are shown in Figure 10.16, where it can be clearly seen that THD has been halved by this simple change. To the best of my knowledge this is a new result; if you must work in Class-AB, then keep the emitter resistors as low as possible, to minimize the gain changes.

Having considered the linearity of Classes A and AB, we must not neglect what effect this radical Re change has on Class-B linearity. The answer is not very much (see Figure 10.17, where crossover distortion seems to be slightly higher with Re = 0.2 Ω than for either 0.1 or 0.4 Ω). Whether this is a consistent effect (for CFP stages anyway) remains to be seen.

The detailed mechanisms of bias control and mode switching are described in the next section.

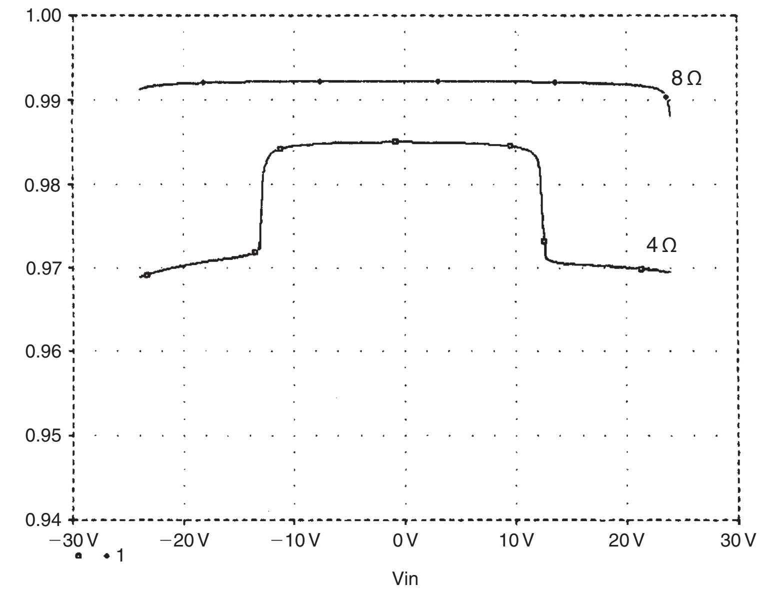

Figure 10.15: CFP output linearity with Re = 0R1, re-biased to keep I_q at 1.5 A. There is slightly poorer linearity in the flat-topped Class-A region than for Re = 0R22, but the 4 Ω AB steps are halved in size at 0.012 units. Note that both gains are now closer to unity. Same scale as in Figure 9.14

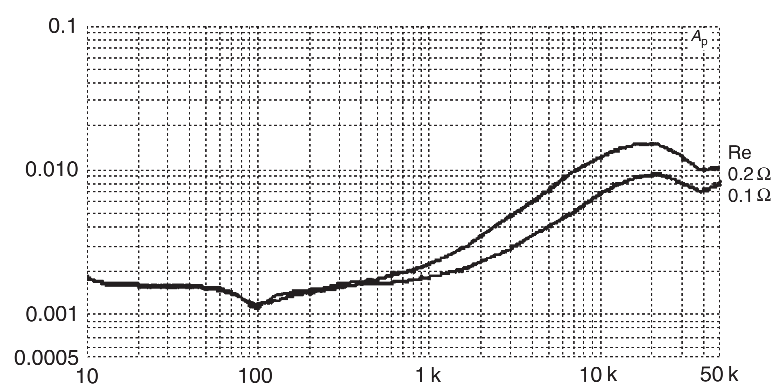

Figure 10.16: Distortion in Class-AB is reduced by lowering the value of Re

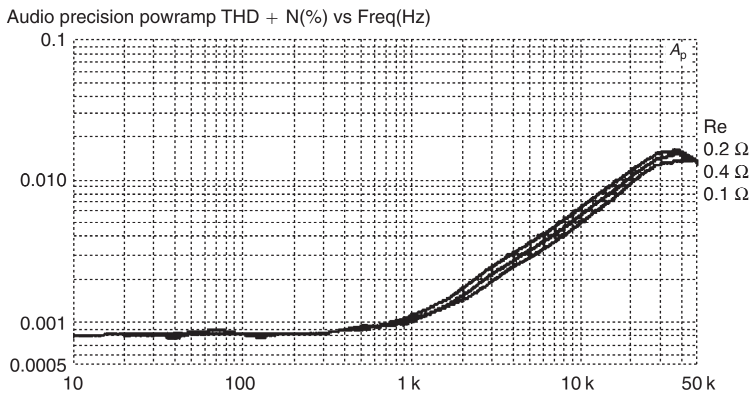

Figure 10.17: Proving that emitter resistors matter much less in Class-B. Output was 20 W in 8 Ω, with optimal bias. Interestingly, the bias does not need adjusting as the value of Re changes

Trimodal Biasing

Figure 10.18 shows a simplified rendering of the Trimodal biasing system; the full version appears in Figure 10.19. The voltage between points A and B is determined by one of two controller systems, only one of which can be in command at a time. Since both are basically shunt voltage regulators sitting between A and B, the result is that the lowest voltage wins. The novel Class-A current controller introduced on page 307 is used here adapted for 0.1 Ω emitter resistors, mainly by reducing the reference voltage to 300 mV, which gives a quiescent current (I_q) of 1.5 A when established across the total emitter resistance of 0.2 Ω.

In parallel with the current controller is the V_{be}-multiplier TR13. In Class-B mode, the current controller is disabled, and critical biasing for minimal crossover distortion is provided in the usual way by adjusting preset PR1 to set the voltage across TR13. In Class-A/AB mode, the voltage TR13 attempts to establish is increased (by shorting out PR1) to a value greater than that required for Class-A. The current controller therefore takes charge of the voltage between X and Y, and unless it fails TR13 does not conduct. Points A, B, X, and Y are the same circuit nodes as in the simple Class-A design (see Figure 10.6c).

Class-A/AB Mode

In Class-A/AB mode, the current controller (TR14, TR15, TR16 in Figure 10.18) is active and TR13 is off, as TR20 has shorted out PR1. TR15, TR16 form a simple differential amplifier that compares the reference voltage across R31 with the V_{bias} voltage across output emitter resistors R16 and R17; as explained above, in Class-A this voltage remains constant despite delivery of current into the load. If the voltage across TR16, TR17 tends to rise, then TR16 conducts more, turning TR14 more on and reducing the voltage between A and B. TR14, TR15, and TR16 all move up and down with the amplifier output, and so a tail-current source (TR17) is used.

I am very aware that the current controller is more complex than the simple V_{be}-multiplier used in most Class-B designs. There is an obvious risk that an assembly error could cause a massive current that

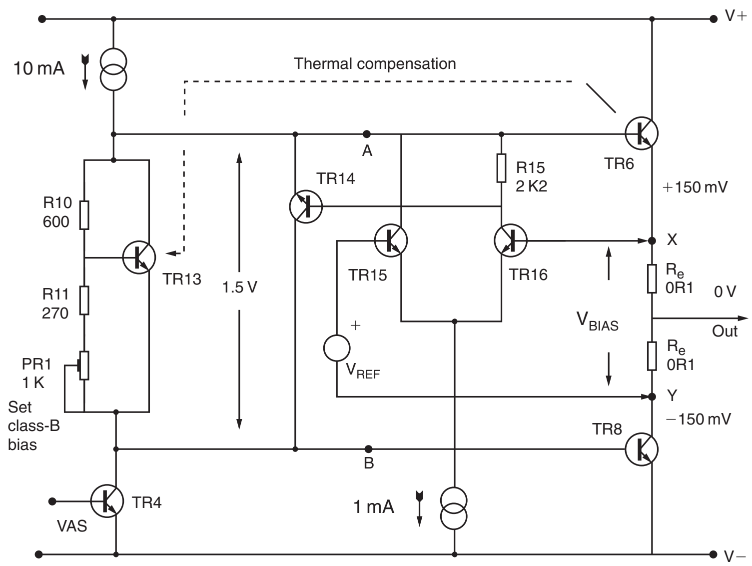

Figure 10.18: The simplified current controller in action, showing typical DC voltages in Class-A. Points A, B, X, and Y are in Figure 10.6

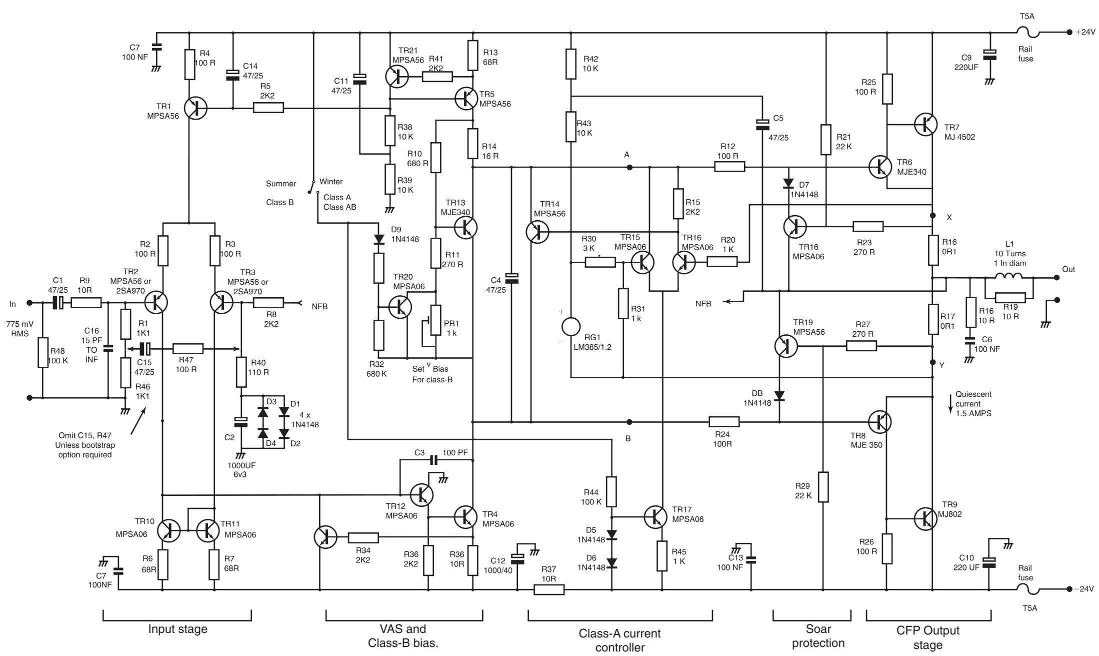

Figure 10.19: The complete circuit diagram of the Trimodal amplifier, including the optional bootstrapping components, R47 and C15

would prompt the output devices to lay down their lives to save the rail fuses. The tail-source TR17 is particularly vulnerable because any fault that extinguishes the tail current removes the drive to TR14, the controller is disabled, and the current in the output stage will be very large. In Figure 10.18 the V_{be}-multiplier TR13 acts as a safety circuit that limits V_{bias} to about 600 mV rather than the normal 300 mV, even if the current controller is completely non-functional and TR14 fully off. This gives a quiescent of 3.0A, and I can testify this is a survivable experience for the output devices in the short term; however, they may eventually fail from overheating if the condition is allowed to persist.

There are some important points about the current controller. The entire tail current for the error amplifier, determined by TR17, is siphoned off from VAS current source TR5, and must be taken into account when ensuring that the upper output half gets enough drive current.

There must be enough tail current available to turn on TR14, remembering that most of TR16 collector current flows through R15, to keep the pair roughly balanced. If you feel moved to alter the VAS current, remember also that the base current for driver TR6 is higher in Class-A than Class-B, so the positive slew rate is slightly reduced in going from Class-A to Class-B.

The original Class-A amplifier used a National LM385/1.2, its output voltage fixed at 1.223V nominal; this was reduced to approximately 0.6V by a 1k–1k potential divider. The circuit also worked well with V_{ref} provided by a silicon diode, 0.6V being an appropriate V_{bias} drop across two 0.22Ω output emitter resistors. This is simple, and retains the immunity of I_q to heat-sink and output device temperatures, but it does sacrifice the total immunity to ambient temperature that a band-gap reference gives.

The LM385/1.2 is the lowest voltage band-gap reference commonly available; however, the voltages shown in Figure 10.18 reveal a difficulty with the new lower V_{bias} value and the CFP stage; points A and Y are now only 960mV apart, which does not give the reference room to work in if it is powered from node A, as in the original circuit. The solution is to power the reference from the V+ rail, via R42 and R43. The mid-point of these two resistors is bootstrapped from the amplifier output rail by C5, keeping the voltage across R43 effectively constant. Alternatively, a current source could be used, but this might reduce positive headroom. Since there is no longer a strict upper limit on the reference voltage, a more easily obtainable 2.56V device could be used providing R30 is suitably increased to 5k to maintain V_{ref} at 300mV across R31.

In practical use, I_q stability is very good, staying within 1% for long periods. The most obvious limitation on stability is differential heating of TR15, TR16 due to heat radiation from the main heat-sink. TR14 should also be sited with this in mind, as heating it will increase its beta and slightly imbalance TR15, TR16.

Class-B Mode

In Class-B mode, the current controller is disabled by turning off tail-source TR17 so TR14 is firmly off, and critical biasing for minimal crossover distortion is provided as usual by V_{be}-multiplier TR13. With 0.1Ω emitter resistors V_{bias} (between X and Y) is approximately 10mV. I would emphasize that in Class-B this design, if constructed correctly, will be as Blameless as a purpose-built Class-B amplifier. No compromises have been made in adding the mode-switching.

As in the previous Class-B design, the addition of R14 to the V_{be}-multiplier compensates against drift of the VAS current-source TR5. To make an old but much-neglected point, the preset should always be in the bottom arm of the V_{be}-divider R10, R11, because when presets fail it is usually by the wiper going open; in the bottom arm this gives minimum V_{bias}, but in the upper it would give maximum.

In Class-B, temperature compensation for changes in driver dissipation remains vital. Thermal runaway with the CFP is most unlikely, but accurate quiescent setting is the only way to minimize crossover distortion. TR13 is therefore mounted on the same small heat-sink as driver TR6. This is often called thermal feedback, but it is no such thing as TR13 in no way controls the temperature of TR6; 'thermal feedforward' would be a more accurate term.

The Mode-Switching System

The dual nature of the biasing system means Class-A/Class-B switching can be implemented fairly simply. A Class-A amplifier is an uneasy companion in hot weather, and so I have been unable to resist the temptation to subtitle the mode switch summer/winter, by analogy with a car air intake.

The switchover is DC-controlled, as it is not desirable to have more signal than necessary running around inside the box, possibly compromising interchannel crosstalk. In Class-A/AB mode, SW1 is closed, so TR17 is biased normally by D5, D6, and TR20 is held on via R33, shorting out preset PR1 and setting TR13 to safety mode, maintaining a maximum V_{bias} limit of 600 mV. For Class-B, SW1 is opened, turning off TR17 and therefore TR15, TR16, and TR14. TR20 also ceases to conduct, protected against reverse-bias by D9, and reduces the voltage set by TR13 to a suitable level for Class-B. The two control pins of a stereo amplifier can be connected together, and the switching performed with a single-pole switch, without interaction or increased crosstalk.

The mode-switching affects the current flowing in the output devices, but not the output voltage, which is controlled by the global feedback loop, and so it is completely silent in operation. The mode may be freely switched while the amplifier is handling audio, which allows some interesting A/B listening tests.

It may be questioned why it is necessary to explicitly disable the current controller in Class-B; TR13 is establishing a lower voltage than the current controller, the subsystem of which will therefore turn TR14 off as it strives in a futile manner to increase V_{bias}. This is true for 8 Ω loads, but 4 Ω impedances increase the currents flowing in R16, R17 so they are transiently greater than the Class-A I_q, and the controller will therefore intermittently take control in an attempt to reduce the average current to 1.5 A. Disabling the controller by turning off TR17 via R44 prevents this.

If the Class-A controller is enabled, but the preset PR1 is left in circuit (e.g. by shorting TR20 base emitter) we have a test mode that allows suitably cautious testing; I_q is zero with the preset fully down, as TR13 overrides the current controller, but increases steadily as PR1 is advanced, until it suddenly locks at the desired quiescent current. If the current controller is faulty then I_q continues to increase to the defined maximum of 3.0 A.

Thermal Design

Class-A amplifiers are hot almost by definition, and careful thermal design is needed if they are to be reliable, and not take the varnish off the Sheraton. The designer has one good card to play; since the internal dissipation of the amplifier is maximal with no signal, simply turning on the prototype and leaving it to idle for several hours will give an excellent idea of worst-case component temperatures. In Class-B the power dissipation is very program-dependent, and estimates of actual device temperatures in realistic use are notoriously hard to make.

Table 10.5 shows the output power available in the various modes, with typical transformer regulation, etc.; the output mode diagram in Figure 10.11 shows exactly how the amplifier changes mode from A to AB with decreasing load resistance. Remember that in this context 'high distortion' means 0.002% at 1 kHz. This diagram was produced in the analysis section of PSPICE simply by typing in equations, and without actually simulating anything at all.

The most important thermal decision is the size of the heat-sink; it is going to be expensive, so there is a powerful incentive to make it no bigger than necessary. I have ruled out fan cooling as it tends to make concern for ultra-low electrical noise look rather foolish; let us rather spend the cost of the fan on extra cooling fins and convect in ghostly silence. The exact thermal design calculations are simple but tedious, with many parameters to enter – the perfect job for a spreadsheet. The final answer is the margin between the predicted junction temperatures and the rated maximum. Once power output and impedance range are decided, the heat-sink thermal resistance to ambient is the main variable to manipulate, and this is a compromise between coolness and cost, for high junction temperatures always reduce semiconductor reliability. This is summarized very roughly in Table 10.6.

Table 10.6 shows that the transistor junctions will be 80°C above ambient, i.e. at around 100°C; the rated junction maximum is 200°C, but it really is not wise to get anywhere close to this very real limit. Note the case-sink thermal washers were high-efficiency material, and standard versions have a slightly higher thermal resistance.

Table 10.5: Power capability

	Load resistance			Distortion
	8 Ω	**6 Ω**	**4 Ω**	
Class-A	20 W	27 W	15 W	Low
Class-AB	n/a	n/a	39 W	High
Class-B	21 W	28 W	39 W	Medium

Table 10.6: Temperature rises resulting in a 100°C junction temperature

	Thermal resistance (°C/W)	Heat flow (W)	Temp. rise (°C)	Temp. (°C)
Junction to TO3 case	0.7	36	25	100 junction
Case to sink	0.23	36	8	75 TO3 case
Sink to air	0.65	72	47	67 heat-sink
Total			80	20 ambient

The heat-sinks used in the prototype had a thermal resistance of 0.65°C/W per channel. This is a substantial chunk of metal, and since aluminum is basically congealed electricity, it's bound to be expensive.

A Complete Trimodal Amplifier Circuit

The complete Class-A amplifier is shown in Figure 10.19, complete with optional input bootstrapping. It may look a little complex, but we have only added four low-cost transistors to realize a high-accuracy Class-A quiescent controller, and one more for mode-switching. Since the biasing system has been described above, only the remaining amplifier subsystems are dealt with here.

The input stage follows my design methodology by using a high tail current to maximize transconductance, and then linearizing by adding input degeneration resistors R2, R3 to reduce the final transconductance to a suitable level. Current-mirror TR10, TR11 forces the collector currents of the two input devices TR2, TR3 to be equal, balancing the input stage to prevent the generation of second-harmonic distortion. The mirror is degenerated by R6, R7 to eliminate the effects of V_{be} mismatches in TR10, TR11. With some misgivings I added the input network R9, C15, which is definitely not intended to define the system bandwidth, unless fed from a buffer stage; with practical values the HF roll-off could vary widely with the source impedance driving the amplifier. It is intended rather to give the possibility of dealing with RF interference without having to cut tracks. R9 could be increased for bandwidth definition if the source impedance is known, fixed, and taken into account when choosing R9; bear in mind that any value over 47 Ω will measurably degrade the noise performance. The values given roll-off above 150 MHz to keep out UHF.

The input stage tail current is increased from 4 to 6 mA, and the VAS standing current from 6 to 10 mA over the original Chapter 7 circuit. This increases maximum positive and negative slew rates from +21, −48 to +37, −52 V/µs; as described in Chapter 8, this amplifier architecture is bound to slew asymmetrically. One reason is feed-through in the VAS current source; in the original circuit an unexpected slew-rate limit was set by fast edges coupling through the current-source c-b capacitance to reduce the bias voltage during positive slewing. This effect is minimized here by using the negative-feedback type of current-source bias generator, with VAS collector current chosen as the controlled variable. TR21 senses the voltage across R13, and if it attempts to exceed V_{be}, turns on further to pull up the bases of TR1 and TR5. C11 filters the DC supply to this circuit and prevents ripple injection from the V+ rail. R5, C14 provide decoupling to prevent TR5 from disturbing the tail current while controlling the VAS current.

The input tail-current increase also slightly improves input stage linearity, as it raises the basic transistor g_m and allows R2, R3 to apply more local NFB.

The VAS is linearized by beta-enhancing stage TR12, which increases the amount of local NFB through Miller dominant-pole capacitor C3 (i.e. C_{dom}). R36 has been increased to 2 k2 to minimize power dissipation, as there seems no significant effect on linearity or slewing. Do not omit it altogether, or linearity will be affected and slewing much compromised.

As described in Chapter 9, the simplest way to prevent ripple from entering the VAS via the V− rail is old-fashioned RC decoupling, with a small R and a big C. We have some 200 mV in hand (see page 314) in the negative direction, compared with the positive, and expending this as the voltage-drop through the RC decoupling will give symmetrical clipping. R37 and C12 perform this function; the low rail voltages in this design allow the 1000 μF C12 to be a fairly compact component.

The output stage is of the CFP type that, as previously described, gives the best linearity and quiescent stability, due to the two local negative-feedback loops around driver and output device. Quiescent stability is particularly important with R16, R17 at 0.1 Ω, and this low value might be rather dicey in a double EF output stage. The CFP voltage efficiency is also higher than the EF version. R25, R26 define a suitable quiescent collector current for the drivers TR6, TR8, and pull charge carriers from the output device bases when they are turning off. The lower driver is now a BD136; this has a higher fT than the MJE350, and seems to be more immune to odd parasitics at negative clipping.

The new lower values for the output emitter resistors R16, R17 halve the distortion in Class-AB. This is equally effective when in Class-A with too low a load impedance, or in Class-B but with I_q maladjusted too high. It is now true in the latter case that too much I_q really is better than too little – but not much better, and AB still comes a poor third in linearity to Classes A and B.

Safe operating area (SOAR) protection is given by the networks around TR18, TR19. This is a single-slope SOAR system that is simpler than two-slope SOAR, and therefore somewhat less efficient in terms of getting the limiting characteristic close to the true SOAR of the output transistor. In this application, with low rail voltages, maximum utilization of the transistor SOAR is not really an issue; the important thing is to observe maximum junction temperatures in the A/AB mode.

The global negative-feedback factor is 32 dB at 20 kHz, and this should give a good margin of safety against Nyquist-type oscillation. Global NFB increases at 6 dB/octave with decreasing frequency to a plateau of around 64 dB, the corner being at a rather ill-defined 300 Hz; this is then maintained down to 10 Hz. It is fortunate that magnitude and frequency here are non-critical, as they depend on transistor beta and other doubtful parameters.

It is often stated in hi-fi magazines that semiconductor amplifiers sound better after hours or days of warm-up. If this is true (which it certainly is not in most cases) it represents truly spectacular design incompetence. This sort of accusation is applied with particular venom to Class-A designs, because it is obvious that the large heat-sinks required take time to reach final temperature, so I thought it important to state that in Class-A this design stabilizes its electrical operating conditions in less than a second, giving the full intended performance. No 'warm-up time' beyond this is required; obviously the heat-sinks take time to reach thermal equilibrium but, as described above, measures have been taken to ensure that component temperature has no significant effect on operating conditions or performance.

The Power Supply

A suitable unregulated power supply is that shown in Figure 9.2; a transformer secondary voltage of 20–0–20 V rms and reservoirs totaling 20,000 μF per rail will give approximately ±24 V. This supply must be designed for continuous operation at maximum current, so the bridge rectifier must be properly heat-sunk, and careful consideration given to the ripple-current ratings of the reservoirs. This is one reason why reservoir capacitance has been doubled to 20,000 μF per rail, over the 10,000 μF that was adequate for the Class-B design; the ripple voltage is halved, which improves voltage efficiency as it is the ripple troughs that determine clipping onset, but in addition the ripple current, although unchanged in total value, is now split between two components. (The capacitance was not increased to reduce ripple injection, which is dealt with far more efficiently and economically by making the PSRR high.) Do not omit the secondary fuses; even in these modern times rectifiers do fail, and transformers are horribly expensive.

The Performance

The performance of a properly designed Class-A amplifier challenges the ability of even the Audio Precision measurement system. To give some perspective on this, Figure 10.20 shows the distortion of the AP oscillator driving the analyzer section directly for various bandwidths. There appear to be internal mode changes at 2 and 20 kHz, causing step increases in oscillator distortion content; these are just visible in the THD plots for Class-A mode.

Figure 10.21 shows Class-B distortion for 20 W into 8 and 4 Ω, while Figure 10.22 shows the same in Class-A/AB. Figure 10.23 shows distortion in Class-A for varying measurement bandwidths. The lower bandwidths misleadingly ignore the HF distortion, but give a much clearer view of the excellent linearity below 10 kHz. Figure 10.24 gives a direct comparison of Classes A and B. The HF rise for B is due to high-order crossover distortion being poorly linearized by negative feedback that falls with frequency.

Further Possibilities

One interesting extension of the ideas presented here is the adaptive Trimodal amplifier. This would switch into Class-B on detecting device or heat-sink over-temperature, and would be a unique example of an amplifier that changed mode to suit the operating conditions. The thermal protection

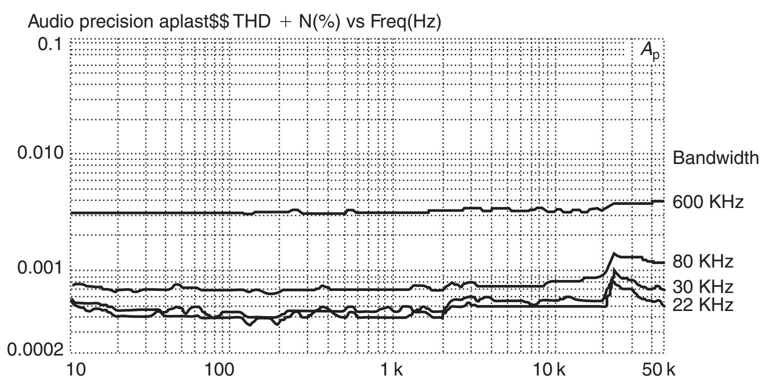

Figure 10.20: The distortion in the AP-1 system at various measurement bandwidths

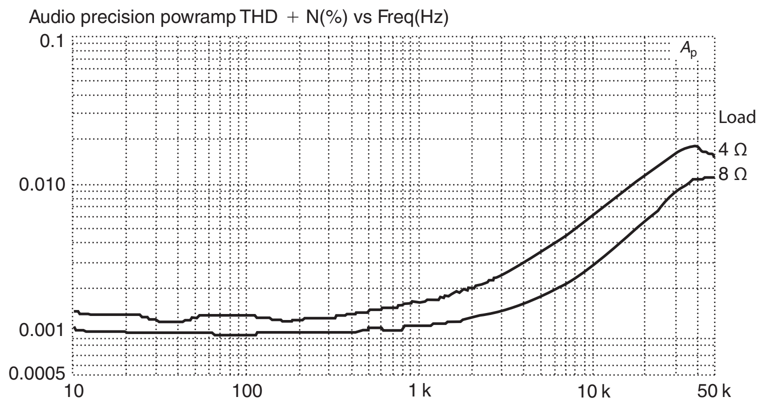

Figure 10.21: Distortion in Class-B (Summer) mode. Distortion into 4 Ω is always worse. Power was 20 W in 8 Ω and 40 W in 4 Ω, bandwidth 80 kHz

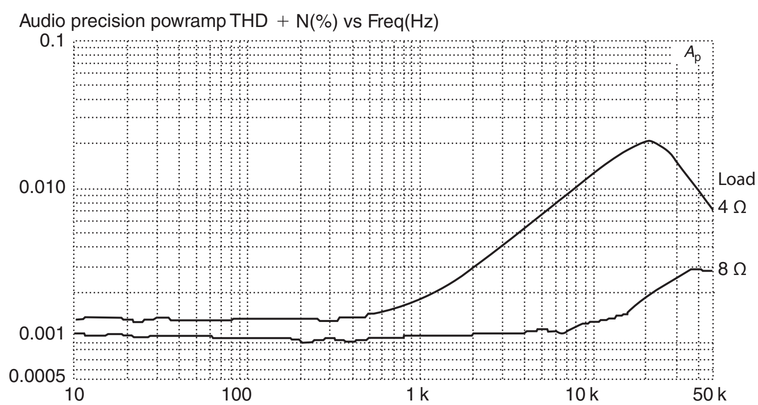

Figure 10.22: Distortion in Class-A/AB (Winter) mode, same power and bandwidth as in Figure 10.21. The amplifier is in AB mode for the 4 Ω case, and so distortion is higher than for Class-B into 4 Ω. At 80 kHz bandwidth, the Class-A plot below 10 kHz merely shows the noise floor

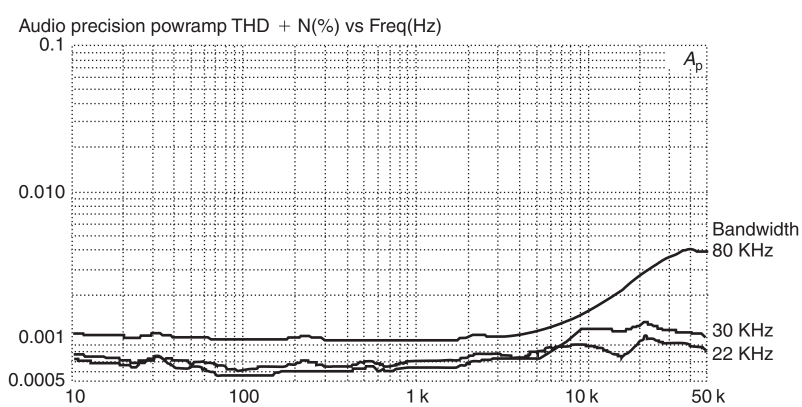

Figure 10.23: Distortion in Class-A only (20 W/8 Ω) for varying measurement bandwidths. The lower bandwidths ignore HF distortion, but give a much clearer view of the excellent linearity below 10 kHz

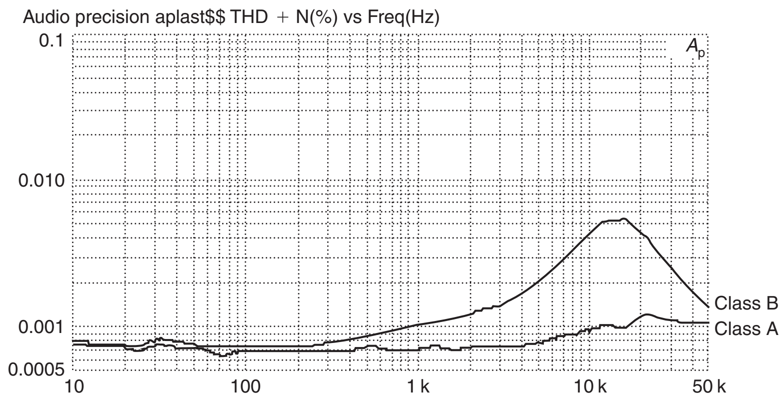

Figure 10.24: Direct comparison of Classes A and B (20 W/8 Ω) at 30 kHz bandwidth. The HF rise for B is due to the inability of negative feedback that falls with frequency to linearize the high-order crossover distortion in the output stage

would need to be latching; flipping from Class-A to Class-B every few minutes would subject the output devices to unnecessary thermal cycling.

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CHAPTER 11

Class-XD™: Crossover Displacement Technology

Class-XD™ is a new output stage technology I have devised which abolishes crossover distortion up to a certain power level, without any accompanying compromises. 'XD' is derived from the phrase 'crossover displacement': the technology is covered by British patent GB2424137B and is proprietary to Cambridge Audio. At the time of writing it has so far been used in the Azur 840A and 840 W power amplifiers, for both of which I did the electronic design. 'Class-XD' is a trademark of Audio Partnership PLC, and it should be pointed out that the use of the Class-XD concept and its trademark is restricted; I have permission from Cambridge Audio to use the term and describe the circuitry but no license to use the technology is implied or granted by the publication of this description.

Having held various posts in companies concerned with audio power amplifiers, I have frequently had to deal with enthusiastic inventors who feel they have come up with an output stage technology that overcomes the crossover distortion problems of conventional Class-B, and who are anxious to sell the idea to me. Two stick in the mind. There was the consortium that took out extensive worldwide patents on an idea that had been disclosed in *Wireless World* a quarter of a century before, and which did not work properly anyway. Then there was the chap who offered me an error-correcting output stage that 'only requires another 140 transistors'. I would have liked to have seen that circuit diagram, but not enough to pay money to do so.

In the light of this sort of thing, anyone is entitled to be skeptical about new and improved amplifier output stages. However, Class-XD is different; it really does work, doing what it claims with total reliability and minimal extra circuitry, as I shall now demonstrate.

One of the main themes of this book is the difficulty of dealing with crossover distortion in a Class-B output stage. I have described various methods of attack, such as the use of multiple output devices to reduce the current changes in each output transistor, and the use of two-pole compensation to increase the global negative-feedback factor. Both methods usefully reduce the amount of crossover distortion but do not eliminate it. As a result, one of the great divides in amplifier technology is still between efficient but imperfect Class-B and beautifully linear but dishearteningly inefficient Class-A. As I demonstrated in my book *Self On Audio*, a Class-A amplifier may theoretically be 50% efficient with a maximum sine-wave output, but when it reproduces a real music signal this falls to 1% or 2%[1]. For those that care at all about the economic utilization of energy, a Class-A amplifier is not an attractive proposition.

Class-B linearity can of course be very good. The Blameless amplifier design methodology, especially in its Load-Invariant form, yields less than 0.001% THD at 1 kHz. The limitation is that a Class-B amplifier inherently generates crossover distortion, and most inconveniently does so at the zero-crossing, so it is always present no matter how low the signal amplitude. At one unique setting of quiescent conditions the distortion produced is at a minimum, and this characterizes optimal Class-B, but at no value can it be made to disappear. It is inherent in the classical Class-B operation of a pair of output transistors.

Given these two alternatives, there has always been a desire for a compromise between the efficiency of Class-B and the linearity of Class-A. The most obvious approach is to turn up the quiescent current of a Class-B stage, to create an area of Class-A operation, with both output transistors conducting, around the zero-crossing. This area widens as the quiescent current increases, until ultimately it encompasses the entire voltage output range of the amplifier, and we have created a pure Class-A design where both output transistors are conducting all the time. There is thus a range of quiescent current between Class-B and Class-A, and this mode of operation is called Class-AB. It is certainly a compromise between Class-A and Class-B, but not a good compromise, as it introduces extra distortion of its own.

This appears when the signal exceeds the limits of the Class-A region. The THD worsens abruptly due to the sudden gain changes when the output transistors turn on and off, and linearity is inferior not only to Class-A but also to optimally biased Class-B. This effect is often called 'g_m-doubling' and is dealt with in detail in Chapter 6. Class-AB distortion can be made very low by proper design, such as using the lowest practicable emitter resistors, but it remains at least twice as high as for the equivalent Class-B situation. The bias control of a Class-B amplifier does *not* give a straightforward trade-off between power dissipation and linearity at all levels, despite the constant repetition this misguided notion receives in some parts of the audio press. To demonstrate this, Figure 11.1 shows THD plotted against output level for Classes AB and B.

What we really want is an amplifier that would give Class-A performance up to the transition level, with Class-B after that, rather than the unsatisfactory Class-AB. This would abolish the abrupt AB gain changes that generate the extra distortion.

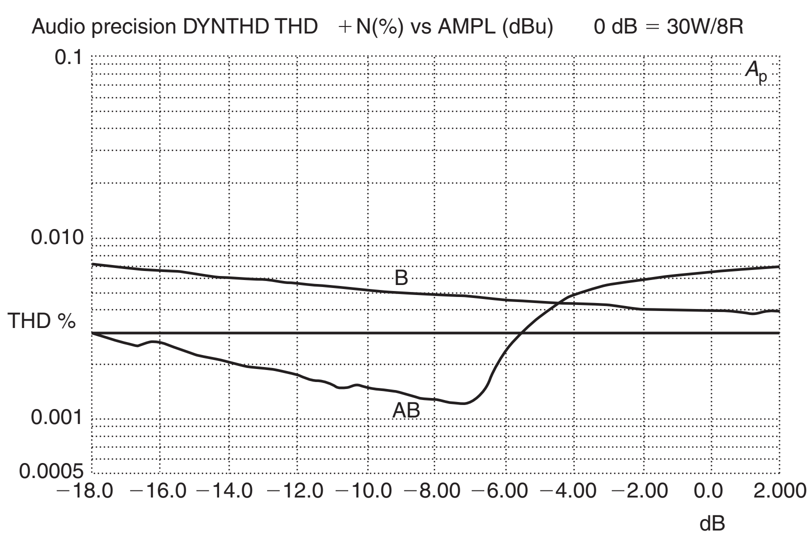


Figure 11.1: THD versus level for Class-B and Class-AB (0 dB is 30 W into 8 Ω)

The Crossover Displacement Principle

When we consider Class-B, it is clear that it would be better if the crossover region were anywhere else rather than where it is. If we can displace the crossover point away from its zero-crossing position, then the amplifier output will not traverse it until the output reaches a certain voltage level. Below this level the performance is pure Class-A; above it the performance is optimal Class-B, the only difference being that crossover discontinuities on the THD residual are no longer evenly spaced. The harmonic structure of the crossover distortion produced is not significantly changed, as explained in more detail below.

The central idea of the crossover displacement principle is the injection of an extra current, either fixed or varying with the signal, into the output point of a conventional Class-B amplifier. This is illustrated in Figure 11.2, where a black box I have called the 'displacer' draws a controlled displacement current from the output and sinks it into the negative rail; sourcing current from the positive rail and injecting it into the output would be equally valid. The displacer current may be constant, or vary with the signal.

The displacement current does not directly alter the output voltage because the output stage has an inherently low output impedance, which is further reduced by the global negative feedback. What it does do is alter the pattern of current flowing in the output devices. The displacement current in the version shown here is sunk to V− from the output, This is arbitrary as the direction of displacement makes no difference. The extra current therefore flows through Re1, and the extra voltage drop across it means the output voltage must go some way negative before the current through Re1 stops and that in Re2 starts. In other words, the crossover point when Q2 hands over to Q4 has been moved to a point negative of the 0 V rail; I refer to this as the 'transition point' between Class-A and Class-B. For output levels below transition both Q2 and Q4 are conducting and no crossover distortion is generated. The resulting change in the incremental gain of the output stage is shown in Figure 11.3. Here the crossover region is moved 8 V negative of ground by a 1 A displacement

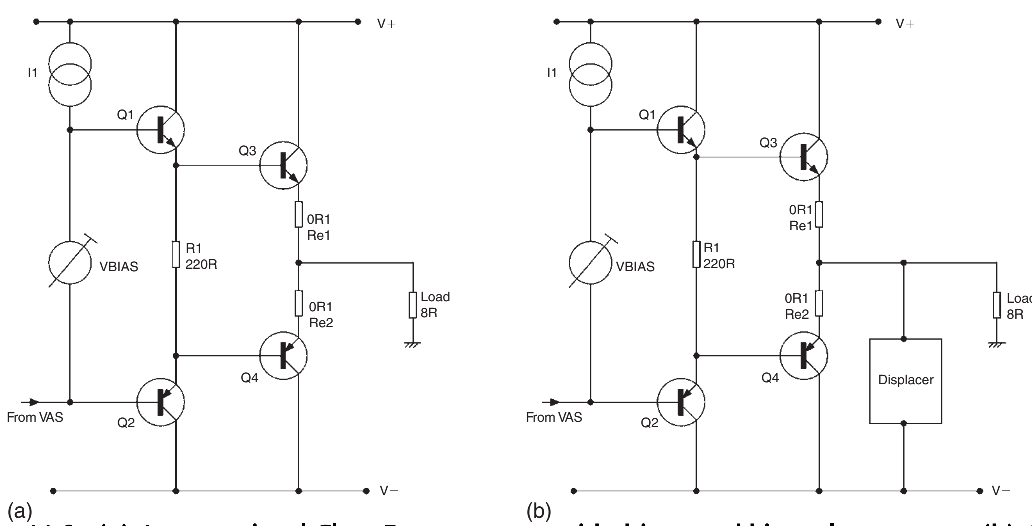


Figure 11.2: (a) A conventional Class-B output stage with drivers and bias voltage source. (b) Adding a displacer system that draws current from the output and sinks it into the negative rail

current; if the displacer had been connected to the positive rail the crossover region would have been pulled upwards. Note that the vertical scale is very much exaggerated, and that the crossover region has been moved but remains the same shape – the existing linearity has not been compromised.

I should emphasize here that crossover displacement in no way renders output stage bias adjustment unnecessary; if it is wrong the same distortion will occur, though only above a certain output level. This could be regarded as making the adjustment less critical, but getting it right costs no more than getting it roughly right, so there is really nothing to be gained by compromising on this.

We now have before us the intriguing prospect of a power amplifier with three output devices, which if nothing else is novel. The operation of the output stage is inherently asymmetrical, and indeed this is the whole point, but it should not cause alarm. Circuit symmetry is often touted as being a prerequisite for either low distortion or respectable operation in general, but this has no real foundation. A perfectly symmetrical circuit may have no even-order distortion, but it may still have frightening amounts of odd-order nonlinearity, such as a cubic characteristic. Odd-order harmonics are normally considered more dissonant than even-order, so circuit symmetry in itself is not enough.

In a conventional optimal Class-B amplifier, the crossover events are evenly spaced in time. In the crossover displacement amplifier, the crossover events are asymmetrical in time and put energy into both even and odd harmonics when operating above the transition point. However, since both even and odd exist already in conventional amplifiers, there is no cause for concern. As always, the real answer is to reduce the distortion, of whatever order, to so far below the noise floor that it could not possibly be audible and you never need to fret about it.

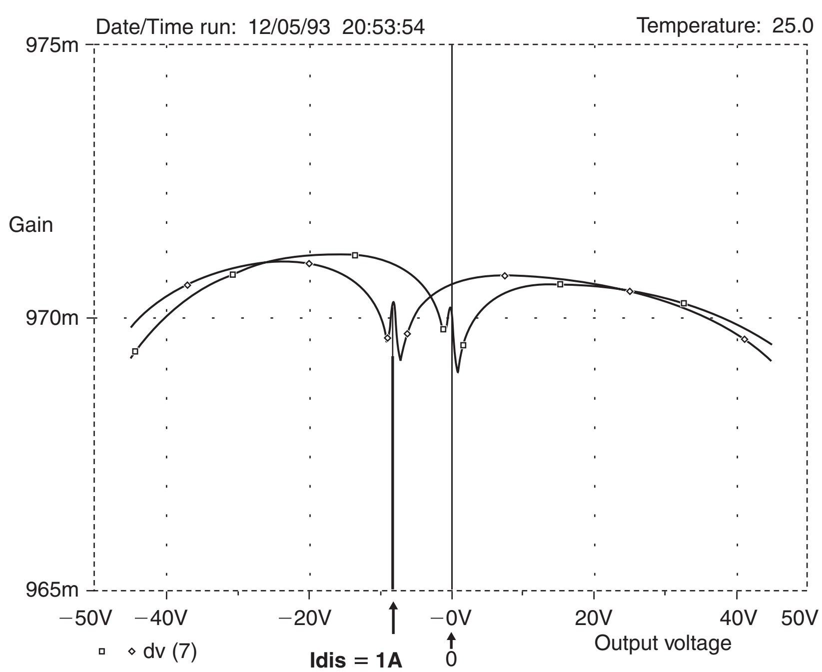


Figure 11.3: SPICE simulation of the output stage gain variation with and without a constant 1 A of displacement current. The central peak is moved left from 0 to −8 V

Crossover Displacement Realization

There are several ways in which a suitable displacement current can be drawn from the main amplifier output node. The simplest method is resistive crossover displacement. Connect a suitable power resistor between the output rail and a supply rail, as shown in Figure 11.4a, and the crossover point will be displaced. In this and all the following examples, the crossover point is displaced negatively by sinking a current into the negative rail.

The resistive method suffers from poor efficiency, as the resistance acts as another load on the amplifier output, effectively in parallel with the normal load. It also threatens ripple-rejection problems as R is connected directly to a supply rail, which in most cases is unregulated and carrying substantial 100 Hz ripple. A regulated supply to the resistor could be used, but this would be relatively expensive and even less efficient due to the voltage drop in the regulator. The resistive system is inefficient because the displacement of the crossover region occurs when the output is negative of ground, but when the output is positive the resistor is still connected and a greater current is drawn from it as the voltage across it increases. This increasing current is of no use in the displacement process and simply results in increased power dissipation in the positive output half-cycles.

This method has the other drawback that the distortion performance of the basic amplifier will be worsened because of the heavier loading it sees, the resistor being connected to ground as far as AC signals are concerned.

A superior solution is constant-current displacement, as shown in Figure 11.4b; here a constant-current source is connected between the output and negative rail. Efficiency is better as no output power is dissipated due to the high dynamic impedance of the current source. The output of the current source does not need to be controlled to very fine limits. Long-term variations in the current

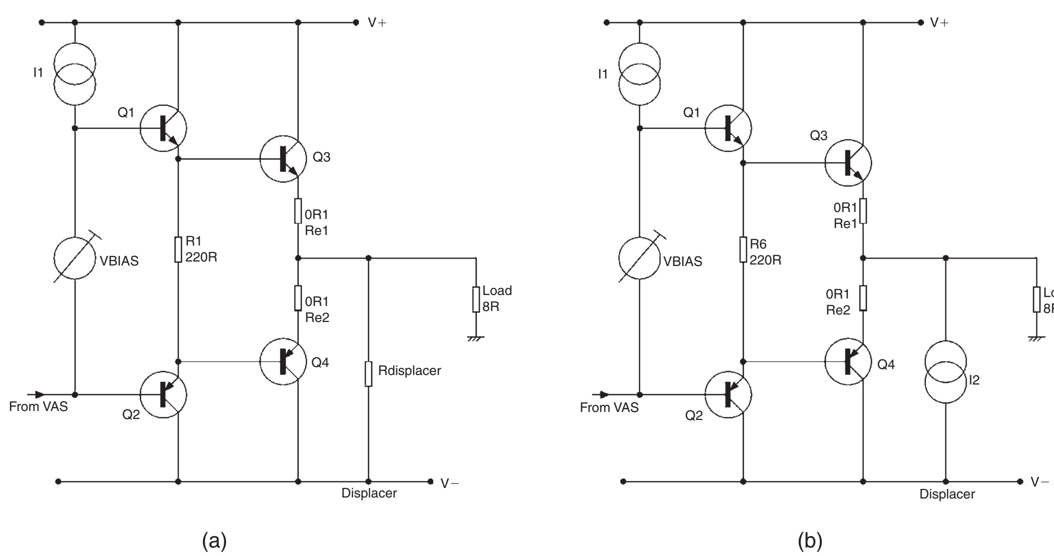


Figure 11.4: (a) The concept of resistive crossover displacement. (b) Constant-current crossover displacement

only affect the degree to which the crossover region is displaced, and this is not a critical parameter. Noise or ripple on the displacement current is greatly attenuated by the very low impedance of the basic power amplifier and its global negative feedback, so sophisticated current-control circuitry is not required. The efficiency of this configuration is greater because the output current of the displacer does not increase as the output moves more positive. The voltage across the current source increases, so its dissipation is still increased, but by a lesser amount than for the resistor. Likewise, the upper output transistor Q3 is passing less current on positive excursions so its power dissipation is less.

Having moved from a simple resistor displacer to a constant-current source, the obvious next step is to move from a constant current to a voltage-controlled current source (VCIS) whose output is modulated by the signal to further improve efficiency. The most straightforward way to do this is to make the displacement current proportional to the output voltage. Thus, if the displacement current is 1 A with the output quiescent at 0 V, it is set to increase to 2 A with the output fully negative, and to reduce to zero with the output fully positive. The displacer current is set by the equation:

$$I_d = I_q\left(1 - \frac{V_{out}}{V_{rail}}\right) \qquad \text{Equation 11.1}$$

where I_q is the quiescent displacement current (i.e. with the output at 0 V) and V_{rail} is the bottom rail voltage, which must be inserted as a negative number to make the arithmetic work. It is not essential for the displacement current to swing from zero to twice the quiescent value; it could be modulated to a lesser extent, and there is in fact a continuum of possible solutions from constant-current displacement to the full push–pull case.

Depending on the design of the VCIS, a scaling factor X is required to drive it correctly (see Figure 11.5). Since a signal polarity inversion is also necessary to get the correct mode of operation, active controlling circuitry is necessary.

The use of push–pull displacement is analogous to the use of push–pull current sources in Class-A amplifiers, where there is a well-known canonical sequence of increasing efficiency, which is fully described in Chapter 10. This begins with a resistive load giving only 12.5% efficiency at full power, moves to a constant-current source with high dynamic impedance giving 25%, and finally to a push–pull controlled current source, giving 50% efficiency. In the push–pull case the sink transistor acts in a sense as a negative resistance, though it is more usefully regarded as a driven source (VCIS) than a pure negative resistance, as the current does not depend on rail voltage. In each of these moves the efficiency doubles. These efficiency figures are ideal, ignoring circuit losses; note that Class-A efficiency is very seriously reduced at output powers less than the maximum. In the same way, there is a canonical sequence of sophistication and efficiency in crossover displacers, though the differences are smaller.

The push–pull displacement approach has another benefit; it also reduces distortion when operating above transition in the Class-B mode. This is because the push–pull system acts to reduce the current swings in the output devices, as the displacement current varies in the correct sense for this. This is equivalent to a decrease in output stage loading; this is the exact inverse of what occurs with resistive

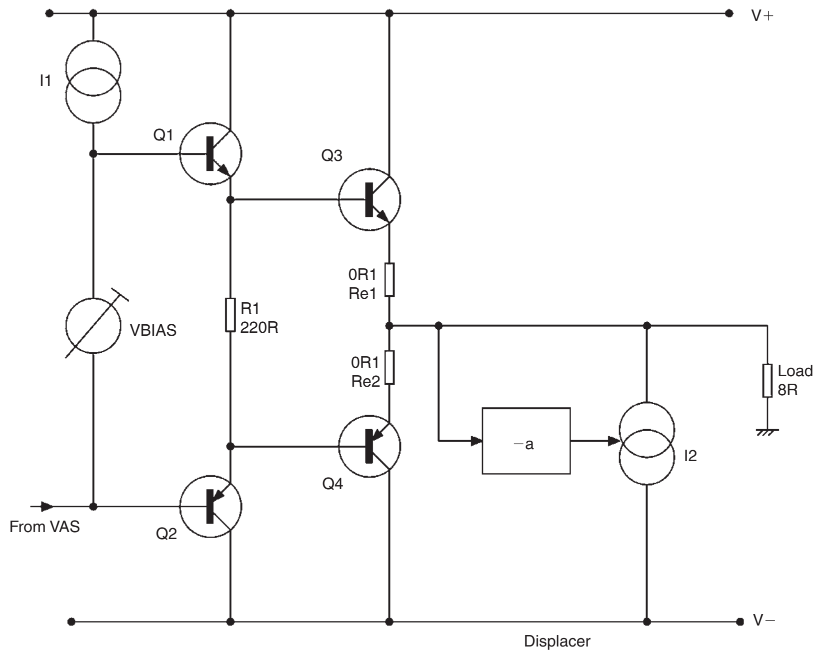


Figure 11.5: The concept of push–pull crossover displacement. The control circuitry implements a scaling factor of −*a* in the signal to the controlled current source

displacement, which increases output loading. Lighter loading is known to make the current crossover between the output devices more gradual, and so reduces the size of the gain wobble that causes crossover distortion; this is described in Chapter 6. In addition the crossover region is spread over more of the output voltage range, so the distortion harmonics generated are lower order and receive more linearization from a negative-feedback factor that falls with frequency. Large-signal nonlinearity (typically experienced with loads of 4 Ω and less) is also somewhat reduced. In push–pull displacement operation, the accuracy of the current variation does not have to be high to get the full reduction of the distortion, because of the low output impedance of the main amplifier, which maintains control of the output voltage. The global feedback around this amplifier is effective in reducing the inherently low output impedance of the output stage in the usual way, being unaffected by the addition of the displacer.

While the constant-current displacement method is simple and effective, the push–pull version of crossover displacement is to be preferred for the best linearity and efficiency; the extra control circuitry required is simple and works at low power so it adds minimally to total amplifier cost.

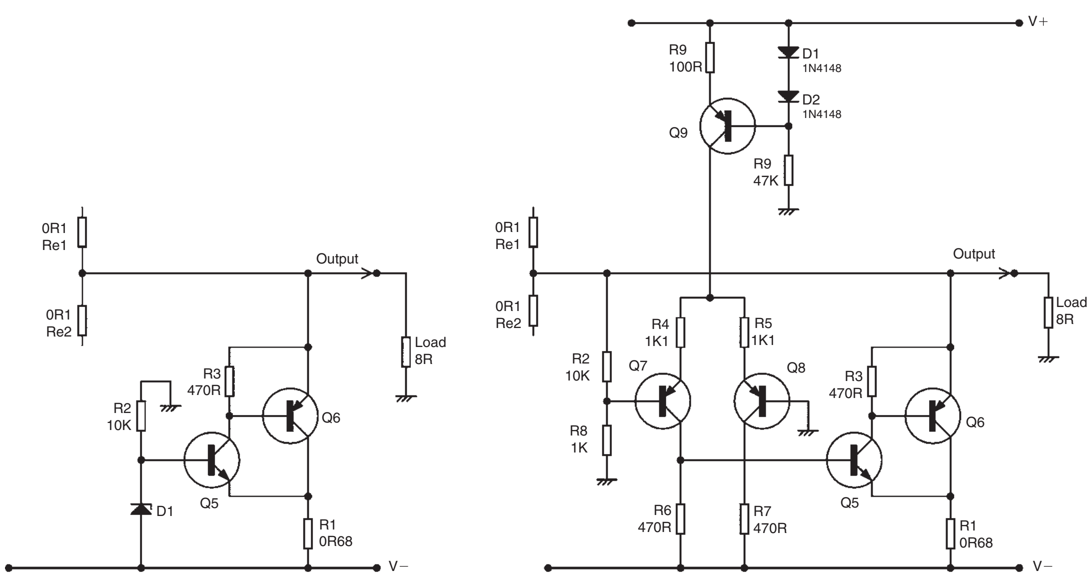
Circuit Techniques for Crossover Displacement

The constant-current displacer is the simplest practical displacement technique, the resistive version being discarded for the reasons given above.

A practical circuit for a constant-current displacer is shown in Figure 11.6a; for clarity the Class-B output stage is omitted. The displacement current typically chosen will be in the region of 0.5–1 A, and therefore a driver transistor Q5 is used, exactly as drivers are used in the main amplifier, so the control circuitry can work at low power levels. The power device Q6 is going to get hot, so its V_{be} must be excluded from having a direct effect on current stability. Therefore the CFP (complementary feedback pair) structure shown is used, so the effect of V_{be} variations is reduced by the negative feedback around the local loop Q5–Q6. The bias for the constant current is shown as a Zener diode D1; if greater accuracy is required a low-voltage reference IC such as the LM385 could be used instead, but there is no real need to do so. The voltage across R1 should not be large enough to limit the output swing; but on the other hand, if it is small compared with the V_{be} of Q5, then the current value may drift excessively with temperature as Q5 warms up.

Power transistor Q6 dissipates significant heat; clearly the greater the crossover displacement required, the greater the displacement current and the greater the dissipation. Q6 is therefore normally mounted on the same heat-sink as the amplifier output devices. This provides the intriguing sight of a power amplifier with an odd number of output transistors, which might conceivably be exploited for marketing purposes.

The push–pull controller drives the displacer so that as the output rail goes positive, the displacer supplies less current. The basic problem is to apply a scaled and inverted version of the output voltage to the displacer. The signal must also have its reference transferred to the negative rail, which can be assumed to carry mains ripple and distorted signal components. Transferring the reference is done by using the high-impedance (like a current source) output from a bipolar transistor collector. As before, a driver transistor Q5 is used to drive the displacer Q6 so the control circuitry can work at low power levels. This not only minimizes total current consumption but also reduces the effect of V_{be} changes due to device heating (see Figure 11.6b).



(a) (b)

Figure 11.6: (a) Constant-current displacer with complementary feedback pair structure. (b) Push–pull displacer with differential pair controller

The controller is simply a differential pair of transistors with one input grounded and the other driven by the main amplifier output voltage, scaled down appropriately by R2, R8. The differential pair has heavy local feedback applied by the addition of the emitter resistors R4, R5, in order to minimize distortion and achieve an accurate gain. The drive to the VCIS displacer is taken from collector load R6, to give the required phase inversion. R7 is present simply to equalize the dissipation in the differential pair transistors to maintain balance.

The tail of the differential pair is fed by the 6 mA constant-current source Q9. This gives good common-mode rejection, which prevents the significant ripple voltages on the supply rails from interfering with the control signal. Since half of the standing current through the differential pair flows through R6, the value of the tail-current source sets the quiescent displacement current. The stability of the current generated by this source therefore sets the stability of the quiescent (no-signal) value of the displacement current. Figure 11.6b shows a simple current source biased by a pair of silicon diodes. This has proven to work well in practice but more sophisticated current sources using negative feedback could be used if greater stability is required. However, even if the tail-current source is perfect, the value of the displacement current still depends on the temperature of Q5. More sophisticated circuitry could be used to remove this dependency; for example, the voltage across R1 could be sensed by an op-amp instead of by Q5. The op-amp used would need to be able to work with a common-mode voltage down to the negative rail, or an extra supply rail would have to be provided.

A further possible refinement is the addition of a safety resistor in the differential pair tail to limit the amount of current flowing in the event of component failure. Such a resistor has no effect on normal operation, but it must be employed with care as its presence may mean that the circuit will not start working until the supply rails have risen to a large fraction of the working value. This is a serious drawback as it is wise to test power amplifiers by slowly raising the rail voltages from zero, and the lower the voltage at which they start working, the safer this procedure is.

A Complete Crossover Displacement Power Amplifier Circuit

Figure 11.7 shows the practical circuit of a push–pull crossover displacement amplifier. The Class-B amplifier is based on the Load-Invariant design and follows the Blameless design philosophy described elsewhere in this book. Conventional dominant-pole compensation is used. The design uses the following robust techniques described in this book to bring the distortion down to the irreducible minimum generated by a Class-B output stage.

1. The local negative feedback in the input differential pair Q1, Q2 is increased by running it at a high collector current, and then defining the stage transconductance and linearizing it by local negative feedback introduced by the emitter resistors R10, R11.
2. The crucial collector-current balance between the two halves of the input differential pair is enforced by the use of a degenerated current-mirror Q3, Q4.
3. The local negative feedback around the voltage-amplifier transistor Q10 is increased by adding the emitter-follower Q11 inside the Miller C_{dom} loop.

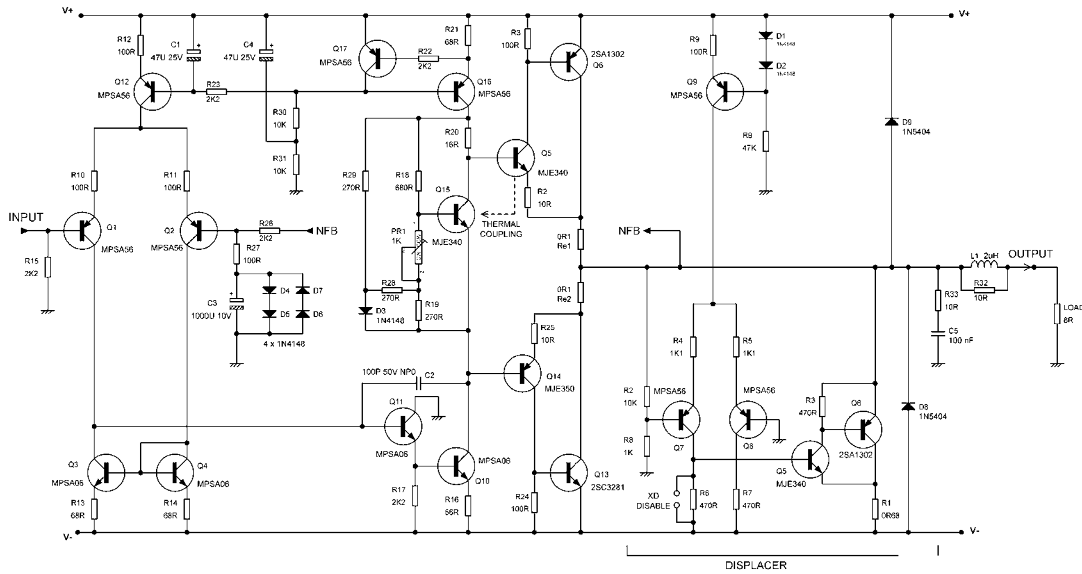


Figure 11.7: Complete circuit of an amplifier using push–pull crossover displacement

4. The output stage uses a complementary feedback pair (CFP) configuration to establish local negative feedback around the output devices. This increases linearity and also minimizes the effect of output junction temperatures on the bias conditions. The bias generator Q15 has its temperature coefficient increased by the addition of D3, R28, R29 to improve the accuracy of the thermal compensation (see Chapter 15 for more details). Q15 is thermally coupled to one of the drivers and preferably mounted on top of it; for this reason Q15 is an MJE340 simply so the packages are the same.

The circuit shown is capable of at least 50 W without modification. Powers above 100 W into 8 Ω will require two paralleled power transistors in the main amplifier output stage. The displacer transistor does not necessarily require doubling; it depends on the degree of crossover displacement desired.

The displacer control circuitry is essentially the same as in Figure 11.6b. A push-on link can be connected across R6 so that the crossover displacement action can be manually disabled to simplify testing and fault-finding.

Note that overload protection circuitry has been omitted from the diagram for simplicity.

The Measured Performance

The measurements shown here demonstrate how crossover displacement not only deals with crossover distortion, but also reduces distortion in general when the push–pull variant is employed. Tests were done with an amplifier similar to that shown in Figure 11.7.

Figure 11.8 shows THD versus frequency for a standard Blameless Class-B amplifier giving 30 W into 8 Ω. The distortion shown only emerges from the noise floor at 2 kHz, and is here wholly due to crossover artefacts; the bias is optimal and this is essentially as good as Class-B gets. The distortion