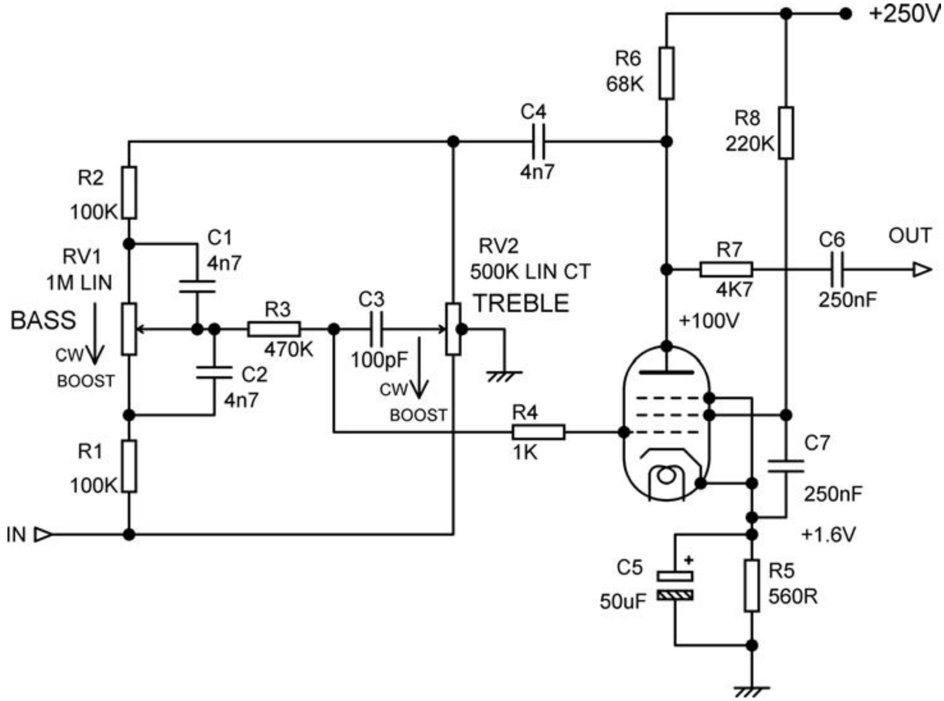

Figure 15.2: The original Baxandall tone control of 1952. The 4k7 series output resistor is to give stability with capacitive loads

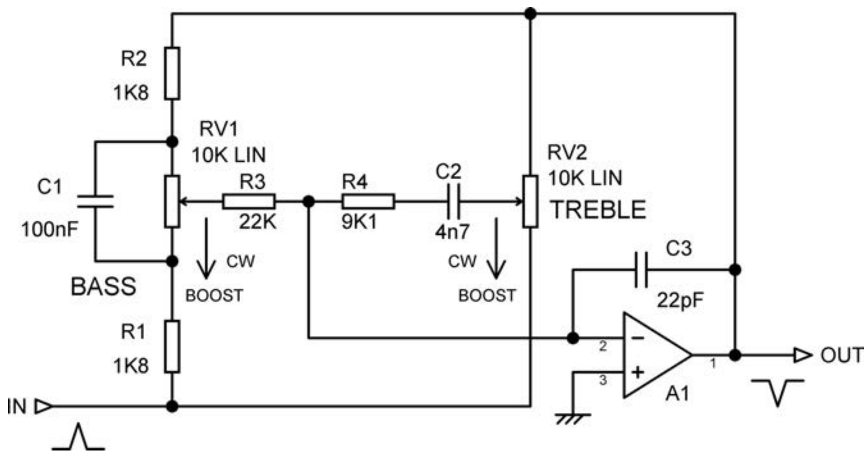

Figure 15.3: The one-capacitor Baxandall tone control

The Baxandall one-LF-capacitor tone control

This is probably the most common form of the Baxandall tone control used today, simply because it saves a capacitor or two. The circuit is shown in Figure 15.3. At high frequencies the impedance of C1 is small and the bass control RV1 is effectively shorted out and R1, R2 give unity gain. At low frequencies RV1 is active and controls the gain, ultimately over a ±16 dB range at very low frequencies. R1, R2 are end-stop resistors which set the maximum boost or cut.

At high frequencies again, C2 has a low impedance and treble control RV2 is active, with maximum boost or cut set by end-stop resistor R4; at low frequencies C2 has a high impedance and so RV2 has no effect. Resistor R3 is chosen to minimise interaction between the controls. The HF network can also be configured with two capacitors, with has some advantages; more on that later.

The one-LF-capacitor version is distinguished by its fixed LF break frequency, as shown by the bass control response in Figure 15.4. The treble control response in Figure 15.5 is similar.

In these figures the control travel is in eleven equal steps of a linear pot (central plus five steps on each side). The bass curves are ±1.0 dB at 845 Hz, while the treble curves are ±1.0 dB at 294 Hz. This sort of overlap is normal with the Baxandall configuration. Phase

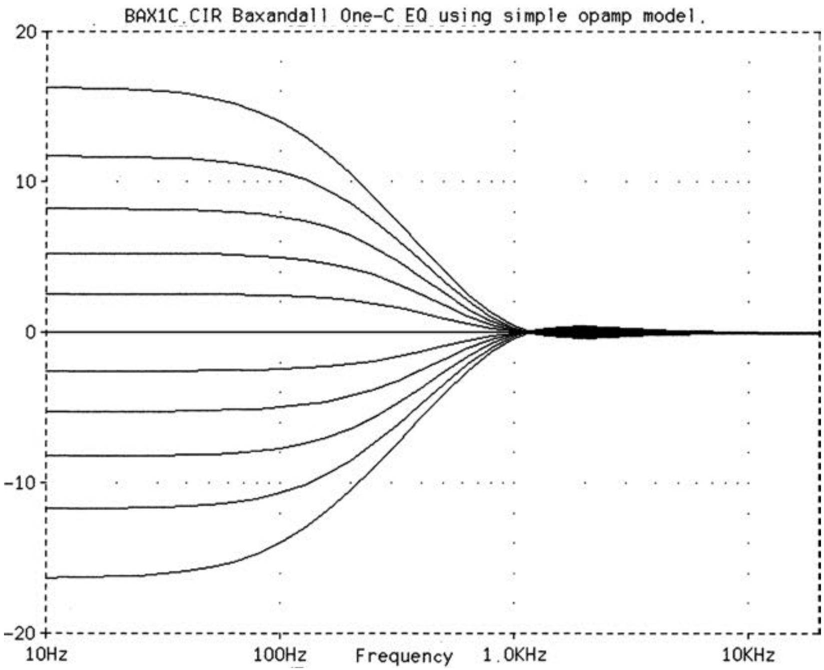

Figure 15.4: Bass control frequency responses. The effect of the LF control above the 'hinge-point' at 1 kHz is very small

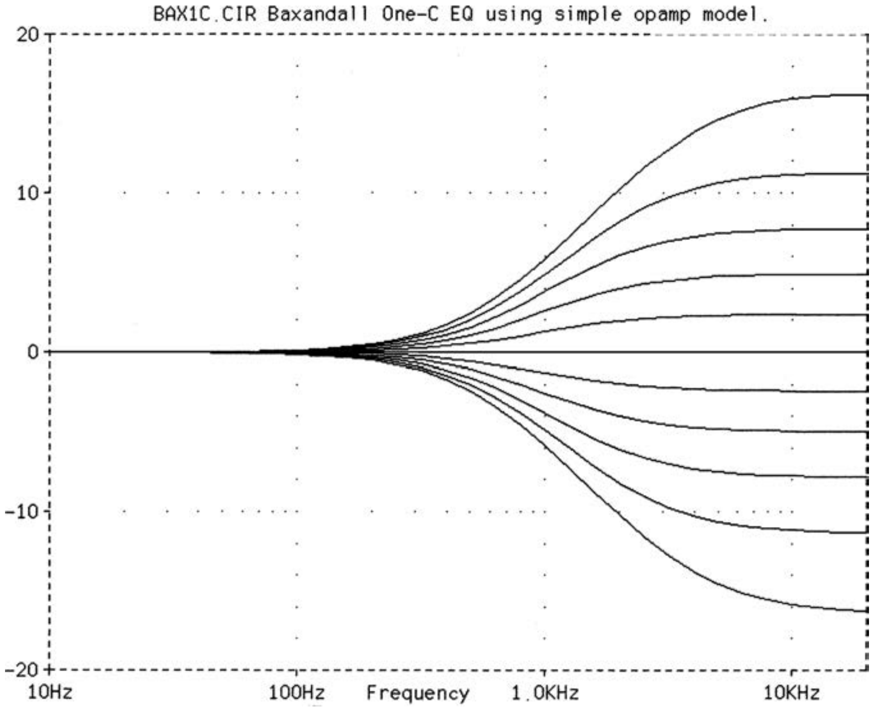

Figure 15.5: Treble control frequency responses

spikes are shown at input and output to underline that this stage phase-inverts, which can be inconvenient; phase spikes will be seen in most of the diagrams that follow in this chapter.

An HF stabilising capacitor C3 is shown connected around A1. This is sometimes required to ensure HF stability at all control settings, depending on how much stray capacitance there is in the physical layout to introduce extra phase shifts. The value required is best determined by experiment. This capacitor is not shown in most of the diagrams that follow, to keep them as uncluttered as possible, but the likely need for it should not be forgotten.

It is important to remember that the input impedance of this circuit varies both with frequency and control settings, and it can fall to rather low values.

Taking the circuit values shown in Figure 15.2, the input impedance varies with frequency as shown in Figure 15.6, for treble (HF) control settings, and as Figure 15.7 for bass (LF) control settings. With both controls central the input impedance below 100 Hz is 2.9 kΩ. At low frequencies C1, C2 have no effect and the input impedance is therefore half the LF pot resistance plus the 1k8 end-stop resistor, adding up to 6.8 kΩ. In parallel is 5 kΩ, half the impedance of the HF pot, as although this has no direct connection to the summing point (C2 being effectively open-circuit), its other end

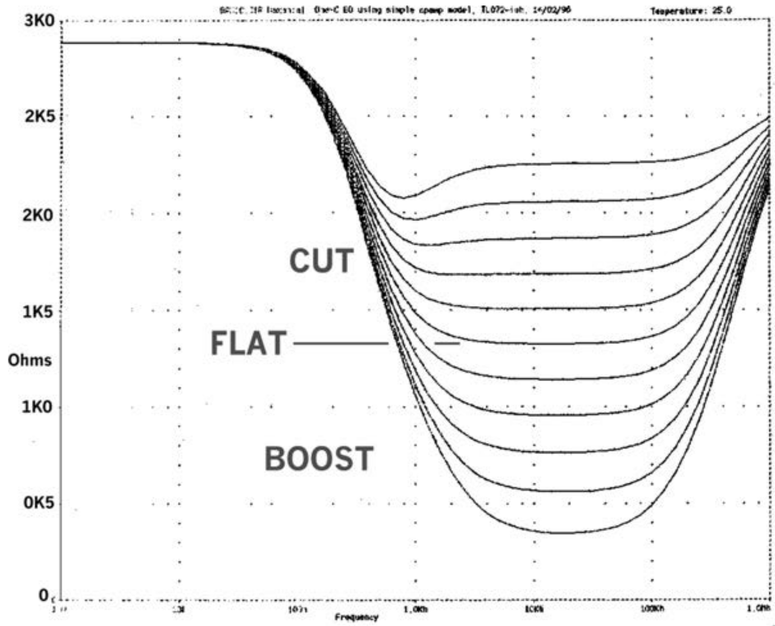

Figure 15.6: Input impedance variation with frequency, for eleven treble control settings

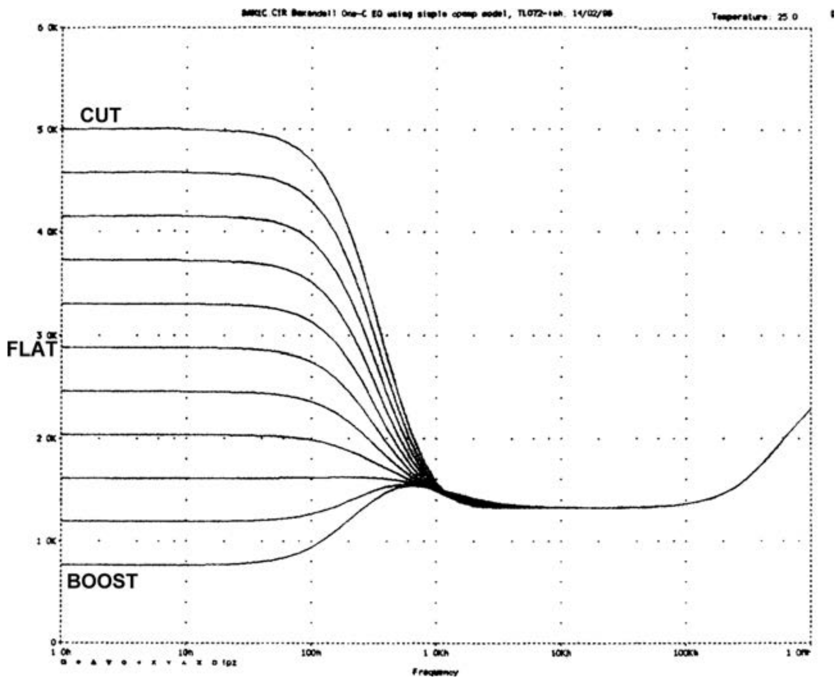

Figure 15.7: Input impedance variation with frequency, for eleven bass control settings

is connected to the output which is the input inverted. Hence the centre of the pot is approximately at virtual-earth. The parallel combination of 6.8 kΩ and 5 kΩ is 2.9 kΩ, and so this is the input impedance at LF.

At frequencies above 100 Hz (still with both controls central) the input impedance falls because C1 is now low impedance and the LF pot is shorted out. The input impedance is now 1.8 kΩ in parallel with 5 kΩ, which is 1.3 kΩ. This is already a significant loading on the previous stage, and we haven't applied boost or cut yet.

When the HF control is moved from its central position, with the LF control central, the HF impedance is higher at full cut at 2.2 kΩ. It is however much lower at 350 Ω at full boost. There are very few opamps that can give full output into such a low impedance, but this is not quite as serious a problem as at first appears. The input impedances are only low when the circuit is boosting; therefore driving the input at the full rail capability is not relevant, for if you do the output will clip long before the stage driving it. Nonetheless, opamps such as the 5532 will show increased distortion driving too heavy a load, even if the level is a long way below clipping, so it is a point to watch.

The input impedances can of course be raised by scaling the impedance of the whole circuit. For example, multiply all resistor values by four, and quarter the capacitor values to keep the frequency response the same. The downside to this is that you have doubled the Johnson noise from the resistors, and quadrupled the effect of opamp current noise, and so made the stage noisier.

When the LF control is moved from its central position, with the HF control central, the input impedance variations are similar, as shown in Figure 15.7. At full LF cut the input impedance is increased to 5.0 kΩ; at full LF boost it falls to 770 Ω at low frequencies. This variation begins below 1 kHz, but is only fully established below 100 Hz.

The figure of 770 Ω requires some explanation. The incoming signal encounters a 1k8 resistor, in parallel with 5 kΩ which represents half of the HF pot resistance, if we assume that its wiper is at virtual earth. The value of this is 1.32 kΩ; so how on earth can the input impedance fall as low as 770 Ω? The answer is that our assumption is wrong. The HF pot wiper is *not* at virtual earth; it has a signal on it only 7 dB less than at the output and this is in phase with the output; in other words, in anti-phase with the input. This causes 'reverse bootstrapping' of the 5 kΩ resistance that is half of the HF pot, and makes it appear lower in value than it is. A similar 'reverse bootstrapping' effect occurs at the cold input of the standard differential amplifier balanced input stage (see Chapter 18).

At low frequencies, more of the input current is actually going into the HF section of the tone-control network than into the LF section. This highlights one of the few disadvantages of the Baxandall type of tone-control – the input impedances are reduced by parts of the circuit that are not actually doing anything useful at the frequency of interest. Other versions of the tone-control network have a somewhat better behaviour in this respect, and this is examined further on in this chapter.

These input impedances also appear as loading on the output of the opamp in the tone-control stage, when the control settings are reversed. Thus at full LF boost there is a 770 Ω load on the preceding stage, but at full LF cut that 770 Ω loading is on the tone-control opamp.

The Baxandall two-LF-capacitor tone control

The two-capacitor version of the Baxandall tone control is shown in Figure 15.8, and while it looks very similar there is a big difference in the LF end response curves, as seen in Figure 15.9. The LF break frequency rises as the amount of cut or boost is increased. It is my view that this works much better in a hifi system, as it allows small amounts of bass boost to be used to correct loudspeaker deficiencies without affecting the whole of the bass region. In contrast, the one-capacitor version seems to be more popular in mixing consoles, where the emphasis is on creation rather than correction.

The other response difference is the increased amount of 'overshoot' in the frequency response (nothing to do with overshoot in the time domain). Figure 15.9 shows how the use of LF boost causes a small amount of cut just above 1 kHz, and LF cut causes a similar boost. The amounts are small and this is not normally considered to be a problem.

Note that the treble control here has been configured slightly differently, and there are now two end-stop resistors, at each end of the pot, rather than one attached to the wiper; the frequency response is identical, but the input impedance at HF is usefully increased.

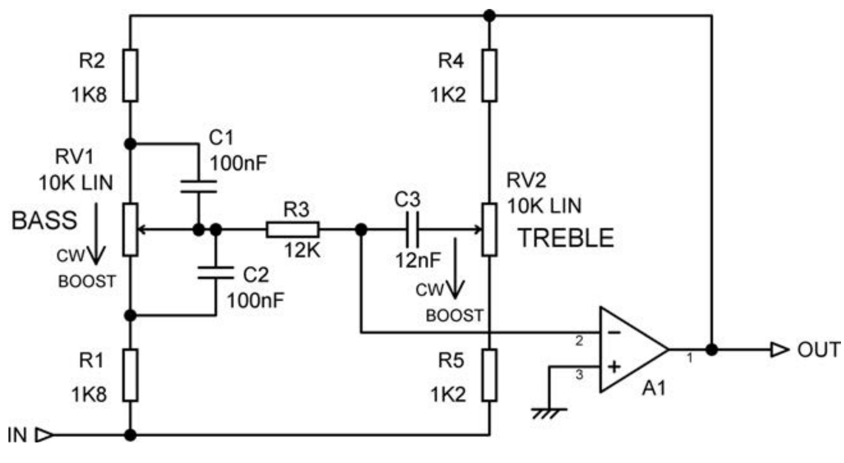

Figure 15.8: The two-LF-capacitor Baxandall tone control

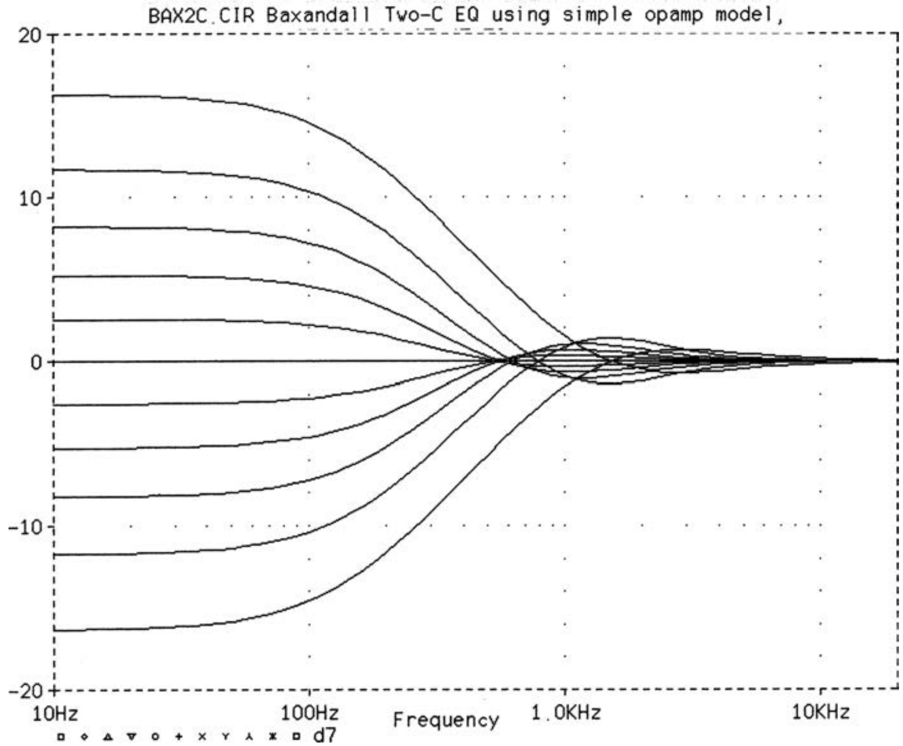

Figure 15.9: Bass control frequency responses for the two-capacitor circuit. Compare Figure 15.4

The input impedance of this version shows variations similar to the one-capacitor version. With controls central, at LF the input impedance is 3.2 kΩ; from 100 Hz to 1 kHz it slowly falls to 1.4 kΩ. The changes with HF and LF control settings are similar to the one-C version, and at HF the impedance falls to 370 Ω.

The Baxandall two-HF-capacitor tone control

The treble control can also be implemented with two capacitors, as in Figure 15.10.

The frequency responses are similar to those of one-HF-capacitor version, but as for the LF control, some 'overshoot' in the curves is introduced. There is a useful reduction in the loading presented to the preceding stage. With controls central, at LF the input impedance is 6.8 kΩ, which is usefully higher than the 2.9 kΩ given by the 1-HF capacitor circuit; from 100 Hz to 1 kHz it slowly falls to 1.4 kΩ.

When the LF control is varied, at full cut the input impedance is increased to 11.7 kΩ, and at full boost it falls to 1.9 kΩ; see Figure 15.11. On varying the HF control, at full cut the

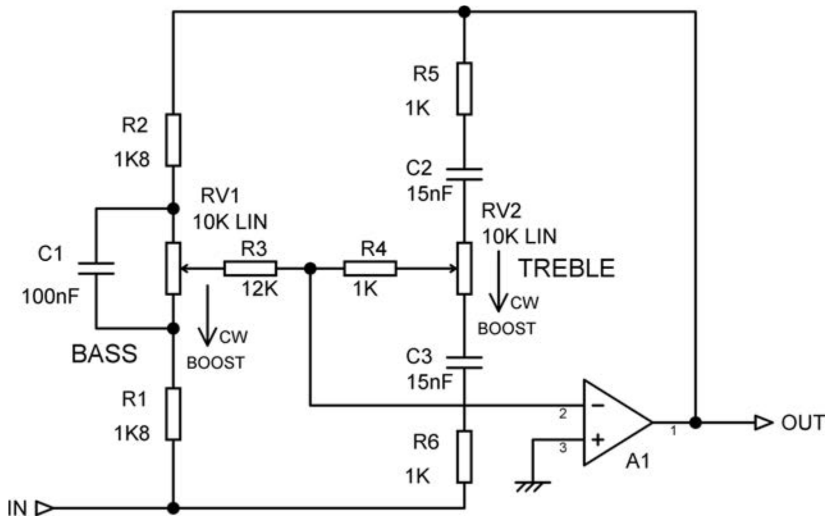

Figure 15.10: Circuit of the two-HF-cap version

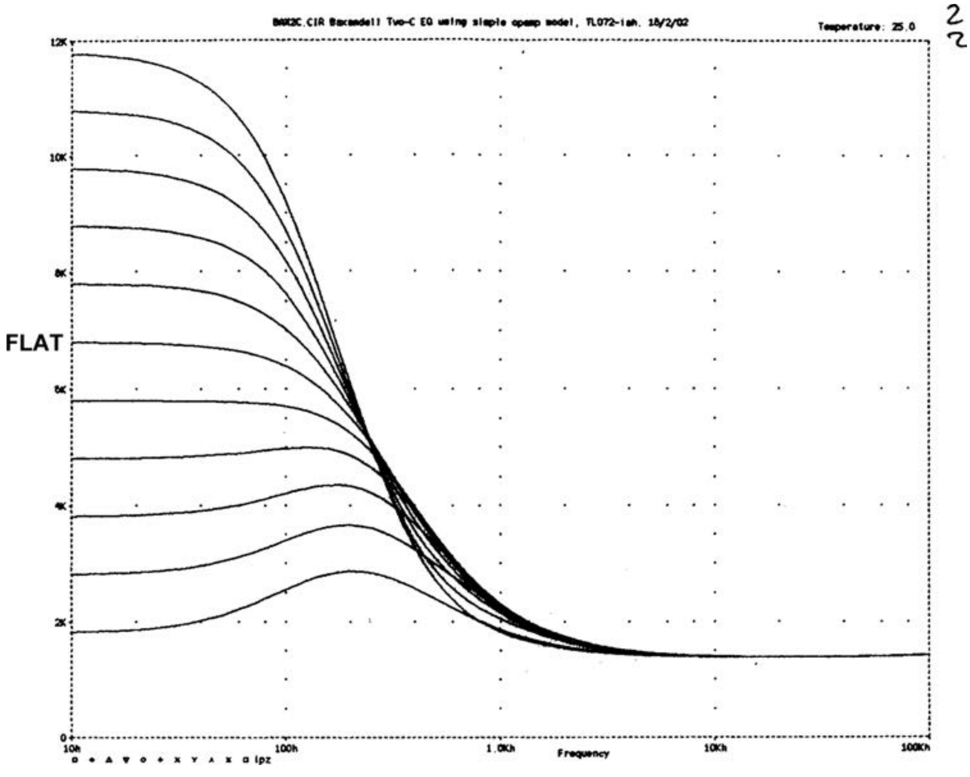

Figure 15.11: Input impedance variation with frequency, for eleven bass control settings; two-HF-capacitor version

input impedance is increased to 2.3 kΩ; at full boost it falls to 420 Ω. These values are higher because with this configuration C3, C4 effectively disconnect the HF pot from the circuit at low frequencies. In some cases the higher input impedance may justify the cost of an extra capacitor. The capacitors will also be about six times larger to obtain the same ±1 dB HF break frequency as the one-HF-capacitor version.

One disadvantage of the Baxandall tone control is that it inherently phase-inverts. This is decidedly awkward, because relatively recently the hifi world has decided that absolute phase is important; in the recording world keeping the phase correct has always been a rigid requirement. The tone-control inversion can however be conveniently undone by a Baxandall active volume control, which also phase-inverts (see Chapter 13). If a balanced input stage is used then an unwanted phase inversion can be corrected simply by swapping over the hot and cold inputs.

A very important point about all of the circuits shown so far is that they assume a FET-input opamp (such as the TL072, or a more sophisticated FET part) will be used to minimise the bias currents flowing. Therefore all the pots are directly connected to the opamp without any explicit provision for preventing DC flowing through them. Excessive DC would make the pots scratchy and crackly when they are moved; this does not sound nice. It is, however, long-established that typical FET bias currents are low enough to prevent such effects in circuits like these; however, there is still the matter of offset voltages to be considered. Substituting a bipolar opamp such as the 5532 will improve the noise and distortion performance markedly, at the expense of the need to make provision for the much greater bias currents by adding DC-blocking capacitors.

The Baxandall tone control: impedance and noise

A major theme in this book is the use of low impedance design to reduce Johnson noise from resistors and the effects of opamp input current noise flowing in them. Let's see how that works with the Baxandall tone control.

The Baxandall controls in Figure 15.12 all give ±10 dB boost and cut, and are equivalent apart from employing 10 kΩ, 5 kΩ, 2 kΩ and 1 kΩ pots, with all other components scaled to keep the frequency responses the same. 1 kΩ is the lowest value in which dual-gang pots can be readily obtained.

Table 15.1 demonstrates that drastically reducing the impedance of the circuit by ten times reduces the noise output by 7.0 dB (controls set flat). Using an LM4562 section instead of a 5532 section shows a similar progression but with somewhat lower noise levels. The improvement on going from 2 kΩ to 1 kΩ pots is small (though reliable) and you may question if it is worth pushing things as far as 1 kΩ, given that, as we saw earlier, Baxandall controls can show surprisingly low input impedance when boosting, with corresponding heavy loads

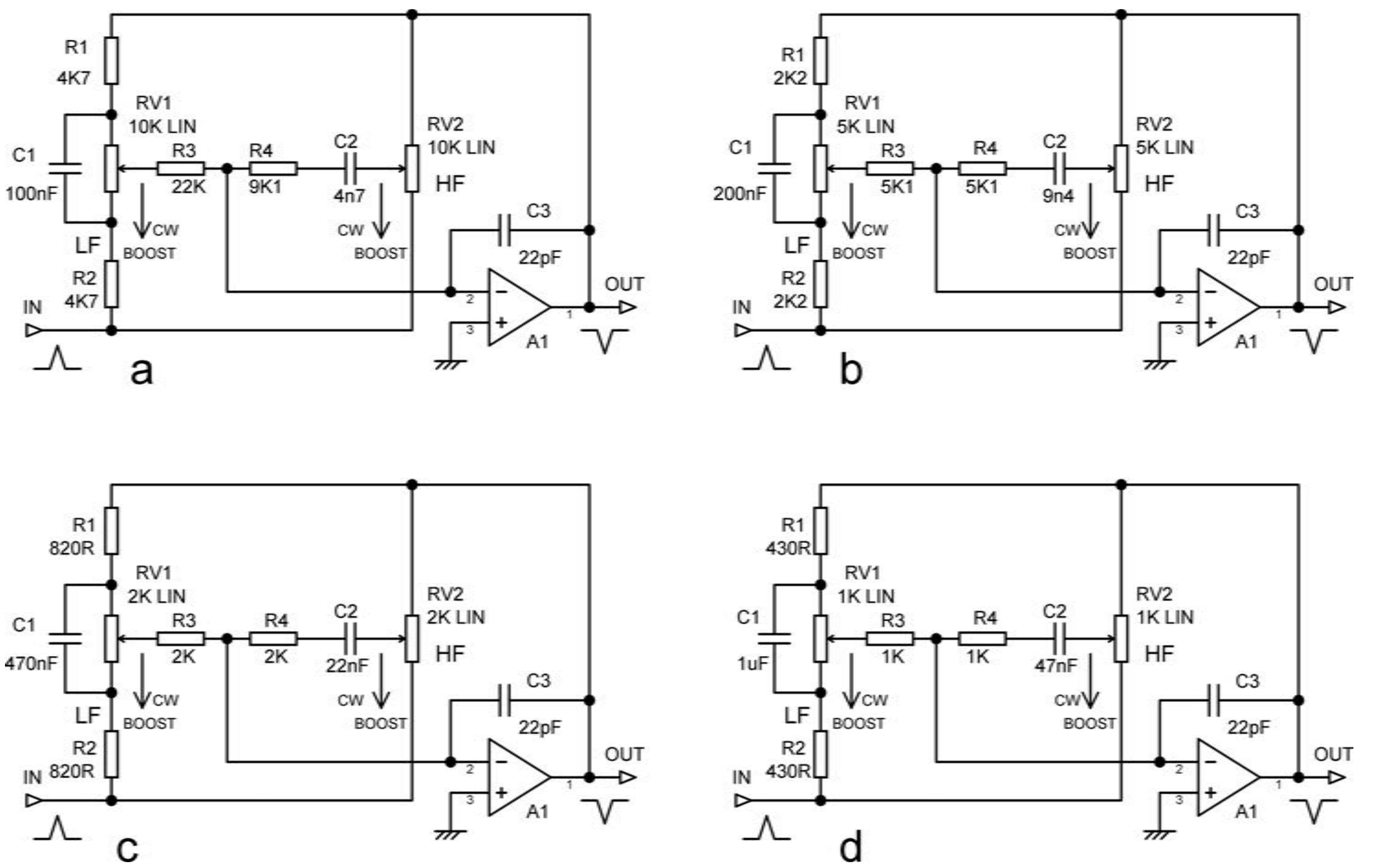

Figure 15.12: Equivalent Baxandall controls giving ±10 dB boost and cut, using 10 k, 5 k, 2 k and 1 k pots

TABLE 15.1 Noise output versus impedance level using a 5532 opamp

Pot value (kΩ)	HF capacitor configuration	Noise output (dBu)
10	1-C	−105.6
5	1-C	−108.9
2	1-C	−112.2
1	1-C	−112.6
1	2-C	−113.1

on the opamp driving the negative feedback path when cutting. The situation becomes proportionally worse when the impedances are scaled down as we have just done. The design process here was driven by a desire to use 1 kΩ pots throughout in the very-low-noise Elektor Preamp 2012. [5]

In the 1 kΩ case, a 5532 is only able to drive 6.7 Vms (±17 V rails) into the negative feedback path before clipping occurs. Clearly we need to either find some way of reducing the loading or increasing the drive capability. In the previous section I described how the Baxandall configuration with two HF capacitors is easier to drive than the one-capacitor version. Figure 15.12d is shown converted to two-HF-capacitor operation in Figure 15.13. The 5532 now clips at 8.8 Vrms, which is much better but the opamp is still overloaded and will not give a good distortion performance. Note that the two HF capacitors are both much larger than the single HF capacitor in Figure 15.12d and will be relatively expensive. On the upside we get the side-benefit of yet lower output noise – see the bottom row of Table 15.1.

Elsewhere in this book I have described the great advantages of using multiple opamps in parallel to drive heavy loads. In this case it is not at all clear how the inverting opamp A1 can

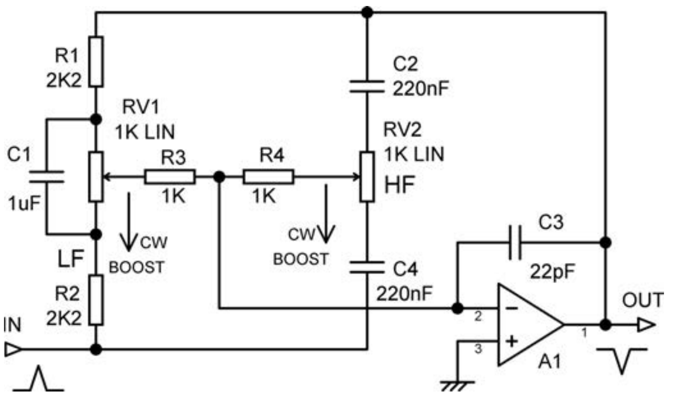

Figure 15.13: Baxandall control with 1 kΩ pots, converted to 2-HF-capacitor configuration to reduce loading

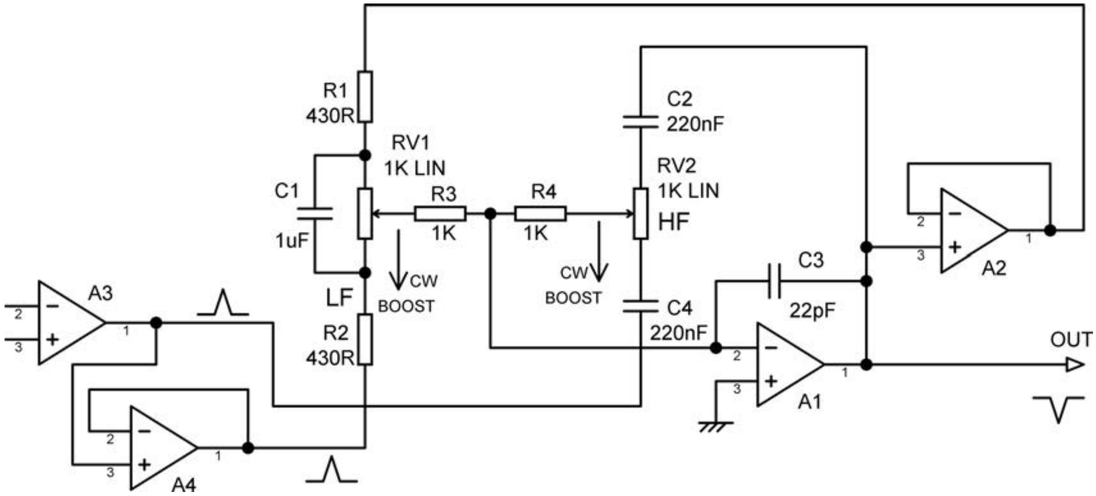

Figure 15.14: Split-drive Baxandall control fed by two parallel opamps and with separate feedback drive to the LF and HF control networks

be multiplied in parallel. My solution is to split the drive to the LF and HF control networks so that the LF section is driven by its own unity-gain buffer A2 (see Figure 15.14). It is the LF section that is driven by the buffer, so that the HF section can be fed directly from A1 and will not suffer phase-shift in A2 that might imperil stability. The output now clips at a healthy 10.8 Vrms, and THD is reduced to the low levels expected.

This scheme is to the best of my knowledge novel; it was used for the first time in the Elektor 2012 preamplifier (using LM4562s) with great success [5]. I call it a Split-Drive Baxandall stage.

The input side of the control also needs at least two opamps to drive it with low distortion. The arrangement here was also used in the Elektor preamp; it makes no assumptions about what A3 is up to (it was part of a balanced line input in the preamp) but simply uses A4 as a buffer to give separate drive to the input of the LF control network.

Switched-HF-frequency Baxandall controls

While the Baxandall approach gives about as much flexibility as one could hope for from two controls, there is often a need for more. The most obvious elaboration is to make the break frequencies variable in some way.

The frequency response of the Baxandall configuration is set by its RC time-constants, and one obvious way to change the frequencies is to make the capacitors switchable, as in an early preamp design of mine [6]. Changing the R part of the RC is far less practical as it would require changing the potentiometer values as well. If fully variable frequencies controlled by pots are wanted then a different configuration must be used, as described later in this chapter.

If a large number of switched frequencies (more than three) are wanted then a relatively expensive rotary switch is required, but if three will do a centre-off toggle switch will certainly take up less panel space, and probably be cheaper; you don't have to pay for a knob. This is illustrated in Figure 15.15, where C1 is always in circuit, and either C2 or C3 can be switched-in parallel. This is an HF-only tone-control with ±1 dB break frequencies at 1 kHz, 3 kHz and 5 kHz; most people will find frequencies much higher than that to be a bit too subtle. A similar approach can be used with the two-HF-capacitor Baxandall control, but twice as much switching is required.

A disadvantage of the centre-off switch is that the maximum rather than the middle frequency is obtained at the central toggle position. If this is unacceptable then C&K make a switch variant (Model 7411) [7] with internal connections that are made in the centre position, so that two switch sections will make up a 1-pole 3-way switch as in Figure 15.16. A stereo version is physically a 4-pole switch. This method is naturally more expensive than the centre-off approach.

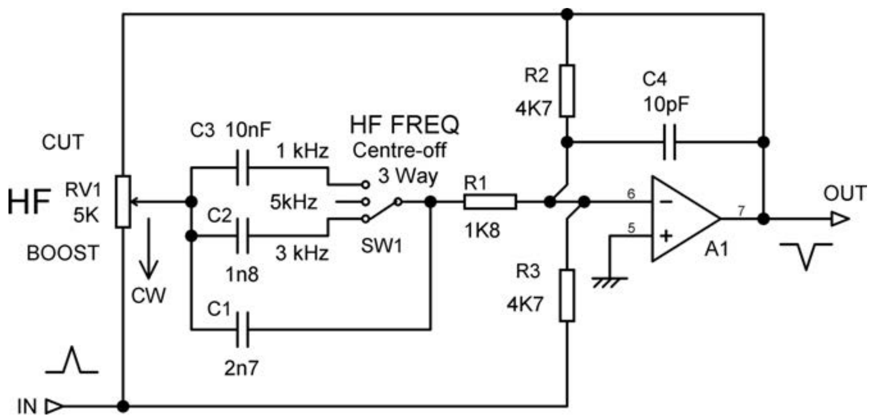

Figure 15.15: A Baxandall HF-only tone control with three switched turnover frequencies

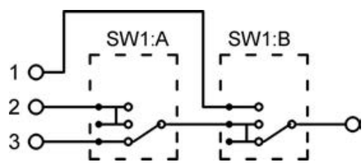

Figure 15.16: Making a 1-pole 3-way switch with two C&K 7411 switch sections

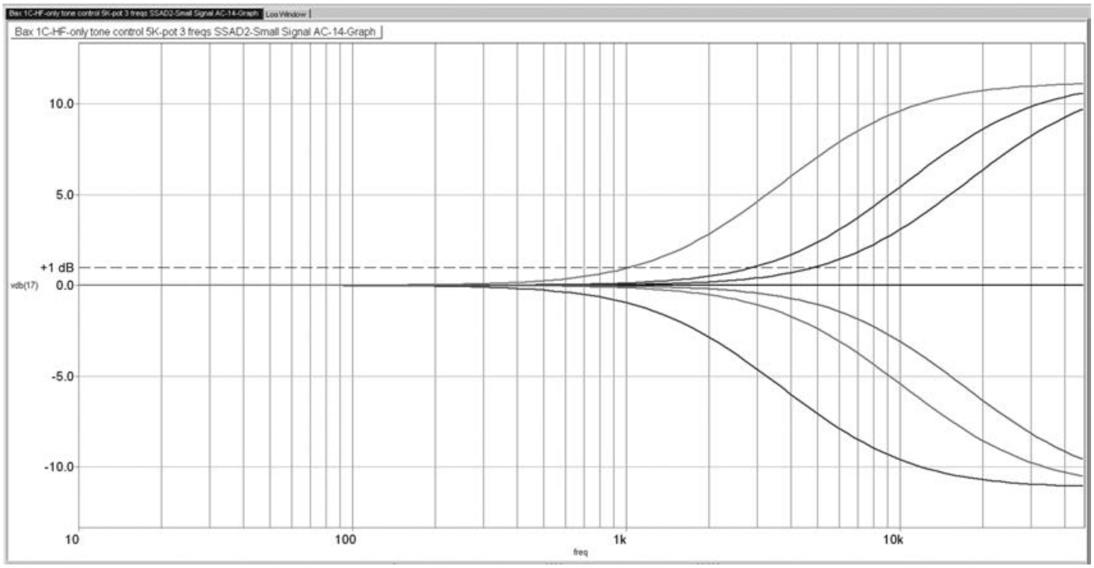

Figure 15.17: Frequency response of tone control with three switched HF break frequencies, at 1, 3 and 5 kHz

The circuit of Figure 15.15 also demonstrates how an HF-only tone-control is configured, with R2 and R3 taking the place of the LF control, and providing negative feedback at DC and LF. If this is combined with an LF-only control, the second phase inversion cancels the first and both stages can be bypassed by a tone-cancel switch without a phase change. The response is shown in Figure 15.17.

A similar approach can be used to switch the LF capacitor in a Baxandall control. The capacitor values tend to be inconveniently large for low break frequencies, especially if you are using low-value pots. A different tone control configuration gives more modest sizes – see the section 'Variable-frequency LF EQ' below.

Variable-frequency HF EQ

Since we are moving now more into the world of mixing consoles rather than preamplifiers, I shall stop using 'bass' and 'treble' and switch to 'LF' and 'HF', which of course mean the same thing. The circuit shown in Figure 15.18 gives HF equalisation only, but with a continuously variable break frequency, and is used in many mixer designs. It is similar to the Baxandall concept in that it uses opamp A1 in a shunt feedback mode so that it can provide either cut or boost, but the resemblance ends there.

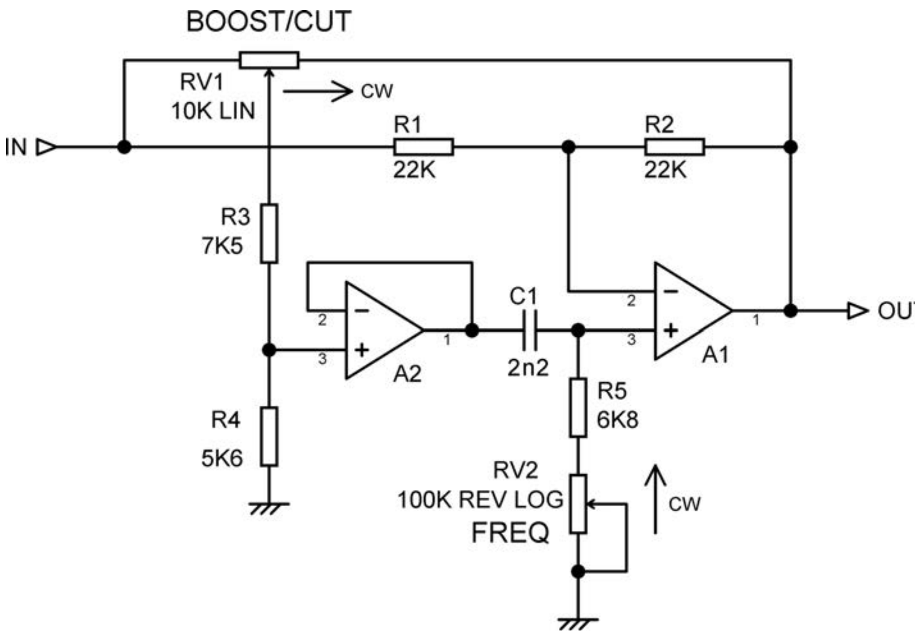

Figure 15.18: A variable-frequency HF shelving circuit. The ±1 dB break frequency range is 400 Hz–6.4 kHz

R1 and R2 set the basic gain of the circuit to −1, and ensure that there is feedback at DC to establish the operating conditions; there is no DC path to the non-inverting input of A1. When the wiper of RV1 is at the output end, positive feedback partially cancels the negative feedback through R2 and the gain increases. When the wiper of RV1 is at the input end, the signal fed through A2 causes partial cancellation of the input signal, and gain is reduced.

The signal tapped off is scaled by divider R3, R4, which set the maximum cut/boost. The signal is then buffered by voltage-follower A2, and fed to the frequency-sensitive part of the circuit, a high-pass RC network made up of C1 and (R5 + RV2). Since only high frequencies are passed, this circuit has no effect at LF. The response is shown in Figure 15.19. The frequency-setting control RV2 has a relatively high value at 100 kΩ, because this allows for a 16:1 of variation in frequency, and a lower value would give excessive loading on A2 at the high frequency end. This assumes TL072 or similar opamps are used to avoid the need to deal with opamp input bias currents. If 5532 or LM4562 opamps are used there is considerable scope for reduction in the circuit impedances, resulting in lower noise.

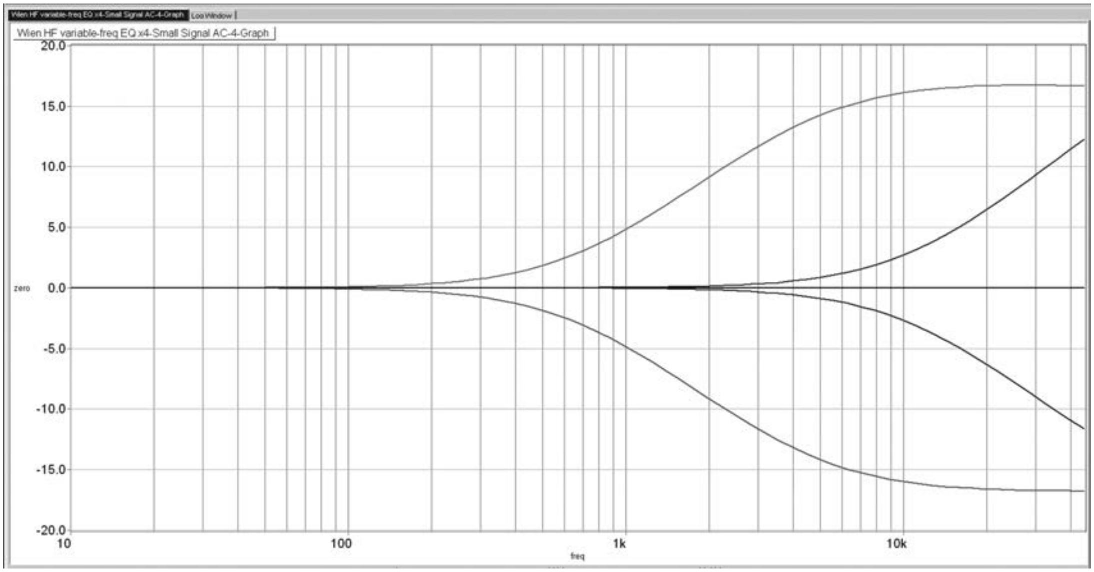

Figure 15.19: The response of the variable frequency HF shelving circuit at extreme frequency settings. Maximum cut/boost is slightly greater than ±15 dB

A2 prevents interaction between the amount of cut/boost and the break frequency. Without it cut/boost would be less at high frequencies because R5 would load the divider R3, R4. In cheaper products this interaction may be acceptable, and A2 could be omitted. In EQ circuitry it is the general rule that the price of freedom from control interaction is not eternal vigilance but more opamps. The frequency range can be scaled simply by altering the value of C1. The range of frequency variation is controlled by the value of end-stop resistor R5, subject to restrictions on loading A2 if distortion is to be kept low.

The component values given are the E24 values that give the closest approach to ±15 dB cut/boost at the control extremes. When TL072 opamps are used this circuit is stable as shown, assuming the usual supply rail decoupling. If other types are used, a small capacitor (say, 33 pF) across R2 may be required.

Variable-frequency LF EQ

The corresponding variable LF frequency EQ is obtained by swapping the positions of C1 and R5, RV2 in Figure 15.18 to give Figure 15.20. Now only the low frequencies are passed through, and so only they are controlled by the boost/cut control. The frequency response is shown in Figure 15.21.

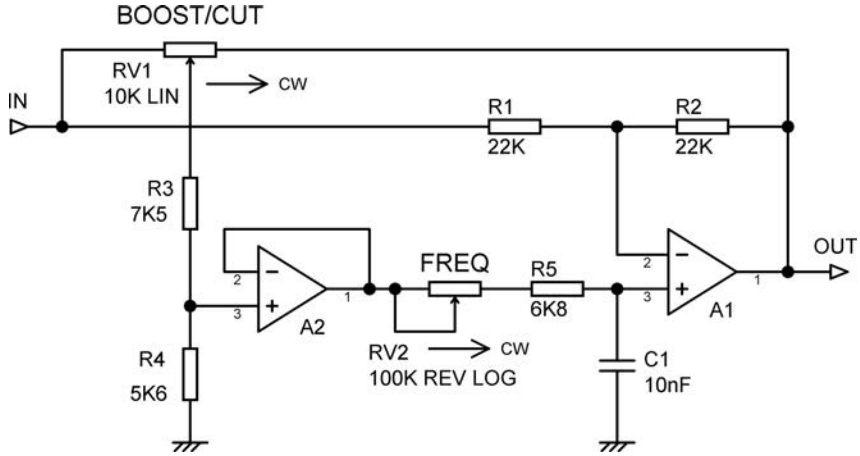

Figure 15.20: A variable frequency LF shelving circuit. The ±1 dB break frequency range is 60 Hz–1 kHz. Boost/cut range of ±15 dB is set by R3 and R4

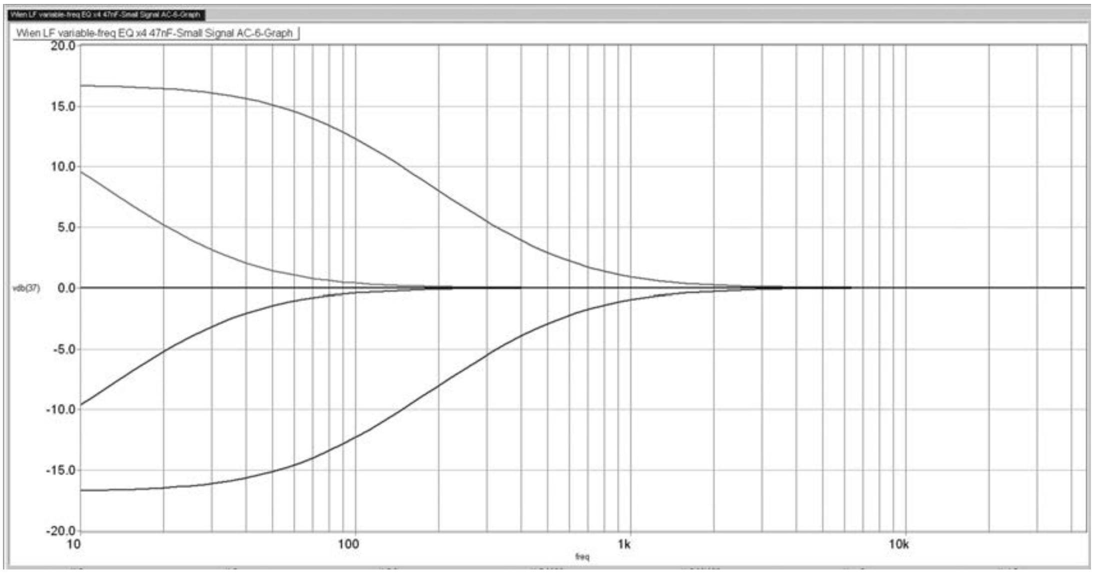

Figure 15.21: The response of the variable frequency LF shelving circuit at extreme frequency settings. Maximum cut/boost is slightly greater than ±15 dB

A new type of switched-frequency LF EQ

In the circuit of Figure 15.20, R3 and R4 attenuate the signal before it passes through the unity-gain buffer A2. If a very low noise stage is required, it may be possible to put the attenuator after the buffer instead, in which case it will attenuate the buffer's noise by

7.4 dB. This is easier to do for a switched-frequency rather than a fully variable frequency EQ. An example is Figure 15.22, which has switched ±1 dB break frequencies at 100 Hz and 400 Hz and gives ±10 dB maximal cut and boost. With SW1 open, R3 and R4 attenuate the signal by the appropriate amount to give ±10 dB. Their combined impedance (R3 and R4 in parallel) gives the required 100 Hz frequency in combination with C1. When SW1 is closed, R5 is now in parallel with R3 and R6 is in parallel with R4; the attenuation is the same but the impedance is reduced so the break frequency increases to 400 Hz.

At first sight it appears that the relatively high values of R3, R4, R5, R6 would make the circuit very noisy. In fact, the large value of C1 means that their noise is largely filtered out. In the first version measured R1 and R2 were 4.7 kΩ, and the output noise (flat) was −109.2 dBu using LM4562 for both opamp sections. Reducing R1, R2 to 2.2 kΩ as shown reduced this handily to −111.1 dBu, which is rather quiet.

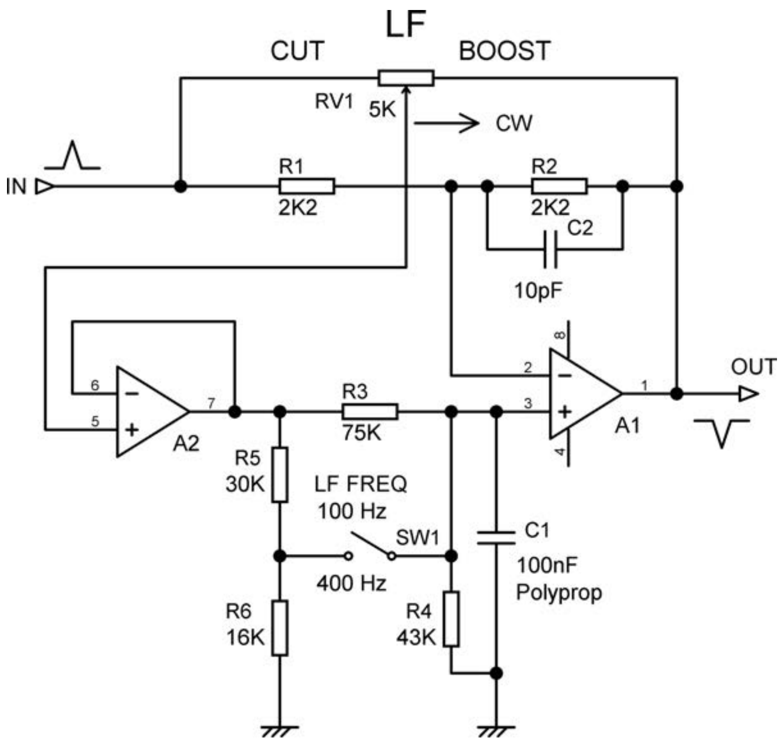

Figure 15.22: Low-noise switched-frequency LF EQ with the attenuation after buffer A2

Variable-frequency HF and LF EQ in one stage

In this section I describe how to combine variable frequency LF and HF shelving EQ in one stage. This gives some component economy because opamp A1 can be shared between the two functions. Despite this the schematic in Figure 15.23 is a little more complex than you might expect from looking at the two circuits above, because this is intended to be premium EQ with the following extra features:

- The response returns to unity gain outside the audio band. This is often called return-to-flat (RTF) operation. The fixed RTF time-constants mean that the boost/cut range is necessarily less at the frequency extremes, where the effect of RTF begins to overlap the variable boost/cut frequencies.
- The frequency control pots all have a linear law for better accuracy. The required logarithmic law is implemented by the electronic circuitry.
- There is an inbuilt tone-cancel switch that does not cause an interruption in the signal when it is operated.
- Low-impedance design is used to minimise noise.

The boost/cut range is ±10 dB, the LF frequency range is 100 Hz–1 kHz, and the HF frequency range is 1 kHz–10 kHz. This design was used in my low-noise preamplifier published in Jan Didden's *Linear Audio* in 2012 [8], being developed from an earlier version used in the Precision Preamplifier 96 [9]. Very few hifi manufacturers have ever offered this facility. The only one that comes to mind is the Yamaha C6 preamp (1980–81) which had LF and HF frequency variable by slider over wide ranges (they could even overlap in the 500 Hz–1 kHz range) and Q controls for each band as well. The Cello Sound Palette (1992) had a reputation as a sophisticated tone control, but it used six boost/cut bands at fixed frequencies that gave much less flexibility [10].

The variable boost/cut and frequencies make the tone-control much more useful for correcting speaker deficiencies, allowing any error at the top or bottom end to be corrected to at least a first approximation. It makes a major difference.

This control was developed for hifi rather than mixer use, and certain features of the tone control were aimed at making it more acceptable to those who think any sort of tone-control is an Abomination. The control range is restricted to ±10 dB, rather than the ±15 dB which is standard in mixing consoles. The response is built entirely from simple 6 dB/octave circuitry, with inherently gentle slopes. The stage is naturally minimum-phase, and so the amplitude curves uniquely define the phase response. The maximum phase-shift does not exceed 40 degrees at full boost, not that it matters because you can't hear phase-shift anyway.

The schematic is shown in Fig 15.23. The tone-control gives a unity-gain inversion except when the selective response of the sidechain paths allow signal through. In the treble and bass frequency ranges where the sidechain does pass signal, the boost/cut pots RV1, RV2 can give

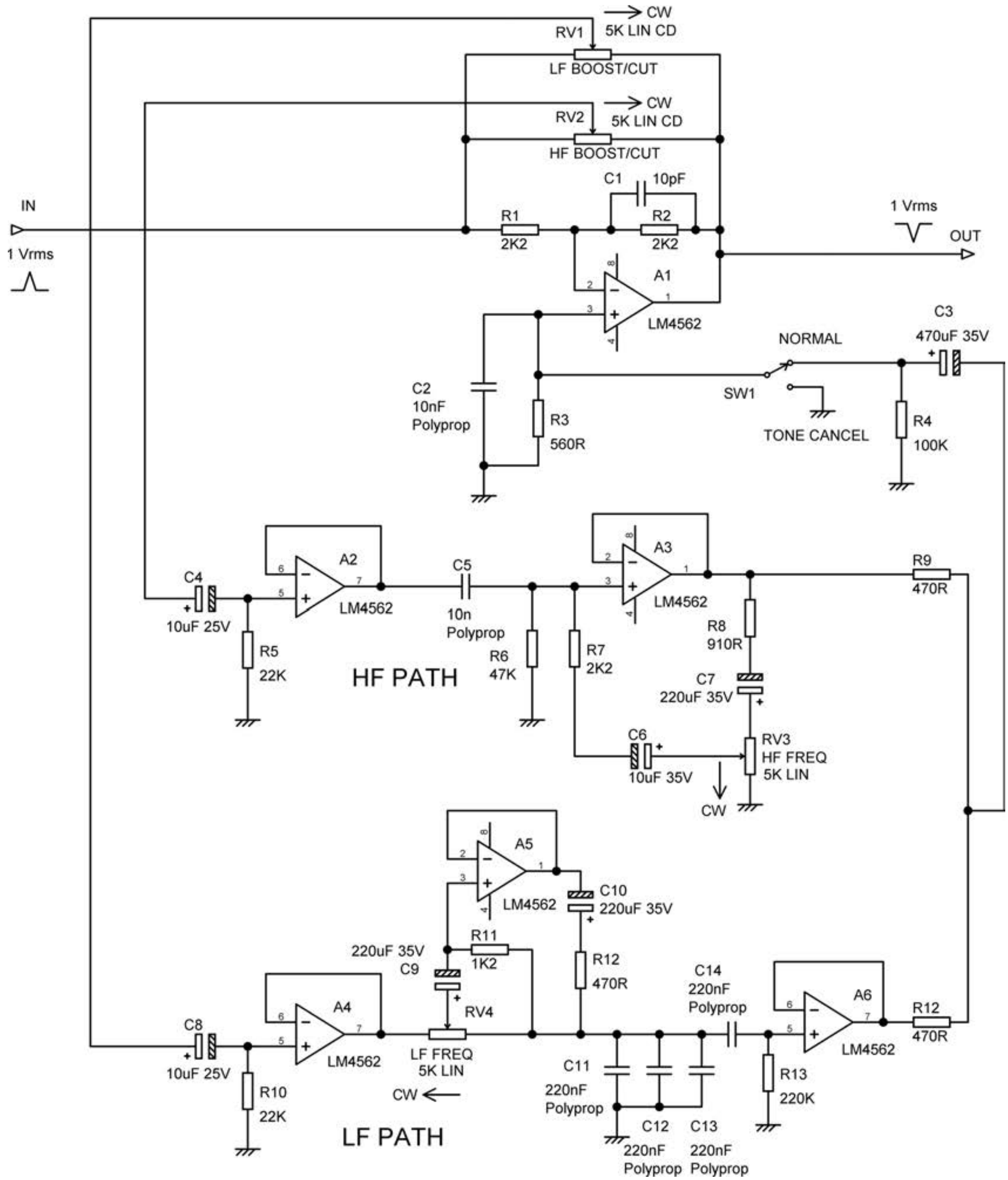

Figure 15.23: The tone-control schematic, showing the two separate paths for HF and LF control. All pots are 5 kΩ linear

either gain or attenuation. When a wiper is central, there is a null at the middle of the boost/cut pot, no signal through that sidechain, and the gain is unity.

If the pot is set so the sidechain is fed from the input then there is a partial cancellation of the forward signal; if the sidechain is fed from the output then there is a partial negative-feedback cancellation, or to put it another way, positive feedback is introduced to counteract part of the NFB. This apparently ramshackle process actually gives boost/cut curves of perfect symmetry. This is purely cosmetic, because you can't use both sides of the curve at once, so it hardly matters if they are exact mirror-images.

The tone control stage acts in separate bands for bass and treble, so there are two parallel selective paths in the sidechain. These are simple RC time-constants, the bass path being a variable-frequency first-order low-pass filter, and the associated bass control only acting on the frequencies this lets through. Similarly the treble path is a variable high-pass filter. The filtered signals are summed and returned to the main path via the non-inverting input, and some attenuation must be introduced to limit cut and boost. Assuming a unity-gain sidechain, this loss is 9 dB if cut/boost is to be limited to ±10 dB. This is implemented by R9, R12 and R4. The sidechain is unity-gain, and has no problems with clipping before the main path does, so it is very desirable to put the loss after the sidechain, where it attenuates sidechain noise. The loss attenuator is made up of the lowest value resistors that can be driven without distortion, to minimise both the Johnson noise thereof and the effects of the current noise of opamp A1.

The Tone Cancel switch disconnects the entire sidechain (5 out of 6 opamps) from any contribution to the main path, and usefully reduces the stage output noise by about 4 dB. It leaves only A1 in circuit. Unlike configurations where the entire stage is by-passed, the signal does *not* briefly disappear as the switch moves between two contacts. This minimises transients due to suddenly chopping the waveform and makes valid tone in/out comparisons much easier.

It is very convenient if all pots are identical. I have used linear 5 kΩ controls, so the tolerances inherent in a two-slope approximation to a logarithmic law can be eliminated. This only presents problems in the tone stage frequency controls, where linear pots require thoughtful circuit design to give the logarithmic action that fits our perceptual processes.

In the HF path, C5, R7 is the high-pass time-constant, driven at low-impedance by unity-gain buffer A2. This prevents the frequency from altering with the boost/cut setting. The effective value of R7 is altered over a 10:1 range by varying the amount of bootstrapping it receives from A3, the potential divider effect and the rise in source resistance of RV3 in the centre combining to give a reasonable approximation to a logarithmic frequency/rotation law. R8 is the frequency endstop resistor that limits the maximum effective value of R7. C2 is the treble RTF capacitor – at frequencies above the audio band it shunts all of the sidechain signal to ground, preventing the HF control from having any effect. The HF sidechain degrades the noise performance of the tone control by 2–3 dB when connected because it passes the HF end of the audio band. The noise contribution is greatest when the HF freq is set to minimum, and the widest bandwidth from the HF sidechain contributes to the main path. This is true even when the HF boost/cut control is set to flat, as the HF path is still connected to A1.

In the LF path, A4 buffers RV1 to prevent boost/frequency interaction. The low-pass time-constant capacitor is the parallel combination of C11 + C12 + C13; the use of three capacitors increases the accuracy of the total value by $\sqrt{3}$. The associated resistance is a combination of RV4 and R11, R12. C14 and R13 make up the RTF time-constant for the LF path, blocking very low frequencies and so limiting the lower extent of LF control action. The bass frequency law is made approximately logarithmic by A5; for minimum frequency RV4 is set fully CCW, so the input of A5 is the same as the C11 end of R12, which is thus bootstrapped and has no effect. When RV4 is fully CW, R11, R12 are effectively in parallel with RV4 and the turnover frequency is at a maximum. R11 gives a roughly logarithmic law; its value is carefully chosen so that the centre of the frequency range is at the centre of the control travel. Sadly, there is some pot-dependence here. The LF path uses three opamps rather than two, but contributes very little extra noise to the tone stage, because most of its output is rolled off by the low-pass action of C11, C12, C13 at HF, eliminating almost all its noise contribution apart from that of A6.

The measured responses for maximum boost and maximum cut at minimum, middle, and maximum frequency settings are shown in Figures 15.24 and 15.25.

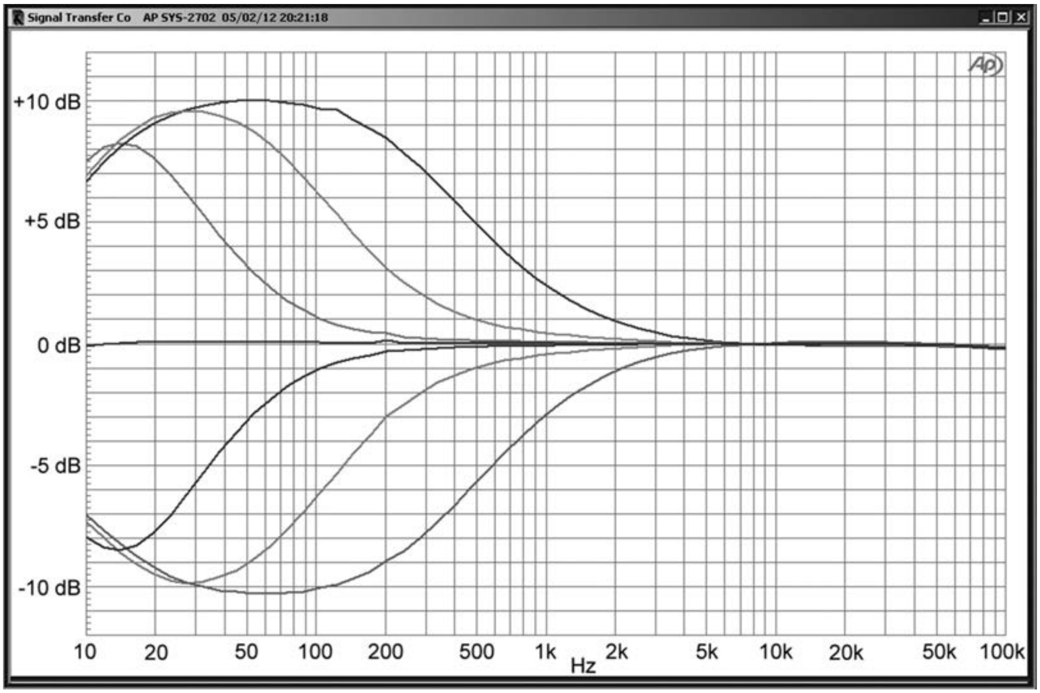

Figure 15.24: LF tone-control frequency response, max cut/boost, at minimum, middle, and maximum frequencies

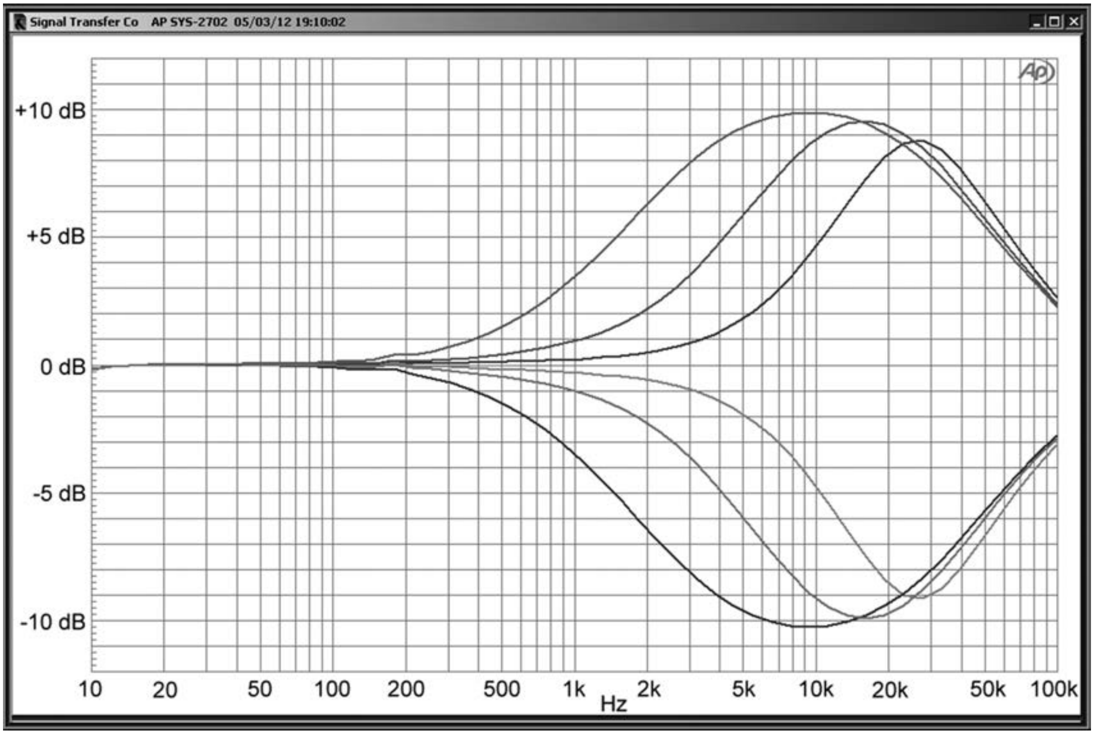

Figure 15.25: HF tone-control frequency response, max cut/boost, at minimum, middle, and maximum frequencies

An important goal in the design of this tone control was low noise. Once the opamps have been chosen, and the architecture made sensible in terms of avoiding attenuation-then-amplification, keeping noise-gain to a minimum, and so on, the remaining way of improving noise performance is by low-impedance design. The resistances are lowered in value, with capacitances scaled up to suit, by a factor that is limited only by opamp drive capability. The 5532 or LM4562 are good for this. A complete evaluation of the noise performance is a lengthy business because of the large number of permutations of the controls. If we just look at extreme and middle positions for each control, we have maximum boost, flat, and maximum cut for both HF and LF, and maximum, middle and minimum for the two frequency controls, yielding $3 \times 3 \times 3 \times 3 = 81$ permutations. The measurements given in Table 15.2 are therefore restricted to those that put the greatest demands on the circuitry, e.g. HF freq at minimum.

Measuring the distortion performance of the tone-control is likewise a protracted affair, requiring the exploration of the permutations of the four controls. There were no surprises so I will not use up valuable space displaying the results; if there is interest I will put them on my website at douglas-self.com. Suffice it to say that THD at 9 Vrms in/out never gets above 0.001%. General THD levels are in the range 0.0003–0.0007%

TABLE 15.2 The noise output of the tone control at various settings

HF level	HF freq	LF level	LF freq	Noise out (dBu)
Flat	Min	Flat	Max	−105.1
Flat	Mid	Flat	Max	−106.2
Flat	Max	Flat	Max	−107.2
Flat	Max	Flat	Mid	−106.8
Flat	Max	Flat	Min	−107.1
Flat	Min	Flat	Min	−105.6
Max boost	Min	Flat	Min	−100.5
Max cut	Min	Flat	Min	−107.4
Max cut	Min	Flat	Max	−107.6
Flat	Min	Max boost	Max	−103.8
Flat	Min	Max cut	Max	−105.9
Flat	Min	Max cut	Min	−105.6
Flat	Min	Max boost	Min	−105.0
Tone-cancel				−110.2

Note: Not corrected for AP noise at −119.2 dBu (difference negligible). Measurement bandwidth 22 Hz–22 kHz, rms sensing, unweighted.

Be aware that circuits like this tone-control can show unexpected input impedance variations. A standard Baxandall tone-control made with 10 kΩ pots can have an input impedance that falls to 1 kΩ or less at high frequencies where the capacitors have a low impedance. It is not obvious but the alternative tone-control configuration used here also shows significant input impedance variations.

Looking at the circuit in Figure 15.23, you might think that because the input terminal connects only to a 4k7 resistor and two 5 kΩ pots the input impedance cannot fall below their parallel combination i.e. $4k7 \| 5k \| 5k = 1.63$ kΩ. You would be wrong; the other end of the 4k7 resistor is connected to virtual ground, but the two 5 kΩ pots are connected to the stage output. When the controls are set flat, or tone-cancel engaged, this carries an inverted version of the input signal. The effective value of the pots is therefore halved, with zero voltage occurring halfway along the pot tracks. The true input impedance when flat is therefore $4k7 \| 2.5k \| 2.5k = 987\ \Omega$, which is confirmed by simulation.

When the tone-control is not set flat, but to boost, then at those frequencies the inverted signal at the output is larger than the input. This makes the input impedance lower than for the flat case; when the circuit is simulated it can be seen that the input impedance varies with frequency inversely to the amount of boost. Conversely, when the tone-control is set to cut, the inverted signal at the output is reduced and the input impedance is higher than in the flat case. This is summarised in Table 15.3, with R1, R2 = 4k7.

Table 15.3 shows that the input impedance falls to the worryingly low figure of 481 Ω at maximum HF and LF boost. In this case the gain is +10 dB, and so the input voltage into this impedance cannot exceed 3 Vrms without the tone-control output clipping. This limits the current required of the preceding stage, and there are not likely to be problems with increased distortion if this is 5532 or LM4562 based.

As mentioned earlier, this tone control was developed from an earlier version designed some 16 years earlier. It is instructive to take a quick look at the changes that were made:

- All opamps changed from 5532 to LM4562 to reduce noise and distortion.
- A general impedance reduction to reduce noise including the use of 5 kΩ linear pots throughout instead of 10 kΩ.
- Frequency control laws improved so the middle of the frequency range corresponds with the middle position of the control.
- Three expensive 470 nF polypropylene capacitors in the LF path replaced with four of 220 nF, giving a cost reduction. The resulting impedance at C14 is quite high at low frequencies and the pick up of electrostatic hum must be avoided.
- DC-blocking capacitor C10 added to the LF path frequency control to eliminate rustling noises.
- C3 increased from 220 μF to 470 μF to remove a trace of electrolytic capacitor distortion at 10 Hz and 9 Vrms out (with max LF freq and max boost). Reduced from 0.0014% to less than 0.0007%.

The electrolytic capacitors in the tone-control are used for DC-blocking only, and have no part at all in determining the frequency response. It would be most undesirable if they did, both because of the wide tolerance on their value and the distortion generated by electrolytics when they have significant signal voltage across them. The design criterion for

TABLE 15.3 Input impedance of the tone control at various settings

HF level	HF freq	LF level	LF freq	Input impedance (Ω)
Flat	Min	Flat	Mid	987
Flat	Mid	Max boost	Mid	481
Flat	Mid	Max cut	Mid	1390
Max boost	Mid	Flat	Mid	480
Max cut	Mid	Flat	Mid	1389

these capacitors was that they should be large enough to introduce no distortion at 10 Hz, the original values being chosen by the algorithm 'looks about right', though I hasten to add this was followed up by simulations to check the signal voltages across them and THD measurements to confirm all was well.

Tilt or tone-balance controls

A tilt or tone-balance circuit is operated by a single control and affects not just part of the audio spectrum but most or all of it. Typically the high frequencies are boosted as the low frequencies are cut, and vice versa. It has to be said that the name 'tone-balance' is unfortunate as it implies it has something to do with interchannel amplitude balance, which it has not. A stereo 'tone-balance control' alters the frequency response of both channels equally and does not introduce amplitude differences between them, whereas a 'stereo balance control' is something quite different. It is clearer to call a tone-balance control a tilt control, and I shall do so.

Tilt controls are (or were) supposedly useful in correcting the overall tonal balance of recordings in a smoother way than a Baxandall configuration, which concentrates more on the ends of the audio spectrum. An excellent (and very clever) approach to this was published by Ambler in 1970 [11] (see Figure 15.26). The configuration is very similar to the Baxandall – the ingenious difference here is that the boost/cut pot effectively swaps its ends over as the frequency goes up. At low frequencies C1, C2 do nothing, and the gain is set by the pot, with maximum cut and boost set by R1, R2. At high-frequencies, where the capacitors are effectively short-circuited, R3, R4 overpower R1, R2 and the control works in reverse. The range available with the circuit shown is ±8 dB at LF and ±6.5 dB at HF. This may seem ungenerous, but because of the way the control works, 8 dB of boost in the bass is accompanied by 6.5 dB of cut in the treble, and a total change of 14.5 dB in the relative

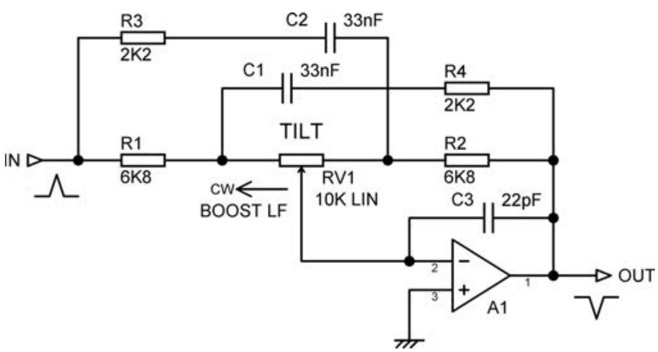

Figure 15.26: Tilt control of the Ambler type

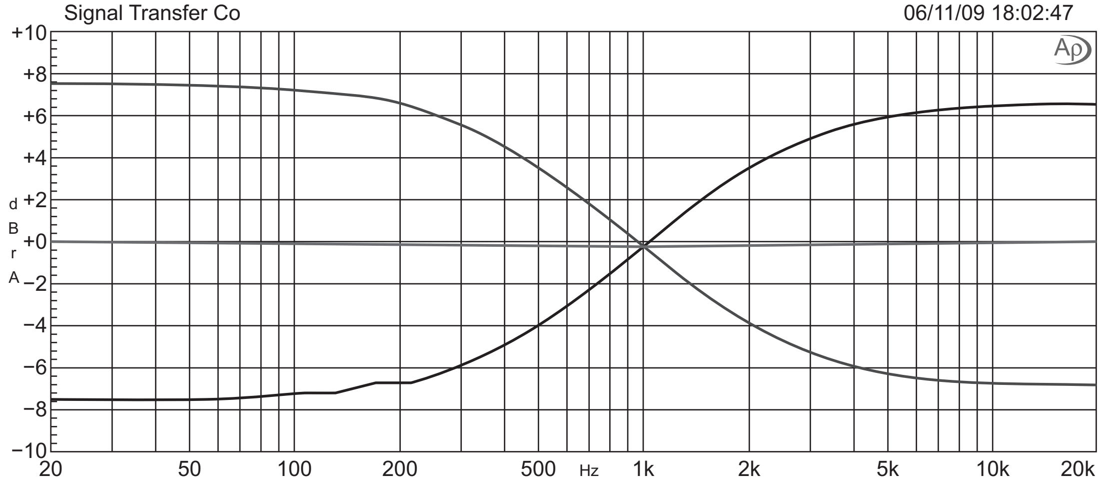

Figure 15.27: Frequency response of Ambler tilt control

level of the two parts of the spectrum should be enough for anyone. The measured frequency response at the control limits is shown in Figure 15.27; the response is not quite flat with the control central due to component tolerances.

The need for one set of endstop resistors to take over from the other puts limits on the cut/boost that can be obtained without the input impedance becoming too low; there is, of course, also the equivalent need to consider the impedance that opamp A1 sees when driving the feedback side of the network.

The input impedance at LF, with the control set to flat, is approx 12 kΩ, which is the sum of R3 and half of the pot resistance. At HF, however, the impedance falls to 2.0 kΩ. Please note that this is not a reflection of the values of the HF endstop resistors R3, R4, but just a coincidence. When the control is set to full treble boost, the input impedance at HF falls as low as 620 Ω. The impedance at LF holds up rather better at full bass boost, as it cannot fall below the value of R1, i.e. 6.8 kΩ.

The impedances of the circuit shown here have been reduced by a factor of ten from the original values published by Ambler, to make them more suitable for use with opamps. The original (1970) gain element was a two-transistor inverting amplifier with limited linearity and load-driving capability. Here a stabilising capacitor C3 is shown explicitly, just to remind you that you might need one.

A famous example of the use of a tilt control is the Quad 44 preamplifier. The tilt facility is combined with a bass cut/boost control in one quite complicated stage, and it is not at all obvious if the design is based on the Ambler concept. Tilt controls have never really caught on and remain rare. One current example is the Classé CP-800 D/A preamplifier, reviewed by

Stereophile in 2012. [12] This gives a maximum of ±6 dB control at the frequency extremes, and is implemented by DSP. In analogue use the signal has to be converted to 24-bit digital data, processed, then converted back to analogue, which to me seems somewhat less than elegant.

Middle controls

A middle control affects the centre of the audio band rather than the bass and treble extremes. It must be said at once that middle controls, while useful in mixers, are of very little value in a preamplifier. If the middle frequency is fixed, then the chances that this frequency and its associated Q corresponds with room shortcomings or loudspeaker problems is remote in the extreme. Occasionally, middle controls appeared on preamps in the Seventies, but only rarely and without much evidence of success in the marketplace. One example is the Metrosound ST60 (1972), which had a 3-band Baxandall tone control – more on this below – with slider controls. The middle control had a very wide bandwidth centred on 1 kHz, and it was suggested that it could be used to depress the whole middle of the audio band to give the effect of a loudness control.

Middle controls come into their own in mixers and other sound-control equipment, where they are found in widely varying degrees of sophistication. In recording applications middle controls play a vital part in 'voicing' or adjusting the timbres of particular instruments, and the flexibility of the equaliser, and its number of controls, defines the possibilities open to the operator. A 'presence' control is centred on the upper-middle audio frequencies so it tends to accentuate vocals when used; nowadays the term seems to be restricted to tone controls on electric guitars.

The obvious first step is to add a fixed middle control to the standard HF and LF controls. Unfortunately, this is not much more useful in a mixer than in a preamplifier. In the past this was addressed by adding more fixed middles, so a line-up with a high-middle and a low-middle would be HF-HMF-LMF-LF, but this takes up more front panel space (which is a very precious resource in advanced and complex mixers, and ultimately defined by the length of the human arm) without greatly improving the EQ versatility.

The minimum facilities in a mixer input channel for proper control are the usual HF and LF controls plus a sweep middle with a useful range of centre frequency. This also uses four knobs but is much more useful.

Fixed frequency Baxandall middle controls

Figure 15.28 shows a middle version of a Baxandall configuration. The single control RV1 now has around it both the time constants that were before assigned to the separate bass and treble controls. R4 and R5 maintain unity gain at DC, and keep the stage biased correctly.

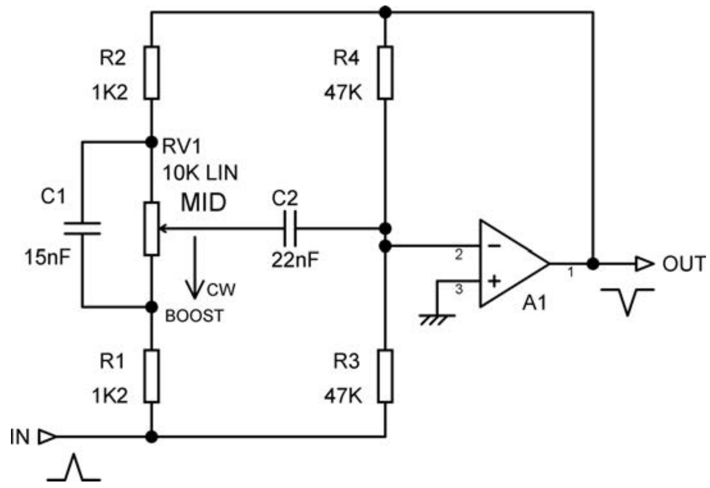

Figure 15.28: Fixed middle control of the Baxandall type. Centre frequency 1.26 kHz

As the input frequency increases from the bottom of the audio band, the impedance of C2 falls and the position of the pot wiper begins to take effect. At a higher frequency, the impedance of C1 becomes low enough to effectively tie the two ends of the pot together, so that the wiper position no longer has effect and the circuit reverts to having a fixed gain of unity. The component values shown give a mid frequency of 1.26 kHz, at a Q of 0.8, with a maximum boost/cut of ±15 dB. The Q value is only valid at maximum boost/cut; with less, the curve is flatter and the effective Q lower. It is not possible to obtain high values of Q with this approach.

This circuit gives the pleasingly symmetrical curves shown in Figure 15.29, though it has to be said that the benefits of exact symmetry are visual rather than audible.

As mentioned above, in the simpler mixer input channels it is not uncommon to have two fixed mid controls; this is not the ideal arrangement, but it can be implemented very neatly and cheaply as in Figure 15.30. There are two stages, each of which has two fixed bands of EQ. It has the great advantage that there are two inverting stages so the output signal ends up back in phase. The first stage needs the extra resistors R5, R6 to maintain DC feedback.

There will inevitably be some control interaction with this scheme. It could be avoided by using four separate stages, but this is most unlikely to be economical for mixers with this relatively simple sort of EQ. To minimise interaction, the control bands are allocated between the stages to keep the frequencies controlled in a stage as far apart as possible, combining HF with LO MID, and LF with HI MID, as shown here.

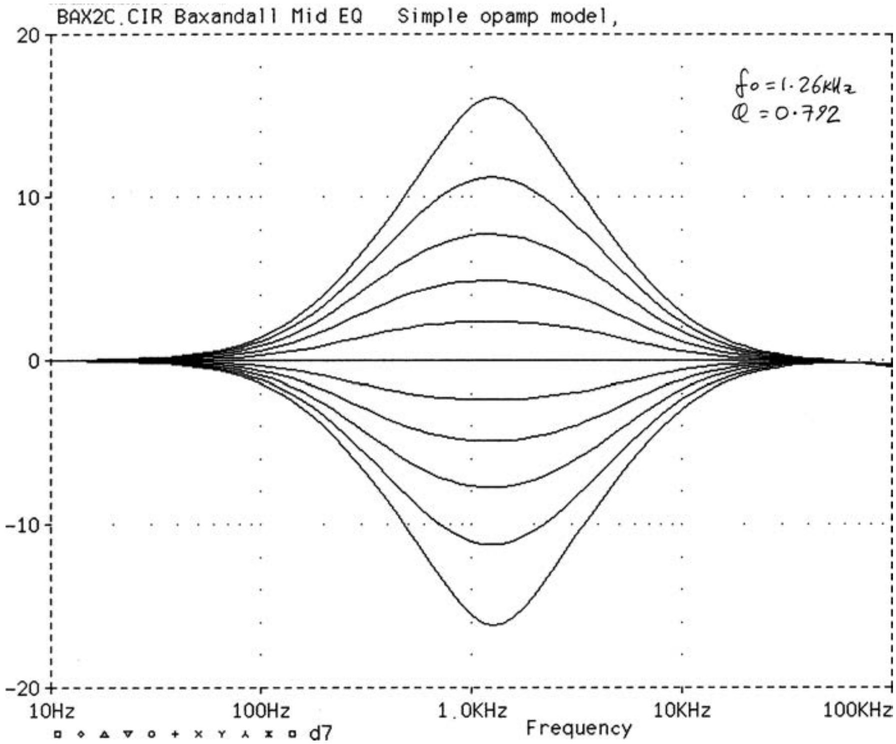

Figure 15.29: The frequency response of the Baxandall middle control in Figure 15.27

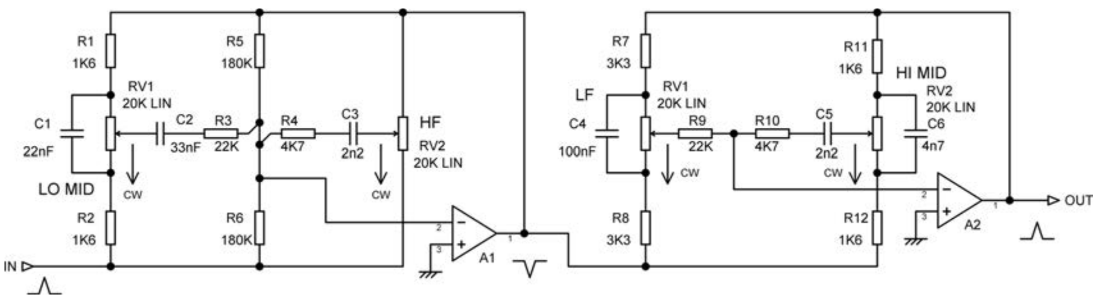

Figure 15.30: A 4-band Baxandall EQ using two stages only

Three-band Baxandall EQ in one stage

The standard Baxandall tone control allows adjustment of two bands with one stage. When a three-band EQ is required, it is common practice to use one such stage for HF and LF and a following one to implement the middle control only. This has the advantage that the two cascaded inverting stages will leave the signal in the correct phase.

When this is not a benefit, because a phase inversion is present at some other point in the signal path, it is economical to combine HF, MID and LF in one quasi-Baxandall stage. This not only reduces component count, but reduces power consumption by saving an opamp. The drawback is that cramming all this functionality into one stage requires some compromises on control interaction and maximum boost/cut. The circuit shown in Figure 15.31 gives boost/cut limited to ±12 dB in each band. The pots are now 20 kΩ to prevent the input and feedback impedances from becoming too low.

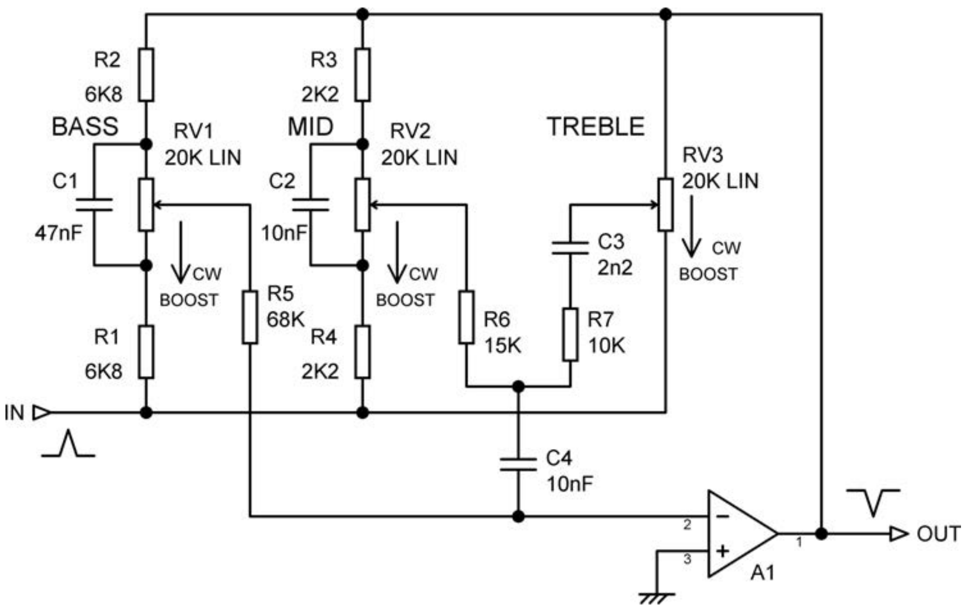

Figure 15.31: The circuit of a Baxandall three-band EQ using one stage only

The frequency responses for each band are given in Figures 15.32, 15.33 and 15.34.

Wien fixed middle EQ

An alternative way to implement a fixed middle control is shown in Figure 15.35. Here the signal tapped off from RV1 is fed to a Wien bandpass network R1, C1, R2, C2, and returned to the opamp non-inverting input. This is the same Wien network as used in audio oscillators.

With the values shown, the centre frequency is 2.26 kHz and the Q at max cut/boost is 1.4; it gives beautifully symmetrical response curves like those in Figure 15.33 above, with a maximal cut/boost of 15.5 dB.

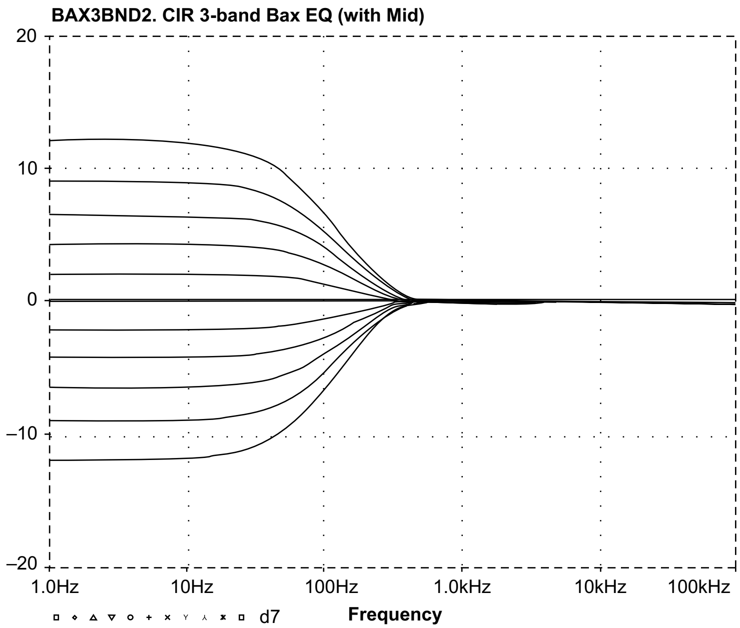

Figure 15.32: The frequency response of a Baxandall three-band EQ. Bass control

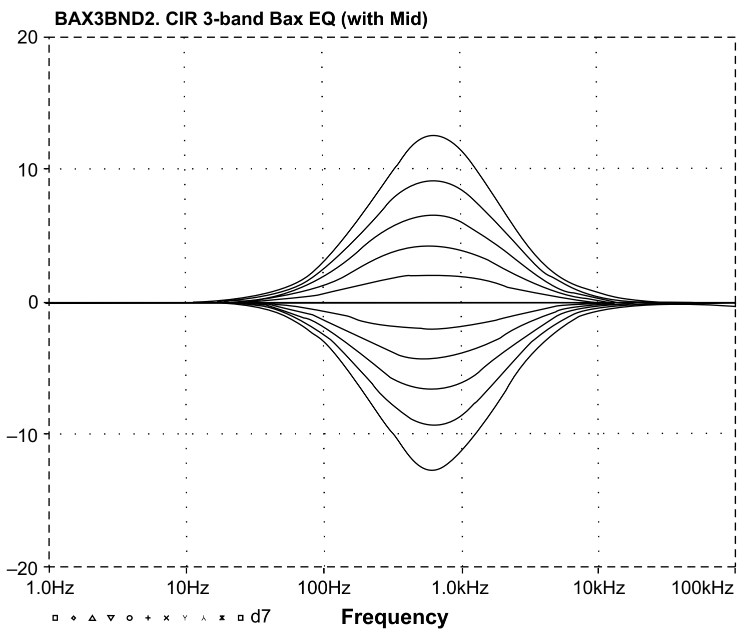

Figure 15.33: The frequency response of a Baxandall three-band EQ. Mid control

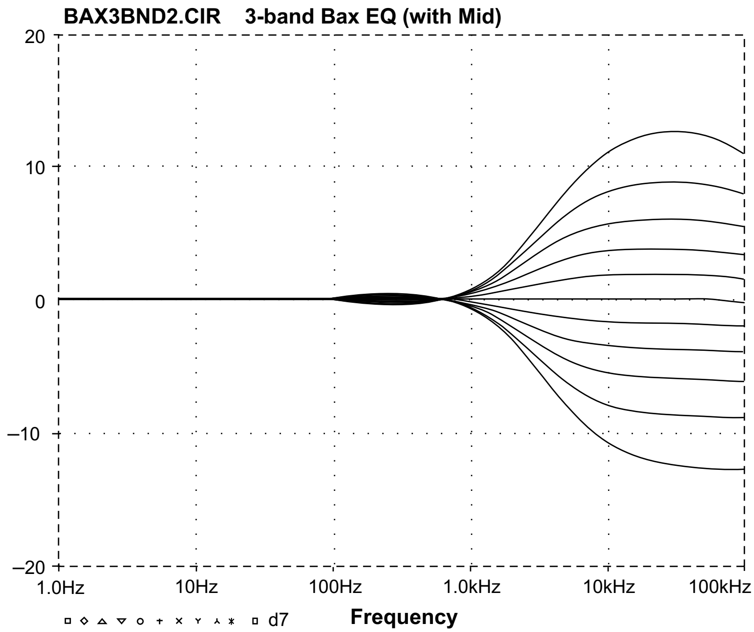

Figure 15.34: The frequency response of the Baxandall three-band EQ. Treble control

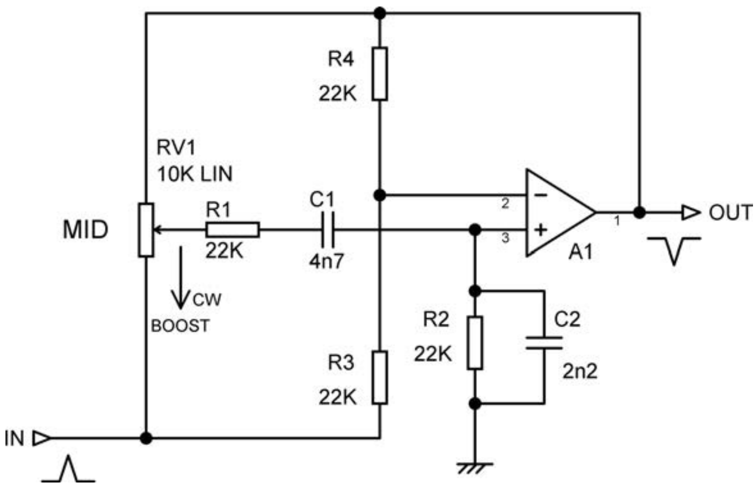

Figure 15.35: The Wien fixed-middle circuit. Centre frequency is 2.26 kHz. The Q at max cut/ boost is 1.4

Variable-frequency middle EQ

A fully variable frequency middle control is much more useful and versatile than any combination of fixed or switched middle frequencies. In professional audio this is usually called a 'sweep middle' EQ. It can be implemented very nicely by putting variable resistances in the Wien network of the stage previously described, and the resulting circuit is shown in Figure 15.36.

The variable load that the Wien network puts on the cut/boost pot RV1 causes a small amount of control interaction, which is normally considered acceptable in middle-range mixers. It could be eliminated by putting a unity-gain buffer stage between RV1 wiper and the Wien network, but in the middle-range mixers where this circuit is commonly used, this is not normally economical.

The Wien network here is carefully arranged so that the two variable resistors RV2, RV3 have four common terminals, reducing the number of physical terminals required from six to three. This is sometimes taken advantage of by pot manufacturers making ganged parts specifically for this EQ application. R1, C1 are sometimes seen swapped in position but this naturally makes no difference.

The combination of a 100k pot and a 6k8 end-stop resistor gives a theoretical frequency ratio of 15.7 to 1, which is about as much as can be obtained using reverse-log Law C pots, without excessive cramping at the high-frequency end of the scale. This will be marked on the control calibrations as a 16 to 1 range.

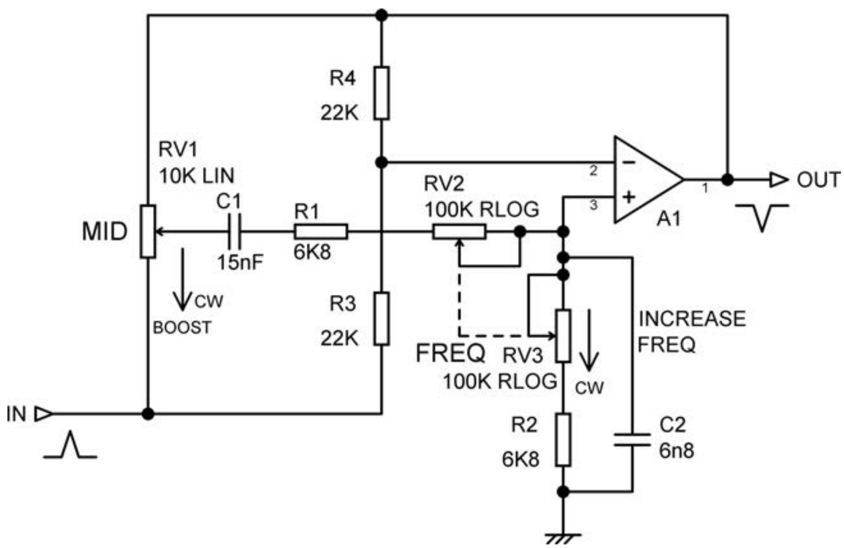

Figure 15.36: A Wien sweep circuit. The centre-frequency range is 150 Hz–2.4 kHz (a 16:1 ratio)

The measured frequency responses at the control limits are shown in Figure 15.37. The frequency range is from 150 Hz to 2.3 kHz; the ratio is slightly adrift due to component tolerances.

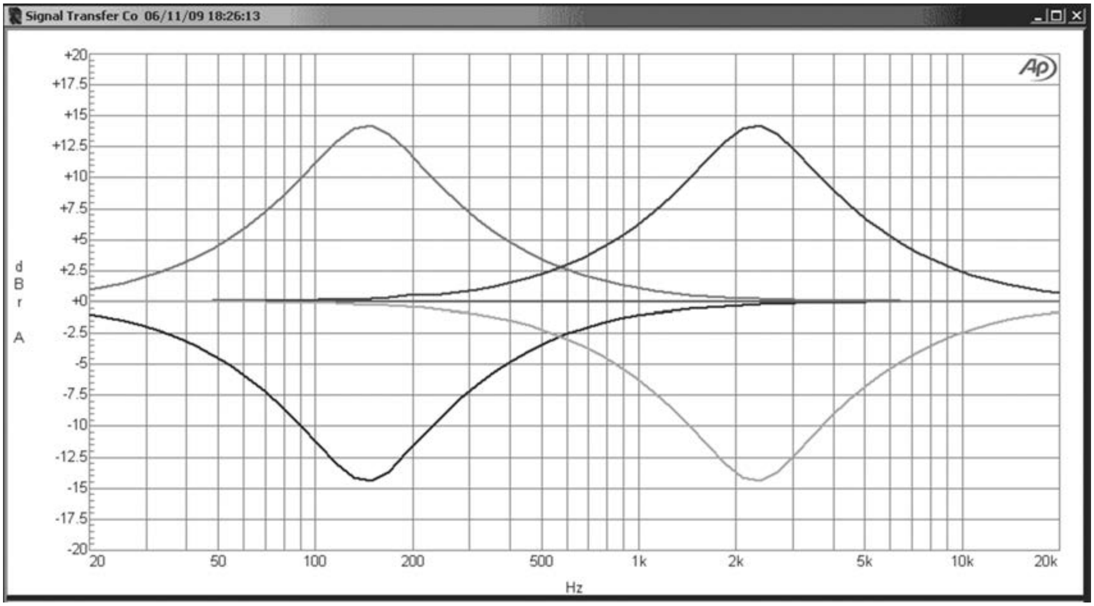

Figure 15.37: The measured response of the sweep middle circuit at control extremes. The cut/boost is slightly short of ±15 dB

Single-gang variable-frequency middle EQ

The usual type of sweep middle requires a dual-gang reverse-log pot to set the frequency. These are not hard to obtain in production quantities, but can be difficult to get in small numbers. They are always significantly more expensive than a single pot.

The problem becomes more difficult when the design requires a stereo sweep middle – if implemented in the usual way, this demands a four-gang reverse-log pot. Once again, such components are available, but only to special order, which means long lead-times and significant minimum order quantities. Four-gang pots are not possible in flat-format mixer construction where the pots are mounted on their backs, so to speak, on a single big horizontal PCB. The incentive to use a standard component is strong, and if a single-gang sweep middle circuit can be devised, a stereo EQ only requires a dual-gang pot.

This is why many people have tried to design single-gang sweep middle circuits, with varying degrees of success. It can be done, so long as you don't mind some variation of Q with centre frequency; the big problem is to minimise this interaction. I too have attacked this problem, and here is my best shot so far, in Figure 15.38

This circuit is a variation of the Wien middle EQ, the quasi-Wien network being tuned by a single control RV2 which not only varies the total resistance of the R5, R7 arm, but also the amount of bootstrapping applied to C2, effectively altering its value. This time a unity-gain buffer stage A2 has been inserted between RV1 wiper and the Wien network; this helps to minimise variation of Q with frequency.

The response is shown in Figure 15.39; note that the frequency range has been restricted to 10:1 to minimise Q variation. The graph shows only maximum cut/boost; at intermediate settings the Q variations are much less obvious.

It is possible to make a yet more economical version of this, if one accepts somewhat greater interaction between boost/cut and Q and frequency. The version shown in Figure 15.40 omits the unity-gain buffer and uses unequal capacitor values to raise the Q of the quasi-Wien network, saving an opamp section. The frequency range is still 10:1.

The response in Figure 15.41 shows the drawback – a higher Q at the centre of the frequency range than for the three-opamp version. Once again the Q variations will be much less obvious at intermediate cut/boost settings.

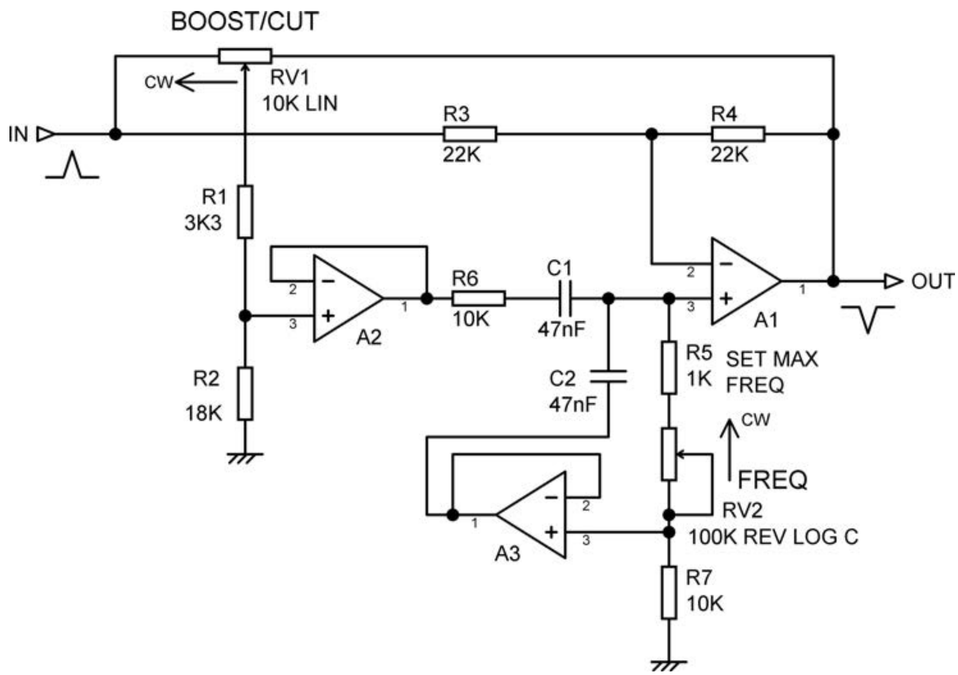

Figure 15.38: My single-gang sweep middle circuit. The centre frequency range is 100 Hz–1 kHz

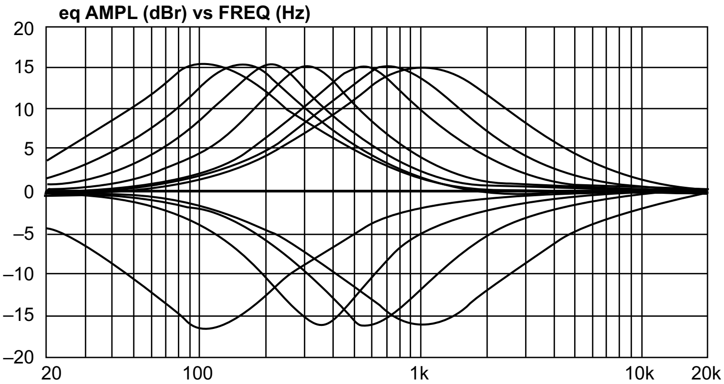

Figure 15.39: The response of the single-gang sweep middle circuit of Figure 15.26. Boost/cut is ±15 dB, and the frequency range is 100 Hz–1 kHz. The Q varies somewhat with centre frequency

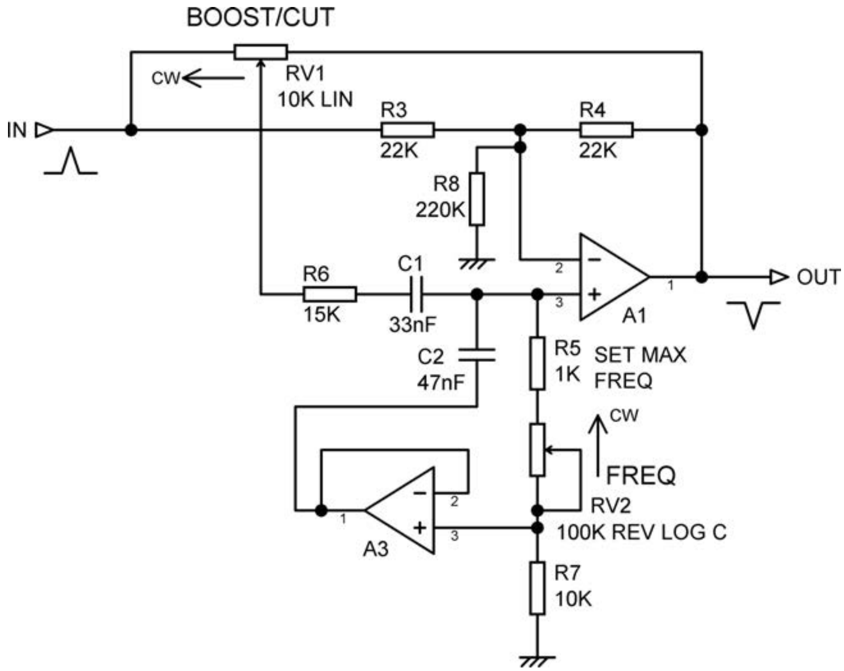

Figure 15.40: My economical single-gang sweep-middle circuit Figure 15.28. The centre frequency range has been changed to 220 Hz–2.2 kHz

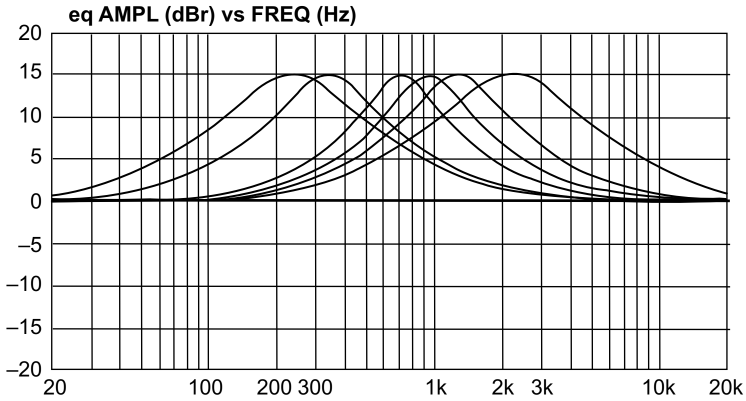

Figure 15.41: The response of the economical single-gang sweep-middle circuit. The response is ±15 dB as before, but the frequency range has now been set to 220 Hz–2 kHz. Only the boost curves are shown; cut is the mirror-image

The question naturally arises as to whether it is possible to design a single-gang sweep-middle circuit where there is absolutely no variation of Q with frequency. Is there an 'existence theorem' i.e. a mathematical proof that it can't be done? At the present time, I don't know . . .

Switched-Q variable-frequency Wien middle EQ

The next step in increasing EQ sophistication is to provide control over the Q of the middle resonance. This is often accomplished by using a full state-variable filter solution, which gives fully variable Q that does not interact with the other control settings, but if two or more switched values of Q are sufficient, there are much simpler circuits available.

One of them is shown in Figure 15.42; here the Wien bandpass network is implemented around A2, which is essentially a shunt-feedback stage, with added positive feedback via R1, R2 to raise the Q of the resonance. When the Q-switch is in the LO position, the output from A2 is fed directly back to the non-inverting input of A1, because R7 is short-circuited. When the Q-switch is in the HI position, R9 is switched into circuit and increases the positive feedback to A2, raising the Q of the resonance. This also increases the gain at the centre frequency, and this is compensated for by the attenuation now introduced by R7 and R8.

With the values shown, the two Q values are 0.5 and 1.5. Note the cunning way that the Q switch is made to do two jobs at once – changing the Q and also introducing the compensating attenuation. If the other half of a two-pole switch is already dedicated to a LED indicator, this saves having to go to a four-pole switch. On a large mixing console with many EQ sections this sort of economy is important.

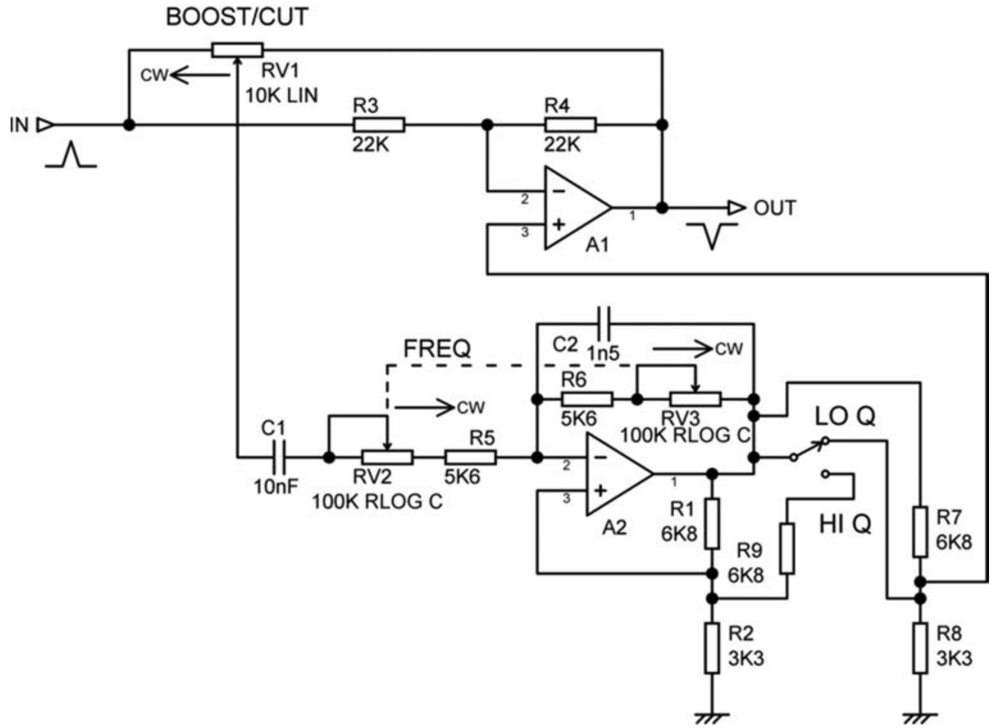

Figure 15.42: Variable-freq switched-Q middle control of the Wien type

Switchable peak/shelving LF/HF EQ

It is frequently desirable to have the highest and lowest frequency EQ sections switchable between a peaking (resonance) mode and shelving operation. The peaking mode allows relatively large amounts of boost to be applied near the edges of the audio band without having a large and undesirable amount occurring outside it.

Figure 15.43 shows one way of accomplishing this. It is essentially a switchable combination of the variable-frequency HF shelving circuit of Figure 15.18 and the sweep-middle circuit of Figure 15.36; when the switches are in the PEAK position the signal tapped off RV1 is fed via the buffer A2 to a Wien bandpass network C2, RV2, R5, C3, R6, RV3, and the circuit has a peak/dip characteristic. When the switches are in the SHLV (shelving) position, the first half of the Wien network is disconnected and C1 is switched in, and in conjunction with R6 and RV3 forms a first-order high-pass network, fed by an attenuated signal because R2 is now grounded. This switched attenuation factor is required to give equal amounts of cut/boost in the two modes because the high-pass network has less loss than the Wien network. R7 allows fine-tuning of the maximum cut/boost; reducing it increases the range.

As always we want our switches to work as hard as possible, and the lower switch can be seen to vary the attenuation brought about by R1, R2 with one contact and switch in C3 with the other. Unfortunately, in this case two poles of switching are required. The response of the circuit at one frequency setting can be seen in Figure 15.44.

When the peaking is near the edges of the audio band, this is called RTF (return-to-flat) operation, or sometimes RTZ (return-to-zero) operation as the gain returns to unity (Zero dB) outside the peaking band.

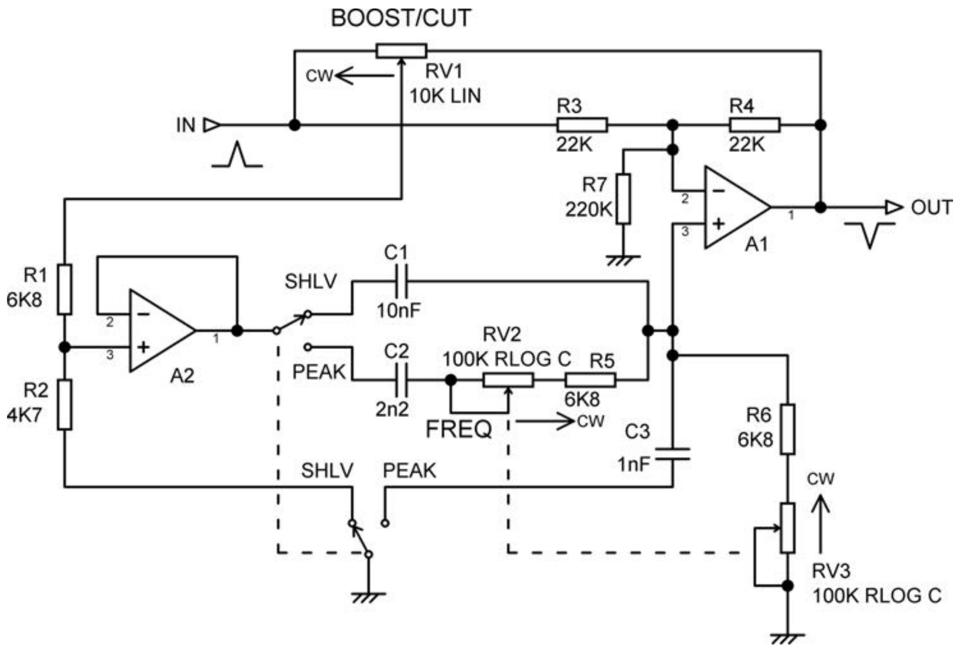

Figure 15.43: Variable-frequency peak/shelving HF EQ circuit

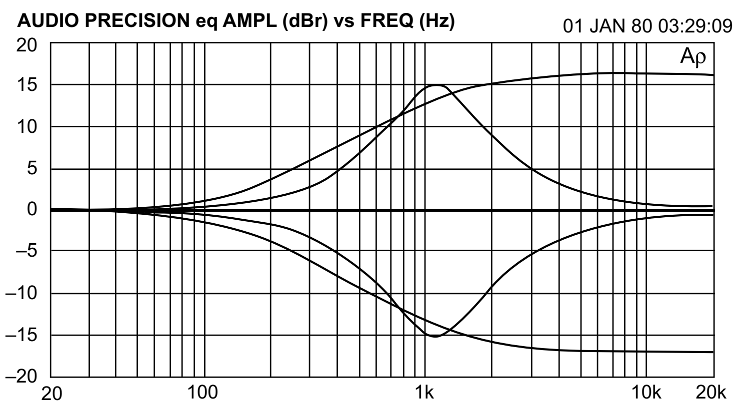

Figure 15.44: The response of the variable-frequency peak/shelving with R7 = 220k; the cut/boost range is thus set to ±15 dB

Parametric middle EQ

A normal second-order resonance is completely defined by specifying its centre frequency, its bandwidth or Q, and the gain at the peak. In mathematical language, these are the parameters of the resonance. Hence an equaliser which allows all three to be changed independently (with a proviso on that coming up soon) is called a parametric equaliser. Upscale mixing consoles typically have two fully parametric middle sections, and usually the LF and HF can also be switched from shelving to peaking mode, when they become two more fully parametric sections.

The parametric middle EQ shown in Figure 15.45 is included partly for its historical interest, showing how opamps and discrete transistor circuitry were combined in the days before completely acceptable opamps became affordable. I designed it in 1979 for a now long-gone company called Progressive Electronics, which worked in a niche market for low-noise mixing consoles. The circuitry I developed was a quite subtle mix of discrete transistor and opamp circuitry which gave a significantly better noise performance than designs based entirely on the less-than-perfect opamps of the day; in time, of course, this niche virtually

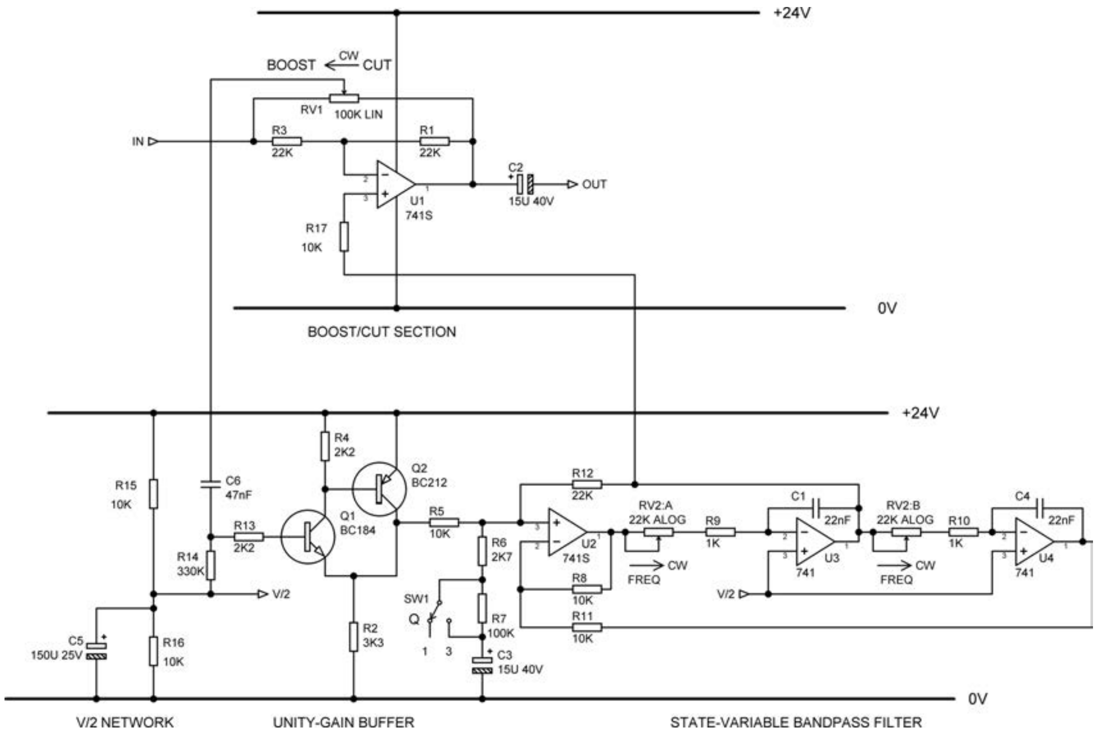

Figure 15.45: An historical parametric middle EQ dating back to 1979

disappeared, as they are wont to do. This parametric middle EQ was used, in conjunction with the usual HF and LF controls, in a channel module called the CM4.

The boost/cut section used an opamp because of the need for both inverting and non-inverting inputs. I used a 741S, which was a completely different animal from the humble 741, with a much better distortion performance and slew rate; it was, however, markedly more expensive and only used where its superior performance was really necessary. The unity-gain buffer Q1, Q2 which ensured a low-impedance drive to the state-variable bandpass filter was a discrete circuit block, as its function is simple to implement. Q1 and Q2 form a CFP emitter-follower. R13 was a 'base-stopper' resistor to make sure that the Q1, Q2 local feedback loop did not exhibit VHF parasitic oscillation. With the wisdom of hindsight, putting a 2k2 resistor directly in the signal path can only degrade the noise performance, and if I was doing it again I would try to solve the problem in a more elegant fashion. The high input impedance of the buffer stage (set by R14) means that C6 can be a small non-electrolytic component.

The wholly conventional state-variable band-pass filter requires a differential stage U2, which once again is best implemented with an opamp, and another expensive 741S was pressed into service. The two integrators U3, U4 presented an interesting problem. Since only an inverting input is required, discrete amplifiers could have been used without excessive circuit complexity; a two or possibly three-transistor circuit (see Chapter 3) would have been adequate. However, the PCB area for this approach just wasn't there, and so opamps had to be used. To put in two more 741S opamps would have been too costly, and so that left a couple of the much-despised 741 opamps. In fact, they worked entirely satisfactorily in this case, because they were in integrator stages. The poor HF distortion and slew rate were not really an issue because of the large amount of NFB at HF, and the fact that integrator outputs by definition do not slew quickly. The indifferent noise performance was also not an issue because the falling frequency response of the integrators filtered out most of the noise. In my designs the common-or-garden 741 was only used in this particular application. Looking at the circuit again, I have reservations about the not inconsiderable 741 bias currents flowing through the two sections of the frequency-control pot RV2, which could make them noisy, but it seemed to work alright at the time.

The filter Q was set by the resistance of R6, R7 to ground. It does not interact with filter gain or centre-frequency. The Q control could easily have been made fully variable by using a potentiometer here, but there was only room on the channel front panel for a small toggle switch. Note the necessity for the DC-blocking capacitor C3, because all the circuitry is biased at V/2 above ground.

The filtered signal is fed back to the boost/cut section through R17, and I have to say that at this distance of time I am unsure why that resistor was present. It could only impair the noise performance. Grounding it would give an EQ-cancel that would also stop noise from the state-variable filter.

It is worth noting that the design dates back to when the use of single-supply rails was customary. In part, this was due to grave and widely-held doubts about the reliability of electrolytic coupling capacitors with no DC voltage across them, which would be the case if dual rails were used. As it happened, there proved to be no real problem with this, and things would have progressed much faster if capacitor manufacturers had not been so very wary of committing themselves to approving non-polarised operation. The use of a single supply rail naturally requires that the circuitry is biased to V/2, and this voltage was generated in the design shown by R15, R16, and C5; it was then distributed to wherever it was required in the channel signal path. The single rail was at +24 V, because 24 V IC regulators were the highest voltage versions available, and nobody wants to get involved with designing discrete power supply regulators if they can avoid it. This is obviously equivalent to a ±12 V dual rail supply, compared with the ±15 V or ±17 V that was adopted when dual-rail powering became universal, and so gave a headroom that was lower by 1.9 dB and 3.0 dB respectively.

This design is included here because it is a good example of making use of diverse circuit techniques to obtain the best possible performance/cost ratio at a given point in time. It could be brought up-to-date quite quickly by replacing all the antique opamps and the discrete unity-gain buffer with 5532s or other modern types.

Modern parametric equalisers naturally use all-opamp circuitry. Figure 15.46 shows a parametric EQ stage I designed back in 1991; it is relatively conventional, with a three-stage state variable filter composed of A2, A3 and A4. There is however an important improvement on the standard circuit topology. Most of the noise in a parametric equaliser comes from the filter path. In this design the filter path signal level is set to be 6 dB higher than usual, with the desired return level being restored by the attenuator R6, R7. This attenuates the

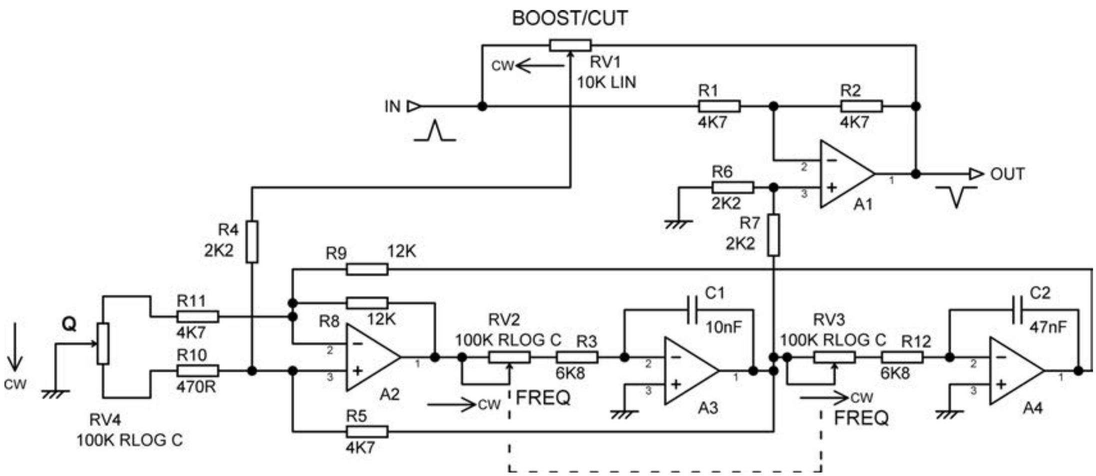

Figure 15.46: Variable-freq variable-Q middle control. State-variable type

filter noise as well, and the result is a parametric section approximately 6 dB quieter than the industry standard. The circuit is configured so that, despite this raised level, clipping cannot occur in the filter with any combination of control settings. If excess boost is applied, clipping can only happen at the output of A1, as usual.

Other features are the Q control RV4, which is configured to give a wide parameter range without affecting the gain. Note the relatively low values for the DC feedback resistors R1 and R2, chosen to minimise Johnson noise without causing excessive opamp loading.

This equaliser section has component values for typical low-mid use, with centre frequency variable over a wide range from 70 Hz to 1.2 kHz, and Q variable from 0.7 to 5. The cut and boost range is the usual ±15 dB

Graphic equalisers

Graphic equalisers are so-called because their cut/boost controls are vertical sliders, the assumption being that a graph of the frequency response will pass through the slider knob positions. Graphic equalisers can have any number of bands, from three to 31, the latter having bands one-third of an octave wide. This is the most popular choice for serious room equalisation work as bands one-third-octave wide relate to the perceptual critical bands of human hearing.

Graphic equalisers are not normally fitted to large mixing consoles, but are often found on smaller powered mixers, usually in the path between the stereo mix and the power amplifiers. The number of bands provided is limited by the space on the mixer control surface, and is usually in the range seven to ten.

There is more than one way to make a graphic equaliser, but the most common version is shown in its basic concept in Figure 15.47, with some typical values. L1, C1 and R3 make up an LCR series resonant circuit which has a high impedance except around its resonant frequency; at this frequency the reactances of L1, C1 cancel each other out and the impedance to ground is simply that of R3. At resonance, when the wiper of RV1 is at the R1 end, the LCR circuit forms the lower leg of an attenuator of which R1 is the upper arm; this attenuates the input signal and a dip in the frequency response is therefore created. When RV1 wiper is at the R2 end, an attenuator is formed with R2 that reduces the feedback factor at resonance and so creates a peak in the response. It is not exactly intuitively obvious, but this process does give symmetrical cut/boost curves. At frequencies away from resonance the impedance of the RLC circuit is high and the gain of the circuit is unity.

The beauty of this arrangement is that two, three or more LCR circuits, with associated cut/boost pots, can be connected between the two opamp inputs, giving us an equaliser with pretty much as many bands as we want. Obviously, the more bands we have, the narrower

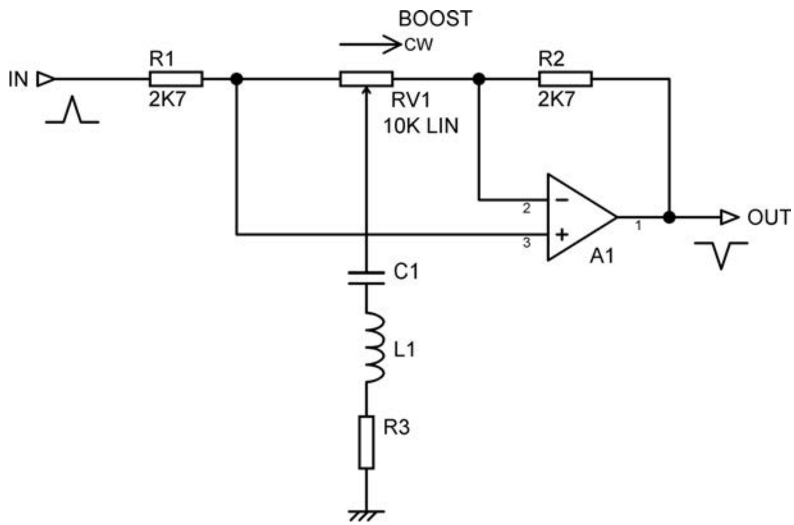

Figure 15.47: The basic idea behind graphic equalisers; gain is unity with the wiper central

they must be to fit together properly. A good example of a classic LCR graphic design was published by Reg Williamson in 1973 [13].

As described in Chapter 2, inductors are in general thoroughly unwelcome in a modern design, and the great breakthrough in graphic equalisers came when the LCR circuits were replaced by gyrator circuits that emulated them but used only resistors, capacitors and a gain element. It is not too clear just when this idea spread, but I can testify that by 1975 gyrators were the standard approach, and the use of inductors would have been thought risible.

The basic notion is shown in Figure 15.48; C1 works as a normal capacitor as in the LCR circuit, while C2 pretends to be the inductor L1. As the applied frequency rises, the attenuation of the high-pass network C2–R1 reduces, so that a greater signal is applied to unity-gain buffer A1 and it more effectively bootstraps the point X, making the impedance from it to ground increase. Therefore we have a circuit fragment where the impedance rises proportionally to frequency – which is just how an inductor behaves. There are limits to the Q values that can be obtained with this circuit because of the inevitable presence of R1 and R2.

The sample values in Figure 15.48 synthesise a grounded inductor of 100 mH (which would be quite a hefty component if it was real) in series with a resistance of 2 kΩ. Note the surprisingly simple equation for the inductor value. Another important point is that the opamp is used as a unity-gain buffer, which means that the early gyrator graphic equalisers could use a simple emitter-follower in this role. The linearity was naturally not so good, but it worked and made graphic equalisers affordable.

A simple seven-band gyrator-based graphic equaliser is shown in Figure 15.49. The maximal cut/boost is ±8 dB. The band centre-frequencies are 63 Hz, 160 Hz, 410 Hz, 1 kHz, 2.5 kHz, 7.7 kHz and 16 kHz. The Q of each band at maximum cut or boost is 0.9.

The response of each band is similar to that shown in Figure 15.33 above. The maximum Q value is only obtained at maximum cut or boost. For all intermediate settings the Q is lower. This behaviour is typical of the straightforward equaliser design shown here, and is usually

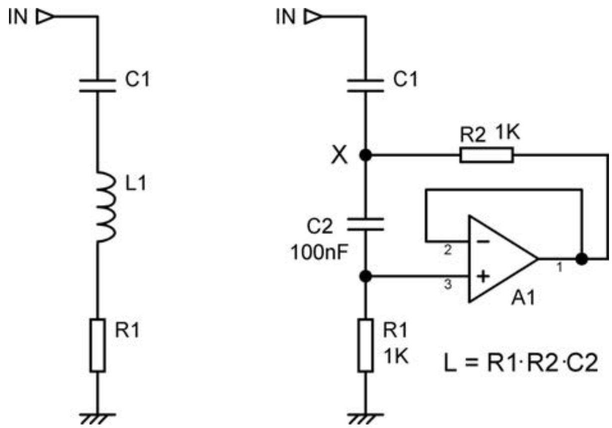

Figure 15.48: Using a gyrator to synthesise a grounded inductor in series with a resistance

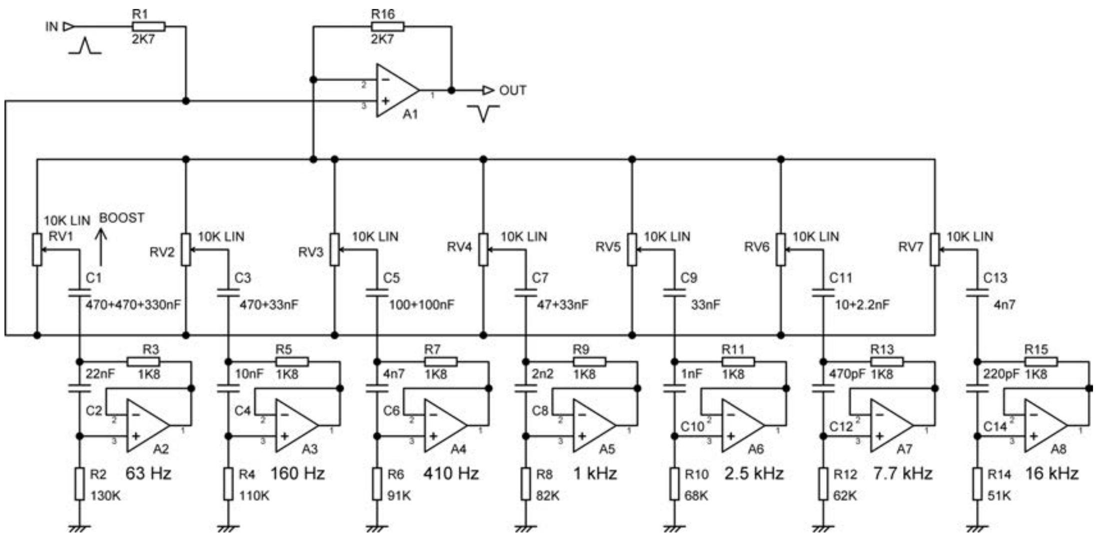

Figure 15.49: A 7-band graphic equaliser

referred to as ‘proportional-Q’ operation; it results in a frequency response which is very different from what might be expected on looking at the slider positions.

There is however another mode of operation called ‘constant-Q’ in which the Q of each band does not decrease as the cut/boost is reduced [14]. This gives a frequency response that more closely resembles the slider positions.

The graphic equaliser described here has a symmetrical response, also known as a reciprocal response; the curves are the same for cut and for boost operations. It is also possible to design an equaliser for an asymmetric or non-reciprocal response, in which the boost curves are as shown above but the cut response is a narrow notch. This is often considered to be more effective when the equaliser is being used to combat feedback in a sound reinforcement system.

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CHAPTER 16

Mixer architectures

Introduction

A large mixing console arguably represents the most demanding area of analogue audio design. The steady advance of digital media demands that every part of the chain that takes music from performer to consumer must be near perfect, as the comfortable certainty that everything will be squeezed through the quality bottleneck of either analogue tape or vinyl disc now looks very old-fashioned.

The technical problems that must be overcome in a professional mixing console are many and varied, and often unique to console design. A large number of signals flow in a small space and they must be kept strictly apart until the operator chooses to mix them; crosstalk must be exceedingly low.

There may be 72 input channels or more, each with many stages that all have the potential to add distortion and noise to the precious signal. Even summing these signals together, while sounding trivially easy, is in practice a major challenge. The quality requirements, especially for recording consoles, are much more demanding than those for the most expensive hifi equipment, because degradation introduced at this stage can never be retrieved.

Performance factors

Two primary requirements of modern consoles are very low noise and minimal distortion. Since a comprehensive console must pass the audio through a large number of circuit stages (perhaps over 100 from microphone to final mixdown) great attention to detail is essential at each stage to prevent a build-up of noise and distortion; one of the most important trade-offs is the impedance of the circuitry surrounding the opamp, for if this is too high Johnson noise will be increased, while if it is too low an opamp will exhibit non-linearity in struggling to drive it.

The choice of opamp is also critical, for cost considerations discourage the global use of expensive ICs. In a console with many stages of signal processing, this becomes a major concern; nonetheless it is possible to keep the right-through THD below 0.002% at 20 dB above the normal operating level, so that at normal levels it is unmeasurable.

I will use the term mixer, mixing console, and mixing desk interchangeably where it will make the text less clunky. In the USA it is sometimes called simply ‘The Board’. I hope I can put ‘mixer’ without anyone thinking I am referring to the front end of a radio or a kitchen appliance, or indeed anything concrete-related. Throughout this chapter the various sections of the mixer – the channels, groups, master etc, will be referred to as ‘modules’, even if they are not constructed as separate physical modules.

In hifi, the preservation of absolute phase has only recently come to be considered necessary; with normal listening material it appears highly unlikely that it actually makes any audible difference, but it unquestionably looks good on the spec. In mixer design the meticulous preservation of phase is a rigid requirement, to avoid the possible cancellation of signals. Both summing amplifiers and most EQ sections use shunt feedback and thus invert the signal, and it is sometimes necessary to include inverting stages that do nothing but get the phase straight again. It is a mark of good design to keep such stages to a minimum.

Mixer internal levels

The internal signal levels of any mixing console are always a compromise between noise and headroom. A higher level means that the signal is less compromised by circuit noise, but makes inadvertent overload more likely, and vice versa. The levels chosen depend on the purpose of the console. If you are recording you only have to get it right once, but you do have to get it exactly right, i.e. with the best possible signal-to-noise ratio, so the internal level is relatively high, very often −2 dBu (615 mV rms) which gives a headroom of about 22 dB. In broadcast work to air you only have one chance to get it right, and a mildly impaired signal-to-noise ratio is much preferable to a crunching overload, especially since the FM or DAB channel has limited noise performance anyway, so the internal levels are considerably lower. The Neve 51 Series broadcast consoles used −16 dBu (123 mV rms) which gives a much increased headroom of 36 dB at the cost of noise performance. At least one manufacturer has used balanced mixing buses to reduce the impact of the lower internal levels on signal-to-noise ratio.

As a sidelight on this issue, I might mention a mixer design I worked on some years ago. The decision was taken (no, not by me) to use an internal level of −6 dBu instead of the usual −2 dBu to improve headroom in live situations, the assumption being that 4 dB greater internal noise would go unnoticed. It most certainly did not, and vociferous complaints during beta-testing led to a rapid redesign to convert it back to the old −2 dBu level. Fortunately we were ready for this one, and the design had been arranged so that the internal level could be altered simply by changing resistor values.

An internal level of −2 dBu does have one unique advantage; doubling the level, as is done by most forms of balanced output stage, gets you directly to the professional +4 dBu level.

Mixer architecture

A mixing console has a number of input channels that can be summed into groups, and these groups again summed into a stereo mix as the final result. Small mixers often have no groups and mix directly to stereo.

The structure and signal flow of a mixing console depends very much on its purpose. A recording console needs track return sections to allow the creation of a monitor mix so that as a track is recorded, those already done can be played back at roughly the same levels they will have in the final mix; this is essential so new material can be synchronised with that already in existence. While it is quite possible to use a recording-oriented console for PA work, the reverse is much more difficult, and usually impractical.

A sound-reinforcement (PA) console does not need monitor sections, so it can be somewhat simpler, but it typically is fitted with a large number of effects returns. The output from a PA console at a large event is not likely to be in stereo – there will be multiple speakers each of which requires an individual feed with different equalisation and delay requirements; often there is a centre output.

In PA work there are often two mixers. The front-of-house (FOH) mixer usually resides in the middle of the audience or in a control-room at the back. If the mixer is in the audience, the less room it takes up the better, as it is occupying seats that could have been paid for. Size is also very much an issue for mixers intended for installation in a control-room, as access to these is often cramped and difficult.

The second type of PA mixer is the so-called monitor mixer. PA work makes much use of 'monitors' i.e. small speakers, often wedge-shaped, that are placed close to the musician so that they can hear their output clearly without straining to pick it out of the main mix. Creating these foldback feeds on a separate mixer where basically each channel consists of a bank of aux sends (from 12 to 32 may be available) gives much greater flexibility than relying on the relatively few aux sends available on the FOH desk. The inputs to the monitor mixers are taken from the inputs to the FOH console by a splitter box, usually transformer-based, at the stage end of the 'snake' cable to the FOH console. Traditionally monitor mixers were placed just out of sight at the side of the stage, so that disgruntled performers could signal to the hapless monitor engineer that the foldback mix was not to their liking. The widespread adoption of intercoms has given more flexibility in monitor desk placement.

Broadcast consoles are more specialised in their layout and no one expects to be able to use them for recording or PA applications.

Recording consoles come in two different formats. The first and more traditional is called a split, separate-group, or European-style arrangement and has input channels arranged in one bank, typically to the left of the master section, with the groups banked to the right of it; on a large console there may be a further set of channels to the right of the groups.

The second format is called the in-line arrangement, and has the channels and groups built into the same module. Normally there is a row of channel/group modules to both the left and right of a central master section. In-line consoles are more compact and make more efficient use of the electronic facilities, but they are conceptually more complex and somewhat harder to use. They are not popular in PA work where you need to do things quickly, in real-time, without juggling a block-diagram in your head.

In the sections that follow we will look at the internals of generic mixers that do not follow any particular design but are configured to bring out as many instructive points as possible.

The split mixing architecture

Figure 16.1 shows the basic arrangement of a split (separate-group) mixer. Its design is oriented to recording, but like many small and medium sized mixers it could also be used effectively for PA work.

Each channel has both a microphone and line-level input, with a wide-ranging gain control. Once it has been raised to the nominal internal level of the console, the signal is processed by the equalisation section, controlled in level by the channel fader, and then routed to a group or pair of groups, or directly to the stereo L–R bus. The position of the channel signal in the overall stereo image is set by the panoramic potentiometer (universally known as a panpot) that controls the proportion of the signal sent to left and right. The channel also has auxiliary send controls that feed the channel signal to the auxiliary buses; sends taken off before the fader are called prefade and are normally used for foldback, i.e. helping the musician hear the sound they are producing. Sends taken off after the fader, so that their level is dependant on the fader position, are called postfade sends and are normally used for adding effects such as reverberation that are expected to vary in level with the original signal.

The final feed from the channel to the buses is the prefade-listen (PFL); when the PFL switch is pressed, the prefade channel signal is sent to the master module via the PFL bus, where it automatically replaces the normal stereo mix feed to the control-room monitor (CRM) loudspeakers.

The group/monitor modules contain both the group section and one or two monitor sections. The group consists of a summing amplifier that collects together all the signals sent to its group bus, and passes it to the outside world via a fader and an output amplifier. When recording, this group signal is sent to one track of the recording machine. The monitor sections allow tracks which are already recorded to be played back to the stereo mix bus via a level control and a panpot, so that a rough idea of what the final mix will sound like can

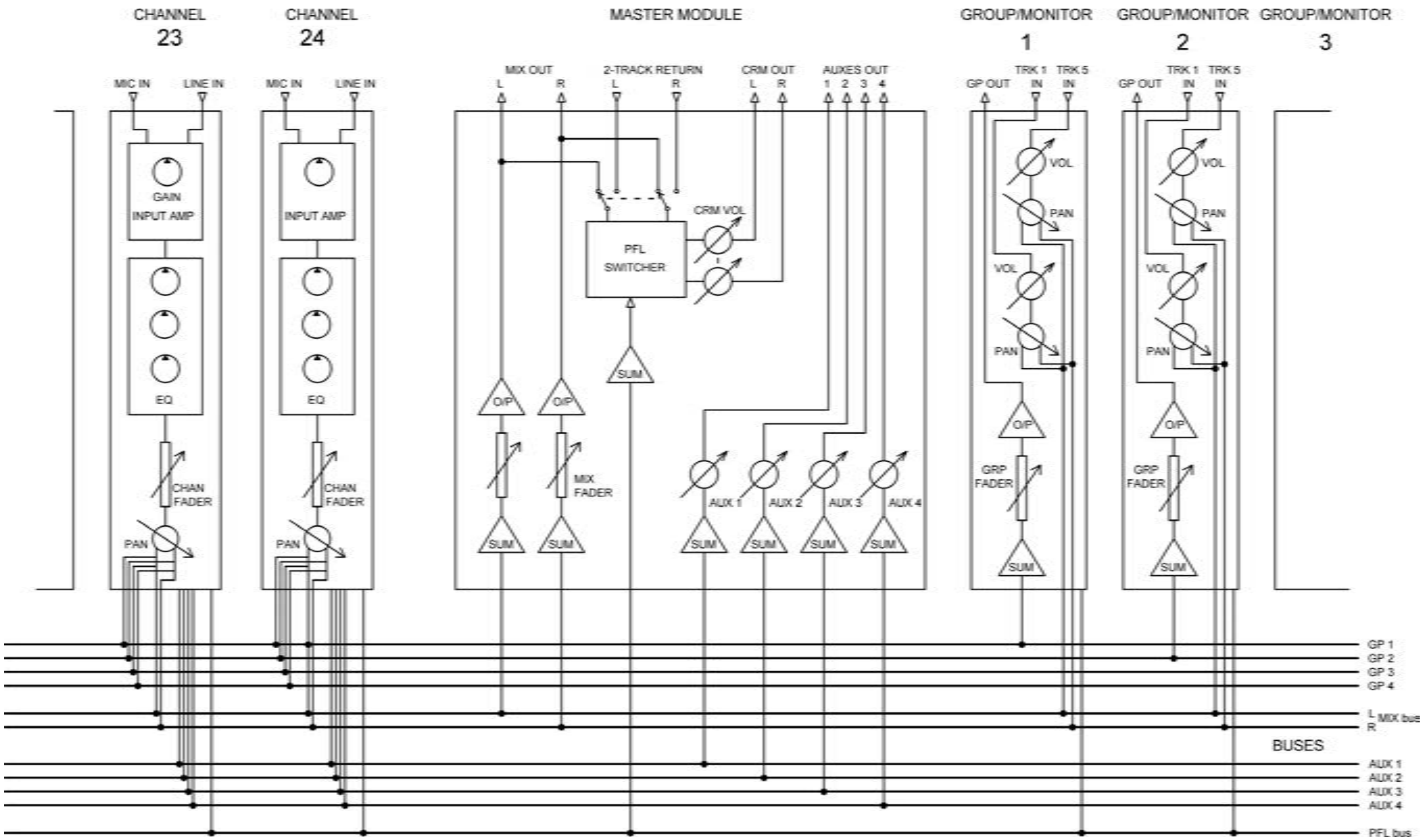

Figure 16.1: Block diagram of a split-format mixer with four groups and four aux sends

be set up as early in the recording process as possible. One of the monitor sections will be switchable between the group output and the track return so the monitor mix can be made up of tracks that are being recorded as well as those which have already been laid down. This group/track switching is not shown in Figure 16.1, to aid clarity, but is fully shown in Figure 16.4 below. Lack of space means that the monitor return sections usually have only simple EQ or none at all, but they almost always have a prefade aux send for foldback purposes. A postfade send is also extremely useful as the rough monitor mix is much more convincing if reverberation can be added as appropriate; this is called ‘wet’ monitoring. The monitor sections often each have a PFL switch.

When recording is finished, and mixdown begins, the full facilities of the channels are required, and so the track return signal is now sent to the channel line input and the channel routed to the stereo mix bus. This is often a connection normalled through the line input jack, so that the input amplifier receives the track return signal at all times unless a jack is inserted. Back in the days of tape machines, this facility was called ‘tape normalling’ but now ‘track return normalling’ describes it rather better. The monitor sections in the group/monitor modules are now no longer needed, but since they are routed to the stereo mix bus they can be used as extra effects returns.

The master section contains the summing amplifiers for the stereo mix bus, which feed a stereo fader and suitable output amplifiers. These provide the final feed to the two-track recording device at mixdown. Normally, this stereo output is also sent to a pair of meters and the control-room monitor (CRM) loudspeakers, the latter via a volume control, but the meters and CRM speakers can also be switched to monitor the return from the two-track recorder. When any PFL switch on the console is pressed, this is detected and the PFL signal (in mono) is fed to the meters and CRM speakers instead, so the signal quality and level from any channel can be assessed.

The master section also contains the summing and level controls for the aux send buses, and may also incorporate talkback facilities and a line-up oscillator; these features are dealt with in more detail later in this chapter.

The in-line mixing architecture

Figure 16.2 shows the basic arrangement of an in-line mixer, essentially intended for recording. There are now no group/monitor modules, their functions being performed by what might be called channel/group/monitor modules. Only four group buses are shown for clarity, but in practice there would normally be a much higher number, to make use of the in-line format.

The group summing amplifier is now part of the channel. It no longer has its own fader; instead the summing amp gain is controlled by a rotary ‘bus trim’ control so overload can avoided. This is normally placed out of the way at the very top of the module. The group signal is sent out

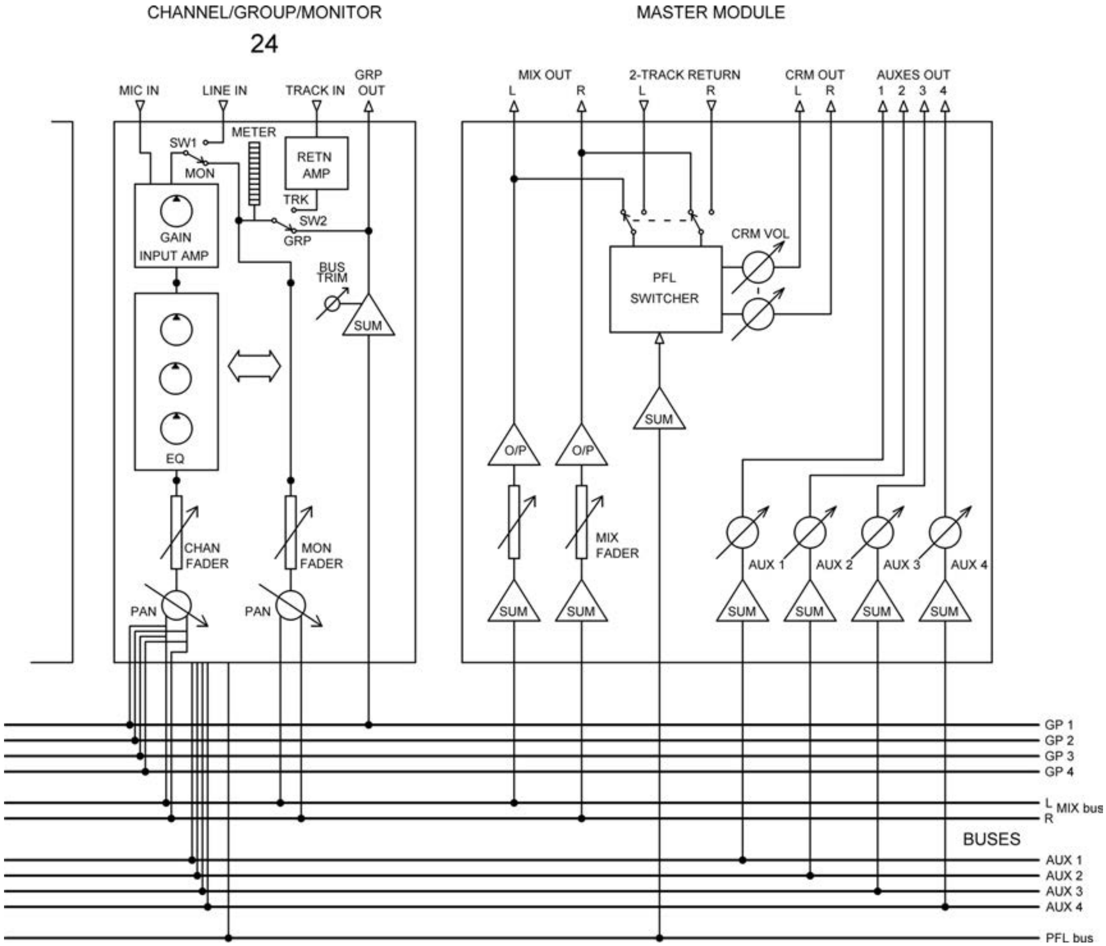

Figure 16.2: Block diagram of an in-line mixer with N groups and four aux sends

to the recording machine, and the track return comes back in, usually via a balanced amplifier. A track/group switch SW2 selects either the group or the track return signal for the metering and the monitor path; as shown in the figure it consists only of a monitor fader and a monitor panpot which send the signal to the stereo mix bus to create a monitor mix. The monitor fader is normally a short fader mounted on the channel, as opposed to the long channel fader which is right at the front of the console. Sometimes a rotary monitor fader is fitted.

There are two slightly differing approaches to the operation of an in-line module:

- In the first approach, at mixdown switch SW1 now selects the track return signal rather than the line input, and the signal reaches the stereo mix bus via the full facilities of the channel. Sometimes additional switching allows the monitor path to be used as an extra effects return, as in split consoles. Typically the line input jack is pressed into service as the input connector for this purpose.

- An alternative design approach is not to switch the source of the channel path to the track return at mixdown, but instead to move the whole EQ section from the channel to the monitor path. This means that the short monitor fader would have to be used for level adjustment, which is not ideal when there is a long fader right in front of you. This has particular force when fader automation is fitted, for it is normally only applied to the long fader. A switch called a 'fader flip' is therefore fitted which swaps over the function of the two faders, allowing the long fader to be used at mixdown. This sort of approach involves a lot of reconfiguration when switching from record to mixdown mode, and on the more sophisticated consoles this is handled entirely by electronic switching, with global control from the master section, so that only one switch needs to be pressed to change the console mode. This second method always seems to me to be unnecessarily complex, but it has seen extensive use.

The master section shown in Figure 16.2 is identical to that shown for the split console in Figure 16.1.

A closer look at split format modules

We will now look in more detail at the channel, group and master modules that make up a split mixing console:

The channel module (split format)

Figure 16.3 shows a typical input channel for a relatively simple split-format mixing console. The input stage provides switchable balanced mic and line inputs; the mic input has an impedance of 1–2 kΩ, which provides appropriate loading for a 200 Ω mic capsule. If a +48 V phantom power facility to power microphones is provided, it is independently switchable on each channel. The line input will have a bridging impedance of not less than 10 kΩ. The mic preamplifier in particular will have a wide range of gain, such as 0 to 70 dB, while the line input tends to have a more restricted range such as +20 to −10 dB. The track return from the recording machine is shown connected through to the line input via the normalling contact of the line jack socket, so that mixdown with the full channel facilities can begin as soon as the mic/line switch is set to 'line', so long as no jack has been inserted into the line input to break the normalling contact. The line input is shown as unbalanced in Figure 16.3 for clarity, but in practice it would usually be a balanced input using the tip and ring connections of the jack socket.

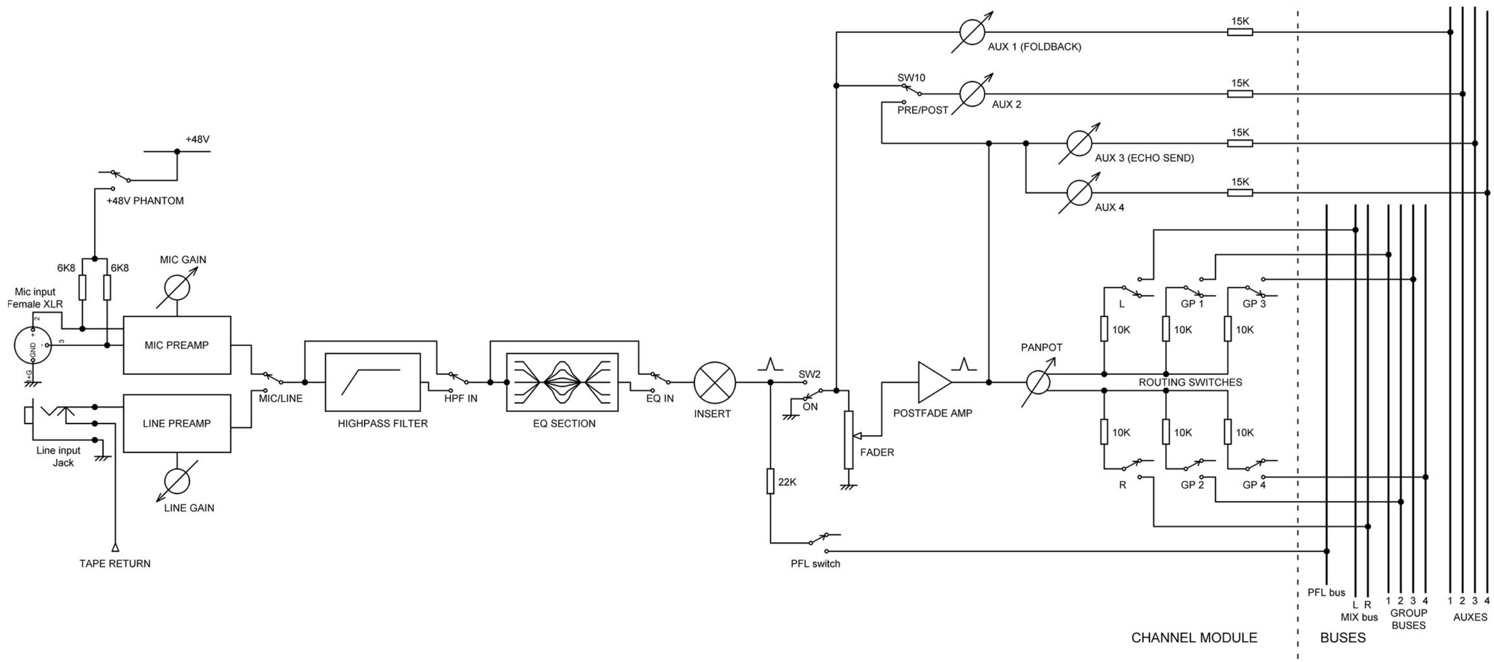

Figure 16.3: Block diagram of a typical channel module for a small mixer

The input stage is followed immediately by a switchable high-pass filter (usually −3 dB at 100 Hz) to remove low-frequency disturbances picked up by the microphone as soon as possible; if not filtered out these can eat up headroom. The filter is usually second or third order, giving a roll-off of 12 or 18 dB per octave respectively.

The tone-control section (universally known in the pro audio business as 'EQ' or equalisation) typically includes one or more mid-band resonance controls as well as the usual shelving Baxandall-type high and low controls. On all but the simplest mixers, the EQ section can be switched out to allow before/after comparisons.

The larger and more sophisticated consoles incorporate dynamics sections into each channel. This is not shown in Figure 16.3 to aid clarity. The dynamics facilities available vary but usually include compression, limiting, and sometimes noise-gating; some consoles have been produced with just the noise-gate function, as it is easier to fit the required electronics into a limited space. A perennial problem with this sort of thing is finding panel room for the extra controls required; the permissible length of a module is limited by the reach of the human arm.

Next comes the insert point, though in some designs it may be configured so it can be placed ahead of the EQ and dynamics section instead. This is a jack or pair of jacks that allow external processing units to be plugged into the signal path. When nothing is plugged in, the 'normalling' contacts on the jack socket allow the signal to flow through unchanged.

The PFL (prefade-listen) switch routes the post-insert signal to the master module and the monitor speakers independently of all other controls; a PFL-detect bus signals the master module to switch the studio monitoring speakers from the normal stereo mix bus to the PFL bus.

Next in the chain comes the channel on switch. This may be either a simple mechanical switch or an electronic mute block. Note the PFL feed is taken off *before* the 'on' switch so the channel signal is always accessible. The channel level in the mix is controlled by a linear fader with a postfade amplifier typically giving 10 dB of gain, and the panpot sets the stereo positioning, odd group numbers being treated as left, and even as right.

The channel shown has three auxiliary sends. The auxiliary sends of a console represent an extra mixing system that works independently of the main groups; the number and configuration of these sends have a large effect in determining the overall versatility of the console. Each send control provides a feed to a console-wide bus; this is centrally summed and then sent out of the console. Sends come essentially in two kinds. Prefade sends are taken from before the main channel fader, and are therefore independent of its setting. Post-fade sends take their feed from after the fader, so that the send level falls or rises according to the fader setting.

Prefade sends are normally used for 'foldback'; i.e. sending the artist a headphone feed of what he/she is perpetrating, which is important if electronic manipulation is part of the creative process, and essential if the artist is adding extra material that must be in time with that already recorded. In the latter case, the existing tracks are played back to the artist via the prefade sends on the monitor sections.

Postfade sends are used as effects sends; their source is after the fader, so that the effect will be faded down at the same rate as the untreated signal, maintaining the same ratio. The sum of all feeds to a given bus is sent to an external effects unit and the output of this returned to the console. This allows many channels to share one expensive device, such as a high-quality digital reverb, and for this sort of purpose is much more appropriate than patching processors into the channel insert points.

There may be anything from one to twelve or more sends available. In the example shown, the first send (Aux 1) is a dedicated prefade send which is typically used for foldback. The third (Aux 3) is a dedicated postfade send, typically used for effects such as reverb. The second send (Aux 2) can be switched so it is either prefade or postfade. In a recording console the greatest need for foldback sends is during track laying, while effect sends are in greatest demand at mixdown; it is therefore common for some or all of the sends on a channel to be switchable prefade or postfade.

On more sophisticated consoles, it is often possible to switch every auxiliary send between pre and post to give maximal flexibility. Traditionally, this meant pressing a switch on every input module, since it is most unlikely that a mixture of pre- and post-sends on the same bus would be useful for anything. On a 64-input console this is a laborious process. More advanced designs minimise the effort by setting pre/post selection for each auxiliary bus with a master switch that controls solid-state pre/post switching in each module. An example of this approach is the Soundcraft 3200 recording console.

Effect return modules

In complex mixdowns, it may be necessary to return a large number of effects to the mix. While, as described above, it is often possible to press unused monitor sections or channel modules into service as effect returns, sometimes this just does not provide enough and specialised 'effects return' modules may be fitted. These usually have facilities intermediate between the channels and the monitor sections, and it is common to fit two, and sometimes even four, into the space occupied by a channel module. The returned effect, which may well now be in stereo (such as the output of a digital reverb unit, for example) is usually added to the stereo mix bus via level and pan controls.

The group module

Figure 16.4 shows a typical group module for a recording desk. Each mix bus has its own summing amplifier; the summed group signal is then sent to an insert so external processing

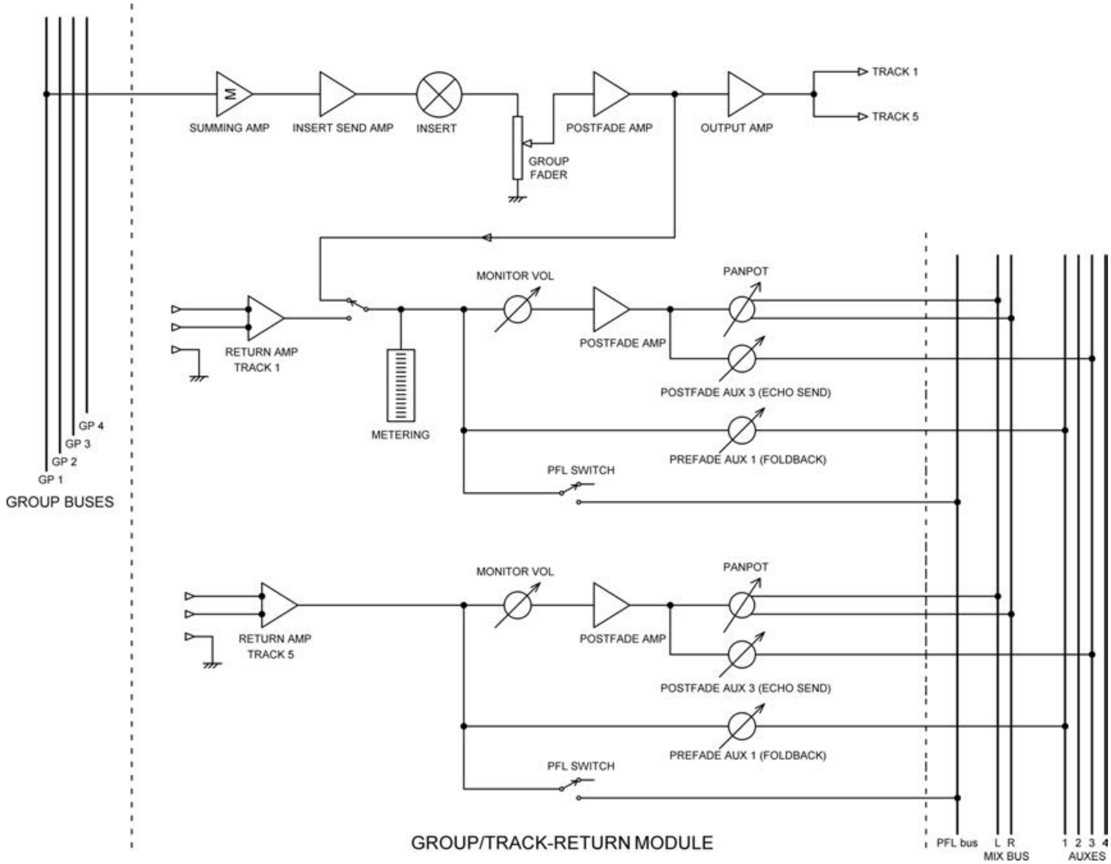

Figure 16.4: Block diagram of a typical group module for a small recording mixer

can be applied. The insert send is very often arranged for ground-cancelling operation, for the simple reason that the summing amp inherently phase-inverts the signal, which must be returned to the correct polarity by another amplifier stage before it meets the outside world. This second stage can be arranged to be ground-cancelling at minimal expense. The operation of ground-cancelling outputs is described in Chapter 19.

The signal from the insert return goes to the group fader. Like that of the channel, it is a linear fader with a postfade amplifier typically giving 10 dB of gain. The signal then goes to an output amplifier suitable for driving the recording device, possibly via lengthy cables. On all but the cheapest designs this is a balanced output.

The second major function of the typical group module is to allow the creation of a monitor mix so that as a track is recorded, those already done can be played back at roughly the same levels they will have in the final mix; this is obviously essential to allow new material to be synchronised with that already in existence. Back in the days of tape recording machines, the monitor playback had to be done by the record head, as the playback head was physically displaced and so would have given a delayed signal; this meant the quality of the monitor mix was compromised as the record head was optimised for recording and not playback. With digital recording this is of course not a problem.

The monitor section typically consists of a balanced line input to prevent ground-loop problems with the recorder, some form of switching between the group and recorder return (so the track being recorded can be incorporated into the monitor mix) and level and panpot controls. EQ is not often fitted for reasons of space, but aux sends for foldback and effects are standard, because adding reverb to the monitor mix makes it much more realistic. A PFL switch for each monitor section is usually provided.

Normally relatively few tracks are recorded at a time, so it is sensible to have more monitor sections than groups. For example, a 4-group mixer can be very effectively used with an 8-track recorder if Group 1 is connected to Track 1 and Track 5, Group 2 is connected to Track 2 and Track 6, and so on. The track to be recorded is selected at the recorder. To make use of this, each group module will contain one actual group section and two monitor sections, which explains why there is not normally room for even basic EQ in the monitor sections.

The master module

Figure 16.5 shows the basics of a typical master section. It contains the summing amplifiers for the stereo mix bus, with their associated insert send amplifiers and insert points. This is followed by a stereo fader (sometimes implemented as two mono faders with the knobs mounted immediately next to each other), a +10 dB post amp, and a balanced or ground-cancelling output stage which feeds the two-track recorder during mixdown.

The stereo mix bus normally drives the L–R meters and the control-room monitor (CRM) loudspeakers, but a manual source-select switch allows this feed to be replaced by the return signal from the two-track recorder for quality checking of the final stereo recording. Whenever a PFL switch anywhere on the console is pressed, the PFL-detect system responds and activates two solid-state switches that replace the stereo monitor signal with the PFL signal. The L–R meter feeds are taken off after the PFL-switched so those meters can be used to check the level at whichever point in the system the PFL switch in question happens to be.

The master section also contains the summing and level controls for the aux send buses. The auxes may have dedicated meters or just a PFL switch each.

The master module will also carry any master status switches for globally changing things such as group/recorder switching for the monitor sections, pre/post operation of auxes, and so on.

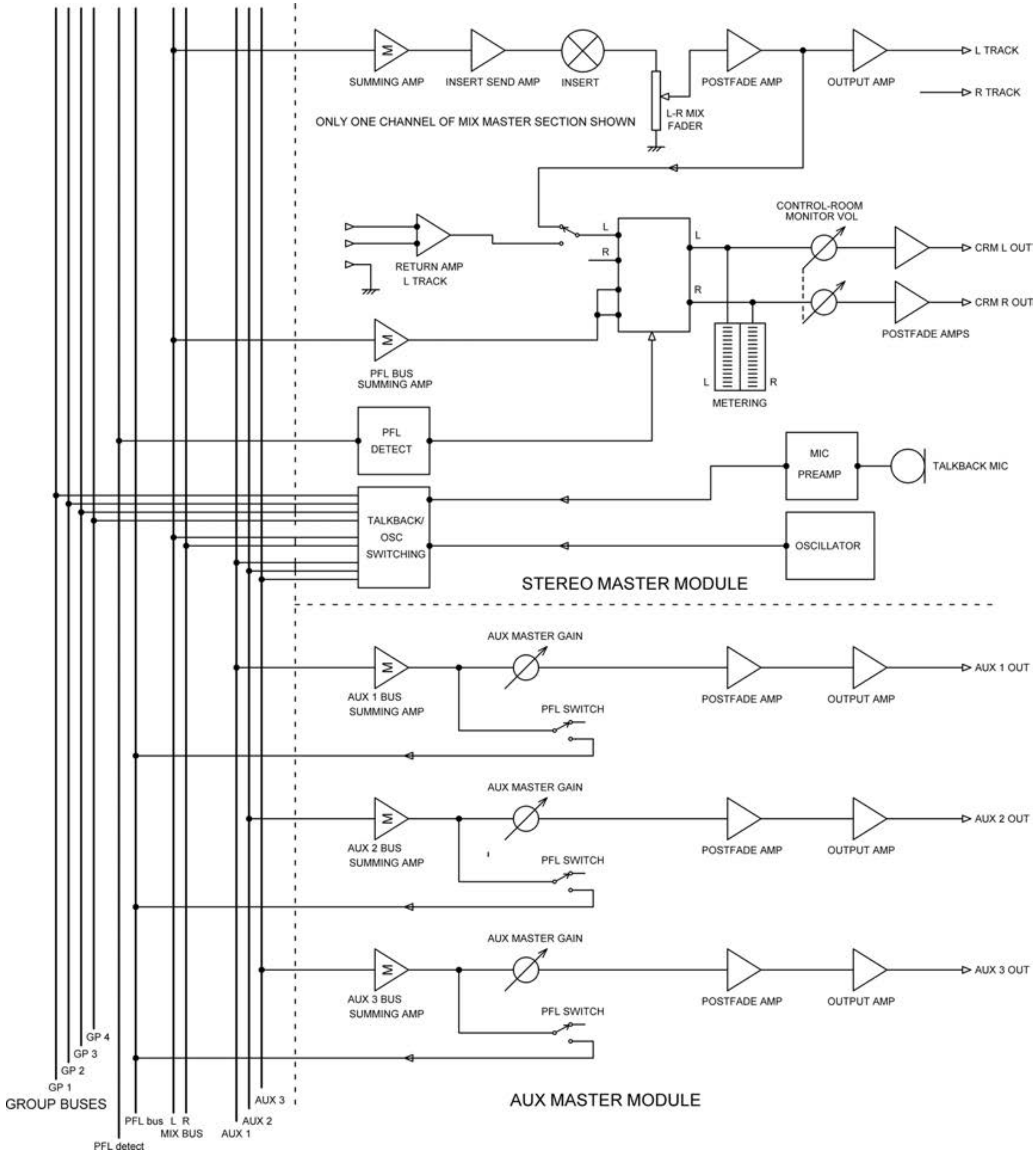

Figure 16.5: Block diagram of a typical master module for a small mixer

Talkback and oscillator systems

The talkback system allows the mixing engineer to talk to the musicians in the studio or add spoken comments to the recording, using a microphone mounted on the console. Back in the day it was considered cool to mount the talkback microphone on a flexible gooseneck, but these get in the way and the modern approach is to have a small electret microphone mounted flush with the master module panel.

For talkback to the studio the microphone feed is routed to the aux buses, on the assumption that one or more of these will be in use for foldback and so will be connected to a loudspeaker in the studio area. Routing may also be provided to a dedicated talkback loudspeaker. In many cases there is a facility to route to the first two auxes only, as these will almost certainly be used for foldback purposes (see the typical system in Figure 16.6).

For recording identification, the microphone feed is routed to the group buses. This facility, which allows the engineer to identify a recording by saying something like 'Spinal Tap- take 147' is often called a 'slate' facility, by analogy with the film industry.

When the talkback facility is used, there is a danger of acoustical feedback. If a microphone in the studio is active and routed through channel and group to the control-room monitors, the monitor speaker output will be picked up by the talkback microphone, fed to the studio . . . and away she goes. It is therefore normally arranged that pressing any of the talkback switches will attenuate or cut completely the monitor speaker signal. On larger consoles the amount of attenuation (or 'dim' as it is usually called in this context) is adjustable.

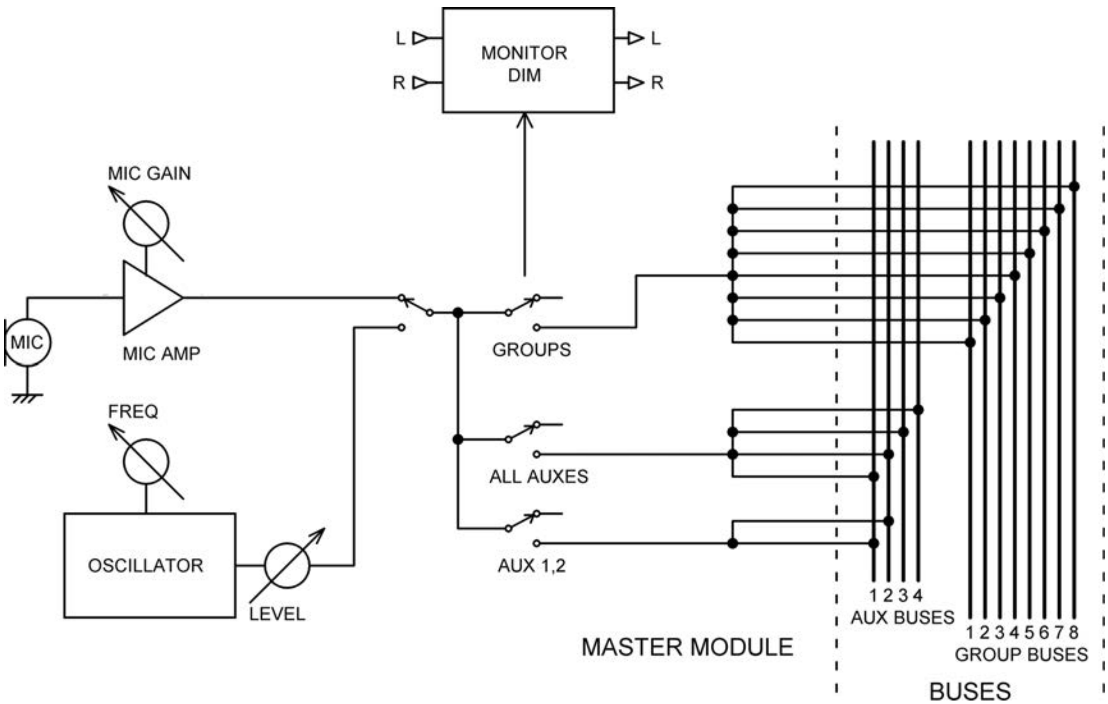

Figure 16.6: Talkback and oscillator system with routing to aux and group buses

If the talkback microphone is mounted in the master panel, the distance between it and the engineer will vary as he moves. More sophisticated consoles sometimes incorporate a talkback compressor to reduce the level variations.

Many of the larger recording consoles include a 'listen' facility whereby the control-room can hear messages from the studio, even if no microphone channels are faded up. A microphone is mounted somewhere out of the way (often suspended from the ceiling) and is routed to the monitors when the 'listen' switch is pressed.

Most mixers of medium size and above incorporate a line-up oscillator which can be routed to all the groups and the stereo mix. This allows the mixer metering to be lined up with external level indication on the recorder. A fixed frequency oscillator running at 1 kHz with a variable output level is the minimum facility; more advanced models have switched or fully-variable frequency controls.

In the typical talkback/oscillator system of Figure 16.6, the line-up oscillator is shown sharing with the talkback system the bank of mix resistors that inject signal into the buses. On a large console this is quite a lot of resistors, and it would not be sensible to design them in twice. The only compromise is that it is not possible to use talkback and the oscillator at the same time, but then why would you want to? The default selection is talkback, because this is used much more frequently than the oscillator.

The in-line channel module

Figure 16.7 shows a typical channel/group/monitor module for an in-line mixing console. This uses the first approach to in-line operation described above, where to enter mixdown mode the source of the channel path is switched to the track return rather than bodily moving the EQ and other facilities into the monitor path.

The group fader no longer appears. More usefully, the summing amplifier gain can be varied by a rotary 'bus trim' control to prevent overload; it is highly desirable that this control alters the gain of the actual summing amplifier, rather than having a low-gain summing amplifier with a level control and post amplifier following it, because the former gives greater protection against clipping and superior noise performance at low gain settings. The bus trim is rarely altered so it can be conveniently placed out of the way at the very top of the in-line module. The group signal from the summing amplifier is sent out to the recording machine, usually by means of a balanced or ground-cancelling output stage to prevent ground loops occurring with the multi-track recorder. The track return usually comes back via a balanced amplifier for the same reason. A track-return/group switch selects either the group or the track return signal for the metering and the monitor path; as shown in the figure this path consists of a monitor fader (the 'short fader') and a monitor panpot which send the signal to

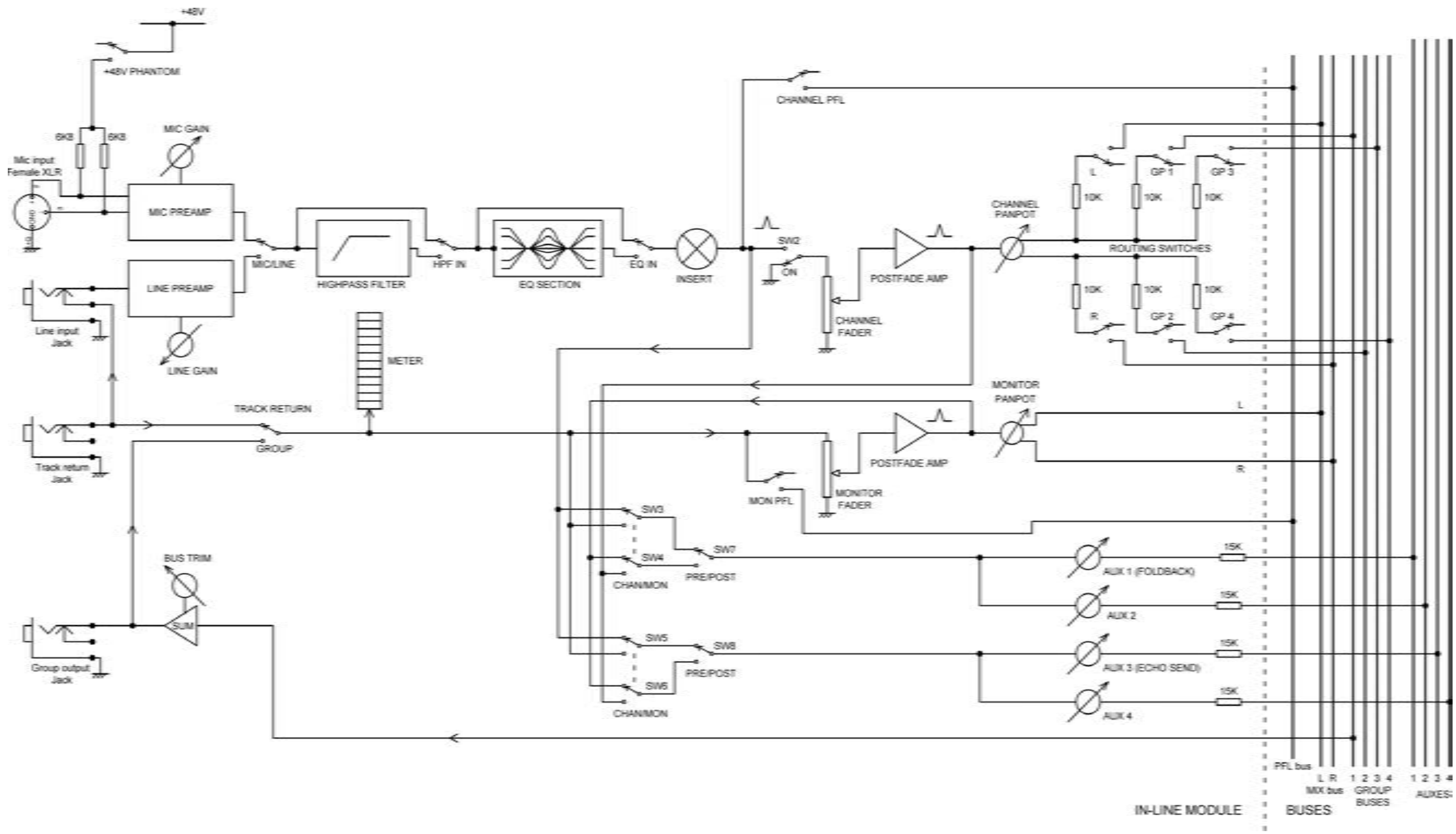

Figure 16.7: A channel/group/monitor module for an in-line recording console

the stereo mix bus to create a monitor mix. Note that every in-line module will have a meter, though it is not usual to provide meters for every channel on a split console. A monitor PFL switch is provided so that an individual track return or group can be conveniently listened to.

As in split consoles, it is convenient to connect the track return to the channel line input through normalling contacts on the line input jack, so switching to mixdown requires the minimum number of operations. As described earlier, in this form of in-line configuration the monitor path is redundant, as the channel fader and panpot are used to control the signal coming back from the multi-track recorder, and so there is usually a way in which the monitor path can be used as an extra effects return, as is commonly done in split consoles. Typically, the line input jack is used as the input connector for this purpose; alternatively an extra insert point may be provided in the monitor path, into which an external line signal can be fed. These arrangements are not shown in Figure 16.7, which is quite complicated enough already; it is one of the drawbacks of the in-line system that it is conceptually more complex than the split console format.

The auxiliary send system is slightly more complex in an in-line module, in the interests of maximum flexibility of working. In the example shown, the sends can be switched in pairs to take their signal either prefade or postfade from either the channel or monitor path. This is implemented by means of the switches SW3–SW8. For example, during recording the effect sends can be switched to be monitor postfade, allowing wet monitoring. Many variations on this aux send are possible; on larger consoles there will be six or eight sends, and sometimes a stereo aux send with its own panpot is provided.

CHAPTER 17

Microphone preamplifiers

Microphone preamplifier requirements

A microphone preamplifier is a serious design challenge. It must provide a gain variable from 0 to 70 dB or more to amplify deafening drum-kits or discreet dulcimers, generate minimal internal noise, and have a high-CMRR balanced input to cancel noise pickup in long cables. It must also be able to withstand +48 V DC of phantom-power suddenly applied to its input while handling microvolt signals.

It is now rare to use input transformers to match a low-impedance (150–200 Ω) microphone to the preamplifier, since the cost and weight penalty is serious, especially when linearity at low frequencies and high levels is important. Both dynamic and capacitor microphones have a low output impedance of this order; the former because of the low number of turns used in the coil, the latter because active buffering of the extremely high capsule impedance is essential.

The low-noise requirement rules out the direct use of opamps, since their design involves compromises that make them at least 10 dB noisier than discrete transistors when faced with low source impedances. The answer, for at least the last 30 years, has been to use hybrid input stages that combine discrete input transistors to give low noise, combined with opamps to provide raw open-loop gain for linearisation, and load-driving capability.

To summarise the requirements of a microphone preamplifier:

1. Variable gain, usually from 0 to +70 dB. Some designs have a gain range extending to +80 dB. The bottom 20 dB section of the range is often accessed by switching in a 20 dB input attenuator.
2. Minimal noise. Taking the source impedance of a microphone as 200 Ω, the Johnson noise from that resistance is −129.6 dBu at 25 °C, 20 kHz bandwidth. This puts an immediate limit on the noise performance; if you set up 70 dB of gain then the noise output from the preamplifier will be −59.6 dBu, even if it is itself completely noiseless. Most mic preamplifiers approach this at maximum gain, often having a noise figure as low as 1 or 2 dB, but depart from it further and further as gain is reduced. In other words,

the noise output does not fall as fast as it would if the preamp was noiseless, because reducing the gain means increasing the effective resistance of the gain control network, creating more Johnson noise.

3. The input must have a high common-mode rejection ratio (CMRR) to reject interference and ground noise. The CMRR should ideally be high, be flat with frequency, and remain high as the input gain is altered over the whole range. In practice, CMRR tends to worsen as gain is reduced. Because of the need for a balanced input, microphone inputs are almost always female XLR connectors.
4. The input must have a constant resistive input impedance of 1 to 2 kΩ, which provides appropriate loading for a 200 Ω dynamic microphone capsule. This is also a suitable load for the internal head amplifiers of capacitor microphones.
5. The input must be proof against the sudden application (or removal) of +48 V DC phantom-power. It must withstand this for many repeated cycles over the life of the equipment.

Transformer microphone inputs

For a long time transformer microphone inputs were the only real option. The cost and weight of a microphone input transformer for every channel is considerable, especially at the quality end of the market, because a large transformer core is needed to handle high levels at low frequency without distortion. Because of the low signal levels, mumetal screening cans were normally used to minimise magnetic interference, and this added to the cost and weight.

Step-up ratios from 1:5 to 1:10 were used, the higher ratios giving a better noise performance but more difficulties with frequency response because of the greater capacitance of a larger secondary winding. The impedance reflected to the input depends on the square of the step-up ratio, so with a 1:10 transformer the secondary had to be loaded with 100 kΩ to get the desired 1 kΩ at the primary input.

The arrangement in Figure 17.1 has an amplifier gain range from to 0.3 to 60 dB to which is added 20 dB from the transformer ratio of 1:10, giving +20 to +80 dB of overall gain. R6 in conjunction with the reverse-log pot shape the control law so it is roughly logarithmic. R4 is the 100 kΩ secondary loading resistor, and C1, R3 form a Zobel network to damp the secondary resonance. In the mid-Seventies, when opamps were still a dubious proposition for quality audio, the input amplifiers were very often discrete transistor stages using four or more transistors. Note R5, which prevents the gain from being reduced to exactly unity. Its presence is testimony to the fact that the discrete preamplifiers were difficult to stabilise at HF if the output and non-inverting input were directly connected together. R1 and R2 feed in phantom-power when the switch is in the 'on' position.

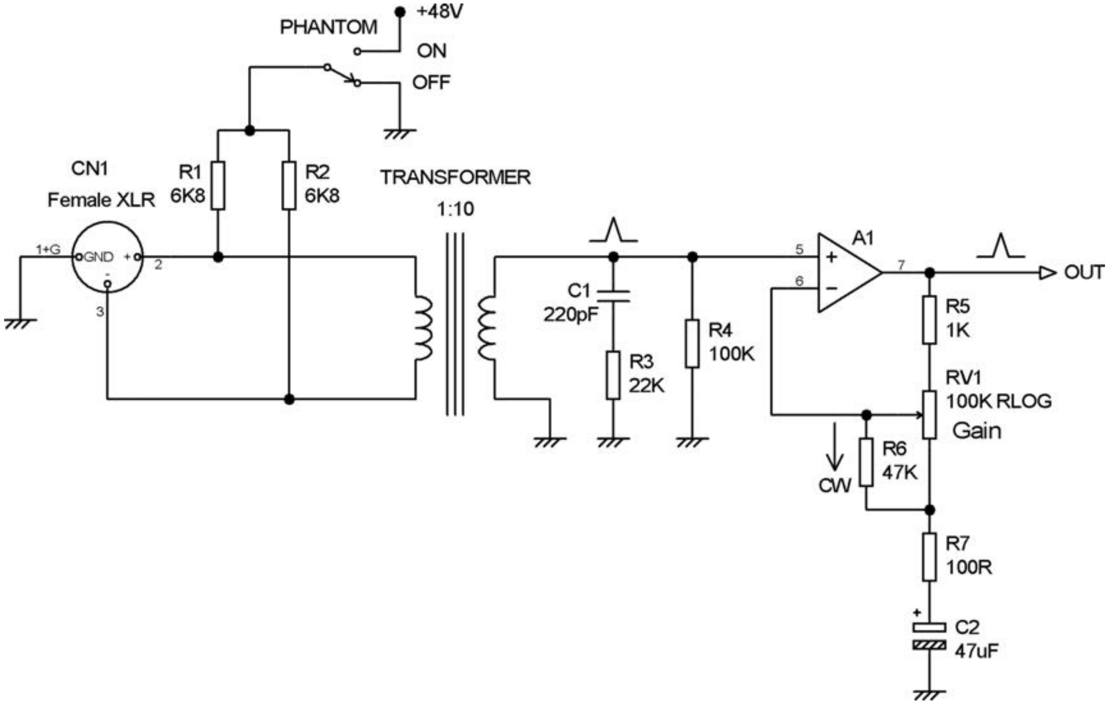

Figure 17.1: A transformer microphone preamplifier. Gain range +20 to +80 dB

Microphone transformer technology had, and still has, its own advantages. A balanced input with a good and constant CMRR was inherent. The step-up ratio, which gave 'gain for free' in electronic if not financial terms, and the associated impedance matching, meant that it was easy to design a quiet input stage. Transformers also gave good RF rejection, and isolated the input preamplifier from phantom-power voltages. For EMC reasons in particular, transformer microphones continued in use in broadcast mixing consoles long after they had disappeared from recording and PA equipment.

The simple hybrid microphone preamplifier

The cost incentive to develop an effective low-noise transformerless microphone input was considerable, and after much experimentation the arrangement shown in Figure 17.2 became pretty much standard for some years, being extensively used in mixers around the period 1978–1984 (in this period the more expensive consoles stuck to transformer microphone amplifiers). The difficulties of getting the noise low enough and the linearity good enough were at first formidable, but after a good deal of work (some of which I did) such stages became almost universal. Q1 and Q2 work as common-emitter stages, with the gain-control

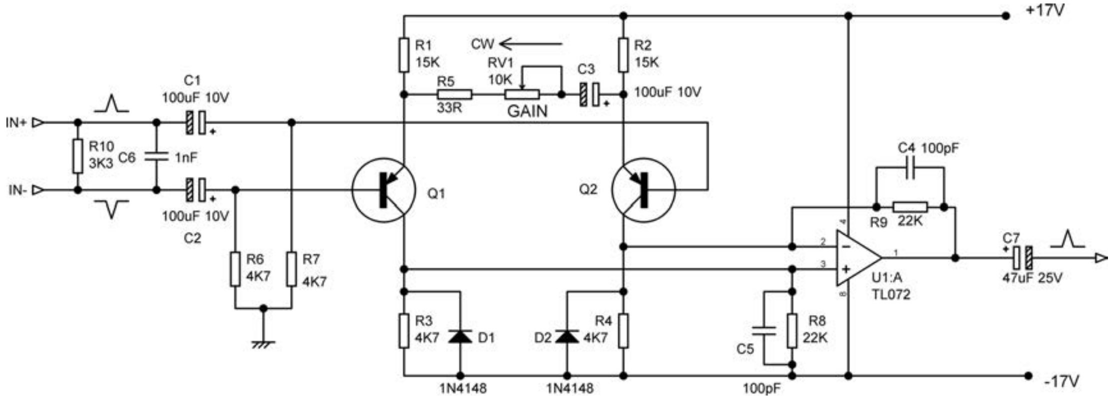

Figure 17.2: The simple hybrid microphone preamplifier

network R5, RV1, C3 connected between the emitters. As the resistance of this network is reduced, the differential gain increases but the common-mode gain remains low. The two signals at the collectors are then summed by the opamp. This does not have to be a low-noise type because the gain in the transistors means they determine the noise performance, and the TL072 was almost universally used in this position, being at the time much cheaper than the 5532. With appropriate choice of transistors (a type with low Rb) and collector current, preamplifiers of this type can give an equivalent input noise (EIN) of −128 dBu at maximum gain, equivalent to a noise figure of only 1.2 dB with a 200 Ω load. The EIN rises as gain is reduced, because the resistance of the gain-control network and its Johnson noise is increased. The noise output still falls as gain is reduced, but not as fast as the gain.

The gain law is very non-linear with Rg; a reverse-log D-law pot helps but there is still some cramping of the calibration at the high gain end, and preamplifier gain ranges of more than 50 dB are not really practicable with this simple arrangement.

Today this approach is considered obsolete for anything except budget mixer purposes because of its mediocre distortion performance. You will note that the input transistor pair has no overall feedback loop closed around them, and their nonlinearity creates significant distortion, especially at high gains.

The two reverse-biased diodes in the transistor collector circuits are to prevent the opamp being damaged by having its input driven below the −17 V rail when phantom-power was applied or removed. Note that the input coupling capacitors C1, C2 shown here are not intended to cope with the phantom voltage if used. On arriving at the input XLR, the microphone signal first encounters the phantom feed resistors, DC blocking capacitors (usually rated at 63 V), a switchable 20 dB attenuator, and possibly a phase-invert switch that swaps over the inputs, before it reaches the preamplifier (this is illustrated in Figure 17.4).

The balanced-feedback hybrid microphone preamplifier (BFMA)

The microphone preamp architecture described in the previous section has the merit of simplicity, but because there is no global feedback loop around the input transistors its distortion performance falls far short of the rest of a mixer, which typically consists of pure opamp circuitry with very low distortion. It is obviously undesirable practically, aesthetically, and in every other way for the very first stage in the signal chain to irretrievably mess up the signal, but this is what happened for several years. In the mid-1980s I decided to do something about this. Many methods of applying feedback to the input stage were tried, but foundered on the fact that if feedback was applied to one of the transistors, the current variations in the other were excessive and created distortion. The use of two feedback paths in anti-phase, i.e. balanced feedback, was my solution to this problem; this meant that one feedback path would have to be via an inverting amplifier, with extra phase shifts that might imperil HF stability.

The basic concept is shown in Figure 17.3. Direct negative feedback to Q2 goes through R12, while the inverted feedback passes through the unity-gain inverting stage U1:B and goes through R11 to Q1. The gain-control network R5, RV1 and C3 is connected between the two feedback points at the Q1, Q2 emitters, and sets the closed-loop gain by controlling the NFB factor. Distortion is pretty much eliminated, even at maximum gain.

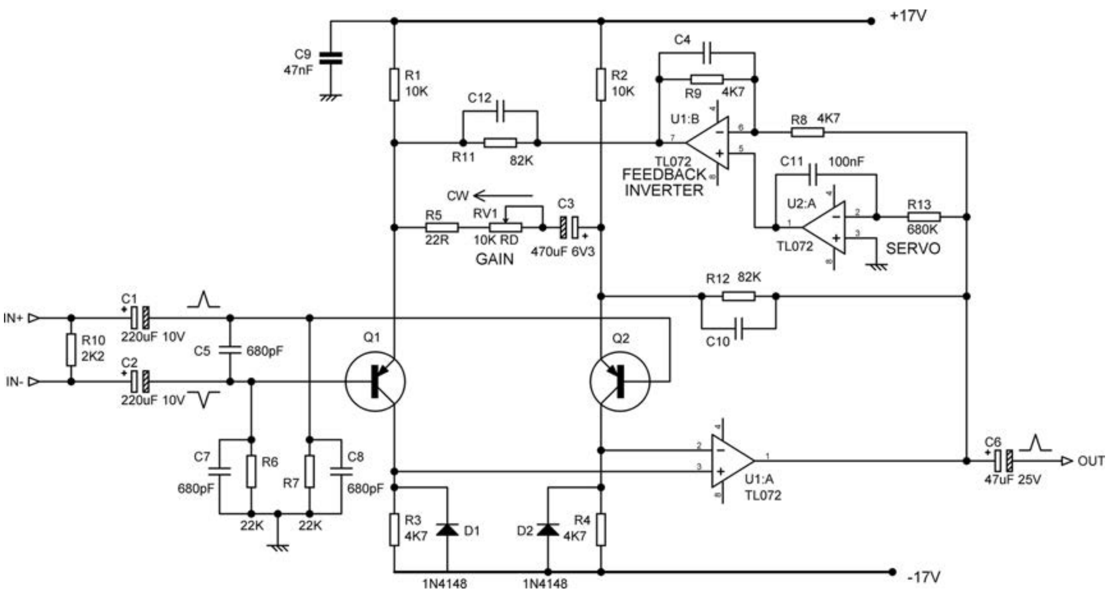

Figure 17.3: The balanced feedback microphone amplifier, known to its friends as the BFMA

The solution to the stability problem is to make sure that the direct HF feedback through C10 dominates that through C12, which has been through the inverter U1:A; this is aided by adding C4 to the inverter to control its HF response. This feedback is therefore not symmetrical at extreme HF, but this has no effect on the functioning of the circuit with audio signals.

$$\text{Gain} = \frac{R_{nfb} + \left(\left(\frac{R_g}{2} R_e\right) \Big/ \left(\frac{R_g}{2} + R_e\right)\right)}{2\left(\frac{R_g}{2} R_e\right) \Big/ \left(\frac{R_g}{2} + R_e\right)} \qquad \text{(Equation 17.1)}$$

The closed-loop gain is given by Equation 17.1, which looks complicated, but becomes clearer when you appreciate that the terms in brackets simply represent the parallel combination of the emitter resistor Re (R1, R2 in the figure) and half the resistance of the gain control network Rg. Rnfb is the value of R11, R12. The 2 on the bottom comes from the fact that we are only using one output of the amplifier – there is an inverted output from the inverter which could be phase-summed to give twice the gain. We don't do that because it is desirable to have as low a minimum gain as possible. With the values shown, the gain range is +22 to +71 dB, which in conjunction with a switchable 20 dB pad gives a useful total range of about 0 to 70 dB.

In order to get enough maximum gain with reasonable values for R5 and C3, the feedback resistors R11, R12 have to be quite high at 82 kΩ. This means that the control of the DC level at the output is not very good. To solve this, the DC servo integrator U2:A was added; R13 is connected to the preamplifier output and the integrator acts via the non-inverting input of the inverter opamp to keep the output at 0 V.

This technology was used on several consoles, such as the Soundcraft TS12, and later appeared in the Yamaha GA 24/12 and GA32/12 mixers in 1998 (a very sincere form of flattery). It had a relatively short life at Soundcraft as I came up with something better – the padless microphone preamplifier described later in this chapter.

Microphone and line input pads

Microphone pads, or attenuators, are used when the output is too high for the mixing console input to cope with; this typically happens when you put a microphone inside a kick-drum. Attenuators are also used when, for reasons of economy, it is desirable that the microphone input doubles up as a line input.

A typical arrangement is shown in Figure 17.4. The preamplifier has a typical gain range of +20 to +70 dB. There is an input XLR and phantom feed resistors R1, R2. C1 and C2 are DC blocking capacitors to stop the phantom voltage from getting into the preamplifier; these should be as large as possible to preserve LF CMRR. Next comes a 20 dB balanced attenuator made up of R3–R6; note that the loading of the preamplifier input resistor

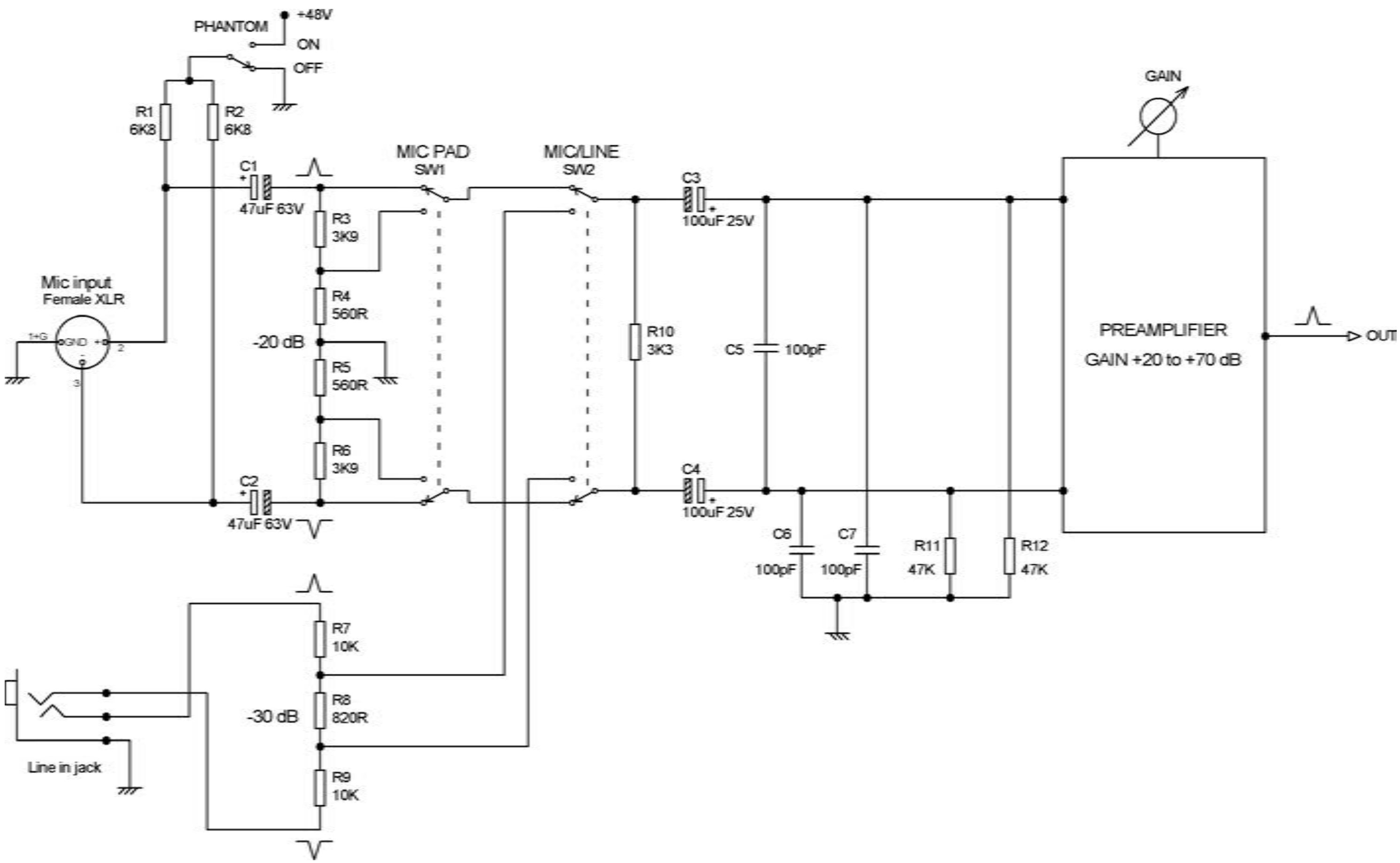

Figure 17.4: Mic and line attenuators at the input of a preamplifier

R10 must be taken into account when designing the attenuator resistor values. One of the functions of this resistor is to prevent the preamplifier input from being open-circuited when the pad switch SW1 or the mic/line switch SW2 is between contacts. C3 and C4 are two further DC blocking capacitors, which prevent the input terminals of the preamplifier, which are not in general at 0 V, from causing clicks in the switching. C5, C6 and C7 increase EMC immunity and also keep the preamplifier from oscillating if the microphone input is left open-circuit at maximum gain. Such oscillation is not an indication that the preamplifier itself is unstable – it normally happens because the insert jacks, which carry the output signal from the preamplifier, are capacitively crosstalking to the microphone input, forming a feedback loop. Ideally the microphone input should be open-circuit stable not only with the input gain at maximum, but also with full treble boost set up on the EQ section. This can be challenging to achieve, but it is possible with careful attention to layout. I have done it many times.

In line mode, the microphone gain is, of course, much too high, and the usual practice is to use a 30 dB attenuator on the line input, which allows a high input impedance to be set by R7, R9 while R8 provides a low source impedance to minimise preamplifier noise. This pad alters the gain range to −10 to +40 dB, which is actually too wide for a line input, and some consoles have another switch section in the mic/line switch, which reduces the gain range of the preamplifier so that the overall range is a more useful −10 to +20 dB. The larger and more expensive consoles usually have separate line input stages which avoid the compromises inherent in using the microphone input as a line input.

There is an important point to be made about the two attenuators. You will have noticed that the microphone attenuator uses four resistors and has its centre connected to ground, whereas the line attenuator uses a more economical three resistors with no ground connection. The disadvantage of the three resistor version is that the wanted differential signal is attenuated, but the unwanted common-mode signal is not, and so the CMRR is much worsened. This does not happen with the four-resistor configuration because the ground connection means that both differential and common-mode signals are attenuated equally. There is no reason why the line attenuator here could not have been designed with four resistors – I just wanted to make the point.

The microphone amplifiers described have a high CMRR, and a problem with attenuators like this is that both types degrade the overall CMRR quite seriously because of their resistor tolerances, even if 1% components are used.

The padless microphone preamplifier

The ideal microphone preamplifier would have a gain range of 70 dB or thereabouts on a single control, going down to unity gain without the inconvenience of a pad switch. It was mentioned in the previous section that resistive pads degrade the overall CMRR, and also the

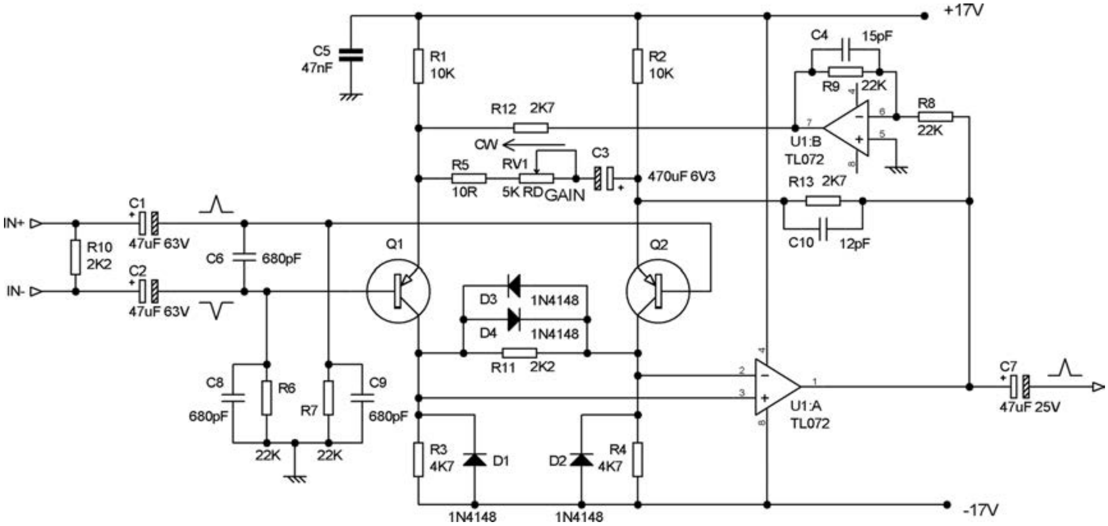

Figure 17.5: The padless balanced feedback microphone preamp: mic input stage

noise performance as an inevitable consequence of following a 20 dB pad with an amplifier having 20 dB of gain. In addition, space on a channel front panel is always in short supply and losing a switch would be very welcome. I therefore invented the padless microphone preamplifier. Looking at the mixer market today (2013), the idea seems to have caught on.

The concept is based on the balanced feedback mic amp described above, but now the total gain is spread over two stages to give a smooth 0–70 dB gain range with the rotation of a single knob.

The first stage shown in Figure 17.5 is based on the BFMA circuit in Figure 17.3 above, but with the feedback resistors reduced to 2k7 to reduce the gain range. The gain-control network R5, RV1 and C3 has also been halved in resistance to reduce Johnson noise, and the net result is a gain range of +1.5 to +49 dB; as before, a reverse-log D-law pot is used. The lower feedback resistors mean that no servo is required to correct the DC conditions. Note that the greater amount of NFB means that under overload conditions it is possible for the common-mode range of the opamp to be exceeded, leading to the well-known TL072 phase reversal and latchup. This is prevented by R11, D3, and D4, which have no effect on linearity in normal operation.

The second stage of the padless mic amp is shown in Figure 17.6. This consists essentially of a variable-gain balanced input stage as described in Chapter 18, configured for a gain range of 0 to +20 dB. The gain pot is once again a reverse-log D-law pot and the combination of the gain laws of the two stages gives a very reasonable law over the almost 70 dB range, though there is still a little cramping at the high-gain end.

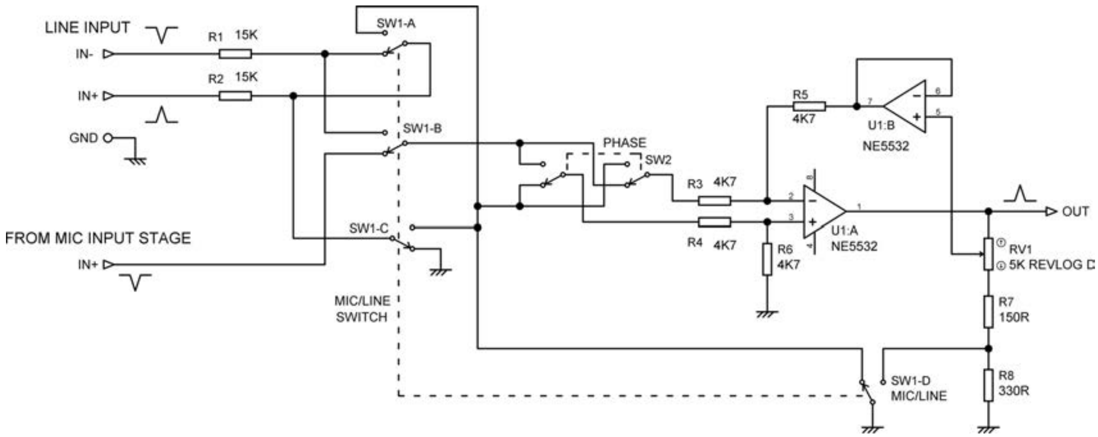

Figure 17.6: The padless balanced feedback microphone preamp: mic/line switching and second stage

The second stage is also used as a line-input stage with a gain range of −10 to +20 dB. The mic/line switching used to do this may look rather complex, but it does a bit more than simply change sources. In Figure 17.6, switch SW1 is shown in the 'mic' position, and the first stage reaches the inverting input of the second via SW1-B; the output of the mic amp is phase-inverted simply by swapping over its inputs. Line input resistors R1, R2 give reduced gain for line input working, and in mic mode they are shorted together by SW1-A to prevent crosstalk from line to mic, which is an important issue when track-normaling is incorporated (see Chapter 16). In several of my designs these resistors were placed on the input connector PCB rather than in the channel, to keep their hot ends away from other circuitry and further reduce line to mic crosstalk. SW1-C shorts the junction of R1, R2 to ground, further improving mic/line crosstalk if the line input is not balanced. SW1-D shorts the unused second stage non-inverting input to ground.

In line mode, R1, R2 are connected to the second stage via SW1-B and SW1-C. In mic mode, SW1-D shorts R8 to ground so that the gain range of the second stage is increased to the required 30 dB. SW2 is a phase-invert switch which simply swaps over the input connections to the second stage.

The padless mic amp gives both a good gain control law and lower noise at low gain settings. The noise performance versus gain for a typical example can be seen in Table 17.1; as described earlier, the EIN and noise figure worsen as the gain is turned down, due to the increased resistance of the gain network. A noise figure of 30 dB may appear to be pretty dire, but the corresponding noise output is only −98 dBu, and this will soon be submerged in the noise from following stages.

TABLE 17.1 Noise performance versus gain of padless mic amp

Gain (dB)	Noise out (dBu)	EIN (dBu)	Noise figure (dB)
1.4	−98.2	−99.6	30.0
6.6	−95.4	−102.0	27.6
15.4	−90.5	−105.9	23.7
31.7	−83.1	−114.8	14.8
44.1	−77.5	−121.6	8.0
51.8	−72.7	−124.5	5.1
60.6	−65.7	−126.3	3.3
69.3	−58.7	−128.0	1.6

This effect can be reduced by reducing the impedance of the feedback and gain control network. This increases the power required to drive them, and because of the square-root in the Johnson noise equation, a reduction by a factor of ten, (which would need some serious electronics) would only give a 10 dB improvement, and that at the low gain end where it is least needed. Specialised outboard mic amps with low-resistance feedback networks, driven by what are in effect small power amplifiers, have been developed but do not seem to have caught on.

Another advantage of the padless approach is that one pair of DC blocking capacitors suffices, rated at 63 V as in Figure 17.5, and this improves the CMRR at low frequencies. The padless microphone preamplifier concept was protected by patent number GB 2242089 in 1991, and was used extensively over many ranges of mixing console.

Capacitor microphone head amplifiers

A capacitor capsule has an extremely high output impedance, equivalent to a very small capacitor of a few picoFarads. It is in fact the highest impedance you are ever likely to encounter in the audio business, and certainly the highest I have ever had to deal with. Special circuit techniques are required to combine low noise and high impedance, working with a strictly limited amount of power. A while ago I designed the electronics for a new capacitor microphone by one of the well-known manufacturers, and the circuitry described here is a somewhat simplified version of that.

The first point is that the microphone capsule had an impedance of about 5 pF, so to get a −3 dB point of 10 Hz the total load impedance has to be no more than 3.2 GΩ. (yes, that's 3200 MΩ). The capsule needs to be fed with a polarising voltage through a resistor, and the

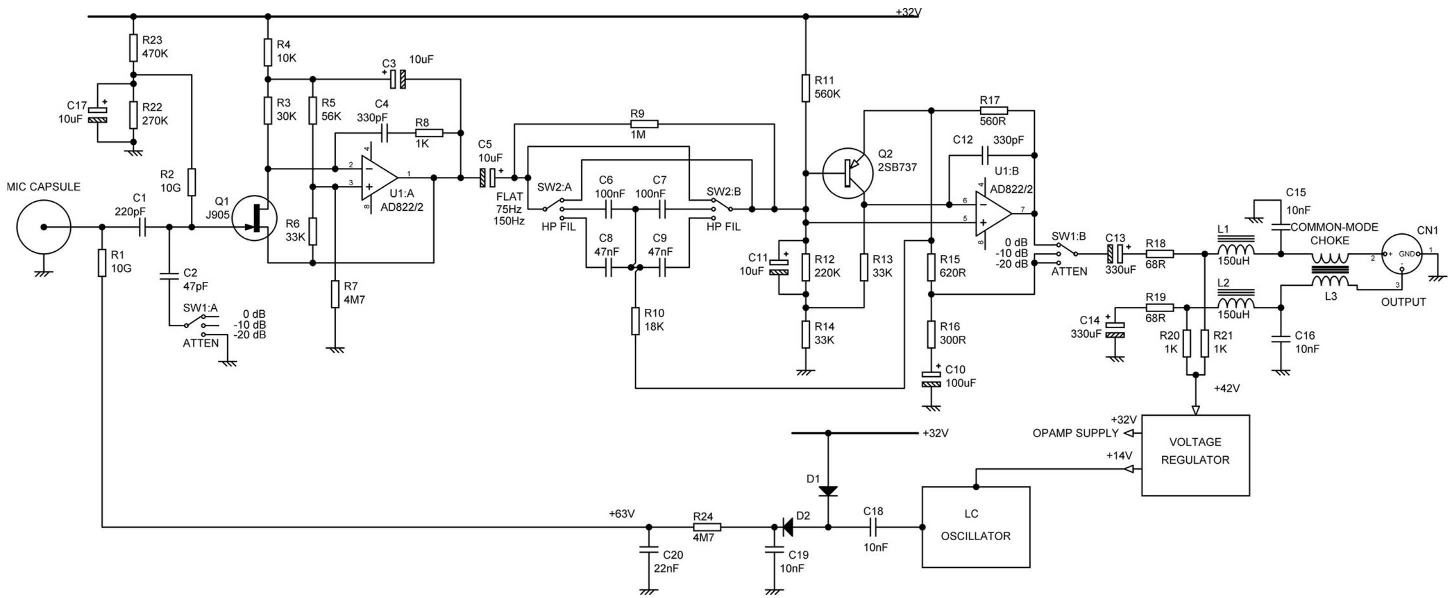

Figure 17.7: A typical head amp system for a capacitor microphone with phantom powering. All of this circuitry is fitted onto the body of the microphone

head amplifier needs a biasing resistor. In this design, both were 10 GΩ, which meant that the input impedance of the amplifier itself had to be not less than 8.9 GΩ. Resistors with these astronomical values are exotic components that come in a glass encapsulation that must be manipulated with tweezers – one touch of a finger and the insulation properties of the glass are fatally compromised.

Figure 17.7 shows my capacitor mic head amp. R1 supplies the capsule polarising voltage and R2 biases the first stage, a unity-gain JFET source-follower augmented by opamp U1:A which provides the gain for a high NFB factor to linearise Q1. The drain of Q1 is bootstrapped via C3 to prevent local feedback through the gate-drain capacitance of the JFET from reducing the input impedance. R3, R5, R6 set the DC conditions for Q1.

The second stage is a low-noise amplifier with gain of +4 dB, defined by the ratio of feedback resistor R17 to R15 and R16. Like the first stage, it is a hybrid design that combines the low noise of low-Rb transistor Q2 with the open-loop gain and load-driving capability of an opamp.

The stage also acts as a unity-gain follower making up a second order Butterworth Sallen and Key high-pass filter when C6, C7 or C8, C9 are switched in; the resistive elements are R10 and R14.

The system has two steps of attenuation; −10 and −20 dB. The first 10 dB step is obtained by using SW1:B to take the output from the junction of R15 and R16 instead of the normal stage output. The second 10 dB step results from switching C2 across the mic capsule, forming a capacitive attenuator that reduces the input to the first stage and prevents overload. The gains available are thus +4 dB, −6 dB and −16 dB. The maximum sound pressure handling is +146 dB SPL, or +155 dB SPL with the −20 dB pad engaged.

The incoming phantom-power is tapped off by R20, R21 and fed to a discrete BJT regulator that gives +32 V to power the opamp, and +14 V to run a small LC oscillator that pumps the +32 V rail up to the +63 V required to polarise the capsule. Total current consumption is 2.2 mA.

The noise output is −120.7 dBu (A-weighted), which may appear high compared with the microphone amplifiers described above, but remember that a capacitor microphone puts out a high signal voltage so that the S/N ratio is actually very good. The mic capsule, being a pure reactance, generates no noise of its own, and the noise output comes only from the Brownian motion of the air against the capsule diaphragm and from the electronics. A larger diameter capsule means lower noise because more energy is absorbed from the coherent sound waves; the same applies to the Brownian motion but this partially cancels, just as with the use of multiple opamps for low noise.

Noise measurement of this technology requires special methods. The impedances are so high that meaningful results can only be obtained by putting the circuitry inside a completely closed metal screening enclosure.

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CHAPTER 18

Line inputs

External signal levels

There are several standards for line signal levels. The −10 dBu standard is used for a lot of semi-professional recording equipment as it gives more headroom with unbalanced connections – the professional levels of +4 dBu and +6 dBu assume balanced outputs which inherently give twice the output level for the same supply rails as it is measured between two pins with signals of opposite phase on them (see Table 18.1).

Signal levels in dBu are expressed with reference to 0 dBu = 775 mVrms; the origin of this odd value is that it gives a power of 1 mW in a purely historical 600 Ω load. The unit of dBv refers to the same level but takes the voltage as the reference rather than the power – a distinction of little interest nowadays. Signal in dBv (or dBV) are expressed with reference to 0 dB = 1.000 Vrms.

These standards are well established, but that does not mean all equipment follows them. To take a current example, the Yamaha P7000S power amplifier requires +8 dBu (1.95 Vrms) to give its full output of 750 W into 8 Ω.

Internal signal levels

In any audio system it is necessary to select a suitable nominal level for the signal passing through it. This level is always a compromise – the signal level should be high so it suffers minimal degradation by the addition of circuit noise as it passes through the system, but not so high that it is likely to suffer clipping before the gain control, or generate undue distortion

TABLE 18.1 Nominal signal levels

	Vrms	dBu	dBv
Semi-professional	0.316	−7.78	−10
Professional	1.228	+4.0	+1.78
German ARD	1.55	+6.0	+3.78

below the clipping level (this last constraint is not normally a problem with modern circuitry, which gives very low distortion right up to the clipping point).

It must always be considered that the gain control may be maladjusted by setting it too low and turning up the input level from the source equipment, making input clipping more likely. The internal levels chosen are usually in the range −6 to 0 dBu, (388 mV to 775 Vrms) but in some specialised equipment such as broadcast mixing consoles, where levels are unpredictable and clipping distortion less acceptable than a bit more noise, the nominal internal level may be as low as −16 dBu (123 mVrms). If the internal level is in the normal −6 to 0 dBu range, and the maximum output of an opamp is taken as 9 Vrms, (+21.3 dBu) this gives from 27 to 20 dB of headroom before clipping occurs.

If the incoming signal does have to be amplified, this should be done as early as possible in the signal path, to get the signal well above the noise floor. If the gain is implemented in the first stage (i.e. the input amplifier, balanced or otherwise) the signal will be able to pass through later stages, at a high level and so their noise contribution will be less significant. On the other hand, if the input stage is configured with a fixed gain, it will not be possible to turn it down to avoid clipping. Ideally the input stage should have variable gain. It is not straightforward to combine this feature with a balanced input, but several ways of doing it are shown later in this chapter.

Input amplifier functions

Firstly, RF filtering is applied at the very front end to prevent noise breakthrough and other EMC problems. It must be done before the incoming signal encounters any semiconductors where RF demodulation could occur, and can be regarded as a 'roofing filter'. At the same time, the bandwidth at the low end is given an early limit by the use of DC-blocking capacitors, and in some cases overvoltage spikes are clamped by diodes. The input amplifier should present a reasonably high impedance to the outside world, not less than 10 kΩ, and preferably more. It must have a suitable gain – possibly switched or variable – to scale the incoming signal to the nominal internal level. Balanced input amplifiers also accurately perform the subtraction process that converts differential signals to single-ended ones, so noise produced by ground loops and the like is rejected. It's quite a lot of work for one stage.

Unbalanced inputs

The simplest unbalanced input feeds the incoming signal directly to the first stage of the audio chain. This is often impractical; for example, if the first stage was a Baxandall tone-control circuit then the boost and cut curves would be at the mercy of whatever source impedance was feeding the input. In addition, the input impedance would be low, and

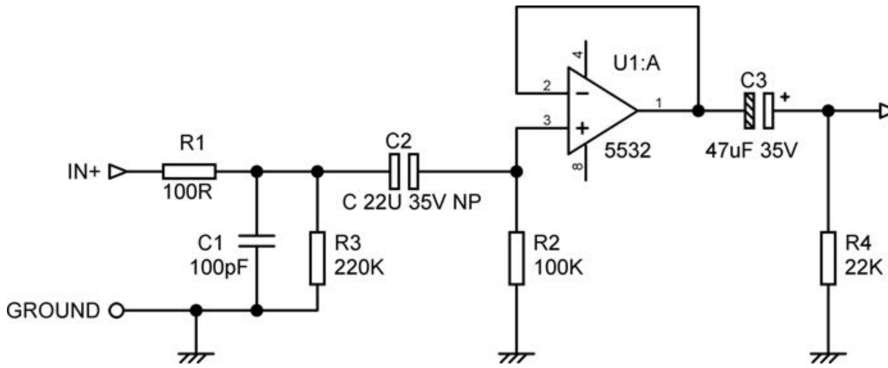

Figure 18.1: A typical unbalanced input amplifier with associated components

variable with frequency and control settings. Some sort of buffer amplifier which can be fed from a significant impedance without ill effect is needed.

Figure 18.1 shows an unbalanced input amplifier, with the added components needed for interfacing to the real world. The opamp U1:A acts as a unity-gain voltage follower; it can be easily altered to give gain by adding two series feedback resistors. A 5532 bipolar type is used here for low noise; with the low source impedances that are likely to be encountered here, an FET-input opamp would be 10 dB or more noisier. R1 and C1 are a first-order low-pass filter to remove incoming RF before it has a chance to reach the opamp and demodulate into the audio band; once this has occurred any further attempts at RF filtering are of course pointless. R1 and C1 must be as close to the input socket as physically possible to prevent RF from being radiated inside the box before it is shunted to ground, and so come before all other components in the signal path.

Selecting component values for input filters of this sort is always a compromise, because the output impedance of the source equipment is not known. If the source is an active preamplifier stage, then the output impedance will probably be around 50 Ω, but it could be as high as 500 Ω or more. If the source is an oxymoronic 'passive preamplifier' – i.e. just an input selector switch and a volume potentiometer, then the output impedance will be a good deal higher (at least one passive preamplifier uses a transformer with switched taps for volume control – see Chapter 13). If you really want to use a piece of equipment that embodies its internal contradictions in its very name, then a reasonable potentiometer value is 10 kΩ, and its maximum output impedance (when it is set for 6 dB of attenuation) will be 2.5 kΩ, which is very different from the 50 Ω we might expect from a good active preamplifier. This is in series with R1 and affects the turnover frequency of the RF filter. Effective RF filtering is very desirable, but it is also important to avoid a frequency response that sags significantly at 20 kHz. Valve equipment is also likely to have a high output impedance.

Taking 2.5 kΩ as a worst-case source impedance and adding R1, then 2.6 kΩ and 100 pF together give us −3 dB at 612 kHz ; this gives a 20 kHz loss of only 0.005 dB, so possibly C1 could be usefully increased; for example, if we made it 220 pF then the 20 kHz loss is still only 0.022 dB, but the −3 dB point is 278 kHz, much improving the rejection of what used to be called 'the medium wave'. If we stick with C1 at 100 pF and assume an active output with a 50 Ω impedance in the source equipment, then together with the 100 Ω resistance of R1 the total is 150 Ω, which in conjunction with 100 pF gives us −3 dB at 10.6 MHz. This is rather higher than desirable, but it is not easy to see what to do about it, and we must accept the compromise. If there was a consensus that the output impedance of a respectable piece of audio equipment should not exceed 100 Ω, then things would be much easier.

Our compromise seems reasonable, but can we rely on 2.5 kΩ as a worst case source impedance? I did a quick survey of the potentiometer values that passive preamplifiers currently employ, and while it confirmed that 10 kΩ seems to be the most popular value, one model had a 20 kΩ potentiometer, and another had a 100 kΩ pot. The latter would have a maximum output impedance of 25 kΩ, and would give very different results with a C1 value of 100 pF – the worst-case frequency response would now be −3 dB at 63.4 kHz and −0.41 dB at 20 kHz, which is not helpful if you are aiming for a ruler-flat response in the audio band.

To put this into perspective, filter capacitor C1 will almost certainly be smaller than the capacitance of the interconnecting cable. Audio interconnect capacitance is usually in the range 50 to 150 pF/metre, so with our assumed 2.5 kΩ source impedance and 150 pF/metre cable, and ignoring C1, you can only permit yourself a rather short run of 3.3 metres before you are −0.1 dB down at 20 kHz, while with a 25 kΩ source impedance you can hardly afford to have any cable at all; if you use low capacitance 50 pF/metre cable you might just get away with a metre. This is just one of many reasons why 'passive preamplifiers' really are *not* a good idea.

Another important consideration is that the series resistance R1 must be kept as low as practicable to minimise Johnson noise; but lowering this resistance means increasing the value of shunt capacitor C1, and if it becomes too big then its impedance at high audio frequencies will become too low. Not only will there be too low a roll-off frequency if the source has a high output impedance, but there might be an increase in distortion at high audio frequencies because of excessive loading on the source output stage.

Replacing R1 with a small inductor to make an LC low-pass filter will give much better RF rejection at increased cost. This is justifiable in professional audio equipment, but it is much less common in hifi, one reason being that the unpredictable source impedance makes the filter design difficult, as we have just seen. In the professional world one *can* assume that the source impedance will be low. Adding more capacitors and inductors allows a 3- or 4-pole LC filter to be made. If you do use inductors then it is essential to check the frequency response to make sure it is what you expect and there is no peaking at the turnover frequency.

C2 is a DC-blocking capacitor to prevent voltages from ill-conceived source equipment getting into the circuitry. It is a non-polarised type as voltages from the outside world are of unpredictable polarity, and it is rated at not less than 35 V so that even if it gets connected to defective equipment with an opamp output jammed hard against one of the supply rails, no harm will result. R3 is a DC drain resistor that prevents the charge put on C2 by the aforesaid external equipment from remaining there for a long time and causing a thud when connections are replugged; as with all input drain resistors, its value is a compromise between discharging the capacitor reasonably quickly, and keeping the input impedance acceptably high. The input impedance here is R3 in parallel with R2, i.e. 220 kΩ in parallel with 100 kΩ, giving 68 kΩ. This is a good high value and should work well with just about any source equipment you can find, including valve technology.

R2 provides the biasing for the opamp input; it must be a high value to keep the input impedance up, but bipolar input opamps draw significant input bias current. The Fairchild 5532 data sheet quotes 200 nA typical, and 800 nA maximum, and these currents would give a voltage drop across R2 of 20 mV and 80 mV respectively. This offset voltage will be reproduced at the output of the opamp, with the input offset voltage added on; this is only 4 mV maximum and so will not affect the final voltage much, whatever its polarity. The 5532 has NPN input transistors, and the bias current flows into the input pins, so the voltage at Pin 3 and hence the output will be negative with respect to ground by anything up to 84 mV.

Such offset voltages are not so great that the output voltage swing of the opamp is significantly affected, but they are enough to generate unpleasant clicks and pops if the input stage is followed by any sort of switching, and enough to make potentiometers crackly. Output DC-blocking is therefore required in the shape of C3, while R4 is another DC drain resistor to keep the output at zero volts. It can be made rather lower in value than the input drain resistor R3 as the only requirement is that it should not significantly load the opamp output. FET-input opamps have much lower input bias currents, so that the offsets they generate as they flow through biasing resistors are usually negligible, but they still have input offsets of a few milliVolts, so DC-blocking will still be needed if switches downstream are to work silently.

This input stage, with its input terminated by 50 Ω to ground, has a noise output of only −119.0 dBu over the usual 22−22 kHz bandwidth. This is very quiet indeed, and is a reflection of the fact that R1, the only resistor in the signal path, has the low value of 100 Ω and so generates a very small amount of Johnson noise, only −132.6 dBu. This is swamped by the voltage noise of the opamp, which is basically all we see; its current noise has negligible effect because of the low circuit impedances.

Balanced interconnections

Balanced inputs are used to prevent noise and crosstalk from affecting the input signal, especially in applications where long interconnections are used. They are standard on professional audio equipment, and are slowly but steadily becoming more common in the world of hifi. Their importance is that they can render ground loops and other connection imperfections harmless. Since there is no point in making a wonderful piece of equipment and then feeding it with an impaired signal, making an effective balanced input is of the first importance.

The basic principle of balanced interconnection is to get the signal you want by subtraction, using a three-wire connection. In some cases a balanced input is driven by a balanced output, with two anti-phase output signals; one signal wire (the hot or in-phase) sensing the in-phase output of the sending unit, while the other senses the anti-phase output.

In other cases, when a balanced input is driven by an unbalanced output, as shown in Figure 18.2, one signal wire (the hot or in-phase) senses the single output of the sending unit, while the other (the cold or phase-inverted) senses the unit's output-socket ground, and once again the difference between them gives the wanted signal. In either of these two cases, any noise voltages that appear identically on both lines (i.e. common-mode signals) are in theory completely cancelled by the subtraction. In real life the subtraction falls short of perfection, as the gains via the hot and cold inputs will not be precisely the same, and the degree of discrimination actually achieved is called the common-mode rejection ratio, of which more later.

It is deeply tedious to keep referring to non-inverting and inverting inputs, and so these are usually abbreviated to ‘hot’ and ‘cold’ respectively. This does *not* necessarily mean that the hot terminal carries more signal voltage than the cold one. For a true balanced connection, the voltages will be equal. The ‘hot’ and ‘cold’ terminals are also often referred to as IN+ and IN−, and this latter convention has been followed in the diagrams here.

The subject of balanced interconnections is a large one, and a big book could be written on this topic alone; one of the classic papers on the subject is by Muncy [1]. To make a start, let us look at the pros and cons of balanced connections.

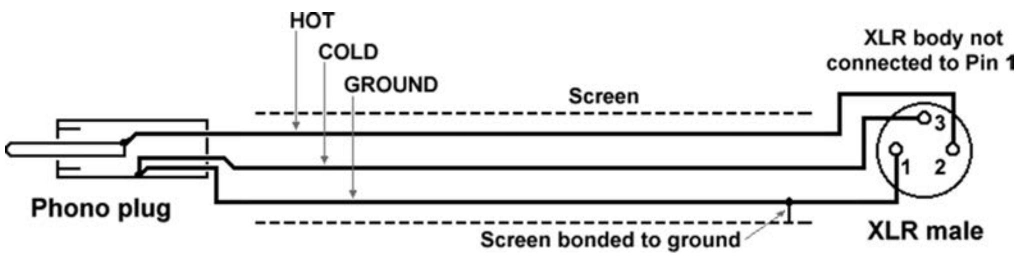

Figure 18.2: Unbalanced output to balanced input interconnection

The advantages of balanced interconnections

- Balanced interconnections discriminate against noise and crosstalk, whether they result from ground currents, or electrostatic or magnetic coupling to signal conductors.
- Balanced connections make ground-loops much less intrusive, and usually inaudible, so people are less tempted to start 'lifting the ground' to break the loop, with possibly fatal consequences. This tactic is only acceptable if the equipment has a dedicated ground-lift switch that leaves the external metalwork firmly connected to mains safety earth. In the absence of this switch the foolhardy and optimistic will break the mains earth (not quite so easy now that moulded mains plugs are standard) and this practice is of course highly dangerous, as a short-circuit from mains to the equipment chassis will result in live metalwork but dead people.
- A balanced interconnection incorporating a true balanced output gives 6 dB more signal level on the line, which should give 6 dB more dynamic range. However, this is true only with respect to *external* noise – as is described later in this chapter, a standard balanced input using 10 kΩ resistors is about 14 dB noisier than the unbalanced input shown in Figure 18.1 above.
- Balanced connections are usually made with XLR connectors. These are a professional 3-pin format, and are far superior to the phono (RCA) type normally used for unbalanced connections (there is more on this below).

The disadvantages of balanced interconnections

- Balanced inputs are inherently noisier than unbalanced inputs by a large margin, in terms of the noise generated by the input circuitry itself rather than external noise. This may appear paradoxical but it is all too true, and the reasons will be fully explained in this chapter.
- More hardware means more cost. Small-signal electronics is relatively cheap; unless you are using a sophisticated low-noise input stage, of which more later, most of the extra cost is likely to be in the balanced input connectors.
- Balanced connections do not of themselves provide any greater RF immunity than an unbalanced input. For this to happen, both legs of the balanced input would have to demodulate the RF in equal measure for common-mode cancellation to occur. This is highly unlikely, and the chances of it happening over a wide frequency range are zero. It remains vital to provide the usual passive RF filtering in front of any electronics to avoid EMC troubles.

- There is the possibility of introducing a phase error. It is all too easy to create an unwanted phase inversion by confusing hot and cold when wiring up a connector, and this can go undiscovered for some time. The same mistake on an unbalanced system interrupts the audio completely and leaves no room for doubt.

Balanced cables and interference

In a balanced interconnection two wires carry the signal, and the third connection is the ground wire which has two functions. Firstly, it joins the grounds of the interconnected equipment together. This is not always desirable, and if galvanic isolation is required a transformer balancing system will be necessary because the large common-mode voltages are likely to exceed the range of an electronic balanced input. A good transformer will also have a very high CMRR, which will be needed to get a clean signal in the face of large CM voltages.

Secondly, the presence of the ground allows electrostatic screening of the two signal wires, preventing both the emission and pick-up of unwanted signals. This can mean:

1. A lapped screen, with wires laid parallel to the central signal conductor. The screening coverage is not total, and can be badly degraded as the screen tends to open up on the outside of cable bends. Not recommended unless cost is the dominating factor.
2. A braided screen around the central signal wires. This is much more expensive, as it is harder to make, but opens up less on bending than a lap screen. Even so, screening is not 100%. It has to be said that it is a pain to terminate in the usual audio connectors. Not recommended.
3. An overlapping foil screen, with the ground wire (called the drain wire in this context for some reason) running down the inside of the foil and in electrical contact with it. This is usually the most effective as the foil is a solid sheet and cannot open up on the outside of bends. It should give perfect electrostatic screening, and it is much easier to work with than either lap screen or braided cable. However, the higher resistance of aluminium foil compared with copper braid means that RF immunity may not be so good.

There are three main ways in which an interconnection is susceptible to hum and noise:

1 Electrostatic coupling

An interfering signal at significant voltage couples directly to the inner signal line, through stray capacitance. The stray capacitance between imperfectly-screened conductors will be a fraction of a pF in most circumstances, as electrostatic coupling falls off with the square of distance. This form of coupling can be serious in studio installations with unrelated signals running down the same ducting.

The three main lines of defense against electrostatic coupling are effective screening, low impedance drive, and a good CMRR maintained up to the top of the audio spectrum. As regards screening, an overlapped foil screen provides complete protection.

Driving the line from a low impedance, of the order of 100 Ω or less, is also helpful because the interfering signal, having passed through a very small stray capacitance, is a very small current and cannot develop much voltage across such a low impedance. This is convenient because there are other reasons for using a low output impedance, such as optimising the interconnection CMRR, minimising HF losses due to cable capacitance, and driving multiple inputs without introducing gain errors. For the best immunity to crosstalk the output impedance must remain low up to as high a frequency as possible. This is definitely an issue as opamps invariably have a feedback factor that begins to fall from a low, and quite possibly sub-audio, frequency, and this makes the output impedance rise with frequency as the negative feedback factor falls, as if an inductor were in series. Some line outputs have physical series inductors to improve stability or EMC immunity, and these should not be so large that they significantly increase the output impedance at 20 kHz. From the point of view of electrostatic screening alone, the screen does not need to be grounded at both ends, or form part of a circuit [2]. It must of course be grounded at some point.

If the screening is imperfect, and the line impedance non-zero, some of the interfering signal will get into the hot and cold conductors, and now the CMRR must be relied upon to make the immunity acceptable. If it is possible, rearranging the cable-run away from the source of interference and getting some properly screened cable is more practical and more cost-effective than relying on very good common-mode rejection.

Stereo hifi balanced interconnections almost invariably use XLR connectors. Since an XLR can only handle one balanced channel, two separate cables are almost invariably used and interchannel capacitive crosstalk is not an issue. Professional systems, on the other hand, use multi-way connectors that do not have screening between the pins and there is an opportunity for capacitive crosstalk here, but the use of low source impedances should reduce it to below the noise floor.

2 Magnetic coupling

If a cable runs through an AC magnetic field, an EMF is induced in both signal conductors and the screen, and, according to some writers, the screen current must be allowed to flow freely or its magnetic field will not cancel out the field acting on the signal conductors, and therefore the screen should be grounded at both ends, to form a circuit [3]. In practice the magnetic field cancellation will be very imperfect and reliance is better placed on the CMRR of the balanced system to cancel out the hopefully equal voltages induced in the two signal wires. The need to ground both ends to possibly optimise the magnetic rejection is not usually a restriction, as it is rare that galvanic isolation is required between two pieces of audio equipment.

The equality of the induced voltages can be maximised by minimising the loop area between the hot and cold signal wires, for example by twisting them tightly together in manufacture. In practice most audio foil-screen cables have parallel rather than twisted signal conductors, but this seems adequate almost all of the time. Magnetic coupling falls off with the square of distance, so rearranging the cable-run away from the source of magnetic field is usually all that is required. It is unusual for it to present serious difficulties in a hifi application.

3 Ground voltages

These are the result of current flowing through the ground connection, and is often called 'common-impedance coupling' in the literature [1]. This is the root of most ground-loop problems. The existence of a loop in itself does no harm, but it is invariably immersed in a 50 Hz magnetic field that induces mains-frequency currents plus harmonics into it. This current produces a voltage drop down non-negligible ground-wire resistances, and this effectively appears as a voltage source in each of the two signal lines. Since the CMRR is finite a proportion of this voltage will appear to be a differential signal, and will be reproduced as such.

Balanced connectors

Balanced connections are most commonly made with XLR connectors, though it can be done with stereo (tip-ring-sleeve) jack plugs. XLRs are a professional 3-pin format, and are a much better connector in every way than the usual phono (RCA) connectors used for unbalanced interconnections. Phono connectors have the great disadvantage that if you are connecting them with the system active (inadvisable, but then people are always doing inadvisable things) the signal contacts meet before the grounds and thunderous noises result. The XLR standard has Pin 2 as hot, Pin 3 as cold, and Pin 1 as ground.

Stereo jack plugs are often used for line level signals in a recording environment, and are frequently found on the rear of professional power amplifiers as an alternative to an adjacent XLR connector. Both full-size and 3.5 mm sizes are used. Balanced jacks are wired with the tip as hot, the ring as cold, and the sleeve as ground. Sound reinforcement systems often use large multiway connectors that carry dozens of 3-wire balanced connections.

Balanced signal levels

Many pieces of equipment, including preamplifiers and power amplifiers designed to work together, have both unbalanced and balanced inputs and outputs. The general consensus in the hifi world is that if the unbalanced output is say 1 Vrms, then the balanced output will be created by feeding the in-phase output to the hot output pin, and also to a unity-gain inverting stage, which drives the cold output pin with 1 Vrms phase-inverted. The total balanced output voltage between hot and cold pins is therefore 2 Vrms, and so the balanced input must have a gain of ½ or −6 dB relative to the unbalanced input to maintain consistent internal signal levels.

Electronic vs transformer balanced inputs

Balanced interconnections can be made using either transformer or electronic balancing. Electronic balancing has many advantages, such as low cost, low size and weight, superior frequency and transient response, and no low-frequency linearity problems. It may still be regarded as a second-best solution in some quarters, but the performance is more than adequate for most professional applications. Transformer balancing does have some advantages of its own, particularly for work in very hostile RF/EMC environments, but serious drawbacks. The advantages are that transformers are electrically bullet-proof, retain their high CMRR performance forever, and consume no power even at high signal levels. They are essential if galvanic isolation between ground is required. Unfortunately, transformers generate LF distortion, particularly if they have been made with minimal core sizes to save weight and cost. They are liable to have HF response problems due to leakage reactance and distributed capacitance, and compensating for this requires a carefully designed Zobel network across the secondary. Inevitably they are heavy and expensive. Transformer balancing is therefore relatively rare, even in professional audio applications, and the greater part of this chapter deals with electronically-balanced inputs.

Common mode rejection

Figure 18.3 shows a balanced interconnection reduced to its bare essentials; hot and cold line outputs with source resistances R_{out+}, R_{out-} and a standard differential amplifier at the input end. The output resistances are assumed to be exactly equal, and the balanced input in the receiving equipment has two exactly equal input resistances to ground R1, R2. The ideal balanced input amplifier senses the voltage difference between the points marked IN+ (hot) and IN− (cold) and ignores any common-mode voltage which are present on both. The amount by which it discriminates is called the common-mode rejection ratio or CMRR, and is usually measured in dB. Suppose a differential voltage input between IN+ and IN− gives an output voltage of 0 dB; now reconnect the input so that IN+ and IN− are joined together and the same voltage is applied between them and ground. Ideally the result would be zero output, but in this imperfect world it won't be, and the output could be anywhere between −20 dB (for a bad balanced interconnection, which probably has something wrong with it) and −140 dB (for an extremely good one). The CMRR when plotted may have a flat section

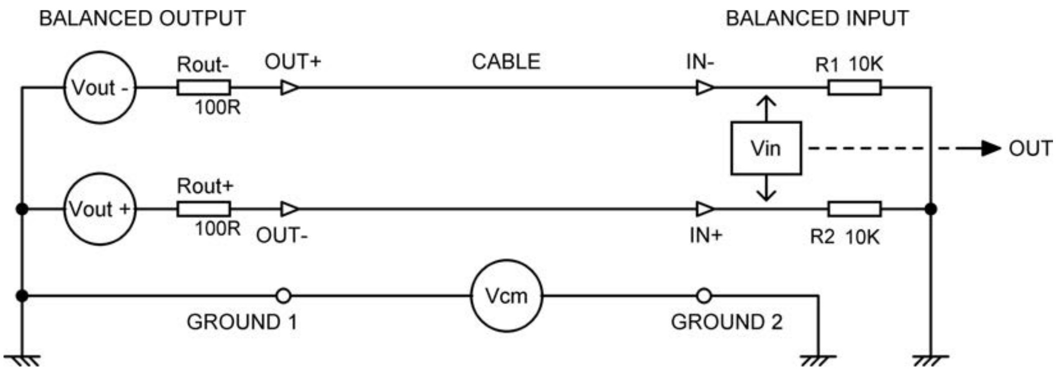

Figure 18.3: A theoretical balanced interconnection showing how the output and input impedances influence CMRR

at low frequencies, but it very commonly degrades at high audio frequencies, and may also deteriorate at very low frequencies. More on that later.

In one respect balanced audio connections have it easy. The common-mode signal is normally well below the level of the unwanted signal, and so the common-mode range of the input is not an issue. In other areas of technology, such as electrocardiogram amplifiers, the common-mode signal may be many times greater than the wanted signal.

The simplified conceptual circuit of Figure 18.3, under SPICE simulation, demonstrates the need to get the resistor values right for a good CMRR, before you even begin to consider the rest of the circuitry. The differential voltage sources V_{out+}, V_{out-} which represent the actual balanced output are set to zero, and V_{cm}, which represents the common-mode voltage drop down the cable ground, is set to 1 volt to give a convenient result in dBV. The output resulting from the presence of this voltage source is measured by a mathematical subtraction of the voltages at IN+ and IN− so there is no actual input amplifier to confuse the results with its non-ideal performance.

Let us begin with R_{out+} and R_{out-} set to 100 Ω and R1 and R2 set to 10 kΩ. These are typical real-life values as well as being nice round figures. When all four resistances are exactly at their nominal value, the CMRR is in theory infinite, which my SPICE simulator rather curiously reports as exactly −400 dB. If one of the output resistors or one of the input resistors is then altered in value by 1%, then the CMRR drops like a stone to −80 dB. If the deviation from equality is 10%, things are predictably worse and the CMRR degrades to −60 dB, as shown in Table 18.2. That would be quite a good figure in reality, but since we have not yet even thought about opamp imperfections or other circuit imbalances, and have only altered one resistance out of the four that will in real circuitry all have their own tolerances, it underlines the need to get things right at the most basic theoretical level before we dig deeper into the circuitry. The CMRR is naturally flat with frequency because our simple model has no frequency-dependent components.

TABLE 18.2 How resistor tolerances affect the theoretical CMRR of the theoretical circuit in Figure 18.3

R_{out+}	R_{out-}	R_{out} deviation	R1	R2	R1, R2 deviation	R_{in}/R_{out} ratio	CMRR (dB)
100	100	0	10k	10k	0	100	∞
100	101	1%	10k	10k	0	100	−80.2
100	110	10%	10k	10k	0	100	−60.2
100	100	0	10k	10.1k	1%	100	−80.3
100	100	0	10k	11k	10%	100	−61.0
100	100	0	100k	101k	1%	1000	−100.1
100	100	0	100k	110k	10%	1000	−80.8
100	100	0	1M	1.01M	1%	10,000	−120.1
100	100	0	1M	1.1M	10%	10,000	−100.8
68	68	0	20k	20.2k	1%	294	−89.5
68	68	0	20k	22k	10%	294	−70.3

The essence of the problem is that we have two resistive dividers, and to get an infinite CMRR they must have exactly the same attenuation. If we increase the ratio between the output and input resistors, by reducing the former or increasing the latter, the attenuation factor becomes closer to unity, so variations in either resistor value have less effect on it. If we increase the input impedance to 100 kΩ, which is quite practical in real life (we will put aside the noise implications of this for the moment) things are ten times better, as the R_{in}/R_{out} ratio has improved from 100 to 1000 times. We now get a CMRR of −100 dB with a 1% resistance deviation, and −80 dB with a 10% deviation. An even higher input impedance of 1 MΩ, which is perhaps a bit less practical, raises R_{in}/R_{out} to 10,000 and gives −120 dB for a 1% resistance deviation, and −100 dB for a 10% deviation.

We can attack the other aspect of the attenuation problem by reducing the output impedances to 10 Ω, ignoring for the moment the need to secure against HF instability caused by cable capacitance, and also return the input impedance resistors to 100 kΩ. R_{in}/R_{out} is 10,000 once more, and as you might suspect the CMRR is once more −120 dB for a 1% deviation, and −100 dB for a 10% deviation. Ways to make stable output stages with very low output impedances are described in Chapter 19; a fraction of an Ohm at 1 kHz is quite easy to achieve.

In conventional circuits, the combination of 68 Ω output resistors and a 20 kΩ input impedance is often encountered; 68 Ω is about as low as you want to go if HF instability is to be absolutely guarded against with the long lines used in professional audio. The 20 kΩ common-mode input impedance is what you get if you make a basic balanced input amplifier with four 10 kΩ resistors. I strongly suspect that this value is so popular because it looks as if it gives standard 10 kΩ input impedances – in fact it does nothing of the sort, and the common-mode input impedance, which is what matters here, is 20 kΩ on each leg; more on that later. It turns out that 68 Ω output resistors and a 20 kΩ input impedance give a theoretical CMRR of −89.5 dB for a 1% deviation of one resistor, which is quite encouraging. These results are summarised in Table 18.2.

The conclusion is simple: we need the lowest possible output impedances and the highest possible input impedances to get the maximum common-mode rejection. This is highly convenient because low output impedances are already needed to drive multiple amplifier inputs and cable capacitance, and high input impedances are needed to minimise loading and maximise the number of amplifiers that can be driven.

The basic electronic balanced input

Figure 18.4 shows the basic balanced input amplifier using a single opamp. To achieve balance R1 must be equal to R3 and R2 equal to R4. It has a gain of R2/R1 (=R4/R3). The standard one-opamp balanced input or differential amplifier is a very familiar circuit block, but its operation often appears somewhat mysterious. Its input impedances are not equal when it is driven from a balanced output; this has often been commented on [4], and some confusion has resulted.

The source of the confusion is that a simple differential amplifier has interaction between the two inputs, so that the input impedance seen on the cold input depends on the signal applied to the hot input. Input impedance is measured by applying a signal and seeing how much current flows into the input, so it follows that the apparent input impedance on each leg varies according to how the cold input is driven. If the amplifier is made with four 10 kΩ resistors, then the input impedances on hot and cold are as Table 18.3.

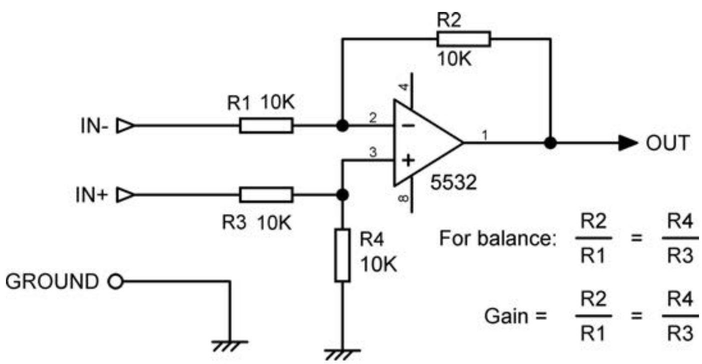

Figure 18.4: The basic balanced input amplifier

TABLE 18.3 The input impedances for different input drive conditions

Case	Pins driven	Hot input res (kΩ)	Cold input res (kΩ)
1	Hot only	20	Grounded
2	Cold only	Grounded	10
3	Both (balanced)	20	6.66
4	Both common-mode	20	20
5	Both floating	10	10

Some of these impedances are not exactly what you might expect, and require some explanation.

Case 1: The balanced input is being used as an unbalanced input by grounding the cold input and driving the hot input only. The input impedance is therefore simply R3 + R4. Resistors R3 and R4 reduce the signal by a factor of a half, but this loss is undone as R1 and R2 set the amplifier gain to two times, and the overall gain is unity. If the cold input is not grounded then the gain is 0.5 times. The attenuate-then-amplify architecture, plus the Johnson noise from the resistors, makes this configuration much noisier than the dedicated unbalanced input of Figure 18.1, which has only a single 100 Ω resistor in the signal path.

Case 2: The balanced input is again being used as an unbalanced input, but this time by grounding the hot input, and driving the cold input only. This gives a phase inversion and it is unlikely you would want to do it except as an emergency measure to correct a phase error somewhere else. The important point here is that the input impedance is now only 10 kΩ, the value of R1, because shunt negative feedback through R2 creates a virtual earth at Pin 2 of the opamp. Clearly this simple circuit is not as symmetrical as it looks. The gain is unity, whether or not the hot input is grounded; grounding it is desirable because it not only prevents interference being picked up on the hot input pin, but also puts R3 and R4 in parallel, reducing the resistance from opamp Pin 3 to ground and so reducing Johnson noise.

Case 3: This is the standard balanced interconnection. The input is driven from a balanced output with the same signal levels on hot and cold, as if from a transformer with its centre-tap grounded, or an electronically balanced output using a simple inverter to drive the cold pin. The input impedance on the hot input is what you would expect; R3 + R4 add up to 20 kΩ. However, on the cold input there is a much lower input impedance of 6.66 kΩ. This at first sounds impossible as the first thing the signal encounters is a 10 kΩ series resistor, but the crucial point is that the hot input is being driven simultaneously with a signal of the opposite phase, so the inverting opamp input is moving in the opposite direction to the cold input due to negative feedback, and what you might call anti-bootstrapping reduces the effective value of the 10 kΩ resistor to 6.66 kΩ. These are the differential input impedances we are examining, the impedances seen by the balanced output driving them. Common-mode signals see a common-mode impedance of 20 kΩ, as in Case 4 below.

You will sometimes see the statement that these unequal differential input impedances 'unbalance the line'. From the point of view of CMRR, this is not the case, as it is the CM input impedance that counts. The line is, however, unbalanced in the sense that the cold input draws three times the current from the output that the hot one does. This current imbalance might conceivably lead to inductive crosstalk in some multi-way cable situations, but I have never encountered it. The differential input impedances can be made equal by increasing the R1 and R2 resistor values by a factor of three, but this degrades the noise performance markedly and makes the common-mode impedances to ground unequal, which is a much worse situation as it compromises the rejection of ground voltages, and these are almost always the main problem in real life.

Case 4: Here both inputs are driven by the same signal, representing the existence of a common-mode voltage. Now both inputs show an impedance of 20 kΩ. It is the symmetry of the common-mode input impedances that determines how effectively the balanced input rejects the common-mode signal. This configuration is of course only used for CMRR testing.

Case 5: Now the input is driven as from a floating transformer with the centre-tap (if any) unconnected, and the impedances can be regarded as equal; they must be, because with a floating winding the same current must flow into each input. However, in this connection the line voltages are *not* equal and opposite: with a true floating transformer winding the hot input has all the signal voltage on it while the cold has none at all, due to the negative feedback action of the balanced input amplifier. This seemed very strange when it emerged in SPICE simulation, but a sanity check with real components proves it true. The line has been completely unbalanced as regards crosstalk to other lines, although its own common-mode rejection remains good.

Even if absolutely accurate resistors are assumed, the CMRR of the stage in Figure 18.4 is not infinite; with a TL072 it is about −90 dB, degrading from 100 Hz upwards, due to the limited open-loop gain of the opamp. We will now examine this effect.

Common-mode rejection: the basic balanced input and opamp effects

In the earlier section on CMRR we saw that in a theoretical balanced line, choosing low output impedances and high input impedances would give very good CM rejection even if the resistors were not perfectly matched. Things are a bit more complex (i.e. worse) if we replace the mathematical subtraction with a real opamp. We quickly find that even if perfectly matched resistors everywhere are assumed, the CMRR of the stage is not infinite, because the two opamp inputs are not at exactly the same voltage. The negative feedback error-voltage between the inputs depends on the open-loop gain of the opamp, and that is neither infinite nor flat with frequency into the far ultra-violet. Far from it. There is also the fact that opamps themselves have a common-mode rejection ratio; it is high, but once more it is not infinite.

As usual, SPICE simulation is instructive, and Figure 18.5 shows a simple balanced interconnection, with the balanced output represented simply by two 100 Ω output resistances connected to the source equipment ground, here called Ground 1, and the usual differential opamp configuration at the input end, where we have Ground 2.

A common-mode voltage V_{cm} is now injected between Ground 1 and Ground 2, and the signal between the opamp output and Ground 2 measured. The balanced input amplifier has all four of its resistances set to precisely 10 kΩ, and the opamp is represented by a very simple model that has only two parameters; a low-frequency open-loop gain, and a single pole frequency that says where that gain begins to roll-off at 6 dB per octave. The opamp input impedances and the opamp's own CMRR are assumed infinite, as in the world of simulation they so easily can be. Its output impedance is set at zero.

For the first experiments, even the pole frequency is made infinite, so now the only contact with harsh reality is that the opamp open-loop gain is finite. That is however enough to give distinctly non-ideal CMRR figures, as Table 18.4 shows.

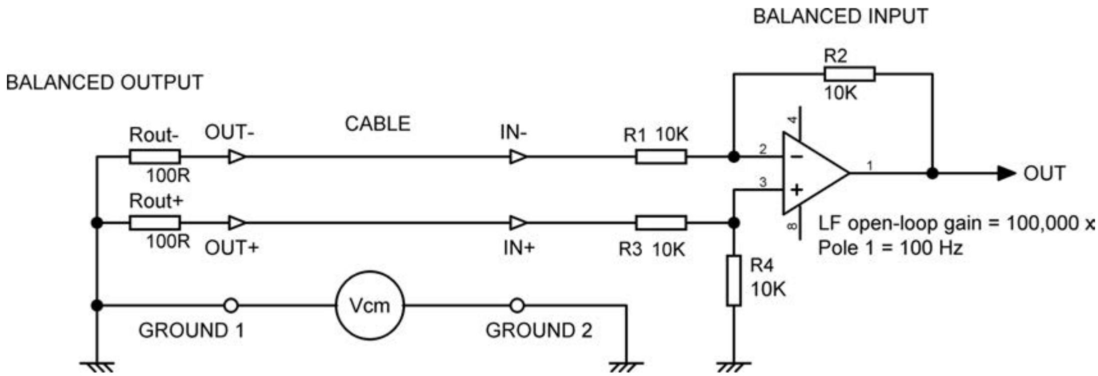

Figure 18.5: A simple balanced interconnection for SPICE simulation to show the effect that opamp properties have on the CMRR

TABLE 18.4 The effect of finite opamp gain on CMRR for the circuit of Figure 18.5

Open-loop gain	CMRR (dB)	CMRR ratio
10,000	−74.0	19.9×10^{-5}
30,000	−83.6	66.4×10^{-6}
100,000	−94.0	19.9×10^{-6}
300,000	−103.6	6.64×10^{-6}
1,000,000	−114.1	1.97×10^{-6}

With a low-frequency open-loop gain of 100,000, which happens to be the typical figure for a 5532 opamp, even perfect components everywhere will never yield a better CMRR than −94 dB. The CMRR is shown as a raw ratio in the third column so you can see that the CMRR is inversely proportional to the gain, and so we want as much gain as possible.

Opamp frequency response effects

To examine these we will set the low-frequency gain to 100,000 which gives a CMRR 'floor' of −94 dB, and then introduce the pole frequency that determines where it rolls-off. The CMRR now worsens at 6 dB/octave, starting at a frequency set by the interaction of the low-frequency gain and the pole frequency. The results are summarised in Table 18.5 which shows that as you might expect, the lower the open-loop bandwidth of the opamp, the lower the frequency at which the CMRR begins to fall off. Figure 18.6 shows the situation diagrammatically.

Table 18.6 gives the open-loop gain and pole parameters for a few opamps of interest. Both parameters, but especially the gain, are subject to considerable variation; the typical values from the manufacturers' data sheets are given here.

TABLE 18.5 The effect of opamp open-loop pole frequency on CMRR for the circuit of Figure 18.5

Pole frequency	CMRR breakpoint freq
10 kHz	10.2 kHz
1 kHz	1.02 kHz
100 Hz	102 Hz
10 Hz	10.2 Hz

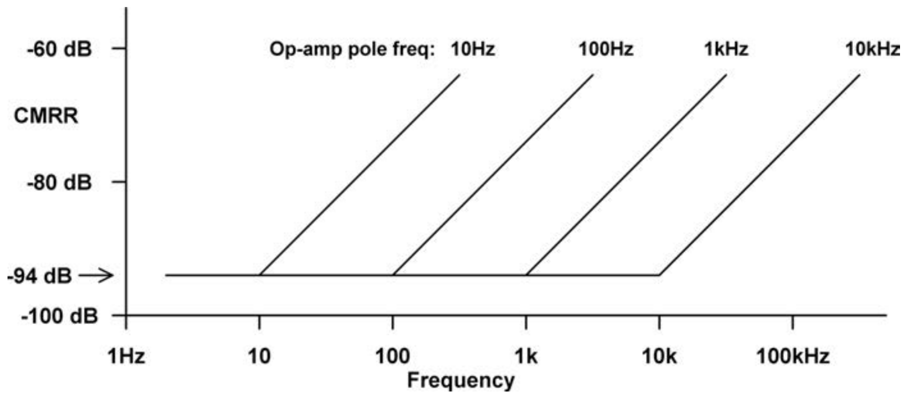

Figure 18.6: How the CMRR degrades with frequency for different opamp pole frequencies. All resistors are assumed to be perfectly matched

TABLE 18.6 Typical LF gain and open-loop pole frequency for some opamps commonly used in audio

Name	Input device type	LF gain	Pole freq (Hz)	Opamp LF CMRR dB
NE5532	Bipolar	100,000	100	100
LM4562	Bipolar	10,000,000	below 10	120
LT1028	Bipolar	20,000,000	3	120
TL072	FET	200,000	20	86
OP27	FET	1,800,000	3	120
OPA2134	FET	1,000,000	3	100
OPA627	FET	1,000,000	20	116

Some of these opamps have very high open-loop gains, but only at very low frequencies. This may be good for DC applications, but in audio line input applications, where the lowest frequency of CMRR interest is 50 Hz, they will be operating above the pole frequency and so the gain available will be less – possibly considerably so, in the case of opamps like the OPA2134. This is not however a real limitation, for even if a humble TL072 is used the perfect-resistor CMRR is about −90 dB, degrading from 100 Hz upwards. This sort of performance is not attainable in practice. We will shortly see why not.

Opamp CMRR effects

Opamps have their own common-mode rejection ratio, and we need to know how much this will affect the final CMRR of the balanced interconnection. The answer is that if all resistors are accurate, the overall CMRR is equal to the CMRR of the opamp [5]. Since opamp CMRR is typically very high (see the examples in Table 18.6) it is very unlikely to be the limiting factor.

The CMRR of an opamp begins to degrade above a certain frequency, typically at 6 dB per octave. This is (fortunately) at a higher frequency than the open-loop pole, and is frequently around 1 kHz. For example the OP27 has a pole frequency at about 3 Hz, but the CMRR remains flat at 120 dB until 2 kHz, and it is still greater than 100 dB at 20 kHz.

Amplifier component mismatch effects

We saw earlier in this chapter that when the output and input impedances on a balanced line have a high ratio between them and are accurately matched we got a very good CMRR; this was compromised by the imperfections of opamps, but the overall results were still very good – and much higher than the CMRRs measured in practice. There remains one place where we are still away in theory-land; we have so far assumed the resistances around the

TABLE 18.7 How resistor tolerances affect the CMRR with some realistic opamp o/l gains

R1 (kΩ)	R1 deviation (%)	Gain ×	CMRR (dB)
10	0	100,000	−94.0
10.001	0.01	100,000	−90.6
10.01	0.1	100,000	−66.5
10.1	1	100,000	−46.2
11	10	100,000	−26.6
10	0	1,000,000	−114.1
10.001	0.01	1,000,000	−86.5
10.01	0.1	1,000,000	−66.2
10.1	1	1,000,000	−46.2
11	10	1,000,000	−26.6

opamp were all exactly accurate. We must now face reality, admit that these resistors will not be perfect, and see how much damage to the CMRR they will do.

SPICE simulation gives us Table 18.7. The situation with LF opamp gains of both 100,000 and 1,000,000 is examined, but the effects of finite opamp bandwidth or opamp CMRR are not included. R1 in Figure 18.5 is varied while R2, R3 and R4 are all kept at precisely 10 kΩ, and the balanced output source impedances are set to exactly 100 Ω.

Table 18.7 shows with glaring clarity that our previous investigations, which took only output and input impedances into account, and determined that 68 Ω output resistors and 20 kΩ input impedances gave a CMRR of −89.5 dB for a 1% deviation in either, were actually quite unrealistic, and even adding in opamp imperfections left us with unduly optimistic results. If a 1% tolerance resistor is used for R1, (and nowadays there is no financial incentive to use anything less accurate) the CMRR is dragged down at once to −46 dB; the same figure results from varying any other one of the four resistances by itself. If you are prepared to shell out for 0.1% tolerance resistors, the CMRR is a rather better −66 dB.

This shows that there really is no point in worrying about the gain of the opamp you use in balanced inputs; the effect of mismatches in the resistors around that opamp are far greater.

The results in the table give an illustration of how resistor accuracy affects CMRR, but it is only an illustration, because in real life – a phrase that seems to keep cropping up, showing how many factors affect a practical balanced interconnection – all four resistors will of course be subject to a tolerance, and a more realistic calculation would produce a statistical distribution of CMRR rather than a single figure. One method is to use the Monte Carlo function in SPICE, which runs multiple simulations with random component variations and collates the results. However you do it, you must know (or assume) how the resistor values are distributed within their tolerance window. Usually you don't know, and finding out by measuring hundreds of resistors is not a task that appeals to all of us.

It is straightforward to assess the worst-case CMRR, which occurs when all resistors are at the limit of the tolerance in the most unfavourable direction. The CMRR in dB is then:

$$\mathrm{CMRR} = 20\log = \left(\frac{1 + R2/R1}{4T/100}\right) \qquad \text{(Eqution 18.1)}$$

Where R1 and R2 are as in Figure 18.5, and T is the tolerance in %.

This deeply pessimistic equation tells us that 1% resistors give a worst-case CMRR of only 34.0 dB, that 0.5% parts give only 40.0 dB and expensive 0.1% parts yield but 54.0 dB. Things are not however quite that bad in actuality, as the chance of everything being as wrong as possible is actually very small. I have measured the CMRR of more of these balanced inputs, built with 1% resistors, than I care to contemplate, but I do not ever recall that I ever saw one with an LF CMRR worse than 40 dB.

There are 8-pin SIL packages that offer four resistors that ought to have good matching, if not accurate absolute values; be very, very wary of these as they usually contain thick-film resistive elements that are not perfectly linear. In a test I did a 10 kΩ SIL resistor with 10 Vrms across it generated 0.0010% distortion. Not a huge amount perhaps, but in the quest for perfect audio, resistors that do not stick to Ohm's Law are not a good start.

To conclude this section, it is clear that in practical use it is the errors in the balanced amplifier resistors that determine the CMRR, though both unbalanced capacitances (C1, C2 in Figure 18.9 below) and the finite opamp bandwidth are likely to cause further degradation at high audio frequencies. If you are designing both ends of a balanced interconnection and you are spending money on precision resistors, you should put them in the input amplifier, not the balanced output. The LF gain of the opamp, and opamp CMRR, have virtually no effect.

Balanced input amplifiers made with four 1% resistors are used extensively in the professional audio business, and almost always prove to have adequate CMRR for the job. When more CMRR is thought desirable, for example in high-end mixing consoles, one of the resistances is made trimmable with a preset, as in Figure 18.7. This means a bit of tweaking in manufacture, but the upside is that it is a quick set-and-forget adjustment that will not need to be touched again unless one of the four resistors needs replacing, and that is extremely unlikely. CMRRs at LF of more than 80 dB can easily be obtained by this method, but the CMRR at HF will degrade due to the opamp gain roll-off and stray capacitances.

Figure 18.8 shows the CMRR measurements for a trimmed balanced input amplifier. The flat line at −50 dB was obtained from a standard fixed-resistor balanced input using four 1% 10 kΩ unselected resistors, while the much better (at LF, anyway) trace going down to

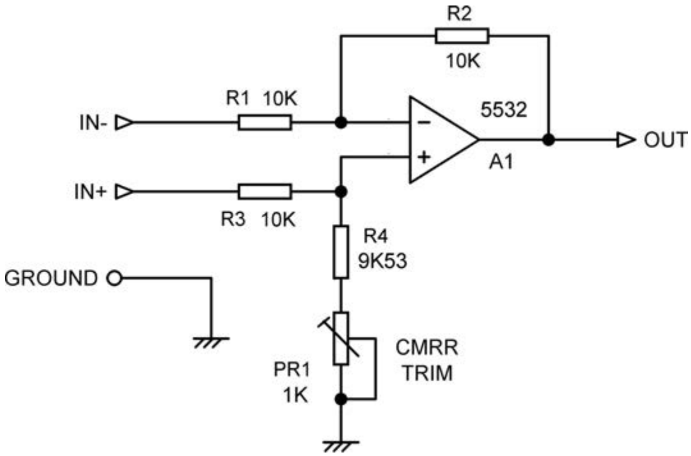

Figure 18.7: A balanced input amplifier, with preset pot to trim for best LF CMRR

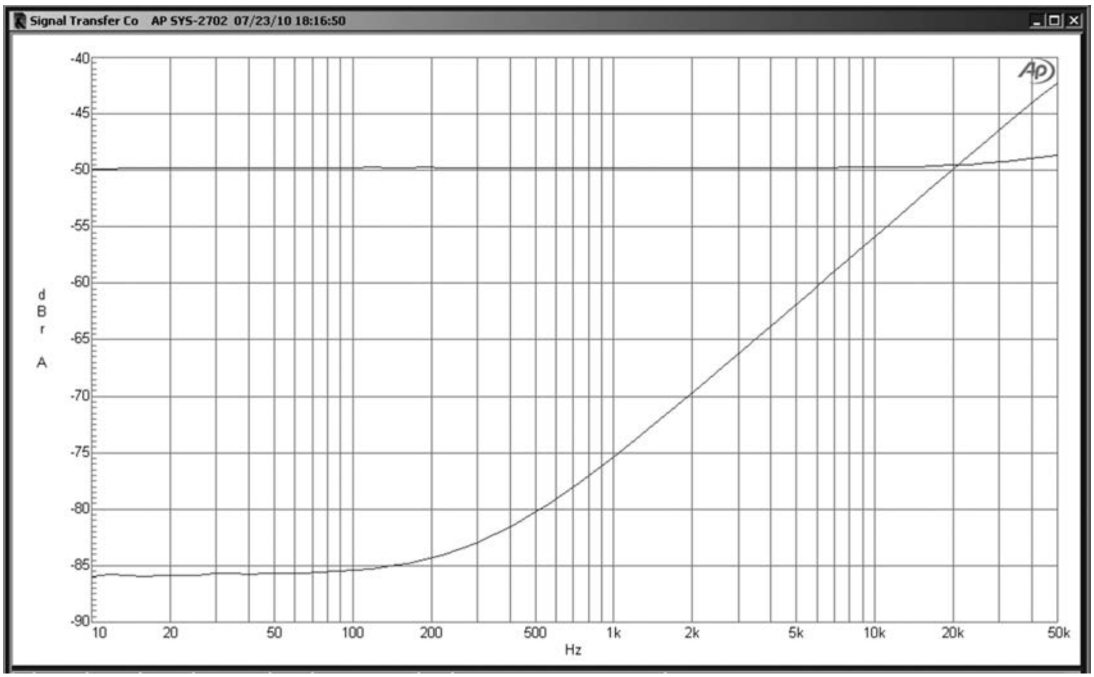

Figure 18.8: The CMRR of a fixed-resistor balanced amplifier compared with the trimmed version. The opamp was a 5532 and all resistors were 1%. The trimmed version gives better than 80 dB CMRR up to 500 Hz

−85 dB was obtained from Figure 18.7 by using a multi-turn preset for PR1. Note that R4 is an E96 value so a 1k preset can swing the total resistance of that arm both above and below the nominal 10 kΩ. The CMRR is dramatically improved by more than 30 dB in the region 50–500 Hz where ground noise tends to intrude itself, and is significantly better across almost all the audio spectrum.

The upward sloping part of the trace in Figure 18.8 is partly due to the finite open-loop bandwidth of the opamp, and partly due to unbalanced circuit capacitances. The CMRR is actually worse than 50 dB above 20 kHz, due to the stray capacitances in the multi-turn preset. In practice the value of PR1 would be smaller, and a one-turn preset with much less stray capacitance used. Still, I think you get the point: trimming can be both economic and effective.

The only thing that could go wrong here is that vibration might affect preset adjustments in on-the-road mixing consoles. This can be prevented by using good quality parts; multi-turn presets stay where they are put, but as noted their stray capacitance can be a problem and they are relatively expensive parts to apply to every line input.

A practical balanced input

The simple balanced input circuit shown in Figure 18.4 is not fit to face the outside world without additional components. Figure 18.9 shows a fully equipped version. Firstly, and most important, C1 has been added across the feedback resistor R2; this prevents stray capacitances from Pin 2 to ground causing extra phase-shifts that lead to HF instability. The value required for stability is small, much less than that which would cause an HF roll-off anywhere near the

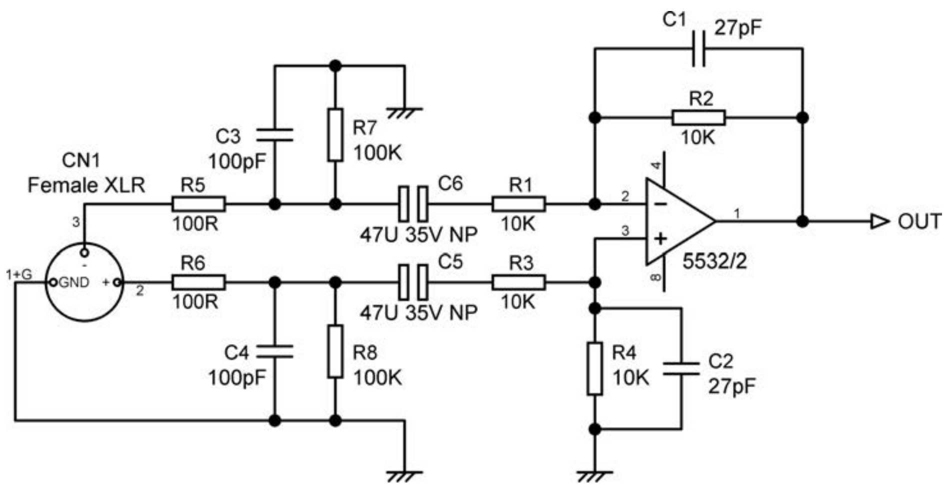

Figure 18.9: Balanced input amplifier with the extra components required for DC blocking and EMC immunity

top of the audio band. The values here of 10k and 27 pF give −3 dB at 589 kHz, and such a roll-off is only down by 0.005 dB at 20 kHz. C2, of equal value, must be added across R4 to maintain the balance of the amplifier, and hence its CMRR, at high frequencies.

C1 and C2 must not be relied upon for EMC immunity as C1 is not connected to ground, and there is every chance that RF will demodulate at the opamp inputs. A passive RF filter is therefore added to each input, in the shape of R5, C3 and R6, C4, so the capacitors will shunt incoming RF to ground before it reaches the opamp. Put these as close to the input socket as possible to minimise radiation inside the enclosure.

I explained earlier in this chapter when looking at unbalanced inputs that it is not easy to guess what the maximum source impedance will be, given the existence of 'passive preamplifiers' and valve equipment. Neither is likely to have a balanced output, unless implemented by a transformer, but either might be used to feed a balanced input, and so the matter needs thought.

In the unbalanced input circuit resistances had to be kept as low as practicable to minimise the generation of Johnson noise that would compromise the inherently low noise of the stage. The situation with a standard balanced input is however different from the unbalanced case as there have to be resistances around the opamp, and they must be kept up to a certain value to give acceptably high input impedances; this is why a balanced input like this one is much noisier. We could therefore make R5 and R6 much larger without a measurable noise penalty if we reduce R1 and R3 accordingly to keep unity gain. In Figure 18.9 R5 and R6 are kept at 100 Ω, so if we assume 50 Ω output resistances in both legs of the source equipment, then we have a total of 150 Ω, and 150 Ω and 100 pF give −3 dB at 10.6 MHz. Returning to a possible passive preamplifier with a 10 kΩ potentiometer, its maximum output impedance of 2.5k plus 100 Ω with 100 pF gives −3 dB at 612 kHz, which remains well clear of the top of the audio band.

As with the unbalanced input, replacing R5 and R6 with small inductors will give much better RF filtering but at increased cost. Ideally a common-mode choke (two bifilar windings on a small toroidal core) should be used as this improves performance. Check the frequency response to make sure the LC circuits are well-damped and not peaking at the turnover frequency.

C5 and C6 are DC-blocking capacitors. They must be rated at no less than 35 V to protect the input circuitry, and are the non-polarised type as external voltages are of unpredictable polarity. The lowest input impedance that can occur with this circuit when using 10 kΩ resistors, is, as described above, 6.66 kΩ when it is being driven in the balanced mode. The low-frequency roll-off is therefore −3 dB at 0.51 Hz. This may appear to be undesirably low, but the important point is not the LF roll-off but the possible loss of CMRR at low frequencies due to imbalance in the values of C5 and C6; they are electrolytics with a significant tolerance. Therefore they should be made large so their impedance is a small part of the total input impedance. 47 μF is shown here but 100 μF or 220 μF can be used to advantage if there is the space to fit them in. The low-end frequency response must be defined somewhere in the input system, and the earlier the better, to prevent headroom or linearity being affected by subsonic disturbances, but this is not a good place to do it. A suitable time-constant immediately after the input amplifier is the way to go, but remember that capacitors used as time-constants may distort unless they are NP0 ceramic, polystyrene, or polypropylene (see Chapter 2 for more on this).

R7, R8 are DC drain resistors to prevent charges lingering on C5 and C6. These can be made lower than for the unbalanced input as the input impedances are lower, so a value of say 100 kΩ rather than 220 kΩ makes relatively little difference to the total input impedance.

A useful property of this kind of balanced amplifier is that it does not go mad when the inputs are left open-circuit – in fact it is actually *less* noisy than with its inputs shorted to ground. This is the opposite of the 'normal' behaviour of a high-impedance unterminated input. This is because two things happen; open-circuiting the hot input doubles the resistance seen by the non-inverting input of the opamp, raising its noise contribution by 3 dB. However, opening the cold input makes the noise gain drop by 6 dB, giving a net drop in noise output of approx. 3 dB. This of course refers only to the internal noise of the amplifier stage, and pickup of external interference is always possible on an unterminated input. The input impedances here are modest, however, and the problem is less serious than you might think. Having said that, deliberately leaving inputs unterminated is always bad practice.

If this circuit is built with four 10 kΩ resistors and a 5532 opamp section, the noise output is −104.8 dBu with the inputs terminated to ground via 50 Ω resistors. As noted above, the input impedance of the cold input is actually lower than the resistor connected to it when working balanced, and if it is desirable to raise this input impedance to 10 kΩ, it could be done by raising the four resistors to 16 kΩ; this slightly degrades the noise output to −103.5 dBu. Table 18.8 gives some examples of how the noise output depends on the resistor value; the third column gives the noise with the input unterminated, and shows that in each case the amplifier is about 3 dB quieter when open-circuited. It also shows that a useful improvement in noise performance is obtained by dropping the resistor values to the lowest that a 5532 can easily drive (the opamp has to drive the feedback resistor), though this usually gives unacceptably low input impedances (there is more on that at the end of the chapter).

TABLE 18.8 Noise output measured from simple balanced amps using a 5532 section

R value (Ω)	50 Ω terminated inputs (dBu)	Open-circuit inputs (dBu)	Terminated/open difference (dBu)
100k	−95.3	−97.8	2.5
10k	−104.8	−107.6	2.8
2k0	−109.2	−112.0	2.8
820	−111.7	−114.5	2.8

Variations on the balanced input stage

I now give a collection of balanced input circuits that offer advantages or extra features over the standard balanced input configuration. The circuit diagrams often omit stabilising capacitors, input filters, and DC-blocking capacitors to improve the clarity of the basic principle. They can easily be added; in particular bear in mind that a stabilising capacitor like C1 in Figure 18.9 is often needed to guarantee freedom from high-frequency oscillation, and C2, of equal value, must also be added to maintain balance at HF.

Combined unbalanced and balanced inputs

If both unbalanced and balanced inputs are required, it is extremely convenient if it can be arranged so that no switching between them is required. Switches cost money, mean more holes in the metalwork, and add to assembly time. Figure 18.10 shows an effective way to implement this. In balanced mode, the source is connected to the balanced input and the unbalanced input left unterminated. In unbalanced mode, the source is connected to the unbalanced input and the balanced input left unterminated, and no switching is required. It might appear that these unterminated inputs would pick up extra noise, but in practice this is not the case. It works very well and I have used it successfully in high-end equipment for two prestigious manufacturers.

As described above, in the world of hifi balanced signals are at twice the level of the equivalent unbalanced signals, and so the balanced input must have a gain of ½ or −6 dB relative to the unbalanced input to get the same gain by either path. This is done here by increasing R1 and R3 to 20 kΩ. The balanced gain can be greater or less than unity, but the

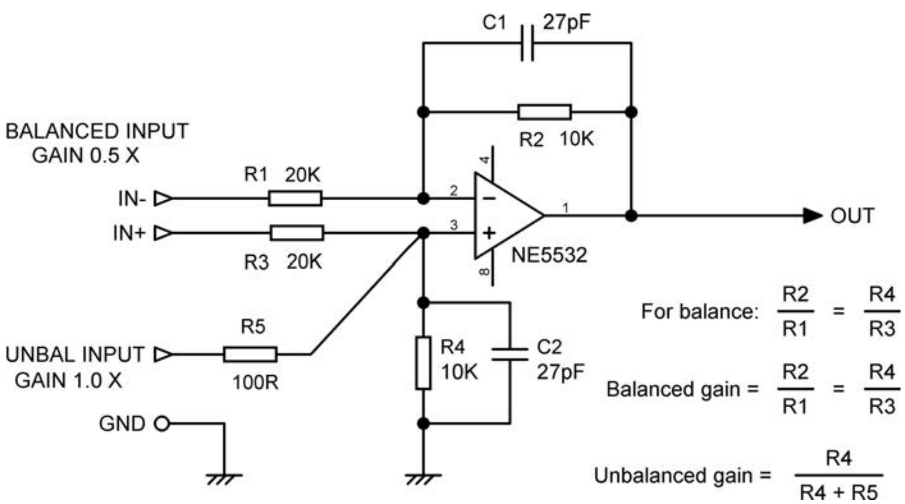

Figure 18.10: Combined balanced and unbalanced input amplifier with no switching required

gain via the unbalanced input is always approximately unity unless R5 is given a higher value to deliberately introduce attenuation.

There are two minor compromises in this circuit which need to be noted. Firstly, the noise performance in unbalanced mode is worse than for the dedicated unbalanced input described earlier in this chapter, because R2 is effectively in the signal path and adds Johnson noise. Secondly, the input impedance of the unbalanced input cannot be very high because it is set by R4, and if this is increased in value all the resistances must be increased proportionally and the noise performance will be markedly worse. It is important that only one input cable should be connected at a time, because if an unterminated cable is left connected to an unused input, the cable capacitance to ground can cause frequency response anomalies and might, in adverse circumstances, cause HF oscillation. A prominent warning on the back panel and in the manual is a very good idea.

The Superbal input

This version of the balanced input amplifier, shown in Figure 18.11, has been referred to as the 'Superbal' circuit because it gives equal impedances into the two inputs for differential signals. It was publicised by David Birt of the BBC [6], but I have been told it was invented by Ted Fletcher at Alice. With the circuit values shown the differential input impedance is exactly 10 kΩ via both hot and cold inputs. The common-mode input impedance is 20 kΩ as before.

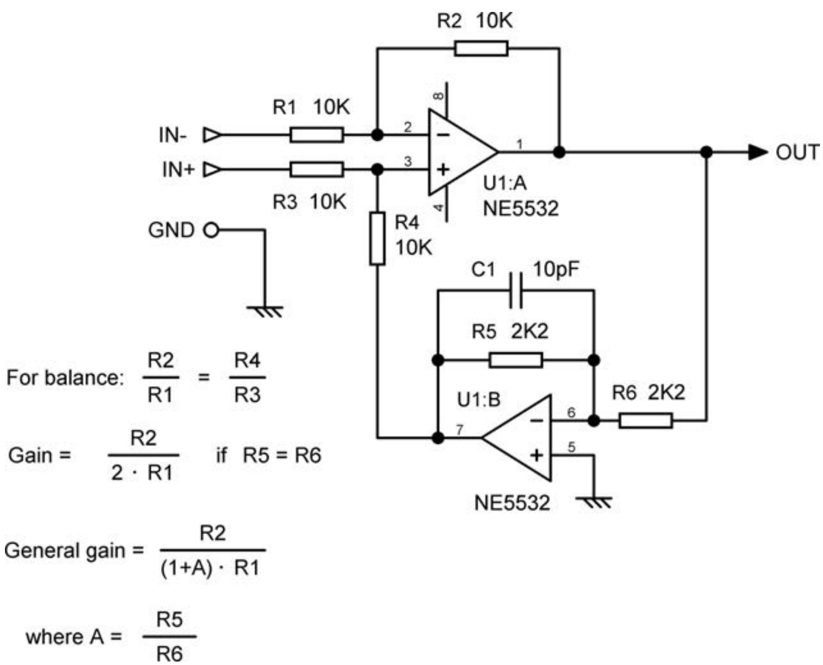

Figure 18.11: The Superbal balanced input amplifier

In the standard balanced input R4 is connected to ground, but here its lower end is actively driven with an inverted version of the output signal, giving symmetry. The increased amount of negative feedback reduces the gain with four equal resistors to – 6 dB instead of unity. The gain can be reduced below −6 dB by giving the inverter a gain of more than one; if R1, R2, R3 and R4 are all equal, the gain is 1/(A+1), where A is the gain of the inverter stage. This is of limited use as the inverter U1:B will now clip before the forward amplifier U1:A, reducing headroom. If the gain of the inverter stage is gradually reduced from unity to zero, the stage slowly turns back into a standard balanced amplifier with the gain increasing from −6 dB to unity and the input impedances becoming less and less equal. If a gain of less than unity is required it should be obtained by increasing R1 and R3.

R5 and R6 should be kept as low in value as possible to minimise Johnson noise; there is no reason why they have to be equal in value to R1, etc. The only restriction is the ability of U1:A to drive R6 and U1:B to drive R5, both resistors being effectively grounded at one end. The capacitor C1 will almost certainly be needed to ensure HF stability; the value in the figure is only a suggestion. It should be kept as small as possible because reducing the bandwidth of the inverter stage impairs CMRR at high frequencies.

Switched-gain balanced inputs

A balanced input stage that can be switched to two different gains while maintaining CMRR is very useful. Equipment often has to give optimal performance with both semi-pro (−7.8 dBu) and professional (+4 dBu) input levels. If the nominal internal level of the system is in the normal range of −2 to −6 dBu, the input stage must be able to switch between amplifying and attenuating, while maintaining good CMRR in both modes.

The brute-force way to change gain in a balanced input stage is to switch the values of either R1 and R3, or R2 and R4, in Figure 18.4, keeping the pairs equal in value to maintain the CMRR; this needs a double–pole switch for each input channel. A much more elegant technique is shown in Figure 18.12. Perhaps surprisingly, the gain of a differential amplifier can be manipulated by changing the drive to the feedback arm (R2 etc.) only, and leaving the other arm R4 unchanged, without affecting the CMRR. The essential point is to keep the source resistance of the feedback arm the same, but drive it from a scaled version of the opamp output. Figure 18.12 does this with the network R5, R6, which has a source resistance made up of 6k8 in parallel with 2k2, which is 1.662 kΩ. This is true whether R6 is switched to the opamp output (low gain setting) or to ground (high gain setting), for both have effectively zero impedance. For low gain the negative feedback is not attenuated, but fed through to R2 and R7 via R5, R6 in parallel. For high gain R5 and R6 become a potential divider, so the amount of feedback is decreased and the gain increased. The value of R2 + R7 is reduced from 7k5 by 1.662 kΩ to allow for the source impedance of the R5, R6 network; this requires the distinctly non-standard value of 5.838 kΩ, which is here approximated by R2

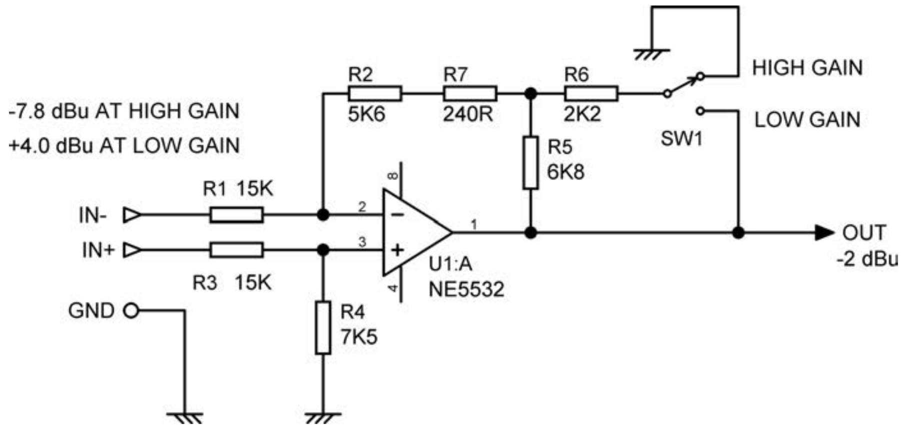

Figure 18.12: A balanced input amplifier with gain switching that maintains good CMRR

and R7 which give 5.6 kΩ + 240 Ω = 5.840 kΩ. This value is the best that can be done with E24 resistors; it is obviously out by 2 Ω, but that is much less than a 1% tolerance on R2, and so will have only a vanishingly small effect on the CMRR.

Note that this stage can attenuate as well as amplify if R1, R3 are set to be greater than R2, R4, as shown here. The nominal output level of the stage is assumed to be −2 dBu; with the values shown the two gains are −6.0 and +6.2 dB, so +4 dBu and −7.8 dBu respectively will give −2 dBu at the output. Other pairs of gains can of course be obtained by changing the resistor values; the important thing is to stick to the principle that the value of R2 + R7 is reduced from the value of R4 by the source impedance of the R5, R6 network. With the values shown the differential input impedance is 11.25 kΩ via the cold and 22.5 kΩ via the hot input. The common-mode input impedance is 22.5 kΩ.

This neat little circuit has the added advantage that nothing bad happens when the switch is moved with the circuit operating. When the wiper is between contacts you simply get a gain intermediate between the high and low settings, which is pretty much the ideal situation.

Variable-gain balanced inputs

The beauty of a variable-gain balanced input is that it allows you to get the incoming signal up or down to the nominal internal level as soon as possible, minimising both the risk of clipping and contamination with circuit noise. The obvious method of making a variable-gain differential stage is to use dual-gang pots to vary R2, R4 together, to maintain CMRR (varying R1, R3 would also alter the input impedances). This is clumsy, and gives a CMRR that is both bad and highly variable due to the inevitable mismatches between pot sections. For a stereo input the required 4-gang pot is an unappealing proposition.

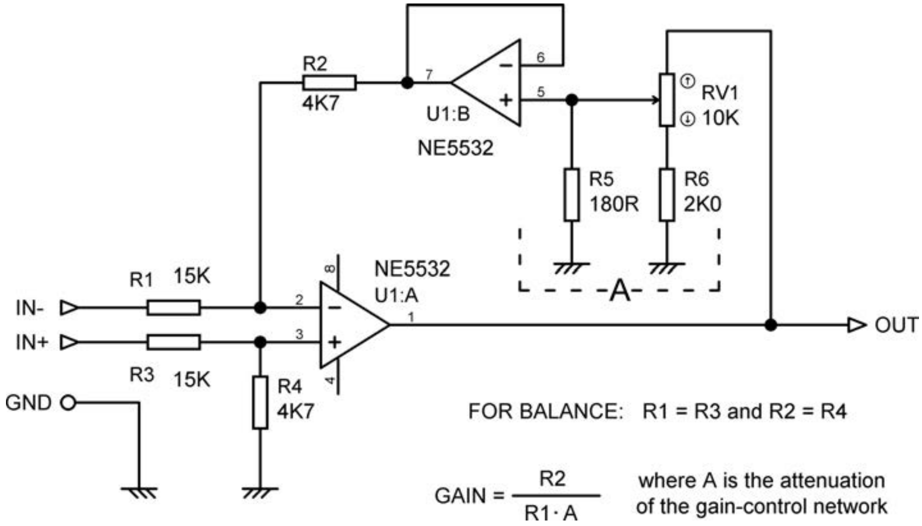

Figure 18.13: Variable-gain balanced input amplifier: gain +20 to −10 dB

There is, however, a way to get a variable gain with good CMRR using a single pot section. The principle is essentially the same as for the switched-gain amplifier above; keep constant the source impedance driving the feedback arm, but vary the voltage applied. The principle is shown in Figure 18.13. To the best of my knowledge I invented this circuit in 1982; any comments on this point are welcome. The feedback arm R2 is driven by voltage-follower U1:B. This eliminates the variations in source impedance at the pot wiper, which would badly degrade the CMRR. R6 sets the maximum gain and R5 modifies the gain law to give it a more usable shape. The minimum gain is set by the R2/R1 ratio. When the pot is fully up (minimum gain) R5 is directly across the output of U1:A so do not make it too low in value. If a centre-detent pot is used to give a default gain setting, this may not be very accurate as it partly depends on the ratio of pot track (no better than ±20% tolerance, and sometimes worse) to 1% fixed resistors.

This configuration is very useful as a general line input with an input sensitivity range of −20 to +10 dBu. For a nominal output of 0 dBu, the gain of Figure 18.13 is therefore +20 to −10 dB, with R5 chosen to give 0 dB gain at the central wiper position. An opamp in a feedback path may appear a dubious proposition for HF stability, because of the extra phase-shift it introduces, but here it is working as a voltage-follower, so its bandwidth is maximised and in practice the circuit is dependably stable.

This configuration can be improved. The maximum gain in Figure 18.13 is set by the end-stop resistor R6 at the bottom of the pot. The pot track resistance will probably be specified at ± 20%, while the resistor will be a much more precise 1%. The variation of pot track resistance can therefore cause quite significant differences between the two channels. It would be nice to find a way to eliminate this, in the same way that the Baxandall volume control eliminates any dependency on the pot track. We could do this if we could get rid of the law-bending resistor R5, and also lose the end-stop resistor R6; the pot would then be working as a pure potentiometer with its division ratio controlled by the angular position of the wiper alone. If the gain range is restricted to around 10 dB, there is no very pressing need for law-bending and R5 can be simply omitted. However if we simply do away with the end-stop resistor R6 the gain is going to be uncontrollably large with the pot wiper at the bottom of its track. We need a way to limit the maximum gain that keeps other resistors away from the pot.

I was just going to bed at dawn (the exigencies of the audio service, ma'am) when it occurred to me that the answer is to add a resistor that gives a separate feedback path around the differential opamp U1:A only. R7 in Figure 18.14 provides negative feedback independently of the gain control and so limits the maximum gain. It is not intuitively obvious (to me, at any rate) but the CMRR is still preserved when the gain is altered, just as for Figure 18.13. You will note in Figure 18.14 that two resistors R4, R8 are paralleled to get a value exactly equal to R2 in parallel with R7. It shows resistor values that give a gain range suitable for combining an active stereo balance control with a balanced line input, in one economical stage. The minimum gain is −3.3 dB and the maximum gain is +6.6 dB. The gain with the pot wiper central is +0.2 dB which I would suggest is close enough to unity for anyone. This concept eliminates the effect of pot track resistance on gain, and I propose to call it the Self Input. Alright?

The gain equation looks more complex than for the previous example, but it is actually just a slight elaboration. The gain is now controlled by the ratio of R1 (=R3) to the parallel

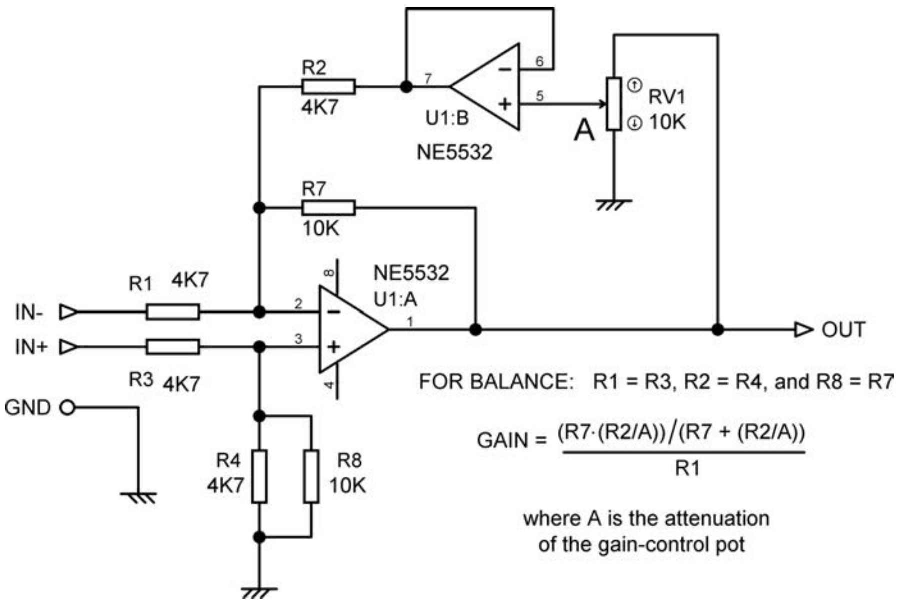

Figure 18.14: The Self Input: a variable-gain balanced input amplifier free from pot-dependence. Gain here is −3.3 to +6.6 dB

combination of R7 and R2, the latter being effectively scaled by the factor A introduced by the pot setting.

The Self Input stage in Figure 18.14 has reasonably high resistor values to allow direct connection to outside circuitry. As for the other configurations in this chapter, the noise performance can be much improved by scaling down all the resistor values and driving the inputs via a pair of unity-gain buffers; the low-value resistors reduce Johnson noise and the effect of opamp input current noise flowing through them; in this form it was used in my recent *Linear Audio* preamp design [7]. There is much more on reducing noise in this way later in this chapter.

High input-impedance balanced inputs

We saw earlier that high input impedances are required to maximise the CMRR of a balanced interconnection, but the input impedances offered by the standard balanced circuit are limited by the need to keep the resistor values down to control Johnson noise and the effects of current noise. High-impedance balanced inputs are also useful for interfacing to valve equipment in the strange world of retro-hifi. Adding output cathode-followers to valve circuitry is expensive and consumes a lot of extra power, and so the output is often taken directly from the anode of a gain-stage, and even a so-called bridging load of 10 kΩ may seriously compromise the distortion performance and output capability of the source equipment.

Figure 18.15 shows a configuration where the input impedances are determined only by the bias resistances R1 and R2. They are shown here as 100 kΩ, but may be considerably higher if opamp bias currents permit. A useful property of this circuit is that adding a single resistor

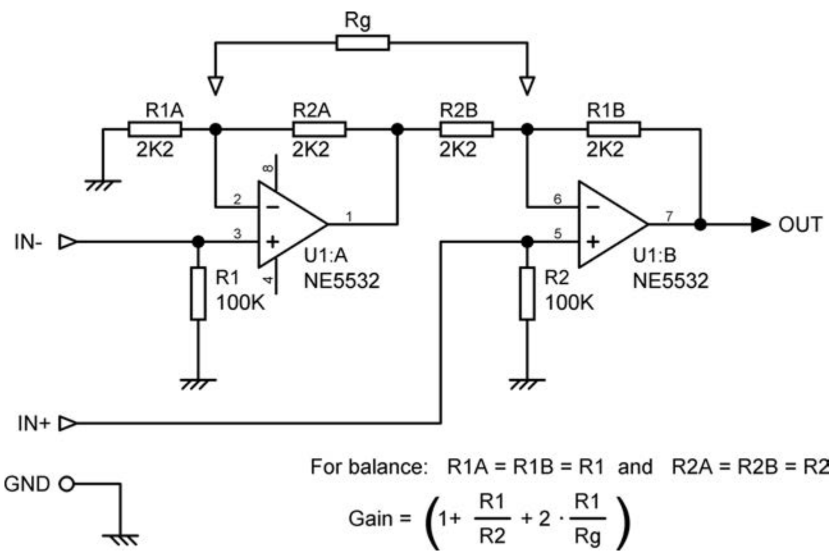

Figure 18.15: High input impedance balanced input

Rg increases the gain, but preserves the circuit balance and CMRR. This configuration cannot be set to attenuate because the gain of an opamp with series feedback cannot be reduced below unity.

It is of course always possible to give a basic balanced input a high input impedance by putting unity-gain buffers in front of each input, but that uses three opamp sections rather than two. Sometimes, however, it is appropriate. Much more on that later.

We noted earlier that the simple balanced input is surprisingly quiet and well-behaved when its input is unterminated. This is not the case with this configuration, which because of its high input impedances will be both noisy and susceptible to picking-up external interference.

The inverting two-opamp input

The configuration depicted in Figure 18.16 has its uses because the hot and cold inputs have the same impedances for differential signals, as well as for common-mode voltages [8]. It is not suited to high input impedances at normal gains because high resistor values would have to be used throughout and they would generate excess Johnson noise, but if it is interfacing with a high voltage source so the gain must be well below unity, R1 and R3 can be made high in value, and R2, R4, R5 set low, the latter components keeping the noise down. The CMRR may degrade at HF because the hot signal has gone through an extra opamp and suffered phase shift, interfering with the subtraction; this can be compensated for by the network

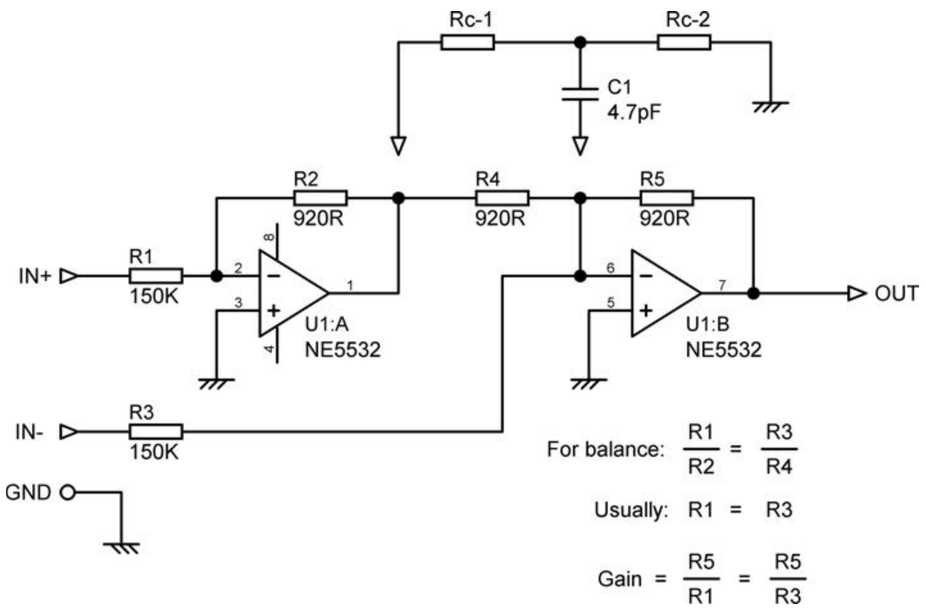

Figure 18.16: Inverting two-opamp balanced input

Rc-1, Rc-2 and C1; the values needed depend on opamp type and must be checked by CMRR measurements.

I have used this configuration for balanced input feeds from a 100 V loudspeaker line of the sort still in use for distributing audio over wide areas. The circuit values shown reduce the 100 V input to a nominal −2 dBu (615 mV) internal level. If this circuit is used for that purpose then effective input overvoltage protection on the inputs is essential as large voltage transients are possible on a loudspeaker line if parts of it are unplugged or plugged with signal present, due to the inductance of the line matching transformers. This protection can be conveniently provided by the usual diode clamps to the supply rails; this is dealt with at the end of this chapter. The input resistors will be of high value in this application so there is very little possibility of excessive inputs overheating the resistors or 'pumping up' the opamp supply rails.

The instrumentation amplifier

Almost every book on balanced or differential inputs includes the three-opamp circuit of Figure 18.17 and praises it as the highest expression of the differential amplifier. It is usually called the instrumentation amplifier configuration because of its undoubted superiority for data-acquisition (specialised ICs exist that are sometimes also called instrumentation amplifiers or in-amps; these are designed for very high CMRR data-acquisition. They are expensive and in general not optimised for audio work.)

The dual input stage U1:A, U1:B buffers the balanced line from the input impedances of the final differential amplifier U2:A, so the four resistances around it can be made much lower

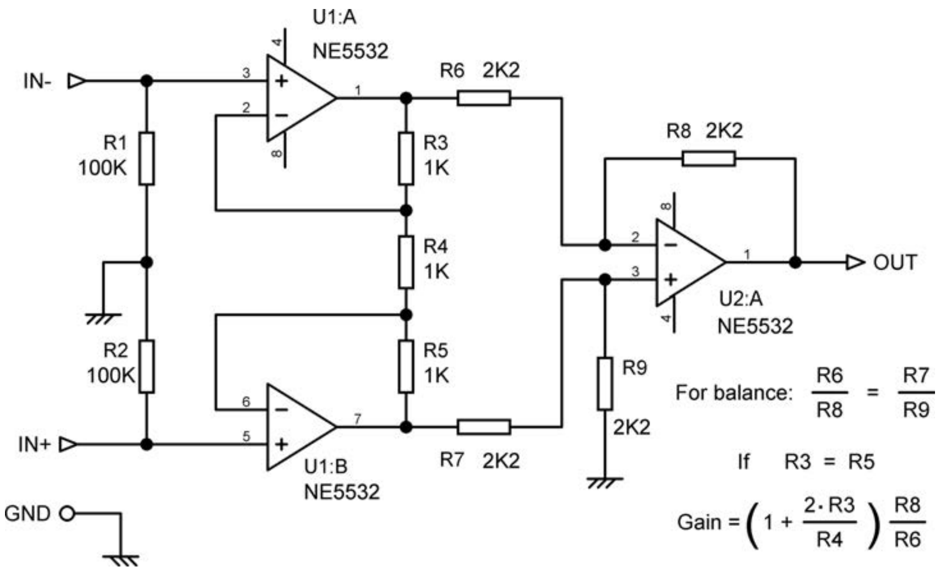

Figure 18.17: The instrumentation amplifier configuration. Gain here is 3 times

in value, reducing their Johnson noise and the effects of opamp current noise by a significant amount, while keeping the CMRR benefits of presenting high input impedances to the balanced line. However, the true beauty of the instrumentation amplifier, much emphasised because of its unquestionable elegance, is that the dual input stage, with its shared feedback network R3, R4, R5, can be set to have a high differential gain by giving R4 a low value, but its common-mode gain is always unity; this property is *not* affected by mismatches in R3 and R5. The final amplifier then does its usual job of common-mode rejection, and the combined CMRR can be very good indeed (over 100 dB) if the first-stage gain is high.

This is all well and good, but it is not immediately obvious that the instrumentation amplifier configuration can usefully improve the CMRR of audio balanced line inputs. A data-acquisition application like ECG monitoring may need a gain of thousands of times; a good portion of this will be put into the first stage, allow a stunning CMRR to be achieved without using precision resistors. Unfortunately, the cruel fact is that in audio usage, this much gain at a line input at this point is simply not wanted, except for microphone amplifiers. In a typical signal path comprised of opamps, the nominal internal level is usually between −6 and 0 dBu, and the line level coming in is at the professional level of +4 dBu; what is needed is 6 dB of attenuation rather than any gain. Gain at this point and attenuation later would introduce what can only be described as a headroom bottleneck. If the incoming level was the semi-pro −7.8 dBu, a small amount of gain could be introduced.

So does this mean that the instrumentation amplifier configuration has no real advantages over an ordinary balanced amplifier with input buffers? No. Stand by for what I believe is a whole new way of looking at this stage.

Instrumentation amplifier applications

Firstly, there are some applications where gain in a balanced input amplifier is either essential or can be made use of. Active crossovers and power amplifiers are probably the two most common.

Active crossovers are commonly used with no volume control or only a limited-range gain trim between them and the power amplifiers, so that it can always be assumed that the power amplifier will clip first. As I pointed out in [9], under these conditions the crossover can be run at an internal level of 3 Vrms or more, with the output level dropped back to say 1 Vrms at the very output of the crossover by a low-impedance passive attenuator. This can give a very impressive noise performance because the signal goes through the chain of filters at a high level [10].

For a typical active crossover application like this, we therefore need an input stage with gain. If the nominal input is 0 dBu (775 mVrms) we want a gain of about four times to get the signal up to 3 Vrms (+11.8 dBu). We will not need a greater output than this because it is sufficient to clip the following power amplifiers, even after the passive attenuators. This demonstrates that you can break the rule 'Do not amplify then attenuate' so long as there is a more severe headroom restriction downstream, and a limited gain control range.

If we use the strategy of input buffers followed by balanced amplifiers working at low impedances, as described later in this chapter (i.e. *not* an instrumentation amplifier as there is no network between the two input opamps), the required gain can be obtained by altering the ratio of the resistors in the balanced amplifier of a 1–1 configuration. With suitable resistor values (R1 = R3 = 560R, R2 = R4 = 2k2) the output noise from this stage (inputs terminated by 50 Ω) is −100.9 dBu, so its equivalent input noise (EIN) is −100.9 − 11.8 = −112.7 dBu, which is reasonably quiet. All noise measurements are 22 Hz–22 kHz, rms sensing, unweighted.

The other area for instrumentation amplifier techniques is that of power amplifiers with balanced inputs. The voltage gain of the power amplifier itself is frequently about 22 times (+26.8 dB) for its own technical reasons. If we take a power output of 100 W into 8 Ω as an example, the maximum output voltage is 28.3 Vrms, so the input level must be 28.3/22 = 1.29 Vrms. If the nominal input level is +4 dBu (1.23 Vrms) then a very modest gain of 1.05 times is required, and we will gain little from using an instrumentation amplifier with that gain in Stage 1. If however the nominal input level is 0 dBu (0.775 Vrms), a higher gain of 1.66 times is needed but that is still only a theoretical CMRR improvement of 4.4 dB. However, we can again make use of the fact that a relatively small signal voltage is enough to clip the power amplifier, and we can amplify the input signal – perhaps by four times again – and then attenuate it, along with the noise from the balanced input amplifier.

The instrumentation amplifier with 4× gain

Since we need a significant amount of gain, the instrumentation amplifier configuration is a viable alternative to the input buffer- balanced amplifier approach, giving a CMRR theoretically improved by 11.8 dB, the gain of its first stage. It uses the same number of opamps – three. A suitable circuit is shown in Figure 18.18. This gives a gain of 3.94 times, which is as close as you can get to 4.00 with E24 values. My tests show that the theoretical CMRR improvement really is obtained. To take just one set of results, when I built the second stage it gave a CMRR of −56 dB working alone, but when the first stage was added the CMRR leapt up to −69 dB, an improvement of 13 dB. These one-off figures (13 dB is actually better than the 12 dB improvement you would get if all resistors were exact) obviously depend on the specific resistor examples I used, but you get the general idea. The CMRR was flat across the audio band.

I do realise that there is an unsettling flavour of something-for-nothing about this, but it really does work. The total gain required can be distributed between the two stages in any way, but it should all be concentrated in Stage 1, as shown in Figure 18.18, to obtain the maximum CMRR benefit.

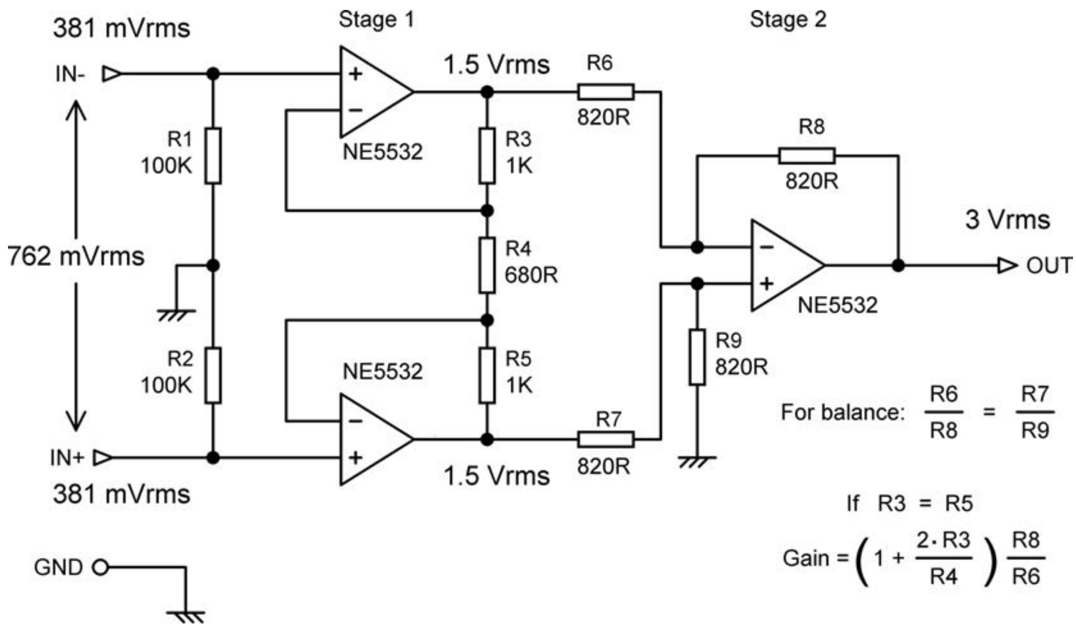

Figure 18.18: Instrumentation-amplifier balanced input stage with gain of 4 times

What about the noise performance? Fascinatingly, it is considerably improved. The noise output of this configuration is usefully lower at −105.4 dBu, which is no less than 4.5 dB better than the –100.9 dBu we got from the more conventional 1–1 configuration described above. This is because the amplifiers in Stage 1 are working in better conditions for low noise than those in Stage 2. They have no significant resistance in series with the non-inverting inputs to generate Johnson noise or turn the opamp current noise into voltage noise. As they implement all the gain before the signal reaches Stage 2, the noise contribution of the latter is therefore less significant.

When noise at −105.4 dBu is put through a 10 dB passive attenuator to get back down to the output level we want, as might typically be done at the output of an active crossover, it will become −115.4 dBu, which I suggest is very quiet indeed.

The circuit in Figure 18.18 requires some ancillary components before it is ready to face the world. Figure 18.19 shows it with EMC RF filtering (R19, C1 and R20, C4), non-polarised DC-blocking capacitors (C2, C3), and capacitors C5, C6 to ensure HF stability. R21, R22 are DC-drain resistors that ensure that any charge left on the DC-blocking capacitors by the source equipment when the connector is unplugged is quickly drained away. These extra components are omitted from subsequent diagrams to aid clarity.

So, here we are with all of the gain in the first stage, for the best CMRR, and this is followed by a unity-gain Stage 2. Now let's extend this thinking a little. Suppose Stage 2 had a gain of less than unity? We could then put more gain in Stage 1, and get a further free and guaranteed

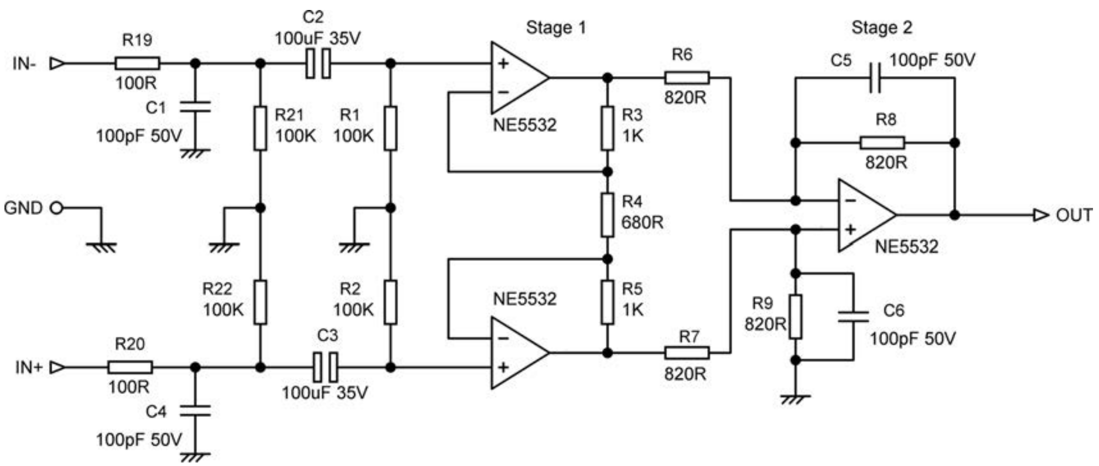

Figure 18.19: Practical instrumentation-amplifier balanced input stage with gain of 4 times

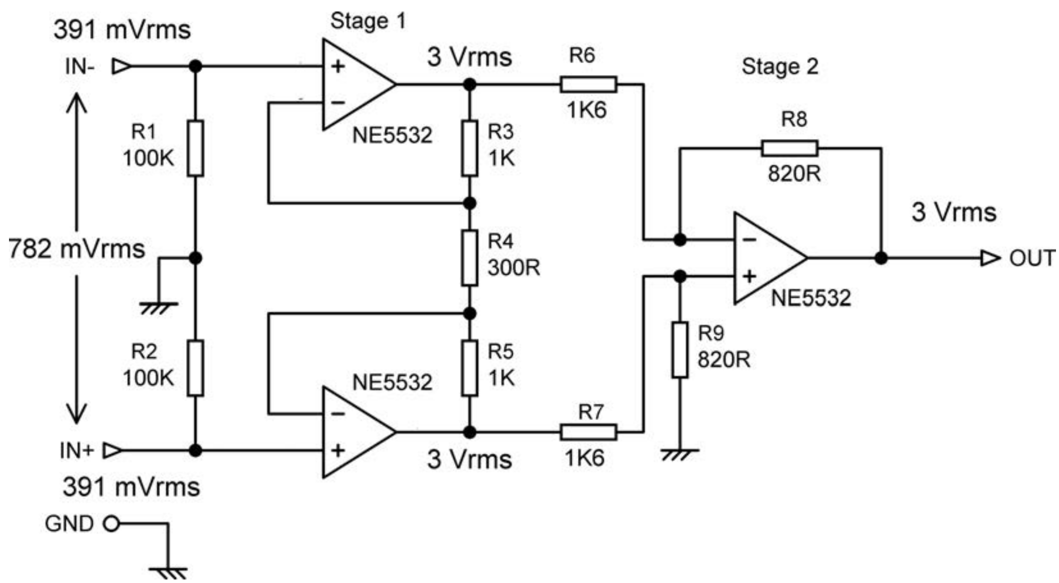

Figure 18.20: Instrumentation-amplifier balanced input stage with gain of 8 times in Stage 1 and 0.5 times in Stage 2

CMRR improvement, while keeping the overall gain the same. But can we do this without causing headroom problems in Stage 1?

The instrumentation amplifier in Figure 18.20 has had the gain of Stage 1 doubled to eight times (actually 7.67 times; 0.36 dB low, but as close as you can get with E24 values) by reducing R4, while the input resistors R6, R7 to Stage 2 have been doubled in value so that stage now has a gain of 0.5 times, to keep the total gain at approx. four times.

In this case the theoretical CMRR improvement is once more the gain of Stage 1, which is 7.67 times, and therefore 17.7 dB. Stage 2 gave a measured CMRR of −53 dB working alone, but when Stage 1 was added the CMRR increased to −70 dB, an improvement of 17 dB, significantly greater than the 14 dB improvement obtained when Stage 1 had a gain of four times. The noise output was −106.2 dBu, which is better than the four-times instrumentation amplifier by 0.8 dB, and better than the somewhat more conventional 1–1 configuration by a convincing 5.3 dB. When the noise level of −106.2 dBu is put through the 10 dB passive attenuator to get back down to the level we want, this will become −116.2 dBu, which is even quieter than the previous version.

You may well be getting nervous about having a gain as high as eight times followed by attenuation; have we compromised the headroom? If we assume that, as before, the maximum output we need is 3 Vrms, then with a total gain of four times we will have 0.775 Vrms between the two inputs, and the corresponding signal voltages at each output of Stage 1 are also 3 Vrms. These outputs are in anti-phase, and so when they are combined in Stage 2, with its gain of 0.5 times, we get 3 Vrms at the final output. This means that in balanced operation all three amplifiers will clip simultaneously, and the input amplifier can pass the maximum opamp output level before clipping. With the usual ±17 V supply rails this is about 10 Vrms.

So far so good; all the voltages are well below clipping. But note that we have assumed a balanced input, and unless you are designing an entire system, that is a dangerous assumption. If the input is driven with an unbalanced output of the same level, we will have 0.775 Vrms on one input pin, and nothing on the other. You might fear we would get 6 Vrms on one output of Stage 1 and nothing on the other; that would still be some way below clipping but the safety margin is shrinking fast. In fact, things are rather better than that. The cross-coupling of the feedback network R3, R4, R5 greatly reduces the difference between the two outputs of stage 1 with unbalanced inputs. With 0.775 Vrms on one input pin, and nothing on the other, the Stage 1 outputs are 3.39 Vrms and 2.61 Vrms, which naturally sum to 6 Vrms, reduced to 3 Vrms out by Stage 2. The higher voltage is at the output of the amplifier associated with the driven input. Maximum output with an unbalanced input is slightly reduced from 10 Vrms to 9.4 Vrms. This minor restriction does not apply to the earlier version with only four times gain in Stage 1.

This shows us that if we can assume a balanced input and stick with a requirement for a maximum output of 3 Vrms, we have a bit more elbow-room than you might think for using higher gains in Stage 1 of an instrumentation amplifier. Taking into account the possibility of an unbalanced input, we could further increase the gain by a factor of 10/3.39 = 2.95 times, on top of the doubling of Stage 1 gain we have already implemented. This gives a Stage 1 gain of 22.6 times. I decided to be conservative and go for a Stage 1 gain of 21 times; the gain of Stage 2 is correspondingly reduced to 0.19 times to maintain a total gain of four times. This version is shown in Figure 18.21; you will note that the feedback resistors R3, R4, R5 in Stage 1 have been scaled up by a factor of two. This is because Stage 1 is now working at

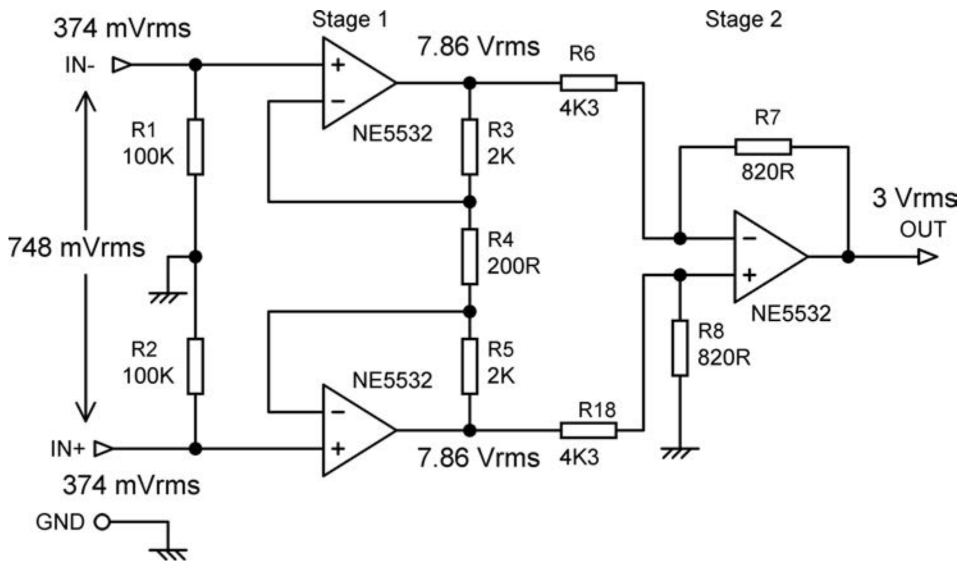

Figure 18.21: Instrumentation-amplifier balanced input stage with gain of 21 times in Stage 1 and 0.19 times in Stage 2. Maximum output 3 Vrms

a higher level; remember also that the resistors R3 and R5 are not the only load on Stage 1, because each of its opamp outputs is looking at the Stage 2 inputs driven in anti-phase, so its inverting input loading is heavier than it at first appears. Leaving R3, R5 at 1 kΩ would significantly degrade the distortion performance; I found it doubled the THD at 10 kHz from 0.0012% to 0.0019% for a 3 Vrms output.

When Stage 2 was tested alone its CMRR was −61 dB; clearly we have been lucky with our resistors. Adding on Stage 1 increased it to a rather spectacular −86 dB, an improvement of no less than 25 dB. This degrades to −83 dB at 20 kHz; this is the first time it has not appeared flat across the audio band. The theoretical CMRR improvement is 26.4 dB, without trim presets or selected components.

The output noise level remains at −106.2 dBu; we have clearly gained all the noise advantage we can by increasing the Stage 1 gain.

Note that throughout this discussion I have assumed that there are no objections to running an opamp at a signal level not far from clipping, and in particular that the distortion performance is not significantly harmed by doing that. This is true for many modern opamps such as the LM4562, and also for the veteran 5532, when it is not loaded to the limit of its capability.

The instrumentation amplifier at unity gain

If you want a unity-gain input stage but also some CMRR improvement, some compromise is required.

Putting R3, R5 = 1 kΩ and R4 = 2 kΩ gives a Stage 1 gain of two times, which is undone by setting R6, R7 to 1k6 in Stage 2 to give a gain of 0.5 times. This will pass the full 10 Vrms maximum opamp level, but only in balanced mode. In unbalanced mode, Stage 1 will clip at the opamp connected to the driven input because when unbalanced the gain to this point from the input is 1.5 times. The maximum signal that can be handled in unbalanced mode is therefore about 7 Vrms. The resulting input amplifier is shown in Figure 18.22. A heavy load is placed on the Stage 1 amplifiers in this situation and for the best distortion behaviour it would be better to make R3, R5 = 2 kΩ and R4 = 4 kΩ.

When I did the measurements, Stage 2 alone had a CMRR of −57 dB; adding on Stage 1 brought this up to −63 dB (this exact correspondence of theoretical and measured improvement is just a coincidence). You can therefore have 6 dB more CMRR free, gratis, and for absolutely nothing, so long as you are prepared to tolerate a loss of 3.5 dB in headroom if the stage is run unbalanced but with the same input voltage. The noise output is −111.6 dBu.

A typical hifi output gives 1 Vrms in unbalanced use, and 2 Vrms when run balanced, as the cold output is provided by a simple inverter, the presence of which doubles the total output voltage. Figure 18.22 shows the signal voltages when our unity-gain input stage is connected to a 2 Vrms balanced input. This sort of situation is more likely to occur in hifi than in professional audio; in the latter simulated 'floating transformer winding' outputs are sometimes found, where the total output voltage is the same whether the output is run balanced or unbalanced (these are described in Chapter 19).

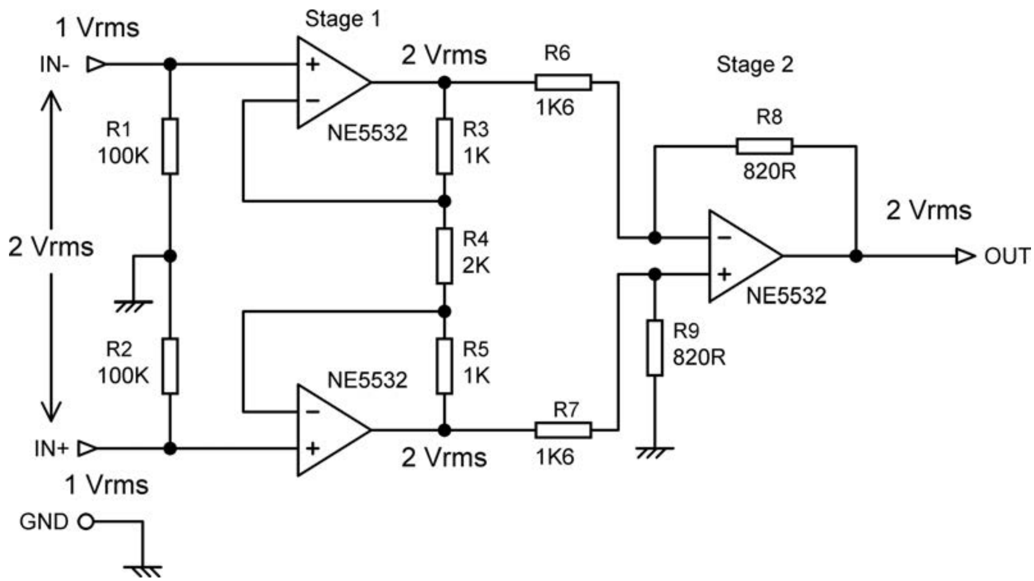

Figure 18.22: Instrumentation-amplifier balanced input stage with gain of 2 times in Stage 1 and 0.5 times in Stage 2 for overall unity gain

In hifi use it may be possible to assume that the instrumentation amplifier will never be connected to a balanced output with a total signal voltage of more than 2 Vrms. If this is the case we can once more put extra gain into Stage 1, and undo it with Stage 2, without compromising headroom. We are now dealing with a restriction on input level rather than output level. We will assume that the 2 Vrms input is required to be reduced to 1 Vrms to give adequate headroom in the circuitry that follows, so an overall gain of 0.5 times is required.

Finally we will look at how the range of gain control between the instrumentation amplifier and the power amplifier affects the headroom situation. Figure 18.23 shows one of the instrumentation amplifiers described above with a maximum output of 3 Vrms, the version in Figure 18.20 with a gain of eight times in Stage 1 and 0.5 times in Stage 2. This is followed by a ±10 dB gain trim network that is normally expected to work at the 0 dB setting, i.e. giving 10 dB of attenuation and so reducing the noise from the input stage. I should point out that this simple network does not give '0 dB' at the centre of the pot travel; that point is more like three-quarters of the way down.

Since the maximum output of this stage is 10 Vrms with a balanced input (slightly less with an unbalanced input) clipping will always occur in the power amplifier first so long as the gain trim is set above −10 dB. At the −10 dB setting (which introduces 20 dB of actual attenuation) the input amplifier and the power amp will clip simultaneously; if more attenuation is permitted then there is the danger that the input amplifier will clip first, and this cannot be corrected by the gain trim. Note the low resistor values in the gain trim network to minimise Johnson noise and also the effect of the power amplifier input noise current. A low source impedance is also required by most power amplifiers to prevent input current distortion [11]. In the worst-case position (approx. wiper central) the output impedance of the network is 275 Ω, and the Johnson noise from that resistance is −128.2 dBu, well below the noise from the input amplifier, which is −106.2 dBu − 6 dB = −112.2 dBu, and also below the equivalent input noise of the power amplifier, which is unlikely to be better than −122 dBu.

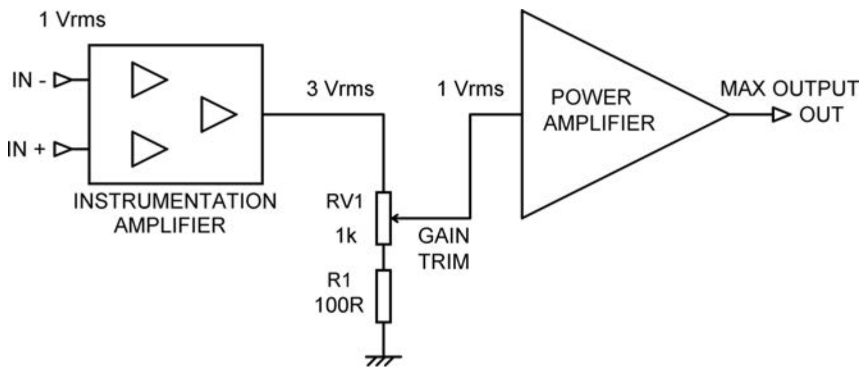

Figure 18.23: Maximum gain trim range possible avoiding headroom restriction in the instrumentation amplifier

Transformer balanced inputs

When it is essential that there is no galvanic connection (i.e. no electrical conductor) between two pieces of equipment, transformer inputs are indispensable. They are also useful if EMC conditions are severe. Figure 18.24 shows a typical transformer input. The transformer usually has a 1:1 ratio, and is enclosed in a metal shielding can which must be grounded. Good line transformers have an inter-winding shield that must also be grounded or the high-frequency CMRR will be severely compromised. The transformer secondary must see a high impedance as this is reflected to the primary and represents the input impedance; here it is set by R2, and a buffer drives the circuitry downstream. In addition, if the secondary loading is too heavy there will be increased transformer distortion at low frequencies. Line input transformers are built with small cores and are only intended to deliver very small amounts of power; they are *not* interchangeable with line output transformers. A most ingenious approach to dealing with this distortion problem by operating the input transformer core at near-zero flux was published by Paul Zwicky in 1986 [12].

There is a bit more to correctly loading the secondary transformer. If it is simply loaded with a high-value resistor there will be peaking of the frequency response due to resonance between the transformer leakage inductance and the winding capacitance. This is shown in Figure 18.25, where a Sowter 3276 line input transformer (a high-quality component) was given a basic resistive loading of 100 kΩ. The result was Trace A, which has a 10 dB peak around 60 kHz. This is bad not only because it accentuates the effect of out-of-band noise, but because it intrudes on the audio frequency response, giving a lift of 1 dB at 20 kHz. Reducing the resistive load R2 would damp the resonance, but it would also reduce the input impedance. The answer is to add a Zobel network, i.e. a resistor and capacitor in series, across the secondary; this has no effect except on high frequencies. The first attempt used R1 = 2k7 and C1 = 1 nF, giving Trace B, where the peaking has been reduced to 4 dB around 40 kHz, but the 20 kHz lift is actually slightly greater. R1 = 2k7 and C1 = 2 nF gave Trace C, which is a

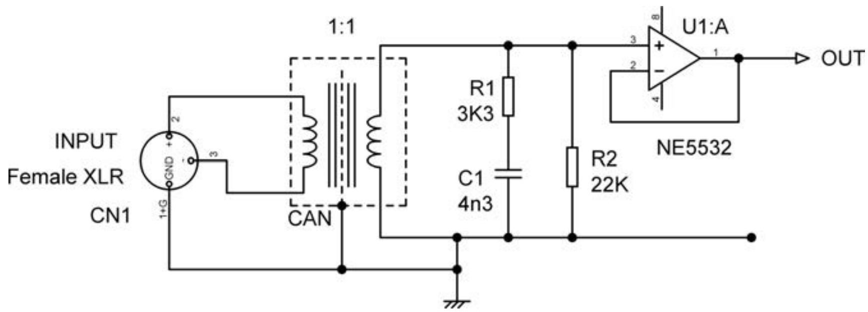

Figure 18.24: A transformer balanced input. R1 and C1 are the Zobel network that damps the secondary resonance

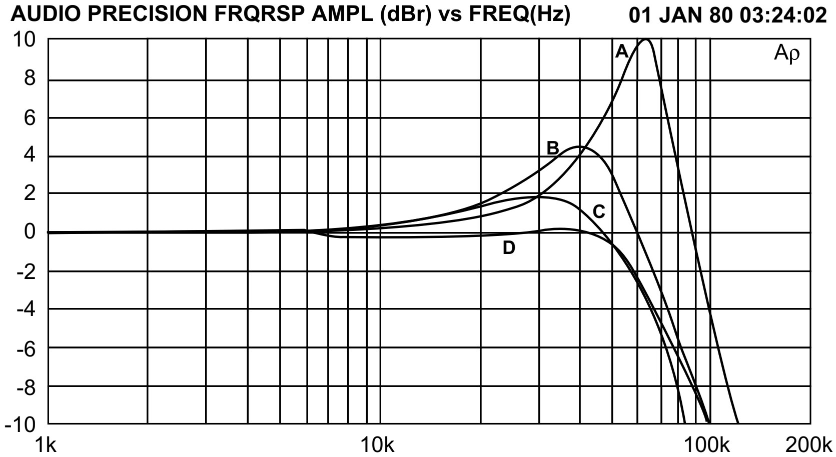

Figure 18.25: Optimising the frequency response of a transformer balanced input with a Zobel network

bit better in that it only has a 2 dB peak. Some more experimentation resulted in R1 = 3k3 and C1 = 4.3 nF (3n3 + 1nF) and yielded Trace D, which is pretty flat, though there is a small droop around 10 kHz. The Zobel values are fairly critical for the flattest possible response and must certainly be adjusted if the transformer type is changed.

Input overvoltage protection

Input overvoltage protection is not common in hifi applications, but is regarded as essential in most professional equipment. The normal method is to use clamping diodes, as shown in Figure 18.26, that prevent certain points in the input circuitry from moving outside the supply rails.

This is straightforward, but there are two points to watch. Firstly, the ability of this circuit to withstand excessive input levels is not without limit. Sustained overvoltages may burn out R5 and R6, or pump unwanted amounts of current into the supply rails; this sort of protection is mainly aimed at transients. Secondly, diodes have a non-linear junction capacitance when they are reverse biased, so if the source impedance is significant the diodes will cause distortion at high frequencies. To quantify this problem here are the results of a few tests. If Figure 18.26 is fed from the low impedance of the usual kind of line output stage, the impedance at the diodes will be about 1 kΩ and the distortion induced into an 11 Vrms 20 kHz input will be below the noise floor. However, if the source impedance is high so the impedance at the diodes is increased to 10 kΩ, with the same input level, the THD at 20 kHz will be degraded from 0.0030% to 0.0044% by adding the diodes. I have thought up a rather elegant way to eliminate this effect completely; anybody interested?

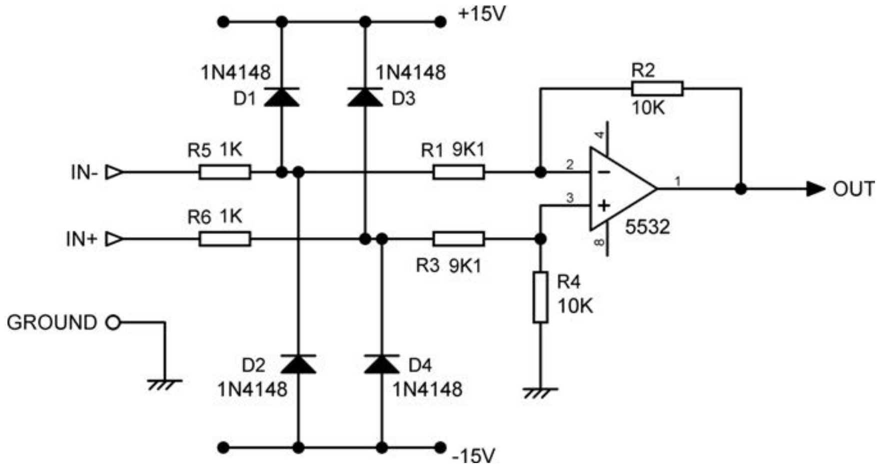

Figure 18.26: Input overvoltage protection for a balanced input amplifier

Low-noise balanced inputs

So far we have not said much about the noise performance of balanced inputs, though we have noted several times that the standard balanced input amplifier with four 10 kΩ resistors as in Figure 18.27a is markedly noisier than an unbalanced input like that in Figure 18.1. The unbalanced input stage, with its input terminated by 50 Ω to ground, has a noise output of −119.0 dBu over the usual 22 Hz–22 kHz bandwidth. If the balanced circuit is built with 10 kΩ resistors and a 5532 section, the noise output is −104.8 dBu with the inputs similarly terminated. This is a big difference of 14.2 dB.

In the hifi world in particular, where an amplifier may have both unbalanced and balanced inputs, many people would feel that this situation is the wrong way round. Surely the balanced stage input, with its professional XLR connector and its much-vaunted rejection of ground noise, should show a better performance in all departments? Well, it does – except as regards internal noise, and a 14 dB discrepancy is both clearly audible and hard to explain away. This section explains how to design it away instead.

We know that the source of the extra noise is the relatively high resistor values around the opamp (see Table 18.8 earlier in the chapter), but these cannot be reduced in the simple balanced stage without reducing the input impedances below what is acceptable. The answer is to lower the resistor values but buffer them from the input with a pair of voltage-followers; this arrangement is shown in Figure 18.27b. 5532s are a good choice for this as they combine low voltage noise with low distortion and good load-driving capability. Since the input buffers are voltage-followers with 100% feedback, their gain is very accurately defined at unity and CMRR is not degraded; CMRR is still defined by the resistor tolerances and by the bandwidth of the differential opamp. I call this a 1–1 configuration, for reasons which will be made clear.

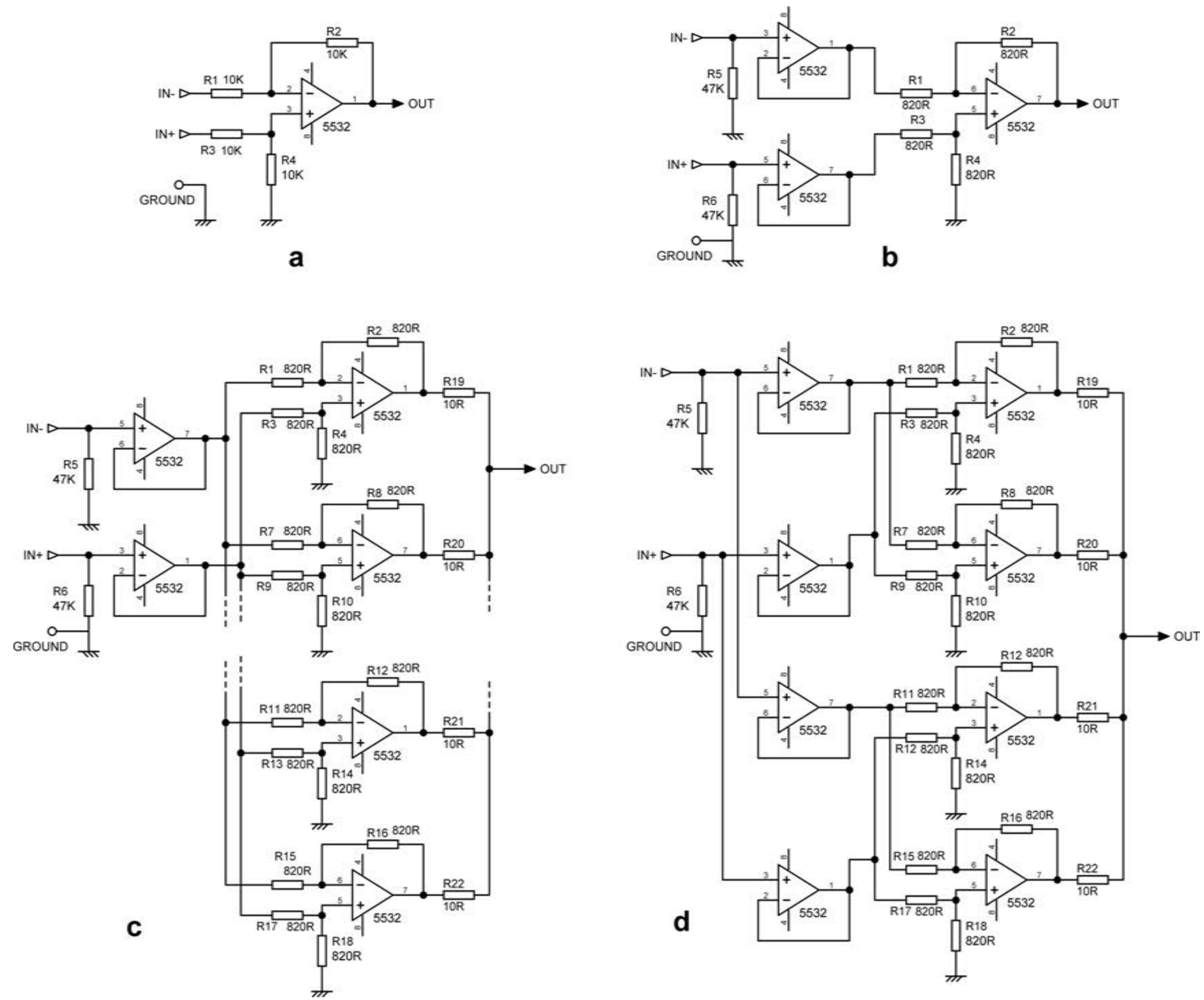

Figure 18.27: Low noise unity-gain balanced inputs using multiple 5532 buffers and differential amplifiers

There is a limit to how far the four resistors can be reduced, as the differential stage has to be driven by the input buffers and it also has to drive its own feedback arm. If 5532s are used a safe value that gives no measurable deterioration of the distortion performance is about 820 Ω, and an 5532 differential stage alone (without the buffers) and 4 × 820 Ω resistors gives a noise output of −111.7 dBu, which is 6.6 dB lower than the standard 4 × 10 kΩ version. Adding the two input buffers degrades this only slightly to −110.2 dB, because we are adding only the voltage noise component of the two new opamps, and we are still 5.1 dB quieter than the original 4 × 10 kΩ version. It is an interesting point that we now have three opamps in the signal path instead of one, but we still get a significantly lower noise level.

This might appear to be all we can do; it is not possible to reduce the value of the four resistors around the differential amplifier any further without compromising linearity. However, there is almost always some way to go further in the great game that is electronics, and here are three possibilities. A step-up transformer could be used to exploit the low source impedance (remember we are still assuming the source impedances are 50 Ω) and it might well work superbly in terms of noise alone, but transformers are always heavy, expensive, susceptible to magnetic fields and of doubtful low-frequency linearity. We would also very quickly run into headroom problems; balanced line input amplifiers are normally required to attenuate rather than amplify.

We could design a hybrid stage with discrete input transistors, which are quieter than those integrated into IC opamps, coupled to an opamp to provide raw loop gain; this can be quite effective but you need to be very careful about high-frequency stability, and it is difficult to get an improvement of more than 6 dB.

Thirdly, we could design our own opamp using all discrete parts; this approach tends to have less stability problems as all circuit parameters are accessible, but it definitely requires rather specialised skills, and the result takes up a lot of PCB area.

Since none of those three approaches are appealing, now what? One of the most useful techniques in low-noise electronics is to use two identical amplifiers so that the gains add arithmetically, but the noise from the two separate amplifiers, being uncorrelated, partially cancels. Thus we get a 3 dB noise advantage each time the number of amplifiers used is doubled. This technique works very well with multiple opamps, as expounded in Chapter 1; let us apply it and see how far it may be usefully taken.

Since the noise of a single 5532-section unity-gain buffer is only −119.0 dBu, and the noise from the 4 × 820 Ω differential stage (without buffers) is a much higher −111.7 dBu, the differential stage is clearly the place to start work. We will begin by using two identical 4 × 820 Ω differential amplifiers as shown in the top section of Figure 18.27c, both driven from the existing pair of input buffers. I call this a 1–2 configuration. This will give partial cancellation of both resistor and opamp noise from the two differential stages if their outputs are summed. The main question is how to sum the two amplifier outputs; any active solution would introduce another opamp, and hence more noise, and we would almost certainly wind up worse off than when we started. The answer is, however, beautifully simple. We just connect the two amplifier outputs together with 10 Ω resistors; the gain does not change but the noise output drops. We are averaging rather than summing. The signal output of both amplifiers is nominally the same, so no current should flow from one opamp output to the other. In practice there will be slight gain differences due to resistor tolerances, but with 1% resistors I have never experienced any hint of a problem. The combining resistor values are so low at 10 Ω that their Johnson noise contribution is negligible.

The use of multiple differential amplifiers has another advantage – the CMRR errors are also reduced in the same way that this noise is reduced. This is similar to the use of multiple parallel capacitors to improve RIAA accuracy, as explained in Chapter 8.

We therefore have the arrangement of Figure 18.27c, with single input buffers (i.e. one per input) and two differential amplifiers. This reduces the noise output by 2.3 dB to −112.5 dBu, which is quieter than the original 4 × 10 kΩ version by an encouraging 7.4 dB. We do not get a full 3 dB noise improvement because both differential amplifiers are handling the noise from the input buffers, which is correlated and so is not reduced by partial cancellation. The contribution of the input buffer noise is further brought out if we take the next step of using four differential amplifiers (1–4 configuration). There is of course nothing special about using amplifiers in powers of two. It is perfectly possible to use three or five differential amplifiers in the array, which will give intermediate amounts of noise reduction. If you have a spare opamp section, then put it to work!

So, leaving the input buffers unchanged, we use them to drive an array of four differential amplifiers. These are added on at the dotted lines in the lower half of Figure 18.27c. We get a further improvement, but only by 1.5 dB this time. The output noise is down to −114.0 dBu, quieter than the original 4 × 10 kΩ version by 8.9 dB. You can see that at this point we are proceeding by decreasing steps, as the input buffer noise is starting to dominate, and there seems little point in doubling up the differential amplifiers again; the amount of hardware could be regarded as a bit excessive, and so would the PCB area occupied. The increased loading on the input buffers is also a bit of a worry.

A more fruitful approach is to tackle the noise from the input buffers, by doubling them up as in Figure 18.27d, so that each buffer drives only two of the four differential amplifiers. This means that the buffer noise will also undergo partial cancellation, and will be reduced by 3 dB (2–4 configuration). There is however still the contribution from the differential amplifier noise, and so the actual improvement on the previous version is 2.2 dB, bringing the output noise down to −116.2 dBu, which is quieter than the original 4 × 10 kΩ version by a thumping 11.1 dB. Remember that there are two inputs, and 'double buffers' means two buffers per hot and cold input, giving a total of four in the complete circuit.

Since doubling up the input buffers gave us a useful improvement, it's worth trying again, so we have a structure with quad buffers and four differential amplifiers, as shown in Figure 18.28, where each differential amplifier now has its very own buffer (4–4 configuration). This improves

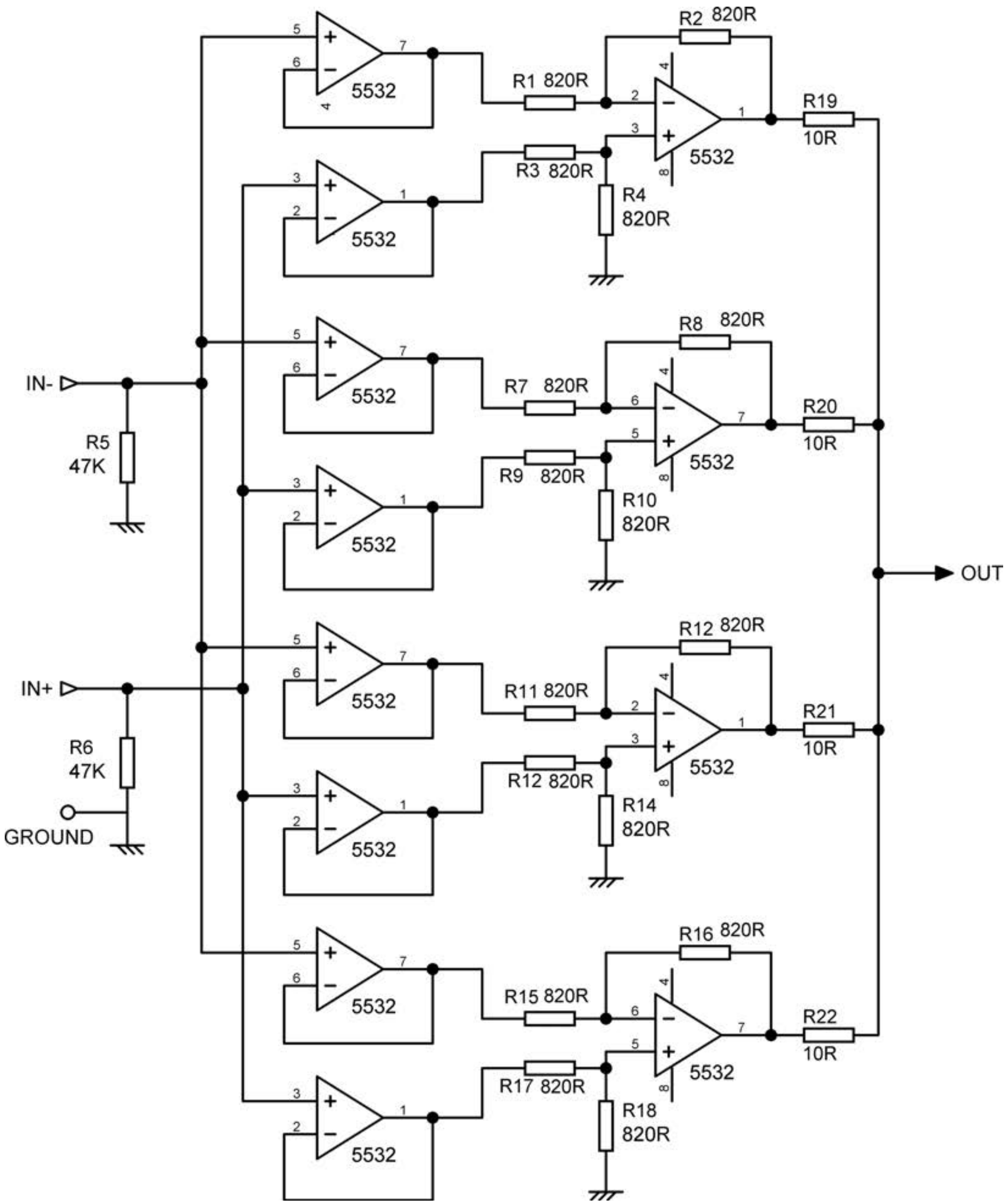

Figure 18.28: The final 5532 low noise unity-gain balanced input stage, with quad input buffers and four differential amplifiers (4–4 configuration). The noise output is only −117.0 dBu

on the previous version by a rather less satisfying 0.8 dB, giving an output noise level of −117.0 dBu, quieter than the original 4 × 10 kΩ version by 11.9 dB. The small improvement we have gained indicates that the focus of noise reduction needs to be returned to the differential amplifier array, but the next step there would seem to be using eight amplifiers, which is not very appealing. Thoughts about ears of corn on chessboards tend to intrude at this point.

This is a good moment to pause and see what we have achieved. We have built a balanced input stage that is quieter than the standard circuit by 11.9 dB, using standard components of low cost. We have used increasing numbers of them, but the total cost is still small compared with enclosures, power supplies, etc. On the other hand, the power consumption is clearly several times greater. The technology is highly predictable and the noise reduction reliable; in fact bullet-proof. The linearity is as good as that of a single opamp of the same type, and in the same way there are no HF stability problems.

What we have not done is build a balanced input that is quieter than the unbalanced one – we are still 2.0 dB short of that target, but at least we have reached a point where the balanced input is not obviously noisier.

These noise results are summarised in Table 18.10 at the end of the chapter.

Low-noise balanced inputs in action

Please don't think that this examination of low-noise input options is merely a voyage off into pure theory. It has real uses and has been applied in practice, for example, in the Cambridge Audio 840 W power amplifier, a design of mine which, I might modestly mention in passing, won a CES Innovation Award in January 2008. This unit has both unbalanced and balanced inputs, and conventional technology would have meant that the balanced inputs would have been significantly the noisier of the two. Since the balanced input is the 'premium' input, many people would think there was something amiss with this state of affairs. We therefore decided the balanced input had to be quieter than the unbalanced input. Using 5532s in an architecture similar to those outlined above, this requirement proved straightforwardly attainable, and the final balanced input design was both economical and quieter than its unbalanced neighbour by a dependable 0.9 dB. Two other versions were evaluated that made the balanced input quieter than the unbalanced one by 2.8 dB, and by 4.7 dB, at somewhat greater cost and complexity. These were put away for possible future upgrades.

The Signal Transfer Company [13] manufactures a low-noise balanced input card based on these principles which has 47 kΩ input impedances, unity gain, and a noise output of only −115 dBu.

Ultra-low-noise balanced inputs

In the section on low-noise balanced inputs above, we reclined briefly on our laurels having achieved an economical balanced input stage with output noise at the extremely low level of −117.0 dBu. Regrettably this is still 2 dB noisier than a simple unbalanced input. It would be wrong to conclude from this that the resources of electronic design are exhausted. At the end of the noise-reduction sequence we were aware that the dominant noise source was currently the differential amplifier array, and we shrank from doubling up again to use eight amplifiers because of issues of cost and the PCB area occupied. We will take things another step by taking a much more relaxed view of cost (as one does at the 'high-end'), and see how that changes the game. We will however retain some concern about PCB area.

An alternative way to make the differential amplifier array quieter is simply to use opamps that are quieter. These will inevitably be more expensive – much more expensive – than the ubiquitous 5532. Because of the low resistor values around the opamps we need to focus on low voltage-noise rather than low current noise, and there are several that are significantly better than the 5532, as shown by the typical noise density figures in Table 18.9.

TABLE 18.9 Voltage and noise densities for low-noise balanced-input opamp candidates

Opamp	Voltage noise density (nV/$\sqrt{\text{Hz}}$)	Current noise density (pA/$\sqrt{\text{Hz}}$)
5532	5	0.7
5534A	3.5	0.4
LM4562	2.7	1.6
AD797	0.9	2
LT1028	0.85	1

Clearly moving to the 5534A will give a significant noise reduction, but since there is only a single opamp per package, and external compensation is needed, the board area used will be much greater. Instead we will try the LM4562, a bipolar opamp which has finally surpassed the 5532 in performance. The input voltage noise density is typically 2.7 nV/$\sqrt{\text{Hz}}$, substantially lower than the 5 nV/$\sqrt{\text{Hz}}$ of the 5532. For applications with low source impedances, this implies a handy noise advantage of 5 dB or more. The LM4562 is a dual opamp and will not take up more space. At the time of writing it is something like 10 times more expensive than the 5532.

Step 1: We replace all four opamps in the differential amplifiers with LM4562s. They are a drop-in replacement with no circuit adjustments required at all. We leave the quad 5532 input buffers in place. The noise output drops by an impressive 1.9 dB, giving an output noise level of −118.9 dBu, quieter than the original 4 × 10 kΩ version by 14.1 dB, and only 0.1 dB noisier than the unbalanced stage.

Step 2: Replace the quad 5532 buffers with quad LM4562 buffers. Noise falls by only 0.6 dB, the output being −119.5 dBu, but at last we have a balanced stage that is quieter than the unbalanced stage, by a small but solid 0.5 dB.

One of the pre-eminent low-noise-with-low-source-resistance opamps is the AD797 from Analog Devices, which has a remarkably low voltage noise at 0.9 nV/$\sqrt{\text{Hz}}$ (typical at 1 kHz) but it is a very expensive part, costing between 20 and 25 times more than a 5532 at the time of writing. The AD797 is a single opamp, while the 5532 is a dual, so the cost per opamp is actually 40 to 50 times greater, and more PCB area is required, but the potential improvement is so great we will overlook that . . .

Step 3: We replace all four opamps in the differential amplifiers with AD797s, putting the 5532s back into the input buffers in the hope that we might be able to save money somewhere. The noise output drops by a rather disappointing 0.4 dB, giving an output noise level of −119.9 dBu, quieter than the original 4 × 10 kΩ version by 15.1 dB.

Perhaps putting those 5532s back in the buffers was a mistake? Our fourth and final move in this game of electronic chess is to replace all the quad 5532 input buffers with dual (not quad) AD797 buffers. This requires another four AD797s (two per input) and is once more not a cheap strategy. We retain the four AD797s in the differential amplifiers. The noise drops by another 0.7 dB yielding an output noise level of −120.6 dBu, quieter than the original 4 × 10 kΩ version by 15.8 dB, and quieter than the unbalanced stage by a satisfying 1.6 dB. You can do pretty much anything in electronics with a bit of thought and a bit of money.

You are probably wondering what happened to the LT1028 lurking at the bottom of Table 18.9. It is true that its voltage noise density is slightly better than that of the AD797, but there is a subtle snag. As described in Chapter 8, the LT1028 has bias-current cancellation circuitry which injects correlated noise currents into the two inputs. These will cancel if the impedances seen by the two inputs are the same, but in moving-magnet amplifier use the impedances differ radically and the LT1028 is not useful in this application. The input conditions here are more benign, but the extra complication is unwelcome and I have never used the LT1028 in audio work. In addition, it is a single opamp with no dual version.

This is not of course the end of the road. The small noise improvement in the last step we made tells us that the differential amplifier array is still the dominant noise source, and further development would have to focus on this. A first step would be to see if the relatively high current-noise of the AD797s is significant with respect to the surrounding resistor values. If so, we need to see if the resistor values can be reduced without degrading linearity at full output. We should also check the Johnson noise contribution of all those 820 Ω resistors; they are generating −123.5 dBu each at room temperature, but of course the partial cancellation effect applies to them as well.

All these noise results are also summarised in Table 18.10 below.

TABLE 18.10 A summary of the noise improvements made to the balanced input stage

Buffer type	Amplifier	Noise output (dBu)	Improvement on previous version (dB)	Improvement over 4 × 10 kΩ diff amp (dB)	Noisier than unbal input by: (dB)
	5532 voltage-follower	−119.0			0.0 dB ref
None	Standard diff amp 10k 5532	−104.8	0	0.0 dB ref	14.2
None	Single diff amp 820R 5532	−111.7	6.9	6.9	7.3
Single 5532	Single diff amp 820R 5532	−110.2	5.4	5.4	8.8
Single 5532	Dual diff amp 820R 5532	−112.5	2.3	7.4	6.5
Single 5532	Quad diff amp 820R 5532	−114.0	1.5	9.2	5.0
Dual 5532	Quad diff amp 820R 5532	−116.2	2.2	11.4	2.8
Quad 5532	Quad diff amp 820R 5532	−117.0	0.8	12.2	2.0
Quad 5532	Quad diff amp 820R LM4562	−118.9	1.9	14.1	0.1
Quad LM4562	Quad diff amp 820R LM4562	−119.5	0.6	14.7	−0.5
Quad 5532	Quad diff amp 820R AD797	−119.9	0.4	15.1	−0.9
Dual AD797	Quad diff amp 820R AD797	−120.6	0.7	15.8	−1.6

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CHAPTER 19

Line outputs

Unbalanced outputs

There are only two electrical output terminals for an unbalanced output – signal and ground. However, the unbalanced output stage in Figure 19.1a is fitted with a three-pin XLR connector, to emphasise that it is always possible to connect the cold wire in a balanced cable to the ground at the output end and still get all the benefits of common-mode rejection if you have a balanced input. If a two-terminal output connector is fitted, the link between the cold wire and ground has to be made inside the connector, as shown in Figure 18.2 in the chapter on line inputs.

The output amplifier in Figure 19.1a is configured as a unity-gain buffer, though in some cases it will be connected as a series feedback amplifier to give gain. A non-polarised DC blocking capacitor C1 is included; 100 µF gives a −3 dB point of 2.6 Hz with one of those notional 600 Ω loads. The opamp is isolated from the line shunt-capacitance by a resistor R2, in the range 47–100 Ω, to ensure HF stability, and this unbalances the hot and cold line impedances. A drain resistor R1 ensures that no charge can be left on the output side of C1; it is placed *before* R2, so it causes no attenuation. In this case the loss would only be 0.03 dB, but such errors can build up to an irritating level in a large system and it costs nothing to avoid them.

If the cold line is simply grounded as in Figure 19.1a, then the presence of R2 degrades the CMRR of the interconnection to an uninspiring −43 dB even if the balanced input at the other end of the cable has infinite CMRR in itself and perfectly matched 10 kΩ input impedances. See Chapter 18.

To fix this problem, Figure 19.1b shows what is called an impedance-balanced output. There are now three physical terminals, hot, cold, and ground. The cold terminal is neither an input nor an output, but a resistive termination R3 with the same resistance as the hot terminal output impedance R2. If an unbalanced input is being driven, this cold terminal can be either shorted to ground locally or left open-circuit. The use of the word 'balanced' is perhaps unfortunate as when taken together with an XLR output connector it implies a true balanced output with anti-phase outputs, which is *not* what you are getting. The impedance-balanced approach is not particularly cost-effective, as it requires significant extra money to be spent

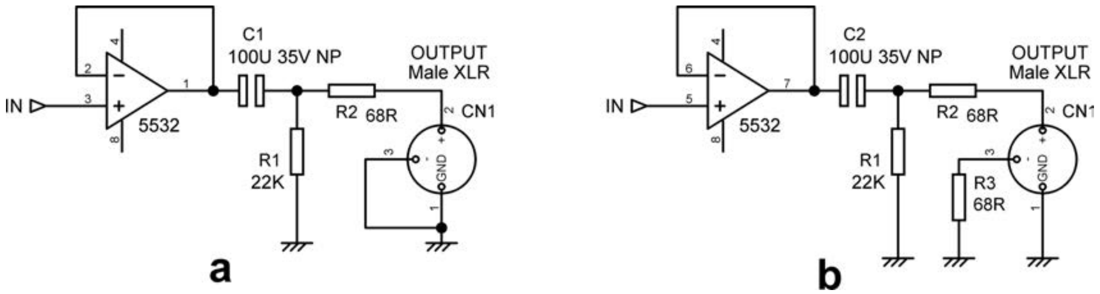

Figure 19.1: Unbalanced outputs: a) simple output, b) impedance-balanced output for improved CMRR when driving balanced inputs

on an XLR connector. Adding an opamp inverter to make it a proper balanced output costs little more, especially if there happens to be a spare opamp half available, and it sounds much better in the specification.

Zero-impedance outputs

Both the unbalanced outputs shown in Figure 19.1 have series output resistors to ensure stability when driving cable capacitance. This increases the output impedance, which can impair the CMRR of a balanced interconnection, and can also lead to increased crosstalk via stray capacitance in some situations. One such scenario that was fixed by the use of a so-called ‘zero-impedance’ output is described in Chapter 22 in the section on mixer insert points. The output impedance is of course not exactly zero, but it is very low compared with the average series output resistor.

Figure 19.2a shows how the zero-impedance technique is applied to an unbalanced output stage with 10 dB of gain. Feedback at audio frequencies is taken from outside isolating resistor R3 via R2, while the HF feedback is taken from inside R3 via C2 so it is not affected by load capacitance and stability is unimpaired. Using a 5532 opamp, the output impedance is reduced from 68 Ohms to 0.24 Ohms at 1 kHz – a dramatic reduction that would reduce capacitive crosstalk by 49 dB. Output impedance increases to 2.4 Ohms at 10 kHz and 4.8 Ohms at 20 kHz as opamp open-loop gain falls with frequency. The impedance-balancing resistor on the cold pin has been replaced by a link to match the near-zero output impedance at the hot pin. More details on zero-impedance outputs are given later in this chapter in the section on balanced outputs.

Figure 19.2b shows a refinement of this multiple-feedback scheme with three paths. In Chapter 2 we saw how electrolytic coupling capacitors can introduce distortion even if the time-constant is long enough to give a flat LF response. In Figure 19.2b most of the feedback is now taken from outside C1, via R5, so it can correct capacitor distortion. The DC feedback goes via R2,

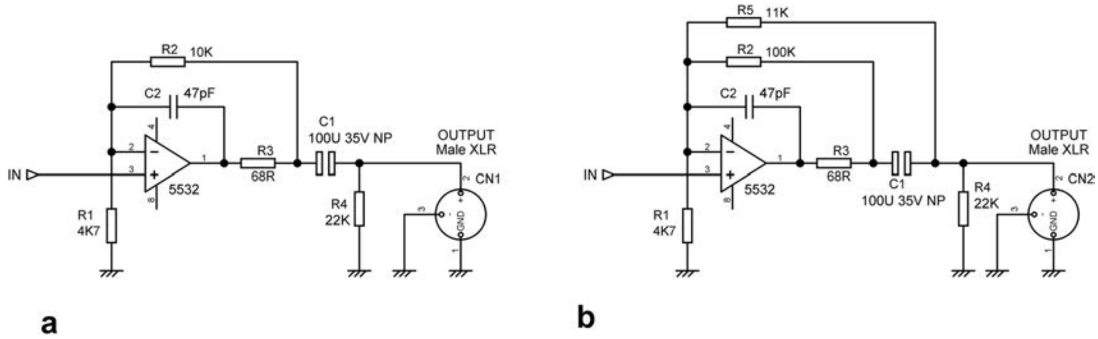

Figure 19.2: a) Zero-impedance output, b) zero-impedance output with NFB ground output capacitor

now much higher in value, and the HF feedback goes through C2 as before to maintain stability with capacitive loads. R2 and R5 in parallel come to 10 kΩ so the gain is the same. Any circuit with separate DC and AC feedback paths must be checked carefully for frequency response irregularities, which may happen well below 10 Hz. A function generator is useful for this.

Ground-cancelling outputs: basics

This technique, also called a ground-compensated output, appeared in the early 1980s in mixing consoles. It allows ground voltages to be cancelled out even if the receiving equipment has an unbalanced input; it prevents any possibility of creating a phase error by mis-wiring; and it costs virtually nothing except for the provision of a three-pin output connector.

Ground-cancelling (GC) separates the wanted signal from the unwanted ground voltage by addition at the output end of the link, rather than by subtraction at the input end. If the receiving equipment ground differs in voltage from the sending ground, then this difference is added to the output signal so that the signal reaching the receiving equipment has the same ground voltage superimposed upon it. Input and ground therefore move together and the ground voltage has no effect, subject to the usual effects of component tolerances. The connecting lead is differently wired from the more common unbalanced-out balanced-in situation, as now the cold line is joined to ground at the *input* or receiving end. This is illustrated in Figure 19.3, which compares a conventional balanced link with a GC link.

An inverting unity-gain ground-cancel output stage is shown in Figure 19.4a. The cold pin of the output socket is now an input, and has a unity-gain path summing into the main signal going to the hot output pin to add the ground voltage. This path R3, R4 has a very low input impedance equal to the hot terminal output impedance, so if it *is* used with a balanced input, the line impedances will be balanced and the combination will still work effectively. The 6 dB of attenuation in the R3–R4 divider is undone by the gain of two set by R5, R6. It is

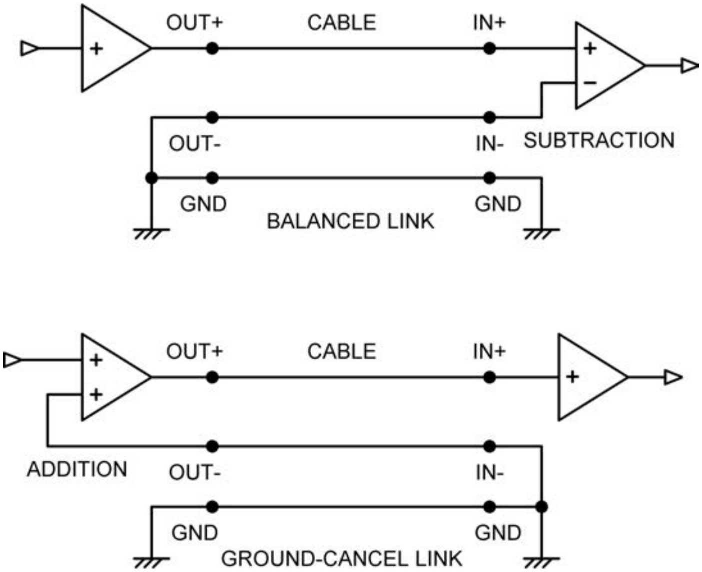

Figure 19.3: A balanced link uses subtraction at the receiving end to null ground noise, while a ground-cancel link uses addition at the sending end

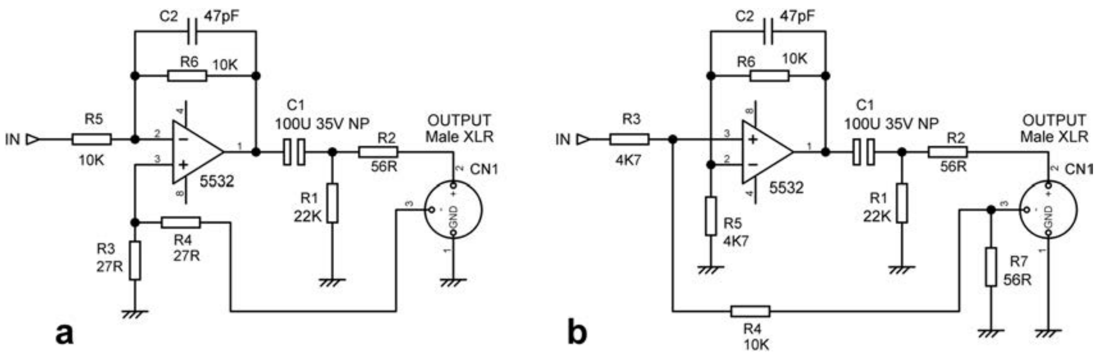

Figure 19.4: a) Inverting ground-canceling output, b) non-inverting ground-cancelling output

unfamiliar to most people to have the cold pin of an output socket as a low impedance input, and its very low input impedance minimises the problems caused by mis-wiring. Shorting it locally to ground merely converts the output to a standard unbalanced type. On the other hand, if the cold input is left unconnected, then there will be a negligible increase in noise due to the very low input resistance of R3.

This is the most economical GC output, and is very useful to follow an inverting summing amplifier in a mixer as it corrects the phase. However, a phase-inversion is not always convenient, and Figure 19.4b shows a non-inverting GC output stage with a gain of 6.6 dB. R5 and R6 set up a gain of 9.9 dB for the amplifier, but the overall gain is reduced by 3.3 dB

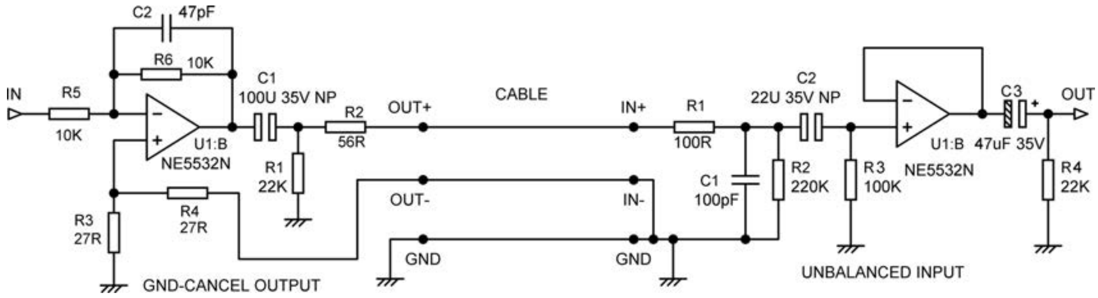

Figure 19.5: The complete circuit of a GC link using an inverting ground-cancel output

by attenuator R3, R4. The cold line is now terminated by R7, and any signal coming in via the cold pin is attenuated by R3, R4 and summed at unity gain with the input signal. The stage must be fed from a very low impedance such as an opamp output to work properly. There is a slight compromise on noise performance here because attenuation is followed by amplification.

Figure 19.5 shows the complete circuit of a GC link using an inverting ground-cancel output stage. EMC filtering and DC-blocking are included for the unbalanced input stage.

Ground-cancelling outputs: CMRR

In a balanced link, the CMRR is a measure of how accurately the subtraction is performed at the receiving end, and so of how effectively ground noise is nulled. A GC link also has an equivalent CMRR that measures how accurately the addition is performed at the sending end. Figure 19.6 shows (for the first time, I think) how to measure the CMRR of a ground-cancelling link. It is slightly more complicated than for the balanced case.

In Figure 19.6, a 10 Ω resistor R17 is inserted into the ground of the interconnection. This allows the signal generator to move the output ground of the sending amplifier up and down with respect to the global ground, via R18. Quite a lot of power has to be supplied so that R17 can be kept low in value.

If the send amplifier is working properly, the signal applied to OUT– will cancel the signal on the output ground, so that as far as the input of the receiving amplifier is concerned, it does not exist. Be clear that here we are measuring the CMRR of the sending amplifier, not the receiving amplifier.

Just as the CMRR of a balanced link depends on the accuracy of the resistors and the open-loop gain of the opamp in the receiving amplifier, the same parameters determine the CMRR of a ground-cancel send amplifier. The measured results from the arrangement in Figure 19.6 are given in Figure 19.7; the CMRR as built (with 1% resistors) was −50 dB, flat up to

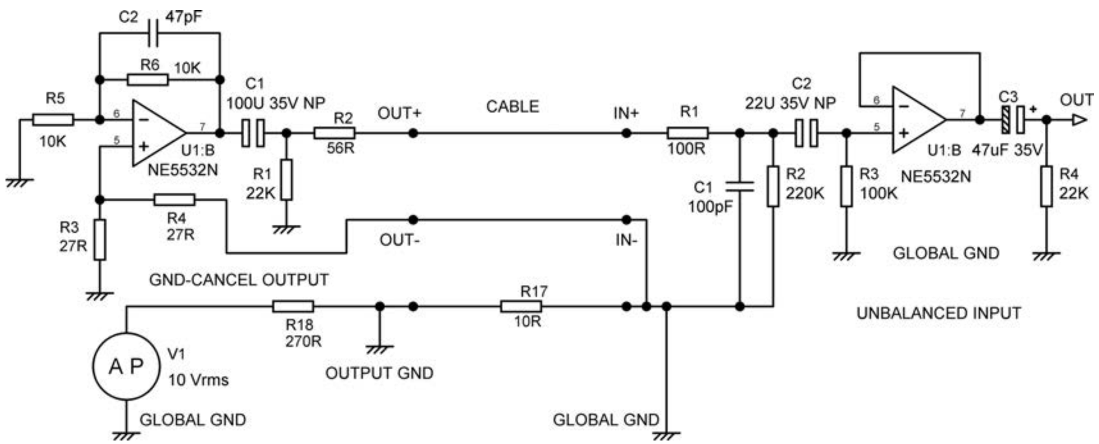

Figure 19.6: Measuring the CMRR of a GC link by inserting resistor R17

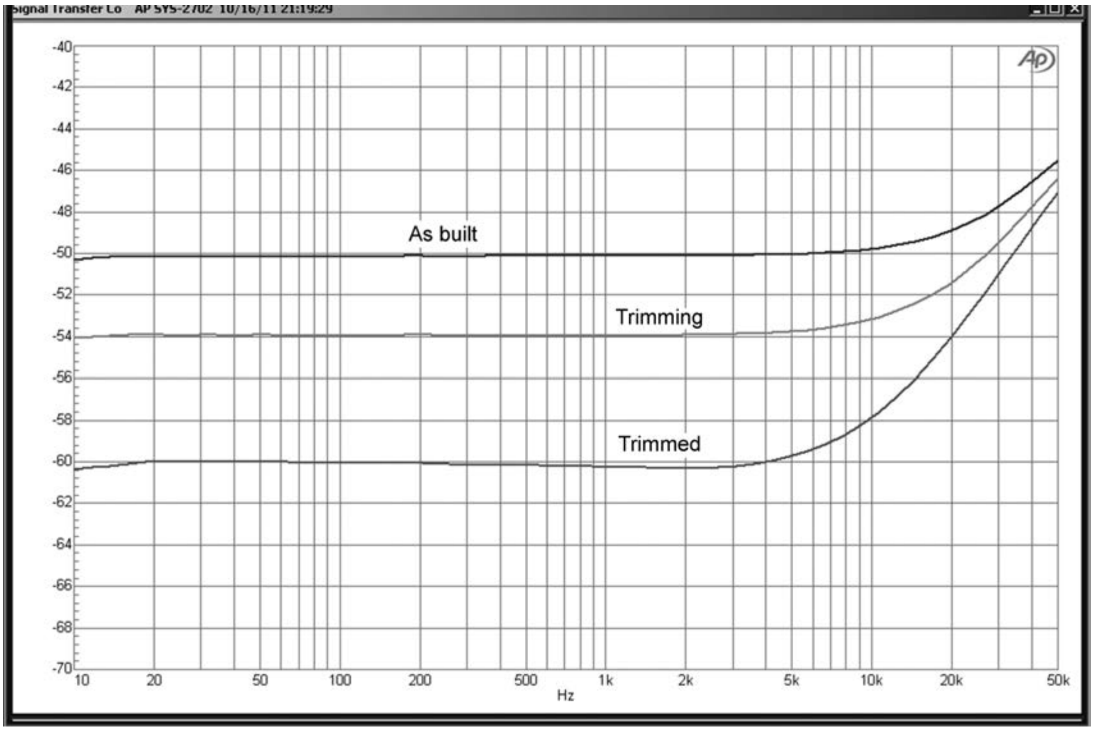

Figure 19.7: Optimising the ground-cancel send amplifier CMRR

10 kHz. The two lower traces were obtained by progressively trimming the value of R6 to minimise the output. If high CMRR is required, a preset adjustment can be used, just as in balanced line input amplifiers.

Ground-cancelling outputs are an economical way of making ground-loops innocuous when there is no balanced input, and it is rather surprising they are not more popular; perhaps it is because people find the notion of an input pin on an output connector unsettling. In particular, GC outputs would appear to offer the possibility of a quieter interconnection than the standard balanced interconnection because a relatively noisy balanced input is not required (see Chapter 18). Ground-cancelling outputs can also be made zero-impedance using the techniques described earlier.

Ground-cancelling outputs: send amplifier noise

In Figure 19.5, the main resistors R5 and R6 have the relatively high value of 10 kΩ, for comparison with the 'standard' balanced input amplifier made with four 10 kΩ resistors. Reducing their value in accordance with the principles of low-impedance design gives useful reductions in noise, as shown in Table 19.1. A 5532 opamp was used.

A very useful 4.3 dB reduction in noise is gained by going to 2k2, at zero cost, but after that the improvements become smaller, for while Johnson noise and the effects of current noise are reduced, the voltage noise of the opamp is unchanged.

It is instructive to compare the signal/noise ratio with that of a balanced link. We will put 1 Vrms (+2.2 dBu) down a balanced link. The balanced output stage raises the level by 6 dB, so we have +8.2 dBu going down the cable. A conventional unity-gain balanced input made with 10 kΩ resistors and a 5532 section has a noise output of −104.8 dBu, so the signal/noise ratio is 8.2 + 104.8 = 113.0 dB.

In the ground-cancel case, if we use 1 kΩ resistors as in the fourth row of Table 19.1, the noise from the ground-cancel output stage is −112.2 dBu. Adding the noise of the

TABLE 19.1 GC output noise improvement by reducing the value of R5, R6 (5532 opamp)

Value of R5, R6	Noise output (dBu)	Improvement (dB)
10 kΩ	−107.2	0.0
2k2	−111.3	−4.3
1 kΩ	−112.2	−5.0
560 Ω	−112.6	−5.4

5532 buffer at the receiving end in Figure 19.5, which is −125.7 dBu, we get −112.0 dBu as the noise floor. The signal/noise ratio is therefore 2.2 + 112.0 = 114.2 dB. This is 1.2 dB quieter than the conventional balanced link, and it only uses two opamp sections instead of three.

The noise situation could easily be reversed by using a low-impedance balanced input with buffers to make the input impedance acceptably high, as described in the previous chapter, but we are then comparing a ground-cancel link using two opamp sections with a balanced link using five.

Balanced outputs: basics

Figure 19.8a shows a balanced output, where the cold terminal carries the same signal as the hot terminal but phase-inverted. This can be arranged simply by using an opamp stage to invert the normal in-phase output. The resistors R3, R4 around the inverter should be as low in value as possible to minimise Johnson noise, because this stage is working at a noise-gain of two, but bear in mind that R3 is effectively grounded at one end and its loading, as well as the external load, must be driven by the first opamp. A unity-gain follower is shown for the first amplifier, but this can be any other shunt or series feedback stage as convenient. The inverting output if not required can be ignored; it must *not* be grounded, because the inverting opamp will then spend most of its time clipping in current-limiting, probably injecting unpleasant distortion into the preamp grounding system. Both hot and cold outputs must have the same output impedances (R2, R6) to keep the line impedances balanced and the interconnection CMRR maximised.

A balanced output has the advantage that the total signal level on the line is increased by 6 dB, which will improve the signal-to-noise ratio if a balanced input amplifier is being driven, as they are relatively noisy. It is also less likely to crosstalk to other lines even if they are unbalanced, as the currents injected via the stray capacitance from each line will tend to cancel; how well this works depends on the physical layout of the conductors. All balanced outputs give the facility of correcting phase errors by swapping hot and cold outputs. This is however a two-edged sword, because it is probably how the phase got wrong in the first place.

There is no need to worry about the exact symmetry of level for the two output signals; ordinary tolerance resistors are fine. Such gain errors only affect the signal-handling capacity of the interconnection by a small amount. This simple form of balanced output is the norm in hi-fi balanced interconnection, but is less common in professional audio, where the quasi-floating output described later gives more flexibility.

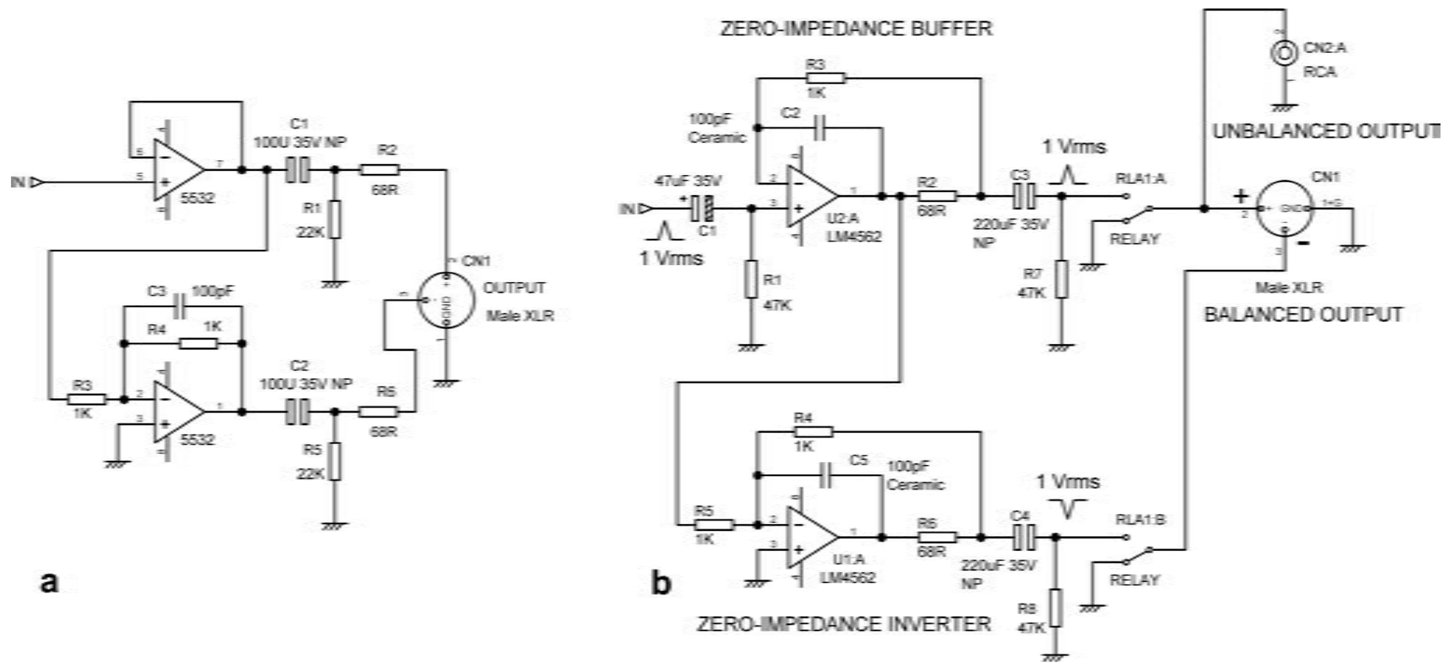

Figure 19.8: a) A conventional balanced output, b) a zero-impedance balanced output with relay muting

Balanced outputs: output impedance

The balanced output stage of Figure 19.8a has an output impedance of 68 Ω on both legs because this is the value of the series output resistors; however, as noted earlier, the lower the output impedance the better, so long as stability is maintained. This balanced output configuration can be easily adapted to have two zero-impedance outputs, as shown in Figure 19.8b. The unity-gain buffer that drives the hot output is configured in the same way as the +10 dB zero-impedance stage described earlier. The zero-impedance inverter that drives the cold output works similarly, but with shunt negative feedback via R4.

The quickest way to measure normal output impedances is to load each output as heavily as is practical (say with 560 Ω) and measure the voltage drop compared with the unloaded state. From this the output impedance is simply calculated. However, in the case of zero-impedance outputs the voltage drop is very small and so measurement accuracy is poor.

A more sophisticated technique is the injection of a test signal current into the output and measuring the voltage that results. This is much more informative; the results for the hot output in Figure 19.8b are given in Figure 19.9. A suitable current to inject is 1 mA rms,

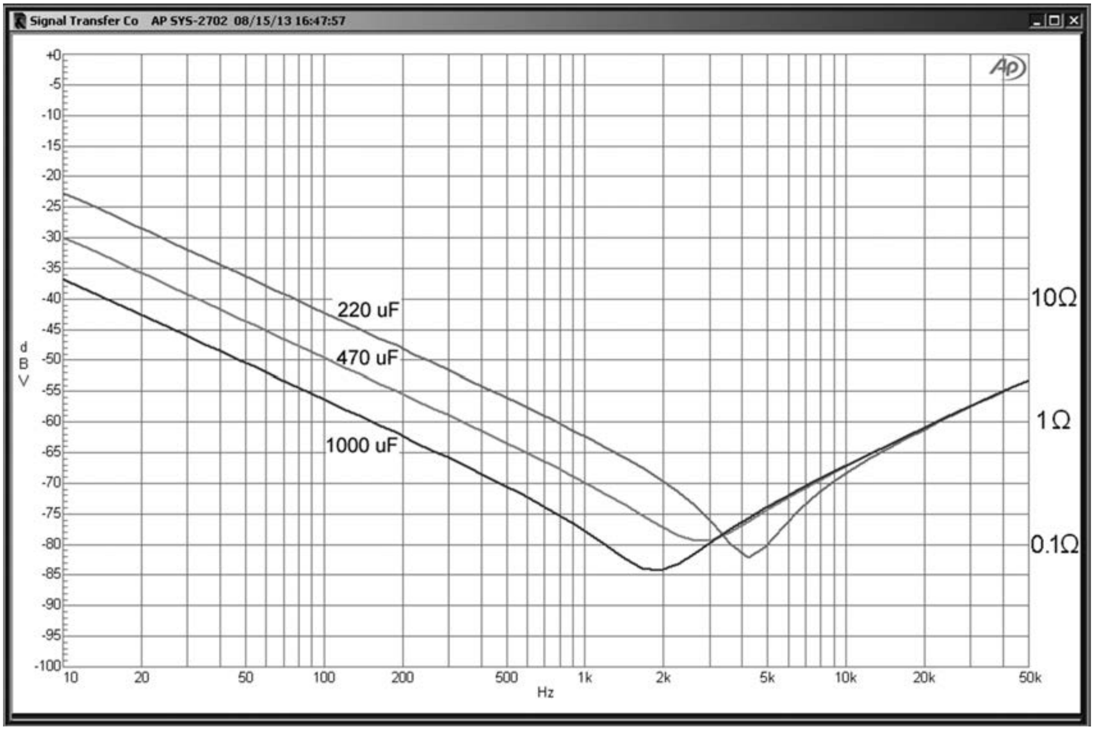

Figure 19.9: The signal voltage at the hot output of the balanced amplifier for 1 mA rms injection current, and the output impedance modulus. Three values of output capacitor

defined by applying 1 Vrms to a 1 kΩ injection resistor. A 1 Ω output impedance will therefore givxe an output signal voltage of 1 mVrms (−60 dBV), which can be easily measured; see Figure 19.9. Likewise −40 dBV represents 10 Ω, and −80 dBV represents 0.1 Ω; a log scale is useful because it allows a wide range of impedance to be displayed and gives convenient straight lines on the plot. The opamp sections were both LM4562.

In Figure 19.9 the modulus of the output impedance falls to below 0.1 Ω between 1 and 5 kHz. The rise above the latter frequency is due to the effect of C2 in Figure 19.8b, and the open-loop gain of the opamp falling as frequency increases. Below 3 kHz the impedance increases steadily as frequency falls, doubling with each octave due to the reactance of the output capacitor. This effect can be reduced simply by increasing the size of the capacitor. In Figure 19.9 the results are shown for 220 μF, 470 μF, and 1000 μF output capacitors. For 1000 μF a broad null occurs reaching down to 0.03 Ω, and the output impedance is below 1 Ω between 150 Hz and 20 kHz. Bear in mind that output capacitors should be non-polarised types, as they may face external DC voltages of either polarity, and should be rated at no less than 25 V; this means that a 1000 μF capacitor can be quite a large component.

The output impedance plot for the cold output was identical to Figure 19.9. This was rather a surprise as the inverter works at a noise-gain of two times, as opposed to the buffer which works at a noise-gain of unity, and so I expected it to show twice the output impedance above 5 kHz. It is not currently known why this is not so.

Since wiring resistances internal to the equipment are likely to be in the region 0.1 to 0.5 Ω there seems nothing much to be gained by reducing the output impedances any further than is achieved by this simple zero-impedance technique.

You may be thinking that the zero-impedance output is a bit of a risky business; will it always be stable when loaded with capacitance? In my wide experience of this technique, the answer is yes, so long as you design it properly. If you are attempting something different from proven circuitry, it is always wise to check the stability of zero-impedance outputs with a variety of load capacitances. In the example of Figure 19.8b both hot and cold outputs were separately checked using a 5 Vrms sweep from 50 kHz to 10 Hz, with load capacitances of: 470 pF, 1 nF, 2n2, 10 nF, 22 nF and 100 nF. At no point was there the slightest hint of instability. A load of 100 nF is of course grossly excessive compared with real use, being equivalent to about 1000 metres of average screened cable, and curtails the output swing at HF.

Balanced outputs: noise

The noise output of the zero-impedance balanced output of Figure 19.8b was measured with 0 Ω source resistance, rms response, unweighted; measured at two bandwidths to demonstrate the absence of hum; the opamp sections were both LM4562. See Table 19.2.

TABLE 19.2 Noise output of the zero-impedance balanced output of Figure 19.8b

Output	Noise out (dBu)	Bandwidth
Hot output only	−113.5	22 Hz–22 kHz
Hot output only	−113.8	400 Hz–22 kHz
Balanced output (hot and cold)	−110.2	22 Hz–22 kHz
Balanced output (hot and cold)	−110.5	400 Hz–22 kHz

Quasi-floating outputs

The purely electronic output stage in Figure 19.10 emulates a floating transformer winding; if both hot and cold outputs are driving signal lines, then the outputs are balanced, as if a centre-tapped output transformer were being used, though clearly the output is not galvanically isolated from ground. If, however, the cold output is grounded, the hot output doubles in amplitude so the total level hot-to-cold is unchanged. This condition is detected by the current-sensing feedback taken from the outside of the 75 Ω resistor R10, and the current driven into the shorted cold output is automatically reduced to a low level that will not cause problems.

Similarly, if the hot output is grounded, the cold output doubles in amplitude and remains out of phase; the total hot–cold signal level is once more unchanged. This system has the advantage that it can give the same level into either a balanced or unbalanced input, given an appropriate connector at the input end. 6 dB of headroom is however lost when the output is used in unbalanced mode. It is most useful in recording studios where various bits of equipment may be temporarily connected; it is of less value in a PA system with a fixed equipment line-up.

When an unbalanced output is being driven, the quasi-floating output can be wired to work as a ground-cancelling connection, with rejection of ground noise no less effective than the true balanced mode. This requires the cold output to be grounded at the remote (input) end of the cable. Under adverse conditions this might cause HF instability, but in general the approach is sound. If you are using exceptionally long cable, then it is wise to check that all is well.

If the cold output is grounded locally, i.e. at the sending end of the cable, then it works as a simple unbalanced output, with no noise rejection. When a quasi-floating output stage is used unbalanced, the cold leg *must* be grounded, or common-mode noise will degrade the noise floor by at least 10 dB, and there may be other problems with increased distortion.

Quasi-floating outputs use a rather subtle mixture of positive and negative feedback of current and voltage. This performs the required function quite well, but a serious drawback is that it

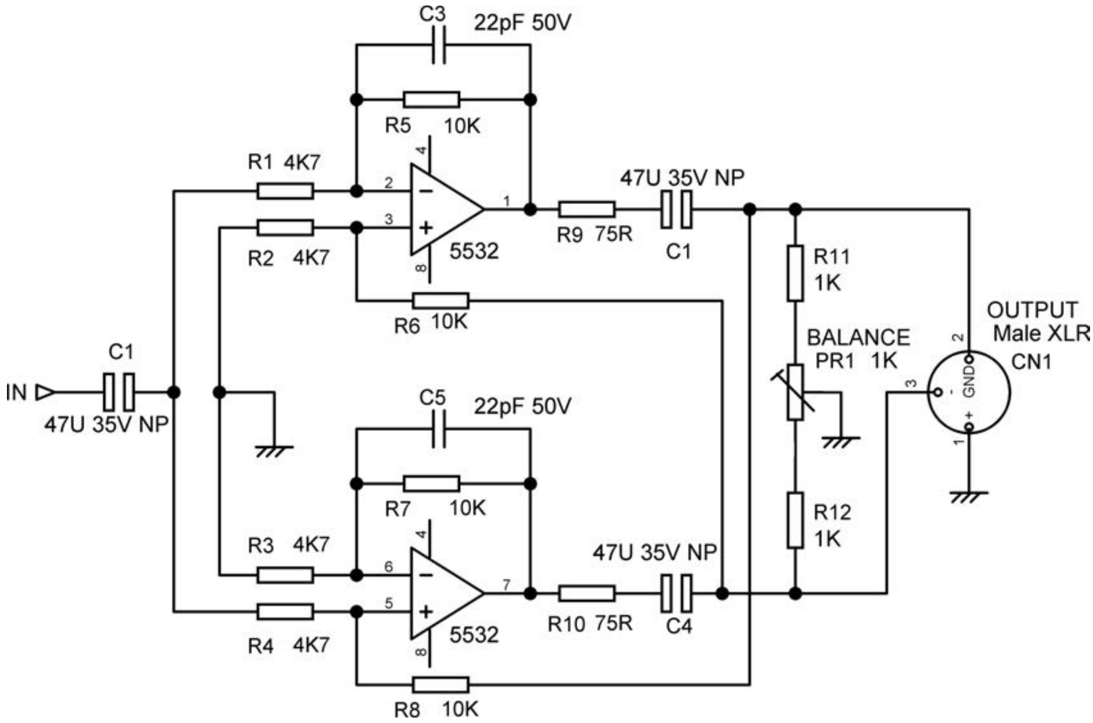

Figure 19.10: Quasi-floating balanced output

accentuates the effect of resistor tolerances, and so a preset resistor is normally required to set the outputs for equal amplitude; the usual arrangement is shown in Figure 19.10. If it is not correctly adjusted one side of the output will clip before the other and reduce the total output headroom. This is a set-and-forget adjustment unless it becomes necessary to change any of the resistors in the circuit.

Transformer balanced outputs

If true galvanic isolation between equipment grounds is required, this can only be achieved with a line transformer, sometimes called a line isolating transformer. You don't use line transformers unless you really have to because the much-discussed cost, weight, and performance problems are very real, as you will see shortly. However, they are sometimes found in big sound reinforcement systems (for example, in the mic-splitter box on the stage) and in any environment where high RF field strengths are encountered.

A basic transformer balanced output is shown in Figure 19.11a; in practice A1 would probably be providing gain rather than just buffering. In good-quality line transformers

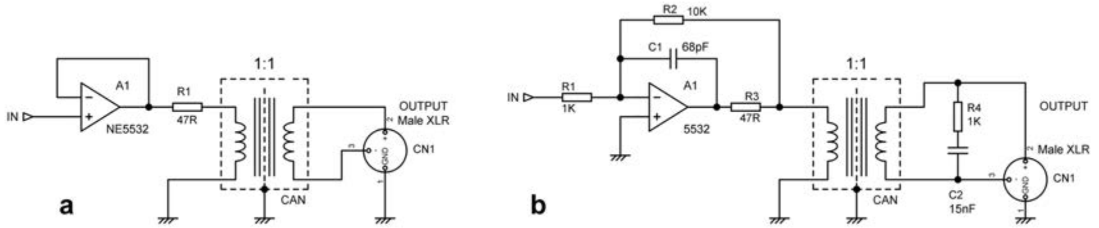

Figure 19.11: Transformer balanced outputs; a) standard circuit, b) zero-impedance drive to reduce LF distortion, also with Zobel network across secondary

there will be an inter-winding screen, which should be earthed to minimise noise pickup and general EMC problems. In most cases this does *not* ground the external can and you have to arrange this yourself, possibly by mounting the can in a metal capacitor clip. Make sure the can is grounded as this definitely does reduce noise pickup.

Be aware that the output impedance will be higher than usual because of the ohmic resistance of the transformer windings. With a 1:1 transformer, as normally used, both the primary and secondary winding resistances are effectively in series with the output. A small line transformer can easily have 60 Ω per winding, so the output impedance is 120 Ω plus the value of the series resistance R1 added to the primary circuit to prevent HF instability due to transformer winding capacitances and line capacitances. The total can easily be 160 Ω or more, compared with, say, 47 Ω for non-transformer output stages. This will mean a higher output impedance and greater voltage losses when driving heavy loads.

DC flowing through the primary winding of a transformer is bad for linearity, and if your opamp output has anything more than the usual small offset voltages on it, DC current flow should be stopped by a blocking capacitor.

Output transformer frequency response

If you have looked at the section in Chapter 18 on the frequency response of line input transformers, you will recall that they give a nastily peaking frequency response if the secondary is not loaded properly, due to resonance between the leakage inductance and the stray winding capacitances. Exactly the same problem afflicts output transformers, as shown in Figure 19.12; with no output loading there is a frightening 14 dB peak at 127 kHz. It is high enough in frequency to have very little effect on the response at 20 kHz, but this high-Q resonance isn't the sort of horror you want lurking in your circuitry. It could easily cause some nasty EMC problems.

The transformer measured was a Sowter 3292 1:1 line isolating transformer. Sowter are a highly respected company, and this is a quality part with a mumetal core and housed in

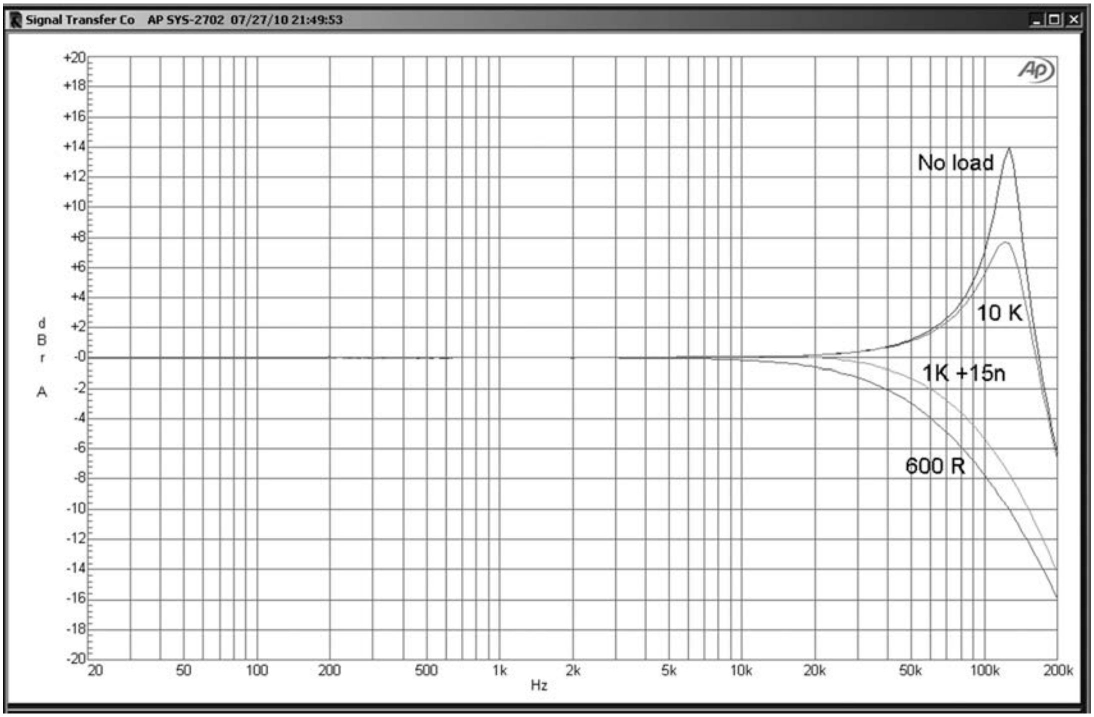

Figure 19.12: Frequency response of a Sowter 3292 output transformer with various loads on the secondary. Zero-impedance drive as in Figure 19.8b

a mumetal can for magnetic shielding. When used as the manufacturer intended, with a 600 Ω load on the secondary, the results are predictably quite different, with a well-controlled roll-off that I measured as −0.5 dB at 20 kHz.

The difficulty is that there are very few, if any, genuine 600 Ω loads left in the world, and most output transformers are going to be driving much higher impedances. If we are driving a 10 kΩ load, the secondary resonance is not much damped and we still get a thoroughly unwelcome 7 dB peak above 100 kHz, as shown in Figure 19.12. We could, of course, put a permanent 600 Ω load across the secondary, but that will heavily load the output opamp, impairing its linearity, and will give us unwelcome signal loss due in the winding resistances. It is also a profoundly inelegant way of carrying on.

A better answer, as in the case of the line input transformer, is to put a Zobel network, i.e. a series combination of resistor and capacitor, across the secondary, as in Figure 19.11b. The capacitor required is quite small and will cause very little loading except at high frequencies where signal amplitudes are low. A little experimentation yielded the values of 1 kΩ in series with 15 nF, which gives the much improved response shown in Figure 19.12. The response is almost exactly 0.0 dB at 20 kHz, at the cost of a very gentle 0.1 dB rise around 10 kHz; this could probably be improved by a little more tweaking of the Zobel values. Be aware that a different transformer type will require different values.

Output transformer distortion

Transformers have well-known problems with linearity at low frequencies. This is because the voltage induced into the secondary winding depends on the rate of change of the magnetic field in the core, and so the lower the frequency, the greater the change in magnitude must be for transformer action [1]. The current drawn by the primary winding to establish this field is non-linear, because of the well-known non-linearity of iron cores. If the primary had zero resistance, and was fed from a zero source impedance, as much distorted current as was needed would be drawn and no one would ever know there was a problem. But there is always some primary resistance, and this alters the primary current drawn so that third-harmonic distortion is introduced into the magnetic field established, and so into the secondary output voltage. Very often there is a series resistance R1 deliberately inserted into the primary circuit, with the intention of avoiding HF instability; this makes the LF distortion problem worse, and a better means of isolation is a low value inductor of say 4 μH in parallel with a low-value damping resistor of around 47 Ω. This is more expensive and is only used on high-end consoles.

An important point is that this distortion does not appear only with heavy loading – it is there all the time, even with no load at all on the secondary; it is not analogous to loading the output of a solid-state power amplifier, which invariably increases the distortion. In fact, in my experience transformer LF distortion is slightly better when the secondary is connected to its rated load resistance. With no secondary load, the transformer appears as a big inductance, so as frequency falls the current drawn increases, until with circuits like Figure 19.11a, there is a sudden steep increase in distortion around 10–20 Hz as the opamp hits its output current limits. Before this happens, the distortion from the transformer itself will be gross.

To demonstrate this I did some distortion tests on the same Sowter 3292 transformer that was examined for frequency response. The winding resistance for both primary and secondary is about 59 Ω. It is quite a small component, 34 mm in diameter and 24 mm high and weighing 45 gm, and is obviously not intended for transferring large amounts of power at low frequencies. Figure 19.13 shows the LF distortion with no series resistance, driven directly from a 5532 output (there were no HF stability problems in this case, but it might be different with cables connected to the secondary) and with 47 and 100 Ω added in series with the primary. The flat part to the right is the noise floor.

Taking 200 Hz as an example, adding 47 Ω in series increases the THD from 0.0045% to 0.0080%, figures which are in exactly the same ratio as the total resistances in the primary circuit in the two cases. It's very satisfying when a piece of theory slots right home like that. Predictably, a 100 Ω series resistor gives even more distortion, namely 0.013% at 200 Hz, and once more proportional to the total primary resistance.

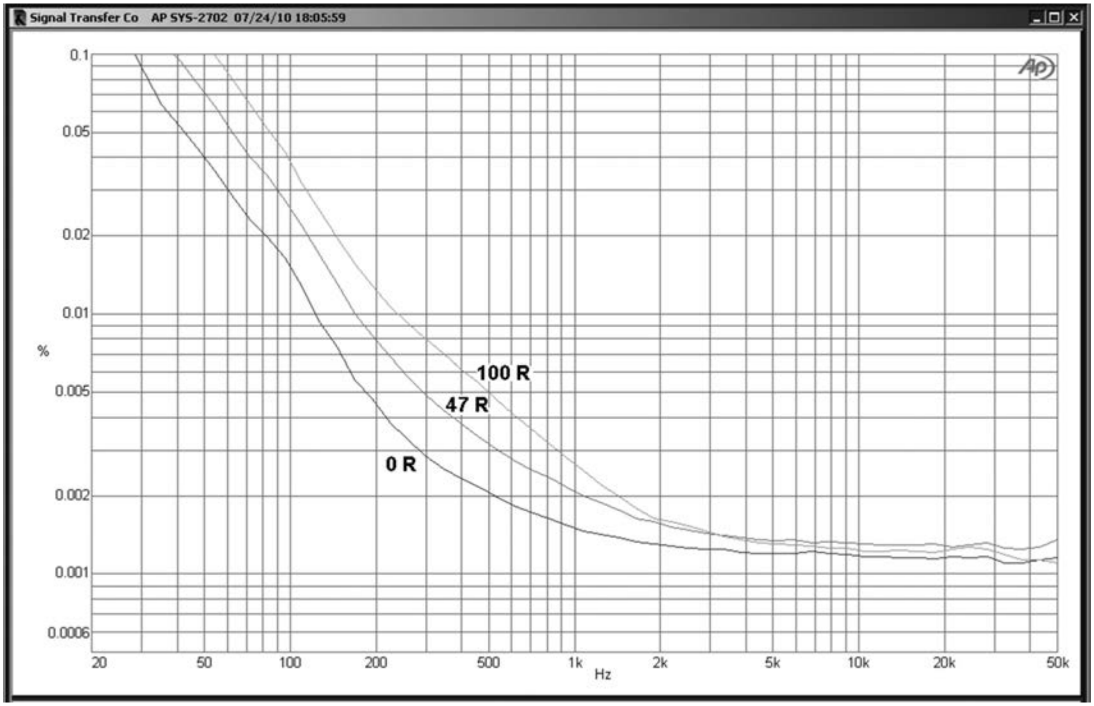

Figure 19.13: The LF distortion rise for a 3292 Sowter transformer, without (0R) and with (47 Ω and 100 Ω) extra series resistance. Signal level 1 Vrms

If you're used to the near-zero LF distortion of opamps, you may not be too impressed with Figure 19.13, but this is the reality of output transformers. The results are well within the manufacturer's specifications for a high-quality part. Note that the distortion rises rapidly to the LF end, roughly tripling as frequency halves. It also increases fast with level, roughly quadrupling as level doubles. Having gone to some pains to make electronics with very low distortion, this non-linearity at the very end of the signal chain is distinctly irritating. The situation is somewhat eased in actual use as signal levels in the bottom octave of audio are normally about 10–12 dB lower than the maximum amplitudes at higher frequencies.

Reducing output transformer distortion

In audio electronics, as in so many other areas of life, there is often a choice between using brains or brawn to tackle a problem. In this case 'brawn' means a bigger transformer, such as the Sowter 3991, which is still 34 mm in diameter but 37 mm high, weighing in at 80 gm. The extra mumetal core material improves the LF performance, but you still get a distortion plot very much like Figure 19.13 (with the same increase of THD with series resistance) except now it occurs at 2 Vrms instead of 1 Vrms. Twice the metal, twice the level – I suppose it makes sense. You can take this approach a good deal further with the Sowter 4231, a much bigger open-frame design tipping the scales at a hefty 350 gm. The winding resistance for the primary is 12 Ω and for the secondary 13.3 Ω, both a good deal lower than the previous figures.

Figure 19.14 shows the LF distortion for the 4231 with no series resistance, and with 47 and 100 Ω added in series with the primary. The flat part to the right is the noise floor. Comparing it with Figure 19.13, the basic distortion at 30 Hz is now 0.032%, compared with about 0.10% for the 3292 transformer. While this is a useful improvement it is gained at considerable expense. Now adding 47 Ω of series resistance has dreadful results – distortion increases by about five times. This is because the lower winding resistances of the 4231 mean that the added 47 Ω has increased the total resistance in the primary circuit to five times what it was. Predictably, adding a 100 Ω series resistance approximately doubles the distortion again. In general bigger transformers have thicker wire in the windings, and this in itself reduces the effect of the basic core non-linearity, quite apart from the improvement due to more core material. A lower winding resistance also means a lower output impedance.

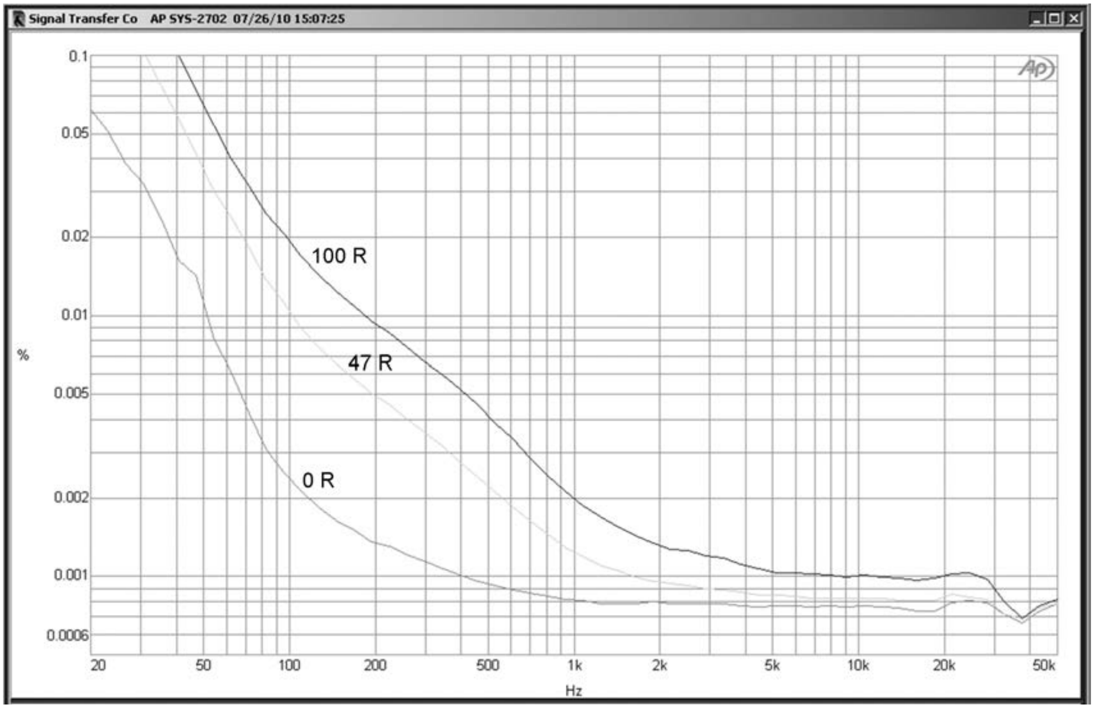

Figure 19.14: The LF distortion rise for the much larger 4231 Sowter transformer, without and with extra series resistance. Signal level 2 Vrms

The LF non-linearity in Figure 19.14 is still most unsatisfactory compared with that of the electronics. Since the 'My policy is copper and iron!' [2] approach does not really solve the problem, we'd better put brawn to one side and try what brains we can muster.

We have seen that adding series resistance to ensure HF stability makes things definitely worse, and a better means of isolation is a low value inductor of say 4 μH paralleled with a low-value damping resistor of around 47 Ω. However inductors cost money and a more economic solution is to use a zero-impedance output as shown in Figure 19.11b above. This gives the same results as no series resistance at all, but with wholly dependable HF stability. However, the basic transformer distortion remains because the primary winding resistance is still there, and its level is still too high. What can be done?

The LF distortion can be reduced by applying negative feedback via a tertiary transformer winding, but this usually means an expensive custom transformer, and there may be some interesting HF stability problems because of the extra phase-shift introduced into the feedback by the tertiary winding; this approach is discussed in reference [3]. However, what we really want is a technique that will work with off-the-shelf transformers.

A better way is to cancel out the transformer primary resistance by putting in series an electronically-generated negative resistance; the principle is shown in Figure 19.15, where a zero-impedance output is used to eliminate the effect of the series stability resistor. The 56 Ω resistor R4 senses the current through the primary, and provides positive feedback to A1, proportioned so that a negative output resistance of twice the value of R4 is produced, which

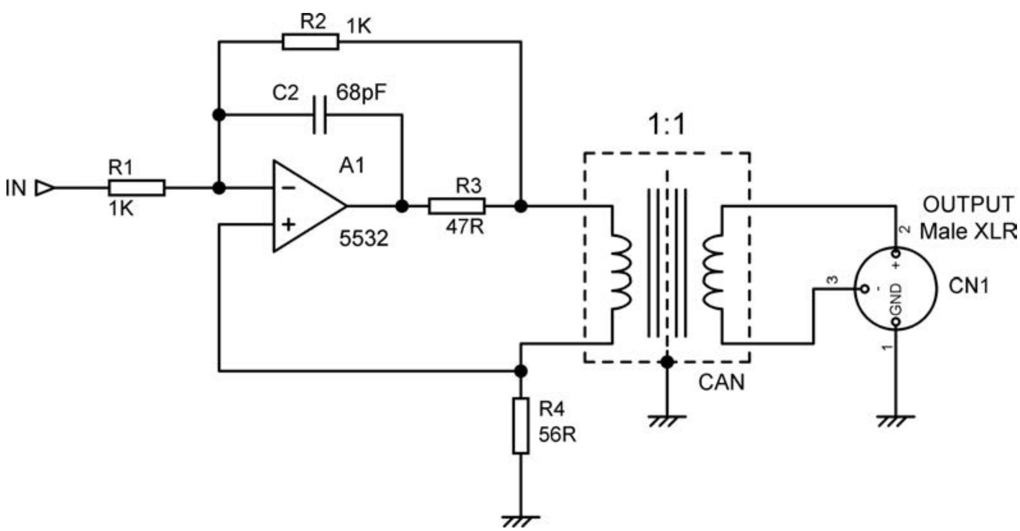

Figure 19.15: Reducing LF distortion by cancelling out the primary winding resistance with a negative resistance generated by current-sensing resistance R4. Values for Sowter 3292 transformer

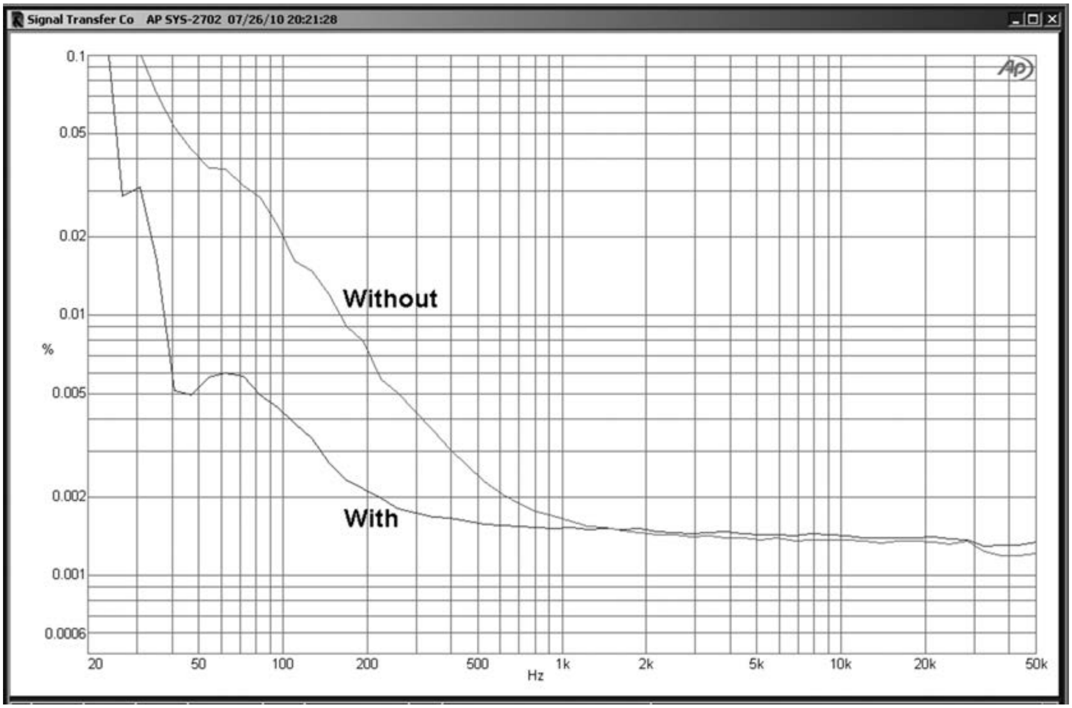

Figure 19.16: The LF distortion rise for a 3292 Sowter transformer, without and with winding resistance cancellation as in Figure 19.15. Signal level 1 Vrms

will cancel out both R4 itself and most of the primary winding resistance. As we saw earlier, the primary winding resistance of the 3292 transformer is approx 59 Ω, so if R4 was 59 Ω we should get complete cancellation. But . . .

It has always necessary to use positive feedback with caution. Typically it works, as here, in conjunction with good old-fashioned negative feedback, but if the positive exceeds the negative (this is one time you do *not* want to accentuate the positive) then the circuit will typically latch up solid, with the output jammed up against one of the supply rails. R4 = 56 Ω in Figure 19.15 worked reliably in all my tests, but increasing it to 68 Ω caused immediate problems, which is precisely what you would expect. No input DC-blocking capacitor is shown in Figure 19.15 but it can be added ahead of R1 without increasing the potential latch-up problems. The small Sowter 3292 transformer was used.

This circuit is only a basic demonstration of the principle of cancelling primary resistance, but as Figure 19.16 shows it is still highly effective. The distortion at 100 Hz is reduced by a factor of five, and at 200 Hz by a factor of four. Since this is achieved by adding one resistor, I think this counts as a definite triumph of brains over brawn, and indeed confirmation of the old adage that size is less important than technique.

The method is sometimes called 'mixed feedback' as it can be looked at as a mixture of voltage and current feedback. Since the primary resistance is cancelled, there is a second advantage as the output impedance of the stage is reduced. The secondary winding resistance is, however, still in circuit, and so the output impedance is usually only halved. The principle can also be applied when a balanced drive to the output transformer is used.

If you want better performance than this – and it is possible to make transformer non-linearity effectively invisible down to 15 Vrms at 10 Hz – there are several deeper issues to consider. The definitive reference is Bruce Hofer's patent, which covers the transformer output of the Audio Precision measurement systems [4]. There is also more information in the Analog Devices *Opamp Applications Handbook* [5].

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CHAPTER 20

Headphone amplifiers

Driving heavy loads

This chapter focuses on techniques for driving loads heavier than can be handled by a single 5532 or LM4562 opamp–basically any impedance less than 500 Ω. This is typically a headphone load, the impedance of headphones varying widely from 600 Ω to 16 Ω or less, with sensitivity inversely related so the higher impedance models need more drive voltage. The need to drive impedances below 500 Ω is also likely to arise in very low-noise designs, so that special measures have to be taken just to drive the next stage. A good example is my Preamp 2012 design for Elektor [1]. It is assumed here that the heavy loads must be driven to the same high standards of noise and distortion that are required for lighter loading.

In writing this chapter I have become conscious that it would be easy to find enough material for a complete book on headphone amplifiers. Inevitably I have had to be very selective, but I think you will find plenty of new information here.

Driving headphones

The wide range of load impedances is accompanied by a wide range of sensitivity. Headphones of 600 Ω impedance require ten times or more voltage to achieve the same sound pressure level (SPL) than do those of 30 Ω impedance, though there is much variation. The sensitivity of most of the headphones currently on the market is covered by a range of +110 to +130 dB SPL per Volt. If a rather loud +110 dB SPL is taken as the maximum level required, voltages between 110 mV and 1.1 V will be needed. These are low voltages compared with the maximum of about 10 Vrms that can be obtained from standard ±17 V rails, and you may be thinking that it would be more efficient to run a headphone amplifier off much lower rails, such as ±5 V. While this is true, it is rarely done, as a) it is much easier to get low distortion with the standard rails, and b) extra power supplies are expensive to provide.

The traditional solution to the wide impedance/sensitivity range is a resistor of the order of 50 Ω to 100 Ω in series with the amplifier output, which reduces the maximum output voltage available as the load impedance is reduced. This has a valuable safety function as applying the full voltage required for 600 Ω headphones to 30 Ω units will almost certainly wreck them, as well as generating SPLs that will cause hearing damage. The series-resistor technique has been used extensively, despite the fact that it reduces the so-called 'damping factor' of the amplifier-headphone combination to about 0.3. The 'damping factor' of a power amplifier-loudspeaker combination is a meaningless ratio because the amplifier output impedance is a tiny fraction of the loudspeaker voice-coil resistance, so the latter utterly dominates the transient behaviour; here the amplifier output impedance is greater than that of the transducer, and therefore could be expected to have a serious effect on its frequency response and transient behaviour. Nevertheless, this technique has been in wide use for many decades; there seems to be a general feeling that 'damping factor' does not apply to headphones, though it is far from clear why this should be the case. It also seems to be the general view that capacitor-coupled outputs are acceptable for headphones, but not full-scale power amplifiers. There is some technical justification for this as large electrolytic capacitors can introduce distortion when passing large signal currents, even at mid-frequencies [2].

The series-resistor approach has two other important advantages. Firstly, so long as the resistor has a sufficient power rating, the headphone amplifier is inherently short circuit proof without any need for protection circuitry that might, if ill-conceived, degrade the distortion performance by premature operation. DC-offset protection may still be required. Secondly, the amplifier should be stable into any conceivable load without the need for Zobel networks or output inductors.

A solution to the output impedance problem is to retain the series resistor but take the negative feedback from the load side of it. This will give a very low output impedance (see the zero-impedance outputs in Chapter 19) but retains the power-limiting action and inherent immunity to short circuits. However, a Zobel network and output inductor are now likely to be required for reliable stability with all loads.

Special opamps

Opamps do exist with a greater load-driving capability (for example, the NJM4556A from JRC is capable of driving 150 Ω) but this device seems to have achieved little market penetration, possibly because its linearity even into light loads is distinctly inferior to the 5532 and LM4562. By 2009 none of the usual distributors were carrying it, and it is not clear if it is still in production.

Multiple opamps

In some circumstances, paralleled opamp stages are the simplest answer. This is sometimes called an opamp-array or multipath amplifier. There is much more information on this in Chapter 1, focusing mainly on the noise benefits; here I will concentrate on load-driving

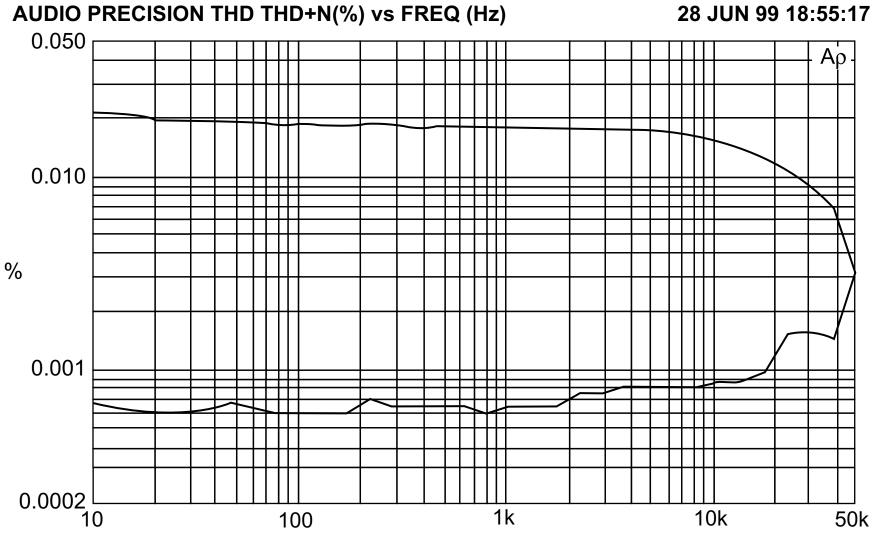

Figure 20.1: One and two 5532 sections attempting to drive 5 Vrms into 220 Ω

ability. In Figure 20.1 the upper trace shows a single 5532 section attempting to drive 5 Vrms into 220 Ω. The distortion is very high for a 5532, and it is clearly running out of current capability. Adding a second section in parallel with the outputs coupled by 10 Ω resistors, as shown in Figure 1.10 in Chapter 1, drops the THD back to the familiar low levels. The gains of the paralleled voltage-follower stages are very closely equal, so the small series output resistors can allow for any microscopic differences, and no significant currents pass from opamp to opamp.

This can be a very simple and cost-effective way to solve the problem of driving medium loads, and it can be extended by connecting more 5532 sections in parallel, essentially without limit. Figure 20.2 shows four parallel 5532 sections driving 100 Ohms to 5 Vrms with very little distortion. An array of 12 parallel 5532 sections can drive a 25 Ω load with 5 Vrms very effectively. Being me, I took the opamp-array technique to its logical conclusion, and possibly beyond, in my power amplifier for Elektor [3] which used 32 opamp packages, giving 64 parallel sections of 5532 joined with 1 Ω sharing resistors. This can easily drive an 8 Ω loudspeaker to 15 Wrms, with the usual very low 5532 distortion. The opamp supply rails limit the power output, so another version was designed that used two of these amplifiers bridged to give about 60 W/8 Ω; the number of opamps in each amplifier was doubled to cope with the doubled load currents, so there were 128-off 5532 sections in parallel with 1 Ω sharing resistors. This project was completely straightforward to design and test, and I think demonstrates conclusively that you can use as many parallel opamps as you like.

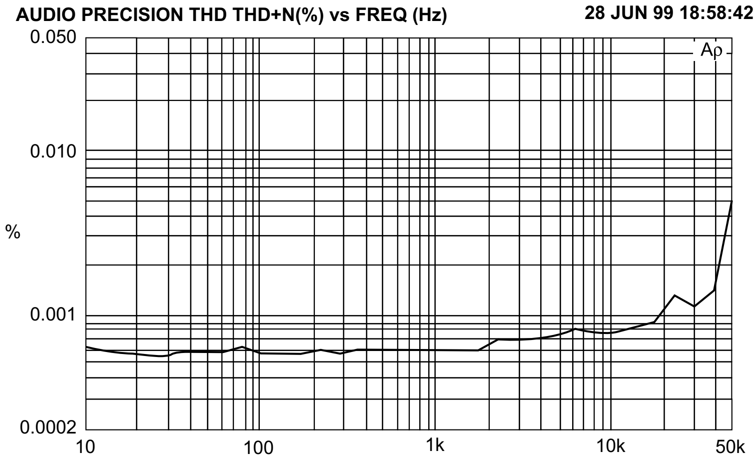

Figure 20.2: Four parallel 5532 sections driving 100 Ω with 5 Vrms. The THD plot is almost identical to two 5532 sections driving 220 Ω

Since the 5532 is usually the cheapest opamp available, using a number of them is not costly. The LM4562 will unquestionably give lower distortion but at the time of writing is significantly more expensive.

The best linearity is given by separating the gain and load driving functions, as shown in Figure 20.5 further on. The first opamp provides all the voltage gain, so the output amps can be operated as voltage-followers with 100% negative feedback to give the best linearity.

As described in Chapter 1, paralleled opamps not only increase drive capability but also reduce noise and allow a lower effective output impedance when driving capacitive loads.

Opamp-transistor hybrid amplifiers

When lower headphone impedances are to be driven, the number of multiple opamps required can become unwieldy, taking up an excessive amount of PCB area, and it becomes more economical to adopt a hybrid circuit (these are 'hybrid' circuits in that they combine IC opamps with discrete transistors; the name does not refer to thick-film construction or anything like that). The usual procedure is to add a Class-AB output stage after the opamp. A Class-A output stage is perfectly feasible if ultimate quality is required, though the power consumption is naturally much higher, and the efficiency with real signals something like 1%. The discrete devices could be either bipolar transistors or power FETs,

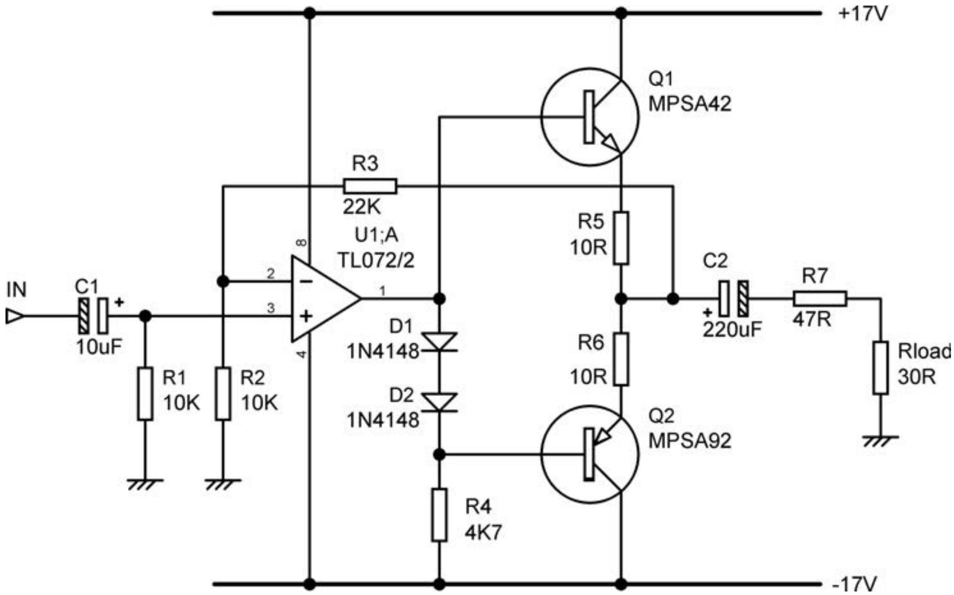

Figure 20.3: Traditional hybrid headphone amplifier combining a TL072 with a discrete Class-B output stage. Note the series output resistor R7, with typical value

but I invariably use bipolar transistors because their greater transconductance translates into better linearity, and they are totally predictable in every parameter that matters for this application.

I have always made it clear that Class-B and Class-AB in a power amplifier are not the same thing. Class-B is that unique quiescent condition where the distortion is a minimum. Class-AB is more highly biased so that it is effectively Class-A for small outputs, but above that extra distortion is generated as the output devices switch on and off. Here the situation is very different; instead of the driver-output device pairs of a power amplifier, there is a single transistor that turns on and off more gradually. There is no real distinction between Class-B and AB, and no need for critical biasing.

The arrangement in Figure 20.3 has been used for many years, dating back to when using a TL072 for the opamp made a significant cost saving. It is economical and dependable and I have used it to drive headphones in low-cost mixing consoles. The output stage is the simplest possible version of Class-AB, but it really works quite well, though its distortion characteristics, as seen in Figure 20.4, are not of the highest standard. The output devices are TO92 small-signal types with high beta, and this is crucial to an acceptable distortion performance; the bias is set by D1, D2 and there is no adjustment. The gain is 3.2 times. Inserting extra stages into opamp feedback loops must be done with

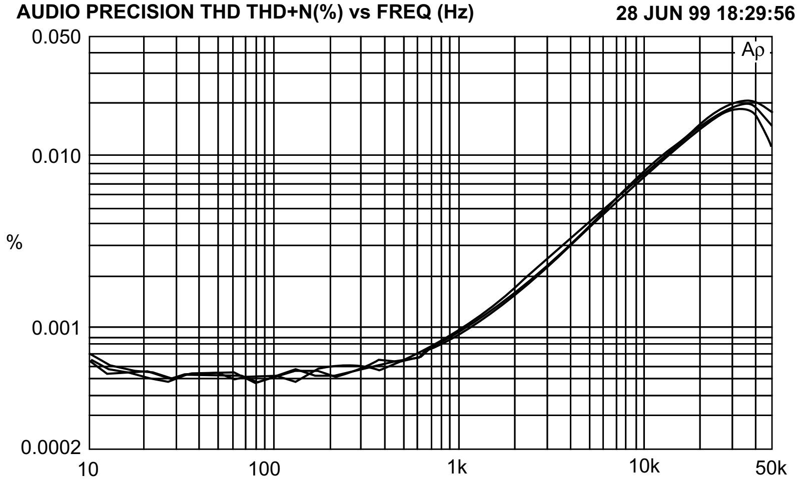

Figure 20.4: THD against load for TL072 with discrete Class-B output stage. No-load, 470R and 220R loading. Gain 3×, output level 7.75 Vrms (before o/p series resistor)

care, to avoid adding extra phase-shifts that may cause HF instability; here there are no problems, and no extra stabilising components are required. The output is AC coupled by C2 so that DC-offset protection is not required, and output resistor R7 takes care of short circuit protection, and caters for the different drive voltages required by 600 Ω and 30 Ω headphones.

Figure 20.5 shows a more sophisticated version, with lower distortion and greater output into lower impedances. I have deployed it in preamplifiers with a headphone output. There are three TO92 output pairs giving greater drive power; no heatsinks are required. The gain is higher than in Figure 20.3 at 5.4 times (+14.6 dB), and is provided by A1, allowing A2 to work at unity closed-loop gain so there is maximal negative feedback for correcting errors in the output stage. Bootstrapping is added to R4, R7 to maintain a more constant current through bias components D1, R5. C3 aids stability; at very high frequencies the NFB is taken from before the output stage and does therefore not suffer phase-shifts from it. This circuit is reliably HF stable with a 5532 for A2; an LM4562 in this position showed hints of instability but this may be curable by adjusting C3. The noise output was measured at −107.0 dBu, so the EIN is −107.0 −14.6 = −121.6 dBu, which is pleasingly low.

The distortion performance is shown in Figure 20.6, without use of a series output resistor. The 5 Vrms trace is higher due to the relatively higher noise level with a smaller output.

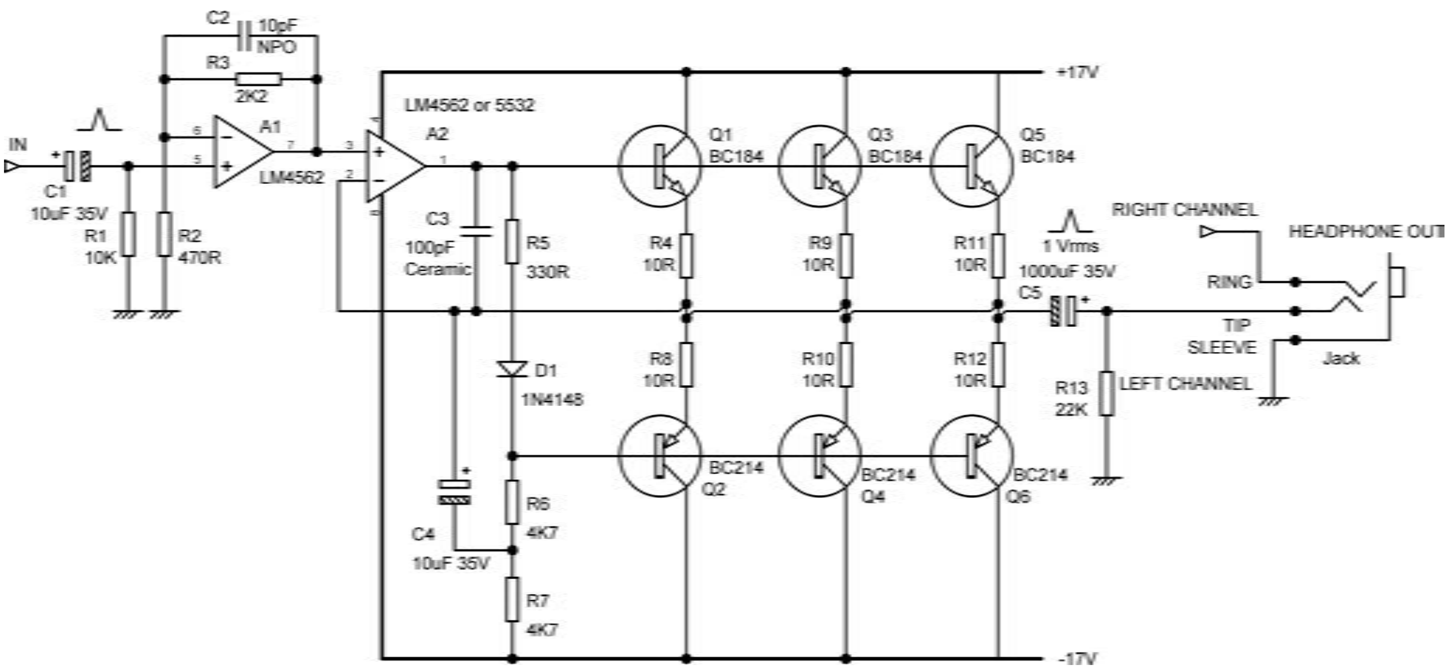

Figure 20.5: A more sophisticated hybrid headphone amplifier with the voltage gain provided by separate stage A1, and three output pairs

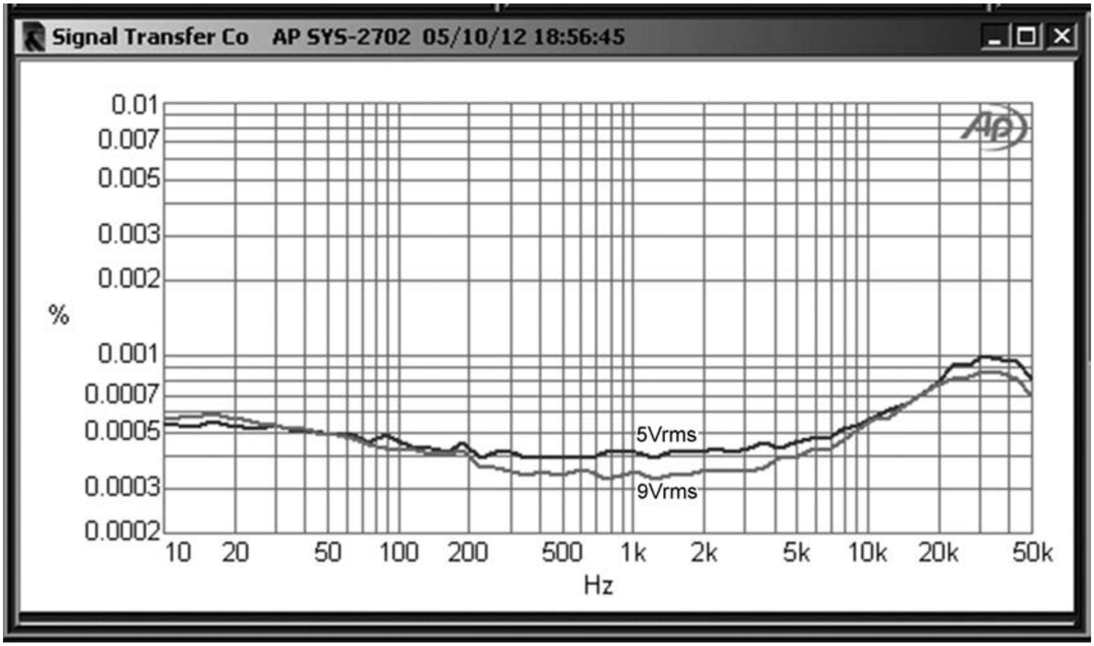

Figure 20.6: The hybrid headphone amplifier of Figure 20.5 driving a 60 Ω load. Upper trace 5 Vrms output, lower trace 9 Vrms output

Discrete Class-AB headphone amplifiers

The hybrid circuit of Figure 20.5 is relatively complicated, and one starts to wonder if a wholly discrete solution might be more economical. Figure 20.7 shows a discrete headphone amplifier based on the discrete opamp designs of Chapter 3. The output stage is a simple complementary Class-AB configuration with fixed bias.

First we will look at the basic linearity with light loading. With a 3 kΩ load, the distortion residual at 1 kHz looks nothing like the characteristic crossover distortion seen in a power amplifier. It is pure third harmonic. This situation continues all the way up to 20 kHz, the residual remaining as pure third harmonic but increasing in relative level.

The distortion with light loading can be reduced dramatically by increasing the quiescent current in the output stage, by adding a biasing resistor in series with the two diodes in the VAS collector circuit, as shown in Figure 20.7. Figure 20.8 demonstrates that HF distortion falls steadily as the value of the bias resistor increases, with no obvious sign of an optimal value. A 10 Ω resistor only gives a modest improvement but 27 Ω reduces HF distortion by a factor of 5, and 56 Ω seems to give most of the improvement that is attainable. The latter corresponds to a quiescent current of 4.6 mA in the output devices. While in this form it makes a very useful model amplifier for studying the small-signal stages of power amplifiers, its distortion increases rapidly when heavier loads are driven.

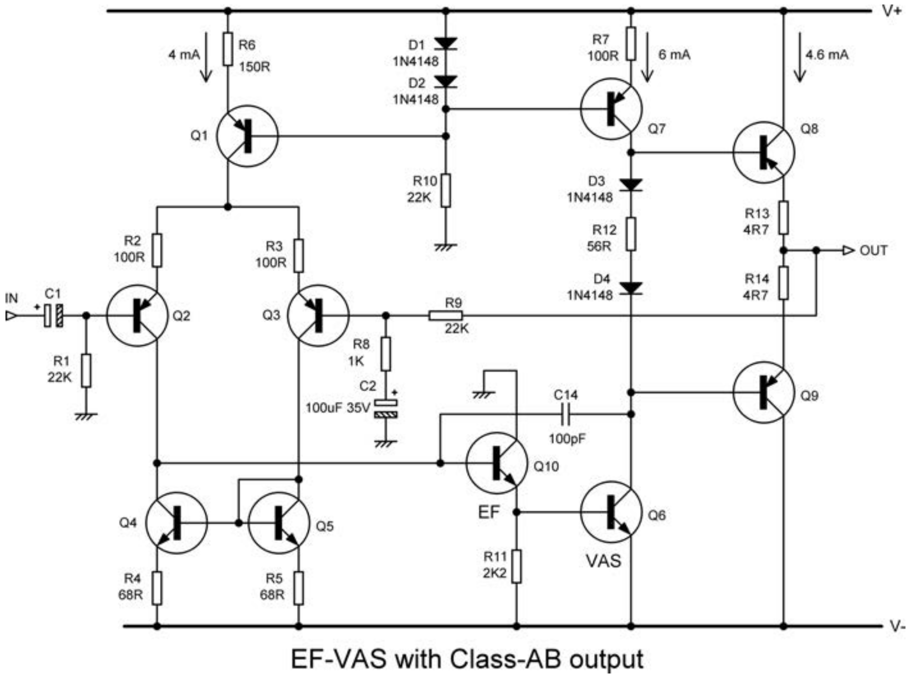

Figure 20.7: Discrete headphone amplifier with Class-AB output stage. Output devices are MJE340/350, all others are MPSA06/56

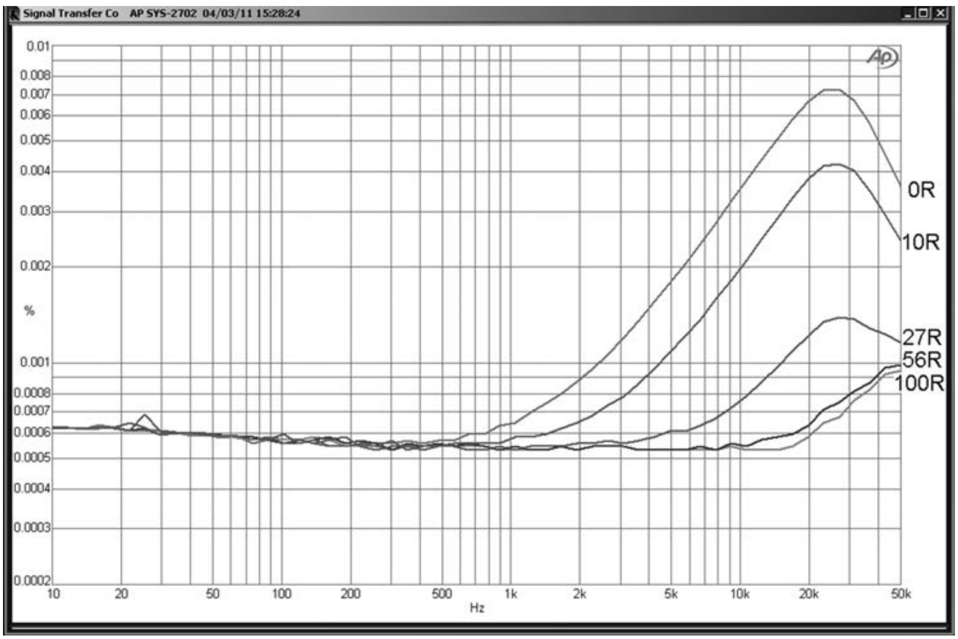

Figure 20.8: Effect of increasing output bias on discrete Class-AB amplifier with load of 3 kΩ. Added bias resistor 0R, (top trace) 10R, 27R, 56R, 100R, +20 dBu output, ±20 V rails

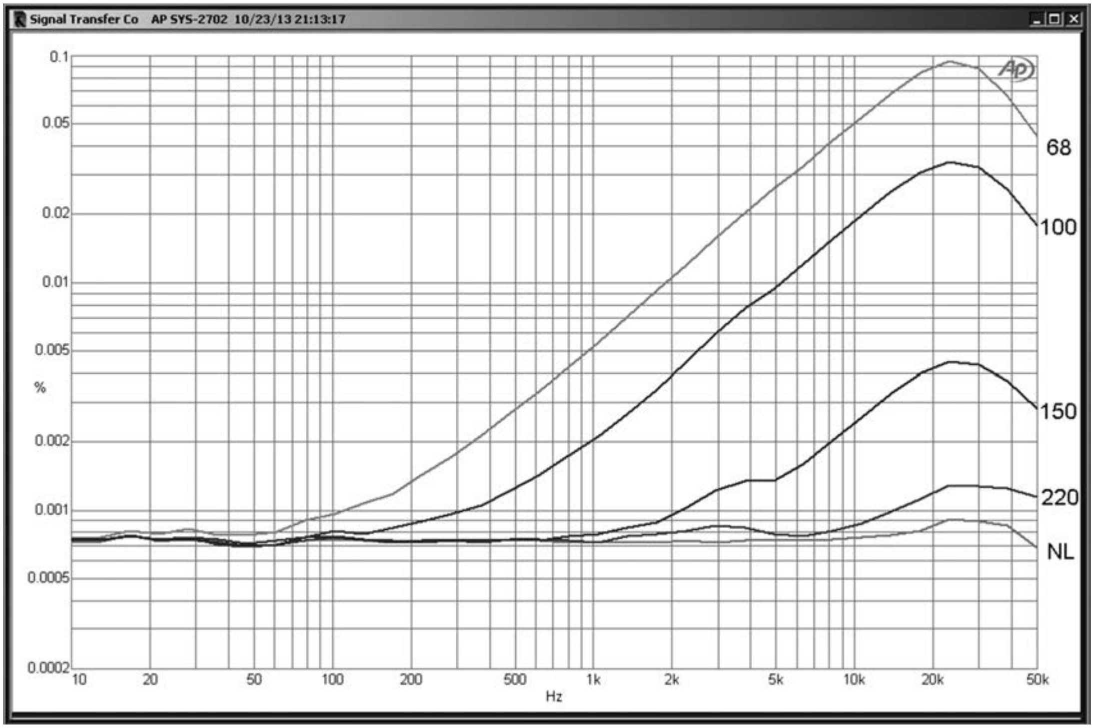

Figure 20.9: Distortion of discrete Class-AB amplifier into no load (NL), 220 Ω, 150 Ω, 100 Ω and 68 Ω, +20 dBu output, ±20 V rails

Figure 20.9 shows distortion remains low for a 220 Ω load, but increases rapidly for heavier loading as the output stage is taken out of its Class-A region, and the THD residual becomes a complex mixture of higher harmonics, increasing at 6 dB/octave with frequency. When comparing this with other designs, be aware that the closed-loop gain is higher at 23 times.

Discrete Class-A headphone amplifiers

Because the power requirements of headphone amplifiers are quite modest, it is quite practical to adopt a headphone amplifier design that works in pure Class-A. This assumes that push-pull Class-A is used to get the maximum efficiency. Even so, with real music signals the efficiency is likely to be as low as 1% at full volume, and even less at lower levels.

Figure 20.10 shows another development of the discrete opamp using a Class-A output stage. The high quiescent dissipation means that TO-220 output transistors must be used; since these have relatively low beta they are paired with TO-92 driver transistors, the result looking

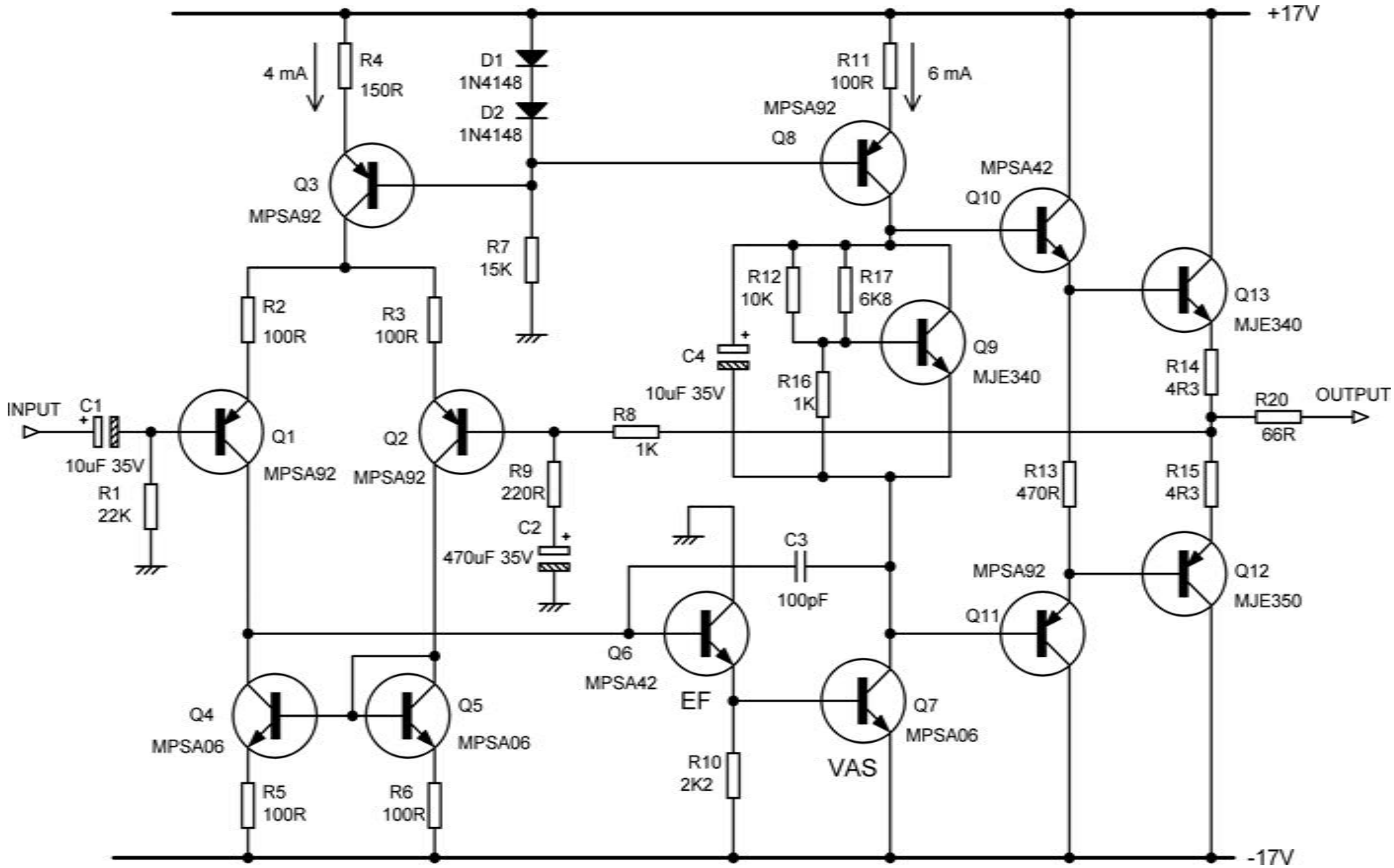

Figure 20.10: Class-A hybrid headphone amplifier with EF output stage

very like a full-scale power amplifier. The quiescent current is set at 90 mA by the bias generator Q9 and fixed resistors R12, R16, R17. No bias adjustment is required as in Class-A the quiescent current is not critical.

In the prototype a small vertical heatsink with a thermal resistance of 10 C/W (type SW38) was used for each output device, with the bias transistor Q9 mounted not just on the same heatsink, but actually on top of output device Q12 for the best thermal coupling. The standard clip can still be used to hold the devices to the heatsink. From cold turn-on to full operating temperature the quiescent remains within ±10%. A feedback control system could hold the quiescent current much more closely than this [4], but its complexity is hardly necessary at such modest powers. No short circuit protection is required because the output resistance R20, made up of two 33 Ω 750 mW resistors in series, can safely absorb the power if the output is shorted.

The distortion driving a 30 Ω load at 3 Vrms out is shown in Figure 20.11. The THD at 10 kHz is 0.0008%, comfortably below 0.001%, and predominantly second-harmonic. The maximum output is 3.3 Vrms, due to the effect of the output resistor.

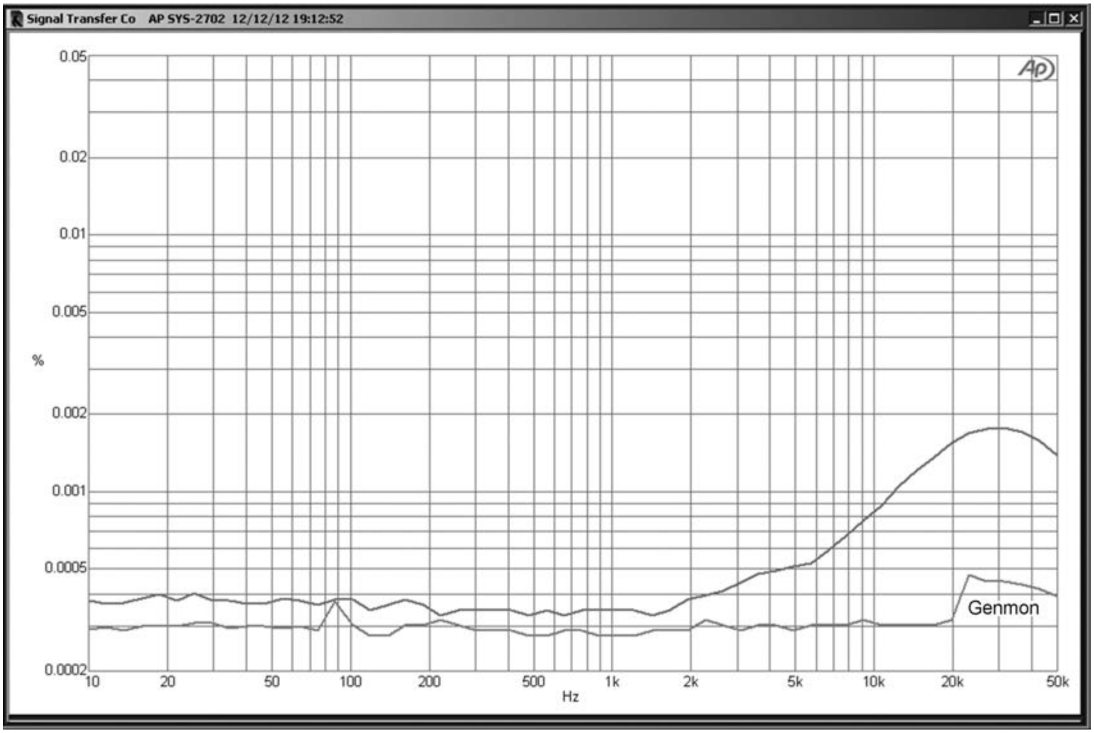

Figure 20.11: Class-A hybrid headphone amplifier distortion performance. The lower trace is the testgear residual

Because of the high performance of this relatively simple circuit, I think it is worthwhile to look at the measurements in more detail than usual:

Gain		
Measured gain	5.52 times	+14.83 dB
Measured gain	4.97 times	+13.93 dB (600 Ω load)
Measured gain	1.74 times	+4.81 dB (30 Ω load)

Maximum output	
10.5 Vrms	No load
9.4 Vrms	600 Ω load
3.3 Vrms	30 Ω load

Noise performance		
Noise measurements bandwidth 22 Hz–22 kHz, unweighted, RMS sensing, input terminated with 40 Ω.		
Noise out	−107.7 dBu	No output load
Equivalent input noise (EIN)	−122.5 dBu	
Noise out	−115.8 dBu	Output load 30 Ω
Equivalent input noise (EIN)	−120.6 dBu	
Noise out is lower with the 30 Ω load because, with the series output resistor, it forms an attenuator.		

Power consumption	
92 mA per amplifier from ±17 V rails	(3.2 Watts)

Balanced headphone amplifiers

It is believed by a small minority that a balanced drive to headphones can be of advantage. It is difficult to think of any technical reason why this might be so, because the moving coil speaker elements are not connected to anything except the connections feeding them, so the presence of a common-mode voltage is irrelevant.

Most headphones will need to be rewired for balanced use, as the typical arrangement for high-quality headphones is to have a 3-wire cable running from a ¼-inch jack plug to one ear-piece, with a 2-wire cable running from there to the other ear-piece. This means that part of the ground connection is common to the left and right signals, and since it has some resistance it will cause interchannel crosstalk. Paradoxically, cheap headphones with a figure-of-eight cable that splits off to each ear-piece may be better in this respect because there are separate grounds to each ear-piece, and these only meet up at the jack plug, so there is the minimum amount of common ground resistance.

Four-way jack plugs do exist (with tip, two rings, and sleeve) but connection to headphones wired for balanced use is normally made with two 3-pin XLR connectors.

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CHAPTER 21

Signal switching

The switching and routing of analogue signals is a fundamental part of signal processing, but not one that is easily implemented if accuracy and precision are required. This article focuses on audio applications, but the basic parameters such as isolation and linearity are equally relevant in many fields.

Mechanical switches

A mechanical switch normally makes a solid unequivocal connection when it is closed, and it is as 'on' as the resistance of its contacts and connections allows; these are small fractions of an Ohm and are unlikely to cause trouble in small-signal audio design. Switches are, however, in general terms a good deal less 'off'. The insulation resistance may be measured in peta-ohms, but what does the damage is the inevitable capacitance between contacts. This is usually small in pF, but quite large enough to dominate the degree of offness obtainable at high audio frequencies. Its effects naturally depend on the impedance at the 'receiving' side of the switch. For all the tests discussed here this was 10 kΩ.

Using an ALPS SPUN type push switch, at 10 kHz the offness is only −66 dB, and grounding the unused side of the switch only improves the offness by about 2 dB. A graph of the result can be seen in Chapter 22; the offness naturally degrades by 6 dB/octave. Switch capacitance is an important issue in designing mixer routing systems.

In another test a miniature 3-way slide switch gave −70 dB at 10 kHz. Once again, grounding the unused contact at the end of the switch only gave a 4 dB improvement, and it is wise to assume that in general grounding unused switch sections will not help much.

Switch inter-contact capacitance is quite easy to determine; measure the offness, i.e. the loss of the RC circuit, and since R is known C can be calculated easily. Once it is known for a given switch construction, it is easy to calculate the offness for different loading resistances. Interestingly, the inter-contact capacitance of switches seems to be relatively constant, even though they vary widely in size and construction. This seems to be because the smaller switches have smaller contacts, with a smaller area, but on the other hand they are closer together.

Input-select switching

Some time ago, Morgan Jones [1] raised the excellent point of crosstalk in the input-select switching of preamplifiers. If the source impedance is significant, then this may be a serious problem. While I agree that his use of a rotary switch with twice the required number of positions and grounding alternate contacts is slightly superior to the conventional use of rotary switches, measuring a popular Lorlin switch type showed the improvement to be only 5 dB. I am also unhappy with all those redundant 'mute' positions between input selections, so when I design a preamp I normally choose interlocked push-switches rather than a rotary switch. A 4-changeover format can then be used to reduce crosstalk.

The problem with conventional input select systems like Figure 21.1a is that the various input tracks necessarily come into close proximity, with significant crosstalk through capacitance

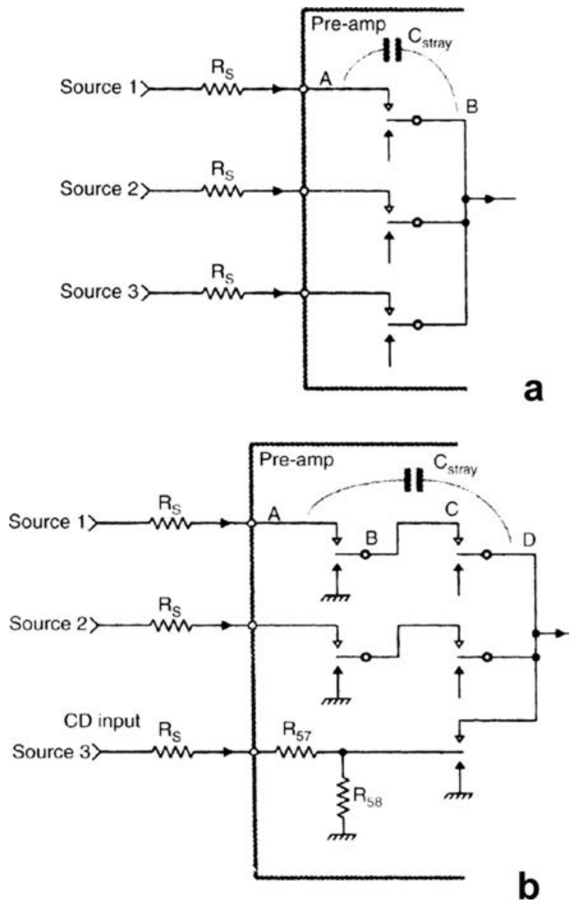

Figure 21.1: a) Two-changeover select switches give poor signal rejection due to switch capacitance C_{stray}, b) using four-changeover switches improves offness by 21 dB at 10 kHz. Note CD input attenuator

TABLE 21.1 Offness of various switch configurations (10 kHz)

	(dB)
Simple rotary switch	−71
Rotary with alternate contacts grounded	−76
Two changeover switch	−74
Four changeover switch	−95

C_{stray} to the common side of the switch, i.e. from A to B. Using two changeovers per input side (i.e. four for stereo) allows the intermediate connection B-C to be grounded by the NC contact of the first switch section, and keeps the 'hot' input A much further away from the common input line D, as shown in Figure 21.1b. C_{stray} is now much smaller.

The crosstalk data in Table 21.1 was gathered at 10 kHz, with 10 KΩ loading resistances:

The emphasis here is on minimising crosstalk between different sources carrying different signals, as interchannel (L–R) crosstalk is benign by comparison. Interchannel isolation is limited by the placement of left and right channels on the same switch, with the contact rows parallel, and limits L–R isolation to −66 dB at 10 kHz with a high 10 kΩ source impedance. Actual source impedances are likely to be lower, with both inter-source and inter-channel crosstalk proportionally reduced; so a more probable 1 kΩ source gives 115 dB of intersource rejection at 10 kHz for the 4-changeover configuration.

The third input of Figure 21.1b has a resistive attenuator intended to bring CD outputs down to the same level as other sources. In this case inter-source crosstalk can be improved simply by back-grounding the attenuator output when it is not in use, so only a 2-pole switch is required for good isolation of this input.

The Virtual Contact

What do you do if you need a changeover switch – to select one of two signal sources – but only have a make contact? Here is a technique that can be a lifesaver when you have screwed up on ordering a switch with a long leadtime, or if you live in a Dilbertian world of last-minute spec changes.

Figure 21.2 demonstrates the principle. With the switch S open, source A goes through voltage-follower A and R_{feed} to the output voltage-follower. With the switch closed, the much lower impedance output of voltage-follower B takes over and the contribution from A is now negligible. To give good rejection of A, the output impedance of follower B must be much, much lower than the value of R_{feed}; so an opamp output must be used directly. Figure 21.3 shows how good the rejection of A can be using 5532 opamps as the voltage-followers.

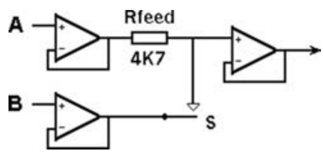

Figure 21.2: The virtual contact concept. When the switch S is closed the signal B overrides signal A

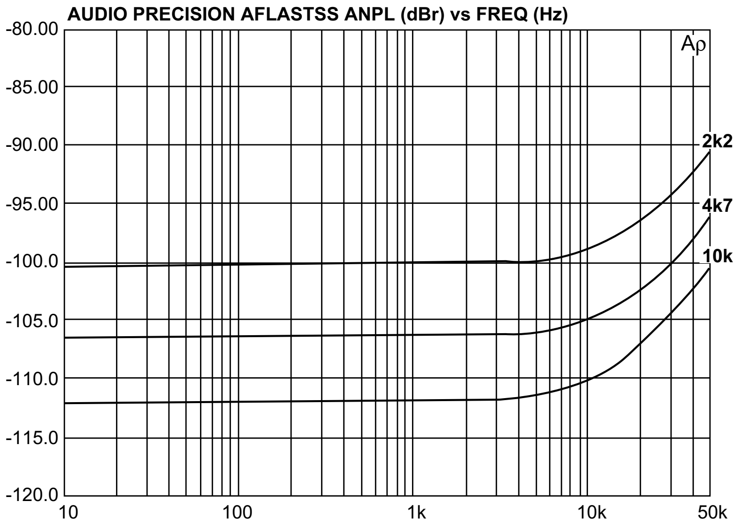

Figure 21.3: Rejection of signal A for 2k2, 4k7 and 10 kΩ Rfeed resistors, using 5532 opamps

At first this technique looks a bit opamp-intensive. However, there is often no need to use dedicated voltage-followers if a similar low-impedance feed is available from a previous stage that uses an opamp with a large amount of negative feedback. Likewise, the output voltage-follower may often be dispensed with if the following load is reasonably high.

There is also the rejection of B when the switch is open to consider. The impedance of R_{feed} mean there is the potential for capacitative crosstalk across the open switch contacts. The amount depends on the value of R_{feed} and on switch construction.

If the offness of B is more important than the offness of A, then R_{feed} should be a lower value, to minimise the effects of the capacitance. Do not make R_{feed} too low as A drives through it into effectively a short circuit when B is selected.

If the offness of A is more important, R_{feed} should be higher to increase its ratio to the output impedance of B; be aware that making it too high may introduce excessive Johnson noise.

The rejection of A shown in Figure 21.3 worsens at high frequencies, as the dominant-pole of opamp B reduces its open-loop gain and the output impedance rises. The slopes are 6 dB/octave as usual.

This technique is particularly useful for switching between three sources with a centre-off toggle switch.

Relay switching

Any electronic switching technique must face comparison with relays, which are still very much with us. Relays give total galvanic isolation between control and signal, zero contact distortion, and in audio terms have virtually unlimited signal-handling capability. They introduce negligible series resistance and shunt leakage to ground is usually also negligible. Signal offness can be very good, but as with other kinds of switching, this depends on intelligent usage. There will always be capacitance between open contacts, and if signal is allowed to crosstalk through this to nominally off circuitry, the 'offness' will be no better than other kinds of switching.

Obviously relays have their disadvantages. They are relatively big, expensive, and not always as reliable as more than a hundred years of development should have made them. Their operating power is significant, though it can be reduced by circuitry that applies full voltage to pull in the relay and then a lower voltage to keep it closed. Some kinds of power relay can introduce disastrous distortion if used for switching audio because the signal passes through the magnetic soft-iron frame; however, such problems are likely to be confined to the output circuits of large power amplifiers. For small-signal switching the linearity of relay contacts can normally be regarded as perfect.

Electronic switching

Electronic switching is usually implemented with CMOS analogue gates, of which the well-known 4016 is the most common example, and these are examined first. However, there are many special applications where discrete JFETs provide a better solution, so these are dealt with in the second part.

Switching with CMOS analogue gates

CMOS analogue gates, also known as transmission gates, are quite different from the CMOS logic gates in the 4000 series, though the underlying process technology is the same. Analogue gates are bilateral, which means that either of the in/out leads can be the input or output; this is emphatically not true for logic gates. The 'analogue' part of the name emphasises that they are not restricted to fixed logic levels, but pass whatever signal they are given with low distortion. The 'low' there requires a bit of qualification, as will be seen later.

When switched on, the connection between the two pins is a resistance which passes current in each direction as usual, depending on the voltage between the two terminals. Analogue gates have been around for a long time, and are in some ways the obvious method of electronic switching. They do however have significant drawbacks.

Analogue gates like the 4016 are made up of two MOSFETs of opposite polarity connected back to back. The internal structure of a 4016 analogue gate is shown in Figure 21.4. The two transmission FETs with their protective diodes are shown on the right; on the left is the control circuitry. A and B are standard CMOS inverters whose only function is to sharpen up the rather soggy voltage levels that 4000-series CMOS logic sometimes provides. The output of B directly controls one FET, and inverter C develops the anti-phase control voltage for the FET of opposite polarity, which requires an inverted gate voltage to turn it on or off.

MOSFETS are of the enhancement type, requiring a voltage to be applied to the gate to turn them on (in contrast JFETs work in depletion mode and require a gate voltage to turn them *off*) so as the channel approaches the gate voltage, the device turns off more. An analogue gate with only one polarity of FET would be of little use because R_{on} would become very high at one extreme of the voltage range. This is why complementary FETs are used; as one polarity finds its gate voltage decreasing, turning it off, the other polarity has its gate voltage increasing, turning it more on. It would be nice if this process cancelled out so the R_{on} was constant, but sadly it just doesn't work that way. Figure 21.5 shows how R_{on} varies with input voltage, and the peaky curve gives a strong hint that something is turning on as something else turns off.

Figure 21.5 also shows that R_{on} is lower and varies less when the higher supply voltage is used; since these are enhancement FETs the on-resistance decreases as the available

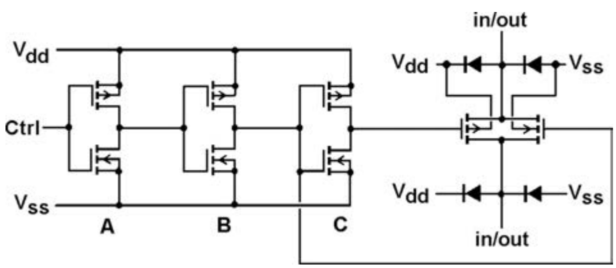

Figure 21.4: The internal circuitry of a 4000-series analogue gate

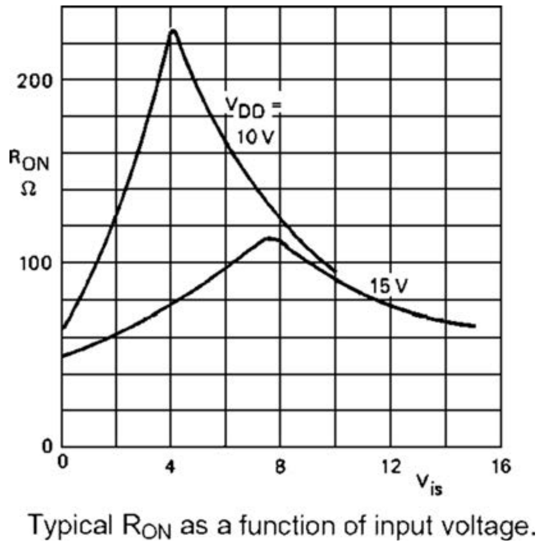

Figure 21.5: Typical variation of the gate series resistance RON for the 4016

control voltage increases. If you want the best linearity then always use the maximum rated supply voltage.

Since R_{on} is not very linear, the smaller its value the better. The 4016 R_{on} is specified as 115 Ω typical, 350 Ω max, over the range of input voltages and with a 15 V supply. The 4066 is a version of the 4016 with lower R_{on}, 60 Ω typical, 175 Ω max under the same conditions. This option can be very useful both in reducing distortion and improving offness, and in most cases there is no point in using the 4016. The performance figures given below all assume the use of the 4066 except where stated.

CMOS gates in voltage mode

Figure 21.6 shows the simplest and most obvious way of switching audio on and off with CMOS analogue gates. This series configuration is in a sense the 'official' way of using them; the only snag being that it doesn't work very well.

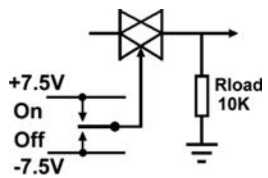

Figure 21.6: Voltage-mode series switching circuit using analogue gate

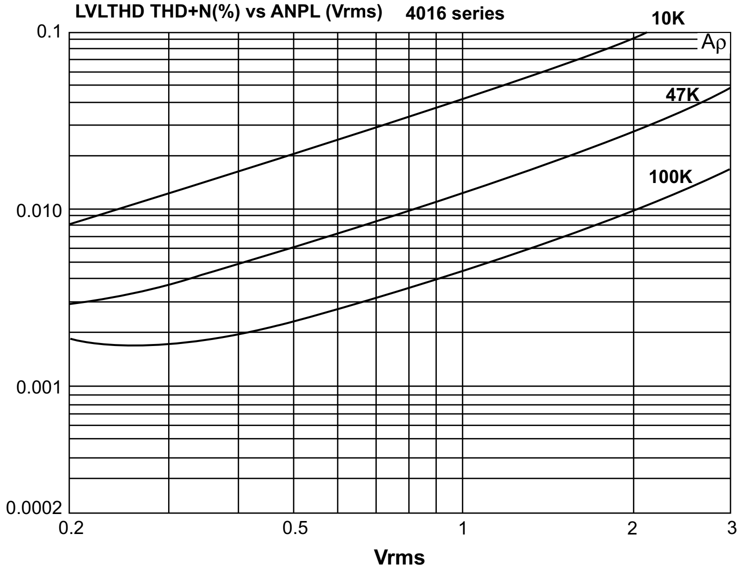

Figure 21.7: 4016 series-gate THD versus level, with different load resistances

Figure 21.7 shows the measured distortion performance of the simple series gate using the 4016 type. The distortion performance is a long way from brilliant, exceeding 0.1% just above 2 Vrms. These tests, like most in this section, display the results for a single sample of the semiconductor in question. Care has been taken to make these representative, but there will inevitably be some small variation in parameters like R_{on}. This may be greater when comparing the theoretically identical products of different manufacturers.

Replacing the 4016 gate with a 4066 gives a reliable improvement due to the lower R_{on}. THD at 2 Vrms (10 kΩ load) has dropped to a third of its previous level (see Figure 21.8). There seems to be no downside to using 4066 gates instead of the more common and better-known 4016, and they are used exclusively from this point on, unless otherwise stated. Likewise, using multiple gates in parallel reduces distortion; see Figure 21.9.

The distortion is fairly pure second harmonic, except at the highest signal levels where higher-order harmonics begin to intrude. This is shown in Figures 21.8 and 21.9 by the straight line plots beginning to bend upwards above 2 Vrms.

Analogue gate distortion is flat with frequency as far as audio is concerned, so no plots of THD versus frequency are shown; they would just be a rather uninteresting set of horizontal lines.

This circuit gives poor offness when off, as shown by Figure 21.10. The offness is limited by the stray capacitance in the package feeding through into the relatively high load impedance.

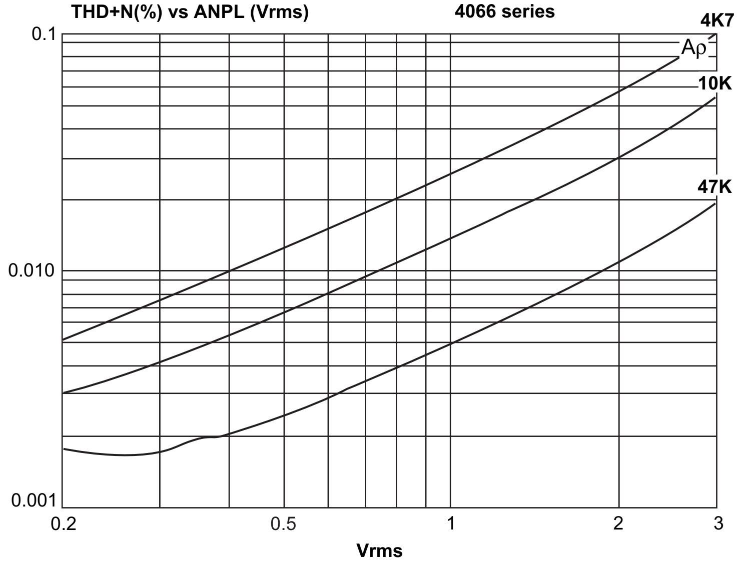

Figure 21.8: 4066 THD versus level, with different load resistances

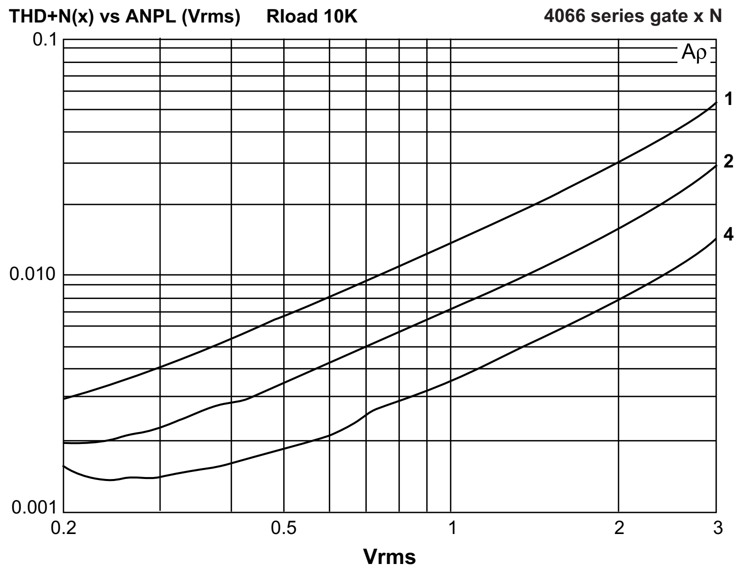

Figure 21.9: THD versus level, for different numbers of paralleled 4066 gates

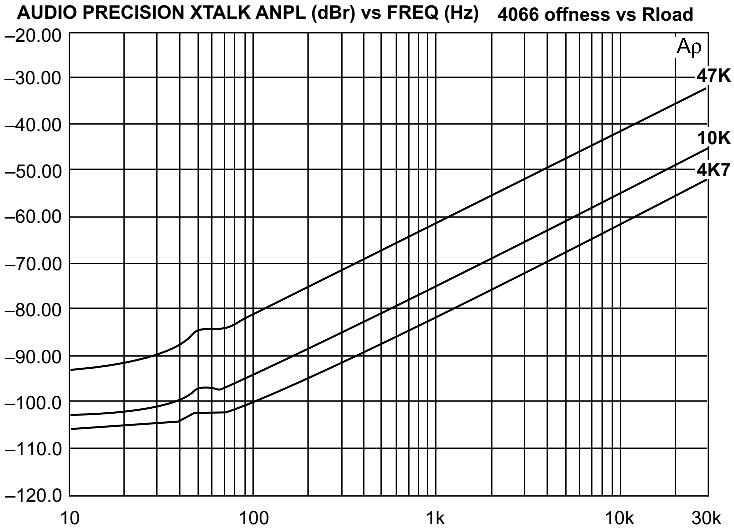

Figure 21.10: 4066 offness versus load resistance. −48 dB at 20 kHz with a 10 kΩ load

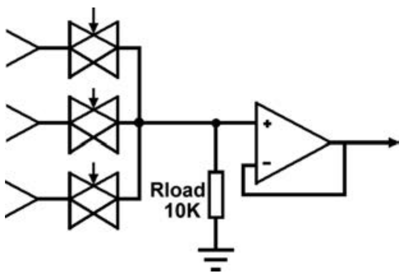

Figure 21.11: A one-pole, three way switch made from analogue gates

If this is 10 kΩ the offness is only −48 dB at 20 kHz, which would be quite inadequate for most applications. The load impedance could be reduced below 10 kΩ to improve offness – for example, 4k7 offers about a 7 dB improvement – but this degrades the distortion, which is already poor at 0.055% for 3 Vrms, to 0.10%.

Using 4066 gates instead of 4016s does not improve offness in this configuration. The internal capacitance that allows signals to leak past the gate seems to be the same for both types. The maximum signal level that can be passed through (or stopped) is limited by the CMOS supply rails and conduction of the protection diodes. While it would in some cases be possible to contrive a bootstrapped supply to remove this limitation, it is probably not a good route to head down.

Figure 21.11 shows a CMOS three-way switch. When analogue gates are used as a multi-way switch, the offness problem is much reduced, because capacitative feedthrough of the unwanted inputs is attenuated by the low R_{on} looking back into the (hopefully) low impedance of the active input, such as an opamp output. If this is not the case then the crosstalk from nominally off inputs can be serious.

In this circuit the basic poor linearity is unchanged, but since the crosstalk problem is much less, there is often scope for increasing the load impedance to improve linearity. This makes R_{on} a smaller proportion of the total resistance. The control voltages must be managed so that only one gate is on at a time, so there is no possibility of connecting two opamp outputs together.

It may appear that if you are implementing a true changeover switch, which always has one input on, the resistor to ground is redundant, and just a cause of distortion. Omitting it is however very risky, because if all CMOS gates are off together even for an instant, there is no DC path to the opamp input and it will register its displeasure by snapping its output to one of the rails. This does not sound nice.

Figure 21.12 shows the offness of a changeover system, for two types of FET-input opamps. The offness is much improved to −87 dB at 20 kHz, an improvement of 40 dB over the simple series switch; at the high-frequency end however it still degrades at the same rate of 6 dB/octave. It is well-known that the output impedance of an opamp with negative feedback increases with frequency at this rate, as the amount of internal gain falls, and this effect is an immediate suspect. However, there is actually no detectable signal on the opamp output (as shown by the lowest trace) and is also not very likely that two completely different opamps would have exactly the same output impedance. I was prepared for a subtle effect, but the

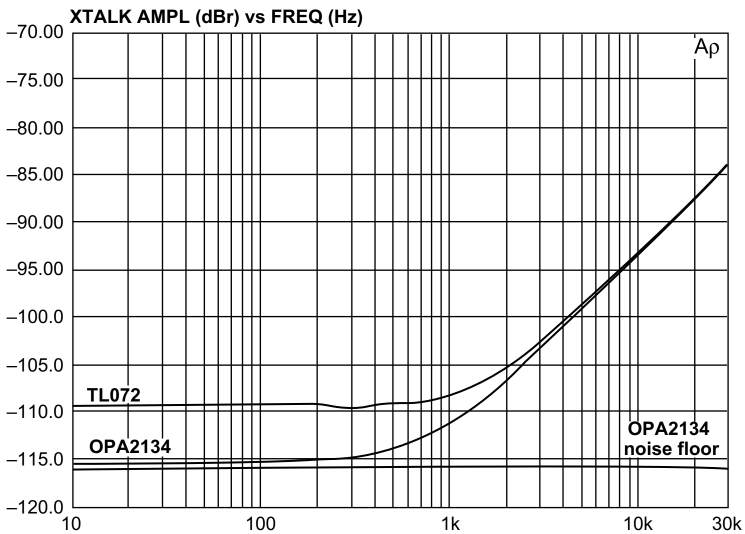

Figure 21.12: Voltage-mode changeover circuit offness for TL072 and OPA2134. 10 kΩ load

true explanation is that the falling offness is simply due to feedthrough via the internal capacitance of the analogue gate.

It now remains to explain why the OPA2134 apparently gives better offness in the flat low-frequency region. In fact it does not; the flat parts of the trace represent the noise floor for that particular opamp. The OPA2134 is a more sophisticated and quieter device than the TL072, and this is reflected in the lower noise floor.

There are two linearity problems. Firstly, the on-resistance itself is not totally linear. Secondly, and more serious, the on-resistance is modulated when the gates move up and down with respect to their fixed control voltages.

It will by now probably have occurred to most readers that an on/off switch with good offness can be made by making a changeover switch with one input grounded. This is quite true, but since much better distortion performance can be obtained by using the same approach in current mode, as explained below, I am not considering it further here.

Figure 21.13 shows a shunt muting circuit. This gives no distortion in the ON state because the signal is no longer going through the R_{on} of a gate. However the offness is limited by the R_{on}, forming a potential divider with the series resistor R; the latter cannot be very high in value or the circuit noise will be degraded. There is however the advantage that the offness plot is completely flat with frequency. Note that the ON and OFF states of the control voltage are now inverted.

Table 21.2 gives the measured results for the circuit, using the 4066. The offness can be improved by putting two or more of these gates in parallel, but since doubling the number N only gives 6 dB improvement, it is rarely useful to press this approach beyond four gates.

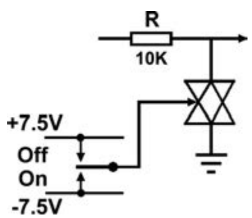

Figure 21.13: Voltage-mode shunt CMOS muting circuit

TABLE 21.2 Offness versus number of shunt 4066 analogue gates used, with R = 10kΩ

N gates	Offness (dB)
1	−37
2	−43
4	−49

CMOS gates in current mode

Using these gates in current mode – usually by defining the current through the gate with an input resistor and dropping it into the virtual-earth input of a shunt-feedback amplifier – gives much superior linearity. It removes the modulation of channel resistance as the gate goes up and down with respect to its supply rails and, in its more sophisticated forms, can also remove the signal voltage limit and improve offness.

Figure 21.14 shows the simplest version of a current-mode on/off switch. An important design decision is the value of R_{in} and R_{nfb}, which are often equal to give unity gain. Too low a value increases the effect of the non-linear R_{on}, while too high a value degrades offness, as it makes the gate stray capacitance more significant, and also increases Johnson noise. In most cases 22 KΩ is a good compromise.

Table 21.3 gives the distortion for +20 dBu (7.75 Vrms) in/out, and shows that it is now very low compared with voltage-mode switchers working at much lower signal levels; compare the table data with Figures 21.8 and 21.9 above. The increase in THD at high frequencies is due to a contribution from the opamp. However, the offness is pretty poor, and would not be acceptable for most applications. The problem is that with the gate off, the full signal voltage appears at the gate input and crosstalks to the summing node through the package's internal capacitance. In practical double-sided PCB layouts the inter-track capacitance can usually be kept very low by suitable layout, but the internal capacitance of the gate is inescapable.

In Figures 21.14 and 21.15, the CMOS gate is powered from a maximum of ±7.5 V. This means that in Figure 21.14, signal breakthrough begins at an input of 5.1 Vrms. This is

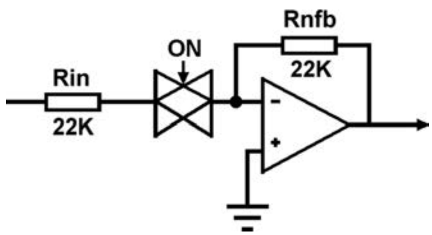

Figure 21.14: The simplest version of a current-mode on/off switch

TABLE 21.3 Distortion produced by a current-mode switch using 4016 gates, showing the gate contribution is small

	1 kHz	10 kHz	20 kHz
THD, via 4016, +20 dBu	0.0025%	0.0039%	0.0048%
THD, 4016 shorted, +20 dBu	0.0020%	0.0036%	0.0047%
Offness	−68 dB	−48 dB	−42 dB

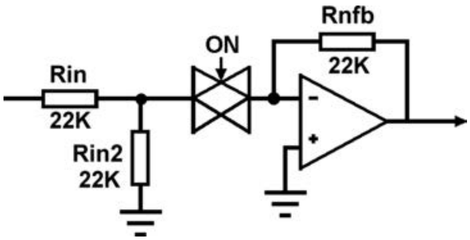

Figure 21.15: Current-mode switch circuit with breakthrough prevention resistor R_{in2}

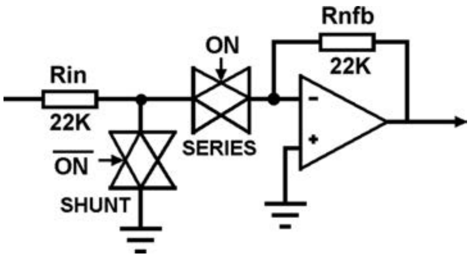

Figure 21.16: A series-shunt current-mode switch

much too low for opamps running off their normal rail voltages, and several dB of headroom are lost.

Figure 21.15 shows a partial cure for this. Resistor R_{in2} is added to attenuate the input signal when the CMOS gate is off, preventing breakthrough. There is no effect on signal gain when the gate is on, but the presence of R_{in2} does increase the noise gain of the stage.

As with all shunt-feedback stages, this circuit introduces a phase-inversion, which is sometimes convenient, but usually not.

CMOS series-shunt current mode

We now extravagantly use two 4016 CMOS gates, as shown in Figure 21.16.

When the switch is on, the series gate passes the signal through as before; the shunt gate is off and has no effect. When the switch is off the series gate is off and the shunt gate is on, sending almost all the signal at A to ground so that the remaining voltage is very small. The exact value depends on the 4016 specimen and its R_{on} value, but is about 42 dB below the input voltage. This deals with the offness (by greatly reducing the signal that can crosstalk through the internal capacitance) and also increases the headroom by several dB, as there is now effectively no voltage signal to breakthrough when it exceeds the rails of the series gate.

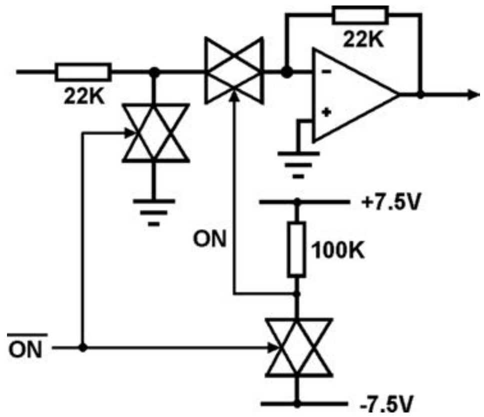

Figure 21.17: Generating antiphase control signals with a spare analogue gate

TABLE 21.4 Distortion levels with series-shunt switching

	1 kHz	10 kHz	20 kHz
THD, via 4016 × 1, +20 dBu	0.0016%	0.0026%	0.0035%
THD, via 4016 × 2, +20 dBu	0.0013%	0.0021%	0.0034%
THD, 4016 shorted, +20 dBu	0.0013%	0.0021%	0.0034%
Offness 4016 × 1	−109 dB	−91 dB	−86 dB
Offness 4016 × 1, J111	< −116 dB	−108 dB	−102 dB

Two antiphase control signals are now required. If you have a spare analogue gate it can generate the inverted control signal, as shown in Figure 21.17.

The distortion generated by this circuit can be usefully reduced by using two gates in parallel for the series switching, as in Table 21.4; this gate-doubling reduces the ratio of the variable R_{on} to the fixed series resistor and so improves the linearity. Using two in parallel is sufficient to render the distortion negligible (the higher distortion figures at 10 kHz and 20 kHz are due to distortion generated by the TL072 opamp used in the measurements).

As before the input and output levels are +20 dBu, well above the nominal signal levels expected in opamp circuitry; measurements taken at more realistic levels would show only noise.

Discrete FETs have lower R_{on} than analogue gates. If a J111 JFET is used as the shunt switching element the residual signal at A is further reduced, to about 60 dB below the input level, with a consequent improvement in offness, demonstrated by the bottom row in Table 21.4. This could also be accomplished by using two or more CMOS gates for the shunt switching.

Control voltage feedthrough in CMOS gates

When an analogue gate changes state, some energy from the control voltage passes into the audio path via the gate-channel capacitance of the switching FETs, through internal package capacitances, and through any stray capacitance designed into the PCB. Since the control voltages of analogue gates move snappily, due to the internal inverters, this typically puts a click rather than a thump into the audio. Attempts to slow down the control voltage going into the chip with RC networks are not likely to be successful for this reason. In any case, slowing down the control voltage change simply converts a click to a thump; the FET gates are moving through the same voltage range, and the feedthrough capacitance has not altered, so the same amount of electric charge has been transferred to the audio path – it just gets there more slowly.

The only certain way to reduce the effect of transient feedthrough is to soak it up in a lower value of load resistor. The same electric charge is applied to a lower resistor value (the feedthrough capacitance is tiny, and controls the circuit impedance) so a lower voltage appears. Unfortunately reducing the load tends to increase the distortion, as we have already seen; the question is if this is acceptable in the intended application.

CMOS gates at higher voltages

Analogue gates of the 4016/4066 series have a voltage range of ±7.5 V, so they cannot directly switch signals from opamps running from the usual ±15 V or ±17 V rails. The classic device for solving this problem is the DG308 from Maxim. The absolute maximum supply rails are ±22 V, and the maximum signal range is specified as ±15 V. The R_{on} is typically between 60 and 95 Ω with ±15 V supply rails, comparable with a 4016 on ±7.5 V rails. The DG308 is not pin-compatible with the 4016/4066.The DG308 has been around for a long time, and it has always been significantly more expensive than 4016/4066. That situation continues today.

CMOS gates at low voltages

So far we have generally assumed that the CMOS analogue gates will be run from the maximum rated rail voltages of ±7.5 V, to maximise the linearity of types such as the 4016/4066. As noted in Chapter 5, there is nowadays much interest in designing audio paths that can give decent quality when run from +5 V, the power typically being drawn from a USB port or similar digital source; what I call Five Volt Fidelity. The 4016/4066 gates work poorly under these conditions; the supply voltage is reduced to a third of the maximum, and this gives very bad linearity.

The answer is to use something designed for the job. The DG9424 from Vishay-Siliconix is rated for maximum supplies of ±6 V, and works well running from a single 5 V supply. In this case its typical R_{on} is 3.4 Ω, which is much superior to the range of R_{on} shown for the 4016

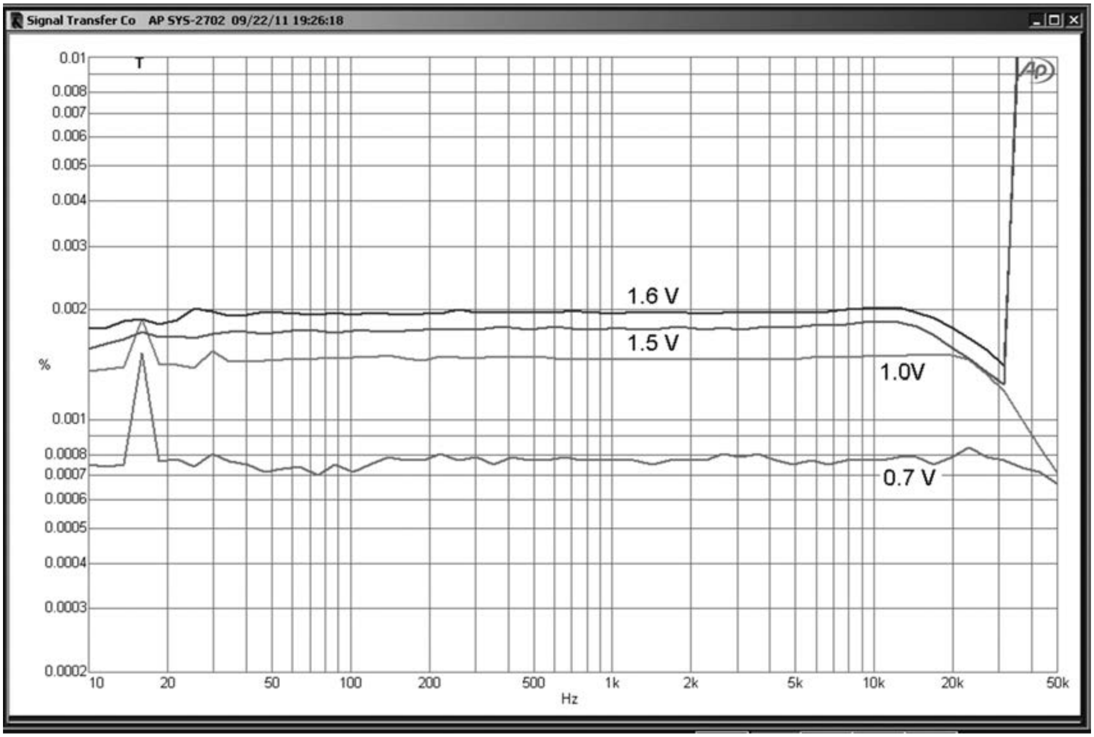

Figure 21.18: Distortion from DG9424 analogue gates powered from +5 V and driving a 10 kΩ load at various levels

in Figure 21.5. Once again a higher supply voltage means a lower R_{on} and less distortion; running the DG9424 from ±5 V reduces the typical R_{on} to 2 Ω, and using the maximum ±6 V rails reduces it further to typically 1.8 Ω. This is a considerable advance on the older parts.

Figure 21.18 shows the distortion in series mode at various signal levels when driving a 10 kΩ load. When making the measurements I found that for outputs greater than 1.5 V and above 32 kHz the analogue switches entered a high-distortion mode that gave about 0.06% THD. Whether that is a typical finding I am not sure, but it is unlikely to cause problems unless you are handling high-level ultrasonic signals. I also found that distortion was at a minimum for a single +8 V supply, which does not fit in well with the official R_{on} specs; the reason for this is currently unknown. The DG9424 is pin-compatible with the DG308.

CMOS gate costs

Table 21.5 shows the prices for SMD format in small quantities at June 2013. The prices will obviously be much lower for quantity production, but the ratios between them should be roughly the same.

TABLE 21.5 Analogue gate prices

Analogue gate	Price each (GBP)
4016	0.49
4066	0.215
DG308	1.32
DG9424	1.52

You can see that the DG308 is a good deal more costly than the humbler 4016. The DG9424, being something of a specialised part, is more expensive again. The surprise is that the 4016 comes out as more than twice as costly as the 4066. The only explanation I can think of is that, as mentioned earlier, the 4066 always works better and there seems no reason why the 4016 should ever be used. Perhaps that means they are manufactured in smaller quantities and are therefore pricier.

Discrete JFET switching

Having looked in detail at analogue switching using CMOS gates, and having seen how well they can be made to work, you might be puzzled as to why anyone should wish to perform the same function with discrete JFETs. There are at least two advantages in particular applications.

Firstly, JFETs can handle the full output range of opamps working from maximum supply rails, so higher signal levels can often be switched directly without requiring opamps to convert between current and voltage mode.

Secondly, the direct access to the device gate allows relatively slow changes in attenuation (though still measured in milliseconds, for reasons that will emerge) rather than the rapid on-off action which CMOS gates give as a result of their internal control-voltage circuitry. This is vital in creating mute circuits that essentially implement a fast fade rather than a sharp cut, and so do not generate clicks and thumps by abruptly interrupting the signal. The downside is that they require carefully-tailored voltages to drive the gates, and these cannot always be conveniently derived from the usual opamp supply rails.

The series JFET switch in voltage mode

The basic JFET series switching circuit is shown in Figure 21.19. With the switch open there is no other connection to the gate other than the bootstrap resistor R_{boot}, V_{gs} is zero, and so the FET is on. When the switch is closed, the gate is pulled down to a sufficiently negative voltage to ensure that the FET is biased off even when the input signal is at its negative limit.

TABLE 21.6 Characteristics of the J111 JFET series

	J111	J112	J113
$V_{gs(off)}$ min	−3.0 V	−1.0 V	−0.5 V
$V_{gs(off)}$ max	−10 V	−5.0 V	−3.0 V
$R_{ds(on)}$	30	50	100

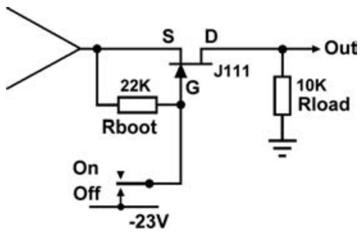

Figure 21.19: The basic JFET switching circuit, with gate bootstrap resistor

The JFET types J111 and J112 are specially designed for analogue switching and pre-eminent for this application. The channel on-resistances are low and relatively linear. This is a depletion-mode FET, which requires a negative gate voltage to actively turn it off. The J111 requires a more negative V_{gs} to ensure it is off, but in return gives a lower $R_{ds(on)}$ which means lower distortion.

The J111, J112 (and J113) are members of the same family – in fact they are the device, selected for gate/channel characteristics, unless I am much mistaken. Table 21.6 shows how the J111 may need 10 V to turn it off, but gives a 30 Ω on-resistance or $R_{ds(on)}$ with zero gate voltage. In contrast the J112 needs only 5.0 V at most to turn it off, but has a higher $R_{ds(on)}$ of 50 Ω. The trade-off is between ease of generating the gate control voltages, and linearity. The higher the $R_{ds(on)}$, the higher the distortion, as this is a non-linear resistance.

FET tolerances are notoriously wide, and nothing varies more than the V_{gs} characteristic. It is essential to take the full range into account when designing the control circuitry.

Both the J111 and J112 are widely used for audio switching. The J111 has the advantage of the lowest distortion, but the J112 can be driven directly from 4000 series logic running from ±7.5 V rails, which is often convenient. The J113 appears to have no advantage to set against its high $R_{ds(on)}$ and is rarely used – I have never even seen one.

The circuits below use either J111 or J112, as appropriate. The typical version used is shown, along with typical values for associated components.

Figure 21.19 has Source and Drain marked on the JFET. In fact the J111 devices appear to be perfectly symmetrical, and it seems to make no difference which way round they are connected, so further diagrams omit this. As JFETs, in practical use they are not particularly static-sensitive.

The off voltage must be sufficiently negative to ensure that V_{gs} never becomes low enough to turn the JFET on. Since a J111 may require a V_{gs} of −10 V to turn it off, the off voltage must be 10 V below the negative saturation point of the driving opamp – hence the −23 V rail. This is not exactly a convenient voltage, but the rail does not need to supply much current and the extra cost in something like a mixing console is relatively small.

To turn a JFET on, the V_{gs} must be held at zero volts. That sounds simple enough, but it is actually the more difficult of the two states. Since the source is moving up and down with the signal, the gate must move up and down in exactly the same way to keep V_{gs} at zero. This is done by bootstrap resistor R_{boot} in Figure 21.19. When the JFET is off, DC flows through this resistor from the source; it is therefore essential that this path be DC-coupled and fed from a low impedance such as an opamp output, as shown in these diagrams. The relatively small DC current drawn from the opamp causes no problems.

Figure 21.20 is a more practical circuit using a driver transistor to control the JFET (if you had a switch contact handy, you would presumably use it to control the audio directly). The pull-up resistor R_c keeps diode D reverse-biased when the JFET is on; this is its sole function, so the value is not critical. It is usually high to reduce power consumption. I have used anything between 47 kΩ and 680 kΩ with success.

Sometimes DC-blocking is necessary if the opamp output is not at a DC level of 0 V. In this case the circuit of Figure 21.21 is very useful; the audio path is DC-blocked but not the bootstrap resistor, which must always have a DC path to the opamp output. R_{drain} keeps the capacitor voltage at zero when the JFET is held off.

Figure 21.22 shows the distortion performance with a load of 10 kΩ. The lower curve is the distortion from the 5532 opamp alone. The signal level was 7.75 Vrms (+20 dBu).

Figure 21.23 shows the distortion performance with various heavier loadings, from 10 kΩ down to 1 kΩ. As is usual in the world of electronics, heavier loading makes things worse. In this case, it is because the non-linear R_{on} becomes a more significant part of the total circuit resistance. The signal level was 7.75 Vrms (+20 dBu).

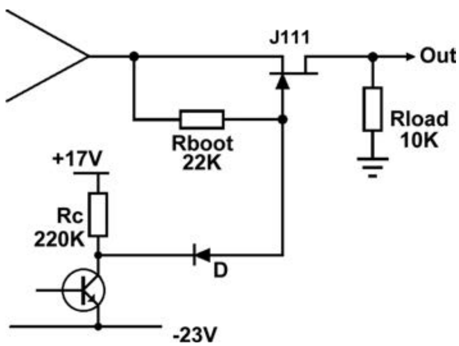

Figure 21.20: Using a transistor and diode for gate control

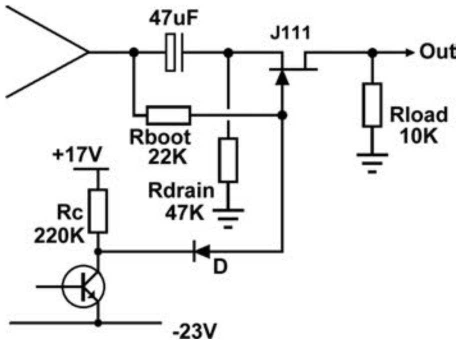

Figure 21.21: The JFET switching circuit with a DC blocking capacitor

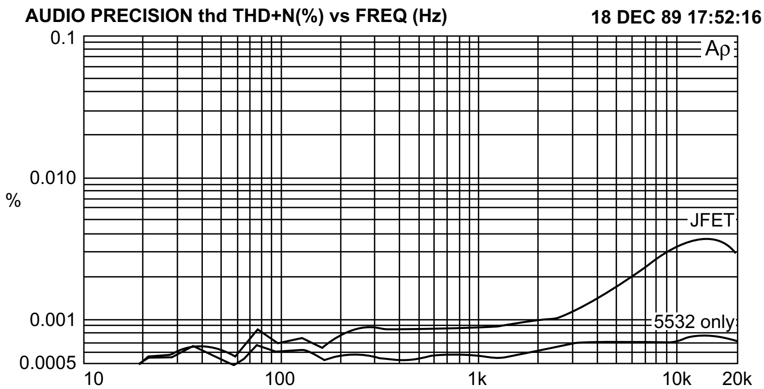

Figure 21.22: The JFET distortion performance with a load of 10 kΩ

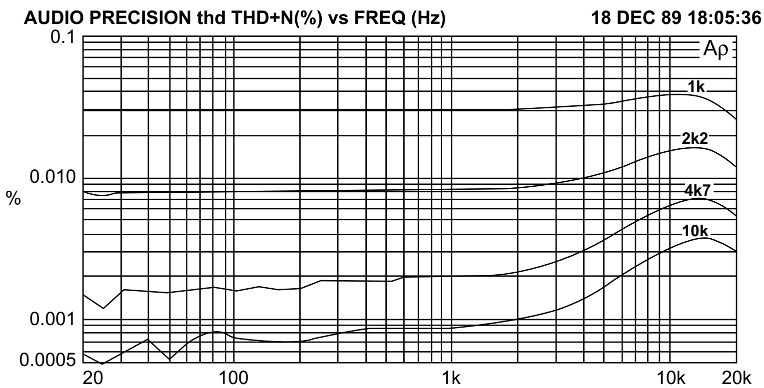

Figure 21.23: The JFET distortion performance versus loading

Figure 21.24 shows the distortion performance with different values of bootstrap resistor. The lower the value, the more accurately the drain follows the source at high audio frequencies, and so the lower the distortion. The signal level was 7.75 Vrms (+20 dBu) once again. There appears to be no disadvantage to using a bootstrap resistor of 22 kΩ or so, except in special circumstances, as explained below.

Two series JFET switches can be simply combined to make a changeover switch, as shown in Figure 21.25. The valid states are A on, B on, or both off. Both on is not a good option because the two opamps will then be driving each other's outputs through the JFETs.

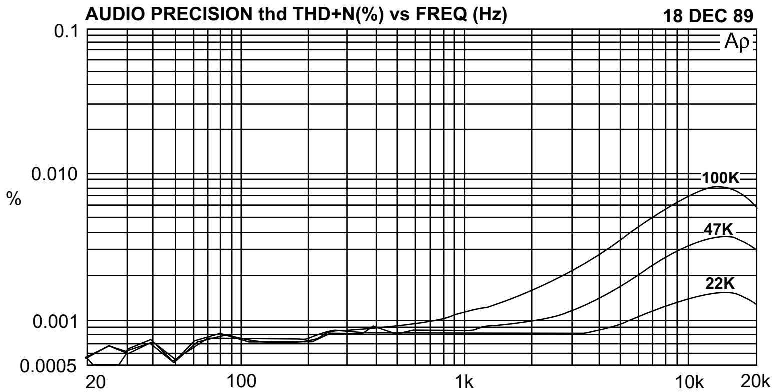

Figure 21.24: The distortion performance with different values of bootstrap resistor

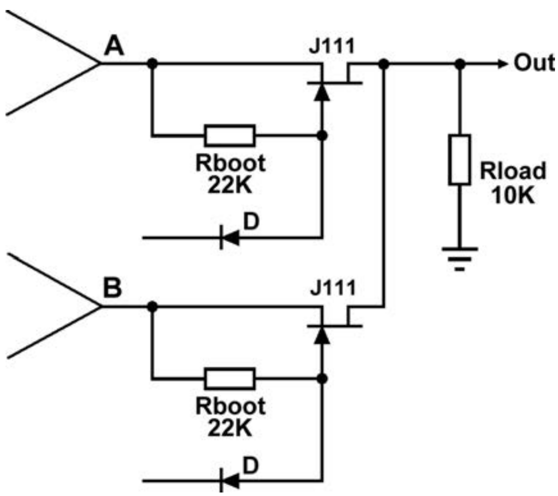

Figure 21.25: A JFET changeover switch

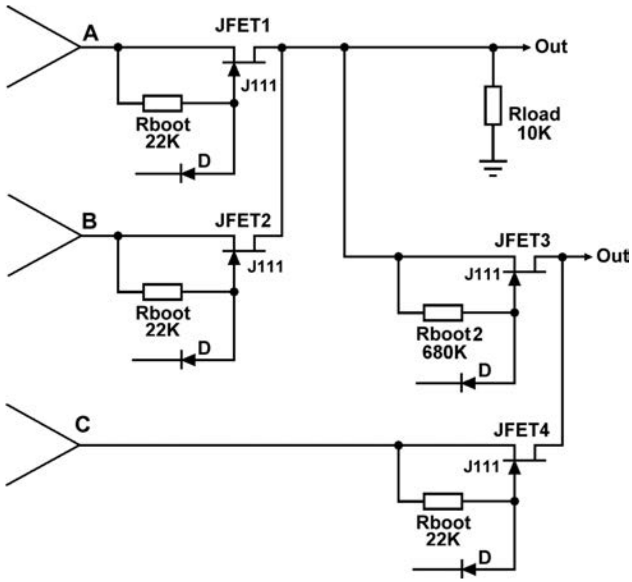

Figure 21.26: Cascaded FET switches

It is possible to cascade FET switches, as in Figure 21.26, which is taken from a real mixing console application. Here the main output is switched between A and B as before, but a second auxiliary output is switched between this selection and another input C by JFET3 and JFET4. The current drawn by the second bootstrap resistor R_{boot2} must flow through the $R_{ds(on)}$ of the first FET, and will thus generate a small click. R_{boot2} is therefore made as high as possible to minimise this effect, accepting that the distortion performance of the JFET3 switch will be compromised at HF; this was acceptable in the application as the second output was not a major signal path. The bootstrap resistor of JFET4 can be the desirable lower value as this path is driven direct from an opamp.

The shunt JFET switch in voltage mode

The basic JFET shunt switching circuit is shown in Figure 21.27. Like the shunt analogue gate mute, it gives poor offness but good linearity in the ON state, so long as its gate voltage is controlled so it never allows the JFET to begin conducting. Its great advantage is that the depletion JFET will be in its low-resistance state before and during circuit power-up, and can be used to mute switch-on transients. Switch-off transients can also be effectively muted if the drive circuitry is configured to turn on the shunt FETs as soon as the mains disappears, and keep them on until the various supply rails have completely collapsed.

I have used the circuit of Figure 21.27 to mute the turn-on and turn-off transients of a hifi preamplifier. Since this is an output that is likely to drive a reasonable length of cable, with its attendant capacitance, it is important to keep R1 as low as possible, to minimise the possibility of a drooping treble response. This means that the $R_{ds(on)}$ of the JFET puts a limit on the offness possible. The output series resistor R1 is normally in the range 47–100 Ohms, when it has as its only job the isolation of the output opamp from cable capacitance. Here it has a value of 1K, which is a distinct compromise – it is not suited for use with very long cables. Even with this value the muting obtained was not quite adequate at −27 dB so two J111s were used in parallel, giving a further −6 dB of attenuation. The resulting −33 dB across the audio band was sufficient to render the transients inaudible. The offness is not frequency dependent as the impedances are low and so stray capacitance is irrelevant.

JFETS in current mode

JFETS can be used in the current mode, just as for analogue gates. Figure 21.28 shows the basic muting circuit, with series FET switching only. R_{in2} attenuates the signal seen by the FET when it is off, to prevent breakthrough; its presence means that the gain of the circuit is somewhat less than unity with the values shown, but the gain may be readily adjusted by

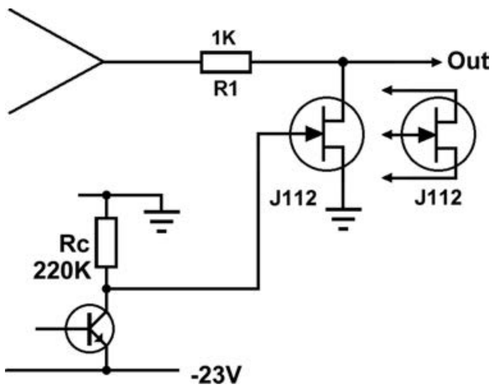

Figure 21.27: The basic JFET shunt switching circuit. Adding more JFETs in parallel increases the offness, but each −6 dB requires doubling their number

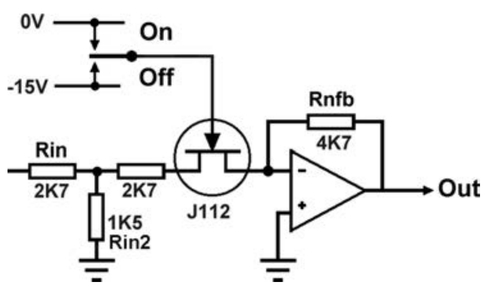

Figure 21.28: The simplest FET mute circuit: single-FET muting forces a crosstalk/linearity trade-off

altering the value of R_{nfb}. Figure 21.29 illustrates the distortion performance and Figure 21.30 the offness of this circuit. Neither is startlingly good.

In designing a mute block, we want low distortion *and* good offness at the same time, so the series-shunt configuration, which proved highly effective with CMOS analogue gates, is the obvious choice. The basic circuit is shown in Figure 21.31, the distortion performance is illustrated in Figure 21.32, and the offness in Figures 21.33 and 21.34. Capacitor C1 across the feedback resistor is usually required to ensure HF stability, due to the FET capacitances hanging on the summing node at D.

Due to the shunt-feedback configuration, this circuit introduces a phase-inversion. I have often been forced to follow this circuit with another inverting stage that does nothing except get the phase right again. In this situation, it is sometimes advantageous to put the inverting

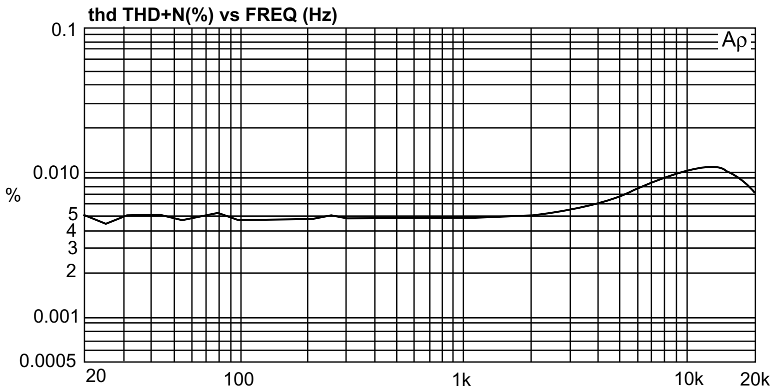

Figure 21.29: THD of the single-FET circuit in Figure 21.28

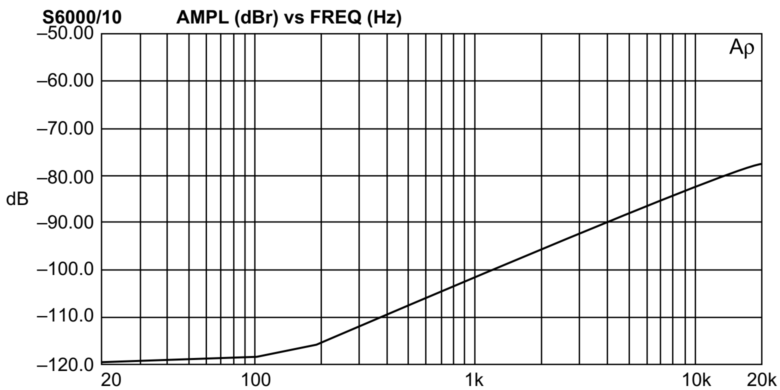

Figure 21.30: Offness of the single-FET circuit in Figure 21.28. It only manages −982 dB at 10 kHz

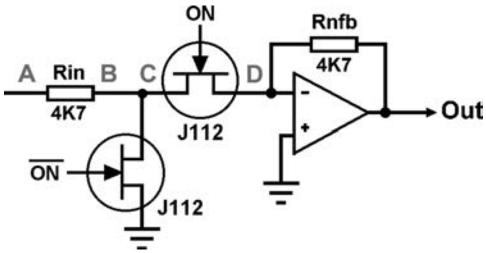

Figure 21.31: Series-shunt mode mute bloc circuit

stage *before* the mute block, so that any crosstalk to its sensitive summing node is muted with the signal by the following mute block.

The control voltages to the series and shunt JFETs are complementary as with the CMOS version, but now they can be slowed down by RC networks to make the operation gradual, as shown in Figure 21.35 below. The exact way in which the control voltages overlap is easy to control, but the V_{gs}/resistance law of the FET is not (and it is about the most variable FET parameter there is) and so the overlap of FET conduction is rather variable. However, I should say at once that this system does work, and works well enough to go in top-notch mixing consoles. As you go into the muted condition the series JFET turns off and the shunt JFET turns on, and if the overlap gets to be too much in error, the following bad things can happen:

1. If the shunt FET turns on too early, while the series JFET is still mostly on, a low-resistance path is established from the opamp VE point to ground, causing a large but brief rise in stage noise gain. This produces a 'chuff' of noise at the output as muting occurs.

2. If the shunt FET turns on too late, so the series JFET is mostly off, the large signal voltage presented to the series FET causes visibly serious distortion. I say 'visibly' because it is well-known that even quite severe distortion is not obtrusive if it occurs only briefly. The transition here is usually fast enough for this to be the case; it would not however be a practical way to generate a slow fade. The conclusion is that we should err on the side of distortion rather than noise.

Reducing distortion by biasing

The distortion generated by this circuit block is of considerable importance, because if the rest of the audio path is made up of 5532 opamps – which is likely in professional equipment – then this stage can generate more distortion than the rest of the signal path combined, and dominate this aspect of the performance. It is therefore worth examining any way of increasing the linearity that we can think of.

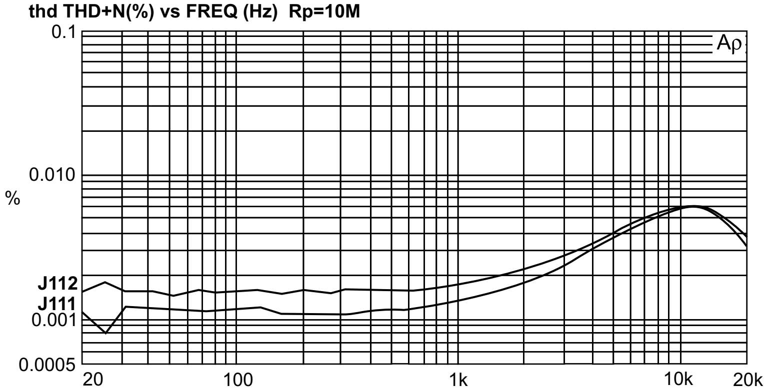

Figure 21.32: The THD of the mute bloc in Figure 21.35 with R = 4k7 and Rp = 10 MΩ. The increase in JFET distortion caused by using a J112 rather than a J111 is shown. The rising distortion above 1 kHz comes from the opamp

We have already noted that to minimise distortion, the series JFET should be turned on as fully as possible to minimise the value of the non-linear $R_{ds(on)}$. When a JFET has a zero gate-source voltage, it is normally considered fully on. It is, however, possible to turn it even more on than this.

The technique is to put a small positive voltage on the gate, say about 200–300 mV. This further reduces the $R_{ds(on)}$ in a smoothly continuous manner, without forward biasing the JFET gate junction and injecting DC into the signal path. This is accomplished in Figure 21.35 by the simple addition of R_p, which allows a small positive voltage to be set up across the 680K resistor R1. The value of R_p is usually in the 10–22 MΩ range, for the circuit values shown here.

Care is needed with this technique, because if temperatures rise the JFET gate diode may begin to conduct after all and DC will leak into the signal path, causing thumps and bangs. In my experience 300 mV is about the upper safe limit for equipment that gets reasonably warm internally, i.e. about 50 °C. Caution is the watchword here, for unwanted transients are much less tolerable than slightly increased distortion.

As with analogue CMOS gates, an important consideration with this circuit is the impedance at which it works, i.e. the values of R_{in} and R_{nfb}. These are usually of equal resistance for unity gain so we will call their value R:

1. Raising R reduces distortion because it minimises the effect of $R_{ds(on)}$ variation in the series JFET.

2. Lowering R reduces the noise generated by the circuit, and improves offness as it reduces the effect of stray capacitances. It also reduces the effect of control-voltage feedthrough via the gate-channel capacitances.

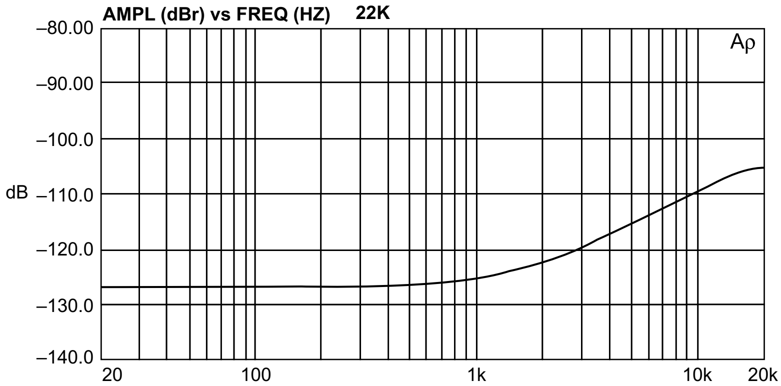

Figure 21.33: Offness of mute bloc in Figure 21.35 with $R_{in} = R_{nfb} = 22k$

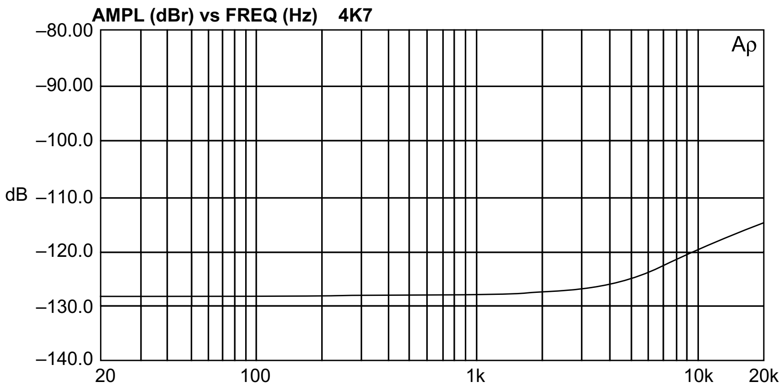

Figure 21.34: Offness of mute bloc in Figure 21.35 with $R_{in} = R_{nfb} = 4k7$. Offness is better and the noise floor (the flat section below 2 kHz) has been lowered by about 2 dB

Figures 21.33 and 21.34 examine how the offness of the circuit is affected by using values of 4k7 and 22 kΩ. The latter gives −110 dB rather than −120 dB at 10 kHz. In my not inconsiderable experience with this circuit, R = 4k7 is the best choice when J112s are used. Values below 4k7 are not usual as distortion will increase as the JFET $R_{ds(on)}$ becomes a larger part of the total resistance in the circuit. The loading effect of R_{in} on the previous stage must also be considered.

JFET drive circuitry

The series-shunt mute bloc requires two complementary drive voltages, and these are most easily generated by 4000-series CMOS running from ±7.5 V rails. NAND gates are shown here as they are convenient for interfacing with other bits of control logic, but any standard CMOS output can be used. It is vital that the JFET gates get as close to 0 V as possible,

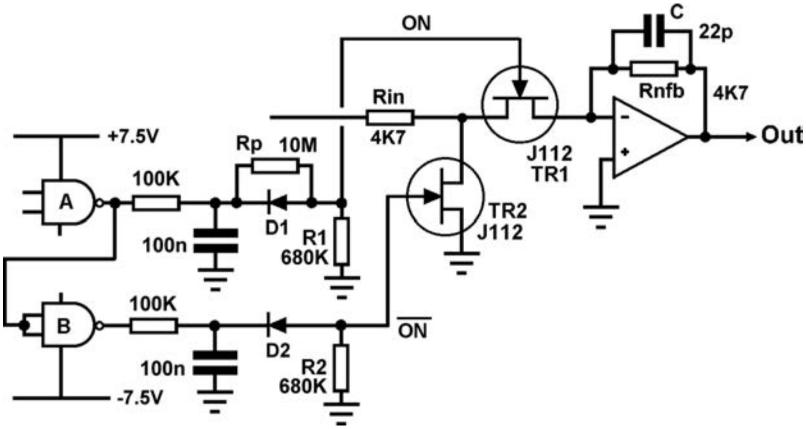

Figure 21.35: Circuitry to generate drive voltages for series-shunt JFET mute bloc

ensuring that the series gate can be fully on and give minimum distortion, so the best technique is to run the logic from these ± rails and use diodes to clamp the gates to 0 V.

Thus, in Figure 21.35, when the mute bloc is passing signal, the signal from gate A is high, so D1 is reverse-biased and the series JFET TR1 gate is held at 0 V by R1, keeping it on (the role of R_p will be explained in a moment). Meanwhile, D2 is conducting as the NAND-gate output driving it is low, so the shunt JFET TR2 gate is at about −7 V and it is firmly switched off. This voltage is more than enough to turn off a J112, but cannot be guaranteed to turn off a J111, which may require −10 V (see Table 21.6 above). This is one reason why the J112 is more often used in this application – it is simpler to generate the control voltages. When the mute bloc is off, the conditions are reversed, with the output of A low, turning off TR1, and the output of B high, turning on TR2.

When switching audio signals, a instantaneous cut of the signal is sometimes not what is required. When a non-zero audio signal is abruptly interrupted there is bound to be a click. Perhaps surprisingly, clever schemes for making the instant of switching coincide with a zero-crossing give little improvement. There may no longer be a step-change in level, but there is still a step-change in slope and the ear once more interprets this discontinuity as a click.

What is really needed is a fast-fade over about 10 msec. This is long enough to prevent clicks, without being so slow that the timing of the event becomes sloppy. This is normally only an issue in mixing consoles, where it is necessary for things to happen in real time. Such fast-fade circuits are often called ‘mute blocks’ to emphasise that they are more than just simple on-off switches. Analogue gates cannot be slowly turned on and off due to their internal circuitry for control-voltage generation, so discrete JFETs must be used. Custom chips to perform the muting function have been produced, but the ones I have evaluated have been

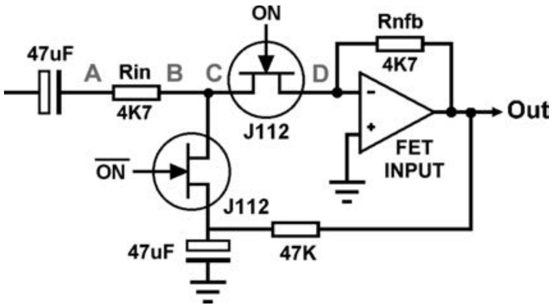

Figure 21.36: Circuit of JFET mute showing stray capacitances and DC handling

expensive, single-source, and give less than startling results for linearity and offness; this situation is of course subject to change.

In Figure 21.35 the rate of control-voltage is determined by the RC networks at the NAND gate outputs. Ingenious schemes involving diodes to make the up/down rates different have been tried many times but my general conclusion is that they give little, if any, benefit.

Physical layout and offness

The offness of this circuit is extremely good, providing certain precautions are taken in the physical layout. In Figure 21.36 there are two possible crosstalk paths that can damage the offness. The path C–D, through the internal capacitances of the series JFET, is rendered innocuous as C is connected firmly to ground by the shunt JFET. However, point A is still alive with full amplitude signal, and it is the stray capacitance from A to D that defines the offness at high frequencies.

Given the finite size of R_{in}, it is often necessary to extend the PCB track B–C to get A far enough from D. This is no problem if done with caution. Remember that the track B–C is at virtual earth when the mute bloc is on, and so vulnerable to capacitative crosstalk from other signals straying into the area.

Dealing with the DC conditions

The circuits shown so far have been stripped down to their bare essentials to get the basic principles across. In reality, things are (surprise) a little more complicated. Opamps have non-zero offset and bias voltages and currents, and if not handled properly these will lead to thumps and bangs. There are several issues:

1. If there is any DC voltage at all passed on from the previous stage, this will be interrupted along with the signal, causing a click or thump. The foolproof answer is of course a DC-blocking capacitor, but if you are aiming to remove all capacitors from the signal

path, you may have a problem. DC servos can partly make up the lack, but since they are based on opamp integrators they are no more accurate than the opamp, while DC blocking is foolproof.

2. The offset voltage of the mute bloc opamp. If the noise gain is changed when the mute operates (which it is) the changing amplification of this offset will change the DC level at the output. The answer is shown in Figure 21.36. The shunt FET is connected to ground via a blocking capacitor to prevent gain changes. This capacitor does not count as 'being in the signal path' as audio only goes through it when the circuit is muted. Feedback of the opamp offset voltage to this capacitor via the 47 kΩ resistor renders it innocuous.

3. The input bias and offset currents of the opamp. These are much more of a problem and are best dealt with by using JFET opamps such as TL072 or OPA2134, where the bias and offset currents are negligible at normal equipment temperatures. All of the distortion measurements in this chapter were made with TL072 opamps in place.

A soft changeover circuit

This circuit (Figure 21.37) is designed to give a soft changeover between two inputs – in effect a fast crossfade. It is the same mute bloc but with two separate inputs, either or both of which can be switched on. The performance at +20 dBu in/out is summarised in Table 21.7.

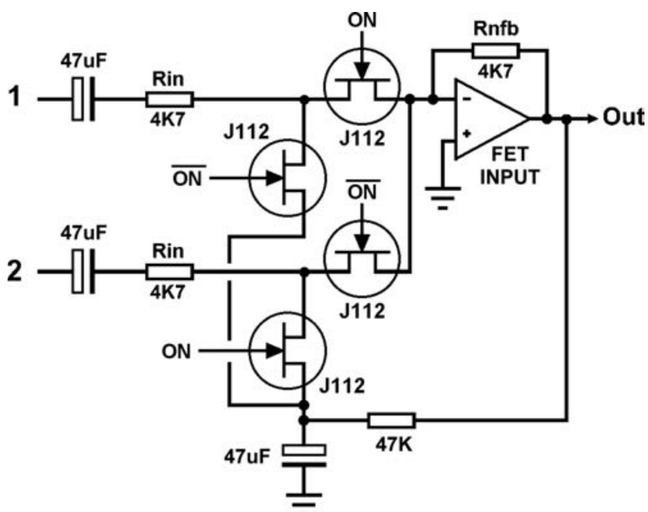

Figure 21.37: Circuit of JFET soft changeover switching

TABLE 21.7 Distortion produced by JFET soft changeover switch

	1 kHz	10 kHz	20 kHz
THD +20 dBu	0.0023%	0.0027%	0.0039%
Offness	−114 dB	−109 dB	−105 dB

The THD increase at 20 kHz is due to the use of a TL072 as the opamp. J112 JFETs are used in all positions.

This circuit is intended for soft-switching applications where the transition between states is fast enough for a burst of increased distortion to go unnoticed. It is not suitable for generating slow crossfades in applications like disco mixers, as the exact crossfade law is not very predictable.

Control voltage feedthrough in JFETS

All discrete FETs have a small capacitance between the gate and the device channel, so changes in the gate voltage will therefore cause a charge to be transferred to the audio path, just as for CMOS analogue gates. As before, slowing down the control voltage change tends to give a thump rather than a click; the same amount of electric charge has been transferred to the audio path, but more slowly. Lowering the circuit impedance reduces the effects of feedthrough, but halving it only reduces the amplitude of transients by 6 dB, and such a reduction is likely to increase distortion.

Reference

[1] Jones, M. 'Designing Valve Preamps', *Electronics World* (March 1996), p. 193.

CHAPTER 22

Mixer sub-systems

Introduction

This chapter deals with the specialised circuit blocks that make up mixing console. Some functions, such as microphone amplification, line input and output, and equalisation have already been explored in previous chapters. The other useful blocks are presented in the order that a signal encounters them as it goes through the console, but many of them are applicable to all the various kinds of module – input channels, groups, and master sections. For example, fader post-amplifiers and insert points will be found in all three types of module.

Mixer bus systems

In all but the smallest mixers there is a need to connect together all the modules so they have access to the mixing buses, power supply rails, and logic and control lines.

There are three basic ways of connecting the modules together. The smallest mixers are usually constructed on a single large PCB lying parallel to a one-piece front panel, and here 'modules' means repeated circuitry rather than physical modules. The all-embracing PCB minimises the money spent on connectors, and the time plugging them in during assembly, but there are obvious limitations to the size of mixer you can build in this way. A definite problem is the need to run summing buses laterally, as this results in them winding their way between controls and circuit blocks, threatening a mediocre crosstalk performance. The use of double-sided PCBs helps greatly with this, but very often there are still awkward points such as the need for the feed to and from the faders to cross over the mix bus area. This can easily wreck the crosstalk performance; one solution is to use what I call a 'three-layer board'. The mix buses are on the bottom of the PCB, the top layer above it carries a section of ground plane, and the fader connections are made by wire links above that. Given a tough solder-resist, no further insulation of the links is necessary; if you have doubts, then laying a rectangle of component-ident screen print under the links will give another layer of insulation without adding labour cost.

Medium-sized mixers are commonly made with separate modules, connected together with ribbon cable bearing insulation-displacement connectors. The advent of IDC ribbon cables (a long time ago, now) had a major effect on the affordability of mixing consoles. These cables naturally join Pin 1 to Pin 1 on every module, and so on, leading to a certain inflexibility in design.

Large mixers use a motherboard system, where each module plugs into a PCB at the bottom of the frame, which is typically divided into 'bins' holding eight or 12 modules. This provides (at considerably increased expense) total flexibility in the running of buses and the interconnection of modules.

To me, it is a 'mixing bus' or a 'summing bus'. I realise that some of the world spells it 'buss' and I am probably wasting my time pointing out that the latter is wrong but I am still going to do it. The term dates from the dawn of electrical distribution, when circuits were connected together by copper bars called 'omnibus bars'. This inevitably got shortened to 'bus-bars' and in mixing consoles it was further abbreviated to 'bus' which somehow turned into 'buss'. The Oxford English Dictionary says buss means 'to kiss' (*archaic*) which seems somehow not quite appropriate. Unless of course you're recording Kiss. I have grave doubts if my protest here will make any difference to common usage, but sometimes you've just got to make a stand.

Input arrangements

Most mixer channels have both microphone and line inputs. On the lower-cost consoles these are usually switched to a single amplifier with a wide gain range, the line input being attenuated to a suitable level first. This approach is covered in Chapter 17 on microphone amplifiers. High-end consoles have separate line input amplifiers, removing some compromises on CMRR and noise performance. Dedicated line input amplifiers are dealt with in Chapter 18.

Equalisation

Mixer input channels have more or less sophisticated tone controls to modify the frequency response, either to correct imperfections or produce specific effects. This subject is fully dealt with in Chapter 15.

Insert points

The addition of effects for general use, such as reverberation, is normally handled by an effects send system. However, if a specific effect (say, flanging) is going to be used on one channel only then it is far more efficient and convenient to connect the external effects unit in series with the signal path of the channel itself. This is done by means of an insert point (usually just called an 'insert') which is a jack with normalling contacts arranged so that the signal flows to it and back to the channel again when nothing is plugged in. When a jack is inserted the normalling connection is broken and the signal flows through the external unit.

Inserts are also often fitted to groups. Inserts come in two versions, illustrated in Figure 22.1. The single jack version is economical in panel space but is restricted to unbalanced operation. The two-jack version is superior because it allows the use of balanced send and returns, and in addition the OUT jack socket can be used as a direct output because inserting a jack in it does not break the signal path; only inserting a jack into the IN jack socket does that.

When the mixer has a patchbay, the insert sends (and indeed, other console outputs) are likely to find their way there through a quite considerable length of ribbon cable, which has significant capacitance between its conductors. It is easy to get into a situation where the crosstalk performance of the console is limited by capacitive crosstalk between outputs, despite their low impedance. Output amplifiers commonly have a series resistor to isolate the amplifier from the capacitance of the cabling and prevent HF instability, and the minimum safe value of this resistor defines the output impedance, which is usually in the region of 47 to 100 Ω. Things get worse when the layers of ribbon cable are laid together in a 'lasagne' format; this is very often necessary because of the sheer number of signals going to and from the patchbay. In some cases layers of grounded screening foil are interleaved with the cables, but this is rather expensive and awkward to do, and does not greatly reduce crosstalk between conductors in the same piece of ribbon. The only way to do this is to reduce the output impedance.

In a particular mixer design project, the crosstalk between the insert sends from the channels, with an output impedance of 75 Ω, was found to be −96 dB at 10 kHz. This may not sound like a lot, but I didn't get where I am today by designing consoles with measurable crosstalk,

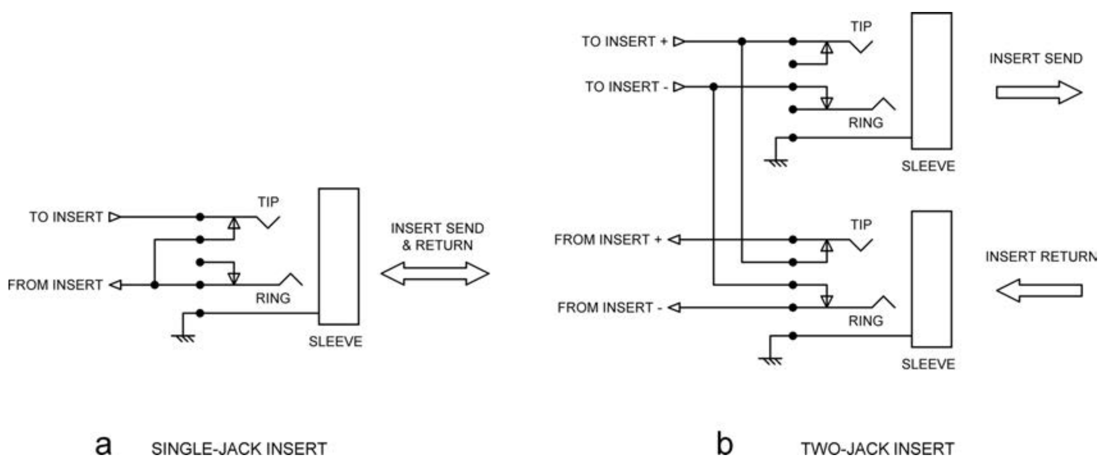

Figure 22.1: a) The single-jack insert and b) the two-jack insert

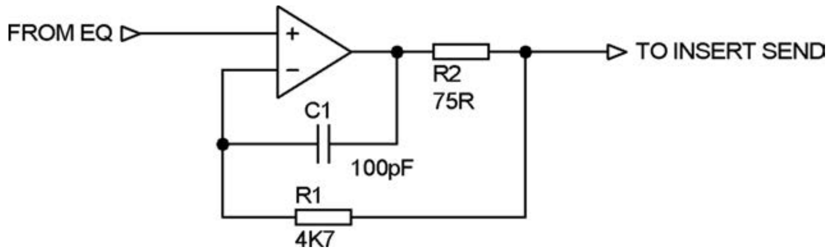

Figure 22.2: A typical non-inverting insert-send amplifier with 'zero-impedance' output

so something had to be done. An effective way to obtain a near-zero output impedance is shown in Figure 22.2. Here the main negative feedback for the opamp goes through R1, from the outside end of isolating resistor R2 and so reduces the output impedance, while the stabilising HF feedback is taken through C1 from the inside end, where it is not subject to phase-shift because of load capacitance. With this insert send stage the output impedance was reduced from 75 Ω to less than 1 Ω and the crosstalk disappeared below the noise floor. Very similar circuitry can be used with stages that have gain. Arrangements like this must always be carefully checked to make sure that HF stability with a capacitive load really *is* maintained; this circuit is stable when driving a 22 nF load, which represents 220 metres of 100 pF/metre cable.

This arrangement is sometimes called a 'zero-impedance' output; the impedance is certainly much lower than usual but it is not, of course, actually zero.

In a group module, an inverting insert send amplifier is often used to correct the phase inversion introduced by the summing amplifier. A zero-impedance version of this is illustrated in the section below on summing amplifiers.

How to move a circuit block

In the more sophisticated and versatile mixers it is often possible to rearrange the order of blocks in the signal path. A typical example is the facility to move an insert point from before to after the EQ section, depending on what sort of external processing is being plugged into the insert. Dynamics sections are also often movable to pre or post the EQ.

Figure 22.3 shows two ways to move a circuit block from one position to another in the signal path. The version at Figure 22.3a uses all the sections of a four-changeover switch to do the job. Normally the insert is before the EQ section, but when the switch (which would probably be labelled 'insert post' or whatever abbreviation of that which can be fitted in) is pressed the signal passes along path A and reaches the EQ first; it then goes back through

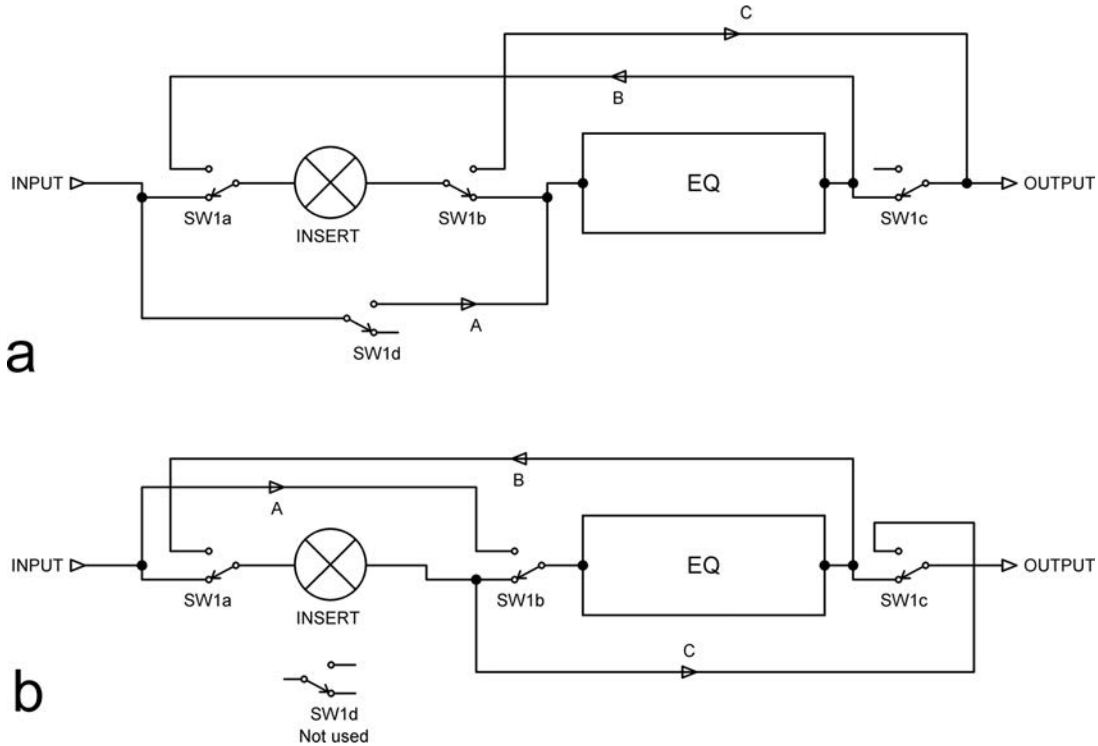

Figure 22.3: Two ways to move a block in the signal path. That at b) requires only three switch sections instead of four

path B, through the insert, and then to the output along path C. However, the existence of two unused switch contacts gives us a broad hint that there may be a more efficient way to do it.

There certainly is; Figure 22.3b shows how to do it with only three changeovers. When the switch is pressed the signal passes along path A and reaches the EQ first, goes back through path B, to the insert, and then out via path C. You might think that this economy is quite pointless because either way you will need to use a physical four-changeover switch, but in fact the spare switch section in Figure 22.3b comes in very handy indeed to operate a switch status indicator LED. Alternatively the spare switch section can be connected in parallel with one of the other sections in the hope of improving reliability, but . . .

Paralleling switch sections is not actually as useful for enhancing reliability as you might think. The classic switch problem is contamination of silver-plated contacts; the silver is converted to non-conductive silver sulphide by the action of hydrogen sulphide in the atmosphere. This can come from industrial pollution, but another source is diesel engine exhaust so virtually nowhere can be assumed to be free of it. If one set of contacts in a switch is affected then the other set right next to it is certain to be affected also, and paralleling the contacts is actually of little or no use. Sad but true.

Faders

So far as circuit design is concerned, a slide fader can normally be regarded as simply a logarithmic potentiometer, though in fact its internal construction may be quite complicated. The technology of faders is dealt with in Chapter 13 on volume controls.

Postfade amplifiers

The postfade amplifier (sometimes called just the post amp) comes just after the fader, as its name implies. Its primary function is to allow the fader section to give gain. A +10 dB post amplifier, which is the most common amount of gain, allows the fader to be calibrated up to +10 dB. The 0 dB setting is then some way down the scale and it is much easier to adjust the channel level up and down with respect to the rest of the mix. Without this feature, if a channel was fully faded up at 0 dB, to increase its relative level it would be necessary to pull all the other faders down. When you consider that a large mixer may have 64 channel faders you can see that this would be something of a nuisance.

The second function of the post amp is to buffer the fader from the heavy loading of the panpot, mix resistors and aux sends. The fader control law is carefully chosen for optimal controllability, and that law would be seriously distorted if all that loading came directly on the fader wiper.

It is important that the post amp has good noise performance, because its noise contribution is not much reduced when the fader is pulled down, and it must also have a good load-driving capability at low distortion. For these two reasons, the 5534/2 opamp found a home in post amps quite early in its history, when its high cost ruled it out for most other circuit functions, which were still handled by TL072s. The other place that 5534/2s appeared early was in summing amplifiers, which often work at a high noise gain.

However, in budget designs TL072s or equivalent FET-input opamps are still used because they are not only cheaper themselves but allow several other parts to be omitted. In Figure 22.4a, the opamp input can be directly connected to the fader wiper because the input bias current is negligible and does not cause significant noise when the wiper is moved along the track. Note that the gain-determining resistors R2, R3 have to be quite high to minimise the loading on the opamp output. C3 improves stability.

In a more sophisticated design as in Figure 22.4b, the use of a 5534 with its significant bias current means that DC blocking capacitor C1 is required to prevent unpleasant sounds when the

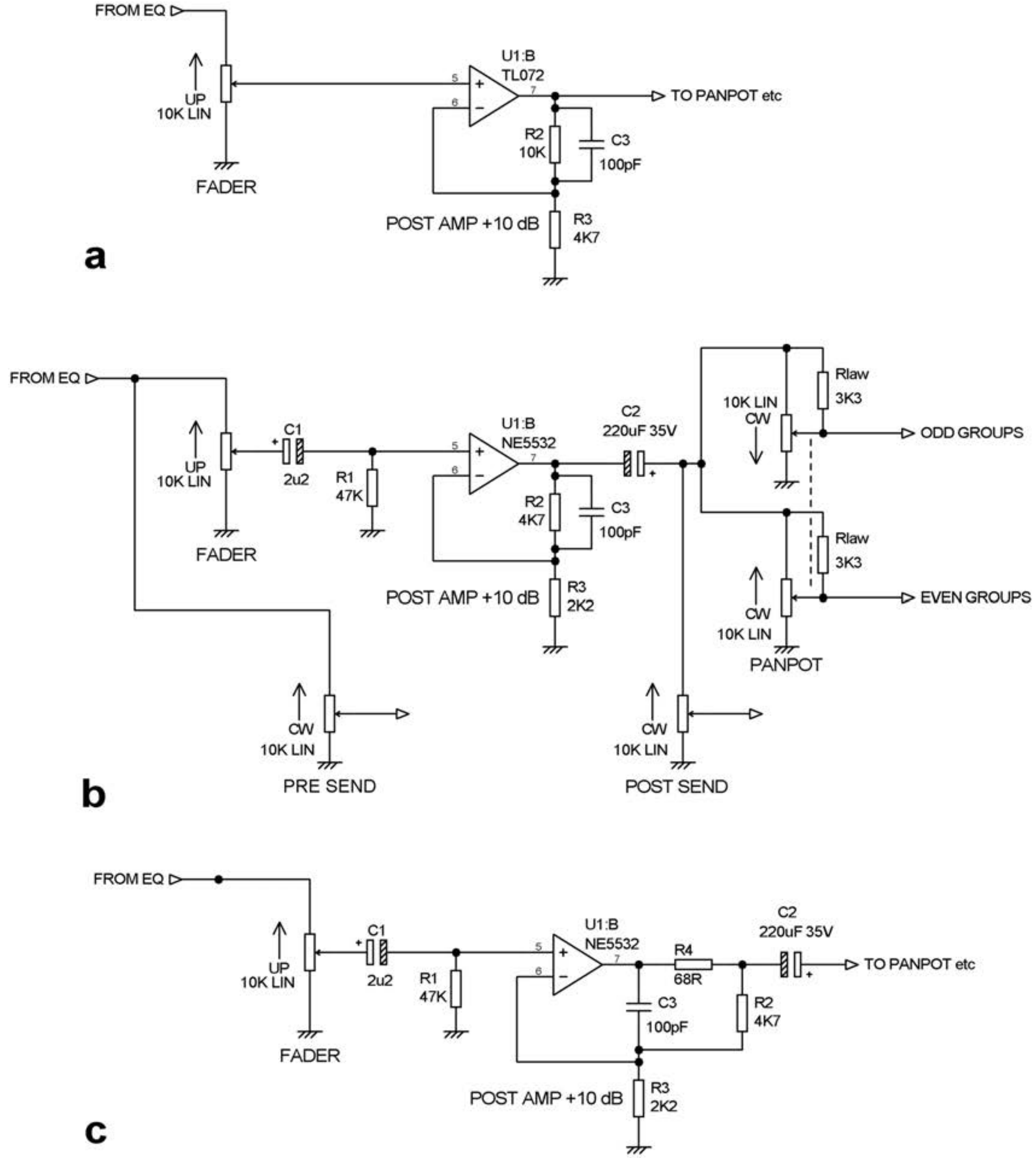

Figure 22.4: Three versions of a postfade amplifier. That at a) is the cheap and cheerful version, b) is the more up-market version, with pre and post sends and the panpot section of a mixer channel shown, c) protects the opamp against capacitive loading while still giving a 'zero-impedance' output

fader is adjusted; in turn this means that R1 is required to bias the opamp. Another consequence flows from this: R1 has to have a relatively high value so it does not load the fader and distort its control law, and so the 5534 input bias current causes a relatively large offset voltage at the opamp input. The 5534 max bias current is 800 nA, which means a possible drop across R1 of −37.6 mV. The 5534 has NPN input devices, so the bias current flows into the input pins, and so this voltage is negative, and that is why C1 is the way round it is. This input offset appears multiplied by three at the post amp output, giving −113 mV. This is not enough to significantly affect the available output swing, but it does mean that C2 must be introduced to keep DC out of the panpot and postfade sends. Since the loading is quite heavy, C2 must be large to prevent it creating distortion (see Chapter 2). Note also the polarity of C2. We now have not only a more expensive opamp with a higher power consumption, but three extra components. Their number is multiplied by N, the number of channels, so you can see that the use of the 5534 as a post amp was actually a more serious cost decision than it might appear.

You will note that in Figure 22.4b the resistors R2, R3 in the feedback network have been reduced in value by a factor of about two to improve the noise performance. However, they are still not particularly low in value and it is true that the noise performance could be improved if they were reduced to, say, 680 Ω and 330 Ω, which in itself would be well within the drive capability of a 5532. However, in simpler desks this opamp has to drive not only its own feedback network, but also the panpot, routing matrix and the postfade sends, and most of its drive capability has to be reserved for this duty.

Figure 22.4c shows a post amp utilising the 'zero-impedance' approach described above for insert send amplifiers. In some layouts the panpot, routing and postfade sends are physically spread out so that the stray capacitance seen by the post amp output is enough to imperil its stability. An output isolating resistor R4 cures the problem but if simply stuck in the output line causes the level after it to vary as changing panpot and send settings alter the load on it. The feedback resistor R2 is therefore fed from after R4, preserving the 'zero-impedance' output, but C3 is fed from before it, maintaining the HF stability.

Fader post amps almost invariably have a fixed gain; if they did not the carefully-designed fader law would no longer be obtained. However, there are places where making the gain effectively variable by the use of positive feedback allows the post amp to have a low gain when its associated control is at a low setting; this minimises the noise contribution of the post amp. There is more on this technique in the section on aux masters.

Direct outputs

In more complex consoles, a direct out is available from each channel. This is a postfade signal, which can be fed directly to a recording device without sending it through the routing and summing systems. This gives a minimum signal path which will exhibit less noise and possibly less distortion. In all except the most elaborate consoles the direct out tends to be unbalanced, to save hardware, and reliance is placed on the recording device having balanced inputs.

Panpots

The word 'panpot' is short for 'panoramic potentiometer'. It is the control that places a monaural source in the desired place in the left-right stereo scene. It is an extremely fortunate property of human hearing that this can be done effectively simply by altering the proportion of the mono signal that is sent to the left and right channels of the stereo output. Changing the perception of up and down is considerably more complicated, and outside the scope of mixer design.

The earliest attempt at 'pan-potting' dates back to before the use of electrical amplification. In 1903, the French engineer, Leon Gaumont, was granted patents for loudspeaker systems to go with his sound-on-gramophone-disc talking films. Gaumont was the first to suggest placing loudspeakers behind the screen, with men carrying them about to follow the images on the screen. This procedure has never found favour in the sound industry, especially amongst those who might be asked to do the carrying.

To give smooth stereo panning without unwanted level changes, the panpot should theoretically have a sine/cosine characteristic, as shown in Figure 22.5; with this law, both

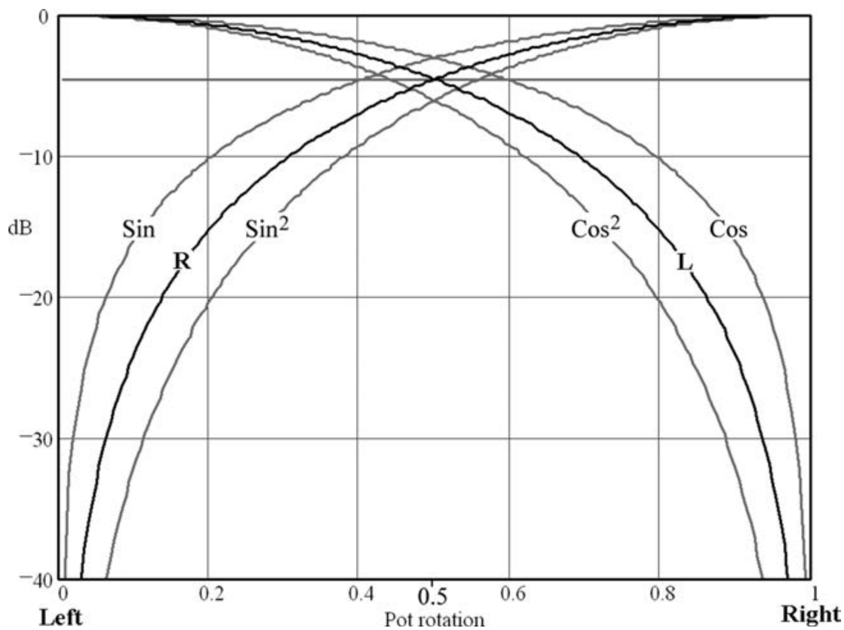

Figure 22.5: Panpot laws. The ideal compromise between the sine and sine-squared laws is the arithmetical mean of the two, shown here

signals are attenuated by 3 dB when the panpot is set centrally. This is because when listening to stereo over loudspeakers, the signals that sum at each ear are not correlated, and so with equal amplitude the two signals give a result that is only 3 dB above the level of each of the components. The summation of two uncorrelated white noise sources to give a combined signal that is 3 dB greater, and not 6 dB, works on exactly the same principle. The sine/cosine law is therefore appropriate for stereo.

However, when you are listening in mono, such as via AM radio, the left and right signals have been summed when they are still in the electrical domain, and therefore are in phase and sum together arithmetically. In this case, the pan law needs to be −6 dB down at the centre rather than −3 dB if level changes are to be avoided, so in this case a sine-squared/cosine-squared law is required. Since it is highly desirable to cope with both cases, the standard answer is to use a compromise between the two laws. This is entirely satisfactory in practice, not least because moving a component of the mix around during the performance is completely inappropriate for most forms of music, and I have yet to hear a music reviewer complain of 'bad panpotting'.

Figure 22.5 shows both the sine and sine-squared theoretical laws, and also the lines L, R which indicate the ideal compromise law; it is the arithmetical mean of sine and sine-squared and so is −4.5 dB down with the panpot central. So – how do we obtain the compromise law we want?

Passive panpots

Figure 22.6 shows three different ways to make a panpot. Panpot circuits can be made with single pots with the wiper grounded, as shown in Figure 22.6a, but this has several disadvantages, not least the poor offness when panned hard over, caused by the current flowing through the wiper contact resistance to ground.

Potentiometers with an accurate sine/cosine law used to exist, which in theory allowed the use of the simple panpot circuit in Figure 22.6b but they were always bulky and prohibitively expensive for audio use, and seem to have disappeared altogether now that analogue computers are no longer at the cutting edge of technology. Hifi preamplifiers sometimes use a control with a 'balance law' for adjusting channel balance (see Chapter 14 for more on that) but this usually has no attenuation at all at the centre point and would make a poor panpot.

The traditional answer to the problem has always been to use dual linear pots and bend the linear law into an approximation of the compromise law we want by the use of fixed pull-up resistors connected to the wipers, as shown in Figure 22.6c. Here the panpot wipers are loaded with mix resistors of 10 kΩ, which being connected to virtual-earth buses are connected to ground as far as the panpot is concerned, and the pull-up resistors R_{law} are made 3k9 to give the desired 4.5 dB of attenuation at the centre position of the panpot.

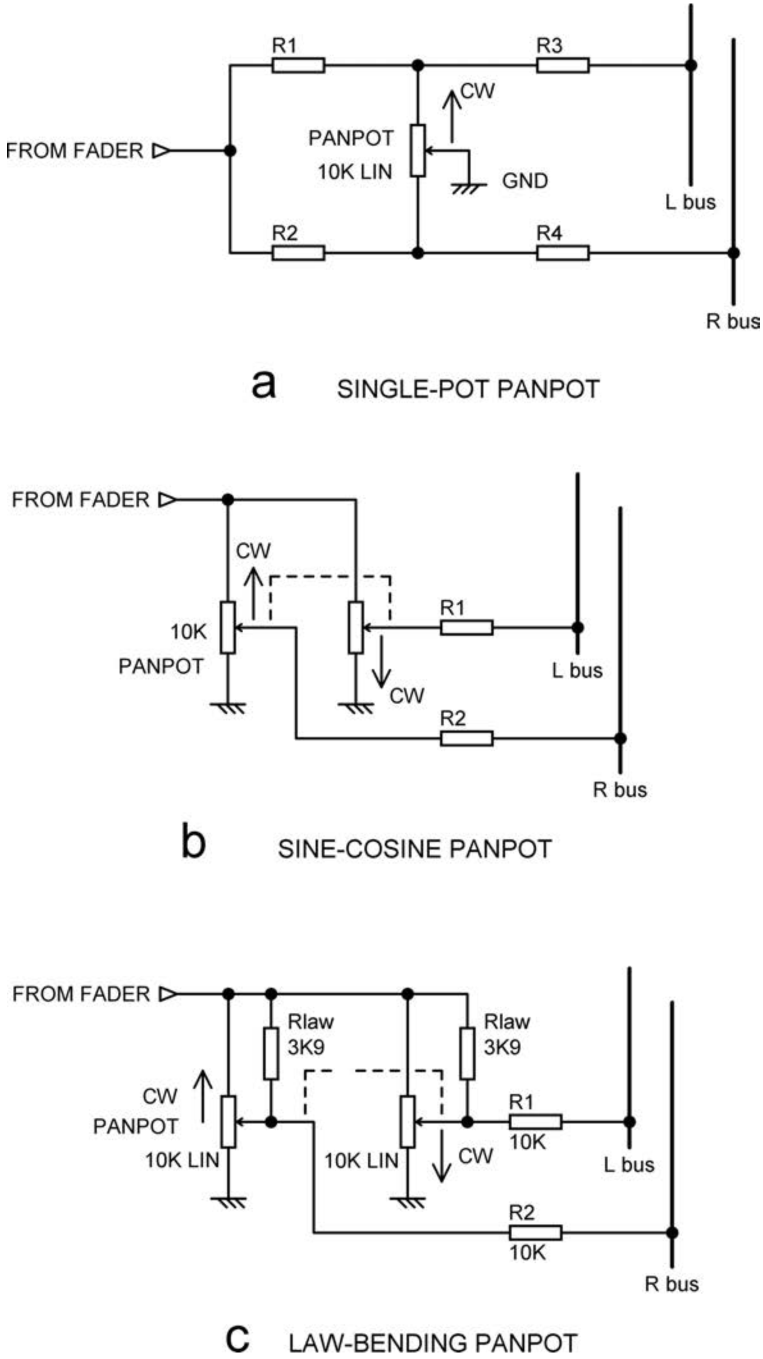

Figure 22.6: Three ways to make a panpot; c) uses a linear pot with law-bending resistors and is the most popular

There is an important factor in the design of panpot circuitry, and here a significant difference between low-cost and high-cost mixers intrudes. A low-cost mixer will drive the routing resistors directly from the panpot wipers, as shown in Figure 22.7a, and the panpot law is affected by the loading to ground represented by the routing resistors; they work in opposition to the law-bending pull-up resistors, which therefore have to be reduced in value (here to 3k3) to get the mid-point attenuation back to −4.5 dB. As the wiper loading increases and the value of the pull-up resistors is decreased with respect to the panpot track, the more the resulting law tends to flatten out around the central position, as shown in Figure 22.8. Table 22.1 shows how the law-bend resistors change with loading when a 10 kΩ pot is used, and the resulting pan laws can be seen in Figures 22.8 and 22.9.

The distorted law that results from excessive loading gives a panpot that does little as it is moved through the central position, and has an unpleasant 'dead' feeling; there are also level

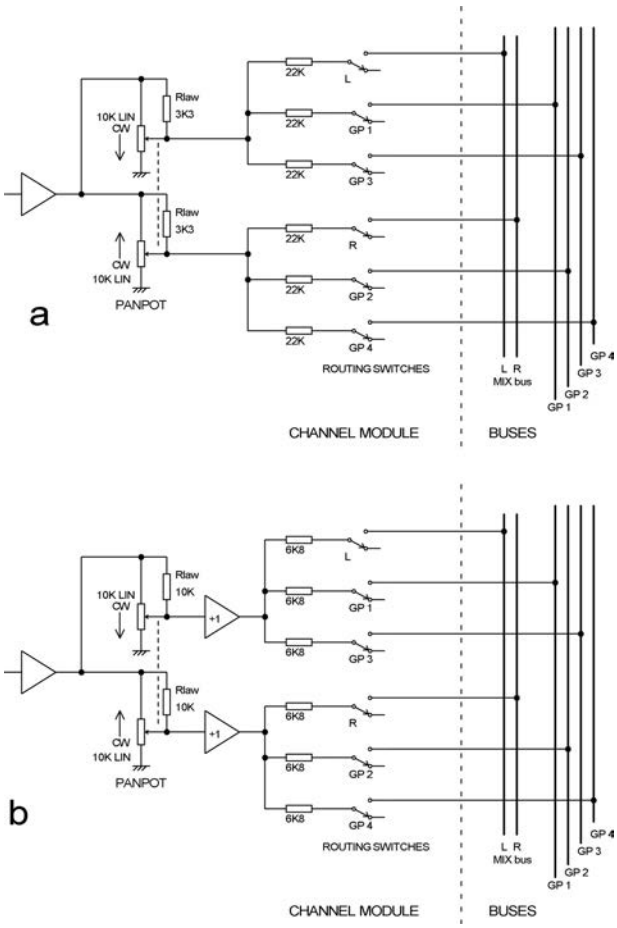

Figure 22.7: A panpot with law-bending resistors; unbuffered and buffered

TABLE 22.1 Wiper loading and pull-up resistor values in Figures 22.8 and 22.9

Trace in Figures 17.8, 17.9	Wiper loading	Pull-up resistor
1	15k	5k
2	4k	2.2k
3	3k	1.8k
4	2k	1.2k

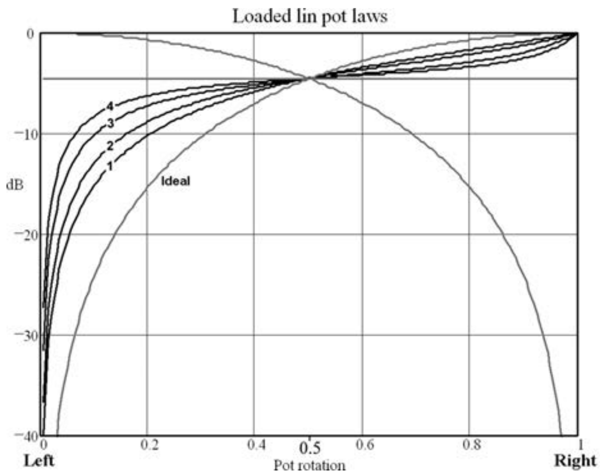

Figure 22.8: How the law of a panpot varies with loading on the wipers; in each case the law-bending resistors have been adjusted to give −4.5 dB at the centre. See Table 22.1

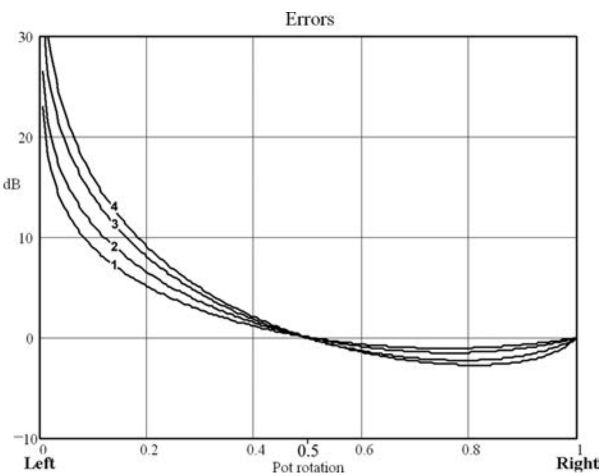

Figure 22.9: The difference between the ideal compromise law and the actual law that results from law bending. See Table 22.1

problems, as previously described. In addition, the lower the pull-up resistors, the greater loading they place on the stage upstream, which is usually the fader post amp; the loading is at its worst when the panpot is hard over to either side. This way of using a panpot presents a compromise on top of a compromise, and puts a limit on the number of mix resistors that can be driven directly from a panpot.

Figure 22.9 shows the differences between the ideal compromise law and the actual law that results from law bending, for various wiper loadings as in Table 22.1. The amplitude error of the right output gets enormous as the loading increases; when the panpot begins to move away from hard left the level shoots up much too quickly.

Since the value of the pull-up resistor required to get the best approach to the required law depends on the loading on the panpot wiper, this is a very good reason to use routing methods that place a constant load on the panpot, regardless of the number of buses that are being routed to, as this at least allows the value of the law-bending resistors to be optimised at one value. There is more on this later.

A more sophisticated approach is used in high-cost mixers, which puts unity-gain (or near-unity gain) buffers between the panpot and the loading of the routing resistors, as shown in Figure 22.7b. Here there are two points to note: the panpot wipers now have a negligible loading to ground, and the pull-up resistors therefore have a much higher value of 10 kΩ, which will improve offness as less current goes through the wiper contact resistance. Also, the routing resistor values can be reduced, as they no longer load the panpot, and this reduces the Johnson noise generated in the summing system.

This law-bending technique is a workable solution, and has been very widely used, but unfortunately the addition of the law-bending resistor introduces another problem. A reasonable quality pot has an offness of about −90 dB with reference to fully up, due to the end-of-track resistance. The pull-up current from the law-bend resistors, however, passes through the wiper contact resistance, which is usually greater than the end-of-track resistance, and this extra resistance severely limits the attenuation the panpot can provide when set hard left or right, degrading the offness of the panpot from approximately −90 dB to −65 dB when it is hard over. The problem is made worse because when the wiper is at the bottom of the track, the whole of the signal voltage across the pot is also across the law-bend resistor, which is lower in resistance than the pot track and so passes more current through the wiper contact resistance. The exact value of offness obtained depends on the component values and the construction of the pot. The way this works is shown in Figure 22.10.

This reduction in offness is not really a problem in stereo use, as an attenuation of −65 dB is more than enough to pan a sound so it is subjectively completely to one side (in fact about −20 dB will do that) but it is a very serious issue for mixers that route to groups in pairs,

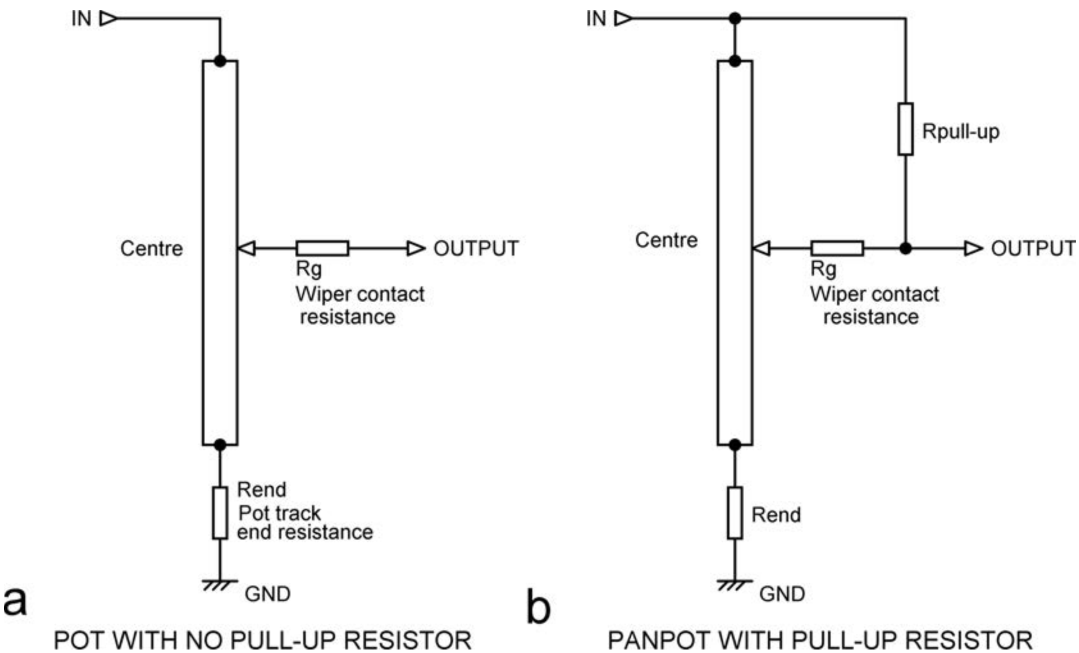

Figure 22.10: How the law-bending resistors degrade the offness of a panpot

or, in other words, have one switch to route to Groups 1 and 2, another switch for Groups 3 and 4, and so on. Simpler mixers route to groups in pairs, as this makes good use of two-changeover switches (and, as mentioned before, the channels are by a long way the most cost-sensitive part of a mixer) but it means that if you are, say, routing to Group 1 by pressing the 1–2 switch and then panning hard left, then the signal being sent to Group 2 will only be −65 dB down. In bigger and more expensive consoles each group has its own routing switch, with the spare switch section typically used for an indicator LED, so panpot offness is not such an important issue.

The active panpot

I think that it has made it clear that a conventional panpot is a bundle of compromises, and I decided in 1989 to do something about it. The active panpot was one of those ideas that strikes with such force that you remember exactly where and when it happened: in this case on the fifth floor of a tower block in Walthamstow.

The active panpot shown in Figure 22.11 works by replacing the simple pull-up resistor with a negative-impedance-converter that modulates the law-bending effect in accordance with the panpot setting, making a closer approach to the sine law possible. This sounds intimidating but is actually very simple. The wiper of the left half of the panpot is connected

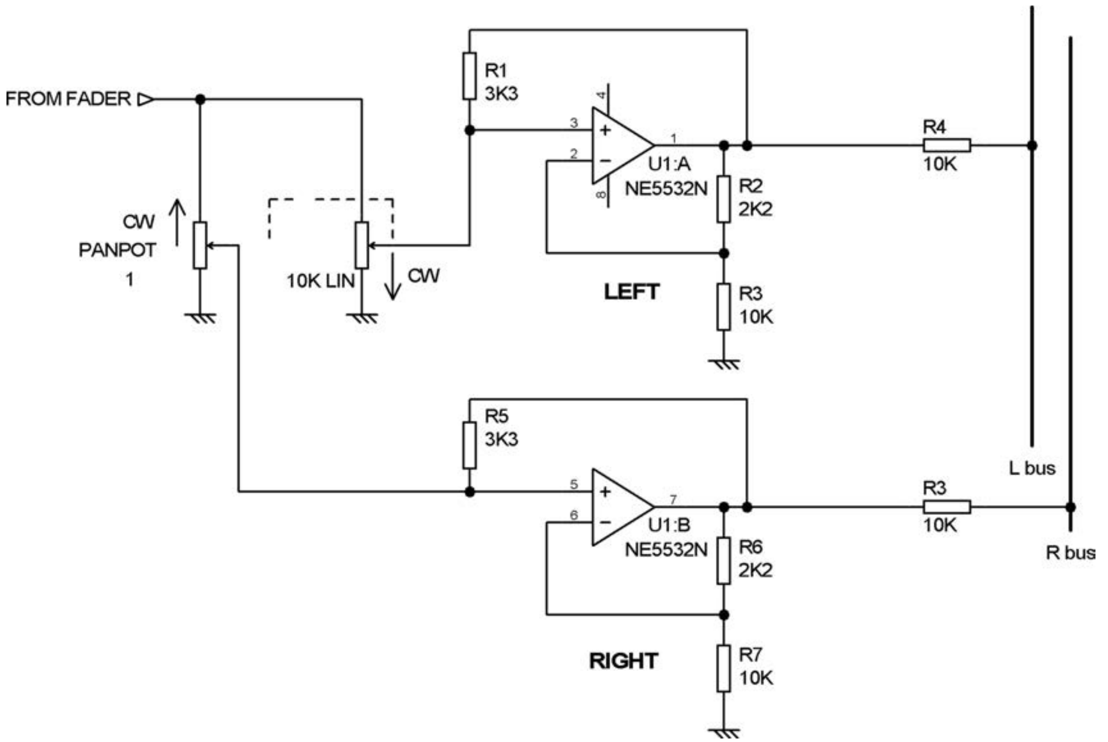

Figure 22.11: The active panpot: the law-bend resistor is driven from an amplifier stage with a gain of about 2 dB

to a series-feedback amplifier U1:A that gives a modest gain of about 2 dB, set by R2 and R3; the exact value shown in Figure 22.11 is 1.73 dB. The law-bending resistor R1 is driven from the output of this amplifier rather than the top of the panpot, and so, when the panpot wiper is at the lower end of its travel, there is little or no signal to the amplifier and therefore no pull-up action when it is not required. There is, therefore, no signal current flowing through the wiper contact resistance, so the left-right isolation using a good-quality pot is greatly improved from approx. −65 dB to −90 dB, being now limited solely by the end-of-track resistance as shown earlier in Figure 22.10a. As an extra benefit, if the amplifier gain and the law-bend resistor are correctly chosen, the pan law is much closer to the desired sin/sin^2 compromise law than can be obtained by the simple law-bending method described above; this is illustrated in Fig 22.12. This concept was protected by patent number GB 2214372 in 1989.

The active panpot is obviously a more expensive solution, as we have added a dual opamp and four resistors to every channel. However, we get other benefits for our money as well as better pan offness and a better pan law; if the dual opamp is a 5532 (which it should be to keep the postfade noise down) then its good drive capability means that the value of the routing resistors can be reduced, which reduces both capacitive crosstalk and noise in the summing system.

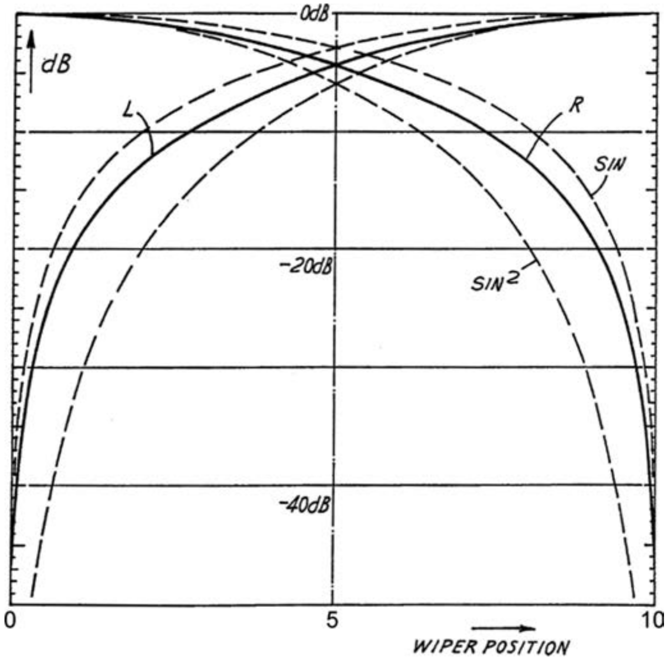

Figure 22.12: The active panpot: the resulting law is much closer to the desired sin/sin2 compromise. From Patent number GB 2 214 372

LCR panpots

A relatively recent innovation in the world of panpots is the LCR panpot, where LCR stands for left-centre-right. This has three outputs; the left and right in general working the same way as for a normal panpot. As the panpot is moved from extreme left, the centre output rises from zero, reaches a maximum with the panpot central, and then falls to zero again as the panpot reaches extreme right. The LCR panpot is used with speaker systems that have a central cluster as well as left and right speaker banks. This means that centrally panned signals are always heard as in the centre of the sound field, regardless of listening position, and this is useful when vocals or speech are combined with music, as typically occurs in religious installations. It is also used for panning to a discrete centre channel to create surround and 5.1 mixes. A switch is normally provided by which the LCR pot can be converted to conventional L/R stereo operation.

There are two possible ways to handle LCR panning. In what I call the ‘full-width’ method, as shown in Figure 22.13, the left and right panpot outputs work much as they do in a normal stereo panpot, only falling to zero at the extreme end of control rotation.

The alternative is the ‘half-width’ method in which both left and right outputs fall to zero at the centre setting, and stay there for further control rotation; this is shown in Figure 22.14.

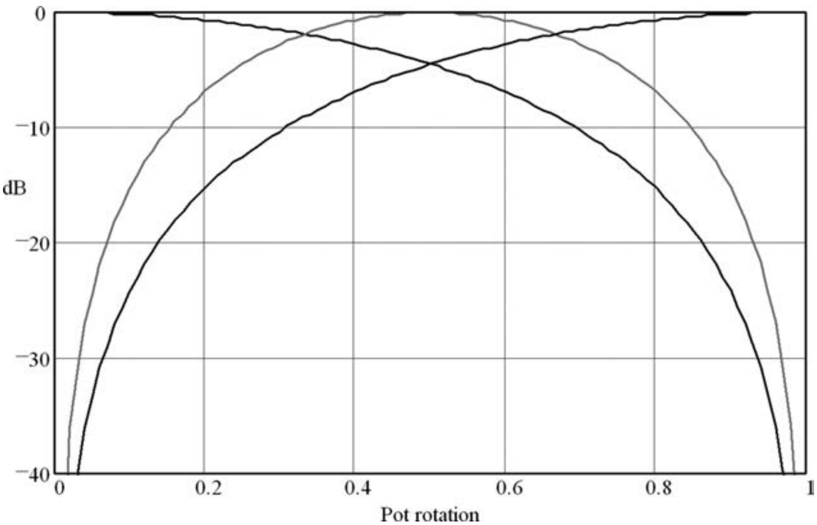

Figure 22.13: A full-width LCR panpot law. Left and right cross over at −4.5 dB in the centre as before, while C is 0 dB at the centre, but zero at each extreme

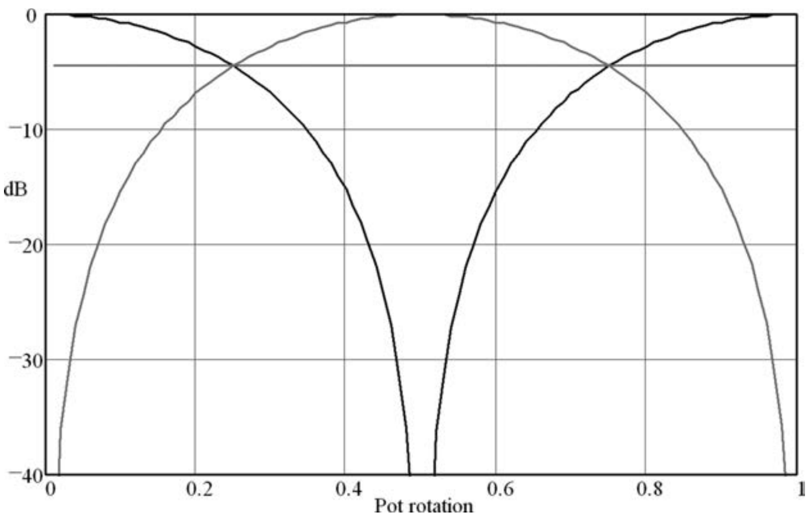

Figure 22.14: A half-width LCR panpot law. Left and right fall to zero at the centre setting and stay there for further control rotation. The crossover points between left and centre and right and centre are now at −4.5 dB

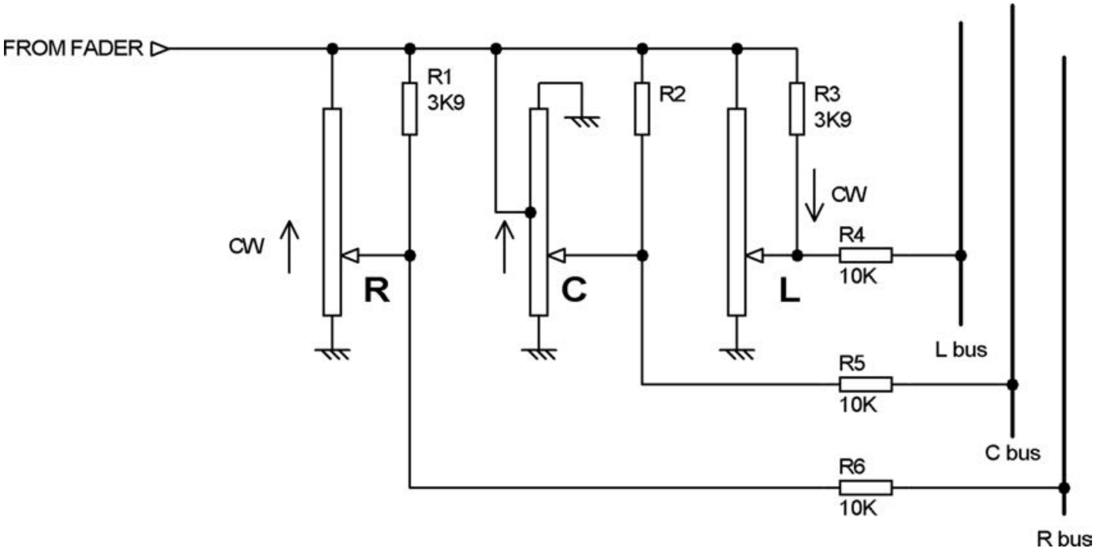

Figure 22.15: A full-width LCR panpot system

Both Figures show laws which are a sin/sin^2 compromise as described earlier, and are therefore 4.5 dB down at the centre – in the case of the half-width method these 'centres' are at pot positions of 25% and 75% rotation. Which LCR panning method is most appropriate depends on the application; some LCR panpot systems can switch between the full-width and half-width modes.

Figure 22.15 shows a full-width LCR panpot arrangement. The L and R sections are as for a normal L–R panpot, but the C pot track is grounded at each end and fed with signal via its centre-tap; an LCR panpot always requires specially-designed pots with centre-taps on the tracks. These are available from major manufacturers like Alps but they are inevitably made in small quantities and are relatively expensive.

Figure 22.16 shows a half-width LCR panpot. The L and R sections are now also centre-tapped (more expense!) and those taps are grounded so that the output falls to zero at the central control setting. The C pot track is grounded at both ends and fed with signal via its centre-tap as before.

LCR panpots, like ordinary panpots, give much superior performance in terms of offness and a better pan law when the active panpot system is applied to them. This is straightforward to do, and is arranged exactly as for the stereo version.

Consoles with LCR panning systems include the Soundcraft Series 5, the Yamaha M2500-56, the D&R Orion, and the Amek 9098i. True LCR panning is not common as it is a relatively high-cost facility.

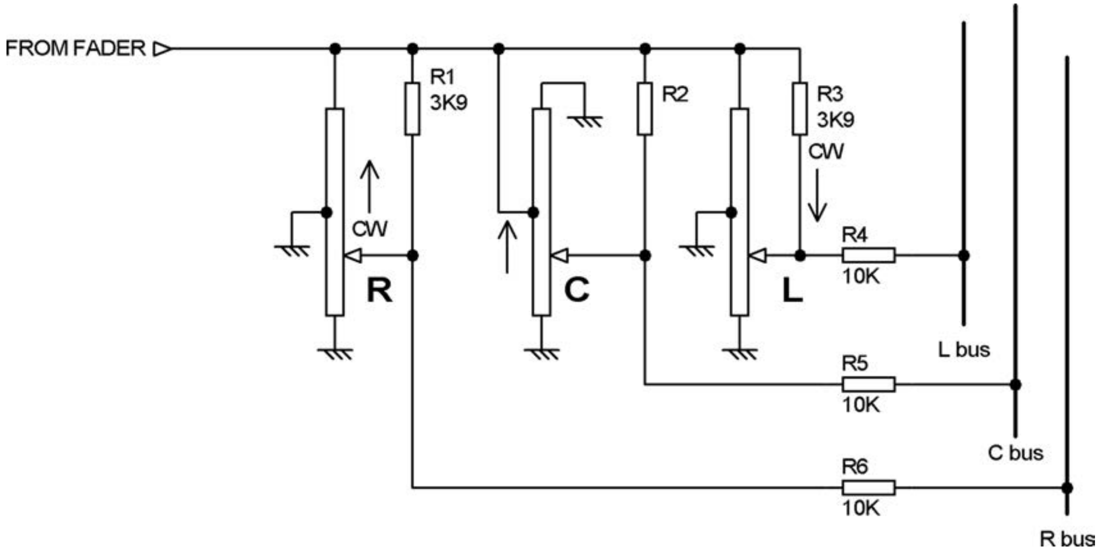

Figure 22.16: A half-width LCR panpot system. The L and R sections are now centre-tapped also

Routing systems

The routing system selects which mix bus the channel signal goes to. In a simple N into 2 stereo mixer the panpot does all the routing that is required, but as soon as more groups are introduced some sort of switching is required for bus selection.

Figure 22.17a shows the conventional routing system as commonly used up until about 1980. There are two problems with this method. Firstly, the capacitance between the switch contacts when they are open is significant, and this severely limits the crosstalk performance of the console. To get a feel for the problem, look at Figure 22.18 which shows the crosstalk performance of an ALPS SPUN two-changeover switch working into a mix bus with a 10 kΩ feedback resistor in the summing stage. This is a conventional push-switch with two parallel sections.

At 10 kHz the offness is only −66 dB, and at 20 kHz it is barely −60 dB. The problem is capacitive crosstalk across the switch contacts in the off position. Note that grounding the second switch section in the pious hope of improving things only reduces the crosstalk by 2 dB.

This problem can be completely eliminated by using the arrangement in Figure 22.17b, which is sometimes called the back-grounding system. The mix resistors are now connected to the channel ground when not in use and so there is no possibility of capacitive crosstalk – so long as the switches, of which one contact still carries signal voltage, are not too close to the mix buses. But this plan has a terrible drawback. The crosstalk performance is good, but back-grounding the mix resistors means that since relatively few routing switches are on at a time, most of the

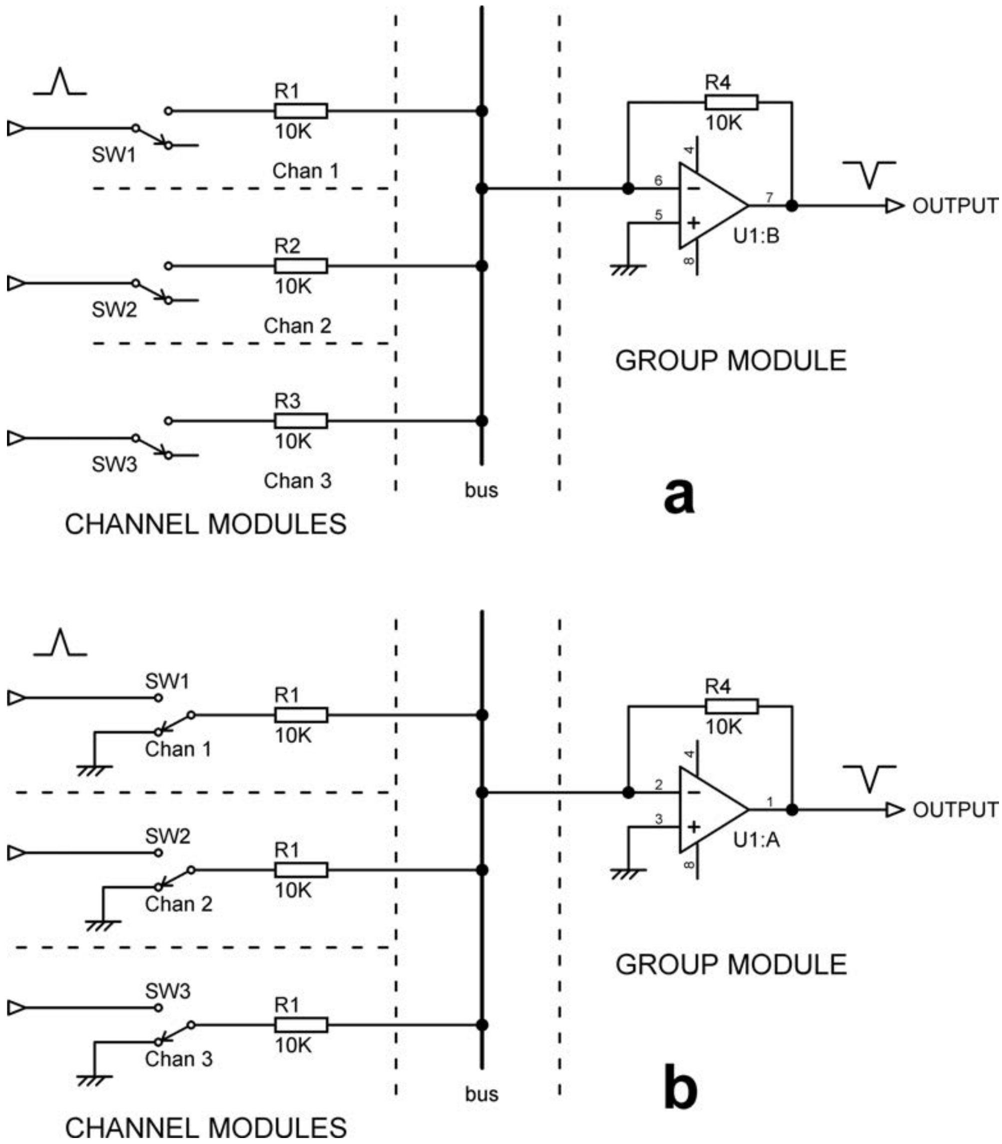

Figure 22.17: Two mediocre routing systems with virtual-earth summing, as used in the 1970s: a) suffers from poor offness, b) has bad noise performance

resistors are grounded and the summing amp is working at or near maximum noise gain, and always picks up the maximum possible ground noise.

It might be wise to take a moment here to explain the noise gain of a summing system, though this material really belongs in the next section on summing systems. A simple inverting amplifier with equal input and feedback resistors has a noise gain of 2 times or 6 dB,

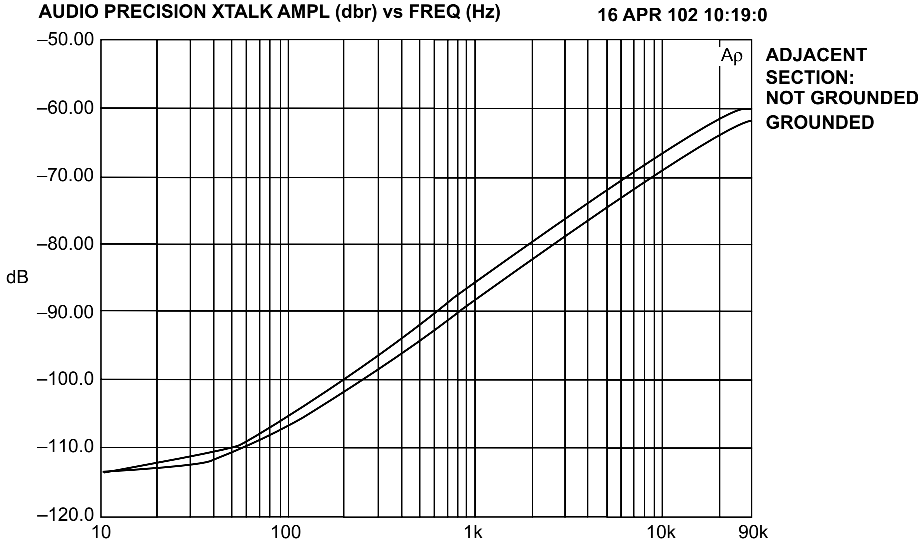

Figure 22.18: The poor high-frequency offness of a switch in a conventional routing system with a 10 KΩ summing feedback resistance

because the noise referred to the amplifier input sees effectively a non-inverting amplifier with a gain of two, and so the noise at the output is twice that of the noise-generating mechanisms at the amplifier input. The important point is that the noise gain is greater than the signal gain, which is unity.

If a second mix resistor of equal value is connected to the summing point, the amplifier input now sees effectively a non-inverting amplifier with a gain of three, and the noise gain is increased to almost 10 dB, though the signal gain via either mix resistor remains unity. The more channels that are routed to a mix bus, the worse the noise performance is, as summarised in Table 22.2. It is therefore clear that the arrangement in Figure 22.17b will be working at the worst-case noise gain all the time. Not only will the noise of the summing amp receive maximum amplification, but any ground noise in the console which puts the channel grounds at slightly different potentials will be picked up as effectively as possible. This is not a good plan, especially in large consoles.

Obviously a routing system that combined the advantages of the first two systems – good offness and minimum noise gain – would be a great improvement.

So, finally, here is the most satisfactory routing system. In the arrangement shown in Figure 22.19, the routing resistors are once more grounded when not being fed by the channel, but the topology has been turned around so that the grounded resistors are now connected to the channel rather than the mix bus. The switches rather than the routing

TABLE 22.2 Noise gain of a summing amp versus the number of mix resistors connected

Number of mix resistors	Noise gain (×)	Noise gain (dB)
1	2	6.02
2	3	9.54
4	5	13.98
8	9	19.08
12	13	22.28
16	17	24.61
24	25	27.96
32	33	30.37
48	49	33.80
56	57	35.12
64	65	36.26

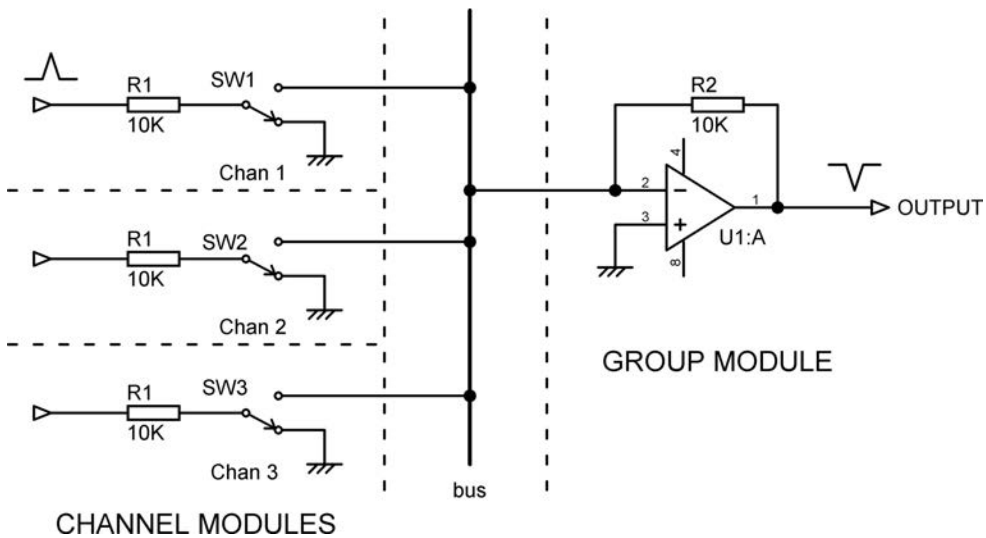

Figure 22.19: The routing system that I introduced in 1979 which gives both good routing offness and optimal noise performance. Virtual-earth summing system also shown

resistors are connected directly to the summing buses and, when a routing switch is not active, the feed from the routing resistor is grounded. So long as the hot end of the mix resistor can be kept away from the mix buses, the offness is truly excellent. With a bit of care in the physical construction, the offness can easily be better than −100 dB at 20 kHz. A possible objection is that all the grounded resistors put a heavier load on the channel circuitry upstream; this is quite true, but it has the countervailing advantage that since each mix resistor is either grounded directly or connected to a virtual–earth mix bus, the loading remains constant, and so if, for example, the resistors are driven from an unbuffered panpot, you will get a predictable panning law, if not necessarily a very good one. This routing method retains the advantage of working the summing system at the minimum noise gain.

I have a difficulty here. To the best of my knowledge I invented this system in the backroom of a shop in Leyton, London in 1979. I recall being nervous about the apparently iffy business of having the routing switch at virtual-earth, because of increased stray capacitance to ground from the virtual-earth bus, but in fact that caused no problems. However, since very little has been published about the details of mixer design, it being a proprietary art confined to a relatively small number of companies, it is not currently possible to say if I was the first to invent it.

The set of routing switches in each channel (often called the routing matrix) is relatively expensive and is, of course, multiplied by the number of channels. It is therefore common on all but the most expensive consoles for routing to be done in pairs; in other words, the first switch routes the channel to Groups 1 and 2, the second to Groups 3 and 4, and so on. This is simple to do as many varieties of push-switch come in a two-changeover format as standard. If routing to only one group is required then the signal has to be steered to the appropriate bus with the panpot. The offness of the panpot when set hard over is then crucial to obtaining good crosstalk figures; the active panpot scheme described above greatly improves this aspect of performance.

This system solves the electrical problems, but it relies on there being adequate physical separation between conductors carrying signal and the mix buses, to keep the capacitance between them low. This can present some interesting topological challenges. The obvious way to arrange the routing matrix is to put the switches next to the front panel, with the mix resistors behind them. Assuming we are using a double sided PCB, the tracks from the panpot to the mix resistors are on the other side of the board from the mix buses, but inevitably they have to cross over at some point, separated only by fibreglass. Back in the dark days when the cost penalty of double-sided PCBs was very serious, a single-sided PCB would be used with various convolutions of tracks and topside links used to get signals to the right places; this was inevitably a bundle of crosstalk compromises. Even a double-sided layout falls a long way short of perfection, as there is no screening between the panpot tracks and the mix bus tracks at the point where they cross. Keeping the track width to a minimum at this point reduces the capacitance between the tracks, but it remains the major limitation on crosstalk, i.e. 'routing offness'. If the routing resistors are through-hole, the pads and component leads on the same PCB side as the mix buses will also contribute to unwanted coupling.

The solution to this problem, introduced by me in the Series 6000 in 1987 at Soundcraft, was to reduce the size of the routing resistors from ¼ W to ⅛ W, which made them small enough to fit between the actuators of the routing switches, right at the front edge of the PCB. This placed the mix resistors well away from the rear of the switches, which was the furthest extent of the mix buses, and likewise the panpot tracks could be run along the very front of the PCB. The use of a double-sided PCB allowed a ground plane to be placed on the other side of the mix buses. This form of construction was very successful, giving a major improvement in routing offness, and was later used in the famous 3200 console in 1988. Nowadays the mix resistors would probably be metal-film SMD, and could be fitted in very tight spaces at the front of the routing switches.

Auxiliary sends

These are straightforward. There is often a pre/post switch before the send pot so the send signal can be taken from before or after the fader. In many designs, added flexibility is given in the form of push-on links so that other take-off points for the send signal can be used, such as before the EQ section. Changing all these links on a large console is naturally not a light matter.

If there is an on/off switch for the send then it should disconnect the mix resistor from the bus, just as with the group routing switches described above. This reduces summing noise dramatically if only a few auxes are in use, and gives an offness much better than that at the end of the aux pot travel.

Group module circuit blocks

Most of the circuit blocks used in group modules carry out the same functions as they do in the input channel modules. The great exception is, of course, the summing amplifier, which has a technology all of its own. Before diving into the details of practical summing amplifiers, it is advantageous to look at the various methods of performing that apparently simple but actually rather demanding task – adding signals together. The most fundamental function of a mixer, as its name suggests, is to combine two or more signals in the desired proportion. As with many areas of electronics, an extremely simple definition of the job to be done (addition – how hard can that be?) rapidly shows itself to have ever-deeper levels of complexity.

There are several intriguing techniques to look at:

Voltage summing

Virtual-earth summing

Balanced summing

Ground-cancel summing

Distributed summing

Summing systems: voltage summing

The earliest mixers used voltage summing, or passive summing as it is sometimes called, as shown in Figure 22.20; note that only half of each panpot is shown, for clarity. The great drawback with this system is that there is a significant voltage on the mix buses, so that a signal can be fed onto the bus by one channel and it can get back into another channel, as shown in the figure, from where it may, depending on the control settings, find its way onto another mix bus where it is definitely not wanted. This is demonstrated by the arrow A in the diagram. Channel 1 is feeding a signal onto Group 3 bus, and this is sidling back into Channel 2, which has both routing switches engaged and the panpot central, allowing the signal to turn round and get onto Group 1 bus. This sort of thing can, of course, be completely suppressed by suitable buffer stages after the panpot, and that is exactly what was done in the large consoles of the day. This means a lot of extra electronics, and this mixing system is not suitable for low or medium-cost designs. Another big snag is that since the mix buses have significant voltages on them, they must be carefully and expensively screened from each other to prevent capacitive crosstalk. Running the buses in a piece of low-cost ribbon cable is simply not an option.

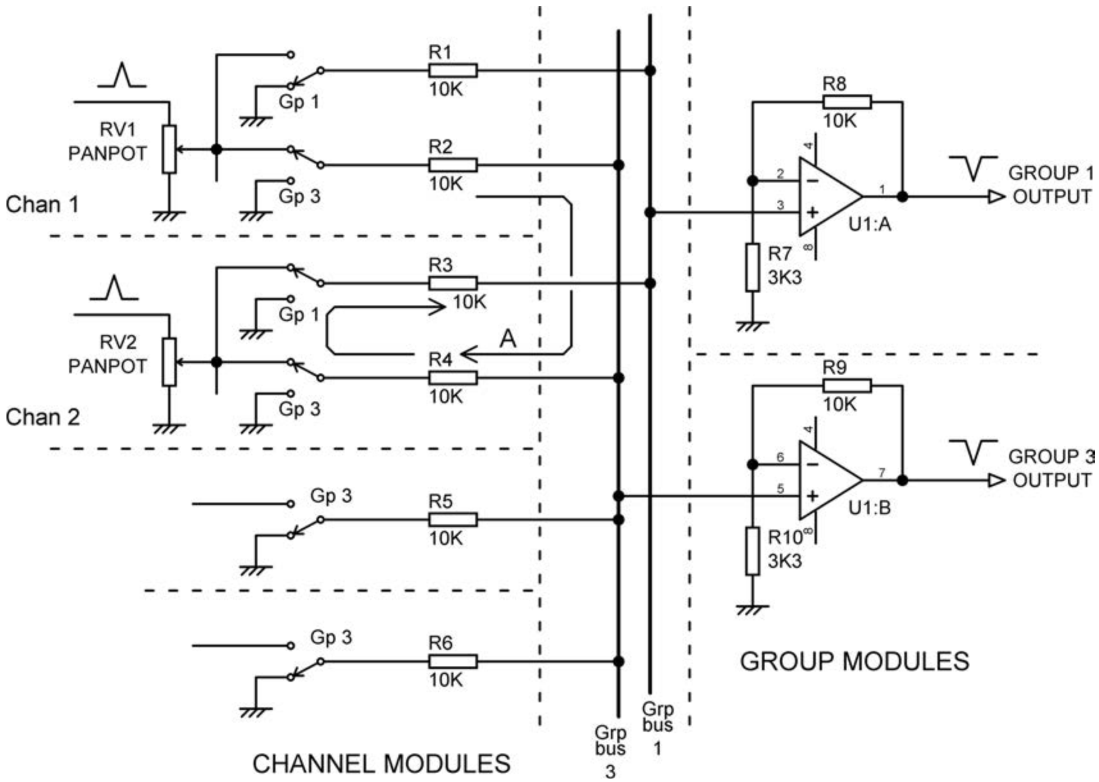

Figure 22.20: Voltage summing system as used in the 1960s

You will have noted that all the routing switches are back-grounded, so there is a constant impedance from the mix bus to ground; this is essential as otherwise the gain of the summing system would vary with the number of channels routed to a bus. This in turn means that the amplifiers that get the low level on the bus back up to the nominal internal level are working at substantial gain and are relatively noisy.

Summing systems: virtual-earth summing

Summing today is done almost universally by virtual-earth techniques, as shown above in Figure 22.19. An opamp or equivalent amplifier with shunt feedback is used to hold a long mixing bus at apparent ground, generating a sort of audio black hole; signals fed into this via mixing resistors apparently vanish, only to reappear at the output of the summing amplifier as they have been summed in the form of current. The great elegance of virtual-earth mixing, as opposed to the voltage-mode summing technique in Figure 22.20, is that signals cannot be fed back out of the bus to unwanted places as it is effectively grounded, and this can save massive numbers of buffer amplifiers in the input channels. The near-zero bus impedance also means that the gain does not alter as varying numbers of channels are routed to a bus. The appearance of the virtual-earth mixing concept was highly significant, as it essentially made low-cost mixing consoles viable.

There is, however, danger in being dazzled by this elegance into assuming that a virtual earth is a perfect earth; it is not. A typical opamp-based summing amp loses open-loop gain as frequency increases, making the inverting input null less effective. The 'bus residual' (i.e. the voltage measurable on the summing bus) therefore increases with frequency and can cause inter-bus crosstalk in the classic situation with adjacent buses running down an IDC cable.

As we saw in the last section on routing, a virtual-earth summing system operates at quite a high noise gain when many inputs are routed to it. Minimising the noise is therefore of prime importance, and one obvious step is to keep the impedance of the summing system as low as possible to reduce Johnson noise in the mix resistors. I have been involved in more mixing console designs than I care to contemplate, of all sizes and types, and it is interesting to recall that the mix resistors were always in the range of 22 kΩ to 4.7 kΩ, which seems like a rather narrow range of 4.7 to 1. There is a reason for this. 22 kΩ mix resistors are used in small budget consoles because a small number of them (say six) can be driven directly from a panpot without distorting its law unacceptably. Now consider a big expensive 32-group console where the panpot is buffered, but it has to drive 16 mix resistors on each side with the buffer, and making them of very low resistance would present enormous loading. The mix resistors will therefore be something like 6.8 kΩ; the value has only been reduced by a factor of about three, and Johnson noise will only be reduced by the square root of this, or 5.3 dB.

Mix resistors lower than this are only used for the critical L-R mix buses, where there are only two to be driven, and 4.7 kΩ is a common value. It actually only gives an advantage of 1.6 dB over 6.8 kΩ. The lowest mixing resistors I am aware of are in fact 4.7 kΩ, and this value was used in the Neve VR consoles, which in their largest formats had a possible 144 inputs going to the L-R mix bus (72 modules each with two paths to the mix bus).

This consideration of resistance demonstrates the difficulty of reducing the Johnson noise beyond a certain point by reducing the summing impedance. Halving the mix resistors doubles the drive power required, but only improves the noise by a factor of the square root of two. Johnson noise is of course only one factor; there is also the noise from the summing amplifier itself. The current-noise component of this is also reduced by reducing the bus impedance.

Apart from the noise performance, another advantage of a low summing system impedance is that it makes the mix buses less sensitive to capacitive crosstalk. Reducing mix resistors from 22 kΩ to 6.8 kΩ may only give a 5.3 dB improvement in noise, but it gives a 10 dB reduction in capacitive crosstalk.

Balanced summing systems

As a console grows larger, the mix bus system becomes more extensive, and therefore more liable to pick up internal capacitive crosstalk or external magnetic fields, which are poorly screened by the average piece of sheet steel. The increased physical size means longer ground paths with non-negligible signal voltage drops across them. An expensive, but thoroughly effective, answer to this is the use of a balanced summing system. The basic idea is shown in Figure 22.21; only half of the channel panpot is shown, for clarity.

Each mix bus is now double, having a hot (in-phase) and cold (out-of-phase) bus, exactly like the hot and cold wires in a balanced line connection. These run physically as close together as possible. At the channel end, each side of the panpot has an inverting buffer A2 that drives a second set of mix resistors that feed the cold mix buses. Two sections are now required for

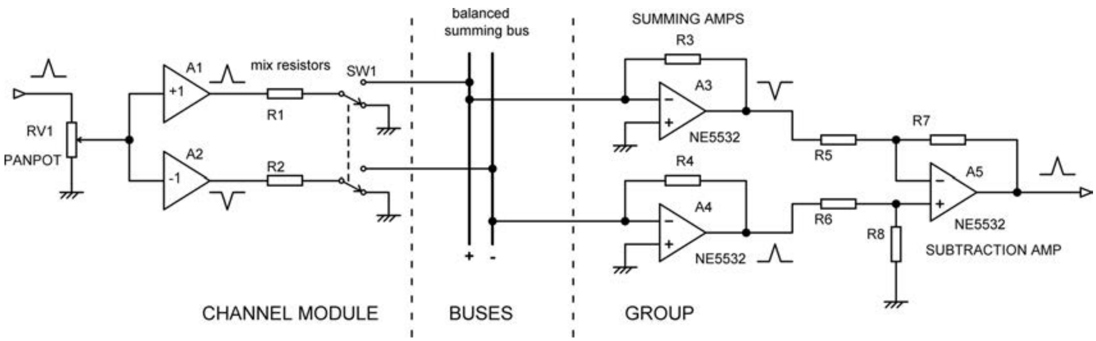

Figure 22.21: A balanced summing system

the routing switch SW1, which is fine as most push switches are in a two-changeover format, but if you want switch status indication, which you normally will on a console large and sophisticated enough to employ balanced summing, four-changeover switches are needed – lots of them – and the cost jumps up. Each bus connects to a summing amplifier A3, A4 and their outputs are subtracted by A5 to cancel the common-mode signals.

A quite separate advantage of a balanced summing system is that it has a better noise performance, because having two buses gives you 6 dB more signal but only 3 dB more noise, because the noise from the two summing amplifiers is uncorrelated and partially cancels. This gives you a 3 dB noise advantage that it would be difficult to get by other methods. Against this must be counted the extra noise from the channel inverters.

Because of the cancellation of capacitive crosstalk that occurs, the offness of the routing matrix can be made very good indeed. On the last console of this type that I designed the offness was a barely measurable −120 dB at 20 kHz.

As I just noted, probably all consoles of this type will have routing switch status indication and so use four-changeover switches. This leaves a spare switch section doing nothing, and if you have followed this book so far you will have gathered it is not in my nature to leave a situation like that alone. In the last balanced console I designed, this fourth switch section was used to detect when *no* routing switches were active; this happens more often than you might think on a large desk. In this condition an FET switcher removed the signal feed to the buffers driving the mix resistors. This prevented a large number of grounded resistors being pointlessly driven, reducing power consumption significantly, and further improving routing offness.

It may have occurred to you that a considerable simplification would result if the channel inverters were done away with and the cold resistors simply connected to ground. The signal level in the summing system would be 6 dB lower, exacting a 3 dB noise penalty, but the rejection of ground voltages and the cancellation of capacitive crosstalk would be just as good. This is not a sound plan. Much of the cost of balanced summing lies in the doubled number of mix resistors, routing switch contacts, and mix buses. The saving on inverter cost is relatively small. The large currents flowing in the mix resistors are now unbalanced and are much more likely to cause troublesome voltage drops in ground connections.

A variation on this approach can however be very useful in specific applications; see the section on ground-cancel summing systems below.

Ground-cancelling summing systems

Ground-cancel summing systems are very useful in auxiliary send systems where all sends are routed permanently to the bus. It gives some of the advantages of a balanced summing system at a tiny fraction of the cost, and it is particularly useful in consoles where the modules are connected together by ribbon cabling with a relatively high ground resistance. The basic principle is shown in Figure 22.22, which represents a single channel with two aux sends. The ground potential of each channel is read by a low-value resistor R3, and summed into the GC bus; this is not a virtual-earth bus as such, but it is at a very low impedance and is essentially immune to capacitive crosstalk. The GC bus is then used as the ground reference for the aux summing system, and by subtraction it gives excellent rejection of the channel ground potentials. If a channel is at the remote end of the ribbon cable, then its ground resistance is relatively high and the local channel ground will carry a large signal potential, which is fed into the GC bus; conversely a channel at the near end of the ribbon cable will have a much lower ground potential, also fed into the GC bus. The end result is that good aux offness figures are obtained for both those channels. In such a system, the aux offness is defined not by the grounding arrangements but by the offness of the aux send pot itself, typically −90 dB with reference to fully up.

A couple of design points. The GC resistors R3 should be very low in value so that their Johnson noise contribution is negligible; the only requirement is that they are large with respect to the ground resistances to allow accurate subtraction. R6 is a safety resistor fitted in the aux master module so that it will keep working if the GC bus becomes disconnected. Without it the entire aux send system is dependant on the single connection A between the bus structure and the aux master module, and such a situation is not good practice. The only requirement is that R6 is high with respect to the total GC bus impedance so that the subtraction remains accurate. If an aux send has an on/off switch that disconnects the mix resistor from the bus, to minimise noise gain, the switch must also disconnect the GC resistor to maintain correct cancellation.

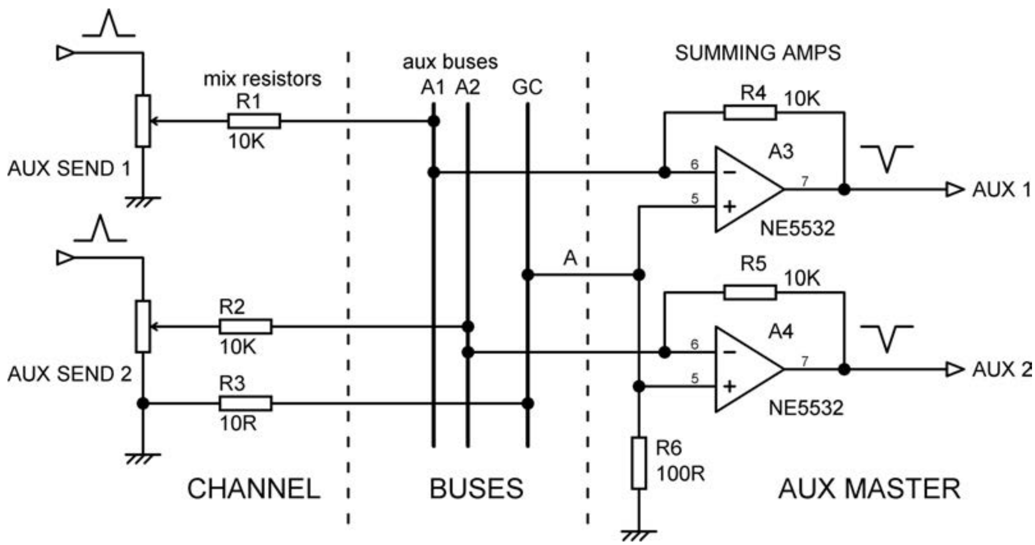

Figure 22.22: A typical ground-cancelling summing system for aux sends

In some ribbon-cable consoles the grounding is reinforced by heavy (32/02) ground wires with push-on connectors plugged into every channel module, or in some cases every fourth or fifth module. Even in this case the use of a ground-cancel summing system will normally give a worthwhile improvement.

Distributed summing systems

Distributed summing (sometimes called devolved summing or devolved mixing) is a deeply cunning way of improving the noise performance of a summing system, with reference to the noise from the summing amps themselves. In what follows, the noise from an input channel and Johnson noise from the summing resistors are for the time being ignored.

In a distributed summing system, the contributory signals are summed in two separate stages. Thus if you are summing together 24 channels, you might sum them in blocks of eight, to get three sub-groups which are then summed together to get a single output, as shown in Figure 22.23a. Unlikely as it may seem, this two-layer summing system gives a definite noise advantage. Assume that every summing amplifier creates the same amount of input-referred voltage noise V_n; the result of this at the amplifier output then depends only on the noise gain at which it is working. If we sum our 24 channels into one in the normal virtual-earth manner, the output noise will be 25 V_n, and that is our reference case. If we sum eight inputs together, the output noise will be only 9 V_n; three of these sub-mix outputs are now summed together, and you may have seen this coming – the signals sum arithmetically but the summing amp noise partially cancels, so the combined noise output is 9 V_n times $\sqrt{3}$, which is 15.6 V_n. To this we must add the noise of our second summing amp; this is working at a noise gain of only four, and so its own output noise contribution is 4 V_n. Adding the uncorrelated 15.6 V_n to 4 V_n rms-fashion gives a total of 16.1 V_n, which is 1.55 times or 3.83 dB quieter than the straightforward all-in-one-go summation of 24 inputs into one. Clever, eh?

The downside to this is of course that a bit more hardware is required – we are using four summing amplifiers rather than one. This is however a trivial extra expense compared with the electronics involved in 24 input channels. A more serious potential problem with this approach is that there is now a hidden layer of summing amplifiers which might clip without there being any indication on the console control surface.

You are by now, I hope, pondering if there is anything special about the 24 into 3 into 1 (24:3:1) structure, or whether other variations might be better; the answer is 'yes'. Three different possibilities are shown in Figure 22.23. Summing the 24 inputs in batches of six into four sub-mixes and then summing them to one (24:4:1) gives an advantage of 4.51 dB, while summing four at a time into six sub-mixes and then summing to one (24:6:1) gives an advantage of 4.97 dB, and this is the optimal solution for 24 inputs.

The results are summarised in Table 22.3, which shows that there is not much difference between 24:6:1 and 24:8:1, except that the latter uses a bit more hardware and is a negligible

TABLE 22.3 The noise advantage of two-layer distributed summing with 24 inputs

No. of inputs	No. of sub-mixes	Noise advantage (dB)
24	2	2.56
24	3	3.83
24	4	4.51
24	6	4.97
24	8	4.76
24	12	3.53

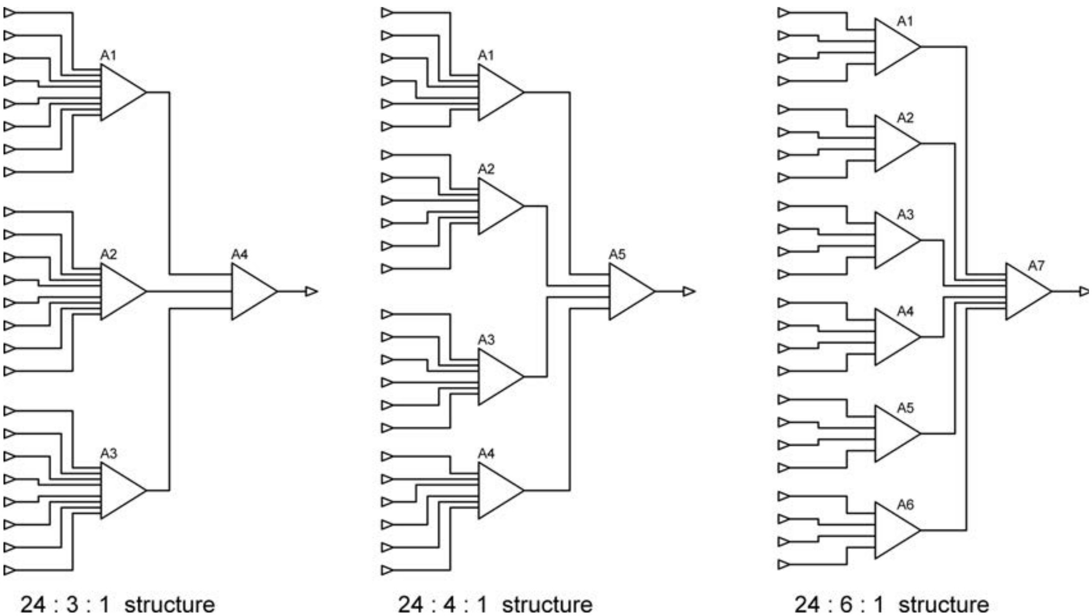

Figure 22.23: Three different distributed summing schemes for 24 inputs. 24:6:1 is the quietest

0.2 dB noisier. It is hard to see why anyone would want to use the 24:12:1 structure, because of all those extra summing amps, but it is interesting to note that it still gives a theoretical noise improvement of 3.53 dB.

Table 22.4 illustrates the noise advantages with various forms of two layer summing for 32 inputs; the maximum advantage is now a very useful 5.88 dB, and once again the best results come from first summing the inputs in batches of four. There are fewer options this time because 32 has less factors than 24.

Table 22.5 shows the optimal structures giving minimum noise for various numbers of inputs; at first it looks as if there is something special about doing the first layer of summing in batches of four inputs, but as the total number of inputs reaches 72 the optimal number suddenly starts to rise.

The figures in Table 22.5 do show a rather unsettling irregularity; for example the optimal structure for 80 inputs is no quieter than that for 72 inputs. This seems to be because some numbers of inputs have more factors than others, and so give a greater choice of structures to choose from.

TABLE 22.4 The noise advantage of two-layer distributed summing with 32 inputs

No. of inputs	No. of sub-mixes	Noise advantage (dB)
32	2	2.68
32	4	4.94
32	8	5.88
32	16	4.01

TABLE 22.5 The optimal two-layer summing structures for various numbers of inputs

Total no. of inputs	No. of inputs in each sub-mix	Optimal structure	Noise advantage (dB)
16	4	16:4:1	3.64
24	4	24:6:1	4.97
32	4	32:8:1	5.88
48	4	48:12:1	7.09
56	4	56:14:1	7.52
64	4	64:16:4	7.88
72	6	72:12:1	8.48
80	8	80:10:1	8.48
88	8	88:11:1	8.84
96	6	96:16:1	9.43
104	8	104:13:1	9.46
112	8	112:14:1	9.73
120	6	120:20:1	10.13
128	8	128:16:1	10.21
136	8	136:17:1	10.43
144	6	144:24:1	10.67

Two-layer distributed summing is particularly adapted to consoles built in sections, typically as bins of eight or 12 modules. It avoids virtual-earth buses extending over large physical areas, which makes them more susceptible to magnetic fields and ground voltage-drops, and can cause difficulties with HF stability. Each bin can now be connected by balanced low-impedance line connections, with the final summing done at whatever point is convenient. Both Neve and Focusrite have produced consoles with distributed summing systems in the past.

As I'm sure you have appreciated by now, I am a great one for following a train of thought until it derails at the points and scatters passengers all over the track. If two layers of distributed summing give a better noise performance, *would three layers be even better?* The answer is generally, yes, but not by much; basically it's not worth it. For example, the optimal two-layer structure 24:6:1 gives a noise advantage of 4.97 dB, while the optimal three-layer structure 24:8:2:1 gives 5.24 dB; the improvement over two layer summing is a very small 0.27 dB. Noise performance is actually very slightly *worse* for the 16-input case, because you have added more summing amps creating noise without enough partial-cancellation going on to outweigh that.

See Table 22.6 for the optimal three-layer structures and their results. (Note that in the three-layer structures the second figure denotes the *total* number of first-layer summing amps rather than the

TABLE 22.6 The optimal three-layer summing structures for various numbers of inputs

Total no. of inputs	Optimal structure	Noise advantage (dB)	Advantage over 2-layer summing (dB)
16	16:4:2:1	3.57	20.07
24	24:8:2:1	5.24	0.27
32	32:8:2:1	6.24	0.36
48	48:16:4:1	7.99	0.90
56	56:14:7:1	8.33	0.81
64	64:16:4:1	9.06	1.18
72	72:18:9:1	9.27	0.79
80	80:20:5:1	9.97	1.49
88	88:22:11:1	10.00	1.16
96	96:32:8:1	10.74	1.28
104	104:26:13:1	10.59	1.13
112	112:28:14:1	11.33	1.60
120	120:40:8:1	11.62	1.49
128	128:32:8:1	11.87	1.66
136	136:34:17:1	11.51	1.08
144	144:36:9:1	12.35	1.68

number feeding each summing amp in the second layer.) The greatest advantage over two-layer distributed summing is 1.7 dB, and that is for a monster mixer with 144 paths to the mix bus; getting that improvement involves using 36 + 9 = 45 summing amplifiers, when the optimal two-layer process uses 24. That's almost twice as many, and we must regretfully conclude that three-layer distributed summing is of marginal utility at best.

Four-layer summing anyone? Probably not; you would need a quite enormous number of inputs to make such a structure even faintly worthwhile.

Summing amplifiers

Figure 22.24 shows a practical summing amplifier circuit using a 5534/2 opamp; the accompanying insert send amplifier is included because it gives a good example of how the summing amp phase inversion needs to be corrected before the signal sees the outside world again, and it also demonstrates how to apply the 'zero-impedance' approach to an inverting stage. As mentioned before, the summing amplifiers were one of the first places that 5534/2s appeared in mixers, because the potentially high noise gain makes their noise performance critical. The circuit is very straightforward but there are a few points to watch. Firstly, the DC-blocking capacitor C1 is essential to prevent the bias current of the 5534/2 from making the routing switches clicky. It needs to be bigger than might at first appear because the relevant LF time constant is not C1 and the mix resistor R1 in the channel, but C1 and the parallel combination of all the mix resistors going to that bus. Thus for the 22K mix resistors and 100 μF capacitor shown, if there are eight input channels the LF −3 dB point will be at 0.59 Hz. Table 22.7 illustrates how this works, and you can see that in this case 100 μF is large enough for even a big mixer. If lower values of mix resistor are used then C1 must be scaled up proportionally.

It may appear that C1 is still rather oversized for the job. This is not so. The signal in a mixing console passes through a large number of RC time constants on its journey from input

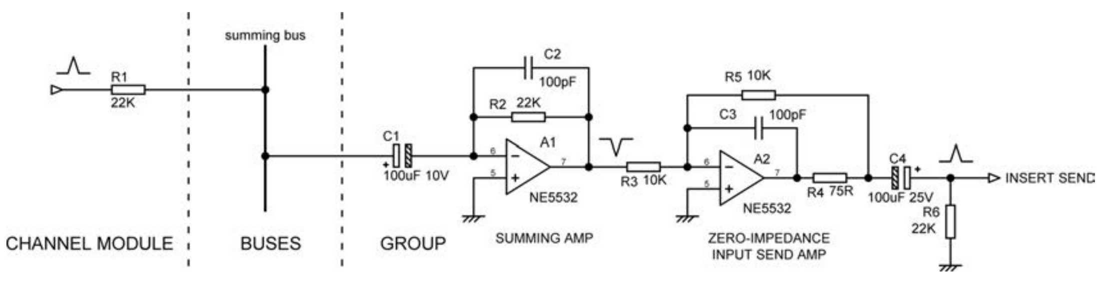

Figure 22.24: Simple opamp summing amplifier, followed by an inverting zero-impedance insert send amplifier that corrects the phase

TABLE 22.7 How the LF frequency response of the summing amp in Figure 22.24 varies with the number of mix resistors connected to the bus

No. of mix resistors	Total mix res to bus (Ω)	−3 dB frequency (Hz)
1	22K	0.072
8	2.75K	0.59
16	1.375K	1.18
24	917	1.74
32	687	2.33
40	550	2.90
48	458	3.44

to output, and even if each one causes only a tiny roll-off at the LF end, the cumulative effect on the frequency will be substantial. It is therefore essential that each LF roll-off is well below the normal audio band. The same considerations apply to HF roll-offs.

There is also the important point that electrolytic capacitors distort with even small signal voltages across them, as described in Chapter 2, and so they must be much larger than might otherwise be required.

There is yet another reason to make C1 large. A summing amplifier aspires to appear as a short-circuit to ground, and even if it is itself perfect, the presence of C1 means that the bus residual voltage will rise at low frequencies, possibly causing increased LF crosstalk.

The capacitor C2 across the summing amp feedback resistor R2 is another vital component; without it the stage is pretty well guaranteed to be unstable at HF. This is because the mix bus has an appreciable capacitance to ground, and this will destabilise the opamp by adding an extra phase shift to the feedback through R2. Adding C2 compensates for this; its value is normally determined by experiment. It clearly must not be so large that it gives an appreciable HF roll-off with R2.

A balanced summing system as in Figure 22.21 above can be easily made using two separate summing amplifiers whose outputs are then subtracted. A more economical method which saves an opamp is shown in Figure 22.25; with equal resistor values the overall gain is −6 dB. The downside is that it does not generate two true virtual earths. The two bus residuals are only low if symmetry holds in every part of the circuit, including the signal levels applied to the summing resistors. I built it with a 5532 and 1% resistors and the bus residuals were both −61 dB (worsening above 1 kHz as usual due to opamp limitations). Changing either of the summing resistors or feedback resistors by 3% of their value increased this to −49 dB. Obviously these are not true virtual earths – more like

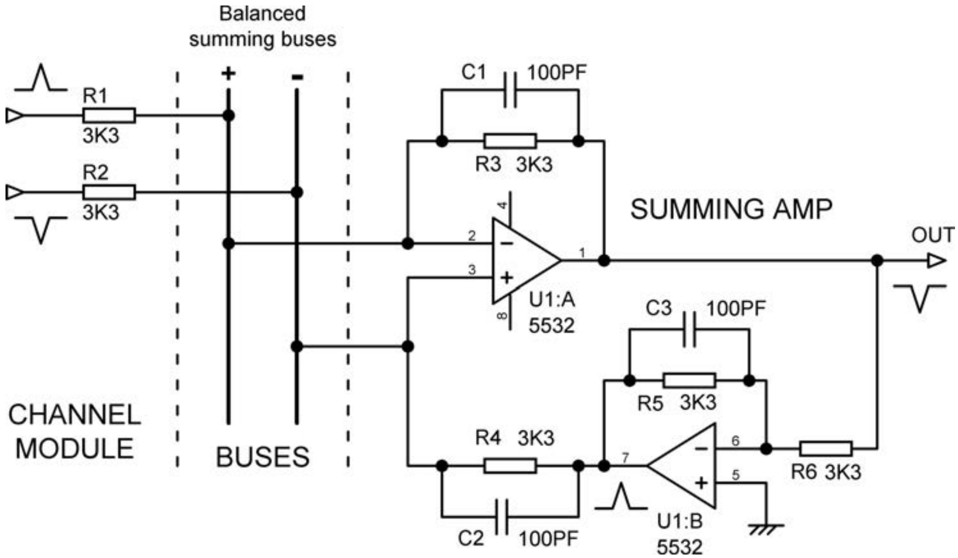

Figure 22.25: Balanced summing amplifier using only two opamps. A low bus residual requires complete symmetry of component values

virtual-virtual-earths. This method was used for the main mix bus in the Mackie Onyx mixer, among others.

Hybrid summing amplifiers

It is well-known that at low source resistances, discrete bipolar transistors are normally quieter than opamps, the exception being specialised opamps like the AD797, which are usually ruled out on the grounds of cost. This applies to the normal run of low-cost transistors, though specialised types such as the 2SB737 are even better. While the advantage obtained varies with the number of summing resistors connected to the bus, and their value, in general a very desirable noise reduction of about 5 dB may be expected in a large console.

The hybrid combination of a discrete bipolar transistor input and an opamp to provide open-loop gain for linearisation and load-driving capability gives the arrangement shown in Figure 22.26. The summing bus is connected to the emitter of Q1 via the DC-blocking capacitor C1; the amplified signal from Q1 collector is passed to the opamp, which has a local feedback loop R3 which controls the DC conditions of Q1. The voltage set up on the non-inverting opamp input by bias network R7, R9, C5 determines the voltage at Q1 collector. C7 acts as a dominant-pole capacitor to give HF stability.

An outer layer of shunt feedback via R4 is used to minimise the effect of C1 in increasing the bus residual at LF, by putting C1 inside this outer feedback loop. C4 prevents the non-zero

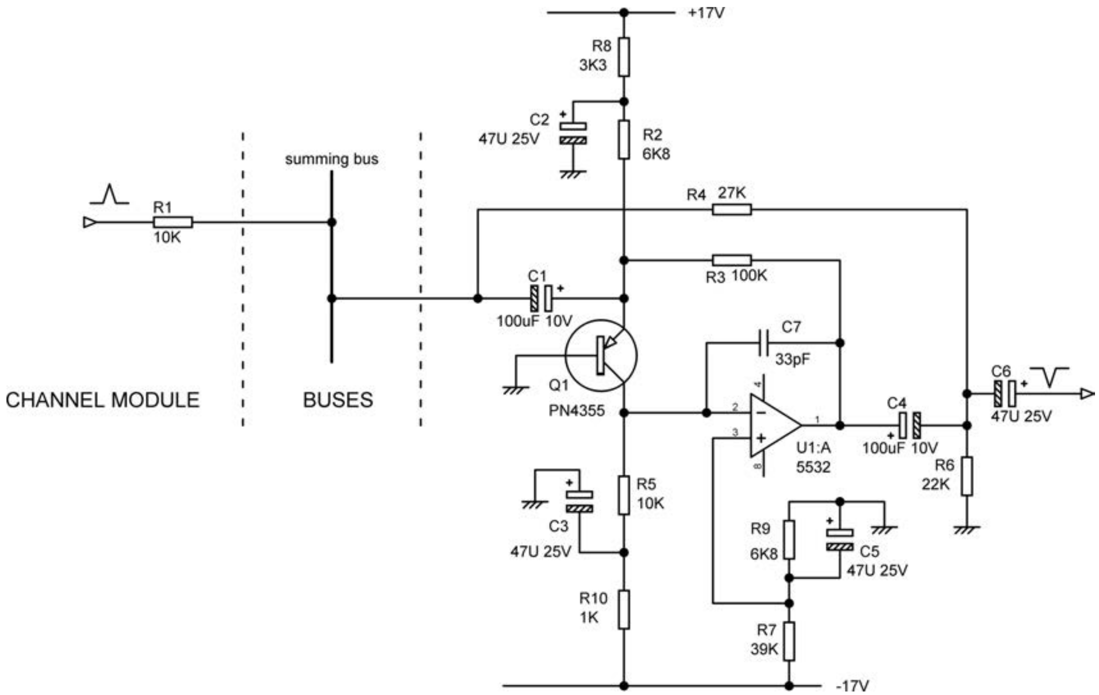

Figure 22.26: One-transistor hybrid summing amplifier

voltage at the opamp output from getting onto the bus via R4, and in the same way C6 is a DC blocking capacitor which prevents the following stage from putting any DC offset voltages on to the bus. This arrangement is effective at doing what it is supposed to, but there are clearly several LF time-constants involved because of the various blocking capacitors, and with such schemes it is essential to check that there is no peaking or other irregularity in the frequency response at the bottom end.

While the one-transistor version gives a good noise performance, it is susceptible to rail noise getting into the emitter or collector circuits of the transistor, and very careful filtering and decoupling is required to prevent this putting a limit on the noise performance. Another potential problem is that there will be LF signal-related voltages on the supply rails, due to the limited effectiveness of decoupling capacitors of reasonable size, and if these get into the summing amplifier they will compromise the LF crosstalk performance. An excellent fix for these issues is the use of a two-transistor circuit as in Figure 22.27. The two transistors are used in a balanced configuration that cancels out rail noise and makes the decoupling requirements much simpler; there is also no need for a biasing network as in the one-transistor version.

There are one or two subtleties to be observed here: the shunt feedback must go to the same transistor as the VE bus, to prevent large current swings in the transistors that would cause

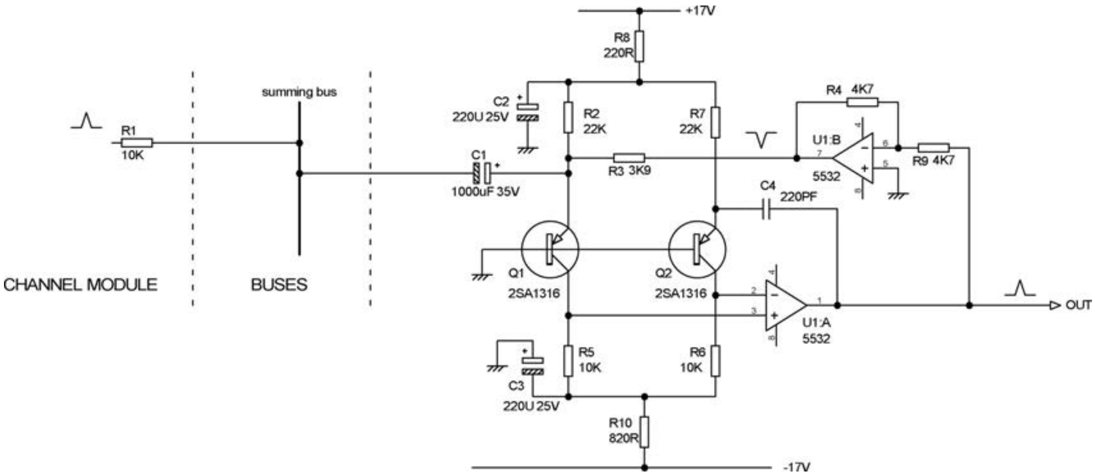

Figure 22.27: Two-transistor hybrid summing amplifier is better at rejecting noise and crosstalk from the supply rails

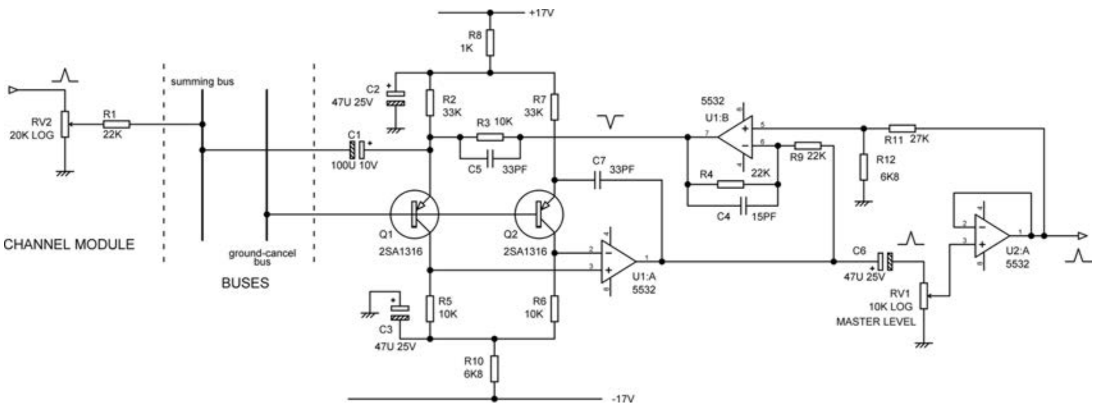

Figure 22.28: Two-transistor hybrid summing amplifier with positive feedback to enhance gain at high control settings only, improving headroom at low gain settings

distortion. This means that the feedback must be inverted to be in the right phase, which is done by U1:B. This introduces stability-threatening extra phase shifts, so HF stability capacitor C4 bypasses the feedback inverter and goes straight to the emitter of Q2. A very useful property of this arrangement is that the output is in phase and so can be fed directly to a group insert.

A more sophisticated version of this summing amplifier, which I thought up a while ago, is shown in Figure 22.28. Here the aux master level control RV1 not only acts as a normal gain control, but also alters the amount of negative feedback around the summing amplifier, effectively giving the stage variable gain. This can be extremely useful, as aux summing amplifiers often have large numbers of inputs feeding them and are more liable to clipping than group summing amplifiers. It is extremely tedious to turn down dozens of aux sends to remedy the situation, so being able to reduce the aux summing amp gain is advantageous.

When RV1 is at minimum, the circuit is essentially that of Figure 22.27, with the main shunt feedback path through the inverter U1:B, and the HF stabilising path through C7. The value of the main feedback resistor R3 is chosen so the sum amp gain is low, in this case −6.8 dB, and overload is unlikely. When RV1 is turned up, a positive feedback signal is sent to the non-inverting input of U1:B via R11 and R12, and the partial cancellation that results reduces the output of the inverter, decreasing the overall amount of negative feedback and increasing the gain of the summing amplifier. The need for a post amplifier with make-up gain is therefore avoided, improving the noise performance.

This technique was used in the Soundcraft Delta mixing console, and I modestly propose that it might be one reason why it won a British Design Award in 1991. My collaborator in the design of the Delta was Mr Gareth Connor.

Balanced hybrid summing amplifiers

In sophisticated consoles, it is desirable to combine the benefits of balanced mixing with low-noise hybrid summing amplifiers. The obvious method of implementing this is to use two hybrid summing amplifiers and then subtract the result, but this uses quite a lot of hardware. A more elegant approach is to use a single balanced hybrid summing amplifier to accept the two antiphase mix buses simultaneously; this reduces noise as well as minimising parts cost and power consumption.

The circuit shown in Figure 22.29 is at first sight very similar to that in Figure 22.27 above, but with the crucial difference that two mix buses now feed into the stage, one to each transistor emitter. To prevent distortion–inducing variations in the transistor currents, and to maintain symmetry, it is therefore now necessary to also apply shunt feedback to both transistor emitters. The feedback to Q1 is taken via an inverter to get the phase right, as in Figure 22.27, but the other feedback path is simply resistor R12. As before, C4 gives HF stabilisation.

Note that the configuration is very similar to that of the balanced microphone amplifier described in Chapter 17, and gives low noise as well as excellent symmetry. This technology was used extensively in the Soundcraft 3200 recording console; once again, my valued collaborator in the design of the 3200 was Mr Gareth Connor.

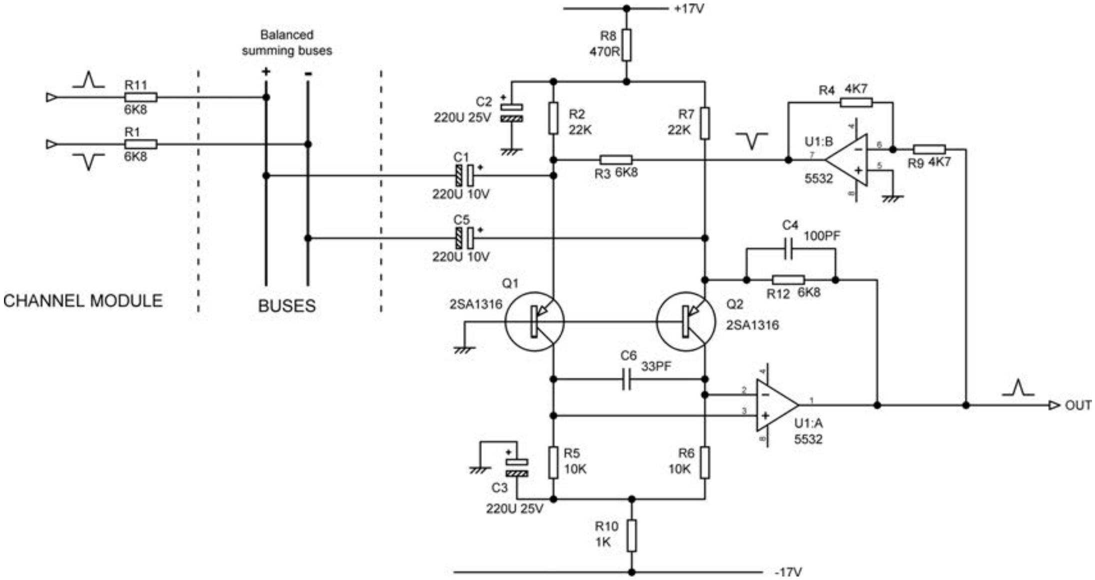

Figure 22.29: Two-transistor balanced hybrid summing amplifier. Note the dual feedback paths via R3 and R12

PFL systems

The prefade listen (PFL) facility goes back a long way in the history of mixers, almost certainly first appearing on broadcast consoles, where it proved extremely useful to be able to listen to a source before unleashing it on an unsuspecting public. It is also very handy in a recording environment, allowing you to listen to one source in isolation without undoing dozens of routing switches. It is invaluable for checking the level in a channel; the PFL feed takes over the main mix metering, which is usually much bigger and better than metering incorporated on a per-channel basis, if indeed there is any at all. On most mixers there is only room for a peak light. As the name suggests, a PFL feed is taken from before the fader, so that the signal is heard at full level even if the fader is down. In the USA it is often called the ‘solo’ facility.

‘PFL’ is often used as a verb, as in ‘Could you PFL channel 23?’. Interestingly, this request is always spelt out as P-F-L and the obvious pronunciation as ‘piffle’ has never achieved favour. When any PFL switch is pressed, a circuit block in the master section switches the left and right monitor outputs so they reproduce the PFL signal in mono, as shown in Figure 22.30. The L/R meters are also taken over by the PFL system so they can be used to check channel prefade levels, allowing the adjustment of input gain when required. It is common to switch

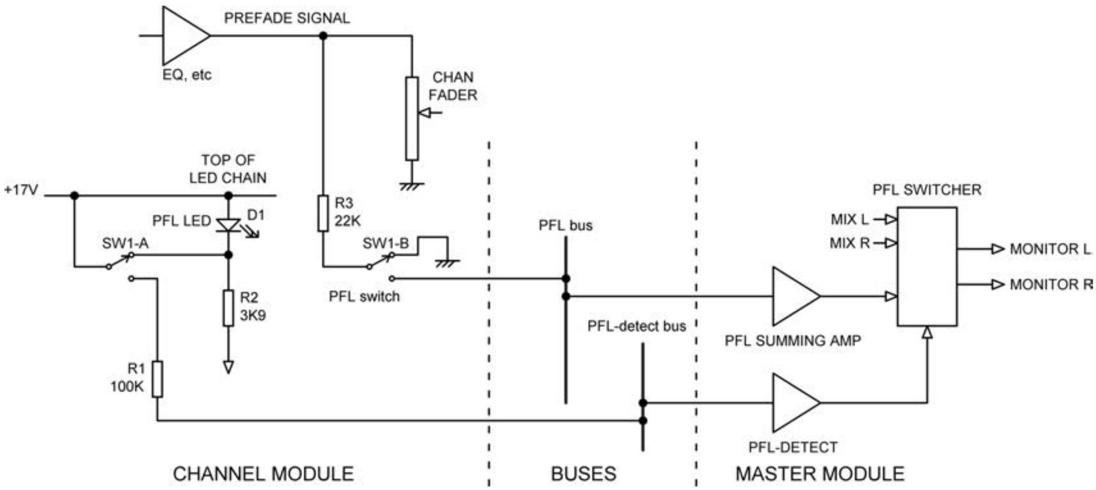

Figure 22.30: A PFL system; showing how to make a two-section PFL switch perform three functions at once

both meters over, though of course they both read the same thing. Historically the PFL switching was done with a relay, but this was taken over by electronic means, specifically 4016 analogue gates, at a relatively early date, as it is not essential for the switching to be absolutely click-free. The later availability of discrete FETs designed for analogue switching improved the linearity and much reduced (but did not wholly eliminate) the switching transients.

All that is required on each channel is a switch section to route the prefade-signal to the PFL bus, which is a simple virtual-earth bus that works in the same way as group buses, and another switch section to signal via a PFL-detect bus that a switch has been pressed and the monitor source must be changed over. The PFL-detect bus simply gathers together all the signals that indicate a PFL switch is pressed; it does not have to be a virtual-earth bus, but there are advantages in making it so, as described below. PFL switches are commonly fitted with indicator lights, even on quite small mixers, because otherwise it is necessary to hunt around the control surface if a PFL switch is inadvertently left pressed in. This does not make a third switch section (and therefore a four-changeover switch) necessary because the ingenious method shown in Figure 22.30 can be used. All the LEDS on a channel module are usually run in a chain from the positive to negative supply rails; so long as there are not too many LEDs a simple series resistor is usually all that is needed to give more-or-less constant-current operation of the chain. If the PFL indicator LED is put at the top of the chain, SW1-A removes the short-circuit from D1 and allows it to turn on, at the same time routing a DC signal into the PFL-detect bus via R1. (Note that the switch section SW1-A that routes the signal to the PFL bus uses the same high-offness configuration described earlier.)

This underlines the point that economy of design in a mixer channel is very important, because every extra component is multiplied by the number of channels. In the master section the need to design out every possible component is less pressing.

PFL summing

There is nothing very difficult about the PFL summing amplifier. It is not necessary for it to have the best possible noise performance as it is only used for quick monitoring checks. The system shown in Figure 22.31 uses a dual opamp, the second half being used to restore the correct phase after the inversion produced by the summing amplifier. C1 provides DC blocking, while R2 keeps the PFL bus at 0 V DC to prevent switch-clicks. R3 sets the gain from channel to PFL summer as unity, and C2 prevents the bus capacitance from causing instability. R4 and R5 set the gain of the re-inverting stage to unity. C3 provides DC blocking at the output to prevent clicks when the PFL switcher operates.

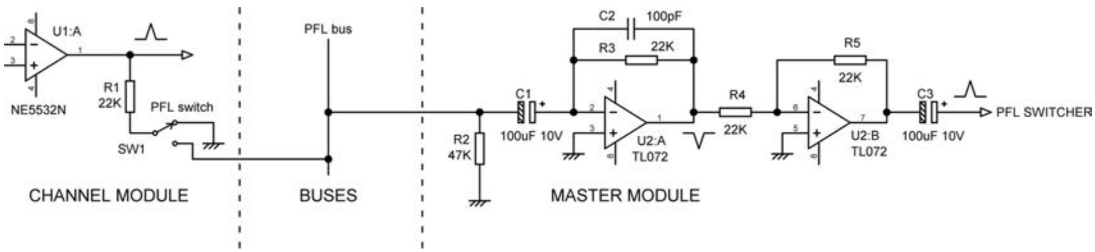

Figure 22.31: An example of a PFL summing system

PFL switching

When the PFL system operates, the normal L-R mix is removed from the monitor outputs and is replaced by a mono feed from the PFL summing system. This is usually done by some form of electronic switching (see Chapter 21 on Signal Switching for more details).

PFL detection

In early consoles, a DC voltage was simply switched into a PFL-detect bus that operated the PFL switching relay directly. Moving that amount of current around inside a mixer is simply asking for trouble, with clicks crosstalking into the summing buses, and very soon a transistor relay driver was added so the voltage and current levels on the PFL-detect bus could be much lower. When solid-state PFL switching was introduced the transistor circuitry drove analogue gates or discrete FETs instead.

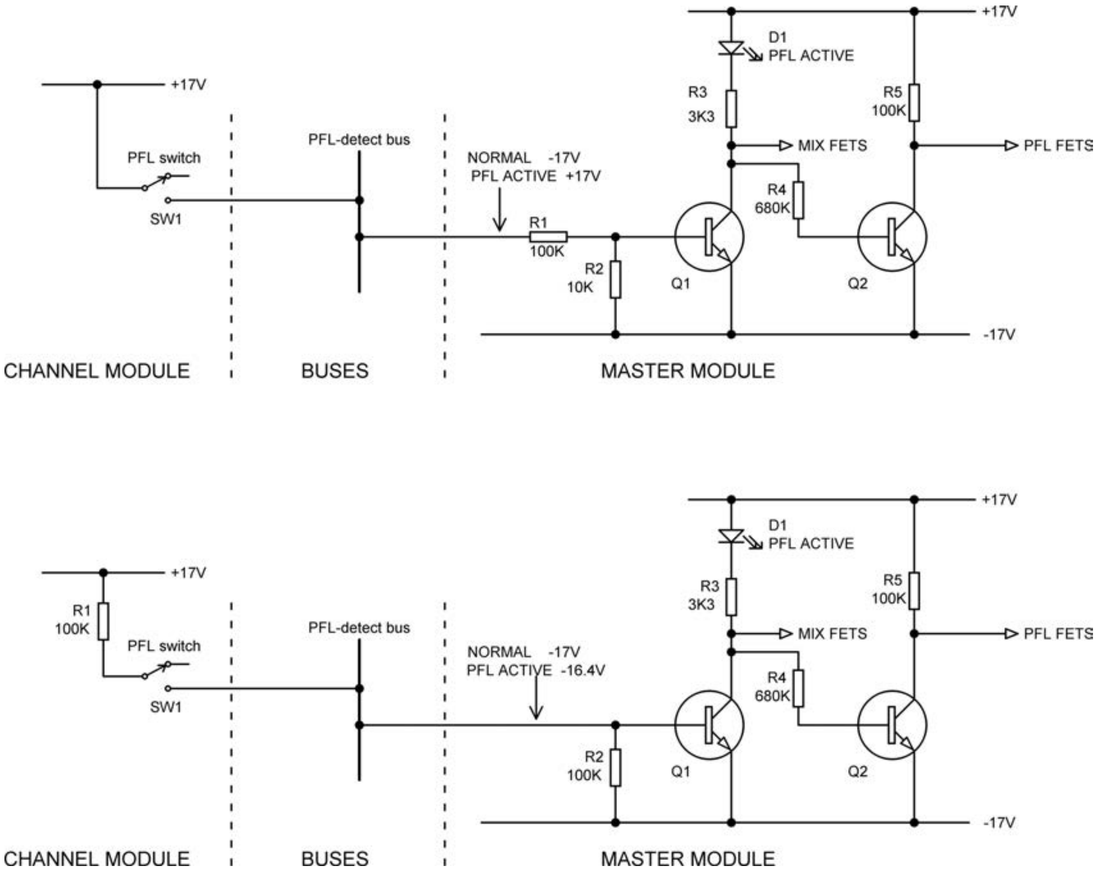

Figure 22.32: Two simple PFL-detect systems. The second version reduces the likelihood of clicks getting into other buses

Two simple methods of PFL detection are shown in Figure 22.32. At 22.32a, how not to do it is demonstrated. When the PFL switch SW1 is closed, +17 V is applied to the PFL-detect bus, turning on Q1 and illuminating the 'PFL ACTIVE' LED D1, and turning Q2 off. The problem is that the 34 V edge on the PFL-detect bus (which normally sits at −17 V) has an excellent chance of crosstalking into the other summing buses.

A better method is shown in Figure 22.32b. Here the resistor R1 has moved to the channel module, and the PFL-detect bus connects directly to the base of the detect transistor Q1. This ensures that the PFL-detect bus can never change by more than 0.6 V, which reduces the possibility of clicks crosstalking capacitively into adjacent summing buses; there should be a 35 dB advantage. A countervailing disadvantage is that the base of Q1 is vulnerable to destruction if the PFL-detect line gets shorted to ground. It is also slightly more expensive to implement as now there has to be a resistor R1 for each PFL switch.

This method was workable when all mixers had motherboards, but the introduction of ribbon-cable interconnections between mixer modules put all the buses closer together, and increased the likelihood of PFL-detect clicks becoming audible.

Virtual-earth PFL detection

The ideal PFL detection system would have no voltage edges on the detect bus at all. This can be done by detecting current instead of voltage – the PFL-detect line becomes a virtual-earth summing bus for DC signals. I thought this up in 1983.

Figure 22.33a shows the principle. U1 is simply a summing amplifier that works at DC. When no PFL switches are pressed, the output of U1:A sits at +10 V because of the bias from resistor R2, Q1 is held on via R4, and Q2 is held off. Negative feedback through R3 keeps the detect bus at 0 V; C1 prevents the bus capacitance to ground from causing instability. The 'PFL ACTIVE' LED has moved to the collector of Q2 to allow for the logical inversion caused by the summing amplifier. When a PFL switch is pressed, the current injected into the detect bus by R1 overcomes that from R2 and the output of U1

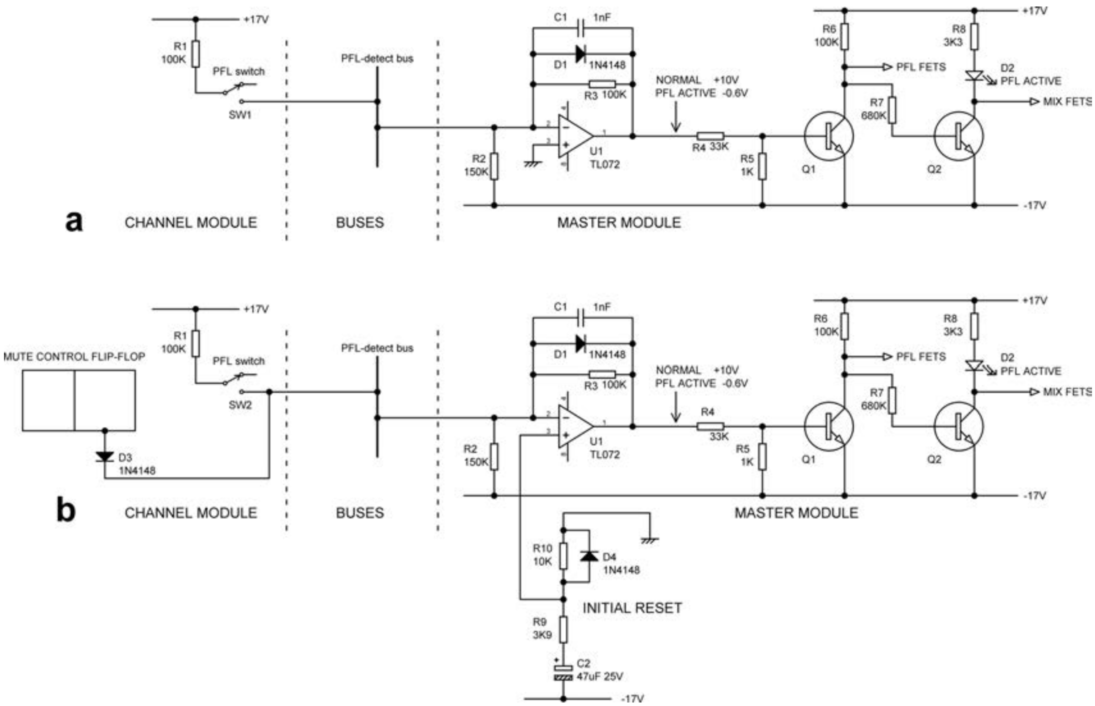

Figure 22.33: Virtual-earth PFL-detect systems. The second version uses the PFL-detect bus to reset channel mute status at power-up

goes below 0 V, and Q1 turns off. Note that the signalling current is only 170 μA, so there is little possibility of inductive crosstalk. It could probably be significantly reduced without problems.

This simple version of the system will silently detect a single PFL button down, but will come unstuck if more are pressed, because U1 output will hit the −V rail and will no longer be able to maintain a virtual earth on the detect bus. This situation is not uncommon, as it gives a rough submix of the channels pressed, and so it is wise to cater for any number of PFLs at once. The circuit of Figure 22.33a overcomes the problem by adding diode D1, which clamps the opamp output so that however many PFL switches are pressed, the opamp output will not hit the −V rail and negative feedback will continue to keep the PFL-detect bus very close to zero volts. The values of R4, and R5 must be carefully chosen so that the level-shift down to the −V rail works reliably and Q1 will always switch correctly despite circuit tolerances.

Note that, once again, all the circuitry works between the +V and −V rails, so there is no chance of transients being injected into the ground system.

This arrangement has another less obvious advantage. It can handle signals in both directions. The ribbon-cable systems referred to earlier come in a number of fixed widths, so you can buy 40-way cable, but if you need 41 ways the next size up is 50-way. Since this increases the cost not only of the cable itself, but of all the connectors on it and all the mating module connectors, increasing the cable size is not a decision taken lightly. In one mixer design there was a need to send an initial-reset signal to mute-control flip-flops on each module, and all 40-ways of the ribbon cable were committed. The only way to avoid 50-way cabling was to make the PFL-detect bus dual-function.

The idea is that by changing the reference voltage applied to the PFL-detection opamp, you can change the DC voltage that it maintains on the detect-bus. In Figure 22.33b, at power-up the initial-reset network R9, R10, C2 pulls the reference voltage going to the non-inverting input of U1 low for a brief period until C2 fully charges, and the PFL-detect bus is therefore also pulled low via R3, and this signal is coupled to each channel mute-control flip-flop by a diode D3; the flip-flop works between positive and negative supply rails so it can drive muting FETs directly. After the initial-reset period the PFL-detect bus settles at its normal level of 0 V and D3 remains reverse-biased. D4 in the initial-reset network ensures rapid recovery when the power is switched off. Obviously the PFL system cannot function during the brief initial-reset period, but after that it works just as described before.

This is a good example of the ingenuity that is sometimes required to fit all the required functionality into a given size of ribbon cable.

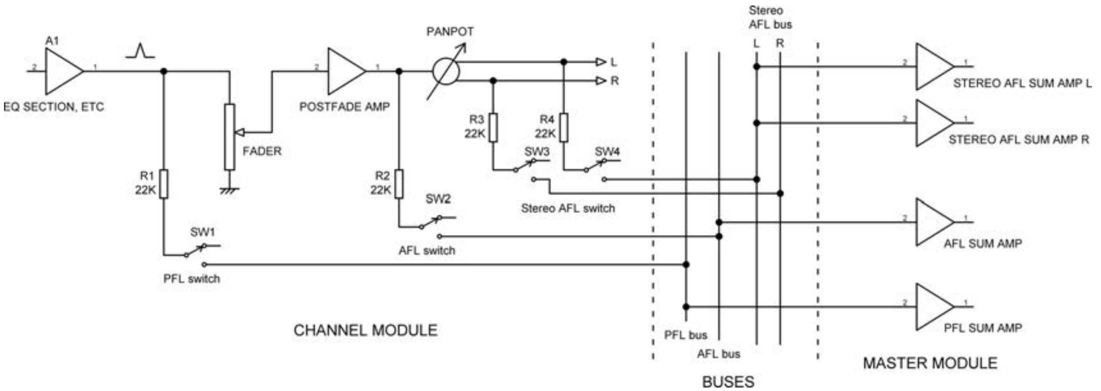

Figure 22.34: Comprehensive system with PFL and mono and stereo AFL

AFL systems

AFL is a very similar concept to PFL, the difference being that an AFL feed is taken from after the fader, so that if a selection of AFL buttons are pressed, the resulting mix has the sources in the right proportion. A stereo AFL system goes one better by taking a stereo feed from after the panpot, so that the sources selected can be heard in their correct positions in the stereo field. Figure 22.34 shows the differing points in the channel from where PFL, AFL, and stereo AFL are taken. Stereo AFL requires a 4-pole switch; two sections route the stereo signal to the AFL buses, and one is required to signal the AFL-detect system. The detection of an AFL condition works in exactly the same way as PFL detection.

It is not common for mixers to be fitted with an AFL system only, as it is much less convenient for checking signal levels in the channel; it will only read the prefade level correctly if the channel fader is at 0 dB. More usually a solo system is fitted that can be switched from PFL to AFL at the master section. The channel module switch will route signals to both the PFL and AFL buses; these are summed by separate summing amplifiers, and which of the two resulting signals is switched to the monitors is selected by the PFL/AFL switch on the master section.

Solo-in-place systems

PFL provides a mono look at a single channel. AFL preserves fader and perhaps pan position, but in both cases the signal that is heard is devoid of effects, which makes critical assessment difficult. A solo-in-place (SIP) system allows a channel or set of channels to be heard in isolation with all the effect sends working normally. This is done simply by muting all the channels except the ones you want to hear. This is obviously not

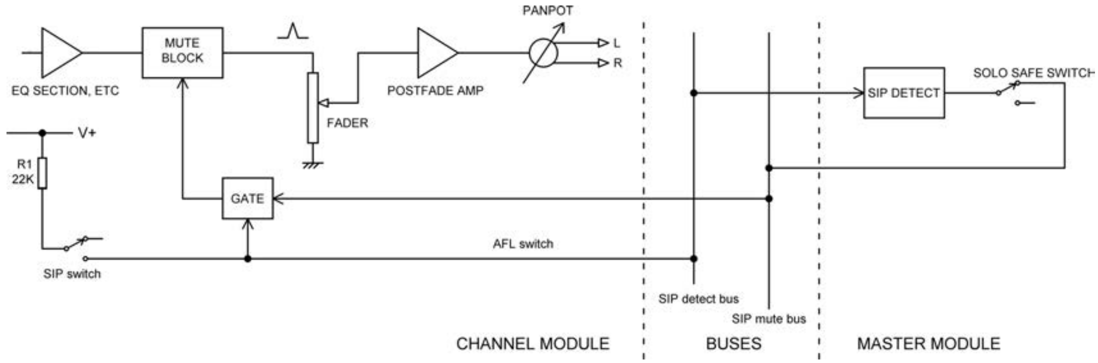

Figure 22.35: Solo-in-place system

applicable to PA work (except during set-up and sound checks) because it is what is called a 'destructive solo' in that the output of the mixer is disrupted while SIP is in use, unlike PFL and AFL systems which take over the monitor output but leave the group and L-R mix outputs unaffected.

A SIP system requires an electronically-controlled mute on each channel. When the SIP button on a channel is pressed, every *other* channel is muted except that one. If two or more channels are solo-ed, they are left unmuted while the rest of the channels are muted.

The basic SIP system therefore has a SIP-detect bus, which works in just the same way as a PFL-detect bus. This determines when one or more SIP switches are pressed and activates a SIP mute bus that potentially mutes all the channels; however, if the SIP switch on a channel is pressed it intercepts the mute signal and prevents it from operating on that channel. The basic scheme is shown in Figure 22.35.

SIP is a destructive solo, and you really don't want to trigger it during a live performance. Therefore a 'solo-safe' switch is usually fitted in the master section to act as a safety-catch or Molly-guard and prevent accidental operation. When it is engaged, the solo system usually acts as a PFL system instead, which is safe to use during performance.

Talkback microphone amplifiers

Most contemporary talkback systems have a small electret microphone mounted flush with the master module panel. The microphones used typically have an internal head amplifier that buffers the high impedance of the electret element, and all else that is needed is an amplifier stage with variable gain over a wide range. A typical arrangement is shown in Figure 22.36: gain is variable from 0 to +55 dB by RV1, with C4 keeping it to unity at DC. Note that the microphone is powered through the filter network R1 and C1.

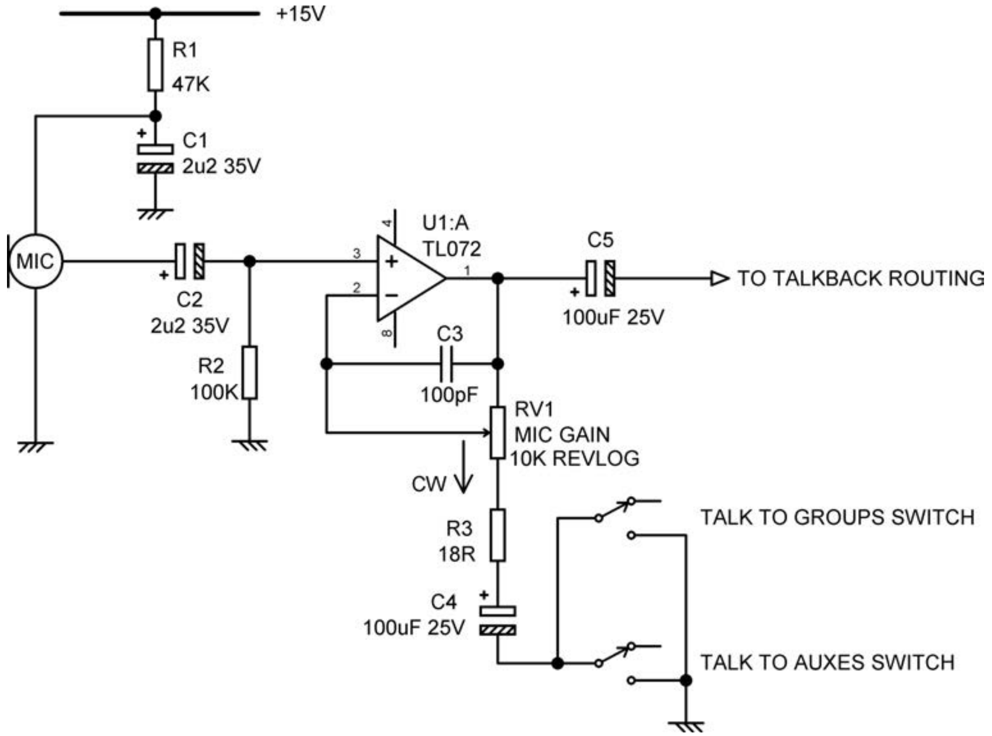

Figure 22.36: A typical talkback microphone amplifier, with gain variable from 0 to +55 dB

Figure 22.36 includes a couple of subtle but important design features. Firstly, the second changeovers of the talkback routing switches are connected into the feedback network so that when none of them are pressed, the amplifier gain is reduced to unity. This prevents the microphone amplifier from clipping continuously if the gain is turned up but talkback is not in use; the resulting distorted waveforms would almost certainly crosstalk crunchily into other parts of the master module, such as the L-R mix summing amplifiers, and this would not be a good thing.

Secondly, you will have noticed that the value of C4 looks rather small compared with the low value of the gain end-stop resistor R3, and in fact gives an LF roll-off that is −3 dB at 90 Hz. This is quite deliberate, and is intended to control the amplitude of LF transients when the gain is set high.

Line-up oscillators

The oscillator output is not required to be a perfect sinewave; it only has to be good enough so that meter calibration is not affected. It also needs to look like a sinewave on an oscilloscope, or customer confidence will be undermined. Figure 22.37 shows a simple but

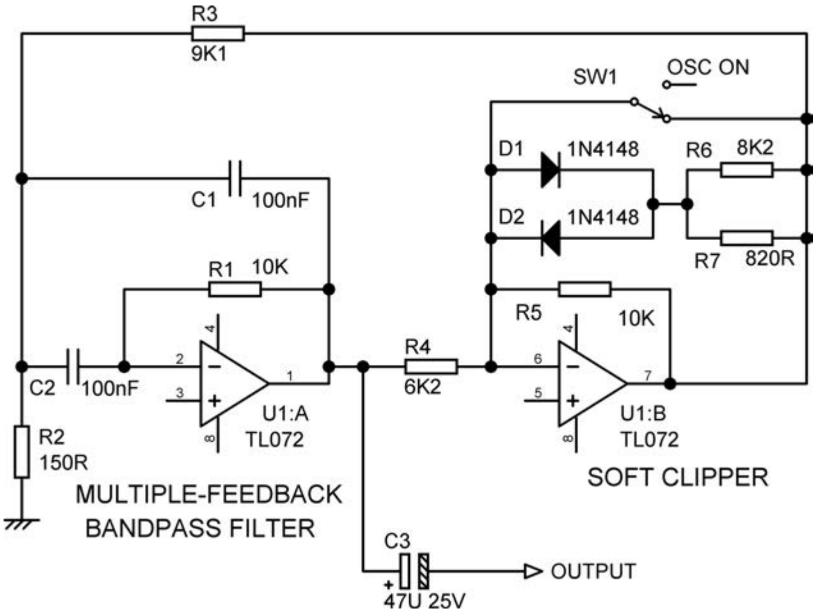

Figure 22.37: 1 kHz oscillator with diode amplitude stabilisation

dependable arrangement for a 1 kHz oscillator which does not require either expensive and fragile thermistors or complicated levelling circuitry.

It consists of a feedback loop containing a bandpass filter and a soft clipping circuit. The waveform at the output of bandpass filter U1:A is amplified by U1:B and then soft-clipped by D1 and D2. This symmetrical clipping introduces odd-order harmonics only, which are more easily filtered out when the signal is fed back to the multiple-feedback bandpass filter U1:A. The distortion is approximately 0.5%. Note that when the oscillator is not in use, it is not routed to the buses, but is also firmly stopped from oscillating at all by the closure of SW1. This is essential to remove any possibility of the oscillator crosstalking into the audio paths during normal operation. When SW1 is opened, this oscillator ramps up in amplitude in a very neat manner.

You will note the use of E24 resistor values to get the frequency as close to 1.0 kHz as possible, and the parallel combination of R6 and R7 to give 745 Ω; this resistance is fairly critical for obtaining the minimum distortion, and the combination shown here gave dependably lower distortion than the E24 value of 750 Ω. The output level is dependant on the forward voltage of the two diodes and so there is some variation with temperature, but it is insignificant in this application.

The next step up in oscillator sophistication is to provide two switched frequencies, usually 1 kHz and 10 kHz. The higher frequency was historically used for checking the azimuth

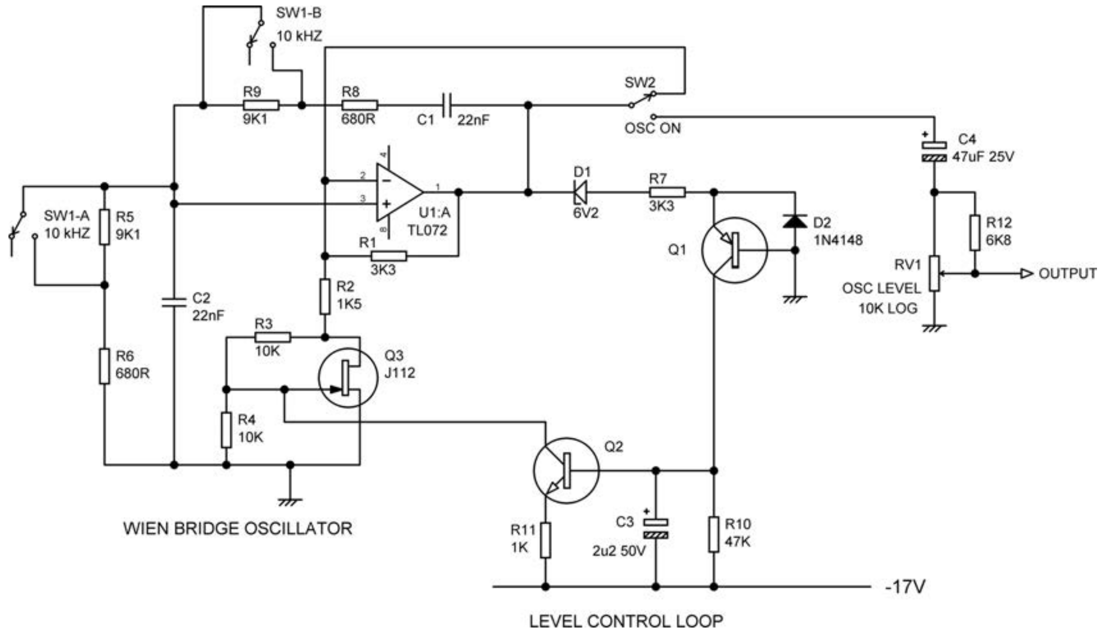

Figure 22.38: Switched 1 kHz/10 kHz oscillator with FET amplitude stabilization

adjustment of tape machine heads (when the azimuth of a tape head is maladjusted, the high frequency response falls off badly). Nowadays it is less useful but can still come in handy for quick frequency response checks. It is therefore very important that the output level is exactly the same for the two frequencies, and the oscillator shown in Figure 22.38 therefore has a more complex level-control loop.

The oscillator is based on the classic Wien bridge configuration, with the oscillation frequency controlled by R8, R9, C1 and R5, R6, C2. When running at 1 kHz R8, R9, and R5, R6 are in circuit. For 10 kHz SW1 is pressed and R5 and R9 are now shorted out, raising the operating frequency. The level control loop operation is as follows: when SW2 is pressed, the short-circuit across R1 in the negative feedback loop R1, R2 which prevents oscillation is removed and the amplitude of oscillation ramps up. When it reaches the desired level, Zener diode D1 begins to conduct on positive peaks and turns on the common-base transistor Q1 (D1 also conducts on negative half-cycles, but Q1 does not respond and is protected from reverse bias by clamp D2). The collector current of Q1 charges C3 and Q2 turns on, pulling down the gate of JFET Q3 and increasing its channel resistance, thus increasing the amount of negative feedback through R1, R2 and regulating the level. Q3 is a J112 FET, a type that is optimised for voltage-controlled resistance (VCR) operation. The network R3, R4 not only acts as the collector load for Q2, but also feeds half the V_{ds} of Q3 to its own gate; this is a classic method of reducing even-order distortion in JFETs and is dealt with in more detail in Chapter 24. C3 and R10 set the time-constant of the control-loop, and their values have a strong effect on oscillator distortion. Q1 and Q2 can be any high-beta transistor types.

Larger mixers often have more sophisticated oscillators with fully variable frequency and sometimes a choice of a squarewave output. Oscillator design is a massive subject on its own, and there is no space to get deeper into it here.

Console cooling and component lifetimes

Large mixing consoles have a large amount of electronics packed into a small space, both to reduce the size of the console so it is easier to install, and to bring all the controls within a reasonable reach. This means that natural convection cooling is often barely adequate. It is made more difficult by the fact that open cooling vents on the top of a console are not popular due to the possibility of various beverages being poured into the works, with an excellent chance of catastrophic results. Elevated internal console temperatures are rarely high enough to cause semiconductor failures, but have two bad consequences:

Firstly, over time the electrolytic capacitors will dry out, and drop in value. This loss of capacitance can be dramatic; a 100 μF component can fall to 10 nF. This in turn causes two further bad consequences: the LF response of the console degrades as coupling capacitors drop in value, and opamps go unstable as rail decoupling capacitors become ineffective. In a large studio console this can mean that its life is only ten years before all the electrolytics need replacing. This 'recapping' procedure, is, as you might imagine, a lengthy and expensive process that can be done only by companies with specialised skills.

The average electrolytic capacitor has a temperature rating of 85 °C. This does not mean that no capacitor degradation occurs below that temperature; degradation is happening all the time, but it accelerates rapidly as temperature rises, roughly doubling in speed with each 10 °C increase. 85 °C is the temperature at which capacitor life has dropped to a nominal 1000 hours, which is only 49 days, and so obviously they must be operated at a much lower temperature than that to get a reasonable equipment lifetime. Electrolytic capacitors rated at 105 °C have recently fallen in price to the point at which they are a viable alternative, and at a given temperature this increases lifetime by a factor of four. The best recapping facilities use low-ESR 105 °C capacitors exclusively for replacements, and in strategic places increase the value fitted so the life is longer before the value falls too low. This obviously requires an intimate knowledge of highly complex consoles and should only be undertaken by professionals such as GJC Designs [1].

Secondly, if silver contacts are used in switches or relays, the rate at which they are corroded by atmospheric hydrogen sulphide, creating non-conducting silver sulphide and causing the contacts to fail, also increases rapidly with temperature. Once again it roughly doubles in speed with each 10 °C increase.

The obvious way to improve console cooling is to add fans to increase the air flow. Fans are very undesirable in consoles installed in recording studios because of the acoustic noise they generate, and are rarely if ever used. They are however sometimes fitted in PA consoles, though still with considerable care to minimise the noise they create. When this is done by running the fan at low speeds, at the low end of its specified voltage range, great caution is needed. After a period of use the fan may stop altogether, as the bearings become worn, and how long this takes is not readily predictable. Sleeve bearings seem to be worse than ball bearings for this. The use of fans is to a large extent a last resort, and every effort should be made to induce adequate cooling by convection alone.

Reference

[1] GJC Designs. Available online at www.gjcdesigns.net/ (accessed October 2013).

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CHAPTER 23

Level indication and metering

Signal-present indication

Some amplifiers and the more sophisticated mixers are fitted with a 'signal-present' indicator that illuminates to give reassurance that a channel is receiving a signal and doing something with it. The level at which it triggers must be well above the noise floor, but also well below the Peak indication or clipping levels. Signal-present indicators are usually provided for each channel, and are commonly set up to illuminate when the channel output level exceeds a threshold something like 20 or 30 dB below the nominal signal level, though there is a wide variation in this.

A simple signal-present detector is shown in Figure 23.1, based on an opamp rather than a comparator. The threshold is −32 dBu, which combined with the −2 dBu nominal level which it was designed for, gives an indication at 30 dB below nominal. Since an opamp is used which is internally compensated for unity gain stability, there is no need to add hysteresis to prevent oscillation when the signal lingers around the triggering point. U1 is configured as an inverting stage with the inverting input biased slightly negative of the non-inverting input by R4 in the bias chain R3, R4, R5. The opamp output is therefore high with no signal, but is clamped by negative feedback through D1 to prevent excessive voltage excursions at the output which might crosstalk into other circuitry; C2 is kept charged via R6 and R7, so Q1 is turned on and LED1 is off. When an input signal exceeds the threshold the opamp output goes low and C2 is rapidly discharged through R6 and D2. R6 limits the discharge current to a safe value; the overload protection of the opamp would probably do this by itself, but I have always been a bit of a belt-and-braces man in this sort of case. With C2 discharged, Q1 turns off and LED1 illuminates as the 8 mA of LED chain current flows through it. When the input signal falls below the threshold, the opamp output goes high again, and C2 charges slowly through R7, giving a peak-hold action. This is a unipolar detector; only one polarity of signal activates it.

The LED chain current is provided by a wholly conventional constant-current source Q3, which allows any number of LEDs to be turned on and off without affecting the brightness of the other LEDs. In this case the clip-detect LED is shown just above the signal-present LED; since it is only illuminated when the signal-present LED is already on, the drive requirements

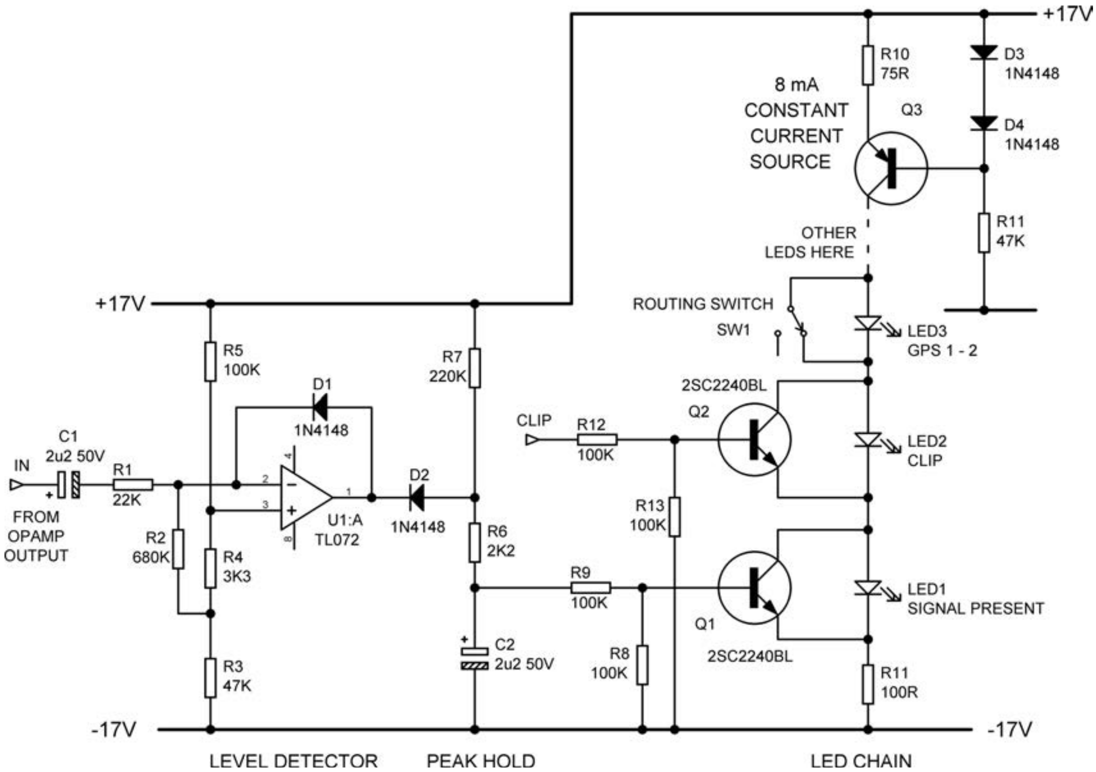

Figure 23.1: A simple unipolar signal-present detector, using an opamp. The threshold is −32 dBu. Other LEDs are run in the same constant-current chain

for Q2 are simple and it can be driven from exactly the same sort of capacitor-hold circuitry. The other LEDs in the chain (here assumed to be routing switch indicators) simply 'float' above the LEDs controlled by transistors. The LED chain is connected between the two supply rails so there is no possibility of current being injected into the ground. This gives a span of 34 V, allowing a large number of LEDs to be driven economically from the same current. The exact number depends on their colour, which affects their voltage drop. It is of course necessary to allow enough voltage for the constant-current source to operate correctly, plus a suitable safety margin.

I have used this circuit many times in mixers and it can be regarded as well-proven.

A vital design consideration for signal-present indicators is that since they are likely to be active most of the time, the operation of the circuitry must not introduce distortion into the signal being monitored; this could easily occur by electrostatic coupling or imperfect grounding if there is a comparator switching on and off at signal frequency. Avoid this.

Peak indication

A mixer has a relatively complex signal path, and the main metering is normally connected only to the group outputs. The mix metering can be used to measure the level in other parts of the signal path by use of the PFL system (see Chapter 22) if fitted, but this can only monitor one channel at a time. It is therefore usual to guard against clipping by fitting peak level indication to every channel of all but the simplest consoles, and sometimes to effect-return modules also.

The Peak indicator is driven by fast-attack, slow-decay circuitry so that even brief peak excursions give a positive display. It is important that the circuitry should be bipolar, i.e. it will react to both positive and negative peaks. The peak values of a waveform can show asymmetry up to 8 dB or more, being greatest for unaccompanied voice or a single instrument, and this is, of course, very often exactly what goes through a mixer channel. This level of uncertainty in peak detection is not a good thing, so only the simplest implementations use unipolar peak detection. Composite waveforms, produced by mixing several voices or instruments together, do not usually show significant asymmetries in peak level.

Figure 23.2 shows a simple unipolar peak LED driving circuit. This only responds to positive peaks, but it does have the advantage of using but two transistors and is very simple and

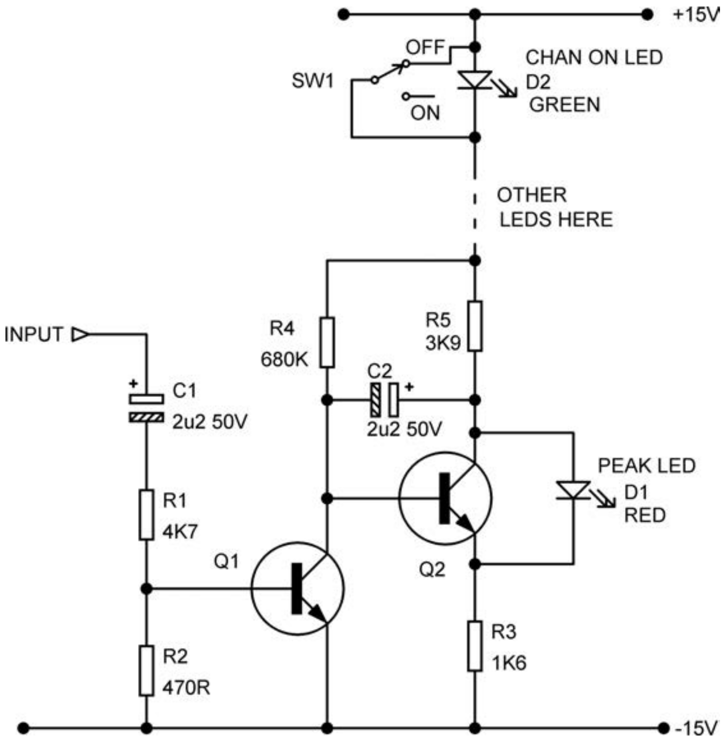

Figure 23.2: A simple unipolar peak detector, including powering for the Chan On LED

cheap to implement. When a sufficient signal level is applied to C1, Q1 is turned on via the divider R1,R2; this turns off Q2, which is normally held on by R4, and Q2 then ceases to shunt current away from peak LED D1. C2 acts as a Miller integrator to stretch the peak hold time; when Q1 turns off again, R4 must charge C2 before Q2 can turn on again. Note that this circuit is integrated into the channel on LED supply, with R5 setting the current through the two LEDs; the channel on LED is illuminated by removing the short placed across it by SW1. R5 is of high enough value, because it is connected between the two supply rails, for there to be no significant variation in the brightness of one LED when the other turns off. If, for some reason, this was a critical issue, R5 could be replaced by a floating constant-current source. Other LEDs switched in the same way can be included in series with the channel on LED.

This peak detect circuit has a non-linear input impedance and must only be driven from a low-impedance point; preferably direct from the output of an opamp. The peak LED illuminates at an input of 6.6 V peak, which corresponds to 4.7 Vrms (for a sinewave) and +16 dBu. For typical opamp circuitry running off the usual supply rails this corresponds to having only 3 or 4 dB of headroom left. The detect threshold can be altered by changing the values of the divider R1, R2.

The Log Law Level LED (LLLL)

The Log Law Level LED or LLLL was evolved for the Elektor preamplifier project of 2012, to aid in the adjustment of a phono input with several different gain options. It is, to the best of my knowledge, a new idea. Usually, a single-LED level indicator is driven by an opamp or a comparator, and typically goes from fully off to fully on with less than a 2 dB change in input level when fed with music (*not* steady sinewaves). It therefore only gives effectively one bit of information.

It would be useful to get a bit more enlightenment from a single LED. More gradual operation could be adopted, but anything that involves judging the brightness of an LED is going to be of doubtful use, especially in varying ambient lighting conditions. The LLLL, on the other hand, uses a comparator to drive the indicating LED hard on or hard off. It incorporates a simple log-converter so that the level range from LED always-off to always-on is much increased, to about 10 dB, the on-off ratio indicating where the level lies in that range. In some applications, such as the Elektor preamplifier, it may be appropriate to set it up so that the level is correct when the LED is on about 50% of the time. This gives a much better indication.

The circuitry of the LLLL is shown in Figure 23.3. The U1:A stage is a precision rectifier circuit that in conjunction with R3 provides a full-wave rectified signal to U1:B; this is another precision rectifier circuit that establishes the peak level of the signal on C1. This is buffered by U2:A and applied to the approximately log-law network around U2:B. As the

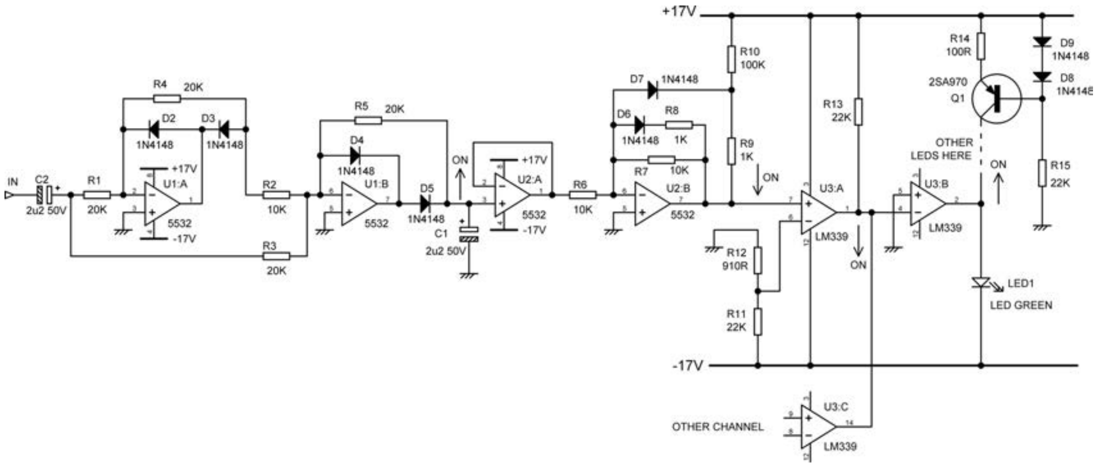

Figure 23.3: The Log Law Level LED or LLLL

signal level increases, first D6 conducts, reducing the gain of the stage, and then at a higher voltage set by R9, R10, D7 conducts and reduces the gain further. If sufficient signal is present (to exceed the threshold set by R11, R12) the output of open-collector comparator U3:A goes low and U3:B output goes high, removing the short across LED1 and allowing it to be powered by the 6 mA current-source Q1. As with other circuitry in this chapter, the LED current is run from rail to rail, avoiding the ground. Many other LEDs can be inserted in the constant-current LED chain. The LLLL has been built in significant numbers and I have never heard of any problems with it.

If a stereo version of the LLLL is required, which will indicate the greater of the two input signals, the output of comparator U3:A is wire-ORed with the output of U3:C, which has the same function in the other channel; the circuitry up to this point is duplicated. A more elegant way to make a stereo version would be to combine the outputs of two peak rectifiers to charge C1. This would save a handy number of components but I have not yet actually tried it out.

I have spent some time testing the operation of this scheme, using various musical genres controlled by a high-quality slide fader. I believe it is a significant advance in signalling level when there is only one LED available, but in the words of Mandy Rice-Davies, ‘Well, he would say that, wouldn’t he’ [1]. Opinions on the value of the LLLL would be most welcome.

Distributed peak detection

When an audio signal path consists of a series of circuit blocks, each of which may give either gain or attenuation – and the classical example is a mixer channel with multiple EQ stages and a fader and post-fade amplifier – it is something of a challenge to make sure that excessive levels do not occur anywhere along the chain. Simply monitoring the level at the end of the chain is no use because a circuit block that gives gain, leading to clipping, may be followed by one that attenuates the clipped signal back to a lower level that does not trip a final peak-detect circuit. The only way to be absolutely sure that no clipping is happening anywhere along the path is to implement bipolar peak detection at the output of every opamp stage. This is, however, normally regarded as a bit excessive, and the usual practice in high-end equipment is to just monitor the output of each circuit block, even though each such block (for example a band of parametric EQ) may actually contain several opamps. It could be argued that a well-designed circuit block should not clip anywhere except at its output, no matter what the control setting, but this is not always possible to arrange.

A multi-point or distributed peak detection circuit that I have made extensive use of is shown in Figure 23.4. It can detect when either a positive or negative threshold is exceeded, at any number of desired points; to add another stage to its responsibilities you need only add another pair of diodes, so it is very economical. However, if one peak detector monitors too many points in the signal path, it can be hard to determine which of them is causing the problem. In most applications I have used the circuit to keep an eye on the output of the microphone preamplifier, the output of the EQ section, and the output of the fader post-amplifier. This means that the location of the clipping can be pinpointed quite easily. If you pull down the fader to 0 dB or below and the peak LED goes out, the problem was at the post-fade amplifier. If that doesn't do the trick, switch out the EQ; this assumes of course that the EQ in/out switch removes the signal feed to the unused EQ section. I always arrange matters so if possible, removing the EQ signal reduces power consumption and minimises the possibility of crosstalk. If that is not the case, then you will have to back off any controls with significant boost and see if that works. Should the peak indication persist, it must be coming from the output of the microphone preamplifier, and you will need to reduce the input gain.

The operation is as follows. Because R5 is greater than R1, normally the non-inverting input of the opamp is held below the inverting input and the opamp output is low. If any of the

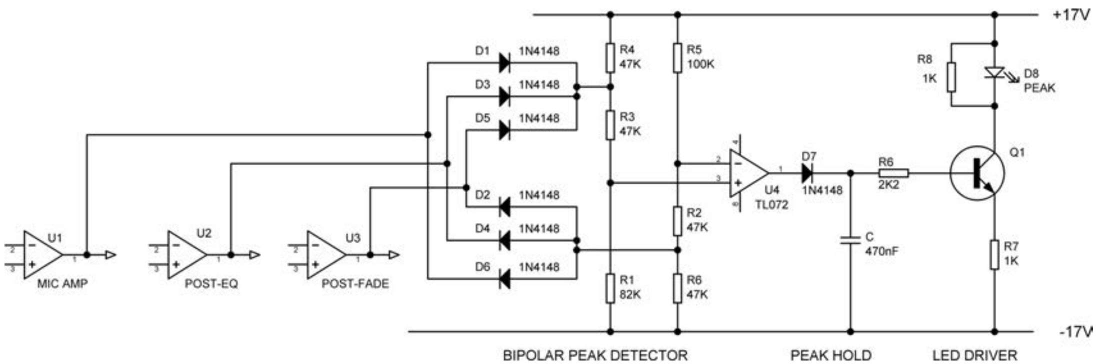

Figure 23.4: A multipoint bipolar peak detector, monitoring three circuit blocks

inputs to the peak system exceed the positive threshold set at the junction of R4, R3, one of D1, D3, D5 conducts and pulls up the non-inverting input, causing the output to go high. Similarly, if any of the inputs to the peak system exceed the negative threshold set at the junction of R2, R6, one of D2, D4, D6 conducts and pulls down the inverting input, once more causing the opamp output to go high. When this occurs C1 is rapidly charged via D7. The output-current limiting of the opamp discriminates against very narrow noise pulses. When C1 charges Q1 turns on, and illuminates D8 with a current set by the value of R7. R8 ensures that the LED stays off when U4 output is low, as it does not get close enough to the negative supply rail for Q1 to be completely turned off.

Each input to this circuit has a non-linear input impedance, and so for this system to work without introducing distortion into the signal path, it is essential that the diodes D1–D6 are driven directly from the output of an opamp or an equivalently low impedance. Do not try to drive them through a coupling capacitor as asymmetrical conduction of the diodes can create unwanted DC-shifts on the capacitor.

The peak-detect opamp U4 must be a FET-input type to avoid errors due to bias currents flowing in the relatively high value resistors R1–R6, and a cheap TL072 works very nicely here; in fact the resistor values could probably be raised significantly without any problems.

As with other non-linear circuits in this book, everything operates between the two supply rails so unwanted currents cannot find their way into the ground system.

Combined LED indicators

For many years there has been a tendency towards very crowded channel front panels, driven by a need to keep the overall size of a complex console within reasonable limits. One apparently ingenious way to gain a few more square millimetres of panel space is to combine the signal-present and peak indicators into one by using a bi-colour LED. Green shows signal-present, and red indicates peak. One might even consider using orange (both LED colours on) for an intermediate level indication.

Unfortunately, such indicators are hard to read, even if with normal colour vision. If you have red-green colour-blindness, the most common kind (6% of males, 0.4% of females), they are useless. Combining indicators like this is really not a good idea.

VU meters

VU meters are a relatively slow-response method of indicating an audio level in 'volume units'. The standard VU meter was originally developed in 1939 by Bell Labs and the USA broadcasters NBC and CBS. The meter response is intentionally a 'slow' measurement which is intended to average out short peaks and give an estimate of perceived loudness. This worked adequately when it was used for monitoring the levels going to an analogue tape machine, as the overload characteristic of magnetic tape is one of soft compression and the occasional squashing of short transient peaks is hard to detect aurally. Digital recorders overload in a much more abrupt and intrusive manner, making even brief overloads unpleasant, and the use of VU meters in professional audio has been in steady decline for many years.

The specifications, particularly of the dynamics, of a standard VU meter are closely defined in the documents British Standard BS 6840, ANSI C16.5-1942, and IEC 60268-17, but there are many cheap meters out there with 'VU' written on them that make no attempt to conform with these documents. The usual VU scale runs from −20 to +3, with the levels above zero being red. 1 VU is the same change in level as 1 dB. The rise and fall times of the meter are both 300 msec so if a sine wave of amplitude 0 VU is applied suddenly, the needle will take 300 msec to swing over to 0 VU on the scale. A proper VU meter uses full-wave rectification so asymmetrical waveforms are measured correctly, but the cheap pretenders normally have a single series diode that only gives half-wave rectification. VU meters are calibrated on the usual measure-average-but-pretend-it's-RMS basis, so a VU meter gives a true reading of RMS voltage level only for a sinewave. Musical signals are usually more peaky than sine waves so the VU meter will read somewhat lower than the true RMS value.

The 0 VU mark on the meter scale is 'zero-reference level' but what that means in terms of actual level depends on the system to which it is fitted. In professional audio equipment, 0 VU is +4 dBu, whereas in semi-pro gear it will be −10 dBv (= −7.8 dBu)

VU meters consist of a relatively low resistance meter winding driven by rectifier diodes, with a series resistor added to define the sensitivity. The usual value is 3k6 which gives 0 VU = +4 dBu. They therefore present a horribly non-linear load to an external circuit, and a VU meter must never be connected across a signal path unless it has near-zero impedance. This is particularly true for cheap ones with half-wave rectification. In practice, a buffer stage is always used between a signal path and the VU meter to give complete isolation and to allow the calibration to be adjusted. Many mixers have a nominal internal level 6 dB below the nominal output level, because the balanced output amplifiers inherently have 6 dB of gain, and so the meter buffer amplifiers must also be capable of giving 6 dB of gain.

Figure 23.5 shows an effective design for a meter buffer stage designed to work with a nominal internal level of −6 dBu, and which must therefore give 10 dB of gain to raise the meter signal to the +4 dBu that will give a 0 VU reading. C1 provides DC blocking, because a VU meter will respond to DC as well as AC. R1 and R2 set the gain range to be +6 to +13 dB, which is an ample range of adjustment. With the preset centralised, the gain is 9.3 dB. The resistor values are unusually high because presenting a high input impedance is more important here than the noise performance. An amplifier which was noisy enough to register directly

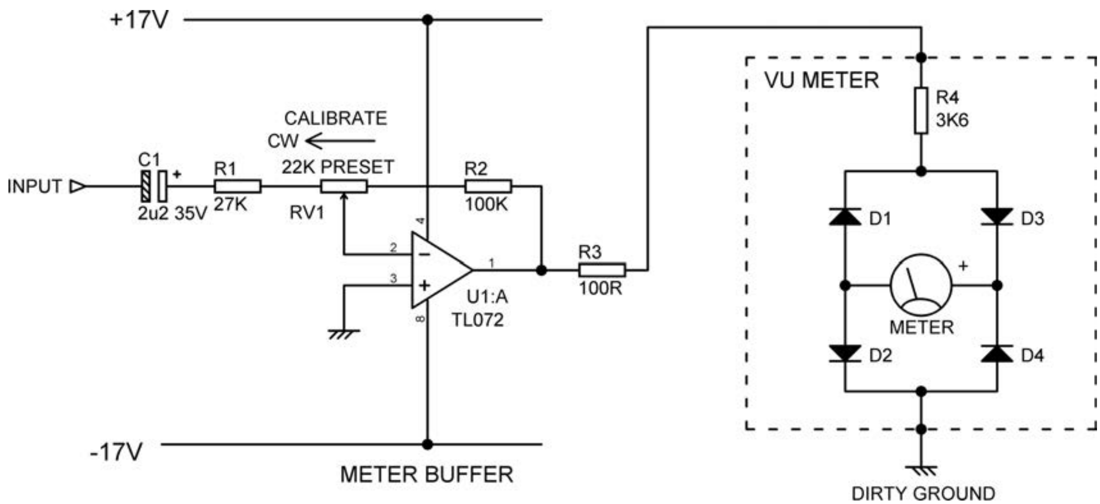

Figure 23.5: A very simple but effective design for a VU meter buffer stage

on a VU meter would probably be better fitted to a life as a white noise generator. R3 is an isolating resistor to make sure that the capacitance of the cable to the VU meter, which may be quite lengthy if the meter is perched up in an overbridge, does not cause instability in the buffer amplifier; its presence in series with the meter resistor R4 is allowed for when the calibration is set. The ground of the meter itself is labelled 'dirty ground' to underline the point that the current through the meter will be heavily distorted by the rectification going on, and must not be allowed to get into the clean audio ground.

Because of their slow response, VU meters are sometimes made with a peak LED projecting through the meter scale. This is driven by a peak-detect circuit of the sort described earlier in this chapter.

PPM meters

Peak programme meters (PPMs) are essentially peak-reading instruments that respond much more quickly than VU meters. They are always a good deal more expensive, partly because of the precisely defined and rather demanding characteristics of the physical meter itself – for example, the needle has to be able to move much faster than a VU needle, but without excessive overshoot – and partly because they need much more complex drive circuitry. The PPM standard was originally developed by the BBC in 1938, as a response to the inadequacy of existing average-responding meters. PPMs have a distinctive scale with white legends and a white needle against a black background, and are marked from 1 to 7. This is a logarithmic scale giving 4 decibels per division, and the accurate and temperature-stable implementation of this characteristic is what makes the drive circuitry expensive.

Nonetheless, PPMs are specifically designed *not* to catch the very fastest of transient peaks, and are therefore sometimes called 'quasi-peak' meters. They only respond to transients sustained for a defined time; the specs give 'Type I' meters an integration time of 5 msec, while 'Type II' meters use 10 msec. The result is that transient levels normally exceed the PPM reading by some 4 to 6 dB. This approach encourages operators to somewhat increase programme levels, giving a better signal-to-noise performance. The assumption (which is generally well-founded) is that occasional clipping of brief transients is not audible. The existence of both Type I and II meters simply reflects differing views on the audibility of transient distortion.

PPMs exhibit a slow fallback from peak deflections, so it is easier to read peak levels visually. Type I meters should take 1.4 to 2.0 seconds to fall back 20 dB while Type II meters should take 2.5 to 3.1 seconds to fall back 24 dB. Type II meters also incorporate a delay of from 75 msec to 150 msec before the needle fallback is allowed to begin; this peak-hold action makes reading easier.

LED bar-graph metering

Bar-graph meters are commonly made up of an array of LEDs. An LED bar-graph meter can be made effectively with an active-rectifier circuit and a resistive divider chain that sets up the trip voltage of an array of comparators; this allows complete freedom in setting the trip level for each LED. A typical circuit which indicates from 0 dB to −14 dB in 2 dB steps with a selectable peak or average-reading characteristic is shown in Figure 23.6 and illustrates some important points in bar-graph design.

U3 is a half-wave precision rectifier of a familiar type, where negative feedback servos out the forward drop of D11, and D10 prevents opamp clipping when D11 is reverse-biased. The rectified signal appears at the cathode of D11, and is smoothed by R7 and C1 to give an average, sort-of-VU response. D12 gives a separate rectified output and drives the peak-storage network R10, C9 which has a fast attack and a slow decay through R21. Either average or peak outputs are selected by SW1, and applied to the non-inverting inputs of an array of comparators. The LM2901 quad voltage comparator is very handy in this application; it has low input offsets and the essential open-collector outputs.

The inverting comparator inputs are connected to a resistor divider chain that sets the trip level for each LED. With no signal input, the comparator outputs are all low and their open-collector outputs shunt the LED chain current from Q1 to −15 V, so all LEDs are off. As the input signal rises in level, the first comparator U2:D switches its output off, and LED D8 illuminates. With more signal, U2:C also switches off and D7 comes on, and so on, until U1:A switches off and D1 illuminates. The important points about the LED chain are that the highest level LED is at the bottom of the chain, as it comes on last, and that the LED current flows from one supply rail down to the other and is not passed into a ground. This prevents

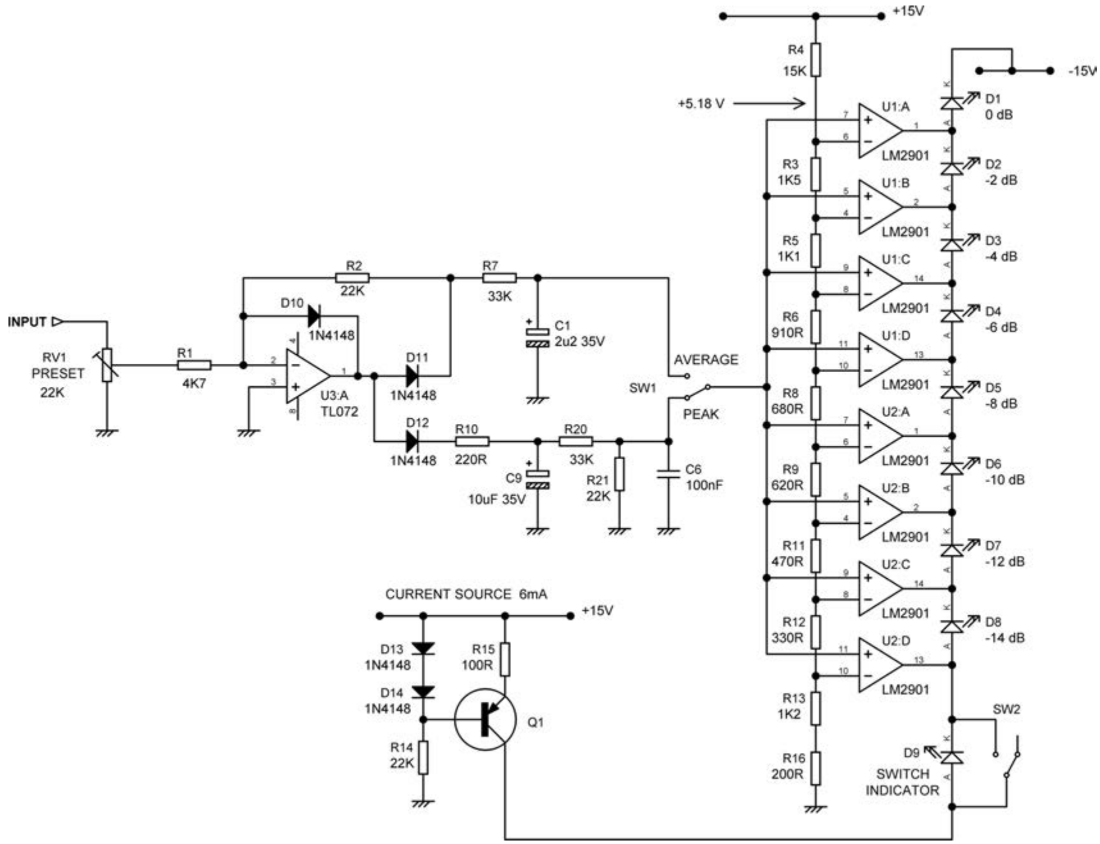

Figure 23.6: LED bar-graph meter with selectable peak/average response

noise from getting into the audio path. The LED chain is driven with a constant-current source to keep LED brightness constant despite varying numbers of them being in circuit; this uses much less current than giving each LED its own resistor to the supply rail and is universally used in mixing console metering. Make sure you have enough voltage headroom in the LED chain, not forgetting that yellow and green LEDs have a larger forward drop than red ones. The circuit shown has plenty of spare voltage for its LED chain, and so it is possible to put other indicator LEDs in the same constant-current path; for example D9 can be switched on and off completely independently of the bar-graph LEDs, and can be used to indicate Channel-on status or whatever. An important point is that in use the voltage at the top of the LED chain is continually changing in 2-volt steps, and this part of the circuit must be kept away from the audio path to prevent horrible crunching noises from crosstalking into it.

This meter can of course be modified to have a different number of steps, and there is no need for the steps to be the same size. It is as accurate in its indications as the use of E24 values in the resistor divider chain allows.

If a lot of LED steps are required, there are some handy ICs which contain multiple open-collector comparators connected to an in-built divider chain. The National LM3914 has 10 comparators and a divider chain with equal steps, so they can be daisy-chained to make big displays, but some law-bending is required if you want a logarithmic output. The National LM3915 also has 10 comparators, but a logarithmic divider chain covering a 30 dB range in 3 dB steps.

A more efficient LED bargraph architecture

The bar-graph meter shown in Figure 23.6 above draws 6 mA from the two supply rails at all times, even if all the LEDs are off for long periods, which is often the case in recording work. This is actually desirable in a simple mixer as the ±15 V or ±17 V rails are also used to power the audio circuitry and step-changes in current taken by the meter could get into the ground system via decoupling capacitors and suchlike, causing highly unwelcome clicks.

In larger mixers, a separate meter supply is provided to prevent this problem and this allows more freedom in the design of the meter circuitry. In the example I am about to recount, the meter supply available was a single rail of +24 V; this came from an existing power supply design and was not open to alteration, negotiation, or messing about with. A meter design with 20 LEDs was required, and an immediate problem was that you cannot power 20 LEDs of assorted colours with one chain running from +24 V; two LED chains would be required and the power consumption of the meter, even when completely dormant, would be twice as great. I therefore devised a more efficient system, which not only saves a considerable amount of power, but also actually economises on components.

The meter circuit is shown in Figure 23.7, and I must admit it is not one of those circuit diagrams where the modus operandi exactly leaps from the page. However, stick with me.

There are two LED chains, each powered by its own constant-current source Q1, Q2. The relevant current source is only turned when it is needed. With no signal input, all LEDs are off; the outputs of comparators U10 and U20 are high (open-collector output off) and both Q1 and Q2 are off. The outputs of all other comparators are low. When a steadily increasing signal arrives, U20 is the first comparator to switch, and LED D20 turns on. With increasing signal, the output of U19 goes high, and the next LED, D19, turns on. This continues, in exactly the same way as the conventional bargraph circuit described above, until all the LEDs in the chain D11–D20 are illuminated. As the signal increases further, comparator U10 switches and turns on the second current source Q2, illuminating D10; the rest of the LEDs in the second chain are then turned on in sequence as before. This arrangement saves a considerable amount of power, as no supply current at all is drawn when the meter is inactive, and only half the maximum is drawn so long as the indication is below −2 dB.

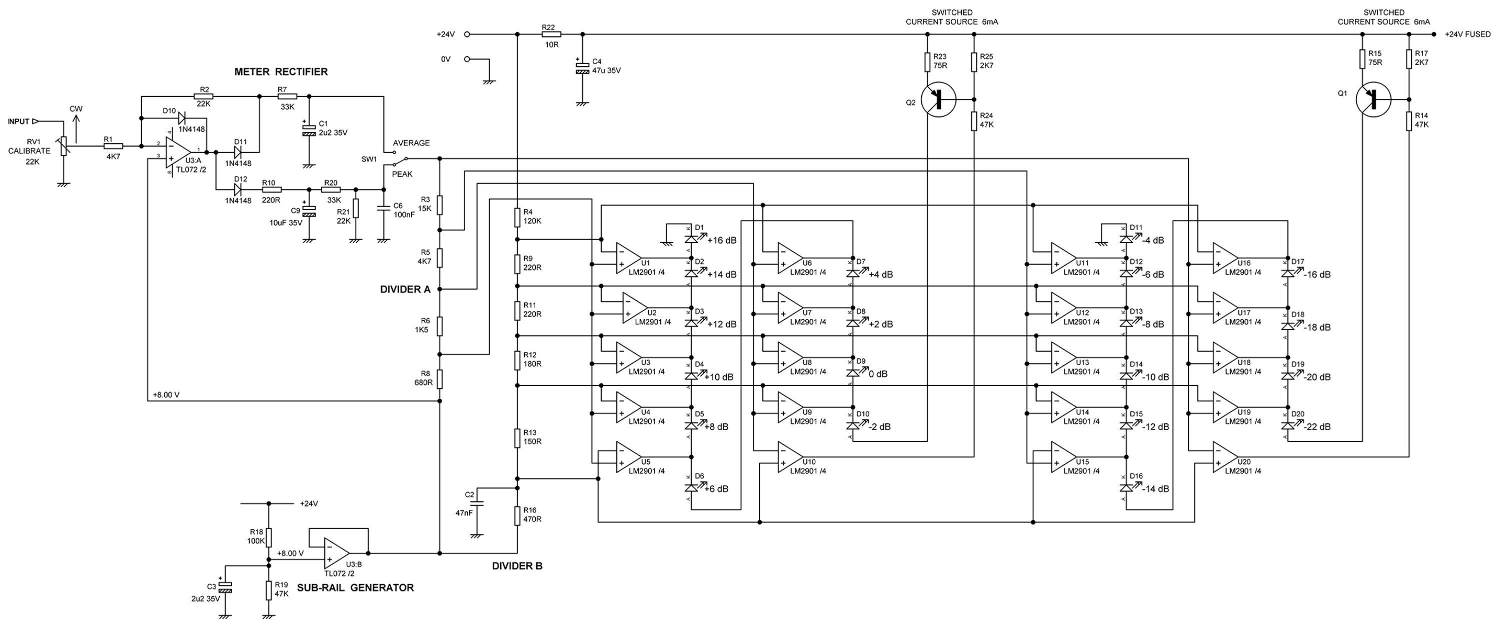

Figure 23.7: A more efficient LED bar-graph meter

There are 10 comparators for each LED chain, 20 in all, so a long potential divider with 21 resistors would be required to provide the reference voltage for each comparator if it was done in the conventional way, as shown in Figure 23.6 above. However, looking at all those comparator inputs tied together, it struck me there might be a better way to generate all the reference voltages required, and there is.

The new method, which I call a 'matrix divider' system, uses only 10 resistors. This is more significant than it might at first appear, because the LEDs are on the edge of the PCB, the comparators are in compact quad packages, and so the divider resistors actually take up quite a large proportion of the PCB area. Reducing their number by half made fitting the meter into a pre-existing and rather cramped meter bridge design possible without recourse to surface-mount techniques. There are now two potential dividers. Divider A is driven by the output of the rectifier circuit, while Divider B produces a series of fixed voltages with respect to the +8.0 V sub-rail. As the input signal increases, the output of the meter rectifier goes straight to comparators U16–U20, which take their reference voltages from Divider B and turn on in sequence as described above. Comparators U11–U15 are fed with the same reference voltages from Divider B, but their signal from the meter rectifier is attenuated by Divider A, coming from the tap between R3 and R5, and so these comparators require more input signal to turn on. This process is repeated for the third bank of comparators U6–U10, whose input signal is further attenuated, and finally for the fourth bank of comparators U1–U5, whose input is still further attenuated. The result is that all the comparators switch in the correct order.

Since in this application there was only a single supply rail, a bias generator is required to generate an intermediate sub-rail to bias the opamps. This sub-rail is set at +8.0 V rather than V/2, to allow enough headroom for the rectifier circuit, which produces only positive outputs; it is generated by R18, R19 and C3 and buffered by opamp section U3:B. The main +24 V supply is protected by a 10 Ω fusible resistor R22, so if a short-circuit occurs on the meter PCB the resistor will fail to open and the whole metering system will not be shut down. This kind of per-module fusing is very common and very important in mixer design; it localises a possibly disabling fault to one module, and avoids having the power supply shut down, which would put the whole mixer out of action. A small but vital point is that the supply for Divider B is taken from outside this fusing resistor; if it was not the divider voltages would vary with the number of LED chains powered, upsetting meter accuracy.

Once again the LM2901 quad voltage comparator is used, as it has low input offset voltages and the requisite open-collector outputs. Q1, Q2 can be any TO-92 devices with reasonable beta; their maximum power dissipation, which occurs with only one LED on in the chain, is a modest 128 mW. This meter system was used with great success.

Vacuum fluorescent displays

Vacuum fluorescent displays (VFDs) are sometimes used as bargraph meters [2]. They are superior to LCDs as they emit a bright light and show good contrast. Their operation is similar to that of a triode valve; a metal-oxide coated cathode gives off electrons when electrically heated, and their movement is controlled by wire grids. If they are directed to hit one of the multiple phosphor-coated anodes, it fluoresces, giving off light. The colour can be controlled by selecting appropriate phosphors. The cathode runs at a much lower temperature than in conventional valves, and does not visibly glow. The cathode heating voltage is usually 2–3 V, and this is often supplied as a balanced centre-tapped drive to minimise brightness variations across the display. This voltage is inconveniently low and the current inconveniently high for many applications so an inverter with a transformer is used. The same transformer may be used to generate the anode voltage, usually +50 to +60 V.

VFDs require hefty tooling charges if you want a custom display and their awkward power requirements make them more complicated to use than LEDs, especially because of the need for an electrically noisy inverter. One of the best known applications of a VFD was in the classic Casio FX-19 scientific calculator [3] that appeared in 1976. I bought mine then and it is still working perfectly; it is my calculator of choice because the display is brilliantly clear under all conditions. VFDs pre-dated LCDs, but they are still very much in use where high contrast is required. My Blu-Ray player has one.

Plasma displays

A plasma display [4] consists of a large number of gas-filled cells between two glass plates, which glow when energised with a high voltage. The cells are addressed by long electrode strips which run in the X-direction on one glass plate and in the Y-direction on the other. When voltages are suitably applied, only the cell at the crossing point of one X and one Y electrode will glow. Plasma displays were used in the SSL 4000 and 5000 consoles, and in the Neve V1, V2 V3 and VR consoles, amongst others. These were neon-filled, giving an orange glow. The plasma modules used by Neve combined PPM and VU characteristic bargraphs in one module; each bargraph was composed of 100 steps. They operate from 248 V DC, a voltage which requires considerable respect. That does not in itself make driving the displays difficult, because such a voltage can be switched with a couple of MPSA42/92 transistors. This only needs to be implemented once as the switching of the steps is done by a multiplexing process.

Since plasma televisions, working on the same basic principle but showing three colours, are available for reasonable prices (£340 for a 43-inch model in June 2013) you might be surprised to learn that a single-channel plasma display module for console use costs £250, presumably due to the small production quantities involved.

Liquid crystal displays

The ultimately versatile metering system is provided by making the meter display from a number of colour LCD display screens [5]. These can be of the size used in the smaller laptop computers, butted side-to-side to make something like a conventional meterbridge, or larger screens that can show three or more rows of bar-graphs at once. All the Calrec digital consoles currently have such metering. The advantages are obviously that you can display any kind of metering that you can think up, and the cost is low because the technology can be based on standard laptop screen displays and graphics chips, which are made in enormous quantities.

References

[1] http://en.wikipedia.org/wiki/Mandy_Rice-Davies (accessed June 2013).
[2] http://en.wikipedia.org/wiki/Vacuum_fluorescent_display (accessed June 2013).
[3] http://www.vintage-technology.info/pages/calculators/casio/casiofx19.htm (accessed June 2013).
[4] http://en.wikipedia.org/wiki/Plasma_display (accessed June 2013).
[5] http://en.wikipedia.org/wiki/Liquid-crystal_display (accessed June 2013).

CHAPTER 24

Level control and special circuits

Gain-control elements

A circuit block that gives voltage-control of gain is a difficult thing to implement. The basic function is simply that of multiplication, but most of the practical techniques introduce significant distortion. In every case distortion and noise can be traded off by altering the operating level, but in some techniques the non-linearity is so great that a compromise that meets modern performance standards is simply not possible.

A brief history of gain-control elements

The difficulty of implementing a good voltage-variable gain element is testified to by the number of technologies that have been tried. A variable attenuator could be made by varying the current flow through diodes, which varied their effective resistance; the control signal inevitably got mixed up with the audio signal, and linearity was poor. Optical combinations of filament bulbs and cadmium-sulphide photoresistors gave very good control-signal isolation but operation was slow because of the thermal inertia of the filament. Later versions used LEDs and were much faster but the linearity of the photoresistors remained a limitation. Chopper systems, which turned the signal hard on and off at an ultrasonic frequency, with a variable mark-space ratio, and then reconstructed the signal with a low-pass filter, gave fast response and reasonable linearity. However it was difficult to get a wide control range and effectively removing all the switching frequencies required a fairly complicated filter. A long time ago I designed just such a variable attenuator, using 4016 analogue gates for the switching, and a clock frequency of 160 kHz. The gain was difficult to control below −40 dB and it never made it to production, which on the whole was probably just as well.

JFETs

The introduction of JFETs (junction-FETs as opposed to MOSFETs) as gain control devices was a great advance. These devices promised good isolation between control-voltage and signal, instantaneous operation, and lower distortion than existing methods. These advantages do exist, but there are some less desirable features, such as a highly non-linear gain/CV law that varies significantly from specimen to specimen, as a result of process tolerances. Signal-CV isolation is absolute in the DC sense, the gate looking like a very-low-leakage reverse-biased diode, but there is always some gate-channel capacitance which means that fast edges on the control voltage can get through to the signal path.

The use of JFETs for on/off signal control rather than voltage-controlled attenuation is covered in Chapter 21 on signal switching.

A JFET used as a voltage-controlled attenuator is operated below pinch-off, i.e. at low values of V_{ds} that allow it to operate more like a resistor than a constant-current source. Some JFET types are better than others for this job, the 2N5457 and the 2N5459 being particularly favoured in the mid-Seventies.

Figure 24.1 shows the most basic voltage-controlled JFET attenuator circuit. When N-channel FETs are used, as is normally the case, the control voltage must go negative of ground to turn off the JFET and give the minimum-attenuation condition. The maximum attenuation is limited by the $R_{ds(on)}$ of the JFET to about 45 dB for practical circuitry. The attenuation law is highly non-linear, and variable between specimens of the same JFET; to some extent this can be trimmed out by applying a constant DC bias to the control voltage, but minor variations in the shape of the control law still remain. One of the advantages of this circuit is that when there is no gain reduction, the JFET is biased hard off and there is no extra distortion introduced, though there will be Johnson noise from the series resistor. This makes JFETs useful in limiter applications where gain reduction occurs only briefly and intermittently; JFETs are much cheaper than VCAs.

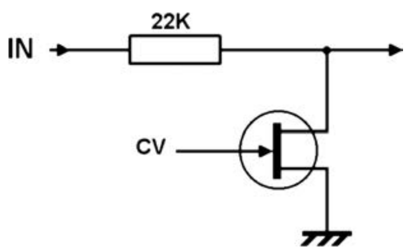

Figure 24.1: The basic voltage-controlled JFET attenuator circuit

With the JFET conducting, the major non-linearity is second-harmonic distortion, which can be much reduced by adding half the drain-source voltage to the gate control voltage, as in Figure 24.2. Note that it is half the drain-source voltage that is applied, not half of the input signal. The ratio is not critical – getting it right within 10% seems to give all the linearisation available. A serious problem with this circuit is the control-voltage feedthrough into the signal path. One answer is to add a DC blocking capacitor C as in Figure 24.3; this stops DC, but does nothing to stop fast transient feedthrough.

Another problem is that the presence of C adds an extra low-pass time-constant to the applied control voltage. In feedback limiters this can cause instability of the control loop, typically

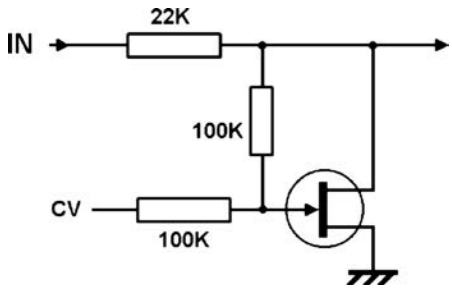

Figure 24.2: Second-harmonic distortion can be much reduced by adding half the drain-source voltage to the gate control voltage

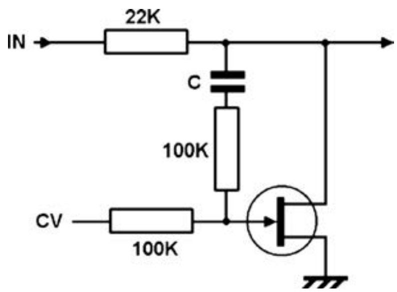

Figure 24.3: Adding blocking capacitor C prevents DC getting into the signal path, but does nothing to stop transient feedthrough

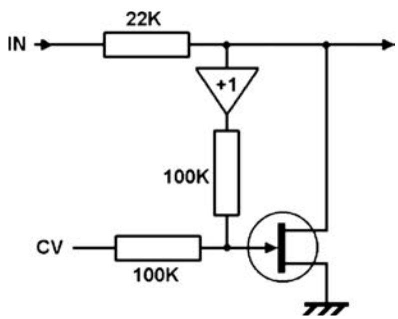

Figure 24.4: Using an active buffer instead of a blocking capacitor prevents both DC and transient feedthrough

causing the gain reduction to suddenly snap to the maximum value, followed by a slow decay back to normal operating conditions. This is not a good thing. Both problems can be solved by using a unity-gain buffer amplifier to isolate the JFET gate network from the signal path, as shown in Figure 24.4.

Because of their unhelpful control-laws, JFETs were most useful in feedback-type compressors and limiters, where the feedback loop linearised the law; more on that later. I produced a compressors/limiter design for *Wireless World* [1] when the application of FETs in this way was relatively new; it was published some years after I designed it. The linearity limitation remains, and JFETs are now rarely used in compressors and limiters (except for those which deliberately embrace obsolescent technologies), having been replaced by VCAs. They are however still useful in noise-gates, because as shown above, they can be configured so that when the noise-gate is open, the JFET is firmly off and introduces no signal degradation at all. A JFET is also much cheaper than a VCA.

Operational transconductance amplifiers

An operational transconductance amplifier (OTA) gives a current output for a voltage difference input, and the amount of current you get for a given input voltage is determined by the current that is made to flow into a control port, giving a variable-gain capability. The output stage is a high-impedance current-source rather than the low-impedance voltage-source of the conventional opamp, and the output current must be converted to a voltage, usually using a simple resistive load. Buffering is then needed to give a low-impedance output. Despite the name – operational transconductance amplifier – this device is not used like a conventional opamp when it is being used to give variable gain. There is no negative feedback around the device, as this would prevent the gain varying, and for acceptable linearity the differential input voltages should be kept to 20 mV or less. Distortion is mostly third harmonic, and comes from the input pair transistors.

The best known operational transconductance amplifier was the CA3080E and a typical voltage-controlled gain circuit for it is shown in Figure 24.5. A1 is the OTA, R1 and R2 reduce the input level so it is suitable for the device input transistors, R3 is the I/V conversion resistor and A2 is a conventional opamp acting as the output buffer. The THD is about 0.15% for a +5 dBu input level. Note the OTA symbol has a current source symbol attached to its output.

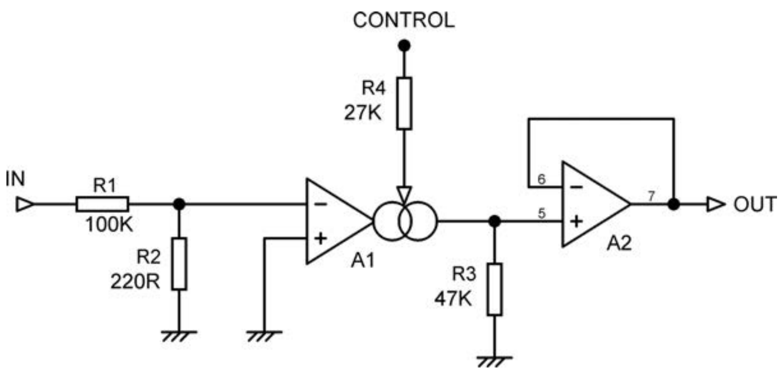

Figure 24.5: A typical gain-control circuit using the 3080E operational transconductance amplifier

Another OTA, called the LM13600, was introduced later by National Semiconductor; this is a dual part and has distortion-compensation diodes in the input stage, which are claimed to have improved linearity fourfold. I have to say that I found that the difference in practical applications was not that large. It also has built-in Darlington output buffers. The LM13600 has now been replaced by the LM13700; the CA3080E went out of production in 2005. The LM13700 is used in much the same way as the CA3080E.

You are probably thinking by now that transconductance amplifiers belong in the history section of this chapter; they are placed here because their flexibility has kept them very popular with builders of analogue synthesisers.

Voltage-controlled amplifiers

Voltage-controlled amplifier (VCA) is the name given to a specific kind of variable gain device. It is essentially a four-quadrant multiplier and is a current-in, current-out device, which therefore needs some support circuitry if it is to interface with the normal voltage-driven world.

Since the input port is a virtual-earth, converting the input voltage into an input current requires only a resistor; see R1 in the basic circuit of Figure 24.6. Converting the output current into an output voltage is less easy and requires a shunt-feedback amplifier A1, as it is essential to avoid signal voltages on the output port. This is a transadmittance amplifier (current-in, voltage-out). An important advantage of this circuit is that is does *not* phase-invert. It looks as if it does – the amplifier A1 certainly inverts – but so does the VCA internally, so the output remains in phase, which is of course essential for mixing console use.

Because of their log/antilog operation VCAs are characterized by an exponential control characteristic, so gain varies directly in decibels with control voltage. This is extremely convenient as it means that a linear fader can be used to control gain over a wide range; however, this is in practice less than ideal as it spreads out the less important higher attenuation range too much. Special 'VCA-law' faders are made that squeeze the high attenuation range so the normal operating range can be expanded.

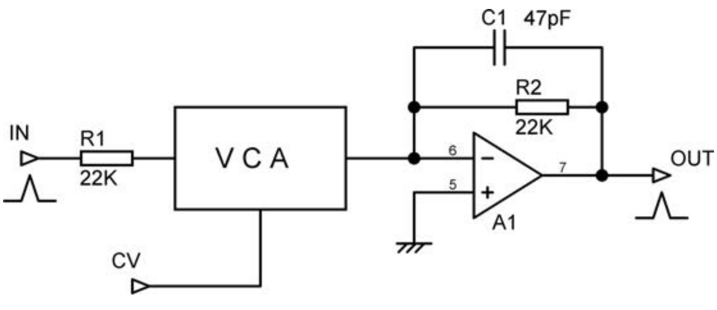

Figure 24.6: The basic gain-control circuit using a VCA

The evolution of VCA technology started in about 1970, based on the 'Blackmer gain cell' developed by David Blackmer of dbx, Inc., and VCA history is a fascinating field in itself; see reference [2]. Another of the very early models was the Allison EGC-101, which gave improved linearity through Class-A operation [3]. The study of the internal operation of VCAs is also a big subject, and regrettably I don't have the space to go into the details here. An article in *Studio Sound* [4] gives a good deal of information on the internals. Other useful sources of information on the technology are references [5] and [6].

For many years the 'standard' VCA was the DBX2150, of which I have deployed more than I care to contemplate in VCA sub-group systems and console automation; more modern ones are represented by parts like the THAT 2181. A typical application circuit suitable for both is shown in Figure 24.7.

You will note that a symmetry trim pot is required; this is set to minimise second harmonic generation. A THD analyser is required to make this adjustment, but on the positive side, once set it stays set, and need never be touched again unless the VCA is replaced. Modern VCAs are very good; the THD at 1 Vrms with 0 dB gain can be as low as 0.002%. The off-isolation (or offness, as I prefer to call it) can be as good as −110 dB at 1 kHz, but will almost certainly be worse at higher frequencies as it depends on stray capacitance. This is why the SIP package is preferred; as shown in Figure 24.7, it keeps the input and output pins as far apart as possible. A crucial layout requirement is that the 'hot' end of R1 is kept as far

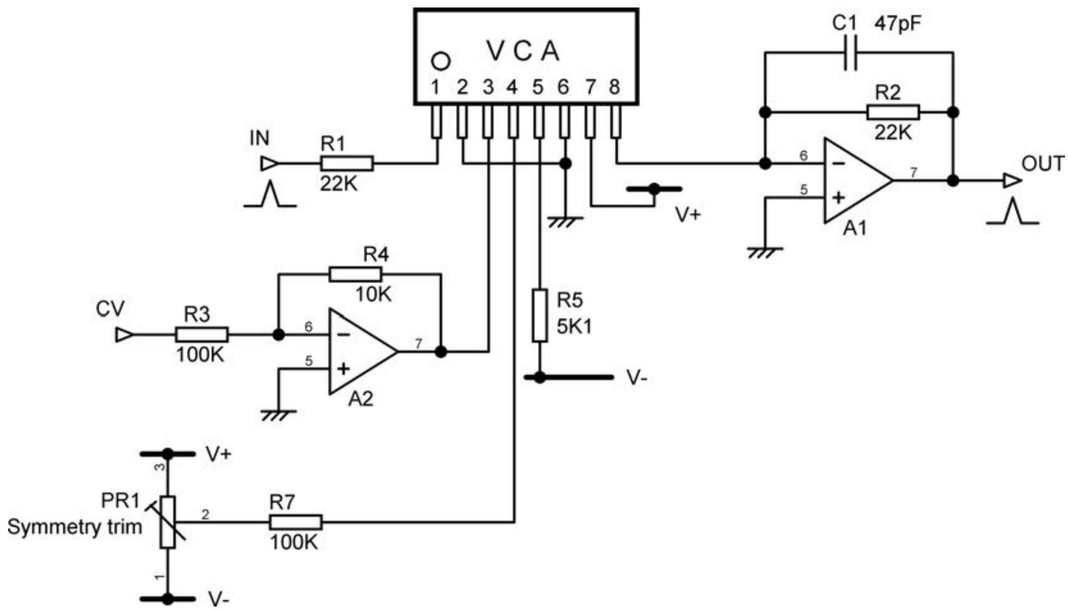

Figure 24.7: A typical gain-control circuit using the 2150 VCA. The 8-pin SIP package is almost always used for VCAs

away as practicable from Pin 8 of the VCA, even if it means extending the track between R1 and VCA Pin 1. This track is at virtual earth and susceptible to capacitive crosstalk so keep unrelated signals well away from it.

The control law (which is set by transistor physics, and is therefore dependable) is 6 mV/dB at the actual VCA control pin, and this is inconveniently low. A scaling amplifier A2 is therefore used so the control voltage has a more useful range, such as 0 to 10 V. The presence of this amplifier also gives the opportunity for control voltage filtering and changes of ground reference. The voltage applied to the control port must be at low impedance (basically an opamp output is the only source that will do) and absolutely free from contamination by any sort of signal or avoidable noise. Contamination with even a trace of the signal being controlled will cause excess distortion.

The capacitor C1 shown across the I–V conversion opamp feedback resistor is always required for HF stability, due to the destabilising capacitance seen looking into the VCA output port.

In other places in this book I have described how noise can be reduced by using multiple transistors or multiple opamps in parallel. This also works with VCAs; you simply put N of them in parallel and connect their outputs to a single shunt-feedback amplifier with a suitably reduced feedback resistor, as shown in Figure 24.8. Two VCAs give a 3 dB improvement and

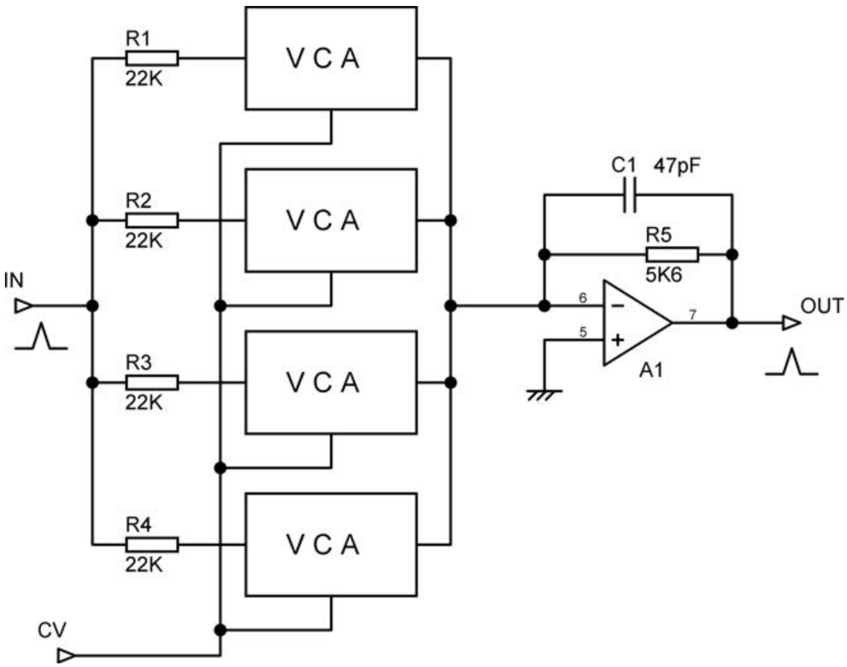

Figure 24.8: Using multiple VCAs in parallel to reduce noise. Each doubling of numbers theoretically reduces noise by 3 dB

four a 6 dB benefit. There are of course limits as to how far you can go with this sort of thing; high-quality VCAs are relatively expensive. Four in parallel have been used in high-quality compressor/limiters such as the Connor. Eight VCAs are employed by That Corporation in their 202 module which is used in high-end consoles such as those by SSL.

Compressors and limiters

Compressors and limiters are devices that control the dynamic range of a signal. The device acts like a rapid volume control that reduces the gain when the signal level becomes excessive.

A compressor reduces the general dynamic range of a signal for the majority of the time. A typical application is control of microphone levels when the talent does not make microphone technique their top priority. This means that it is applying some amount of gain reduction most of the time, so the gain-control element is active and it is important that it is relatively distortion-free and without other audible defects.

A limiter, by contrast, has as its main function the prevention of overload and horribly audible clipping. When loudspeakers and AM transmitters are being driven it may actually prevent expensive equipment damage. Transmitters require the extra protection of a clipper circuit; see the separate section on these. Since a limiter operates relatively rarely (assuming the system is being operated correctly) it is less important that the gain-control element is distortion-free.

Figure 24.9 shows the relationships between input and output levels for compression and limiting. The compressor law begins gain reduction at a lower threshold level, and has a

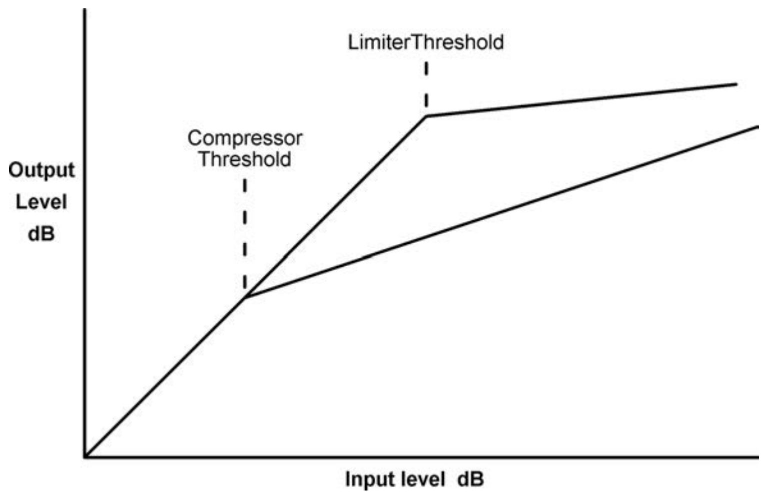

Figure 24.9: Compressor and limiter laws

moderate slope thereafter; the slope shown here is 3:1, which is typical. The limiter law has a higher threshold and only acts when overload is imminent; the slope is much flatter so that even very high signal levels cannot reach the overload point; the slope here is 10:1.

Compressors and limiters do a similar job, using much the same hardware with different parameter settings, so the functions are often combined in a compressor/limiter. Figure 24.10 gives the block diagram of a feedback compressor/limiter. The output signal is amplified, and if it exceeds a certain threshold, applied to a rectifier with a fast-attack slow-decay characteristic. This part of the system is called the sidechain, to emphasise that the signal does not pass through it. The resulting control voltage is applied to the gain-control element and as the signal level increases the gain is reduced, reducing the variations in output level. The sidechain may use either peak or RMS sensing; the latter is considered by some to relate better to our perception of loudness and give a less obtrusive effect.

The laws shown in Figure 24.9 have 'hard knees' as the gain laws change abruptly at the threshold. A 'soft knee' slowly increases the compression ratio as the level increases, giving a curve that gradually attains the desired compression ratio. A 'soft knee' is considered to make the change from uncompressed to compressed less audible, especially for higher compression ratios.

If the threshold is set low, and the sidechain gain is also low, the system works as a compressor, the gain reduction acting to reduce the general dynamic range of the output signal, hopefully without obvious side-effects. If the threshold is set high, and the sidechain gain is also high, we have a limiter instead. The signal will be untouched until it exceeds the threshold, and then gain reduction is applied strongly to prevent the output level significantly increasing.

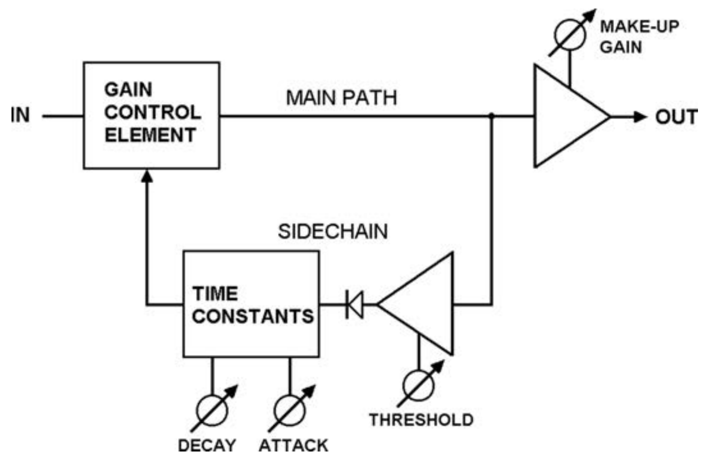

Figure 24.10: Block diagram of a feedback compressor/limiter

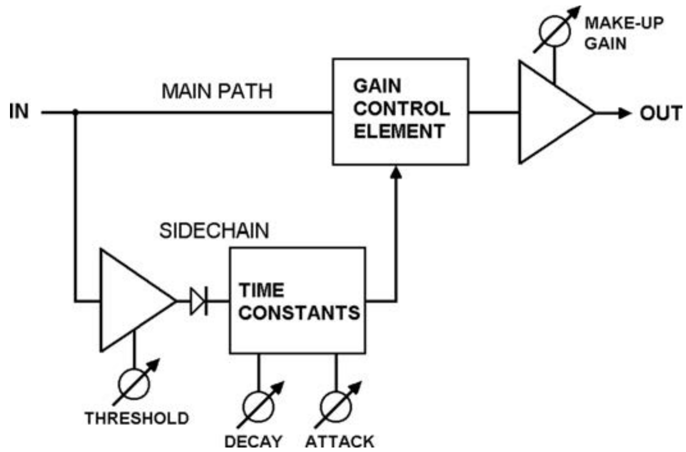

Figure 24.11: Block diagram of a feedforward compressor/limiter

Compressor/limiters can work in either feedforward or feedback modes, each of which has its own advantages and problems. In the feedback mode the compression law tends to be inherently linear because of the feedback around the level-control loop. The feedback configuration has the advantage that the output level set does not depend on the control-voltage law of the gain element.

With feedforward, as in Figure 24.11, the compression law depends on the control-voltage law of the gain element. If this is a VCA with a very predictable law, no problem. However, other gain elements such as FETs sometimes have their uses – with FETs the advantage is that there is no signal degradation when there is no gain reduction. There is, however also a highly non-linear control-voltage law to deal with, and it is not practical to bend this into a more desirable log law.

The way to solve this is shown in Figure 24.12, which assumes that two matched gain elements are easier to make than one with the required law. The sidechain path contains a feedback limiter and the control voltage this generates is fed to the second gain element, which acts in feedforward mode. Typically the gain elements are matched as well as possible by trimming.

So far we have assumed that the sidechain dynamics consist only of two simple time-constants for attack and decay. In many cases this will give rise to objectionable effects, and more sophisticated control measures have been developed to deal with the problems. Even if a gain-control element is completely linear when the control voltage is fixed, rapid CV changes put distortion into the waveform.

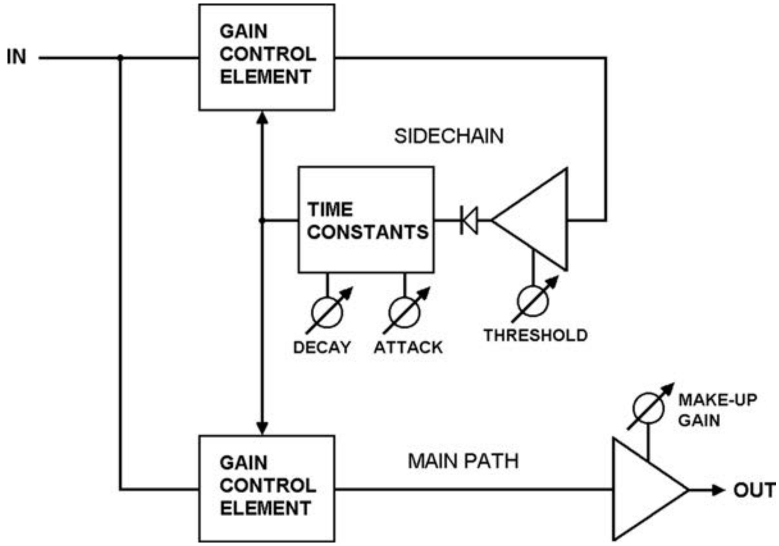

Figure 24.12: A combined feedforward/feedback compressor/limiter

Attack artefacts

When a fast attack is used to control level (typically when the unit is being used as a limiter) and prevent overshoot, it tends to bite chunks out of the controlled waveform, as a natural consequence of a rapid drop in gain. Isolated cases of this usually pass unnoticed, but repeated occurrences are perceived as a crackling noise. This can be controlled by the use of dual attack times, with the faster time switched in if the signal peak exceeds criteria for amplitude and rate of rise, but the only complete solution to this problem is a delay-line compressor/limiter, as seen in Figure 24.13.

A delay in the signal path between the side-chain takeoff point and the gain element allows a feedforward sidechain to react relatively slowly and still have the gain reduced before the signal peak reaches the gain-control element (this is not possible with the feedback compressor/limiter configuration because the sidechain feed is taken *after* the gain-control element, and therefore after the delay). The difficulty is that the signal is delayed at all times; this will cause problems if it is being mixed with undelayed signals, and furthermore the delay section has to be of high quality as it passes the main signal, not just the sidechain information. A well-known BBC design of circa 1967 used a strictly analogue 320 μsec delay-line made up of ten LCR second-order all-pass filter sections [7]. This worked very well but is obviously an expensive technique. An active filter delay-line would be cheaper but still involves a significant amount of circuitry and a number of close-tolerance components. Nowadays a high-quality digital delay using 24-bit converters can be constructed relatively easily.

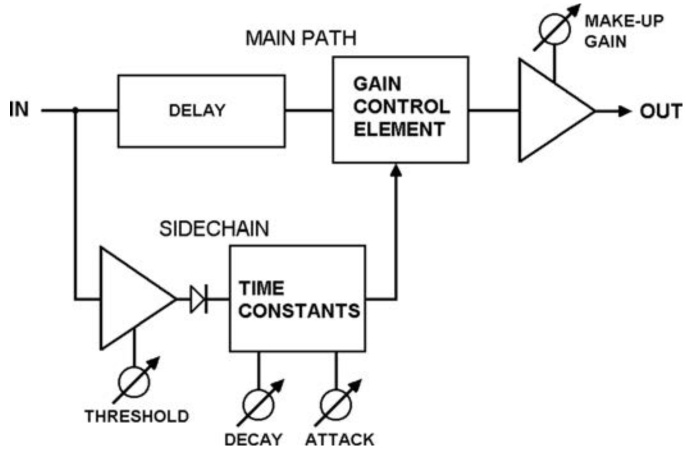

Figure 24.13: Block diagram of a feedforward compressor/limiter with delay

Decay artefacts

Since the sidechain contains either a peak or RMS detecting rectifier, the timing capacitor will have on it a ripple waveform resulting from cyclical charge and discharge, exactly as in a power-supply reservoir capacitor. This waveform modulates the signal path gain and generates distortion; the effect can be severe. In most compressor/limiter usage the attack time has to be fast, but often the decay time can be made much longer, reducing the amplitude of the ripple and reducing distortion. However, in some applications the decay has to be short for rapid recovery from brief level excursions, or the general level of the signal will be unduly depressed. An effective and widely-adopted solution to this problem is the use of dual time-constants. See Figure 24.14 for a simple implementation, where the short time-constant R1, C1 responds to short-term level changes while the long time-constant R2, C2 copes with the general trend.

Another elaboration is a hold circuit which prevents the decay from beginning if any attack events have occurred in, say, the preceding 20 msec. This helps to prevent sudden and obvious gain changes, often called by the highly descriptive term 'pumping'. If the noise background is significant, as with many live news recordings, then even slow gain changes produce a distracting effect called 'breathing', because that's just what it sounds like.

Subtractive VCA control

While modern VCAs are very good, they still introduce more noise and distortion than a typical well-designed chain of 5532 opamps and, as mentioned above, FETs are therefore still favoured for some compressor/limiter applications as they introduce no degradation when

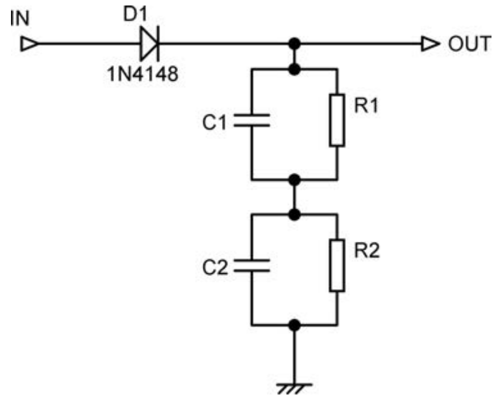

Figure 24.14: Dual decay time-constant circuit

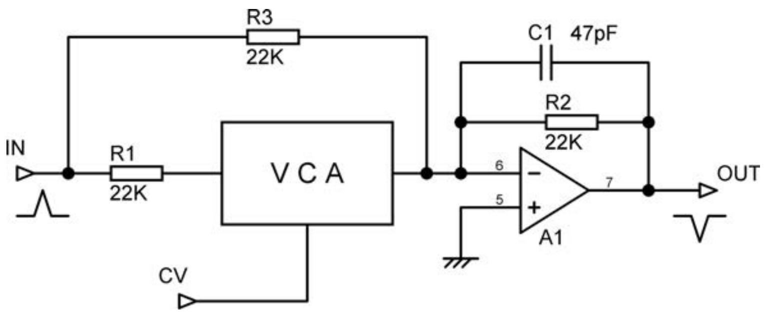

Figure 24.15: A VCA gain control arranged so that no signal passes through it when no gain-reduction is occurring

not bringing about gain reduction. However, the predictable characteristics of VCAs are very attractive, and it would be nice to combine the advantages of both.

This can be done; whether I invented the technique first I have no idea, but it was all a long time ago. The concept, as shown in Figure 24.15, is not to send the signal through a VCA with 0 dB gain when no gain reduction is required, but to have the VCA normally hard off and then turn it on to reduce the gain. When the VCA is off, the circuit acts as a simple unity-gain inverting amplifier, and there is minimal signal degradation.

Note that point about inverting – this circuit phase inverts although the standard VCA circuit does not. As the VCA gain is turned up, it lets more signal current through to the summing point of A1, and since this is out of phase, due to the internal inversion of the VCA, there is partial cancellation and the overall gain falls. When the VCA is set to 0 dB gain theoretically there is complete cancellation and the overall circuit is off. In practice the tolerances in R1, R3 and in the VCA set a limit to the practicable minimum gain, but the range is more than enough for effective dynamics control.

You may have spotted the potential snag. If the gain of the VCA goes above 0 dB (which corresponds with a control voltage of 0 V) the overall gain will begin to rise again. With a feedback type compressor/limiter this could be disastrous as the sidechain will try to reduce the gain, but it will simply increase it further, and the system will latch up solid. This can be prevented by the simple expedient of using an active clamp circuit to prevent the control voltage rising even a millivolt above 0 V. Note that because of the subtractive action the gain law is no longer linear in dBs, and so this plan is better suited to feedback compressor/limiters. I put this system into a broadcast console in 1990 and it worked very well.

Noise gates

A noise gate allows a signal through only when it is above a set threshold; the gate is said to be open or on. When the signal level drops below the threshold, the signal is either attenuated or stopped altogether and the gate is closed or off. For effective noise reduction the level of the signal must be above that of the noise; the threshold is set above the noise level so when there is no signal the gate closes. Noise gates are not actually often used to discriminate against noise as such; in recording they are more likely to be used to, for example, increase the isolation between the signals coming from a multi-mic'd drum kit. They are also used to provide the well-known 'gated reverb' effect, where decaying reverb is cut off suddenly by a noise gate.

A noise gate is a rather different animal from a compressor/limiter. Its operation is much more on/off, and long time-constants are not normally used. This means that the non-linearity of an FET during intermediate degrees of gain reduction can be tolerated. In the traditional form of FET noise-gate, as shown in Figure 24.16 and based on a design I did some years ago, FET distortion is reduced by attenuating the signal considerably – in this case by 32 dB – before applying it to the FET stage. This is done by R1, R4 and the signal is at a low impedance afterwards to keep noise down. The FET (actually two FETs here, to reduce $R_{ds(on)}$ and give a greater gain change) is at the bottom of the NFB network of a low-noise hybrid amplifier stage that restores the signal to its original level when the FET is on. When the FET is off, the gain is reduced to unity. This would only give an offness of −32 dB, which is not enough, so the network R9, PR1, R8 is added, which lets a little signal through to what is effectively an inverting input to the amplifier stage; when PR1 is correctly adjusted there is effective cancellation and the offness can easily exceed −80 dB. FET distortion is further reduced by adding half the drain-source voltage to the gate control voltage via buffer A3, as described earlier.

The low-noise amplifier is a hybrid stage, combining the low-noise of the discrete input transistor Q1 with the open-loop gain and linearity of opamp A1. R2, R6 set up the DC conditions for Q1, while servo integrator A2 defines the DC operating point of A1, keeping its output at 0 V on average. Q2 is a current-source that helps keep rail noise out of the collector circuit of Q1. The alert reader will spot the similarity between this stage and the moving-coil preamplifier stage in Chapter 12. The noise from the amplifier with the gate off is −106 dBu.

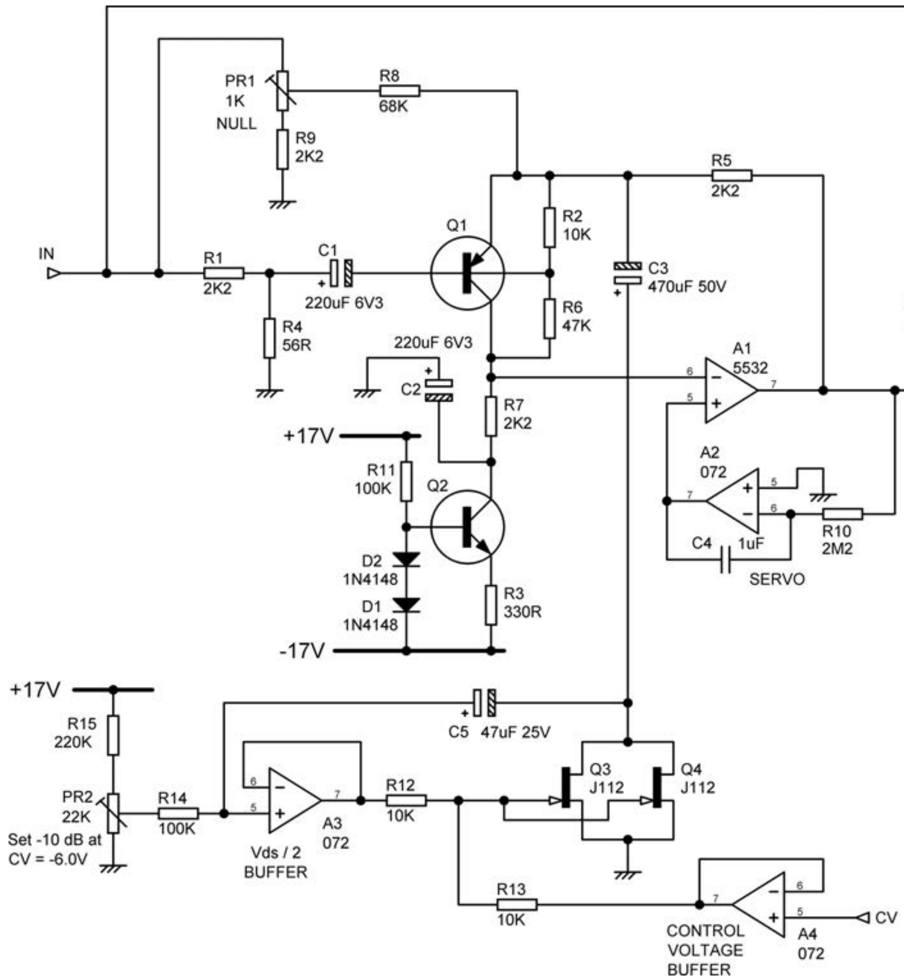

Figure 24.16: A typical FET noise gate

Many noise gates have a 'range control' that allows the offness to be reduced if a more subtle effect is required. This is RV1 in Figure 24.16, an ordinary log pot that has the bottom of its track, which would normally be grounded, connected instead to the noise gate output. Offness is reduced by advancing the control so that some of the signal gets through even when the gate is off.

The sidechain of a noise gate is similar to that of a compressor/limiter, but there is a greater emphasis on fast attack times of 50 µsec or less, and special circuitry is used to charge the timing capacitor as quickly as possible.

Noise gates can also be made with VCAs, and this is the more usual method when a 'dynamics section' which can act as either a noise gate or a compressor/limiter, is squeezed into a mixer channel.

Clipping

Since clipping and its attendant distortion is something we normally strive to avoid, it may be thought perverse to study ways in which to induce it. Clipping circuits do have their uses, however. One application is in the protection of AM transmitters, where even momentary excursions beyond 100% modulation are unacceptable, because they lead to legal problems (sideband splatter causes the regulator to close you down) and technical problems (the transmitter blows up). Either way you're off the air. Limiters are the first line of defence, but they are liable to overshoot and maladjustment. A fixed-level clipping circuit, however, can absolutely prevent any signal exceeding its threshold. Clipping circuitry also has specialised uses in power amplifier design, where it can emulate the performance of a much more complex and expensive regulated power supply [8], or increase the flexibility of bridged amplifiers [9].

To set down some of the requirements for the perfect clipping circuit:

1. Clipping must be at well-defined symmetrical levels, constant with frequency and not dependant upon signal history.
2. The top of the clipped waveform should be absolutely flat, to give tight level control.
3. The clipping level must be arbitrarily settable, without steps.
4. There must be absolutely no degradation of the signal below the clipping threshold – or at any rate no more than would be caused by going through an ordinary single opamp stage.

This last requirement is actually much more demanding than it appears, but it is essential. Let us examine the problem. I appreciate that clean audio clipping may be a bit of a minority interest, but it will be highly instructive to see just what difficulties arise, and how they can be overcome.

Diode clipping

If you look up 'clipping circuits' in the average textbook, you will probably find something like Figure 24.17, where back-to-back diodes conduct when the signal exceeds their conduction threshold.

This fails to meet our requirements in several ways. Firstly, the gradual onset of conduction in the diodes means that the clipping is soft, infringing requirements two and four. And, secondly, it breaks three as well, because the clipping level can only be altered by changing the number of diode pairs used, giving 0.6 V steps.

Figure 24.18 shows the force of this, the result of using various numbers of pairs of 1N4148 silicon signal diodes. The lower line descending from left to right is the noise floor, getting

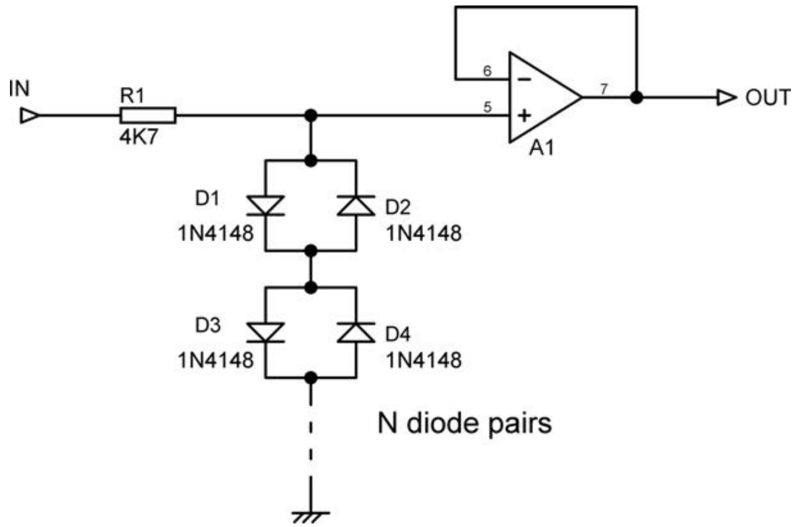

Figure 24.17: A simplest passive clipping circuit using diodes

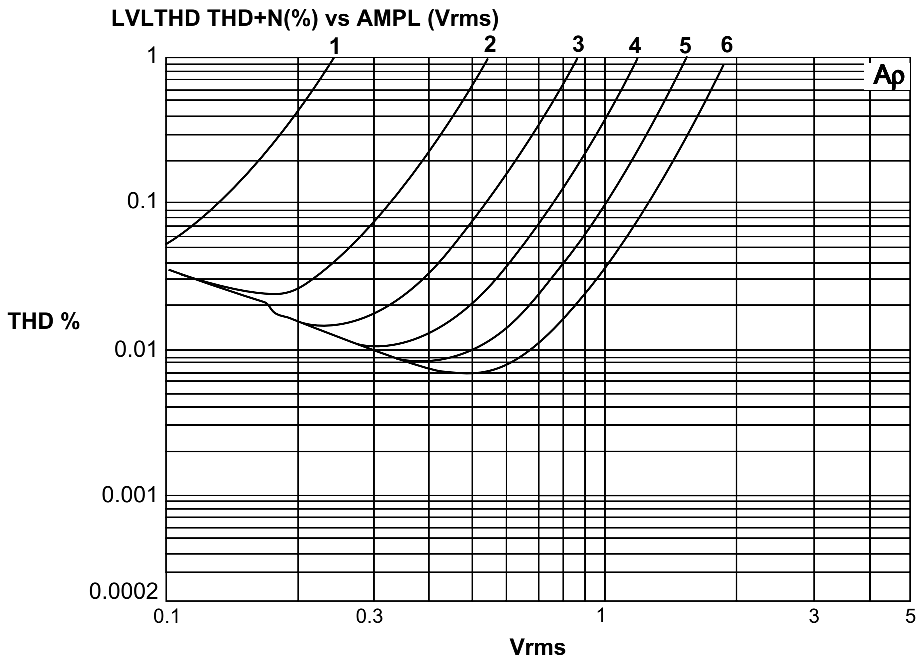

Figure 24.18: Distortion against input level for 1 to 6 pairs of clipping diodes. The uneven spacing of the curves as diode pairs are added is due to the logarithmic X-axis for input level

relatively lower as the input level increases. With one pair of diodes distortion is already clearly above the noise floor for an input as low as 100 mVrms. The gradual increase in distortion after this as level increases shows that the clipping is distinctly soft, and so introduces unacceptable distortion long before it effectively controls the level. An obvious extra snag is that a passive circuit like this has a significant output impedance; in most applications some sort of output buffering like A1 will be required.

The very gradual onset of clipping shown above is due to the slow way in which the exponential conduction law of diodes begins. The clipping action can be made much sharper by connecting each diode to a suitable bias voltage so that it is firmly reverse-biased at low signal levels, and only conducts when the signal is significantly above the bias voltage. The bias chain must have a low impedance to give a steep clipping characteristic, and so diodes are also used to establish the voltages. It would, of course, be possible to generate near-zero impedance supplies by using opamps to buffer resistive dividers (which also allow complete flexibility in setting the clipping threshold). However, if opamps are to be employed there are better ways to use them, as we shall see later in this chapter.

A biased clipping circuit is shown in Figure 24.19, and the resulting distortion performance in Figure 24.20, for both one and two pairs of diodes in the biasing chain. The much steeper rise in distortion shows that the onset of clipping is much sharper; compare Trace 6 in Figure 24.18 with Trace 1 in Figure 24.20.

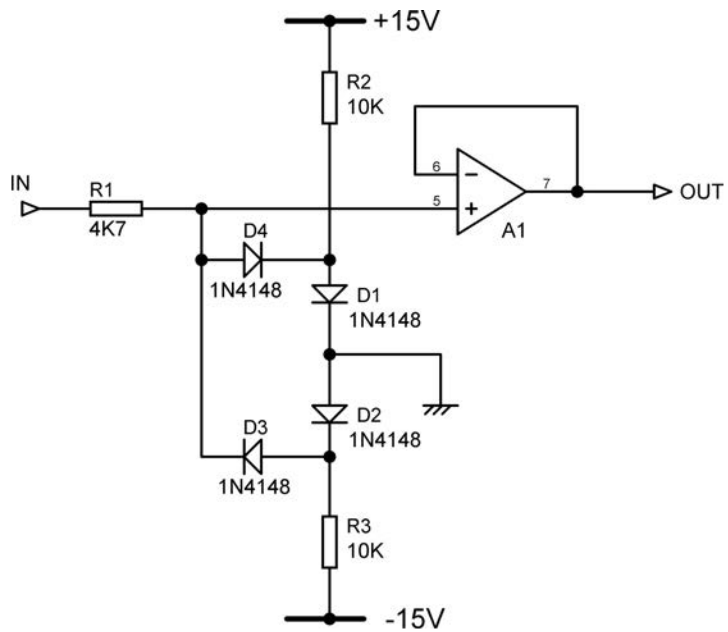

Figure 24.19: A biased diode clipping circuit

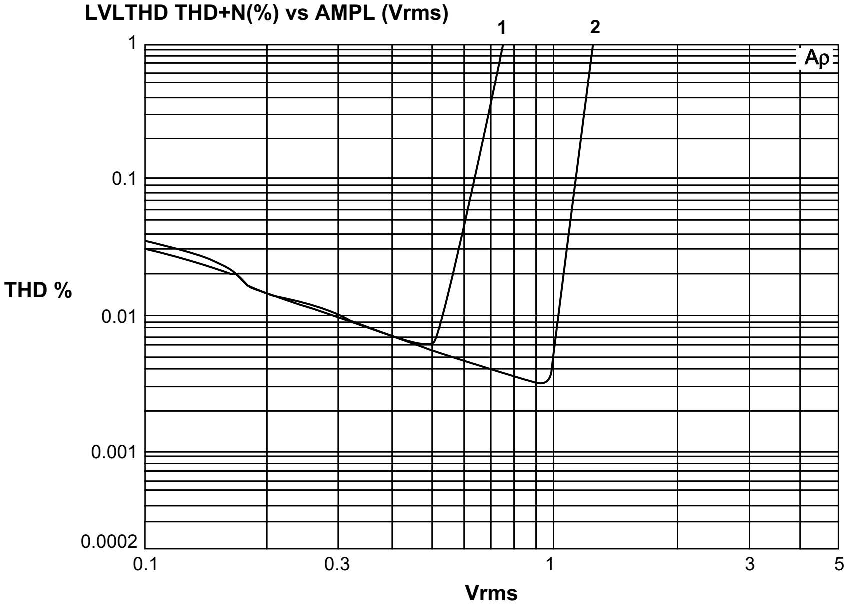

Figure 24.20: Distortion against input level for 1 or 2 pairs of diodes in the biasing chain

The clipping action of simple circuits like this is, however, still some way short of perfect, in that the clipped part of the waveform is not a dead flat horizontal line. Diodes do not suddenly become short-circuits when they start to conduct and the clipped part of the waveform bulges upwards somewhat. Clearly we need active circuitry to sharpen up the diode action.

Active clipping with transistors

The next step up in sophistication from simple diodes is the use of transistors for clipping. This is an active process, in that the voltage applied to the base turns on the collector current. It is not simply a matter of using the base-collector or base-emitter diodes to replace the simple diodes in the previous section.

Figure 24.21a shows a basic transistor clipper with NPN and PNP devices shunting the signal path, and Figure 24.22 shows the distortion performance. A buffer is required for a low-impedance output; trace A in Figure 24.22 shows the result without an output buffer and trace B shows the result with it. In the latter case not only has the noise floor dropped, due to less pickup on what is now a low-impedance output, but the distortion at a given level has risen significantly, as the low-pass action of a capacitive screened cable driven from a medium impedance has been eliminated.

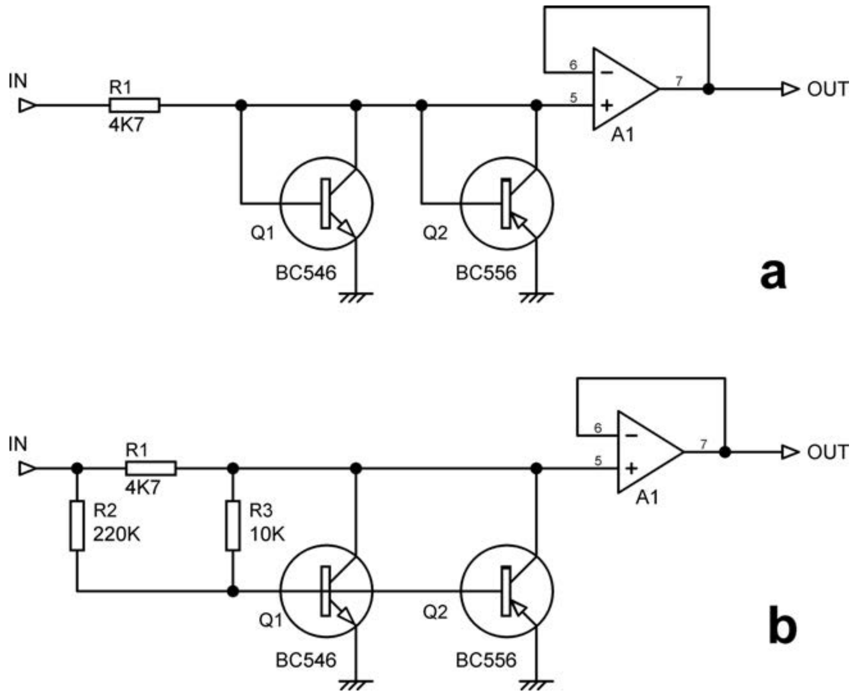

Figure 24.21: a) A simple transistor clipper, b) an enhanced transistor clipper with increased base drive via R2

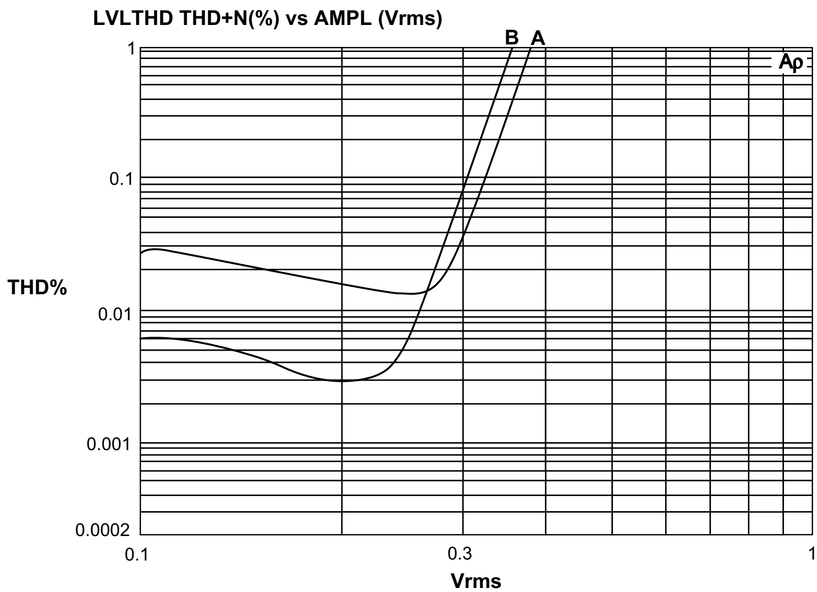

Figure 24.22: Distortion against input level for transistor clipper

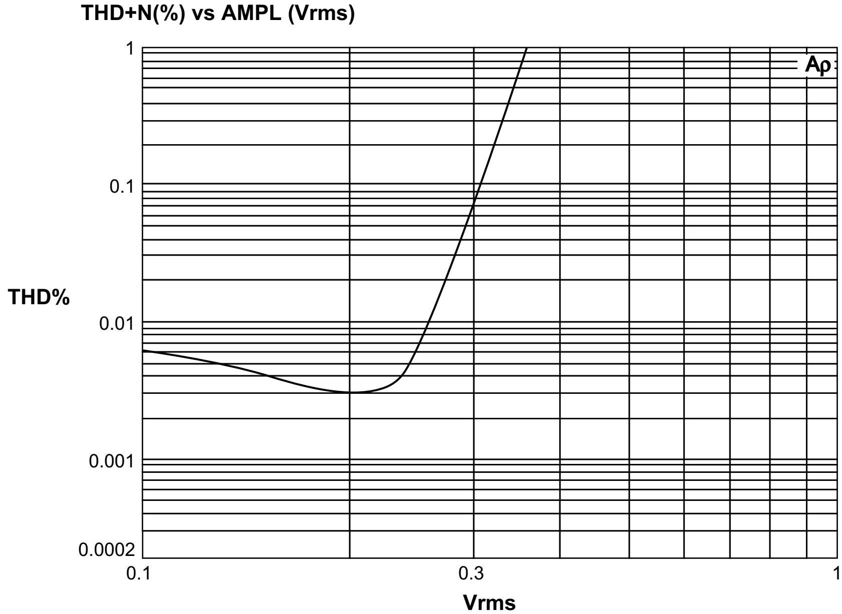

Figure 24.23: Distortion against input level for enhanced transistor clipper

In Figure 24.21b, the circuit is arranged so that extra drive is applied to the transistor bases via R2, because of the voltage drop across R1. This pulls down the clipped part of waveform in the centre, and with a correct choice of values gives improved flatness. The circuit is based on a concept by Stefely. The distortion plot in Figure 24.23 is not very different.

Active clipping with opamps

Opamps are noted for their versatility. However, using them to do precision clipping, with low distortion below the clipping point, is more difficult than it at first appears. Here I examine three ways of doing it, as the limitations of the first two approaches are highly instructive.

Clipping by clamping

The first attempt is the direct descendant of the diode clipping circuits examined above, and uses two active clamps A1, A3 to constrain the voltage at point A, downstream of the 4K7 resistor, as shown in Figure 24.24. Normally this point is between the two clipping thresholds, so the inverting input of A1 is positive of V_{t-}, and the opamp output is saturated negatively. D1 is therefore firmly reverse biased and A1 has no effect on the voltage at A. Likewise, normally the inverting input of A2 is negative of V_{t+}, and D2 is held off.

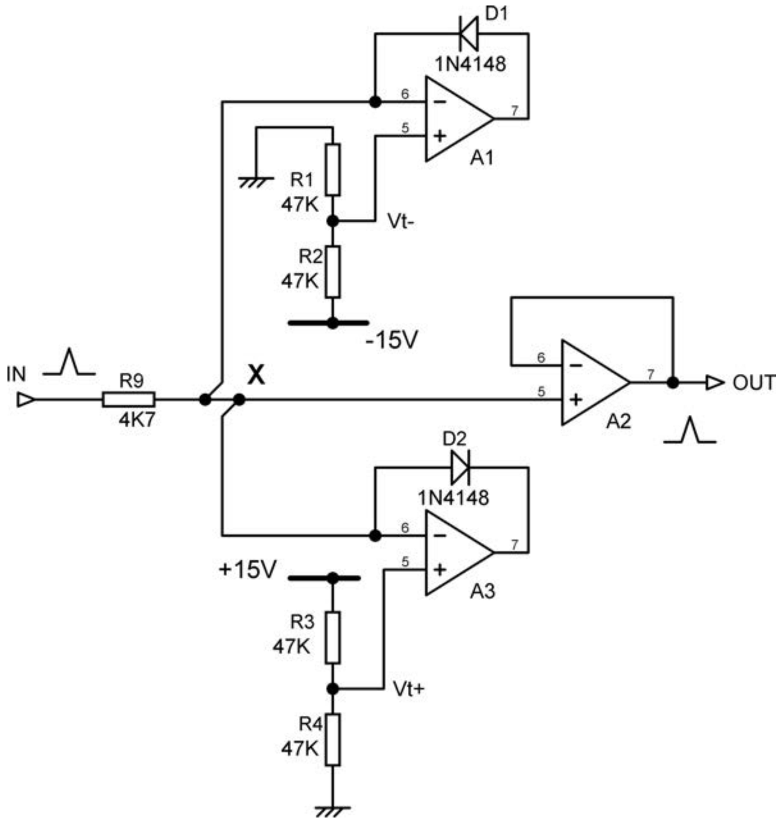

Figure 24.24: An active-clamp clipping circuit, with thresholds set at V_{t+} and V_{t-}. Once again an output buffer A2 is essential to give a low output impedance

When the voltage at X tries to exceed V_{t+}, the output of A2 swings negative and D2 pulls down point A to prevent it. The diode imperfections are servoed out by the open-loop gain of A2, so the clipping threshold is exactly V_{t+}, neglecting opamp offsets and other minor errors. In the same way, if point X tries to go below V_{t-}, the output of A1 swings positive and D1 conducts to clamp the output at this voltage. Like the passive diode clipper above, this circuit has a significant output impedance (4K7 in normal operation) and so buffer A2 is added to give a low output impedance.

The two clipping thresholds can be set to any desired voltage as they are derived from resistive dividers. They can of course be different, though an application where this might be useful is not that easy to visualise.

The clamping-clipper circuit gives very clean clipping, but falls down badly on the distortion it adds to signals below the clipping threshold of 2.2 Vrms. Figure 24.25 shows that distortion reaches an unacceptable 0.035% at 10 kHz and 2 Vrms; the 2 V trace is lower at LF because the relative noise floor is lower. This distortion occurs because point X is at a significant impedance, and is connected to two opamps with JFET inputs, typically TL072s. These have non-linear capacitances to the IC substrate, and cause distortion that worsens with frequency and level, as shown in Figures 24.25 and 24.26. See Chapter 4 for more details of this distortion mechanism. This effect could be eliminated by using opamps with bipolar

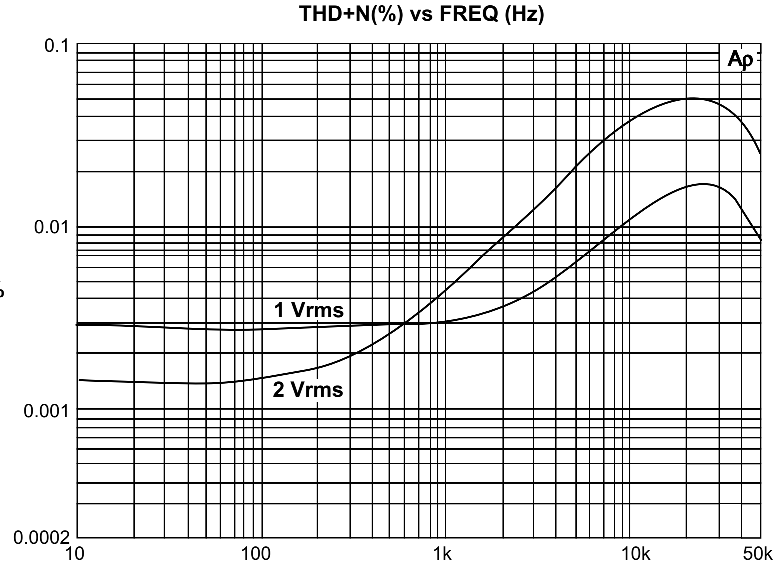

Figure 24.25: Distortion against frequency for the active-clamp clipping circuit

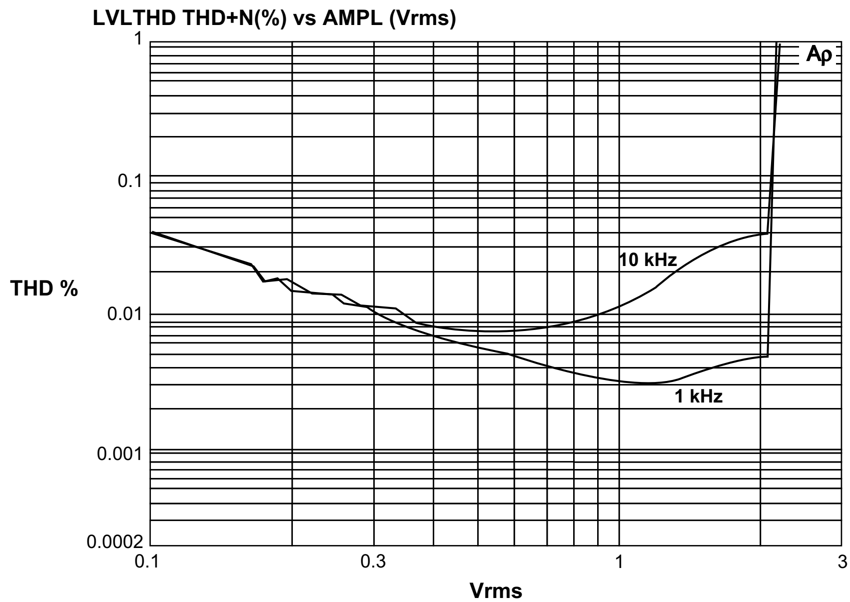

Figure 24.26: Distortion against input level for the active-clamp clipping circuit

inputs but then the bias currents have to be dealt with. One advantage of this circuit is that it does not phase-invert, unlike most active clipper circuits.

Figure 24.26 shows that at 10 kHz distortion is beginning to appear when the input exceeds 300 mVrms, and rises slowly until clipping begins just above 2 Vrms.

Negative-feedback clipping

Having found that clipping in the forward path has significant problems, we move to a configuration where this takes place in the negative-feedback loop, so the clipped output can be taken from an opamp output at low impedance. This also makes interfacing with the next stage simpler.

The circuit in Figure 24.27 works as follows: with no input, point X sits at +5 V and point Y sits at −5 V. When the output heads positive, eventually Y is pulled positive of the 0 V at the A1 inverting input, and D2 starts to conduct, reducing gain and giving clipping. Similarly, on sufficiently large negative output excursions, D1 will conduct. The clipping characteristic is rather soft; the gain cannot be reduced entirely to zero above the threshold because of the diode forward impedance in series with the source resistance of the 2K2–1K bias network. The latter can be reduced by using lower resistance values but this loads the opamp output excessively, and draws more power from the supply rails.

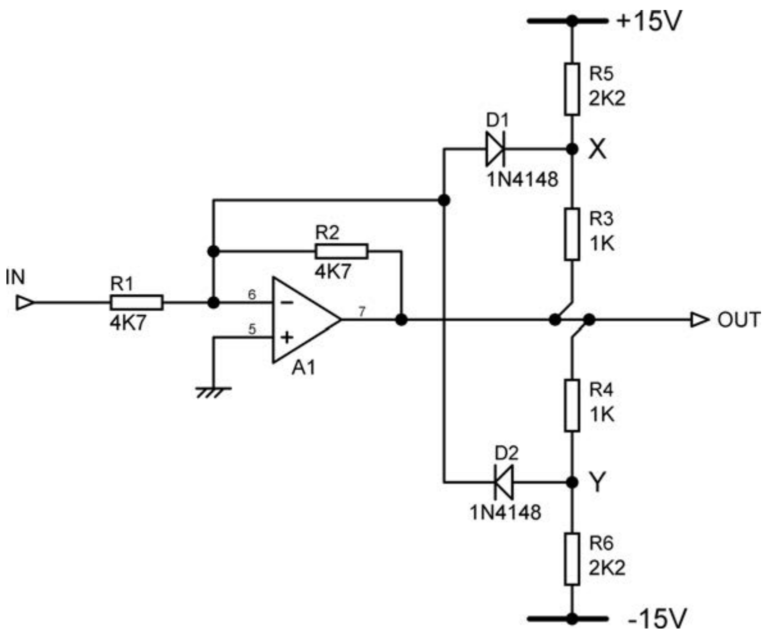

Figure 24.27: A negative-feedback clipper; the clipping threshold is set by the voltages at X and Y

This configuration gives less sub-threshold distortion than the previous circuit by a factor of roughly three at 10 kHz, but it is still a long way from our requirement for 'absolutely no degradation of the signal' which some of you are by now probably thinking was a bit ambitious.

The distortion for a 3 Vrms input, which is well below the clipping threshold for this circuit, is shown by curve A in Figure 24.28, and it is instructive to wheel out the scientific method to find out what is going wrong with the linearity. The previous clamping-clipper circuit ran into trouble by driving non-linear opamp inputs from a non-zero impedance; that cannot be happening here as both opamp inputs are at zero voltage, the inverting input so because there is always a healthy amount of shunt feedback.

So what might be the problem? We know that opamps have their linearity degraded by excessive output loading, and the two bias networks, being connected to rails that are effectively at AC ground, represent a significant load of (2.2K + 1K) in parallel with (2.2K + 1K) which works out to a total load of 1.6K on the output. This is quite enough to degrade the linearity of most opamps (see Chapter 4), but the idea is still only a hypothesis and needs testing.

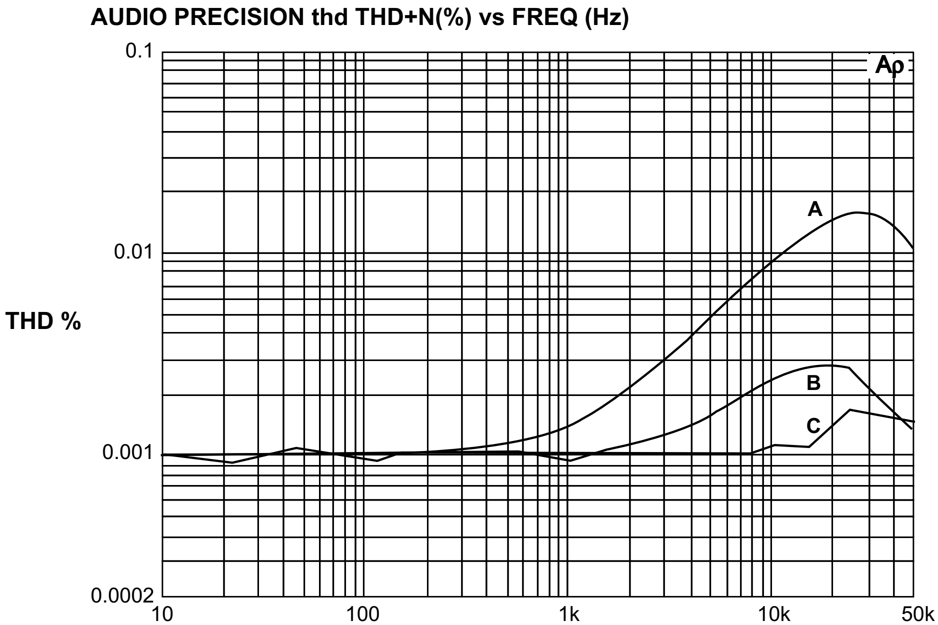

Figure 24.28: The distortion performance of the negative-feedback clipper against frequency for a 3 Vrms input. Plot A: basic distortion of Figure 24.27. Plot B: loading distortion eliminated by buffering A1 output. Plot C: with the diode chain disconnected

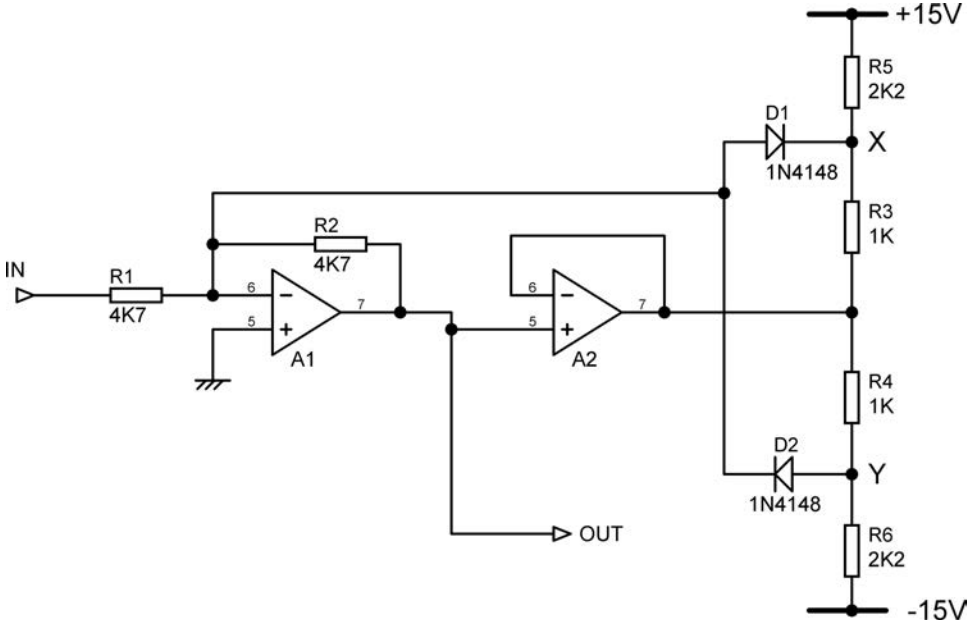

Figure 24.29: A modification of Figure 24.8 that removes the loading distortion from the signal path

This can be done experimentally by means of the circuit in Figure 24.29, which uses a separate voltage-follower A2 to drive the two bias networks, the clean output being taken off before it. Our hypothesis looks good, as eliminating the loading on A1 has much reduced the distortion, giving Plot B in Figure 24.28. This, however, still falls somewhat short of perfect linearity.

The distortion remaining in Plot B is due to negative feedback through the non-linear diode capacitances while they are still reverse-biased, i.e. below the clipping threshold. This can be demonstrated by disconnecting the output of A2 so clipping is disabled and points X and Y do not move. The diodes are still connected to the summing point at the inverting input of A1, which also does not move, and we now get Plot C in Figure 24.28, which is essentially the testgear distortion plus a little circuit noise.

Putting extra opamps in a negative feedback loop is not something to be entered into lightly or inadvisedly, because of the danger that accumulating phase-shifts can make the loop unstable. However, it often works if the extra opamp is configured as a voltage-follower because the 100% local feedback in the follower makes its bandwidth as great as possible and significantly greater than that of an inverting stage, which works at a noise-gain of 2. Here it works dependably.

While this circuit is some way short of perfection, it can be useful sometimes, especially if there happens to be a spare opamp half left over to act as the buffer A2 in Figure 24.29. Don't forget that this circuit gives a phase inversion.

Feedforward clipping

The Attempt 2 clipper demonstrated one way to reduce sub-threshold distortion, but it also highlighted the difficulty of getting hard clipping and tight level control by using negative-feedback techniques. While feedback is a most powerful technique, it is not the only way; sometimes feedforward does it better.

The circuit of a feedforward clipper is shown in Figure 24.30. Below the clipping level it acts simply as a unity-gain inverting stage, with the forward path through R1. This configuration eliminates common-mode distortion in A2 during normal operation.

The clipping circuit consists of two shunt-feedback precision rectifiers A1, A3 that are biased by currents injected into their summing points via R2, R3 so they conduct only above the desired clipping threshold. A1 handles negative clipping; R2 injects 15V/47K = 319 μA into the summing node of A1, and as long as this exceeds the current being pulled out of this node via R5 during negative inputs it will be counteracted by feedback through D2. When the current through R5 exceeds that injected by R2, the opamp output must go positive so it can

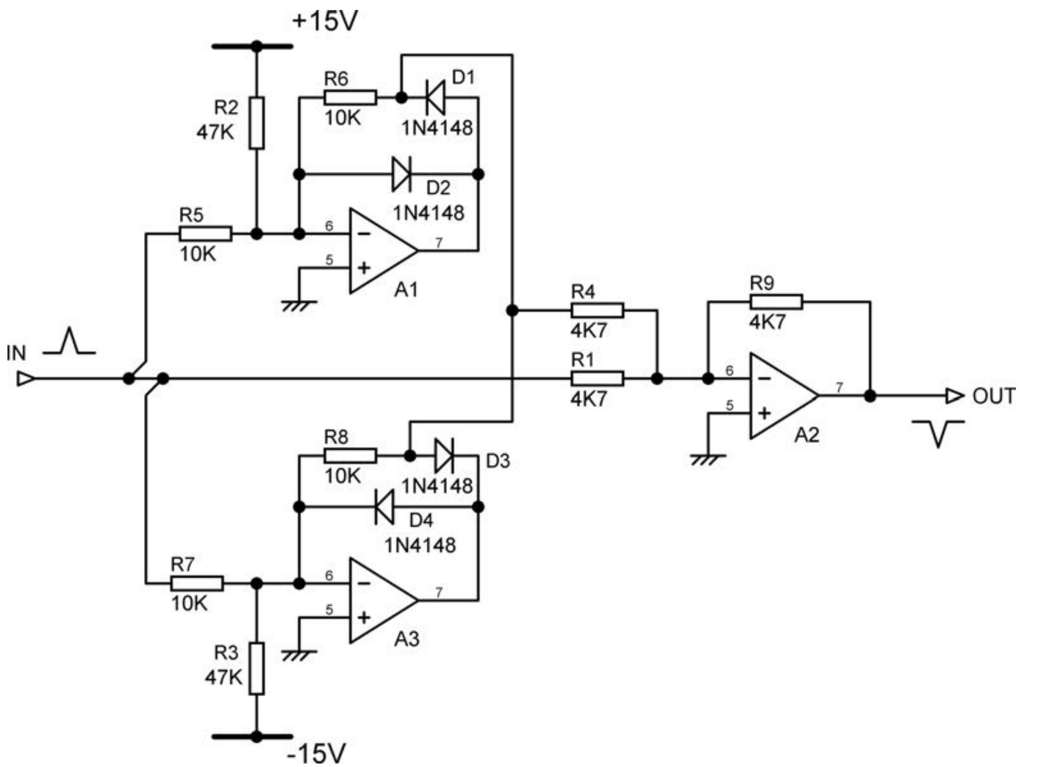

Figure 24.30: A feedforward clipping circuit. The clipping threshold is 2.25 Vrms with the values shown

maintain its inverting input at ground by feedback through D1 and R6. As before, the gain of the opamp A1 is used to 'servo-out' the non-linearity of the diodes D1, D2.

When D1 conducts, a clamped version of the input signal appears at the junction of R6 and D1 and a unity-gain but phase-inverted path is established through R4; the overall gain suddenly drops to zero as the two signals are cancelled at the summing point of A3, so the output cannot move any further and is clipped. The precision rectifier A3 acts in the same way for positive inputs.

Since A1 and A3 are never active at the same time, they can share the resistor R4, which makes for a very elegant circuit. The downside of this economy is that the active precision rectifier has to also drive the feedback resistance (R6 or R8) of the inactive precision rectifier, its other end being connected to virtual ground, and this will use up slightly more power. From this point of view the saving of a resistor is not quite as elegant as it looks; if power economy is paramount then separate cancellation resistors to each precision rectifier remove the problem.

As Figure 24.31 demonstrates, this circuit is a success. There is now no measurable distortion, even just below the clipping threshold (the step around 20 kHz is an artefact of the testgear used). The reason for this is that there are no opamp loading issues and no reverse-biased diodes with signal on one side connected to the audio path on the other side. The two clamp stages have zero output until the threshold is reached, with R4 being connected only to a virtual ground.

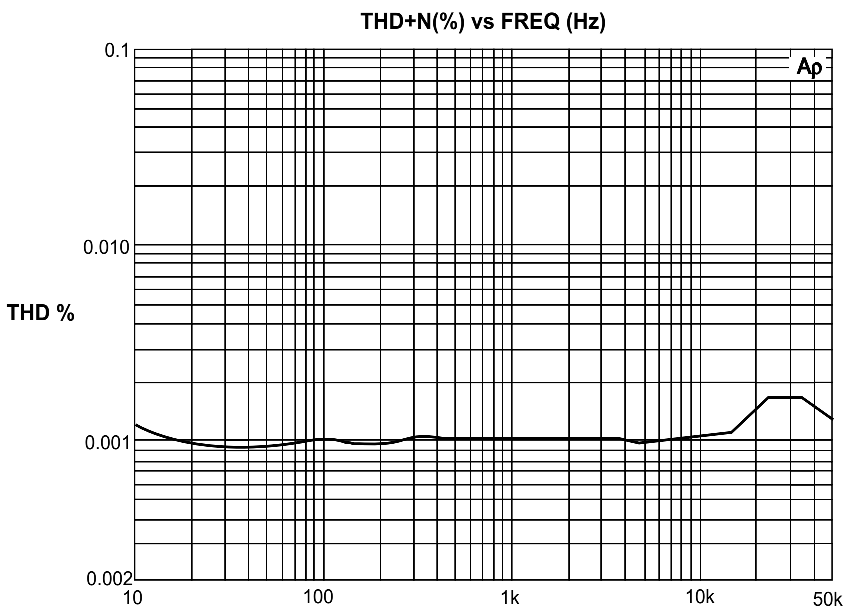

Figure 24.31: Distortion against frequency just below the clipping threshold. Input level 2 Vrms. This is once again the distortion of the testgear plus a little added noise

With perfectly accurate resistors, and indeed perfect opamps, this circuit would give complete cancellation and the top of the clipped waveform would be absolutely flat. Cancellation processes have something of a dubious reputation in engineering circles, because of the need for accurate matching of amplitude and phase to get a good cancellation. Where this depends on several factors – and particularly if semiconductor characteristics enter the equation – it can indeed be difficult to control. However, in this particular case the cancellation accuracy depends only on a few well-matched resistors. 1% tolerance is quite good enough, and most modern resistors are this accurate. The only consequence of mismatching is a very slight curvature of the clipped part of the waveform, and this is not normally enough to be troublesome.

As shown by the phase spikes in the diagram, this circuit gives a phase inversion, and this must be taken into account during the system design.

Noise generators

As with the introduction of deliberate clipping, it may seem perverse to go to a lot of trouble to generate noise when we have spent great efforts so far to minimise it. However, noise sources have their uses, for example in loudspeaker testing, room equalisation, and in analogue synthesisers. Noise comes, as is well known, in various colours. This is not due to synaesthesia ('Tuesdays are red!') but a convenient way of describing the spectral content of various particularly useful kinds of noise.

White noise has equal power in equal bandwidth, so there is the same power between 100 and 200 Hz as there is between 1100 and 1200 Hz. It is the type of noise produced by most electronic noise mechanisms.

Pink noise has equal power in equal ratios of bandwidth, so there is the same power between 100 and 200 Hz as there is between 200 and 400 Hz. The energy per Hz falls at 3 dB per octave as frequency increases. Pink noise is very important as it gives a flat response when viewed on a third-octave or other constant percentage bandwidth spectrum analyzer.

Red noise has its energy per Hz falling at 6 dB per octave; its main use is in synthesisers. There is more on noise of different colours in Chapter 1.

Most noise generators these days are digital, using maximal-length sequences to produce noise that is tightly defined in amplitude and spectral content. That, however, is rather outside the province of this analogue book, so instead we will take a quick look at analogue noise generation.

The most popular method is to use the white noise produced when a bipolar transistor is reverse-biased and allowed to work as a Zener diode, as shown in Figure 24.32. This does the transistor no harm, so long as the current through it is limited, and is surprisingly consistent,

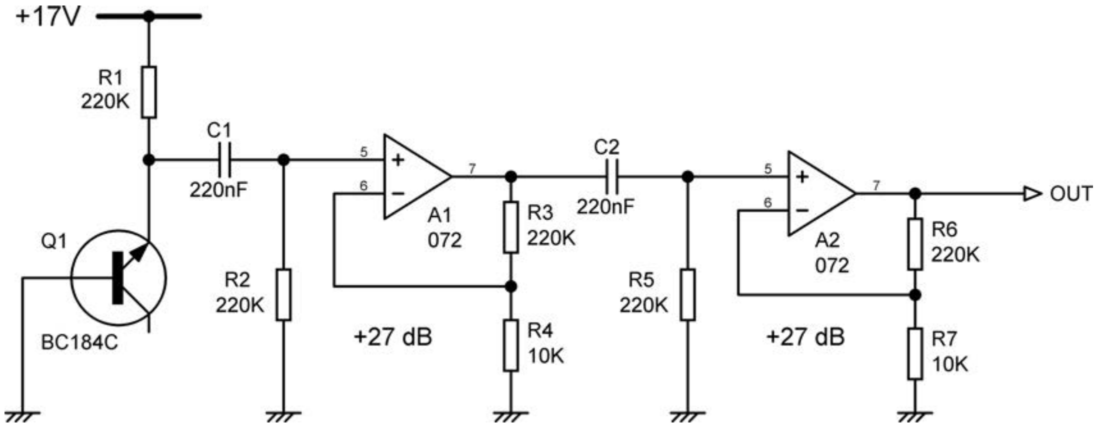

Figure 24.32: A white noise generator using a bipolar transistor as a Zener diode

when you consider that this is not exactly the official way to use a transistor. I took a collection of ten BC184C transistors of varying pedigree and provenance, and found that with an emitter current of 36 μA the noise output varied between −61.5 dBu and −65.6 dBu, with one outlier at −56.9 dBu (bandwidth 22 Hz–22 kHz). The Zener voltage was between 7.7 V and 8.9 V. When I tried a real Zener diode, of 6V2 voltage, the noise output was much less, around −85 dBu. This is presumably because Zeners are designed to produce minimal noise, though they still generate much more than simple silicon diodes; noise is distinctly unwanted in voltage reference applications.

The transistor noise output is well below a millivolt, and a good deal of amplification is needed to get it up to a useful level. Figure 24.32 shows a humble TL072 performing this service; there is of course absolutely no point in using a low noise opamp such as the 5532, and the low bias currents of the TL072 mean that high-value resistors can be used without DC offset troubles. Each stage has a gain of 27 dB. A point to remember is that the TL072 has limited open-loop bandwidth compared with more modern opamps, and if you try to take too much gain in one stage this will lead to a high-frequency roll-off. The last amplifier stage must not be allowed to clip as this will modify the energy spectrum. The output here is about 400 mVrms so clipping will statistically be very rare. Remember there was a 4 dB variation in noise output for different transistor samples, and a preset gain trim may be needed for some applications.

Pinkening filters

Converting white noise into the much more useful pink noise is proverbially difficult because you can't make a filter with a true −3 dB/octave slope, as filter slopes come in multiples of 6 dB/octave only. The only solution when you need a pinkening filter is a series of

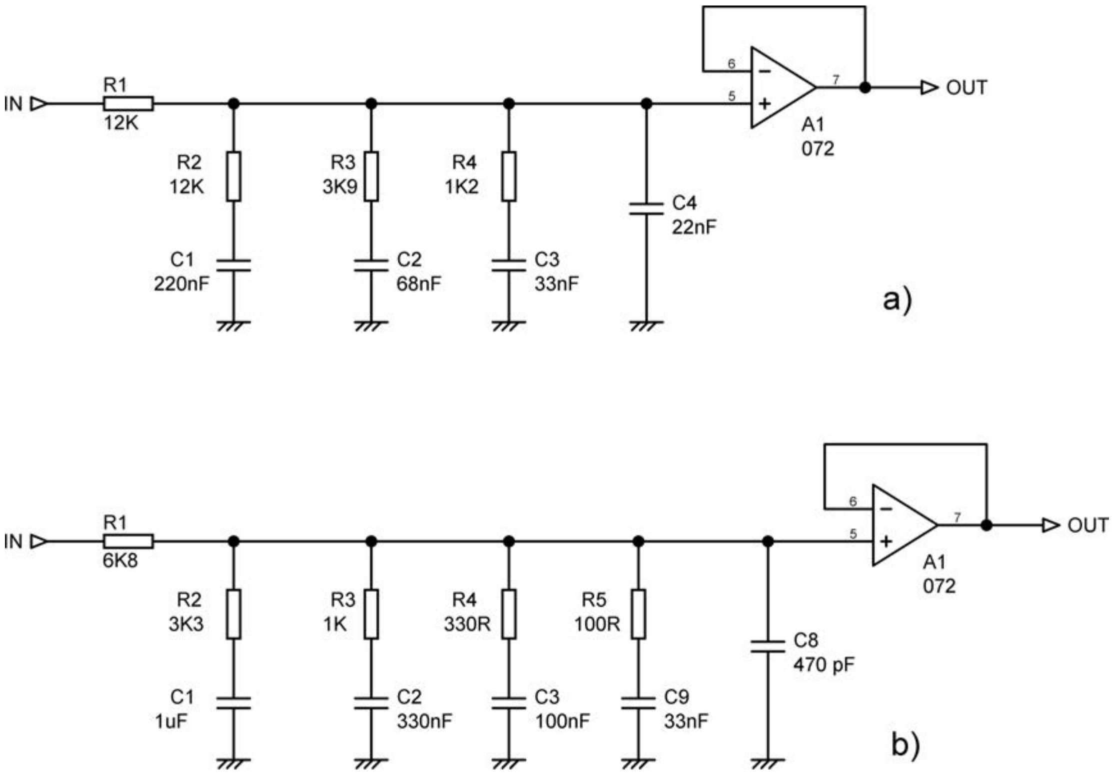

Figure 24.33: Two −3 dB/octave pinkening filters for turning white noise into pink noise

overlapping low-pass and high-pass time-constants (i.e. alternate poles and zeros) that approximate to the required slope. The more pairs of poles and zeros used, the more accurately the response approximates to a −3 dB/octave slope. Two possible versions are shown in Figure 24.33: that in a) uses three pole-zero pairs, with a final unmatched pole introduced by C4; that in b) has four pole-zero pairs, with a final unmatched pole. The closer the pole-zero spacing, the less the wobble on the frequency response.

The response of 24.33b, the more expensive and more accurate version, is shown in Figure 24.34. The measured curve is seen to be closely parallel to the −3 dB/octave line drawn below it on the graph. A still more accurate −3 dB/octave filter can be made by using seven RC networks that fit in with the E6 capacitor series; accuracy is within +0.1 dB over a 10 Hz–20 kHz frequency range. For this and more on pinkening filters, see my book *The Design of Active Crossovers* [10].

Red noise is rarely required, but if needed it can be made from white noise by passing it through a simple integrator, which has the necessary −6 dB/octave slope.

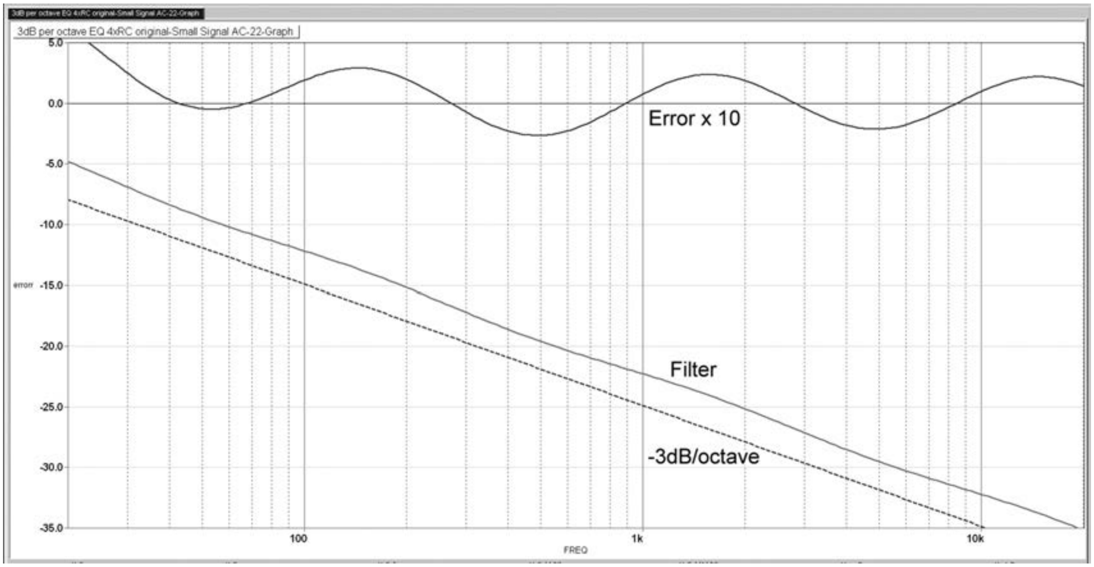

Figure 24.34: The simulated response of the 4× RC −3 dB/octave filter in Figure 24.33b. The error trace at the top (multiplied by ten) shows a maximum error of +0.28 dB over the 20 Hz–20 kHz range

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CHAPTER 25

Power supplies

We thought, because we had power, we had wisdom.

Stephen Vincent Benet: Litany for Dictatorships, 1935

Opamp supply rail voltages

It has been mentioned several times in the earlier chapters of this book that running opamps at the slightly higher voltage of ±17 V rather than ±15 V gives an increase in headroom and dynamic range of 1.1 dB for virtually no cost and with no reliability penalty. Soundcraft ran all the opamps in their mixing consoles at ±17 V for at least two decades, and opamp failures were almost unknown. This recommendation assumes that the opamps concerned have a maximum supply voltage rating of ±18 V, which is the case for the Texas TL072, the new LM4562, and many other types.

The 5532 is (as usual) in a class of its own. Both the Texas and Fairchild versions of the NE5532 have an absolute maximum power supply voltage rating of ±22 V (though Texas also gives a 'recommended supply voltage' of ±15 V), but I have never met any attempt to make use of this capability. The 5532 runs pretty warm on ±17 V when it is simply quiescent, and my view (and that of almost all the designers I have spoken to) is that running it at any higher voltage is simply asking for trouble. This is a particular concern in the design of mixing consoles, which may contain thousands of opamps – anything that impairs their reliability is going to cause a *lot* of trouble. In any case, moving from ±17 V rails to ±18 V rails only gives 0.5 dB more headroom. Stretching things to ±20 V would give 1.4 dB more than ±17 V, and running on the ragged edge at ±22 V would yield a more significant 2.2 dB more than ±17 V, but you really wouldn't want to do it. Pushing the envelope like this is also going to cause difficulties if you want to run opamps with maximum supply ratings of ±18 V from the same power supply.

We will therefore concentrate here on ±17 V supplies for opamps, dealing first with what might be called 'small power supplies' i.e. those that can be conveniently built with TO-220 regulators. This usually means an output current capability that does not exceed 1.5 amps, which is plenty for even complicated preamplifiers, electronic crossovers, etc, but will only run a rather small mixing console; the needs of large consoles are dealt with later in this chapter.

An important question is: how low does the noise and ripple on the supply output rails need to be? Opamps in general have very good power supply rejection ratios (PSRR) and some manufacturer's specs are given in Table 25.1.

The PSRR performance is actually rather more complex than the bare figures given in the table imply; PSRR is typically frequency-dependent (deteriorating as frequency rises) and different for the +V and −V supply pins. It is however rarely necessary to get involved in this degree of detail. Fortunately even the cheapest IC regulators (such as the venerable 78xx/79xx series) have low enough noise and ripple outputs that opamp PSRR performance is rarely an issue.

There is however another point to ponder; if you have a number of electrolytic-sized decoupling capacitors between rail and ground, enough noise and ripple can be coupled into the non-zero ground resistance to degrade the noise floor. Intelligent placing of the decouplers can help – putting them near where the ground and supply rails come onto the PCB means that ripple will go straight back to the power supply without flowing through the ground tracks on the rest of the PCB. This is of limited effectiveness if you have a number of PCBs connected to the same IDC cable, as in many small mixing desks, and in such cases low-ripple power supplies may be essential.

Apart from the opamp supply rails, audio electronics may require additional supplies, as shown in Table 25.2.

TABLE 25.1 PSRR specs for common opamps

Opamp type	PSRR minimum (dB)	PSRR typical (dB)
5532	80	100
LM4562	110	120
TL072	70	100

TABLE 25.2 Typical additional supply rails for opamp based systems

Supply voltage	Function
+5 V	Housekeeping microcontroller
+9 V	Relays
+24 V	LED bar-graph metering systems, discrete audio circuitry, relays
+48 V	Microphone phantom power

It is often convenient to power relays from a +9 V unregulated supply that also feeds the +5 V microcontroller regulator – see later in this chapter. The use of +24 V to power LED metering systems is dealt with in Chapter 23 on metering, and +48 V phantom supplies are examined at the end of this chapter.

Designing a ±15 V supply

Making a straightforward ±15 V 1 amp supply for an opamp-based system is very simple, and has been ever since the LM7815/7915 IC regulators were introduced (which was a long time ago). They are robust and inexpensive parts with both overcurrent and over-temperature protection, and give low enough output noise for most purposes. We will look quickly at the basic circuit because it brings out a few design points which apply equally to more complex variations on the theme. Figure 25.1 shows the schematic, with typical component values; a centre-tapped transformer, a bridge rectifier, and two reservoir capacitors C1, C2 provide the unregulated rails that feed the IC regulators. The secondary fuses must be of the slow-blow type. The small capacitors C7–C9 across the input to the bridge reduce RF emissions from the rectifier diodes; they are shown as X-cap types not because they have to withstand 230 Vrms, but to underline the need for them to be rated to withstand continuous AC stress. The capacitors C3, C4 are to ensure HF stability of the regulators, which like a low AC impedance at their input pins, but these are only required if the reservoir capacitors are not adjacent to the regulators, i.e. more than 10 cm away. C5, C6 are not required for regulator stability with the 78/79 series – they are there simply to reduce the supply output impedance at high audio frequencies.

There are really only two electrical design decisions to be made; the AC voltage of the transformer secondary and the size of the reservoir capacitors. As to the first, you must make sure that the unregulated supply is high enough to prevent the rails dropping out (i.e. letting hum through) when a low mains voltage is encountered, but not so high that either the

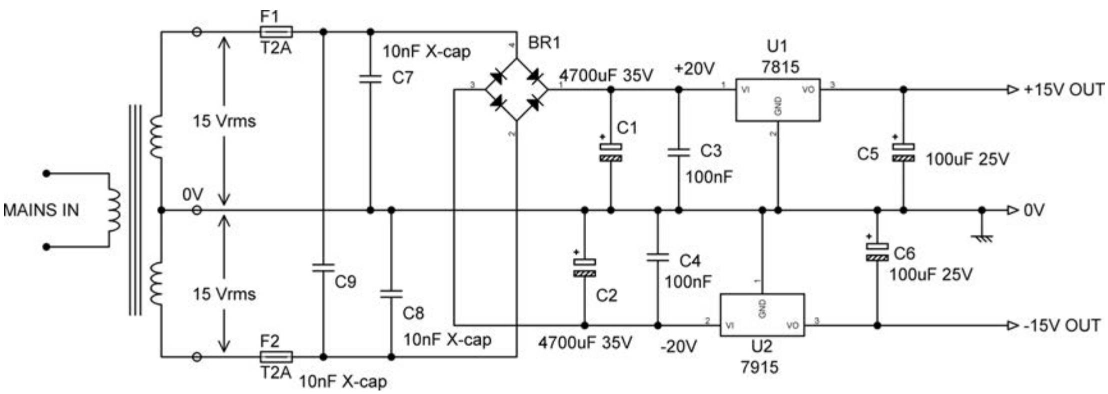


Figure 25.1: A straightforward ±15 V power supply using IC regulators

maximum input voltage of the regulator is exceeded, or it suffers excessive heat dissipation. How low a mains voltage it is prudent to cater for depends somewhat on where you think your equipment is going to be used, as some parts of the world are more subject to brown-outs than others. You must consider both the minimum voltage-drop across the regulators (typically 2 V) and the ripple amplitude on the reservoirs, as it is in the ripple troughs that the regulator will first 'drop out' and let through unpleasantness at 100 Hz.

In general, the RMS value of the transformer secondary will be roughly equal to the DC output voltage.

The size of reservoir capacitor required depends on the amount of current that will be drawn from the supply. The peak-to-peak ripple amplitude is normally in the region of 1 to 2 volts; more ripple than this reduces efficiency as the unregulated voltage has to be increased to allow for unduly low ripple troughs, and less ripple is usually unnecessary and gives excessive reservoir capacitor size and cost. The amount of ripple can be estimated with adequate accuracy by using Equation 25.1

$$V_{\text{pk-pk}} = \frac{I \cdot \Delta t \cdot 1000}{C} \qquad \text{(Equation 25.1)}$$

where $V_{\text{pk-pk}}$ is the peak-to-peak ripple voltage on the reservoir capacitor, I is the maximum current drawn from that supply rail in amps, Δt is the length of the capacitor discharge time, taken as 7 milliseconds, C is the size of the reservoir capacitor in microFarads. The '1000' factor simply gets the decimal point in the right place.

Note that the discharge time is strictly a rough estimate, and assumes that the reservoir is being charged via the bridge for 3 msec, and then discharged by the load for 7 msec. Rough estimate it may be, but I have always found it works very well.

The regulators must be given adequate heatsinking. The maximum voltage drop across each regulator (assuming 10% high mains) is multiplied by the maximum output current to get the regulator dissipation in watts, and a heat sink selected with a suitable thermal resistance to ambient (in °C per watt) to ensure that the regulator package temperature does not exceed, say, 90 °C. Remember to include the temperature drop across the thermal washer between regulator and heatsink.

Under some circumstances it is wise to add protective diodes to the regulator circuitry, as shown in Figure 25.2. The diodes D1, D3 across the regulators are reverse-biased in normal operation, but if the power supply is driving a load with a large amount of capacitance, it is possible for the output to remain higher in voltage than the regulator input as the reservoir voltage decays. D1, D3 prevent this effect from putting a reverse voltage across the regulators. Such diodes are not usually required with normal opamp circuitry, as the amount of rail decoupling, shown as C7, C8 in Figure 25.2, is usually modest.

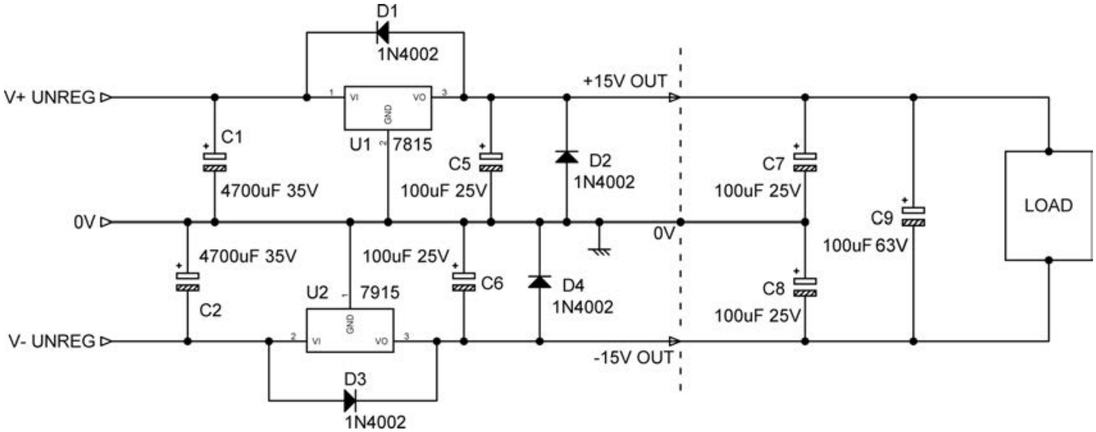
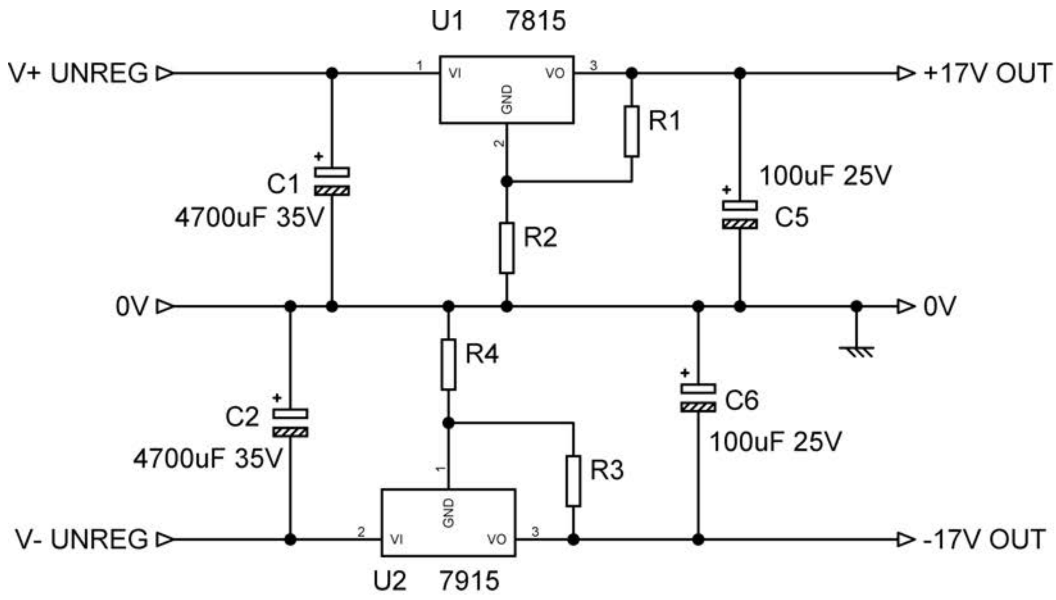


Figure 25.2: Adding protection diodes to a ±15 V power supply. The load has decoupling capacitors to both ground (C7, C8) and between the rails (C9); the latter can cause start-up problems

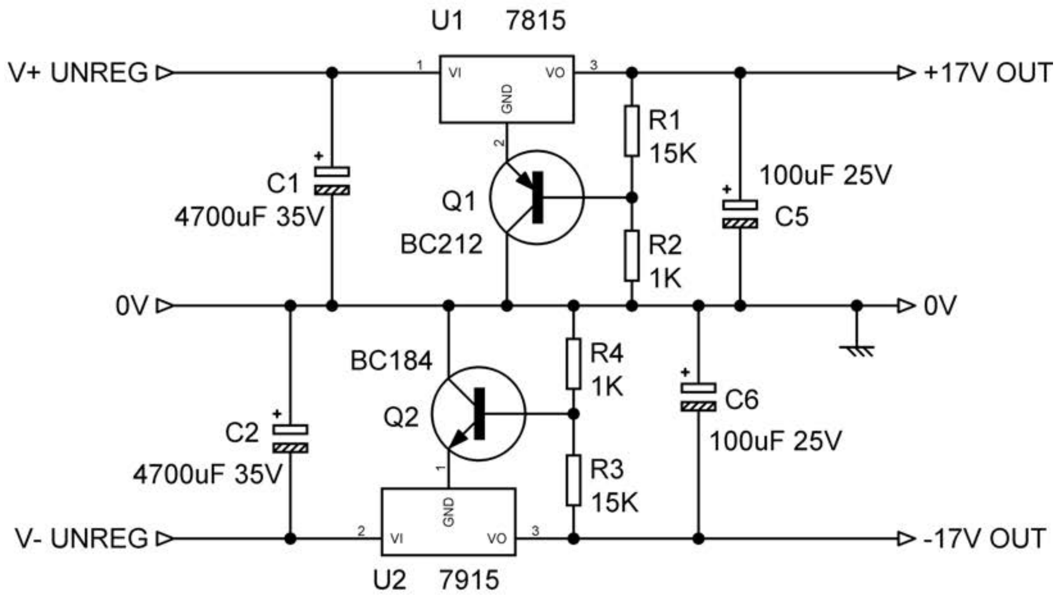
The shunt protection diodes D2, D4 are once again reverse-biased in normal operation. D2 prevents the +15 V supply rail from being dragged below 0 V if the −15 V rail starts up slightly faster, and likewise D4 protects the −15 V regulator from having its output pulled above 0 V. This can be an important issue if rail-to-rail decoupling such as C9 is in use; such decoupling can be useful because it establishes a low AC impedance across the supply rails without coupling supply rail noise into the ground, as C7, C8 are prone to do. However, it also makes a low-impedance connection between the two regulators. D2, D4 will prevent damage in this case, but leave the power supply vulnerable to start-up problems; if its output is being pulled down by the −15 V regulator, the +15 V regulator may refuse to start. This is actually a very dangerous situation, because it is quite easy to come up with a circuit where start-up will only fail one time in 20 or more, the incidence being apparently completely random, but presumably controlled by the exact point in the AC mains cycle where the supply is switched on and other variables such as temperature, the residual charge left on the reservoir capacitors, and the phase of the moon. If even one start-up failure event is overlooked or dismissed as unimportant, then there is likely to be serious grief further down the line. *Every power supply start-up failure must be taken seriously.*

Designing a ±17 V supply

There are 15 V IC regulators (7815, 7915) and there are 18 V IC regulators, (7818, 7918) but there are no 17 V IC regulators. This problem can be effectively solved by using 15 V regulators and adding 2 volts to their output by manipulating the voltage at the REF pin. The simplest way to do this is with a pair of resistors that divide down the regulated output voltage and apply it to the REF pin as shown in Figure 25.3a (the transformer and AC input



a



b

Figure 25.3: Making a ±17 V power supply with 15 V IC regulators: a) using resistors is inefficient and/or inaccurate, b) adding transistors to the voltage — determining resistor network makes the output voltage more predictable and reduces the power consumed in the resistors

components have been omitted in this and the following diagrams, except where they differ from those shown above). Since the regulator maintains 15 V between the OUT and REF pin, with suitable resistor values the actual output with respect to 0 V is 17 V.

The snag with this arrangement is that the quiescent current that flows out of the REF pin to ground is not well controlled; it can vary between 5 and 8 mA, depending on both the input voltage and the device temperature. This means that R1 and R2 have to be fairly low in value so that this variable current does not cause excessive variation of the output voltage, and therefore power is wasted.

If a transistor is added to the circuit as in Figure 25.3b, then the impedance seen by the REF pin is much lower. This means that the values of R1 and R2 can be increased by an order of magnitude, reducing the waste of regulator output current and reducing the heat liberated. This sort of manoeuvre is also very useful if you find that you have a hundred thousand 15 V regulators in store, but what you actually need for the next project is an 18 V regulator, of which you have none.

What about the output ripple with this approach? I have just measured a power supply using the exact circuit of Figure 25.3b, with 2200 μF reservoirs, and I found −79 dBu (87 μVrms) on the +17 V output rail, and −74 dBu (155 μVrms) on the 17 V rail, which is satisfyingly low for inexpensive regulators, and should be adequate for almost all purposes; note that these figures include regulator noise as well as ripple. The load current was 110 mA. If you *are* plagued by ripple troubles, the usual reason is a rail decoupling capacitor that is belying its name by coupling rail ripple into a sensitive part of the ground system and the cure is to correct the grounding rather than design an expensive ultra-low ripple PSU. Note that doubling the reservoir capacitance to 4400 μF only improved the figures to −80 dBu and −76 dBu respectively; just increasing reservoir size is not a cost-effective way to reduce the output ripple.

Using variable-voltage regulators

It is of course also possible to make a ±17 V supply by using variable output voltage IC regulators such as the LM317/337. These maintain a small voltage (usually 1.2 V) between the OUTPUT and ADJ (shown in figures as GND) pins, and are used with a resistor divider to set the output voltage. The quiescent current flowing out of the ADJ pin is a couple of orders of magnitude lower than for the 78/79 series, at around 55 μA, and so a simple resistor divider gives adequate accuracy of the output voltage, and transistors are no longer needed to absorb the quiescent current. A disadvantage is that this more sophisticated kind of regulator is somewhat more expensive than the 78/79 series; at the time of writing they cost something like 50% more. The 78/79 series with transistor voltage-setting remains the most cost-effective way to make a non-standard-voltage power supply at the time of writing.

It is clear from Figure 25.4 that the 1.2 V reference voltage between ADJ and out is amplified by many times in the process of making a 17 V or 18 V supply; this not only increases output ripple, but also output noise as the noise from the internal reference is being amplified. The noise and ripple can be considerably reduced by putting a capacitor C7 between the ADJ pin and ground. This makes a dramatic difference; in a test PSU with a 650 mA load the output noise and ripple was reduced from −63 dBu (worse than 78xx series) to −86 dBu (better than 78xx series) and so such a capacitor is usually fitted as standard. If it is fitted, it is then essential to add a protective diode D1 to discharge C7, C8 safely if the output is short-circuited, as shown in Figure 25.5.

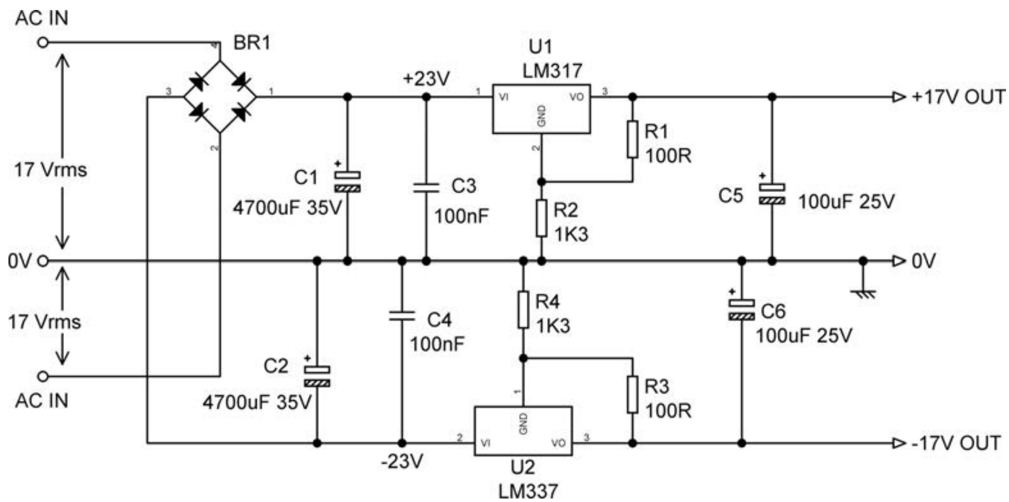


Figure 25.4: Making a ±17 V power supply with variable-voltage IC regulators

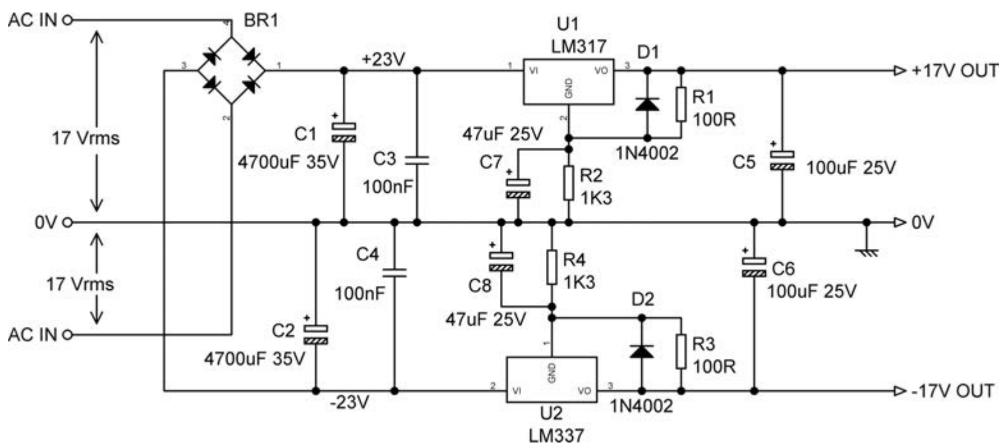


Figure 25.5: Ripple improvement and protective diodes for a variable-voltage IC regulator

TABLE 25.3 Comparing the noise and ripple output of various regulator options

	7815 + transistor (dBu)	LM317 (dBu)	LM317 (μV)
No C on LM317 ADJ pin	−73 (all ripple)	−63 (ripple and noise)	549
47 μF on LM317 ADJ pin	−73 (all ripple)	−86 (ripple and noise)	39
Input filter 2.2 Ω and 2200 μF	−78 (ripple and noise)	−89 (mostly noise)	27
Input filter 2.2 Ω and 4400 μF	−79 (mostly noise)	−90 (all noise)	24

The ripple performance of the aforementioned test PSU, with a 6800 μF reservoir capacitor and a 650 mA load, is summarised for both types of regulator in Table 25.3. Note that the exact ripple figures are subject to some variation between regulator specimens.

Improving ripple performance

Table 25.3 shows that the best noise and ripple performance that can be expected from a simple LM317 regulator circuit is about −86 dBu (39 μVrms) and this still contains a substantial ripple component. The reservoir capacitors are already quite large at 4700 μF, so what is to be done if lower ripple levels are needed? The options are:

1. Look for a higher-performance IC regulator. They will cost more and there are likely to be issues with single sourcing.
2. Design your own high-performance regulator using discrete transistors or opamps. This is not a straightforward business if all the protection that IC regulators have is to be included. There can also be distressing issues with HF stability.
3. Add an RC input filter between the reservoir capacitor and the regulator. This is simple and pretty much bullet-proof, and preserves all the protection features of the IC regulator, though the extra components are a bit bulky and not that cheap. There is some loss of efficiency due to the voltage drop across the series resistor; this has to be kept low and the capacitance large.

The lower two rows of Table 25.3 show what happens. In the first case the filter values were 2.2 Ω and 2200 μF. This has a −3 dB frequency of 33 Hz and attenuates the 100 Hz ripple component by 10 dB. This has a fairly dramatic effect on the output ripple, but the dB figures do not change that much as the input filter does not affect the noise generated inside the regulator. Increasing the filter capacitance to 4400 μF sinks the ripple below the noise level for both types of regulator.

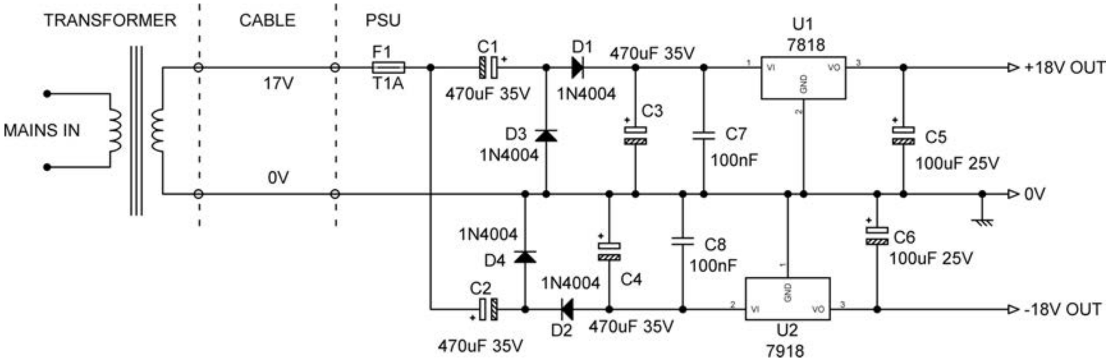


Figure 25.6: A ±18 V power supply powered by a single transformer winding

Dual supplies from a single winding

It is extremely convenient to use third-party 'wall-wart' power supplies for small pieces of equipment, as they come with all the safety and EMC approvals already done for you, though admittedly they do not look appropriate with high-end equipment.

The problem is that the vast majority of these supplies give a single AC voltage on a two-pole connector, so a little thought is required to derive two supply rails. Figure 25.6 shows how it is done in a ±18 V power supply; note that these voltages are suitable only for a system that uses 5532s throughout. Two voltage-doublers of opposite polarity are used to generate the two unregulated voltages. When the incoming voltage goes negative, D3 conducts and the positive end of C1 takes up approximately 0 V. When the incoming voltage swings positive, D1 conducts instead and the charge on C1 is transferred to C3. Thus the whole peak-to-peak voltage of the AC supply appears across reservoir capacitor C3. In the same way, the peak-to-peak voltage, but with the opposite polarity, appears across reservoir C4.

Since voltage-doublers use half-wave rectification, they are not suitable for high current supplies. When choosing the value of the reservoir capacitor values, bear in mind that the discharge time in Equation 25.1 above must be changed from 7 msec to 17 msec. The input capacitors C1, C2 should be the same size as the reservoirs.

Power supplies for discrete circuitry

One of the main reasons for using discrete audio electronics is the possibility of handling larger signals than can be coped with by opamps running off ±17 V rails. The use of ±24 V rails allows a 3 dB increase in headroom, which is probably about the minimum that justifies the extra complications of discrete circuitry. A ±24 V supply can be easily implemented with 7824/7924 IC regulators.

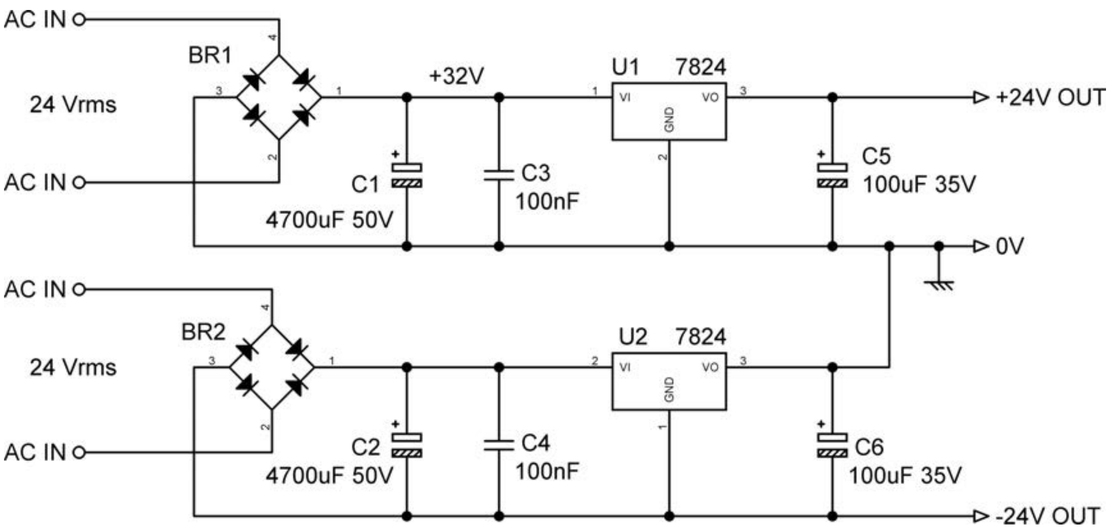


Figure 25.7: A ±24 V power supply using only positive regulators

A slightly different approach was used in my first published preamplifier design [1]. This preamp in fact used two LM7824 + 24 V regulators connected as shown in Figure 25.7 because at the time the LM7924 −24 V regulator had not yet reached the market. The use of a second positive regulator to produce the negative output rail looks a little strange at first sight but I can promise you it works. It can be very useful in the sort of situation described above; you have a hundred thousand +15 V regulators in store, but no −15 V regulators . . . I'm sure you see the point.

Note that this configuration requires two separate transformer windings; it cannot be used with a centre-tapped secondary.

Larger power supplies

So what if you need more than 1 amp of current? This will certainly be the case for all but the smallest mixing consoles. There are of course IC regulators with a greater current capability than 1 Amp. The LM338K variable-voltage 5 amp regulators come in a TO-3 package – which it has to be said is not the easiest format to mount. They were once widely used in mixing console power supplies but have been superseded by more modern devices such as the Linear Tech LT1083CK, which is a 7.5 amp variable-voltage regulator that works very nicely, and can be obtained in a TO-3P package. I have used lots of them in mixing console power supplies. An example of their use, incorporating a mutual-shutdown facility, is shown in Figure 25.8.

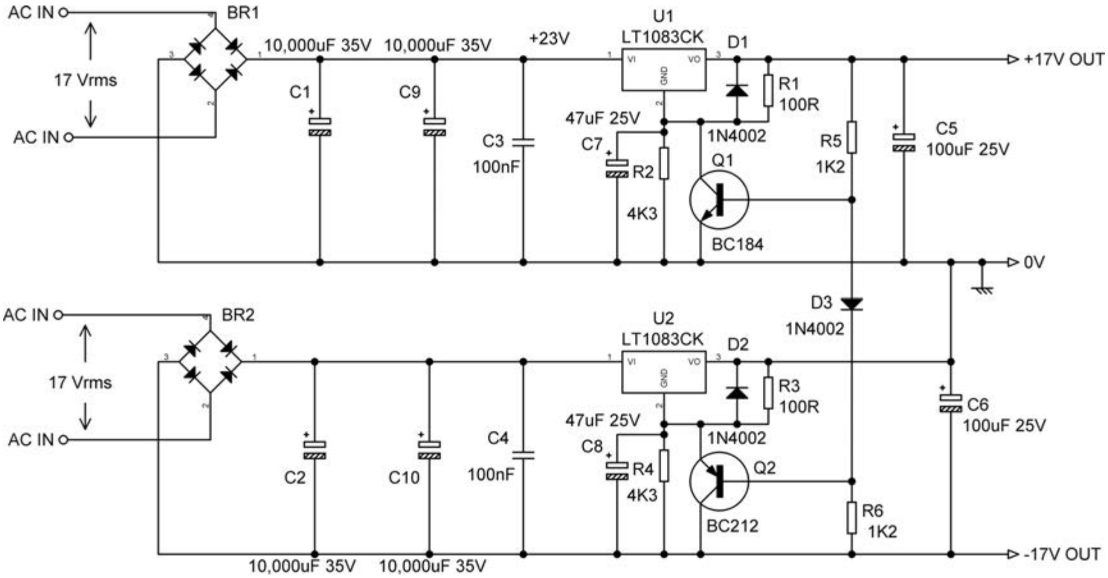


Figure 25.8: A high-current ±17 V power supply with mutual shutdown circuitry

There is no negative version of the LT1083CK, so a ± supply has to be made with two separate secondary windings, just as in Figure 25.7. When using any high-current regulator it pays to be rather cautious about the maximum rated current. Trying to use them too close to the maximum can cause start-up problems.

Mutual shutdown circuitry

It is an awkward quirk of 5532 opamps that if one supply rail is lost and collapses to 0 V, while the other rail remains at the normal voltage, they can under some circumstances get into an anomalous mode of operation that draws large supply currents and ultimately destroys the opamp by over-heating. To prevent damage from this cause, which could be devastating to a large mixing console, the opamp supplies are very often fitted with a mutual shutdown system. Mutual shutdown ensures that if one supply rail collapses, because of overcurrent, over-temperature or any other cause, the other rail will be promptly switched off. The extra circuitry required to implement this is shown in Figure 25.8, which is an example of a high-current supply using 7.5 amp regulators.

The extra circuitry to implement mutual-shutdown in Figure 25.8 is very simple; R5, D3, R6 and Q1 and Q2. Because R5 is equal to R6, D3 normally sits at around 0 V in normal operation. If the +17 V rail collapses, Q2 is turned on by R6, and the REF pin of U2 is pulled down to the bottom rail, reducing the output to the reference voltage (1.25 V). This is not completely off, but it is low enough to prevent any damage to opamps.

If the −17 V rail collapses, Q1 is turned on by R5, pulling down the REF pin of U1 in the same way. Q1 and Q2 do not operate exactly symmetrically, but it is close enough for our purposes.

Note that this circuit can only be used with variable output voltage regulators, because it relies on their low reference voltages.

Very large power supplies

By 'very large' I mean too big to be implemented with IC regulators – say 7 amps and above. This presents a difficult problem, to which there are several possible answers:

1. Split up the system supply rails so that several IC regulators can be used. This is in my view the best approach. The amount of design work is relatively small; in particular the short-circuit protection has been done for you.
2. Use power transistors as series-pass elements, controlled by opamps. This can be a surprisingly tricky technology. The feedback system has to be reliably stable, and the short-circuit protection has to be foolproof. Designing the latter is not too simple.
3. Switch-mode power supplies. Those found in PCs seem to be a mature technology, and are very reliable. Custom designs, of the sort required for this application are, in my experience, another matter. I have seen them explode. There is also the issue of RF emissions.

The first two methods obviously involve increased heat dissipation in proportion to their output current. This usually means that fan cooling is required to keep the heatsinks down to a reasonable size, which is fine for PA work but not welcome in a studio control room. There are no great technical difficulties to powering even a large console over a 20-metre cable (this is the recommended maximum for Neve consoles) so long as remote sensing is used to compensate for voltage drops. 20 metres is usually long enough to allow the power supply to be placed in another room.

The technology of very large audio supplies is a specialised and complicated business, and it would not be appropriate to dig any further into it here.

Microcontroller and relay supplies

It is very often most economical to power relays from an unregulated supply. This is perfectly practical as relays have a wide operating voltage range. If 9 V relays are used then the same unregulated supply can feed a +5 V regulator to power a microcontroller, as shown in Figure 25.9.

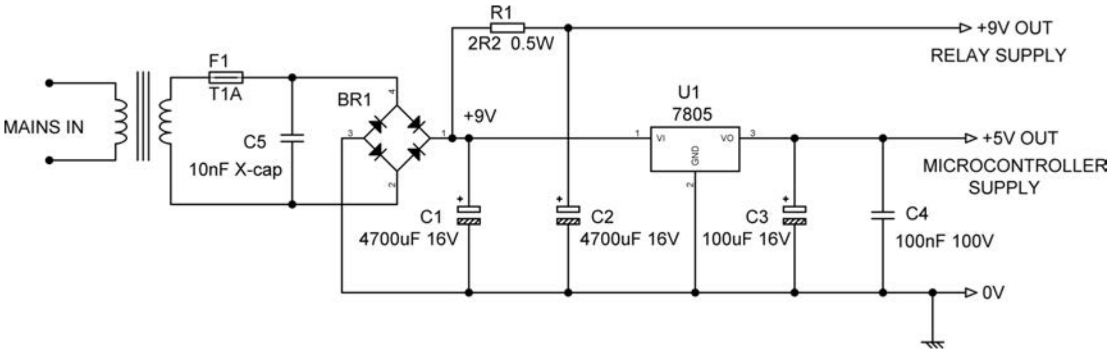


Figure 25.9: A +5 V PSU with an RC smoothed +9 V relay supply

Hum induced by electrostatic coupling from an unregulated relay supply rail can be sufficient to compromise the noise floor; the likelihood of this depends on the physical layout, but inevitably the signal paths and the relay supply come into proximity at the relay itself. It is therefore necessary to give this rail some degree of smoothing, without going to the expense of another regulator and heatsink (there must be no possibility of coupling between signal ground and relay power ground; these must only join right back at the power supply). This method of powering relays is more efficient than a regulated rail as it does not require a voltage drop across a regulator that must be sufficient to prevent drop-out and consequent rail ripple at low mains voltages.

Simple RC smoothing works perfectly well for this purpose. Relays draw relatively high currents, so a low R and a high value C are used to minimise voltage losses in R and changes in the relay supply voltage as different numbers of relays are energised.

The RC smoothing values shown in Figure 25.9 are typical, but are likely to need adjustment depending on how many relays are powered and how much current they draw. R1 is low at 2.2 Ω and C2 high at 4700 µF; fortunately the voltage is low so C2 need not be physically large.

+48 V phantom power supplies

Making a discrete +48 V regulator with the necessary low amounts of noise and ripple is not too hard, but making one that is reliably short-circuit proof is a little more of a challenge, and by far the easiest way to design a +48 V phantom supply is to use a special high-voltage regulator called the TL783C, as shown in Figure 25.10. This extremely handy device can supply 700 mA, subject to power dissipation constraints.

It is a variable-voltage device, maintaining typically 1.27 V between the OUT and ADJ pins. It combines BJT circuitry with high-voltage MOS devices on the same chip,

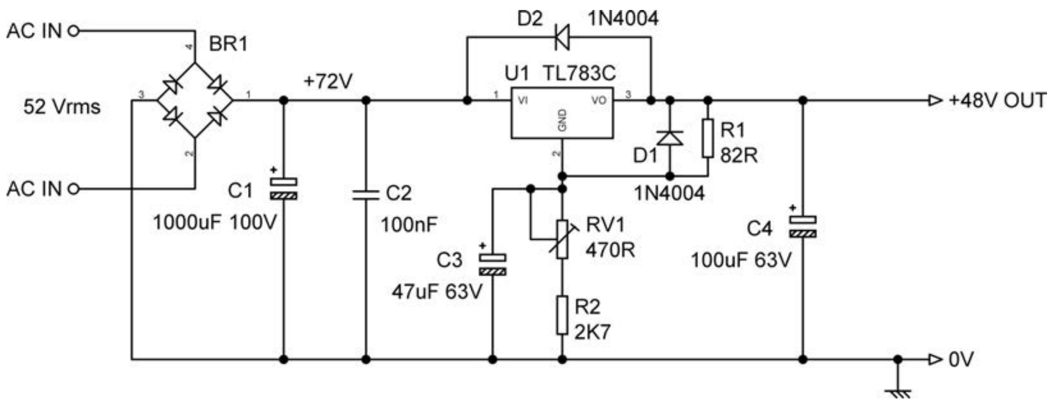


Figure 25.10: A +48 V phantom power supply using the TL783C regulator. Because of the large reference multiplication factor, a preset is required to set the output voltage exactly

allowing it to withstand much higher voltages than standard bipolar regulators. Since MOS devices are not subject to secondary-breakdown or thermal-runaway, the TL783 still gives full overload protection while operating with up to a 125 V voltage drop from the input to the output. The TL783 has current limiting, safe-operating-area (SOA) protection, and thermal shutdown. Even if the ADJ pin is accidentally disconnected, the protection circuitry stays operational. It is a very useful and reliable IC, and I have deployed thousands of them.

As with other variable-voltage regulators, the low voltage maintained between the OUT and ADJ pins needs to be amplified by a considerable ratio to get the desired output voltage, and so the reference voltage tolerances are also amplified. In this case the amplification factor is as high as 37 times, and so a preset is used to adjust the output voltage to exactly +48 V. The filter capacitor C3 is essential for the same reason – without it the ripple is amplified along with the reference voltage.

The unregulated supply can be derived from a completely separate transformer secondary as in Figure 25.10, or alternatively by means of a voltage doubler. The latter is usually more economic, but obviously this depends on the cost of an extra transformer winding versus the cost of the extra capacitor in the doubler.

The arrangement of a voltage-doubler phantom supply is shown in Figure 25.11. Note that the familiar voltage doubler circuit C13, C2, D5, D6 is actually working as a voltage-tripler, because it is perched on the unregulated +23 V supply to the +17 V regulator. If it worked as a true voltage-doubler, based on the 0 V rail, it would generate insufficient unregulated voltage for the phantom regulator. Because of their inherently half-wave operation and relatively poor regulation, voltage-doubler or voltage-tripler methods are not suitable for high-current phantom supplies.

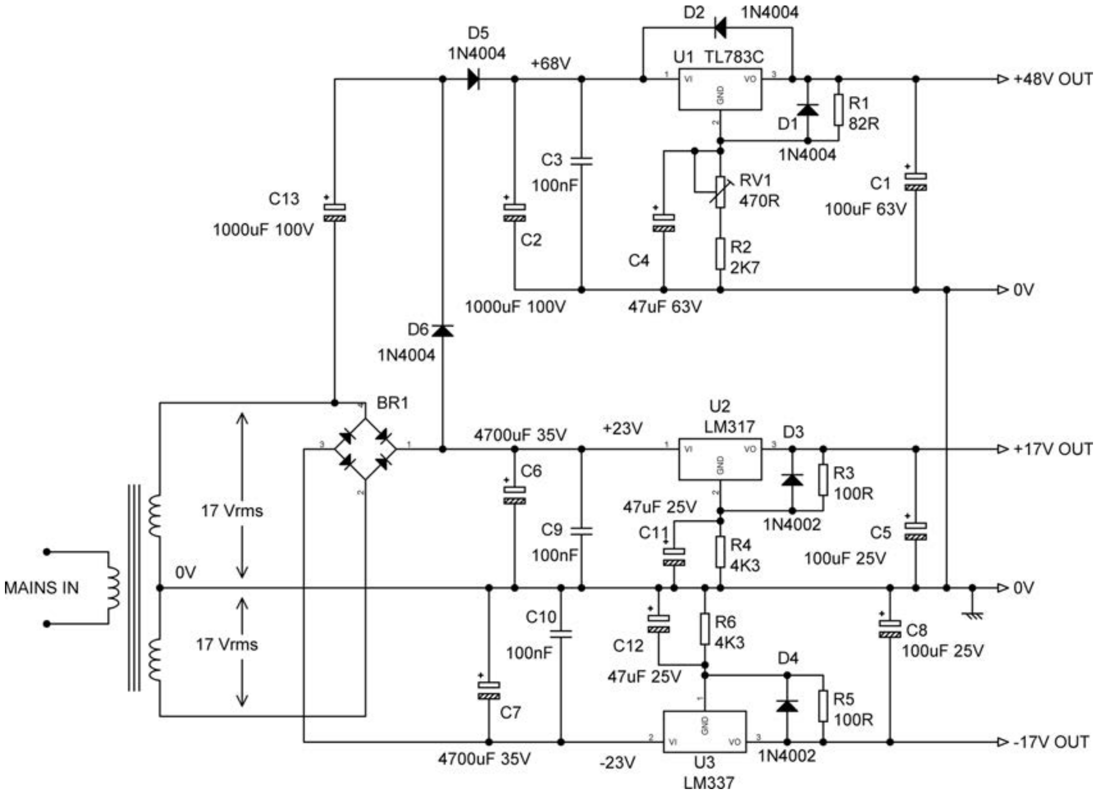


Figure 25.11: +48 V phantom power supply fed by a voltage-doubler

Reference

[1] Self, D. R. G. 'An Advanced Preamplifier Design', *Wireless World* (November 1976).

CHAPTER 26

Interfacing with the digital domain

Introduction

The advance of digital audio has greatly improved the fidelity of audio storage media, and generally made wonderful things possible, but sound waves remain stubbornly analogue, and so conversion from analogue to digital and vice-versa is very necessary. Today's analogue to digital converters (ADCs) and digital to analogue converters (DACs) have excellent performance, with 24-bit accuracy at a 192 kHz sampling rate a commonplace, but to achieve this potential performance in an application there are a good number of factors that need to be appreciated. Some of them, such as the need for effective HF decoupling, are relatively straightforward providing you follow the manufacturer's recommendations; but others, involving the actual interfacing to the analogue input and output pins, are a bit more subtle.

Having said that, contemporary ADCs and DACs are far easier to apply than their ancestors. Oversampling technology means that it is no longer necessary to put a ninth-order brickwall low-pass anti-aliasing filter in front of an ADC, or place a ninth-order brickwall low-pass reconstruction filter after a DAC. If you've ever tried to design a ninth-order filter to a price, you will know that this is a very significant freedom. It has had a major effect in reducing the price of digital equipment, particularly in applications like digital mixers where a large number of ADCs and DACs are required.

ADC and DAC technology moves rapidly; the device examples I have chosen here will probably soon be out of date. The general principles I give should be more enduring, and will be valid for the foreseeable development of the technologies.

PCB layout considerations

The PCB layout for both ADCs and DACs requires observance of certain precautions, which are basically the same for both functions. A double-sided PCB is necessary not only because of the large number of connections that have to be made in a small space, but also because it allows tracks that need to be isolated from each other to be put on opposite sides of the board. A most important consideration is to keep digital signals, particularly fast ones such as clocks, out of the analogue inputs, so use as much physical spacing between these as possible. Critical tracks on opposite sides of the board should be run at right angles to each other to minimise coupling through the PCB. Do not run digital tracks topside under the IC as they may couple noise directly into the die from underneath.

Separate analogue and digital ground planes should be used. Most conversion ICs have their analogue and digital interfaces at opposite ends or opposite sides of the package, facilitating the use of separate ground planes. It is usually best to run the analogue ground plane under the IC to minimise the coupling of digital noise. The two ground planes must of course be connected together at some point, and this should be implemented by a single junction close to the IC. Some manufacturers (e.g. Analog Devices) recommend that the junction should be made through a ferrite bead to filter out high RF frequencies. A maximum-copper (minimum-etch) PCB layout technique is generally the best for ground planes as it gives the most screening possible.

The power supply tracks to the IC should be as wide as possible to give low impedance paths and reduce the voltage effects of current glitches on the power supply lines. Ideally a four-layer PCB (and such boards are now cheaper than they have ever been) should be used so that two layers can be devoted to power supply planes.

Thorough decoupling is always important when using high speed devices such as ADCs and DACs. All analogue and digital supplies should be decoupled to analogue ground and digital ground respectively, using 0.1 μF ceramic capacitors in parallel with 10 μF electrolytic capacitors. Some manufacturers recommend using tantalum capacitors for this. To achieve the best possible decoupling, the capacitors should be placed as physically close to the IC as possible and solidly connected to the relevant ground plane.

When you are designing ADCs or DACs into a system, my experience is that significant time can be saved by doing preliminary testing on manufacturer's evaluation boards; this has particular force when you are using parts from a range you have not used before. Higher authority may urge you to go straight to a PCB layout, but unless you are very sure what you are doing – if for example, you are cut-and-pasting from an existing satisfactory design – it is a relatively high-risk approach. Evaluation boards are usually expensive, as they are produced in small quantities, but in my view it is money very well spent.

Nominal levels and ADCs

The best use of the dynamic range of an ADC is only possible if it is presented with a signal of roughly the right amplitude. Too low a level degrades the signal-to-noise ratio as the top bits are not used, and too high a level will not only cause unpleasant-sounding digital clipping, but can cause damage to the ADC if current flows are not limited. Analogue circuitry is therefore needed to scale the signal to the right amplitude.

A typical application of ADCs is in digital mixing consoles. These must accept both microphone and line input levels. Since the signal level from a microphone may be very low (lute music) or very high (microphone in the kick-drum), an input amplifier with a wide variable gain range is required, typically 70 dB and sometimes as much as 80 dB. The signal level range of line signals is less but still requires a gain range of some 30 dB to cope with all conditions. It is therefore necessary for the operator to adjust the input gain, by reference to a level meter, so that good use is made of the available dynamic range without risking clipping. In live situations with unpredictable levels this is always something of a judgement call.

The signal level required at the ADC input to give maximum output, which is usually referred to as full-scale (FS) varies from manufacturer to manufacturer; this important point is brought out in the next section.

Some typical ADCs

There are a large number of ADCs on the market, and it is necessary to pick out just a few to look at. You will note that the various parts are actually very similar in their application. The inclusion of a device here does not mean that I am giving it any personal recommendation. All the devices mentioned are capable of 24-bit 192 kHz operation. In some cases the input voltage required for FS appears to exceed the supply voltage; this is not so, the quoted peak-to-peak voltage is the difference between two differential input pins. And now in alphabetical order:

The Analog Devices AD1871 is a stereo audio ADC with two 24-bit conversion channels each giving 105 dB of dynamic range, and each having a programmable gain amplifier (PGA) at the front end, a multibit sigma-delta modulator, and decimation filters. The digital details are rather outside our scope here and will not be alluded to further. The PGA has five gain settings ranging from 0 dB to 12 dB in 3 dB steps. The differential input required for full-scale is 2.828 $V_{pk\text{-}pk}$ and the input impedance is 8 KΩ. Like most of its kind, the AD1871 runs its analogue section from +5 V, but the digital section from +3.3 V to save power. This IC is unusual in that it is permissible to run the digital section from +5 V, which can save you a regulator.

The Analog Devices AD1974 is a quad ADC with four differential analogue inputs having a very useful CMRR of 55 dB (typical, at both 1 kHz and 20 kHz). These inputs are not buffered and require special interfacing that will be described later. A differential input of 5.4 $V_{pk\text{-}pk}$ is needed for FS; the input impedance is 8 KΩ. This IC runs from +3.3 V only.

The Burr-Brown PCM1802 is a stereo ADC with single-ended analogue voltage inputs with input buffer amplifiers. It requires 3.0 $V_{pk\text{-}pk}$ to reach full-scale and has a resistive input impedance of 20 KΩ. The analogue section is powered from +5 V, the digital section from +3.3 V.

The Wolfson WM8782 is a stereo ADC with two single-ended analogue inputs with buffer amplifiers. It requires 2.82 $V_{pk\text{-}pk}$ (1.0 Vrms) to reach FS and the input impedance is 10 KΩ. The analogue section is powered from +5 V and the digital core from +3.3 V.

Interfacing with ADC inputs

The issues involved in interfacing with an ADC depend very much on how the ADC input is configured. As we saw in the previous section, some ADCs, such as the Burr-Brown PCM1802 and the Wolfson WM8782 have internal buffer amplifiers that present a relatively high impedance to the outside world (in this case 20 KΩ and 10 KΩ respectively). These inputs are very straightforward to drive. Such buffers are usually only found in ADCs made in a bipolar or BiCMOS process as making good low-noise low-distortion amplifiers in a straight CMOS technology is very difficult.

Others, such as the AD1974, do not have buffering and must be driven from special circuitry. In the case of the AD1974 and similar devices, the differential inputs must be driven from a differential signal source to get the best performance. The basic principle is shown in Figure 26.1. The input pins connect to switched internal capacitors, and these generate glitches. Each input pin must be isolated from the opamp driving it by an external series resistor R5, R6 together with a capacitor C5, C6 connected from input to ground. This capacitor must not generate non-linearity when the voltage across it changes, so ceramic NPO or polypropylene film types must be used; recommended values for the resistors and capacitors are usually given in the application notes. Note that since the external opamps are referenced to ground, and the ADC internals are referenced to half the +5 V rail, blocking capacitors C2, C3 are needed.

Adding external resistance will slow down the charging of input sampling capacitors. These must be allowed to charge for many time-constants if they are to get close enough to the final value to avoid degrading the performance. External resistance increases the time-constant and can degrade accuracy; manufacturers usually provide guidance as to how much external resistance is permissible for a given number of bits of accuracy.

A point that is obvious but easily overlooked is that the inputs must not be driven to excessive levels. This usually means that the input voltage should not go outside the supply rails by more than 300 mV; for example Wolfson specify this restriction for both the analogue and digital inputs of the WM8782 stereo ADC, and the other manufacturers quote similar ratings.

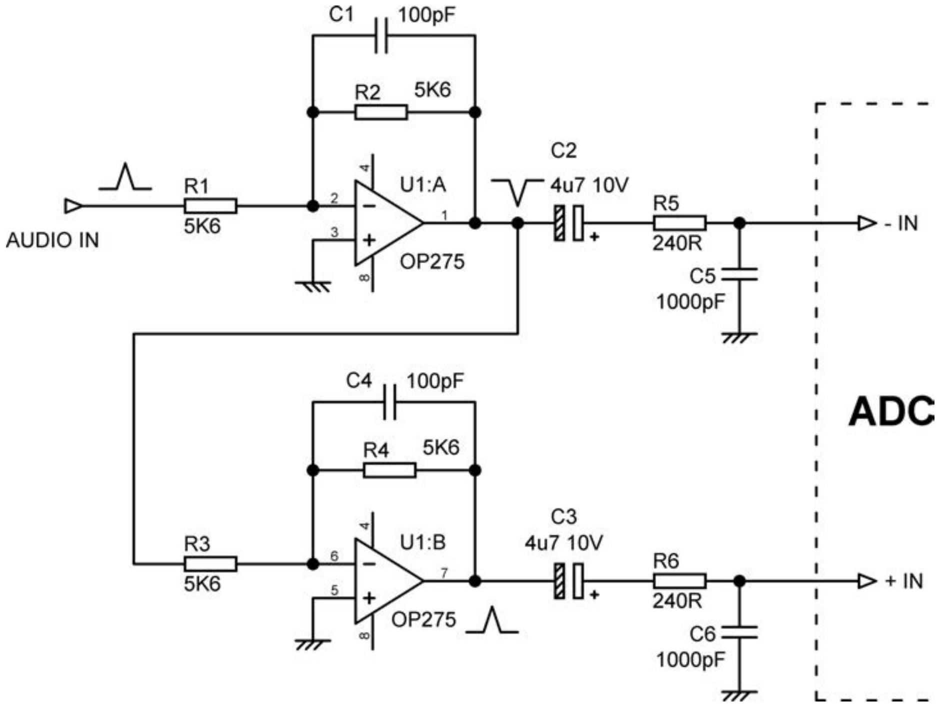


Figure 26.1: A typical drive circuit for an unbuffered differential ADC input

While ADC inputs invariably have clamp diodes for ESD protection that are intended to prevent the inputs moving outside the supply rails, these are small-dimensioned devices that may be destroyed by the output current capability of an opamp. This is why the input voltage should not go outside the supply rails by more than 300 mV – this voltage will not cause a silicon diode to conduct significantly, even at elevated temperatures. The diodes can usually handle 5 mA, but to subject them to anything more is to live dangerously. Obviously the manufacturer's absolute maximum ratings should be followed on this point, but not all manufacturers give a current rating for their clamp diodes.

Bullet-proof protection against input over-voltages is given by running the driving opamp from the same supply rails as the analogue section of the ADC, the opamp saturation voltages ensuring that the input can never reach the supply rails, never mind exceed them. This does, however, restrict your choice of opamp to one that is happy working on low supply voltages; these are likely to be more expensive than the popular audio opamps such as the 5534/5532, which will not give good performance from such low rails.

If you want to stick with the usual audio opamps, working from higher supply rails than the ADC, then an effective means of protection is the use of external clamping diodes, which

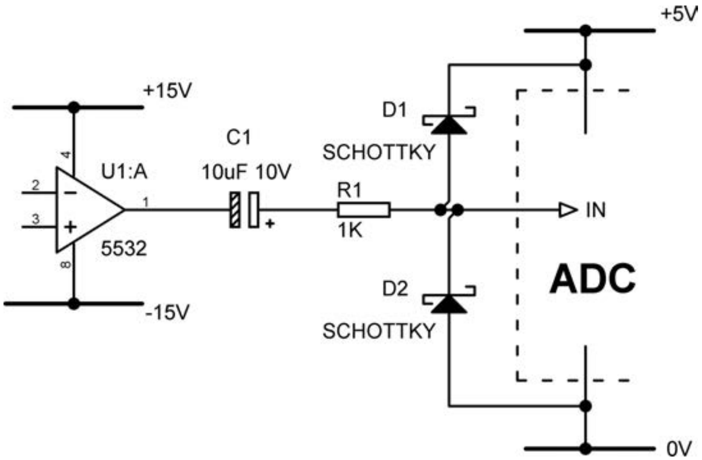


Figure 26.2: Diode clamping circuit to prevent overdriving an ADC input. Note that the diodes must be Schottky types

in conjunction with a series resistance, will limit the voltage swing at the ADC input. The principle is shown in Figure 26.2; if the opamp output exceeds +5 V then D1 will conduct, while if it goes negative of 0 V D2 will conduct, safely clamping the ADC input.

A vital point here is that the clamp diodes must be of the Schottky type, so their forward voltage is substantially less than that of the conventional silicon diodes on-chip, for otherwise they will give little or no protection. The on-chip diodes will be warmer and would conduct before conventional external silicon diodes.

R1 must be large enough to limit the current in D1, D2 to safe levels, but not so large that it causes a roll-off with the ADC input capacitance. It must also not be so large that the non-linear capacitance of the diodes causes significant non-linearity; 1 KΩ should be safe in this respect. Note that R1 is also useful in isolating the opamp output from the ADC input capacitance, which can otherwise erode stability margins.

Some typical DACs

Unlike ADCs, DACs come in two different types – voltage output and current output. Both types of output require some kind of low-pass filtering, but current output DACs also need current-to-voltage (I–V) conversion stages. There are a large number of DACs on the market, and it is essential to be selective in examining a few typical devices. Once again, the inclusion of a device here does not mean that I am giving it my personal recommendation. All the devices mentioned here are capable of 24-bit operation. And now in alphabetical order:

The Analog Devices AD1854 is a stereo audio DAC delivering 113 dB Dynamic Range and 112 dB SNR (A-weighted) at a 48 kHz sample rate. Maximum sample rate is 96 kHz. Differential analogue voltage outputs give a maximum output of 5.6 $V_{pk\text{-}pk}$ at full-scale and the output impedance is less than 200 Ω. It operates from a single +5 V supply rail, though there are separate supply pins for the analogue and digital sections.

The Texas PCM1794A is a stereo audio DAC supporting sample rates up to 192 kHz. It has differential analogue current outputs giving a maximum of 7.8 $mA_{pk\text{-}pk}$ at full-scale. The analogue section is powered from +5 V, the digital section from +3.3 V.

The Wolfson WM8740 is a stereo audio DAC supporting word lengths from 16 to 24 bits and sample rates up to 192 kHz. Differential analogue voltage outputs give a maximum output of 2.82 $V_{pk\text{-}pk}$ at full-scale. It can operate from a single +5 V supply rail, or the digital section can be run from +3.3 V to reduce power consumption.

Interfacing with DAC outputs

Modern DACs use oversampling so that brickwall reconstruction filters are not necessary at the analogue outputs. Nonetheless, some low-pass filtering is essential to remove high-frequency components that could cause trouble downstream from the output.

If you are using a DAC with a current output, the first thing you have to do is convert that current to a voltage. This is usually done with a shunt-feedback stage as shown in Figure 26.3, frequently called an I–V converter. The opamp most popular for this job (and in fact explicitly recommended for the Texas PCM1794A) is no less than our old friend the 5534/5532. The filter capacitors C1, C2 keep down the slew rate required at the outputs of the I–V converters, and with their parallel resistors give a −3 dB roll-off at 88.2 kHz. The current output is simply scaled by the value of R1, R2 and results in a voltage output of 3.20 V peak for 3.9 mA peak out (half of the total 7.8 $mA_{pk\text{-}pk}$ FS output) and when the two anti-phase voltages are combined in the differential amplifier that follows the total output is 6.4 V peak or 4.5 Vrms. The differential amplifier has its own HF roll-off at 151 kHz to give further filtering, implemented by C3 and C4. The capacitors used must be linear; NP0 ceramic, polystyrene or polypropylene are the only types suitable.

Voltage output DACs are somewhat simpler to apply as there is no need for I–V converters and the outputs can drive an active low-pass filter directly. The output is usually differential to obtain enough voltage-swing capability within the limited supply voltage available,

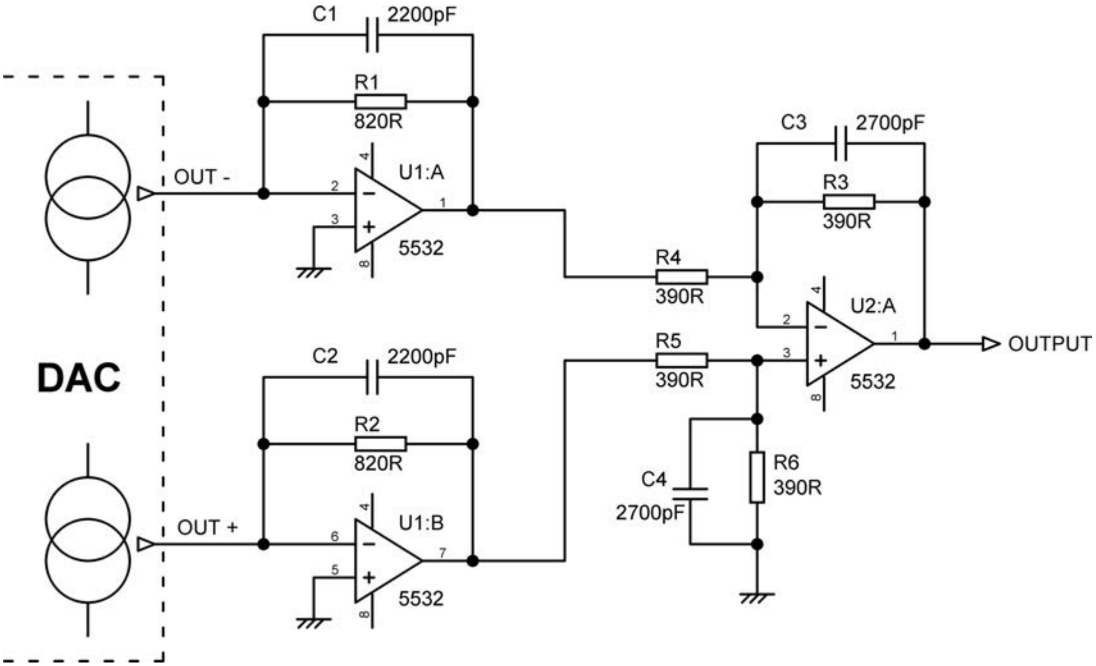


Figure 26.3: A typical output stage for a current-output DAC, with I–V converters and differential output filter

so differential to single-ended conversion is still required, and this is often cunningly implemented in the form of a differential low-pass filter.

Figure 26.4 shows a typical differential low-pass filter system; it has a third-order Bessel characteristic with a corner frequency of 92 kHz. The outputs are combined, and the first two poles are implemented by the differential multiple-feedback filter around U1:A and the third pole is produced by the passive network R7, C5. Note that the circuitry uses E96 resistor values in order to obtain the desired accuracy. Multiple-feedback filters are often preferred for this kind of application because they do not suffer from the failure of attenuation at very high frequencies that afflicts Sallen and Key filters, due to the inability of the opamp to maintain a low impedance at its output when its open-loop gain, and hence its feedback factor, has fallen to a low value.

It must not be assumed from this that all DACs have differential outputs. For example, the Wolfson WM8726, described as a 'low cost stereo DAC' has single-ended voltage outputs; it is recommended they are followed by a second-order low-pass filter.

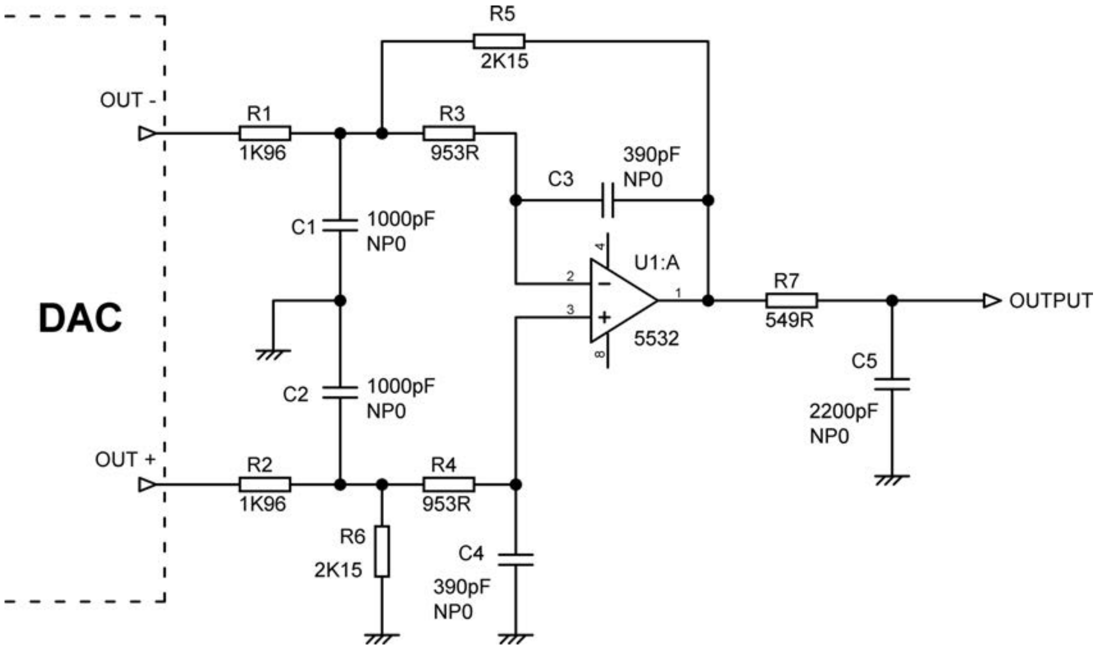


Figure 26.4: A typical output stage for a voltage-output DAC, with a differential output filter

Interfacing with microcontrollers

Having looked at interfacing with ADCs and DACs in digital audio, we turn to the important business of interfacing with microcontrollers. Much audio equipment keeps the signals wholly in the analogue domain, but uses a microcontroller to handle humble but essential housekeeping tasks such as muting relay control at power-up and power-down, and input select relay control. Volume control ICs keep the signal analogue but are controlled by serial data, generated when the microcontroller is driven from a rotary encoder connected to the volume knob. Input select ICs (an array of high-voltage analogue gates), such as the Toshiba TC9163, likewise require serial data to control them. Many pieces of equipment have remote controls, so a microcontroller is essential to decode the incoming RC5 data and activate the correct bit of circuitry. LCD displays typically also take serial data, and with a large display the software can get complex. The microcontroller will also take the unit out of standby when a 12 V trigger is received.

PIC devices are very popular, and the 16F72 is used as an example here. There are many other microcontroller families; the same basic principles apply. The 16F72 is a capable but inexpensive device that incorporates flash program memory, timers, serial ports, and

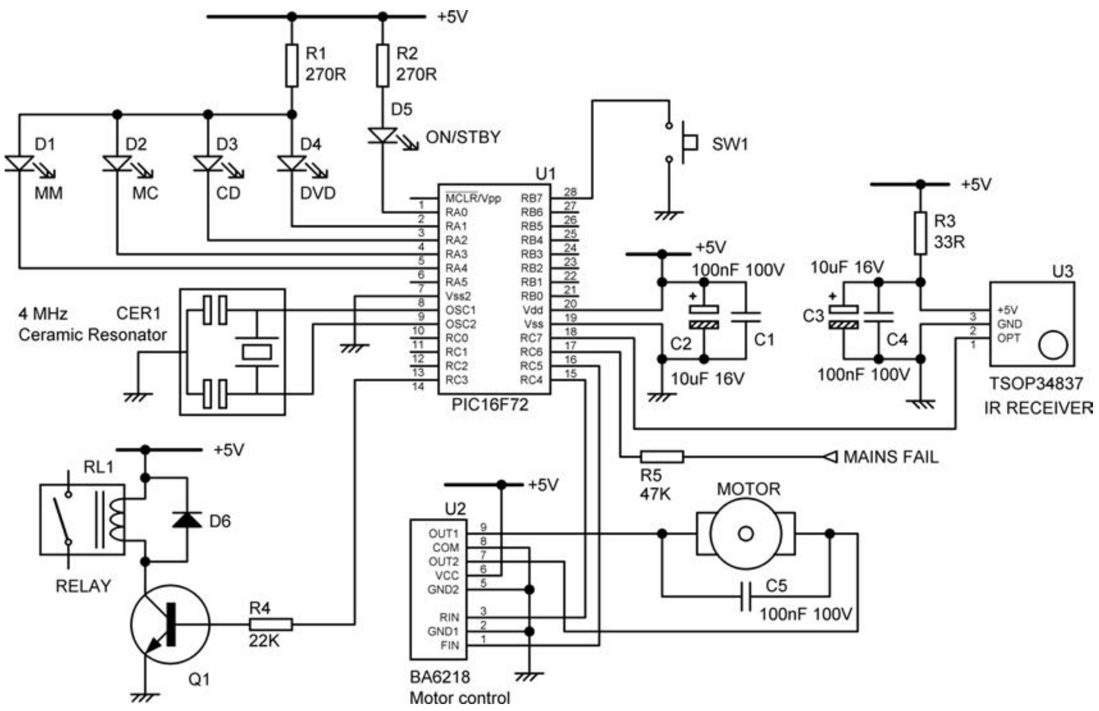


Figure 26.5: Examples of interfacing to a PIC 16F72 microcontroller for preamp housekeeping

an 8-bit A/D converter, but here the three 8-bit I/O ports A, B and C are used to illustrate a preamplifier application. Each port pin can be separately defined to be an input or an output.

Figure 26.5 is not a complete system but illustrates various techniques. The PIC output ports can drive LEDs directly, such as D5 with resistor R2 to set the current. LEDs D1 to D4 indicate which preamp input is selected, and since only one is ever on at a time they can share resistor R1. For loads such as relays that take more current than an output port can handle directly, a transistor Q1 can be used. Note the suppression diode across the relay coil.

The BA6218 control IC for the motorised volume control is driven by the two output pins RC4 and RC5. These are simple logic outputs and not a serial connection. See Chapter 13 for more on motorised volume controls.

Remote control commands are demodulated and filtered by the IR receiver U3, and the serial data output is applied to RC6. The IR commands are then decoded by software in the PIC.

The B port has a useful extra option. It can be set so that each pin has a "weak-pullup" of 250 μA to the +5 V rail; this means that switches like SW1 do not need an external pullup resistor. When there are a large number of switches to be read a row-and-column matrix is much more economical of input pins.

The PIC clock frequency is set by ceramic resonator CER1. These components are much cheaper than quartz crystals, and their 0.5% frequency tolerance (crystals show 0.001%) is more than good enough for setting turn-on delays and so on.

Input port voltages must be clamped to ensure they do not exceed the permissible limits. The PIC has protective clamp diodes on all pins, but these have a limited current capability, and if you plan to use them for clamping as opposed to just static protection, a series resistor of around 47 kΩ is recommended, such as R5 on the MAINS FAIL input.

It is an important principle of interfacing that you should not rely on a low at an output port to prevent, say, the mute relay from energising, perhaps by diverting the drive away from the base of the relay-control transistor. At power-up the output ports will be at high impedance. This is sometimes referred to as being 'tri-stated' as the high-impedance state is neither the high state nor the low state, but a third 'off' state. There may be momentary closure of the relay if it operates faster than the microcontroller can initialise itself and write a low to the output port.

More seriously, at power-down a mute relay must open as quickly as possible, to prevent thumps, *and it must stay off*. The mains-fail circuitry will detect power-down and signal this to the microcontroller, which will write a low to the output port to open the relay. But . . . the output ports will go high impedance when the microcontroller supply falls. Relays have a wide operating voltage range, so the mute relay will close again, letting through all sorts of unpleasant disturbances, and will not finally open until the supply drops below the hold-on voltage, which may take some time. This is a much more likely scenario than misoperation at switch-on.

Avoid configuring things so that either a high or a low on an output port must be asserted to prevent something operating.

Many microcontrollers have a sleep mode which can be used to minimise the chance of electrical interference. I have never found it necessary to employ this in a piece of audio equipment; RF from the microcontroller getting into the audio has never been a problem. I do know of a case where it was done very successfully in an HF communications receiver, where the RF sensitivity is of course very much greater. Be aware that some microcontrollers take a surprisingly long time to come out of sleep mode. Much the same is true of me.

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Note: Page numbers in **boldface** indicate tables; those in *italic* indicate figures.