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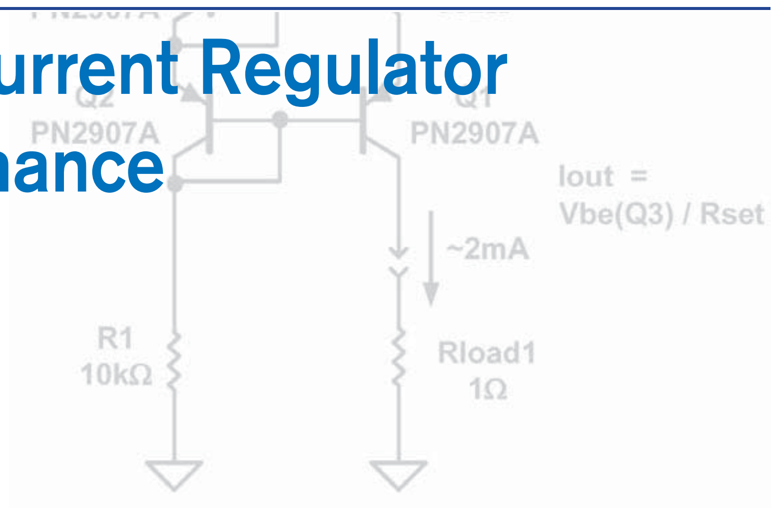
Sources 101: Audio Current Regulator Tests for High Performance

Part 1: Basics of Operation

Noted designer and author of a classic op-amp design cookbook begins his series on quieter power supplies.



By Walt Jung

This article deals with audio current regulator circuits and their performance with regard to power-supply-related noise. It describes a test methodology to characterize these audio current regulation elements and circuits for sensitivity to applied voltage. In terms of audio power supply circuitry within systems, this would be the *line rejection* or *impedance properties*. Because audio current regulators are often referred to as *current sources*, and the performance aspects of such for audio are basic, the article is logically entitled "Sources 101."

But, it isn't the last word on the topic by any means. I hope it can explain to many how to build better current regulator circuits for audio. And, most important, it definitely shows how to test these circuits, and to differentiate their performance.

THE WHYS AND WHEREFORES

While I have been fascinated by current regulators for years, it wasn't until recently that I investigated deeply the wideband AC performance of various topologies operating at many current levels. It was brought home to me that some greater insight could be useful, when a relatively simple "One Vbe" type of current regulator was found lacking, as operated within a shunt regulator design. That experience started this investigation several months ago, and the article here is one fruit of those explorations.

So, there we are. It may seem obvious that it is desirable to have audio circuits with high immunity to power supply potentials, but this area is seldom, if ever, addressed in audio design and construction articles. Typically, a current source or sink circuit is presented without related performance data, so it is difficult to differentiate what types of designs are better for audio. Some relatively simple power-supply rejection tests can help differentiate which circuits perform best and are most immune to power-supply-related distortions. This article details a wide variety of available current regulation circuitry, and documents their relative performance in terms of power supply immunity.

CURRENT REGULATOR PROPERTIES FOR AUDIO

Current regulators have several operating parameters that are key to audio circuit performance. Probably the most important is the *degree of regulation* of the associated current. In other words, the output current, Iout, is maintained at some fixed design level, one which is relatively independent of other circuit conditions. Thus, although the unregulated input voltage varies, the design current stays constant.

As noted, just how well this is done is one of the major figures of merit for a current regulator. You can detail the specification for this in several ways. One is simply the rejection (or line rejection) of noise appearing on the input, expressed most simply in dB, such as, for example, a rejection of 100dB. This means that input noise is reduced by 100dB (100,000/1), or it becomes −100dB with respect to the input (1/100,000).

Because audio is a wideband signal, it is important to quantify a regulator over at least the 20Hz–20kHz audio bandwidth, and preferably even wider. This is because these circuits often have a nonlinearity that can be excited with supersonic signals. This particular performance parameter is also expressed as *impedance*, which I will discuss further.

One form of nonlinearity often found in solid-state audio circuits is nonlinear capacitance. While a pure capacitance is not a distortion producer within an audio circuit, nonlinear capacitance can—and will—produce distortion, particularly when excited with high frequency (HF) noises, such as those typical to rectified-AC supply systems. So, one indirect figure of merit for audio current regulators is the associated capacitance. The lower this is, generally the lower will be any spurious responses to HF noise.

CURRENT *SOURCE* OR CURRENT *SINK*?

One point of potential confusion regarding current regulators for use within audio circuits lies with the terminology used. Two terms you often see are *current source* and *current sink*. These terms are often used interchangeably, but this isn't always technically correct. Some review of the terminology is helpful.

A current *source* circuit is a device (or more complex circuit) that provides current regulation properties, typically operating between a positive rail voltage and common (ground), often using one or more PNP transistors, so as to *source* load current. This is the more popular usage. However, you should note that current regulator circuits operating between a negative rail and common are also sometimes called current "sources."

Regardless of this definition muddling, it is more accurate to describe such circuits as current *sinks*; that is, a current regulator biased with respect to the negative rail, often using one or more NPN transistors to *sink* load current. This is the terminology I will use in this article; i.e., a *current source* operates between a positive rail and the load or common, and a *current sink* operates between a negative rail and the load or common.

To add slightly to the confusion, there are single transistor devices that, because of their unique bias flexibility, can operate as either a current source or sink. An example would be the *common junction field effect transistor*, or JFET. So, in applying such devices, their explicit connection details will determine their exact function. More specifically, N-channel JFET devices such as the J202 and others exhibit this type of flexibility, and will be illustrated shortly.

WHAT TESTS?

While there are many tests potentially useful for audio characterization, this exercise concentrates on power-supply rejection versus frequency, which yields a picture of current regulator impedance versus frequency. These two performances go hand-in-hand. For best immunity to power rail noise components, a wide rejection bandwidth is desirable within the audio circuits. This may not be immediately apparent, because unregulated audio rails usually have predominant 120 or 100Hz ripple, from which it is relatively easy to provide immunity.

A salient point here is that the AC to DC rectification process is by definition a wideband noise generator, by chopping the AC mains waveform into high peak current pulses in the typically used capacitor input filter. A Fourier analysis of the noise components will show there are ample HF components associated with such supply rails, not just the 100/120Hz fundamentals.

Another subtle point is that the audio circuits themselves don't have infinite bandwidth. Thus, while they may have some degree of good supply rejection at lower audio frequencies, in the ultrasonic range this rejection deteriorates, introducing the potential for power noise components to *intermodulate* with the audio. Of course, for high quality reproduction, the possibility of any such intermod must be minimized, either by careful filtering or the development of circuits intrinsically immune to the nonlinearities that can produce the intermodulation.

TEST SETUP

The general setup for this test series is as per Figure 2a of Reference 1, available at www.waltjung.org/PDFs/Regs_ for_High_Perf_Audio_1.pdf. The test is basically a highly sensitive crosstalk measurement done in the analog domain, measuring the output of a circuit as it is related to a 20Hz-200kHz swept sine wave excitation. This method was originally developed for the 1995 series of articles on audio regulators[1].

In this current series of tests, the "regulator" under test is either a current source or sink, which is connected as per **Fig. 1**. Here, the device or circuit under test is a two-terminal (in some cases, three) current source device (or circuit) connected between Vin and the load. The load is simply a 1Ω resistor, Rload1. Note that unlike Fig. 2a of Ref. 1, no output load capacitor is used for this test. The rail driver circuit supplying the +18V DC excitation with a superimposed 1V RMS AC swept frequency signal is shown in Fig. 6 of Ref. 1. It is represented here as the dashed box figure.

Calibration of this test setup is done by first calibrating the Audio Precision test set to the 1V RMS 0dB level, against the applied rail driver Vin(AC) signal. All subsequent measurements are then referenced to this 0dB level. The calibration is completed with a series of test runs, using fixed 1% calibration resistors in the Device Under Test (DUT) position. This step establishes corresponding reference Vout levels of 10kΩ, 100kΩ, and 1MΩ, as shown in the like-named plots of **Fig. 2**. On

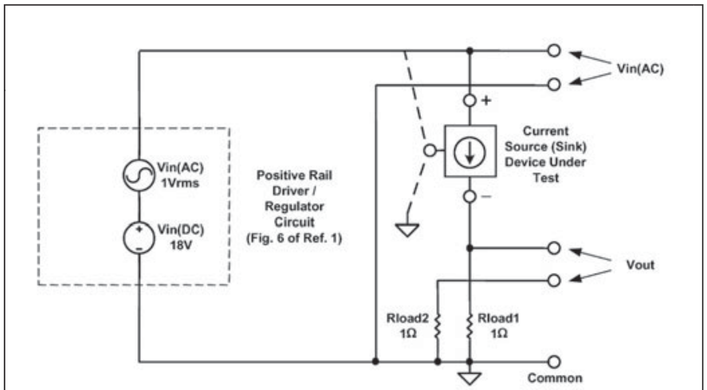


FIGURE 1: Current source/sink DC/AC test circuit.

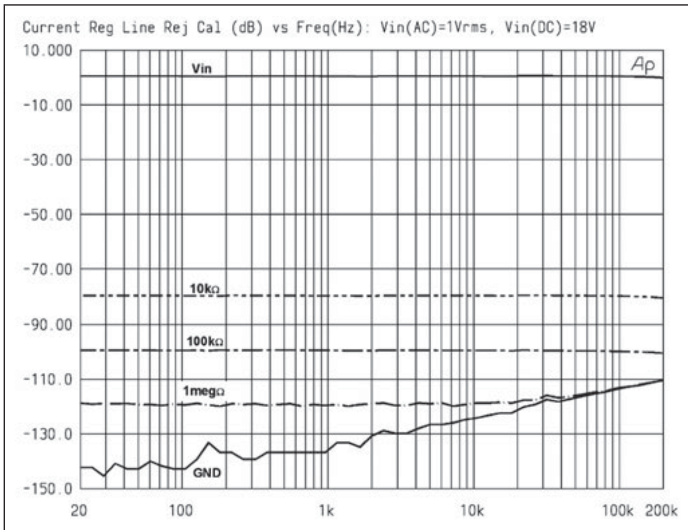


FIGURE 2: Multiple calibration runs with fixed 1% 10kΩ, 100kΩ, and 1MΩ resistors establish measurement validity. The dynamic range limits correspond to Vin and GND.

the dB scale, these correspond to levels of –80dB, –100dB, and –120dB, respectively.

In this figure there is also shown a much lower level trace, marked GND. This trace is the lower limit of the test setup dynamic range, which is the residual noise seen at Vout of **Fig. 1** with the test set active and no DUT connected. In essence, virtually all practical devices/circuits tested show appreciably higher levels at Vout, although some do indeed begin to approach the full dynamic range of more than 140dB.

As you can also note in **Fig. 2**, the plot for the highest fixed calibration impedance of 1MΩ falls about 20dB (or more) higher than the residual noise shown in the GND trace, at the lowest frequencies. At the higher frequencies, setup/system noise shows impedance lowering, corresponding to a dynamic range of ~110dB at 200kHz, which is roughly equivalent to 300kΩ.

THE MEASURED NOISE

The nature of this test is to measure, first of all, *synchronous noise*. This is because it is a modified crosstalk test, and operates by sweeping a measurement bandpass filter, looking at the spectrum appearing across Rload1. The measured noise, however, can be from either of two sources. The obvious one is due to the synchronous components related to Vin; i.e., the "crosstalk" noise components[1] (see "Hofer" box, p. 4).

But, another potential component could be due to any *self-generated* noise in the DUT. The test as configured here really has no means to distinguish one noise from the other—they are simply lumped together. It is true, however, that if and when a very low noise level is measured, both the synchronous as well as the self-generated noises must be low. And, in the many tests that follow, many circuits show very low noise—meaning that they have at or approaching levels of –140dB with respect to 1V RMS; in other words, on the order of 100nV RMS. A potential future investigation might examine self-generated noise more closely, with a higher load impedance and noise analysis software.

WHICH DO YOU PREFER: IMPEDANCE OR NOISE REJECTION?

You should understand that current regulator circuits may be specified in terms of either equivalent dynamic impedance; i.e., "100kΩ," or rejection with respect to some applied voltage reference level. Here, the three calibration impedances equate to rejections of 80dB (10kΩ), 100dB (100kΩ), and 120dB (1MΩ).

For impedances (Z) of more than 10kΩ (rejection of more than 80dB), you can use one of these approximations to convert between the two terminologies.

$$dB \sim 20 * \log(1/Z)$$
or
$$Z \sim 10^{(dB/20)}$$

In the test plots following, the current source or sink can be either a simple device such as a JFET, an IC such as a three-terminal regulator operated as a current source, or a more complex circuit comprised of transistors, resistors, and so on.

Referring to **Fig. 1** again, note that the measurement point of Rload1 is fixed with respect to the common point. For current *source* circuits, this is most appropriate, and the output current is measured directly and accurately. For some current *sink* circuits, a more appropriate current sample point would appear to be with Rload to the (+) lead of the DUT. This test setup doesn't provide for this, but nevertheless current sink type circuits can still be assessed. In the actual measurements that follow, this detail will become somewhat moot.

THE MEASUREMENTS

I measured a wide variety of current regulator circuits, all of which were built up on 8 pin header assemblies and plugged into the **Fig. 1** setup as a DUT. In the performance plots which follow, the Vin(AC) trace is shown at the top, and the GND trace at the bottom. The actual measurement data for a given DUT is contained in one (or more) Vout traces, which fall somewhere between the upper/lower dynamic range extremes. For cases of multiple Vout traces representing different conditions, each is labeled for clarity.

One Vbe Current Source

A "One Vbe" current source is shown in **Fig. 3A**, based on PNP transistors. In this and many circuits that follow, the transistors used are the high gain, TO-92 versions of general-purpose industry standards, either the 2907A series for PNPs, or 2222A series for NPNs. For general purpose work, these parts are preferable, since they represent a sweet spot of performance for these applications, that is excellent linearity for good rejection at lower frequencies, yet still

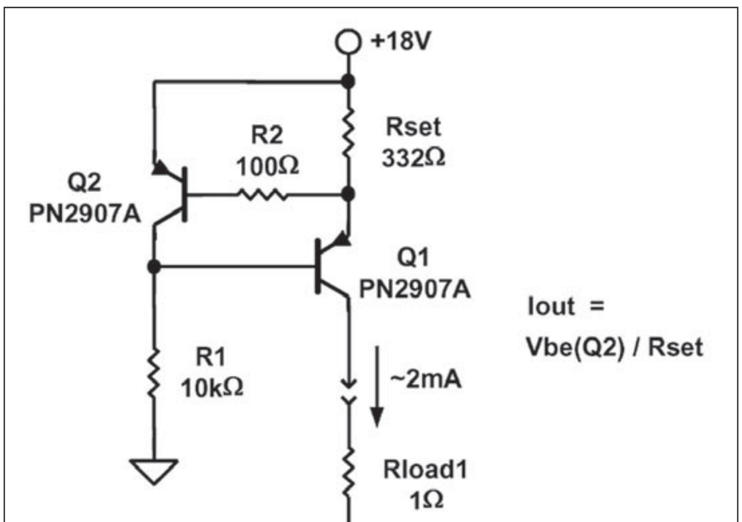


FIGURE 3A: One Vbe current source.

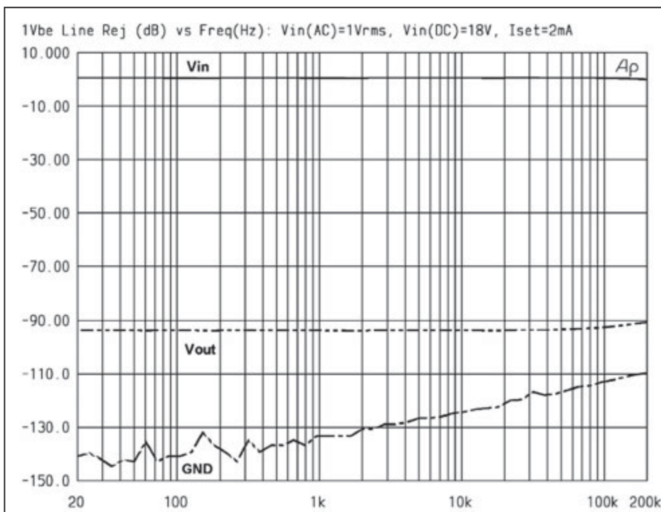


FIGURE 3B: The performance of the *Fig. 3A* "One Vbe" current source shows a rejection characteristic of about 93dB, corresponding to a 45kΩ impedance.

Bruce Hofer, Chaiman and CoFounder of Audio Precision Inc., offered these comments on the tests:

"I would like to note that System One is now an obsolete product having been replaced in the mid 90s with our System Two, which itself has gone through several cycles of evolution. The technique of displaying noise spectrally using a swept 1/3-octave filter is still valid, but some engineers today would probably prefer to look at noise using the FFT. I should also note that analyzer analog input noise has improved (dropped) somewhat since System One, but only slightly. Indeed, many of our competitors today have yet to equal the performance of our original System One! Thus I think your graphs are still quite relevant."

low-to-moderate capacitance, which allows the good performance to hold up well with increasing frequency, at great prices! They allow currents from low µA levels up to 20mA or more, at voltage levels of 40V (or more).

Of course, higher-voltage parts should be used when appropriate. While exotic and super-high-gain parts aren't necessary for very good performance from these circuits, low capacitance devices definitely are preferred (<10pF), a critical point if substituting. For truly excellent performance (at higher cost and reduced availability), you can use select "2S" series parts. Notable examples here are the Sanyo 2SA1016 PNP and 2SC2362 NPN (see www.semiconductor-sanyo.com/discrete/index.htm). These offer lower capacitance than the 2907/2222 families, and are useful up to 120V or more. They can work in the circuits shown, substituting for the PN2907A and PN2222A, respectively.

The familiar circuit of **Fig. 3A** is often used in audio circuits, perhaps due to the relative simplicity. It is configurable over a wide range of output current levels by adjusting the sensing resistor, Rset, which drops one Vbe (that of Q2) in operation. As shown, the output current is about 2mA.

The PN2907A types are useful up to medium voltages and several tens of mA. Q1 may require a heatsink for power dissipation of 0.5W or more. You can make a mirror-image current sink circuit with NPN transistors such as PN2222As for Q1-Q2, referring Rset to a negative supply. Note: This circuit can oscillate readily when using wideband transistors. The tendency to do so is suppressed by R2, which should be used with any form.

Performance of the **Fig. 3A** current source is shown in **Fig. 3B**, a plot of Vout for the cited test conditions. While the impedance offered by this circuit is rather independent of frequency, being flat to nearly 200kHz, it really isn't all that high. The rejection of ~93dB corresponds to an impedance of about 45kΩ.

While at first glance this might appear OK, the performance of this circuit is easily bettered by many others, both more simple and cheaper. For these reasons, plus the stability caveat, this circuit isn't recommended. Except as an example to avoid, perhaps!

Two Diode Current Source

An effective and still relatively simple current source is shown in **Fig. 4A**, again based on PNP transistors. With Q2 and Q3 connected as diodes, this is a two-diode-biased current source. It is often seen with two 1N914 or other diodes functioning just as Q2-Q3 do here. While it does use three transistors, it can still be inexpensive, because standard types such as these PN2907As are less expensive than metal film resistors. Output current is set by Rset and the Vbe of Q3, and is 2mA. As was true with the One Vbe current source, you can implement a current sink with the use of PN2222A NPN transistors.

Performance of the **Fig. 4A** current source is shown in **Fig. 4B**, where the impedance is again independent of frequency. Here the rejection of 100dB corresponds to an impedance of 100kΩ. This circuit is useful for moderate performance at low cost, at the expense of requiring five parts. Or, you could maximize efficiency by using the dual (or quad) packaged 2907A transistor types.

LED Current Sources

Replacing the two diodes of **Fig. 4A** with a single green LED forms the LED current source of **Fig. 5A**. This setup is slightly less complex and inexpensive, and provides good performance for the cost/complexity. Output current is set by the value of Rset, which has about 1.2V across it. The slightly higher voltage of a green versus red LED

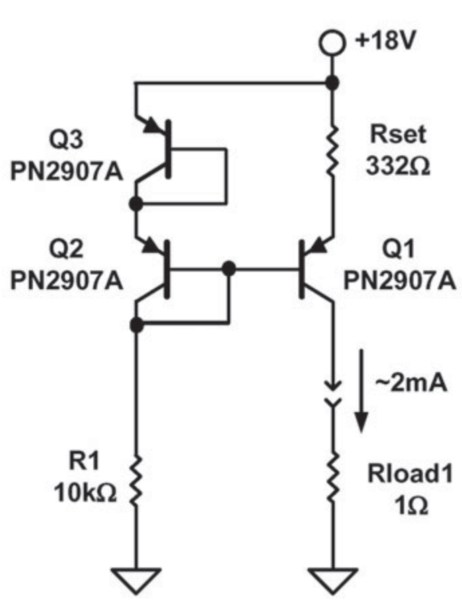


FIGURE 4A: The "Two Diode" current source. Q2-Q3 are diode-connected transistors, but can also be discrete diodes (1N914, and so on).

Iout = Vbe(Q3) / Rset

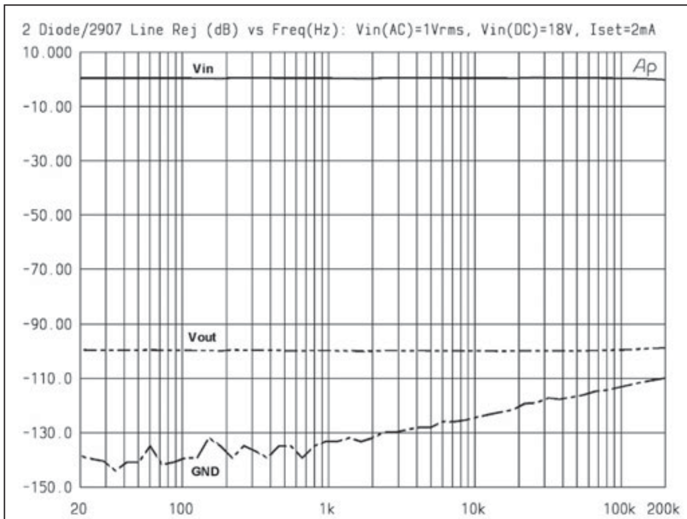


FIGURE 4B: The performance of the *Fig. 4A* Two-Diode current source shows a rejection characteristic of about 100dB, corresponding to a 100kΩ impedance.

provides the 1.2V, but you could also use a red LED. A current sink is also possible, using NPN transistors biased to a negative voltage, plus, of course, an appropriate polarity change for the LED.

Performance of this LED current source is shown in **Fig. 5B**, operating with an output current of 2mA. The rejection is about 105dB, corresponding to an impedance of 177kΩ. This circuit actually achieves better performance than the "two-diode" current source, but with fewer parts!

Reference Diode Current Source

If you replace the LED (with the relatively poor reference voltage) from **Fig. 5A** with a true voltage reference diode, a much higher quality current source is formed (**Fig. 6A**). Note that this form of the circuit has the same number of basic parts as the simple LED source, but is capable of much higher performance. This is due largely to the more stable voltage across the diode, which changes very little with current. An optional cascode connection for Q1 enhances performance considerably.

In this example an LM336Z-2.5V diode is used, resulting in about 1.8-1.9V across Rset. Thus a 1kΩ value for Rset supplies just under 2mA of output current. Note that the circuit is by no means limited to just such lower currents. As long as the power dissipation of Q1 is maintained low (or sufficient heatsinking used), currents up to 10mA or more are allowable from this basic circuit.

Figure 6B shows the performance of this simple current source in two modes, one basic, the other with Q1 operated with the optional cascode. In both cases the resulting performance is excellent. For the basic operation using a simply connected PN2907A for Q1 as shown at the left, the low frequency (LF) rejection approaches 130dB, which would be equivalent to 3MΩ. There is a slight rise in impedance at the upper frequencies, but all in all, the performance is exceptional for such a simple circuit.

When the optional cascode connection shown at the right is used, both LF and HF performance is enhanced. The cascode data, as noted, approaches the residual noise at all frequencies. It is enabled simply by using the optional connection for Q1/J1, which is, in turn, used as Q1 in the main circuit.

Note that this rather elegant connection is *self-biasing*, due to the manner in which the Vgs of J1 automatically provides a collector voltage for Q1. It will do so as long as the Idss of J1 is equal to or greater than

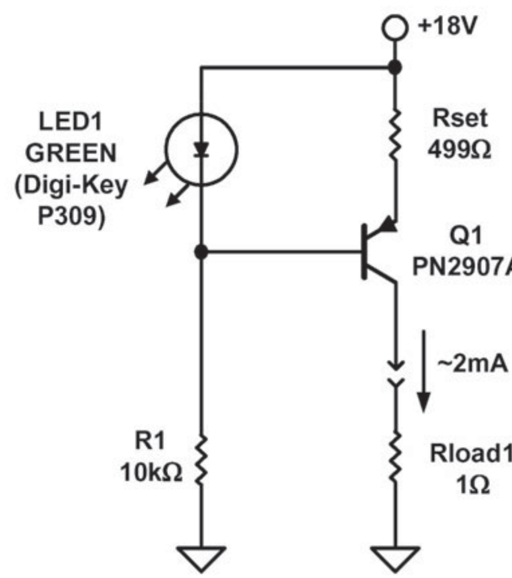


FIGURE 5A: The LED current source. LED1 is here a green type.

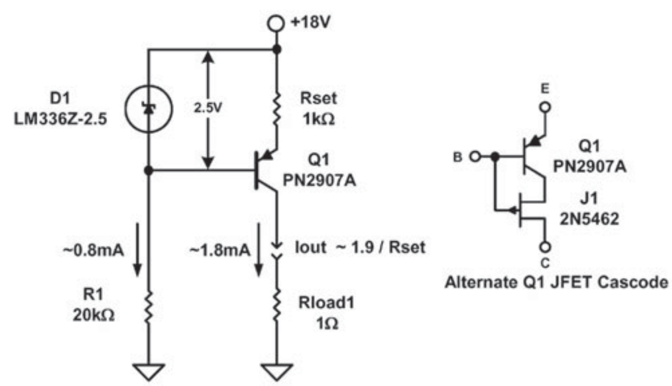


FIGURE 6A: 2.5V reference diode current source (basic, left) with alternate Q1 cascode (right).

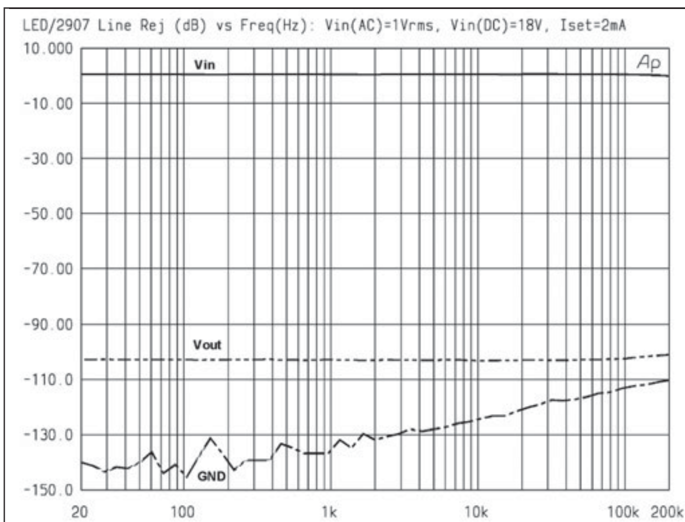


FIGURE 5B: The performance of the *Fig. 5A* LED based current source shows a rejection characteristic of 105dB, corresponding to a 177kΩ impedance.

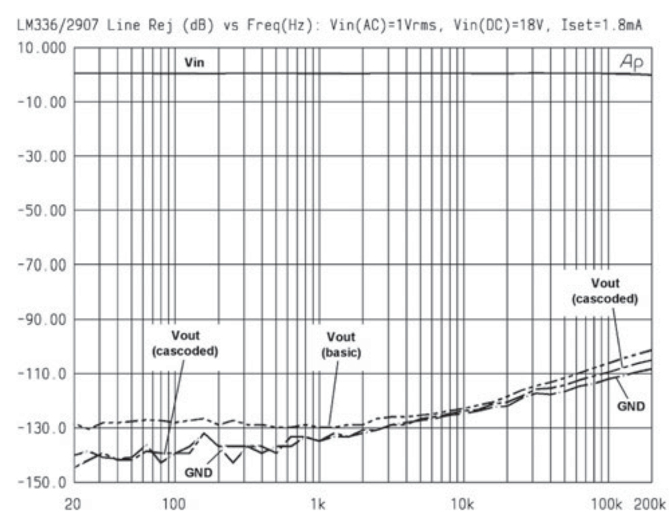


FIGURE 6B: The *Fig. 6A* reference diode current source performance for both basic and cascode operating modes. Cascoding greatly improves low frequency rejection.

the desired output current. For example, the 2N5462 specification for Idss is 4mA(min), which means that this cascode should only be used with lower currents, such as this 1.8mA case.

Alternately, for higher currents, either 2N5462s can be screened for a higher Idss, so that the tested Idss of J1 is always equal to or more than the desired Iout. Or, a basically higher Idss family of P-channel FETs can be used, such as the 2SJ74 V series (being careful to note that these are limited to 25V supplies). But, the 2N5460/5462 series is preferred, both from the standpoint of the low capacitance they offer, ≤10pF, plus their 40V voltage rating. The low noise 2SJ74 audio types have much higher capacitance than 10pF and don't perform as well as HF current sources (although they do excellent below 1kHz).

With a cascode connection such as this, you can expect a loss of output swing, due to the gate-source bias voltage of J1. Take care in testing at higher output swings to ensure that this is not a problem. The circuit of **Fig. 9A** following also discusses cascodes, at higher output currents.

Although the D1 reference diode used in **Fig. 6A** has low TC by itself, as it is applied here the net output current will still change with temperature, following a positive slope due to the Vbe of Q1. Another version of this circuit can also be built, using a 1.2V diode for D1. With this option, the output current will also have a positive slope of about 0.3%/ °C; i.e., the opposite of that of a conventional silicon diode. Although none of the current sources/sinks discussed thus far have low inherent TC, several examples to follow do in fact have both low TC and a predictable output current.

LM334 Current Sources and Sinks

The LM334 is a monolithic IC designed to be used as a current source (or sink), or, alternately, as a temperature transducer. It is shown in **Fig. 7A** (left), used as a basic 2mA current source, requiring only one additional part for functionality, Rset. The device features very simple operation and can be used up to about 5mA of current as shown. It is perhaps one of the more predictable types for output current among the circuits described thus far.

But, note that the LM334 output current is *not* constant with temperature; it does, in fact, vary linearly with a positive slope of about 0.33%/°K. The basic expression for output current is shown in the figure. Note that the very low sense voltage implies efficient utilization of power supply voltage. The part achieves operation at low thresholds, requiring only ~1V across the terminals to operate. Just as shown here, it is configured as a current source. Operation as a current sink is possible with the same number of parts, but with the V- terminal and Rset tied to a negative supply, with a load connected to V+.

Rejection versus frequency performance of the LM334 as both a basic and a cascode current source is shown in **Fig. 7B**. As will be appreciated, these impedance characteristics are exceptional, particularly at the lower frequencies where rejection is a few dB above the setup residual noise, even for the basic curve. This is indicative of an equivalent impedance close to 10MΩ, albeit with deterioration at the higher frequencies.

An improved version is available by adding a cascode-connected JFET, shown as the **Fig. 7A** option at the right. With this version, J1 is preferably a 2N5486 JFET, which is suitable for output currents up to 2mA. There are no other changes to the circuit. The beauty of this cascode connection is that it extends both low and HF impedance, and the resulting composite performance is barely above the residual noise level. The circuit can still be used either as the current source shown or as a current sink, with the load in the drain lead of the JFET.

However, it is important to note that selecting the JFET device for the cascode position really isn't trivial, but requires some careful thought. The 2N5486 series is quite useful because of the low capacitance and medium currents, *but it does have the disadvantage of only a 25V voltage rating*, lower than the LM334's 40V rating. Alternatively, the PN4393 family can operate to 40V, but it does have higher capacitance than the 2N5486 family of RF devices.

Other JFET parts can also work as J2. The basic requirement is that the desired output current be safely less than the minimum device Idss rating, and the JFET Vgs at the operating current is safely greater than the LM334 minimum voltage of 1.2V (~2V recommended). Some general considerations for the JFET selection are included on the LM334 datasheet[2], and in references 3 and 4. This process is also discussed in more detail with the JFET-based current sources to follow.

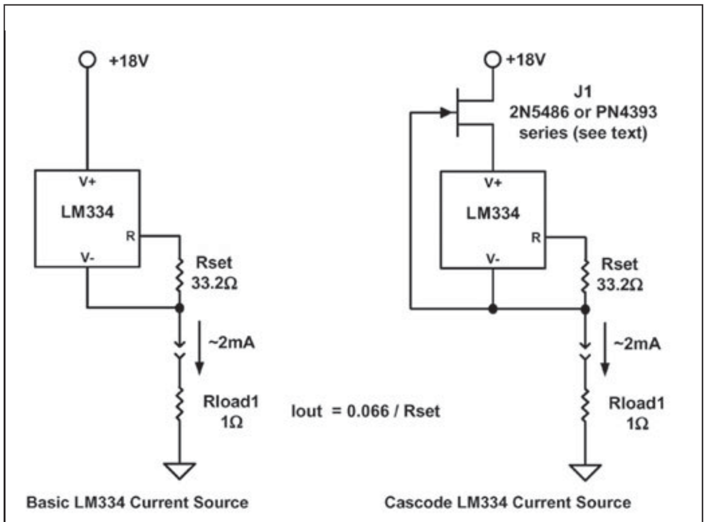


FIGURE 7A: LM334 current source (basic, left) with alternate cascode (right).

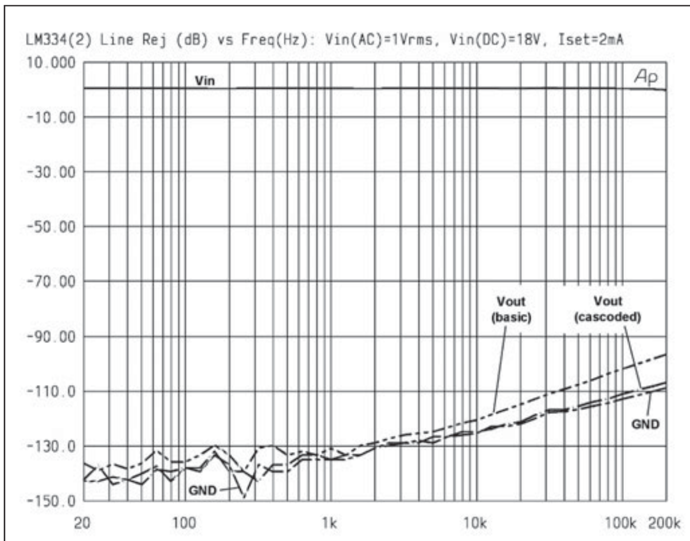


FIGURE 7B: The *Fig. 7A* LM334 current source performance for basic and cascode operating modes. Cascoding improves both low and high frequency rejection.

Some LM334 caveats: As typically operated, an LM334 produces noise components higher than that of a simple bipolar transistor at the same current. This is fundamental to the device design, so take this factor into account before application in low level circuits (see datasheet discussions). It will also have a higher output capacitance than many small signal transistors—about 15pF. Fortunately, this latter characteristic can be mitigated with the use of cascoding, providing that the cascode transistor used has low capacitance.

JFET-Based Current Sources and Sinks

It is fairly well known that many JFET devices make natural current limiters, and thus they can be used as either current sources or sinks. For an N-channel part (the most popular) all you need to do is short the gate and source terminals, apply positive bias to the drain terminal, and connect the gate/source to the load. Under these conditions the JFET will conduct a current equal to the device's Idss (drain current with gate/source common). Once the applied voltage is greater than a minimum voltage equal to Vgs(off), this current will then remain relatively constant with further voltage increases; i.e., the device is operating in a *current-limited* mode.

What is not likely to be as well known are the details appropriate to selecting a JFET part so that this supply voltage independence is maximized. For the testing described here, this implies the highest rejection characteristic. This is a function of both the specific JFET part itself, and how it is used.

A case in point is the simple JFET current source of **Fig. 8A**, using the J202 N-channel FET. While many different FETs can be used in this circuit, the J202 series is well suited to this application, because the rejection characteristics are optimum for the J201 and J202, which are 40V rated devices. They come from a family made with a *long gate* process, which minimizes the device output conductance, and thus maximizes the rejection. *An important note: References 3 and 4 cover this area quite well, and should be considered required reading for anyone building these types of circuits.*

Within all the various JFET process families, there are typically several classifications of devices, sorted as to Idss and Vgs(off) limits. The lower Vgs(off) parts in a given family will have the lowest output conductance, and thus the best rejection. For this family, this would be the J201, which has a max Vgs(off) of 1.5V. From Reference 3, for best rejection, the J1 Vdg > Vgs(off), preferably 2 * Vgs(off). Practically speaking, this means that such FET circuits like lots of voltage to achieve their best performance.

But, it isn't always possible to use such a low Vgs(off) part, as in this case I needed 1mA, which is well above the J201's minimum Idss of 0.2mA. So I used the J202, which, fortunately, still gave great results.

Although the most simple circuit of this type would use an R1 value of zero, I set up these tests with R1 = 100Ω to simplify selection of a device sample to conduct 1mA, the target output current, which occurs with 0.1V across R1. This device was then used in the rejection tests. It actually helps the circuit a bit to use some finite resistance for R1, because this resistance increases the net output impedance slightly. Consider it optional in your final circuit. Store-bought JFET current limiters simply short the gate-source terminals and sell the so-wired part as a two-terminal device (equivalent to the J1 part of the **Fig. 8A** circuit with R1 shorted).

Performance of the J202 as both a basic and cascode JFET current source is shown in **Fig. 8B**, and as you can note, even the basic circuit is excellent, considering the simplicity. The rejection is about 113dB, corresponding to roughly a 446kΩ impedance. While this can be considered very good for such a simple circuit, this is really due to the optimum conditions. There is first the optimum type of device (the J202), then there is also the relatively high bias voltage of 18V, well above the 2 * Vgs(off) rule of thumb. Circuits which bias J1 at lower potentials would be expected to perform more poorly, particularly when Vdg approaches Vgs(off).

While the J202 current source in basic mode is excellent for the simplicity, adding a cascode device makes the rejection characteristics approach the measurement limits. A cascode version of this 1mA current source is shown within the right option of **Fig. 8A**. These measurements used a 2N5486 for J2, and the cascode mode data of **Fig. 8B** is barely discernible from the residual noise.

But, there are more caveats to remember about this cascode circuit. The cascode device must be selected to provide a bias for J1 that is above J1's Vgs(off), preferably well above. For the sample devices used here J2 applied

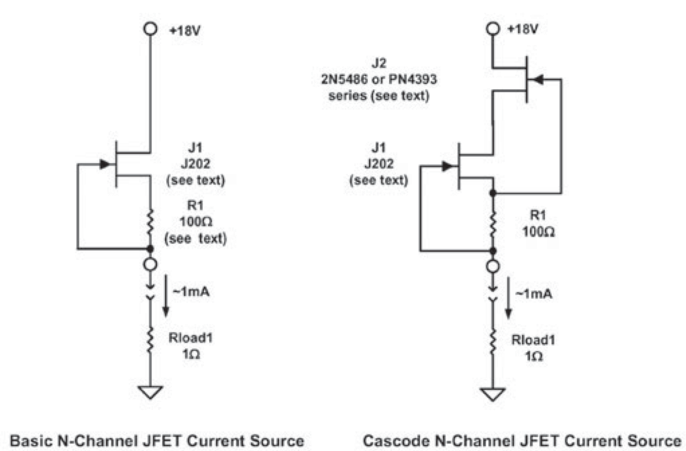


FIGURE 8A: N-channel JFET current source (basic, left) with alternate cascode (right).

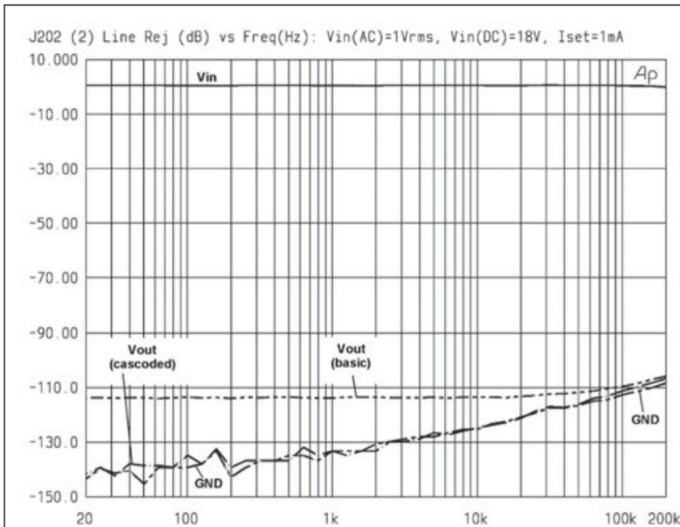


FIGURE 8B: The *Fig. 8A* N-channel JFET current source performance for basic and cascode J202 operation. Cascoding greatly improves low frequency rejection.

a 2.7V bias to J1, indicating that the sample J202 was a low Vgs(off) device. Furthermore, the 2 * Vgs(off) rule should also be applied to J2, and will be more stringent for that part, because it is by necessity a higher Vgs(off) part than is J1. The comments about deterioration of rejection with lower rail voltages apply even more so to this cascoded version, and may require supplies of 12V minimum, for example, for the very best performance.

So, in essence what you have here is a circuit that can work really well, but also exhibits some subtlety to extracting maximum performance. In fact, some J1/J2 preselection may be appropriate to set things up optimally. Fortunately, it appears that once the J1/J2 DC biasing conditions are properly met, the good AC performance falls in place. Watch the voltages on J1 and J2 and use higher voltage parts when necessary for the cascode. A caveat here applies to the 25V 2N5486, as noted.

Finally, I reiterate that both the basic and the cascoded versions of this current regulator can operate as either a source (as shown) or as a sink, with the JFET most negative terminal tied to the negative supply and the load in the positive leg. This is one intrinsic beauty of all JFET-based current regulators; that is, the ability to operate as either a source or a sink, without performance compromise. Note a somewhat subtle minor point here: you need only consider N-channel JFETs for such two-terminal current regulators; not only do they work (well), there are many more of them to choose from than the P-channel counterparts.

JFET Current Regulator Diode Based Current Sources and Sinks

Most of these discussions apply equally to JFET *current regulator diodes* (also called JFET current limiters), because they are JFETs internally wired as two-terminal parts. One series of these is the Vishay/Siliconix J500 family, consisting of individual parts with output current ratings from 0.24mA (J500) to 4.7mA (J511), all rated for 50V operation[5]. Of this series I had some J507s on hand, rated for a nominal 1.8mA operation, and some of these were tested.

Applying this type of current regulator is simplicity itself, as shown in the simple two-part basic test circuit of **Fig. 9A**, using a J507 as J1. Virtually all of the performance characteristics of this circuit are dependent upon the basic J1 part (and to some extent the ap-

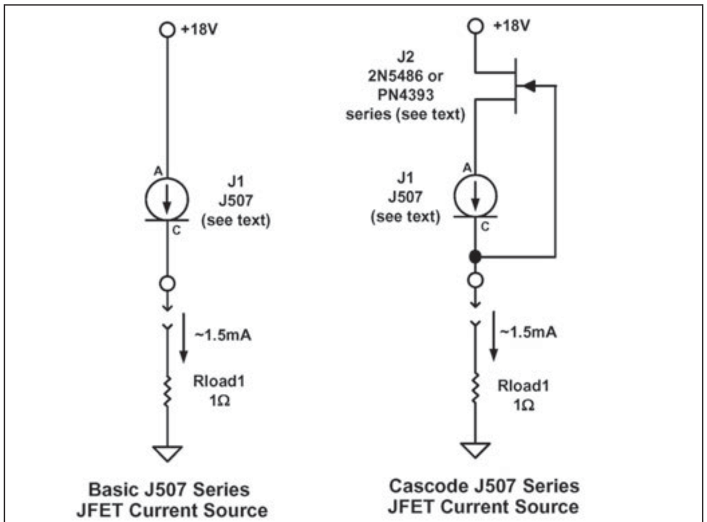


FIGURE 9A: J507 series JFET current source (basic, left) with alternate cascode (right).

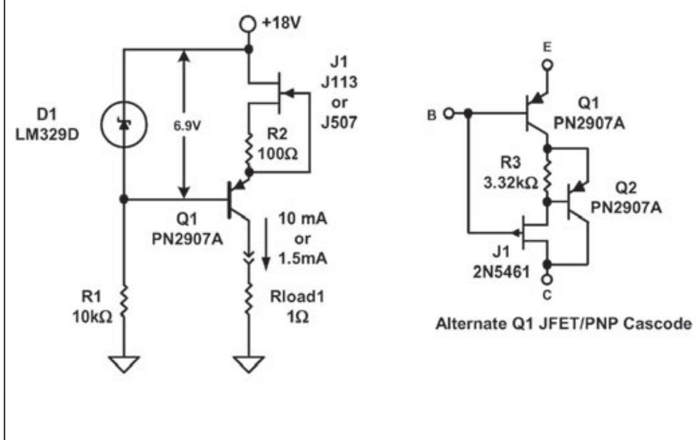


FIGURE 9C: A reference diode buffered JFET current source. Basic operation is shown at the left, with an alternate cascode connection for Q1 shown at the right.

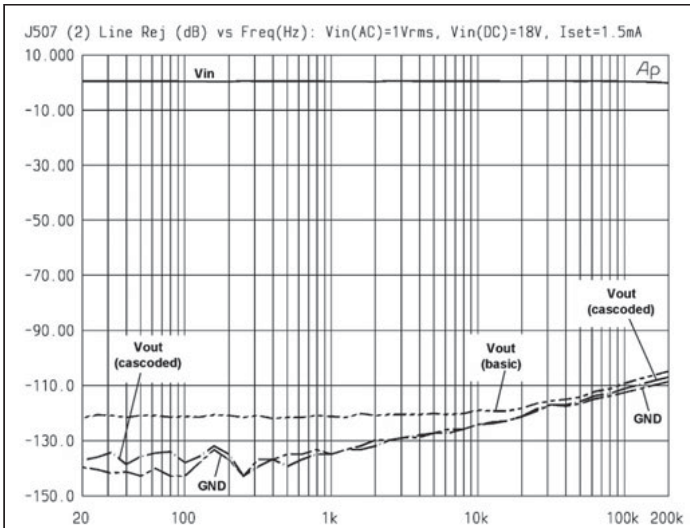


FIGURE 9B: The *Fig. 9A* J507 series current source performance for basic and cascode J507 operation. Cascoding greatly improves low frequency rejection.

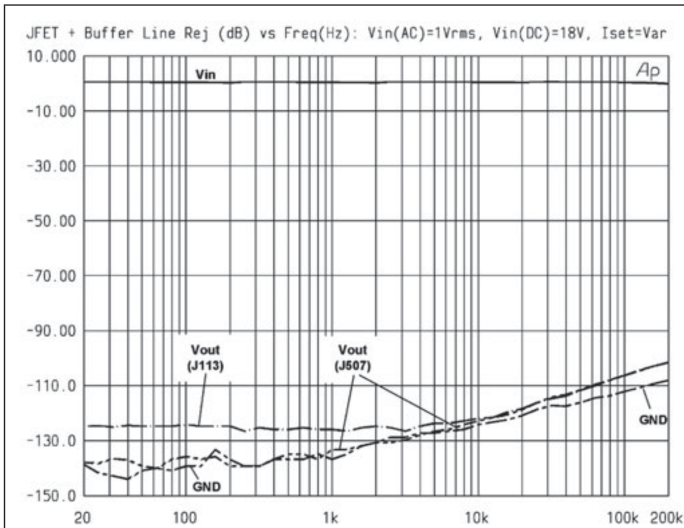


FIGURE 9D: The *Fig. 9C* buffered current source performance for basic 1.5 and 10mA operation. The J113 higher current decreases low frequency performance.

plied voltage). In the case of the J507 sample, an output current of ~1.5mA resulted.

Figure 9B shows the performance of this circuit for both basic and cascode modes. For the basic mode, the LF rejection is 120dB, equivalent to a 1MΩ impedance. Even at the higher frequencies the impedance is still excellent, rising only slightly above the noise level.

The two-terminal J507 is cascoded with a JFET as shown as the right option in **Fig. 9A**, a hookup very similar to the JFET cascode of **Fig. 8C**. For the J500 series, a minimum limiting voltage bias is given on the datasheet, which for the J507 is 2.5V. Thus any cascode circuit should provide a bias from the cascode part above this voltage to achieve best results.

The actual voltage for the parts tested was 2.6V, which was sufficient to allow the performance shown in the cascoded Vout plot. As you will note, this allows a LF rejection of ~140dB (10MΩ), and at higher frequencies performance just above the noise level.

Like the cascoded JFET circuit of **Fig. 8A**, you can apply this cascode to any part of the J500 series or to other similar JFET current regulators. You should choose the cascode device to supply the required minimum voltage across J1. Also take into account voltage limitations for J2, as in the case of the J2 device of **Fig. 8A**, right.

Reference Diode Buffered Current Source

The necessity for very careful attention with the JFET cascode biasing can be a real source of user frustration. Or, perhaps, a much higher voltage capability may also be necessary to get around a low JFET device rating, for example. A solution to both of these points is the reference diode buffered current source of **Fig. 9C**.

Here a 6.9V reference diode is used to bias Q1, a bipolar device, which allows operation up to the Q1 60V Vcb rating. The PN2907A is general purpose, but you can also substitute higher voltage parts, for example the 2SA1016. The 6.9V from D1 is reduced to about 6.3V across J1, which is sufficient to provide operation in the flat portion of the JFET device's output curve, where the rejection is highest. This is true at least for all devices of the J500 series, but some higher Vgs(off) individual JFETs could use even higher voltage bias (~10V).

Performance of this basic circuit is shown in **Fig. 9D** for two basic conditions: 1) a J507 operating at 1.5mA, and 2) a selected J113 JFET operating at 10mA. The J507 performance is generally very close to the noise level, except for a gradual departure at higher frequencies. The J113 is not quite as good, achieving a LF rejection of 125dB, which would be equivalent to 1.8MΩ. This degradation in performance is typical of many current sources when operated at higher currents. Fortunately, there is a solution for it, which is the application of a cascode device for Q1, as originally discussed with **Fig. 6A**.

But, in terms of basic operation, some notes on D1 selection are appropriate here. This diode need not be a high precision part, and even ordinary 1N5230 series (and other similar) zeners will work if chosen for a ~6V breakdown at the final operating current. Note that this can lead to some pre-qualification for proper operating voltage for any ordinary zener; that is, one specified for such higher test currents as 20mA.

It is for this reason that an IC reference diode is preferred here, to alleviate this voltage uncertainty and allow predictable operation at 1-2mA. If an IC diode is used for D1, it can be the loosest tolerance of the family without impacting performance. The LM336Z-5.0 can also be useful in this circuit.

Note that this circuit has similarities to **Fig. 6A**, but **Fig. 6A** depends upon the diode characteristics much more than this circuit. In this **Fig. 9C** circuit, the JFET (or other current source part) in the emitter of Q1 determines the output current, the stability, and so on.

For high currents, a cascode connection can be used for Q1, and will reap performance benefits similar to those noted for **Fig. 6A**, operated with the simple cascode. But the simple cascode using a JFET such as the 2N5462 will be limited to currents of just 4mA within this circuit. To allow a greater degree of freedom in the cascode device selection, you can use the alternate JFET/PNP cascode shown at the right in **Fig. 9C** for Q1. In this configuration, a lower current JFET part (the 2N5461) is used, and is operated at 200µA. In typical operation, this biases the source of J1 about 1V below the gate, sufficient to drive Q2 at currents of 10mA or more, without saturation of Q1. In essence a composite JFET is formed, with a much greater freedom of operating current, but, importantly, still retaining the self-biasing feature of Q1. The alternate cascode at the right replaces Q1 in the circuit at the left, with the C, B, and E terminals as noted.

The performance of this alternate cascode operating with the J113 at 10mA in the circuit of **Fig. 9C** is shown in **Fig. 9E**.

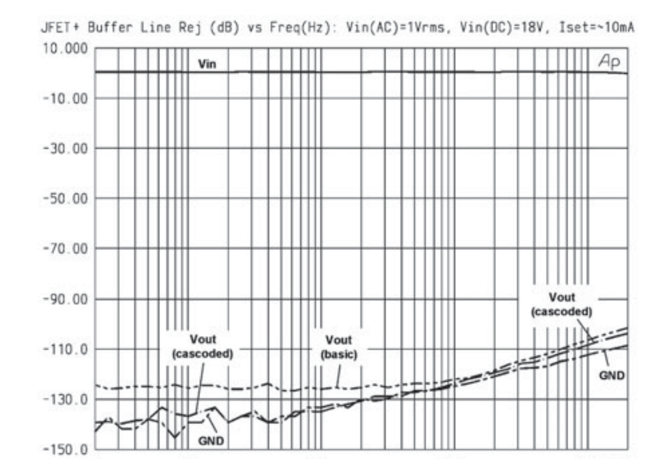


FIGURE 9E: The *Fig. 9C* buffered current source performance for basic and cascoded 10mA, J113 operation. Cascoding greatly increases performance.

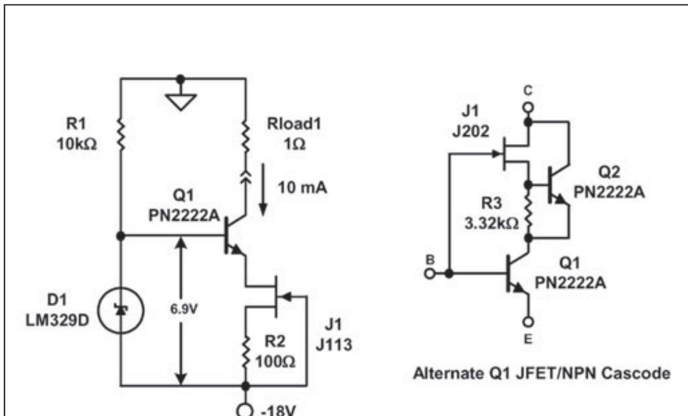


FIGURE 9F: A reference diode buffered JFET current sink. Basic operation is shown at the left, with an alternate cascode connection for Q1 shown at the right.

Now the LF rejection is comparable to the noise level, and the HF performance is also improved. The use of this cascode is recommended for any current above a few mA, or whenever the highest performance is required.

Reference Diode Buffered Current Sink

Comparably operated current regulators can be built to operate as current sinks, and have similar options to those of **Fig. 9C** for cascoding, and so on. Because it is potentially confusing to say simply "use mirror image connections and complementary devices," a full current sink schematic example is shown in **Fig. 9F**. The basic circuit at the left is an exact complement to **Fig. 9C** operated at 10mA, with the J113 setting the current. In the case of this current sink circuit, biasing is from a –18V supply, and the output drives load Rload1.

Cascoding Q1 of this current sink can be done if a JFET/NPN setup is used, as shown in the option at the right of **Fig. 9F**. This alternate JFET/NPN cascode can be used at currents of 5mA and up, without concerns of JFET device preselection. This optional circuit operates similar to the alternate JFET/PNP cascode of **Fig. 9C**, but here uses complementary devices. To employ this cascode, connect the noted C, B, and E terminals within the left basic circuit at the three Q1 nodes.

NEXT TIME. . .

Part 2 will continue these discussions, and will focus on higher current and higher voltage regulators with predictable, low TC output currents, and with very high rejection. It will conclude with some suggestions for applications and user modifications toward optimum use. *aX*

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2. "LM134/234/334 3-Terminal Adjustable Current Sources," *National Semiconductor*, March 2005, www.national.com.

3. Arthur D. Evans, *Designing With Field-Effect Transistors*, McGraw-Hill, ISBN 0-07-057449-9, 1981.

4. "The FET Constant-Current Source/Limiter," Application Note AN103, Vishay/Siliconix, March 10, 1997, www.vishay.com.

5. "J500 Series Current Regulator Diodes," Vishay/Siliconix, July 2, 2001, www.vishay.com.

6. Selected and matched JFETs and JFET current regulator devices, as well as other audio components, are available from Borbely Audio. See www.borbelyaudio.com/audiophile_components.asp.

FAIRCHILD SEMICONDUCTOR®

December 2012

LM336Z25
Programmable Shunt Regulator

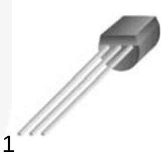
Features

- Low-Temperature Coefficient
- Guaranteed Temperature Stability: 4 mV (Typical)
- 0.2 Ω Dynamic Impedance
- 1.0% Initial Tolerance Available
- Easily Trimmed for Minimum Temperature Drift

Description

The LM336Z25 integrated circuit is a precision 2.5 V shunt regulator. The monolithic I_C voltage reference operates as a low temperature coefficient 2.5 V Zener with 0.2 Ω dynamic impedance. A third terminal on the LM336Z25 allows the reference voltage and temperature coefficient to be trimmed. LM336Z25 is useful as a precision 2.5 V low-voltage reference for digital voltmeters, power supplies, or OP-AMP circuitry. The 2.5 V makes it convenient to obtain a stable reference from low-voltage supplies. Further, since the LM336Z25 operates as a shunt regulator, it can be used as either a positive or negative voltage reference.

TO-92



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Ordering Information

Part Number	Operating Temperature Range	Top Mark	Package	Packing Method
LM336Z25	0 ~ +70°C	LM336Z25	TO-92	Bulk
LM336Z25X		LM336Z25	TO-92	Tape and Reel

Block Diagram

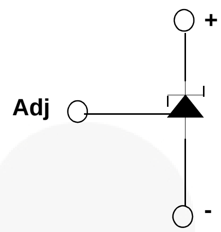


Figure1. Block Diagram

Schematic Diagram

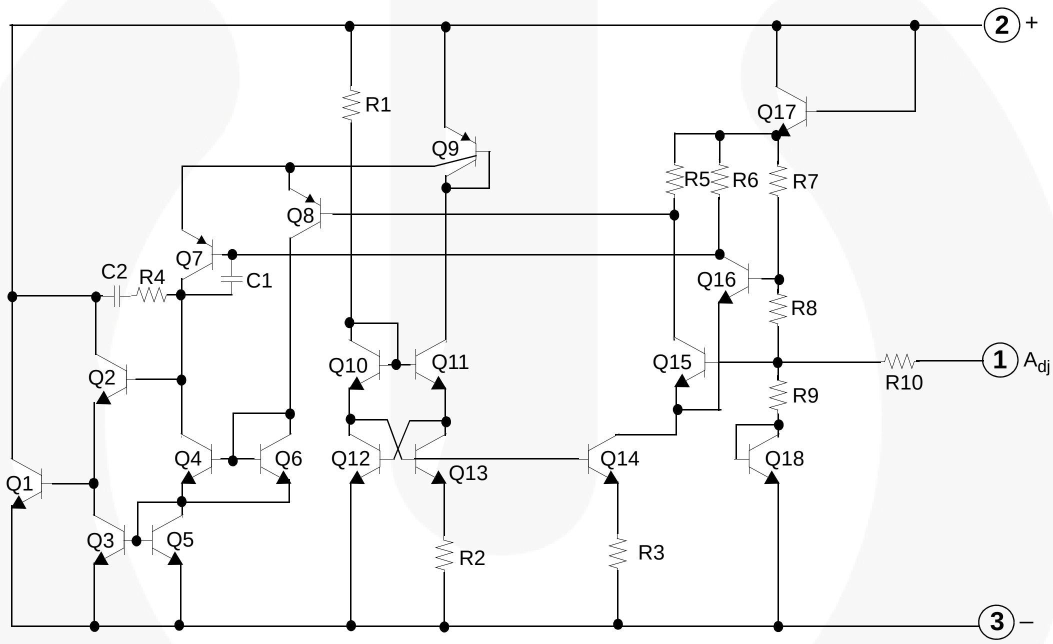


Figure 2. Schematic Diagram

Absolute Maximum Ratings[1]

Stresses exceeding the absolute maximum ratings may damage the device. The device may not function or be operable above the recommended operating conditions and stressing the parts to these levels is not recommended. In addition, extended exposure to stresses above the recommended operating conditions may affect device reliability. The absolute maximum ratings are stress ratings only. Values are at T_A = 25°C unless otherwise noted.

Symbol	Parameter	Value	Unit
I_R	Reverse Current	15	mA
I_F	Forward current	10	mA
T_{OPR}	Operating Temperature Range	0 ~ +70	°C
T_{STG}	Storage Temperature Range	-60 ~ +150	°C

Note:

1. The Absolute Maximum Ratings are those values beyond which the safety of the device cannot be guaranteed. The device should not be operated at these limits. The parametric values defined in the Electrical Characteristics tables are not guaranteed at the absolute maximum rating.

Electrical Characteristics

Values are at 0°C ≤ T_A ≤ +70°C unless otherwise specified.

Symbol	Parameter	Conditions	Min.	Typ.	Max.	Unit
V_R	Reverse Breakdown Voltage	T_A = 25°C, I_R = 1 mA	2.44	2.49	2.54	V
$\Delta V_R / \Delta I_R$	Reverse Breakdown Change with Current	T_A = 25°C, 600µA ≤ I_R ≤ 10 mA		2.6	10.0	mV
Z_D	Reverse Dynamic Impedance	T_A = 25°C, I_R = 1 mA		0.2	1.0	Ω
ST_T	Temperature Stability	I_R = 1mA		1.8	6.0	mV
$\Delta V_R / \Delta I_R$	Reverse Breakdown Change with Current	600 µA ≤ I_R ≤ 10 mA		3.0	12.0	mV
Z_D	Reverse Dynamic Impedance	I_R = 1 mA		0.4	1.4	Ω
ST	Long Term Stability In Reference Voltage	I_R = 1 mA		20.0		ppm/ Khr

Typical Performance Characteristics

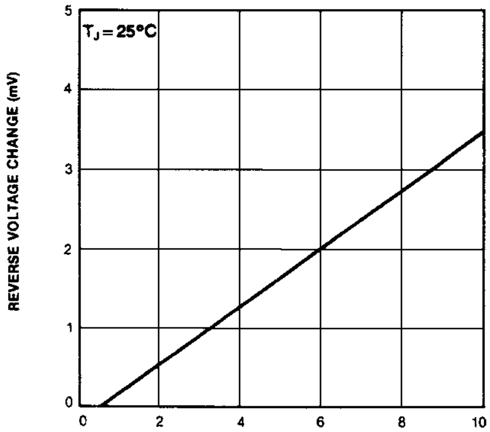


Figure 3. Reverse Voltage Change

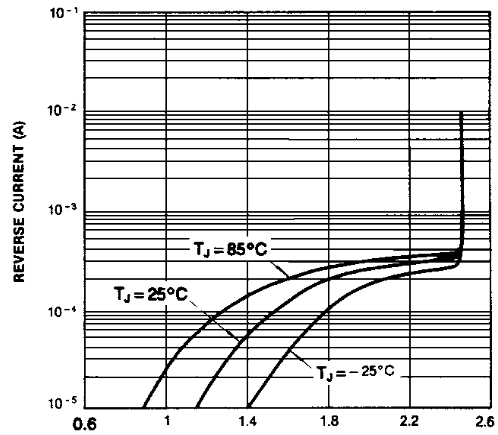


Figure 4. Reverse Characteristics

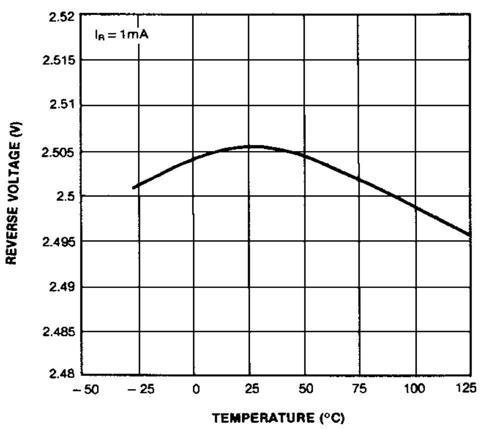


Figure 5. Temperature (°C)

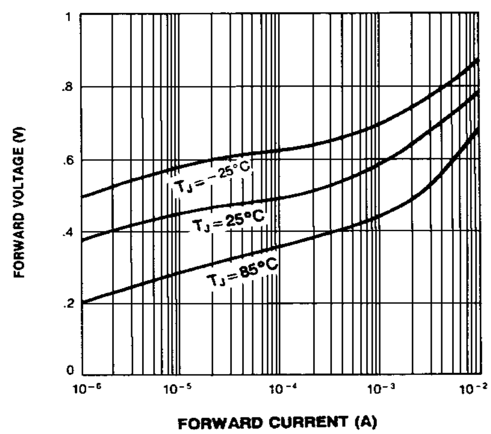
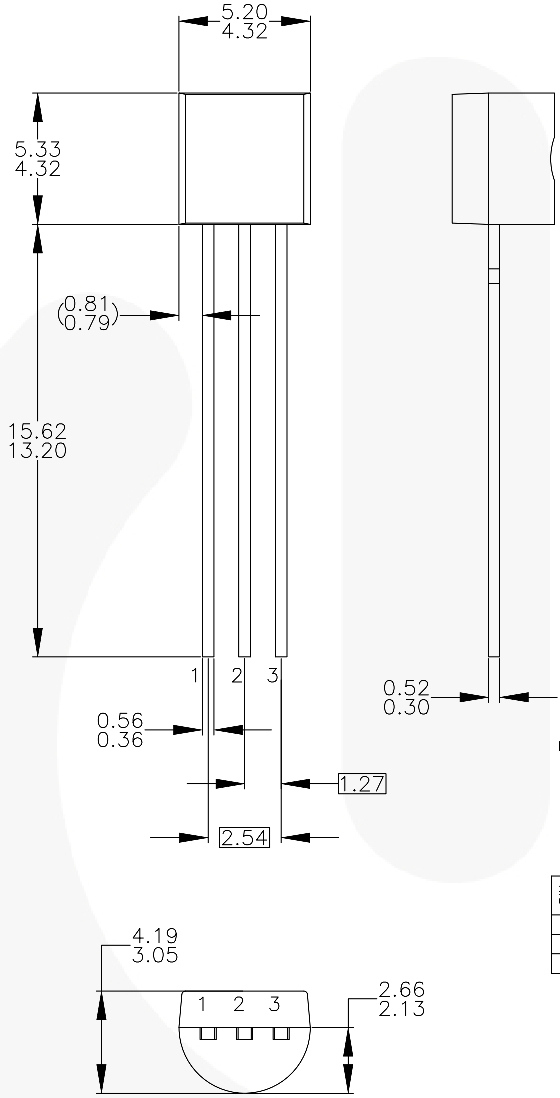


Figure 6. Forward Characteristics

Physical Dimensions

TO-92 Bulk Type



NOTES: UNLESS OTHERWISE SPECIFIED

A) DRAWING WITH REFERENCE TO JEDEC TO−92 RECOMMENDATIONS.
B) ALL DIMENSIONS ARE IN MILLIMETERS.
C) DRAWING CONFORMS TO ASME Y14.5M−1994.
D) TO−92 (92,94,96,97,98) PIN CONFIGURATION:

PIN	92			94			96			97			98		
	P	F	M	P	F	M	B	F	M	P	F	M	P	F	M
1	E	S	S	E	S	S	B	D	G	C	G	D	C	G	D
2	B	D	G	C	G	D	E	S	S	B	D	G	E	S	S
3	C	G	D	B	D	G	C	G	D	E	S	S	B	D	G

LEGEND:
P − BIPOLAR E − EMITTER D − DRAIN
F − JFET B − BASE S − SOURCE
M − DMOS C − COLLECTOR G − GATE

E) FOR PACKAGE 92, 94, 96, 97 AND 98: PIN CONFIGURATION DRAIN "D" AND SOURCE "S" ARE INTERCHANGEAGLE AT JFET "F" OPTION.
F) DRAWING FILENAME: MKT−ZA03DREV3.

Figure 17. 3-Lead, TO-92, Molded, Standard Straight Lead

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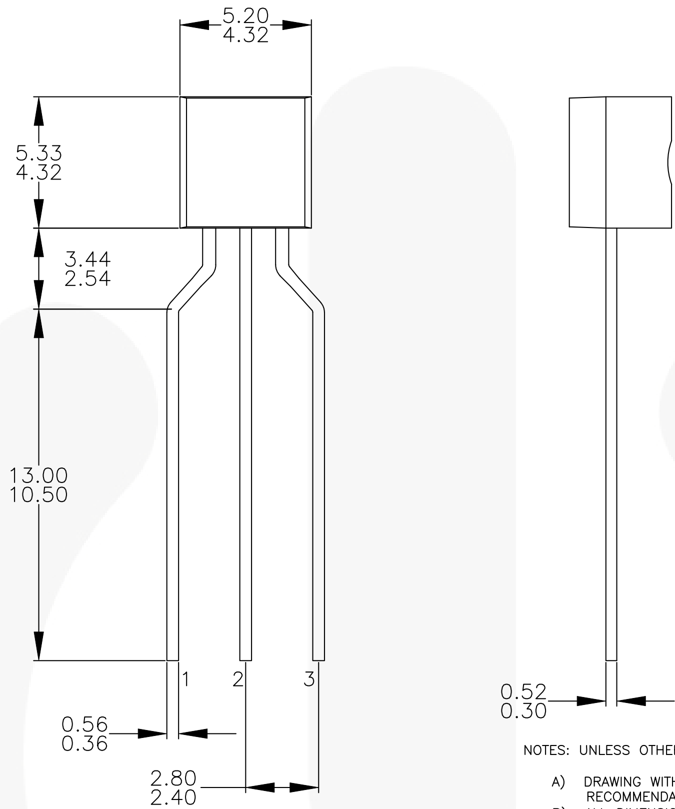
Physical Dimensions (Continued)

TO-92 Tape and Reel Type

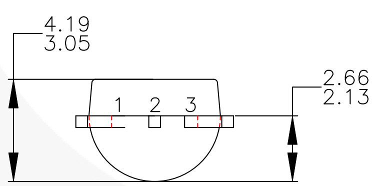


NOTES: UNLESS OTHERWISE SPECIFIED

A) DRAWING WITH REFERENCE TO JEDEC TO–92 RECOMMENDATIONS.
B) ALL DIMENSIONS ARE IN MILLIMETERS.
C) DRAWING CONFORMS TO ASME Y14.5M–1994.
D) TO–92 (92,94,96,97,98) PIN CONFIGURATION:

PIN	92			94			96			97			98		
	P	F	M	P	F	M	B	F	M	P	F	M	P	F	M
1	E	S	S	E	S	S	B	D	G	C	G	D	C	G	D
2	B	D	G	C	G	D	E	S	S	B	D	G	E	S	S
3	C	G	D	B	D	G	C	G	D	E	S	S	B	D	G

LEGEND:
P – BIPOLAR
F – JFET
M – DMOS
E – EMITTER
B – BASE
C – COLLECTOR
D – DRAIN
S – SOURCE
G – GATE



E) FOR PACKAGE 92, 94, 96, 97 AND 98: PIN CONFIGURATION DRAIN "D" AND SOURCE "S" ARE INTERCHANGEAGLE AT JFET "F" OPTION.
F) DRAWING FILENAME: MKT–ZA03FREV2.

Figure 18. 3-Lead, TO-92, Molded, 0.200 in Line Spacing Lead Form

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For current tape and reel specifications, visit Fairchild Semiconductor's online packaging area:
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Article prepared for www.audioXpress.com

Sources 101: Audio Current Regulator Tests for High Performance

Part 2: Precise High Current/Voltage Operation

By Walt Jung

Measurement tests can help reveal which configuration is best for your power supply application.

I will conduct many additional measurements here. Within this phase, the focus is on current regulators that operate at higher voltages, at higher currents, and do so with a higher degree of precision. This implies higher initial accuracy, as well as good temperature stability, for all circuits discussed hereafter, with the exception of those MOSFET based.

LM317 CURRENT SOURCE/SINK

One of the easiest ways to make a quite good audio current source is to simply connect an LM317 IC with a current set resistor (**Fig. 10A**, left). This circuit, which is simplicity personified, cannot be reduced further in functionality. Details of the LM317 operation are described in References 7 and 8 (highly recommended reading). The wide availability of this useful part in a variety of packages at low cost makes it attractive.

The LM317 is a *floating* three-terminal regulator, meaning it can be applied quite flexibly, and no pin inherently needs to be grounded. When operated in a current mode, the internal 1.25V reference voltage appears between the

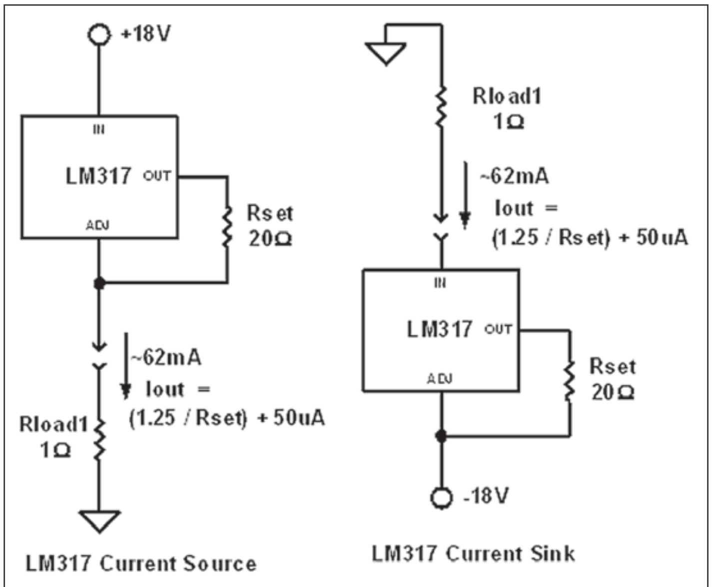

FIGURE 10A: Basic LM317 current source (left) and sink (right).

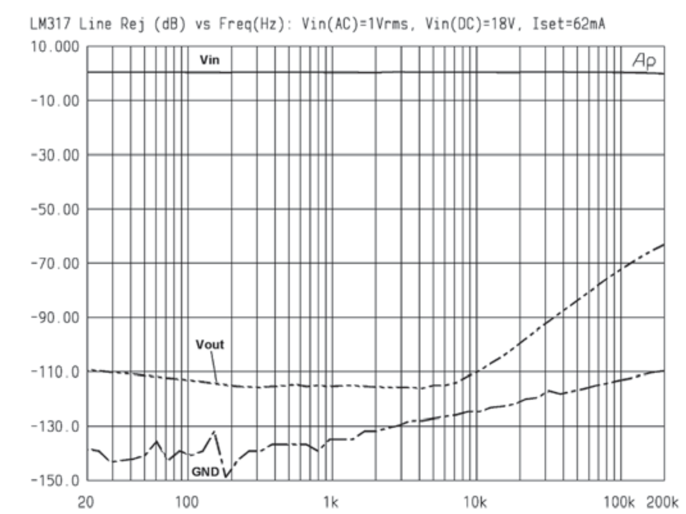

FIGURE 10B: Performance of the LM317 as a 62mA current source shows 110dB or more rejection below 10kHz, but rapid deterioration at higher frequencies.

OUT and ADJ pins, so a simple resistor Rset programs the current into a load. In this case a fixed 20Ω value sets up a 62mA load current. The 1.25V is held to ±50mV, and is stable over temperature.

Thus, an LM317-based current source will be one of the more predictable and stable types for DC. Of course, at such higher currents power dissipation will be an issue, so you should use a TO-220 package part at these current levels, along with the appropriate heatsink.

It may not be obvious at first, but the LM317 can function as both a current source (as in the left case) and as a current sink, shown at the right. In either case, the IC and its Rset resistor are treated as a two-terminal circuit, which is applied between the source and the load. The LM317 current sink is implemented with similar connections shown at the right, with the load connected to the IC's IN pin, and using a negative power supply. Note that in such cases a small AC bypass capacitor may be necessary at this pin, ~1µF.

The LM317 working in this current output mode will require about 2.5V across the IC, plus the 1.25V, for a total of nearly 4V to make it operate. The IC also needs a 10mA minimum of output current for regulation. Practically speaking, this means that Rset should never be any higher than about 125Ω.

Once biased properly, the IC operates reasonably well, as shown in **Fig. 10B**, an AC rejection performance plot of the output. Here the low frequency (LF) rejection is about 110dB, equivalent to an impedance of 316kΩ. There is, however, noticeable deterioration at higher frequencies.

This is one aspect of the LM317's performance that would be desirable to improve, because the rejection at 200kHz is only about 60dB, meaning potentially increased sensitivity to high frequency (HF) intermodulation. A couple of the following circuits address this aspect of the LM317's operation.

LM337 CURRENT SOURCE

A companion device to the LM317 positive regulator IC is the LM337, designed to operate from negative sources. It also has a 1.25V reference voltage and can be configured to regulate current (**Fig. 11A**). The LM337 uses a similar set resistor (Rset) to set up an output current Iout, but it also requires an output capacitor for frequency compensation, C1. A typical value for this capacitor is shown.

While the LM317 and LM337 have complementary functionality, they achieve radically different degrees of rejection versus frequency as operated in a current mode. This is best appreciated by the LM337's AC performance (**Fig. 11B**). While the LM337 rejection is good below a few hundred Hz, it degrades steadily above this, to the point where the rejection is less than 30dB above 100kHz. This is an example of the type of rejection *not* sought for higher performance audio circuits!

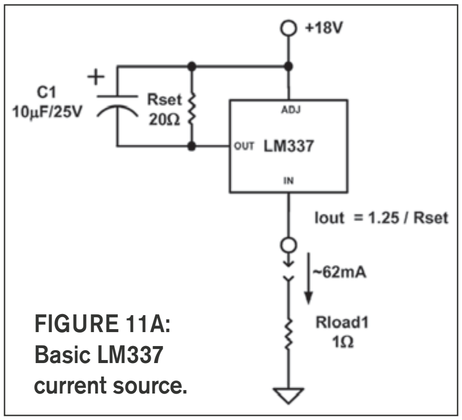

FIGURE 11A: Basic LM337 current source.

A detail worth noting at this point: If complementary source and sink circuits are needed for an application, it is actually better performance-wise to use a pair of LM317s as in **Fig. 10A** left and right, than it would be to use an LM317 and an LM337.

Caveats: A further special point on three-terminal regulator types is to simply be cautious about replacement or "improved" 317-type regulators, especially those designed for low dropout. As a byproduct of their design for low DC dropout voltage, these regulator types can have much worse AC rejection characteristics vis-à-vis the original. For example, two low dropout versions of the 317 were tested for rejection in a current regulator mode similar to **Fig. 10A**, and had responses more like that of **Fig. 11B** than the more desirable LM317 response of **Fig. 10B**. So, this is definitely a case of caveat emptor!

DEPLETION MODE MOSFET CURRENT SOURCE/SINKS

Power MOSFETs are both extremely popular and widely available, and for many years have seen widespread use in audio amplifiers. Typically, these have been the original format, which is that of *enhancement mode* devices. This means simply that they require an applied gate voltage to conduct.

More recently, *depletion mode* MOSFETs have become available, which enables easier use of such parts in audio power supplies. Like the small signal JFETs, a depletion mode MOSFET is fully on with 0V bias, and is controlled to lower degrees of conduction with the applied bias voltage. Thus far the depletion mode MOSFETs that have appeared are N-channel parts. Two TO-220 packaged examples are the DN2540 from Supertex and the IXCP 10M45 from Ixys. See References 10 and 11 for further information.

These TO-220 devices can operate at voltages up to 450V, and at currents from the low mA range up to about 100mA. They are already being found in vacuum-tube-based audio projects where high voltage capability is required. Examples can be found via Reference 12.

In application, a basic current source using either part can be accomplished (**Fig. 12A**). This circuit is exactly the same

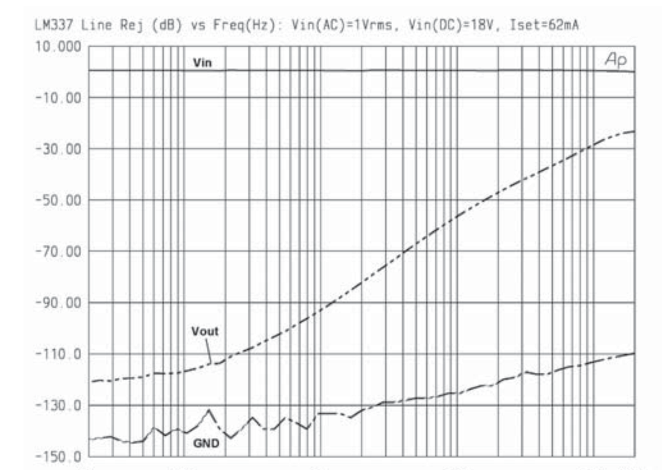

FIGURE 11B: Performance of the LM337 as a 62mA current source shows 110dB or more rejection—but only at the lowest frequencies.

as with a JFET device, save the addition of the gate-stopper resistor R1, and the important fact that the applied voltage can go up to 450V. And, like the JFET counterpart current regulator, this circuit is two-terminal, and so can be used either as a source (shown here), or as a sink, where the load is in the drain lead and negative voltage is applied to the bottom of Rset and R1. The tests described here used an 18V power supply.

For a load current of 30mA, I found that the two resistor values noted for Rset were appropriate. This underscores a basic point: These depletion mode MOSFETs aren't precision devices like the LM317 and other ICs with their fixed reference voltage(s). Rather, the gate bias for these MOSFETs sample to sample will vary, just as it does for other JFET and MOSFET parts. Nevertheless, this circuit still has the utility of extreme simplicity, and Rset is simply chosen to get the required current.

Operated within the test circuit of **Fig. 12A**, the two sample parts produced the data of **Fig. 12B**. Both devices show LF rejections of around 110dB (~316kΩ), with a gradual degradation beginning in the 5–10kHz range. The DN2540 is measurably better in terms of AC rejection at the higher frequencies. This is apparently due to the lower parasitic capacitance of the DN2540 versus the IXCP 10M45, but I cannot precisely confirm this (the latter isn't specified for capacitance).

Nevertheless, these general patterns of AC rejection seemed to be typical for the two devices, and were observed with tests of other samples. The DN2540 is preferred for operation in this circuit, not only because of the better AC rejection at high frequencies, but because the Idss of this part is 150mA, making it more widely applicable.

CASCODE LM317 CURRENT SOURCES

These higher current regulators, like the low-level circuits described in Part 1, can also be enhanced for AC performance by means of cascoding. As the DC current carried by the regulator is increased, the rejection performance inevitably degrades, making the value of an effective cascode circuit more and more important toward good results.

A circuit that can be used to cascode the operation of an LM317 is shown in **Fig. 13A**. This is similar to the basic regulator of **Fig. 10A**, with an additional regulator added—stage U2. The U1 LM317

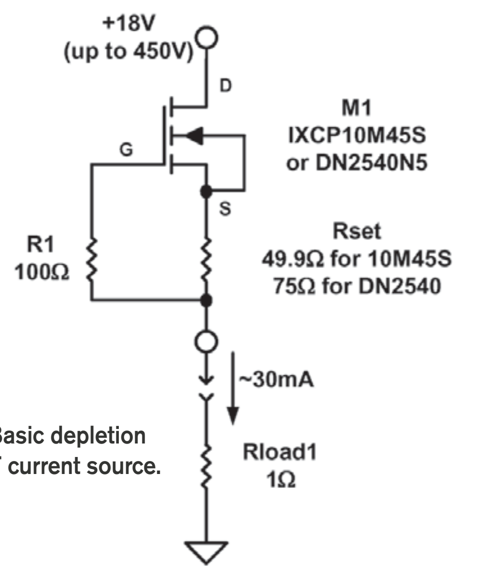

FIGURE 12A: Basic depletion mode MOSFET current source.

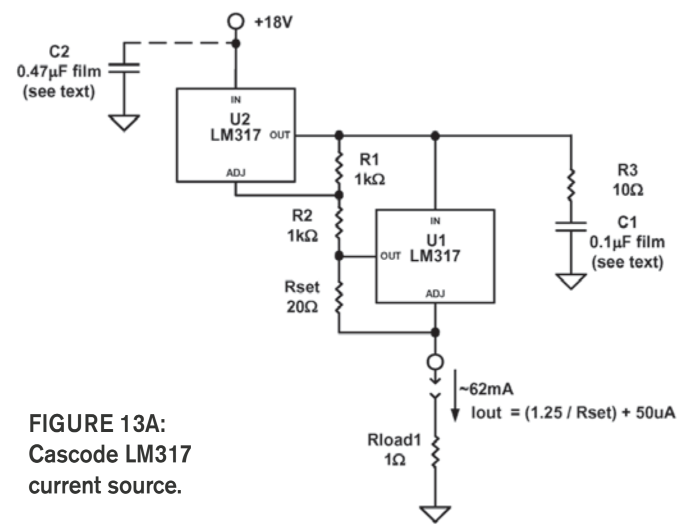

FIGURE 13A: Cascode LM317 current source.

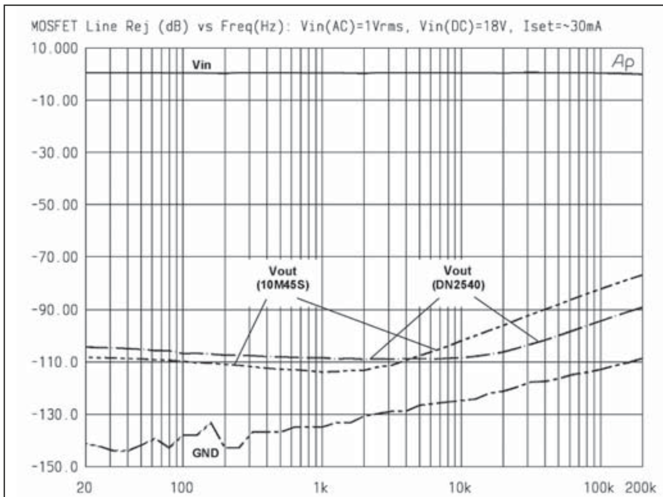

FIGURE 12B: Performance of two depletion mode MOSFET 30mA current sources shows ~110dB rejection below 5-10kHz, then deterioration as frequency increases.

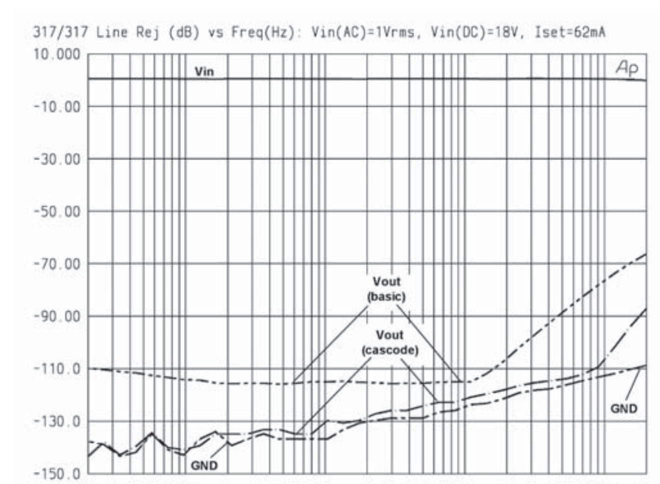

FIGURE 13B: Performance of the LM317/LM317 as a cascode 62mA current source shows much greater rejection than in basic mode, at all frequencies.

operates just as previously, producing an output current as noted, which is proportional to 1.25V and inversely proportional to Rset. The input drive for U1 is derived from cascode IC U2, which floats atop U1's output, 2.5V higher by virtue of resistors R1 and R2. C1 and R3 provide necessary stabilization for the cascode.

I tested the **Fig. 13A** circuit at a current level of 62mA, to be consistent with the basic LM317 operation of **Fig. 10A**. The results are shown in **Fig. 13B** for both the basic and cascode modes of operation. Note that the addition of the cascode reduces the noise down to a level approaching the setup residual at all but the very highest frequencies. Although not shown here, for lower levels of current operation (i.e., ~15mA), this cascode scheme showed even lower noise levels.

Some caveats for the **Fig. 13A** *circuit:* Although the AC rejection properties of this relatively simple circuit could be considered exemplary in some regards, I cannot recommend it unconditionally for several important reasons. One, it has a rather high dropout voltage, requiring ~6.5V across it—just to function! This is due primarily to the basic characteristics of the LM317, and can't be easily reduced. Anticipating potential questions here, using low dropout 317 regulators isn't any real help, either. I tried this, and it does reduce the dropout—but at the expense of rejection.

A second caveat is that the basic LM317 dropout voltage is actually specified as 3V for currents up to 1.5A. Datasheet graphs show it to be typically ~1.7V at a current of 200mA at 25°C. So the scheme here won't really work well at high currents and/or low temperatures.

But, there is still much latitude for use at much lower currents and typical temperatures from 25° C and up. Here operation of U1 is at a fixed input/output voltage of 2.5V, and because this is still somewhat of a gray area, only load currents of <100mA are suggested. Finally, and perhaps most important, this setup can and will oscillate under certain conditions, so be wary. All cascode-type schemes using additional high gain, wide bandwidth parts have this potential and should be rigorously checked. Input bypassing should be used, with a film capacitor such as C2 close to U2, and the C1/R3 network always used.

Fortunately, for all of the cascode schemes tested for this series, only a couple of them showed oscillation tendencies, this one included. The absence of oscillation for an LM317 regulator can be checked by the presence of a stable 1.25V ±50mV output (or, the exact target DC current for this or other precision regulators). If a scope is used, the output should be clean on a scale of a few mV.

For some more carefully selected operating conditions, a cascode LM317 arrangement can be implemented using an LM317 as the control IC and a depletion mode MOSFET as the cascode part. This variation (**Fig. 13C**) can use either the DN2540 or the 10M45 as the cascode device M1. *Note that this circuit will simply not work with a conventional MOSFET!* For the two M1 device types, it has the advantage of workability at very high voltages, up to 450V, making it quite attractive as a simple and precise current source for tube circuits.

This circuit also has some caveats, including the general ones for the 317. For the LM317 to properly function as a regulator, the input/output voltage, labeled here as V317, *must meet the LM317 device dropout limits*. In this circuit V317 is the Vgs of M1, and this should be 2.5V or more. Both the devices listed for M1 typically meet this requirement at lower currents of 10–20mA, and the DN2540 holds up even higher. And, don't forget the RC stabilization network, R2/C1.

AC rejection performance of this circuit operating at 16mA is shown in **Fig. 13D**, and for either of the cascode devices it is nearly ideal. Only a tiny deviation above the noise level at the very highest frequencies can be noted. This exceptional performance makes this a very attractive circuit for such lower currents.

At the higher current of 38mA (**Fig. 13E**), the 10M45 begins to approach the sample device Idss. Therefore, V317 is lower than the minimum required for effective LM317 operation, and as a result, the data for the 10M45 shows noticeable deterioration vis-à-vis lower currents. By contrast, the DN2540, a higher current

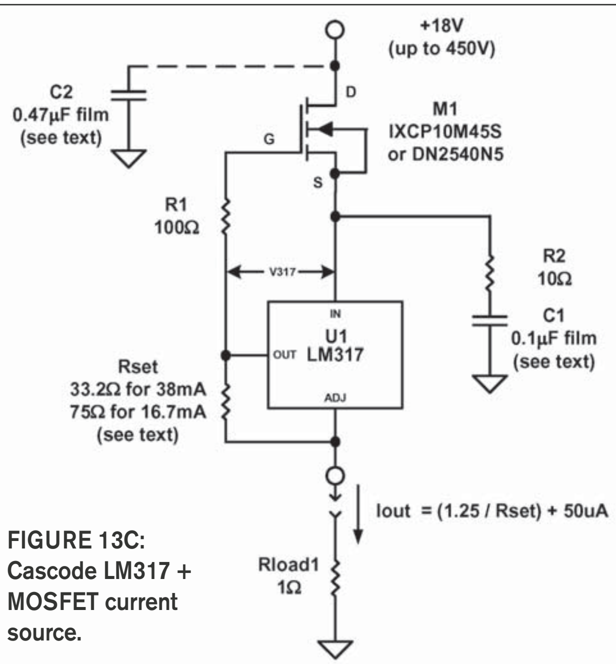

FIGURE 13C: Cascode LM317 + MOSFET current source.

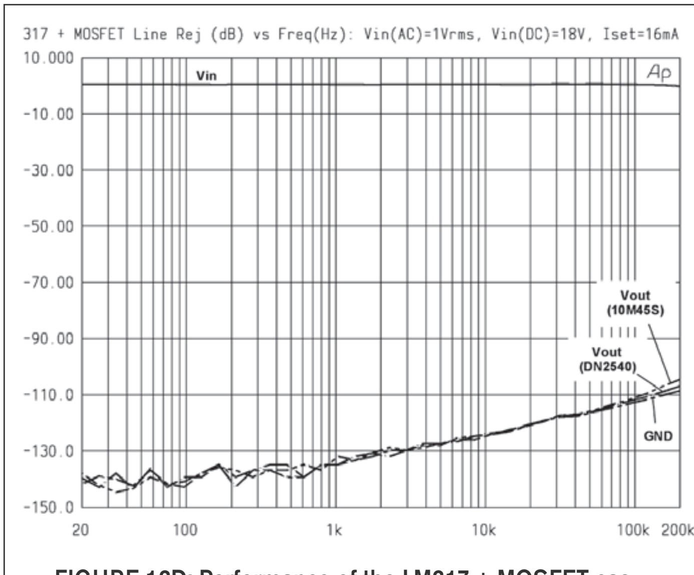

FIGURE 13D: Performance of the LM317 + MOSFET cascode 16mA current source shows excellent rejection compared to basic mode at all frequencies.

device, still shows excellent rejection for these conditions.

A power caveat: While you should always be aware of power dissipation limits for any of these circuits, this boundary can quickly sneak up on you within tube circuits—even at relatively low current levels. For example, a 10mA current in M1 of **Fig. 13C** with 150V across it implies an M1 dissipation of 1.5W, which will definitely require a heatsink. Don't operate under the assumption that a datasheet rating of 1W at 25° C for a TO-220 will guarantee a safe and long life of the part, if it sees 1W of constant power while the room is 25° C. Internally, the part will be much hotter, and it is highly likely a hefty heatsink is in order for a truly reliable design. See Reference 13 for further heatsink information.

TLV431 CURRENT SINK

The TLV431 is a three-terminal IC designed to be used as a programmable shunt regulator, from 1.24 to 6V[14]. It has an uncommitted feedback path, meaning that external active parts can be used with it to extend the basic current and voltage range. As you will see, this part operates as a current regulator referred to the negative rail, thus it is most suited to make current sinks.

The TLV431 reference voltage of 1.24V has a tolerance of ±18mV (1.5%), but A and B suffix parts tighten this to 12mV (1%) and 6mV (0.5%), respectively. The TLV431 is related to the very popular TL431, which offers similar functionality at a reference voltage of 2.5V. Because the TLV431's lower voltage of 1.24V is more desirable for a current regulator (it means lower dropout), I chose it for this test. But note that the same principles applied here for the TLV431 also work for the TL431, except for the higher reference voltage of 2.5V.

Figure 14A is a basic TLV431 current sink that you can use over a range of voltages up to 40V, and currents up to several tens of mA. The final voltage/current rating for this circuit is a function of the transistor type used for Q1 and the heatsinking. Typically the load would be applied between the OUT1 and OUT2 terminals. Note that the OUT1 terminal need *not* be common to the +18V supply as shown; it can (and often will) be a higher voltage.

The U1 IC, a TLV431, regulates with a 1.24V developed between the R and A terminals as noted, so the Rset resistance determines the current flowing into Q1-Q2 and the external load. The feedback path is via terminal K and the base-emitter path of Q1–Q2. Z1 performs as load impedance for IC U1, and can be one of three options, all of which should provide for a current of 100µA, minimum. The simplest option is a 100kΩ resistor (A); next most simple a current source such as the J507 (B); and finally, for highest performance from the circuit as a whole, functioning as a current source, a J202 operating at ~280µA and cascoded with a 2N5486 (similar to *Fig. 8A*, right option, Part one).

Because this circuit is more aptly used as a current sink, the measure of how it performs would best be told by a sense resistor placed at OUT1-OUT2. But, as noted, the test setup here measures current in Rload1, which is tied to ground. Interestingly, however, you can still infer some degree of performance of the circuit by observing the total current in Rload1.

The current in Rload1 has two components, the output current flowing in Rset-Q1/Q2, and Iz1, the bias current of U1, which flows in Z1-U1. When the current in Rload1 is monitored, both of these currents are, in fact, being measured. It would be desirable that only Iz1 be dominant, because this would mean that the Rset-Q1/Q2 current path is noise free.

To a great extent this is indeed true, and is reflected by a related change in Iout rejection, because Z1 is varied. This is shown in **Fig. 14B** for various Z1 conditions. Note that for a finite resistance value for Z1, the net impedance is shown by the Vout (100k) plot (as was true for the calibration plots of Part one of this article). And, as Z1 takes on higher impedance characteristics, such as with the Vout (J507) plot, this condition is reflected in a higher impedance display (i.e., more rejection). The greatest rejec-

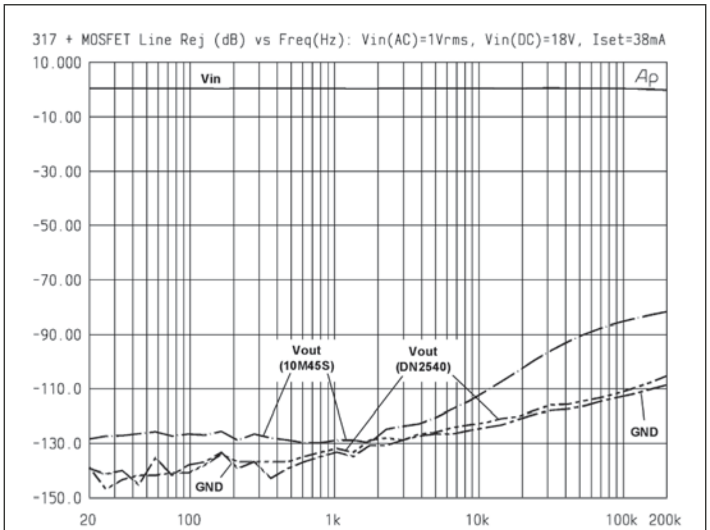

FIGURE 13E: Performance of the LM317 + MOSFET cascode 38mA current source still shows excellent rejection for the DN2540, but deterioration for the 10M45S.

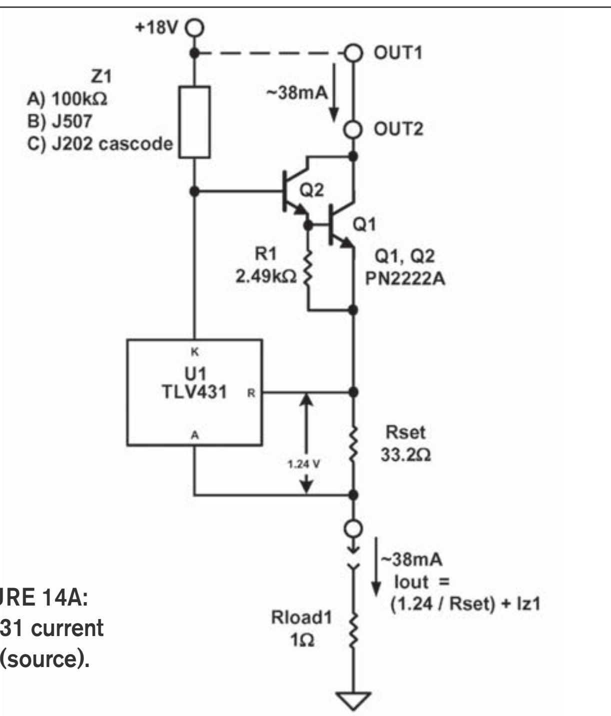

FIGURE 14A: TLV431 current sink (source).

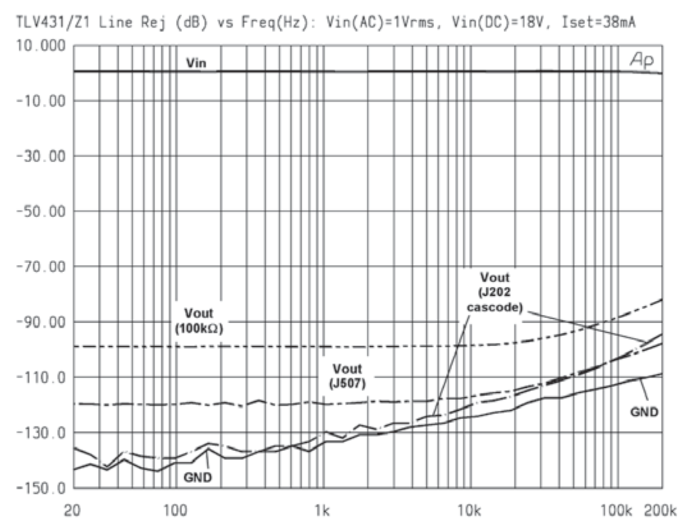

FIGURE 14B: Performance of the TLV431 as a 38mA current source depends upon Z1, but is excellent with a high-Z for Z1. See text on current sink operation.

tion at the lower frequencies is provided by the cascoded J202 setup, while the J507 provides the most rejection with a single component used for Z1.

So, while this test method doesn't directly measure just the current flowing in the collector of Q1/Q2, it still suggests some aspects of relative quality—a good thing, nevertheless. The bottom line is that you can use the circuit as either a current sink, in which case Z1 can likely be the simple 100kΩ resistor, or, alternately, as a current source, whereby the higher impedance choices for Z1 are suggested, such as the J507 or the cascoded J202.

You might ask what the need is for this type of current sink, when previous examples have provided good performance at these currents. The answer lies in the overall flexibility of **Fig. 14A**. Operated as a current sink, and with Q1 properly selected for power and voltage handling, this circuit can handle currents of amperes and voltages as high as the Q1 device rating. Although data isn't shown for this example, with a D44 series power transistor for Q1, output currents of 350mA have been witnessed. This is all available with relative simplicity—Z1 a 100k resistor (Z1) and Rset chosen for the current desired. Or, with Q1/Q2 2SC2362, the OUT1 terminal can operate up to 150V, at low currents, with proper heatsinking.

LM4041 CURRENT SOURCE

The LM4041-ADJ is a three-terminal IC designed to be used as a programmable shunt regulator, from 1.233 to 10V[15]. Like the counterpart TLV431 series, it also has an uncommitted feedback path. And, as with the TLV431, this means external active parts can be used with it to extend the basic current and voltage range.

A key difference in applicability is that the LM4041-ADJ operates with a positive rail common, as opposed to the TLV431, which uses a negative rail common scheme. The two devices can be viewed as complements, performing similar tasks. The basic LM4041-ADJ reference voltage is 1.233V, and the available grades of C and D for this version have initial tolerances of ±0.5% and ±1%, respectively, for Vout = 5V.

Inasmuch as the operation of the LM4041-ADJ is with the positive rail common, you can easily use it to make current sources operating over a wide range. An example is shown in **Fig. 15A**, which is a mirror image of the TLV431 circuit of **Fig. 13A**.

In this current source circuit, the output current is measured in Rload1, which is in series with the Q1-Q2 collectors. There is no error current from the internal amp of the LM4041, thus the rejection characteristics measured at Rload1 are indeed what you get. This is shown in **Fig. 15B**, for conditions as shown and a current of 38mA. The LF rejection is approaching 130dB (3.16MΩ), which, while good, is still well above the noise level. However, the rejection deteriorates above 1kHz.

Cascoding of Q1-Q2 in this circuit did not improve the performance to any great degree, only 2-3dB. At lower cur-

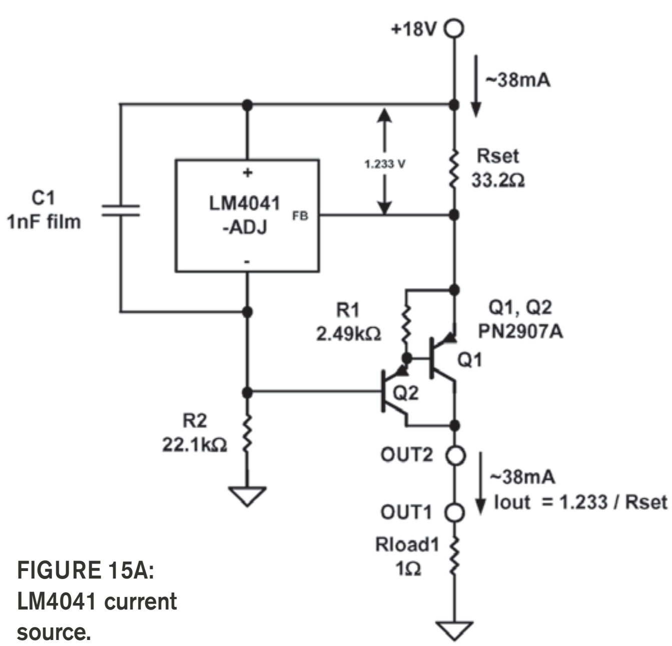

FIGURE 15A: LM4041 current source.

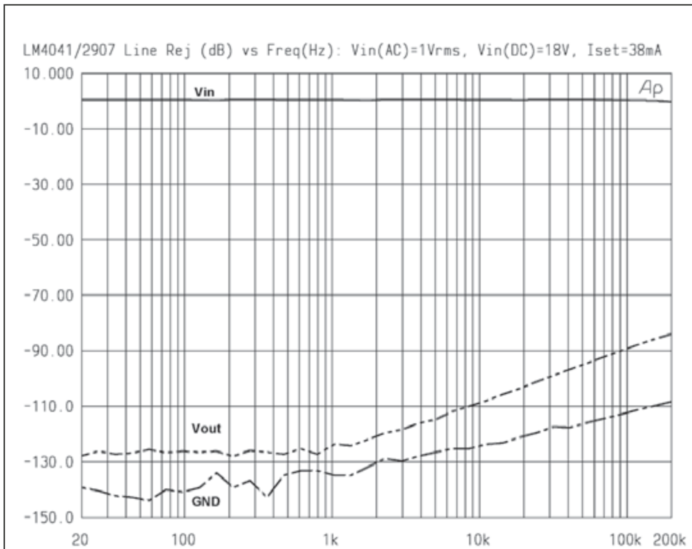

FIGURE 15B: Performance of the LM4041 as a 38mA current source is good, but falls short of excellent, particularly at the higher frequencies.

rent levels of a few mA, the rejection improved to just above the residual noise level. From this, you would conclude that this particular circuit is better used at the lower current levels.

TLV431 BOOSTED CURRENT SOURCE/SINK

As noted, the TLV431 circuits are better suited to use as current sinks, as opposed to sources. But, with some key changes, you can use a TLV431 current regulator either as a source or sink, and/or at high voltages. One scheme to do this is shown in **Fig. 16A**.

This circuit is like **Fig. 14A**, except Q1 uses a standard connection (non-Darlington), and the current source portion represented by Z1 of **Fig. 14A** is replaced by a high current or high voltage equivalent. This has the effect of regulating the current in R1, making the error current flowing from Rset and the TLV431-A pin constant. Therefore, this circuit, operating as a whole, can be used either as a source or as a sink.

With an LM317 for U2, R1 establishes a current of ~800μA, providing drive to Q1 for currents of 50mA or more. Q1 is bootstrapped by the LM317 at the collector and sees less than 2V C-E. It thus does not dissipate high power at 38mA of output or even higher currents. The LM317 will need the heatsink in this circuit long before Q1!

For operating voltages higher than the 40V LM317 rating, you can also use a depletion mode MOSFET by substituting an M1 device at the points marked X and Y. R1 can remain the same, and the current limit for this mode will of necessity be much less than 40mA. But, the voltage limitation becomes that of the M1 device used, or 450V as shown. Take care to use a proper heatsink for M1!

Performance in terms of AC rejection is shown in **Fig. 16B** for all three cascoding options, operating at 38mA. Overall, the best performance is achieved with the LM317, where the errors are only slightly more than residual noise, except for the very highest frequencies. The two MOSFET parts are nearly as good at LF, but deteriorate more rapidly above 1kHz. Of the two MOSFETs, the DN2540 is favored due to lower noise at all frequencies, plus its ability to handle more current. To get higher output currents, Q1 can be operated as parallel devices driven from R1-bottom end, with 10Ω current sharing resistors in the emitters.

CONCLUSIONS, CAVEATS, AND RECOMMENDATIONS

This concludes the testing portions of this series. A future article will explore some example applications of current sources and sinks within audio circuits, and discuss some general power supply system noise reduction techniques.

Some general caveats are appropriate here, beyond those specifically stated. I believe the tests are valid for the conditions cited, and in general can be used to differentiate among the various circuits. Of course, there is an infinite set of different load, voltage, and current operating conditions that you may require. So, you should not expect to duplicate any measurements exactly for other conditions. But, in general the observations should hold up—cascodes work better, JFETs need proper voltages to work best, and so on.

To summarize, here are some principles to keep in mind:

- Select single JFET parts from families with lowest Vgs and thus highest rejection. An example would be the J201/2 series.
- Alternately, select from a specified JFET current regulator device family, such as the J507 series.
- Always operate current regulator circuits with sufficient voltage headroom to maximize rejection.
- Above 4-5mA of current, consider cascode type circuits. At several tens of mA, this should be considered mandatory for good performance.
- For any current regulator circuit, minimize capacitance in whatever active devices are used. This will enhance high frequency noise rejection and minimize the possibility of high frequency intermod.

If I were asked to recommend which of the many current regulators described here to use, I'd try to keep it as simple as possible. The maximum bang-for-the-buck is the cascode LM317 + MOSFET of **Fig. 13C**, assuming your current requirement is 40mA or less. This one

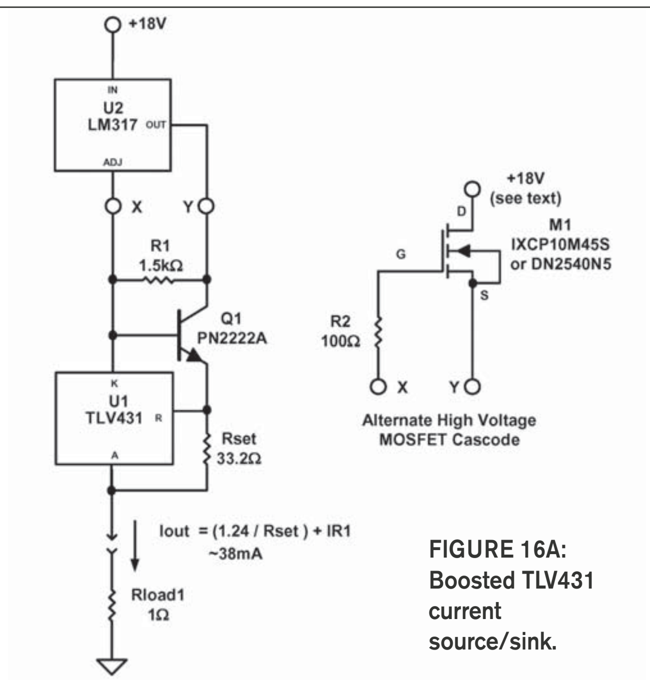

FIGURE 16A: Boosted TLV431 current source/sink.

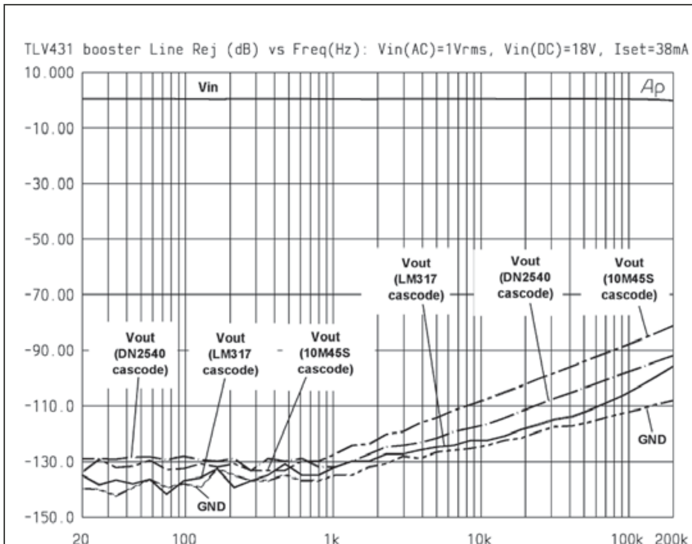

FIGURE 16B: Performance of the boosted TL431 as a 38mA current source or sink ranges from good to excellent, dependent upon the cascode device chosen.

worked great for me within a 12mA current feed for a 24V shunt regulator. For higher currents, the **Fig. 16A** circuit is both flexible as a source or sink and capable of much higher currents when Q1 is appropriately selected. For low currents of just a few mA, single and/or cascoded JFETs are likely best (*Fig. 8A*). Or, you could select the reference diode circuit of *Fig. 6A*.

SOME HOMEWORK ASSIGNMENTS

One manuscript reviewer asked about *very* high output currents, i.e., several amps. My general answer is that yes, this should be possible with minor revisions.

I said, "You could use **Fig. 14A** with a conventional N-channel MOSFET replacing Q1/Q2. The 1-2V Vgs of a MOSFET will bias the K pin of U1 roughly 1-2V above the R pin (but don't forget a 100Ω snubber in the MOSFET gate circuit). This should work OK for ampere outputs. Pick the FET for the required current, voltage, power, and, preferably, lowest C. I'm sure you have a favorite here. One possibility might be the Fairchild FQP4N20L, a TO-220 part, available from Mouser. I think I'll put this idea in at the end of the Part 2, as a reader 'Homework' assignment."

So there you have one assignment for some fun experiments. Let us know what you find out with this MOSFET-boosted current source idea!

Another assignment is to explore a hybrid vacuum tube/solid-state current regulator. For example, you could also use **Fig. 13C** with a power triode in place of M1 (grid to U1-OUT, cathode to U1-IN, and plate to the input voltage). The LM317L might be a possible candidate for U1.

I'd be very interested to hear about your results with these ideas. Write me at *audioXpress* via conventional mail, or contact me via my website, www.waltjung.org/index, and happy current sourcing and sinking! **aX**

Acknowledgments

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Part 1

REGULATORS FOR HIGH-PERFORMANCE AUDIO

By Walt Jung

Power-supply regulation is an important issue when highest audio system performance is your goal. While this may seem obvious, power-supply limitations can be quite pervasive. Power supplies can, to some degree, affect virtually all audio stage performance, since supply rails are a hidden source of crosstalk and noise coupling.

For example, at very low frequencies (<100Hz), op amps have very high power-supply rejection. However, this degrades in the audio range and may also be different for the plus/minus supplies. In another case, wideband IC video amplifiers and some discrete circuits generally have supply rejections which are flat to much higher frequencies, but are also lower—as much as 60–70dB—at the start. Given that they will inevitably have finite power-supply rejection, all amplifying stages can thus benefit from better power supplies.

This article focuses on relatively low-level power supplies, those suitable for analog preamp and line-level stages, digital logic systems, and other +5 to ±20V and less applications with current drains of 300mA or less. It emphasizes extracting the highest performance from both the power-supply regulation stage and its application within a system. The article's scope includes detailed design, testing, and performance information on plus/minus output regulators, from simple fixed and adjustable three-terminal types up through more sophisticated regulators, both discrete and IC-based.

I discuss comparative test performance data for *line rejection* or *LR, noise,* and *output impedance* or Z_O for a wide variety of regulators, but do not include more basic design information on the raw DC supplies to feed these regulators. The article provides an overview perspective of power regulation for audio uses.

Background

This magazine and others have already focused serious attention on power-supply regulation.[1-5] One of the more highly developed regulator models in use since the '60s features a buffered op amp with a self-regulating reference diode in a DC control loop[6] (see references therein, plus those of Chapter 4 of reference 2). However, for audio-oriented use, a more notable variation on this theme is the original Sulzer regulator,[3] and its subsequent relatives.[4,5] These regulator topologies are superb performers, but at the expense of greater complexity as compared with the simpler three-terminal types. It's still unclear, however, how distinct regulator types are differentiated in terms of various performance aspects.

Output impedance tests have covered a wide range of regulators, and generally yield useful comparisons.[7] However, while undoubtedly important, Z_O testing alone simply does not reveal the entire story of regulator performance. With the advent of digital technology and increased awareness of RFI, power regulator LR versus frequency becomes increasingly important and is one additional performance parameter worthy of more complete assessment. Similarly, regulator wideband noise is another useful performance-quality indicator.

With a good working knowledge of these parameters, you can choose a regulator type most appropriate to your desired quality level. This article addresses how to measure these performance factors in general, using sensitive tests with high-performance lab equipment. The tests are useful for both evaluating standard circuits and optimizing newer designs.

While the Sulzer regulator topology has become a standard for high-performance use, newer op amp devices introduced since 1980 offer potential for even further improvements in certain areas. In addition, very careful selection of the pass transistor allows currents well

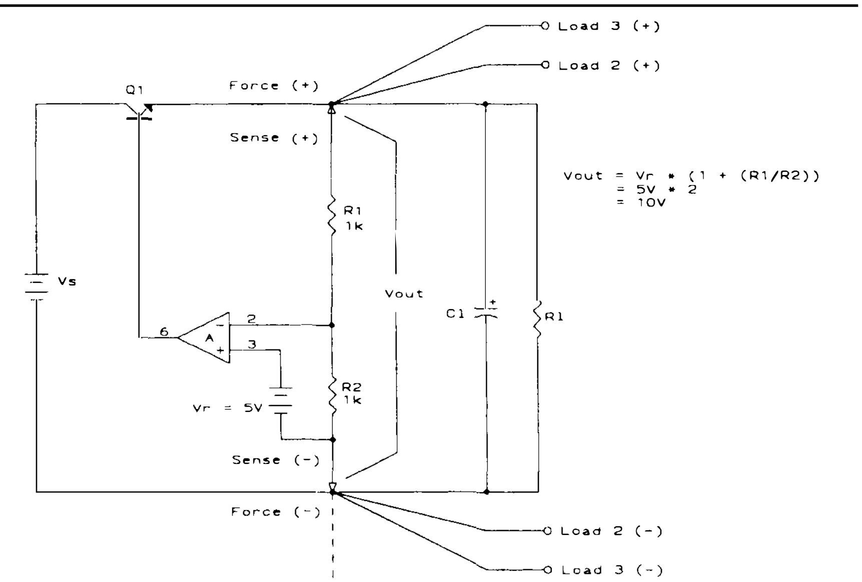

FIGURE 1: Series voltage regulator.

ABOUT THE AUTHOR

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above the ≅100mA of the original design.[5,8] And, lower-voltage headroom regulator designs allow you to employ power-efficient low-voltage regulators within logic systems, where they have a positive impact on jitter and phase noise performance. Finally, in conjunction with cleaner and more noise-free regulation, quieter raw DC supplies also help improve overall system performance by lessening the burden on the regulator for noise rejection.[8,9]

Regulation Basics

A brief review of regulation fundamentals as applicable to audio power supplies is helpful before discussing circuits and their testing. *Figure 1* shows a very general schematic of a *series*-type positive output voltage regulator, so-called because the NPN control transistor Q1 is in series with the load. To obtain a negative voltage, simply reverse V_S and V_r and use a PNP device for Q1.

This circuit produces a stable regulated output voltage V_{OUT}, which is programmed by resistors R1 and R2, given a stable reference voltage V_r. *Figure 1* assumes a V_r of 5V, so 1kΩ R1-R2 values result in a 10V output. The high-gain amplifier A (either an op amp or a discrete circuit) compares the fixed reference V_r against a sample of the output, as selected by the divider resistors. With sufficiently high gain in A, the control loop adjusts the conduction of Q1 so the output remains stable, relatively independent of both load current and variations in V_S.

The key to high performance is the closed feedback loop around Q1, which automatically drives the base to a level that maintains the lowest possible DC and AC errors at the output. This is in distinct contrast to the simpler "emitter-follower"-type power supply (which is technically not a true regulator, since there is no overall feedback and closed loop control). Such circuits simply drive a pass device such as Q1 from a stabilized DC source such as a zener, and the load-dependent V_{BE} variations of Q1 appear at the output.

As you might expect, various *Fig. 1* circuit subtleties impact audio performance. First and foremost, you should note that this regulator only maintains the output voltage constant between the two points sensed by the divider. In this case, the applicable nodes are labeled "Sense," further denoted by their arrowhead connection to the positive and negative rails (amplified by the label "V_{OUT}"). *Note that connections of loads to any other point than these will result in some reduction of performance.* Thus the "star" power connections are preferable for additional loads 2 and 3 (or more).

Regardless of the relative regulation quality, the best performance will always be obtained with the power distribution feeds taken from the sense points. By the same token, the most accurate performance measurement can only be made at these same sense points. At other points along the connecting wires, load-current-proportional voltage errors will occur. The driving connections to the sense nodes, labeled "Force (+)" and "Force (-)," connect to the control transistor Q1 and power supply return, respectively.

In terms of regulator performance, the major specifications mentioned above—LR, noise, and Z_O—are further qualified by how they vary with frequency. In addition, it makes sense to optimize the circuit for a minimum of operating headroom, so you can maintain a useful output for values of the unregulated input V_S just a few volts greater (or less) than V_{OUT}. The specifications for Q1 determine the maximum output current, maximum input voltage, and, to some degree, the operating bandwidth.

Devices most useful here are common-collector connected bipolar power types, which have the lowest output impedance before feedback and a reasonably low-voltage headroom (<1V). While MOSFETs might seem attractive on the surface, their very high voltage thresholds severely limit attempts for low dropout designs, and specified low ON impedances are available only with ampere level conduction, not at the 100–200mA level of these audio regulators. By contrast, an emitter follower's (open loop) output impedance is roughly 1Ω at a current level of only 25mA.

Assessing Performance

Most other regulation performance attributes are determined by the design details of amplifier A. This circuit can be either discrete transistor or op-amp-based, and I've included developed examples of both. Or, the entire regulator function can also be completely integrated, such as in the popular three-terminal IC regulators.

Obviously, a complete regulator in a three-pin package is both space efficient and economical, making such designs attractive where simplicity is paramount. However, like most other things in life, with voltage regulators you get what you pay for (in complexity terms). We'll discover that the more complex regulator circuits definitely offer the highest possible performance on an absolute basis.

To assess regulator performance as applicable to audio use, three test setups are required to exercise a regulator in terms of LR, noise, and Z_O characteristics. These tests are shown in block diagram form in *Figs. 2a–2c*, and are described briefly below. All three tests are derived from my earlier test series developed for reference ICs,[6] which, in turn, is partially based on the impedance measurement procedures of Brimacombe.[10] The common theme of these tests is the use of a

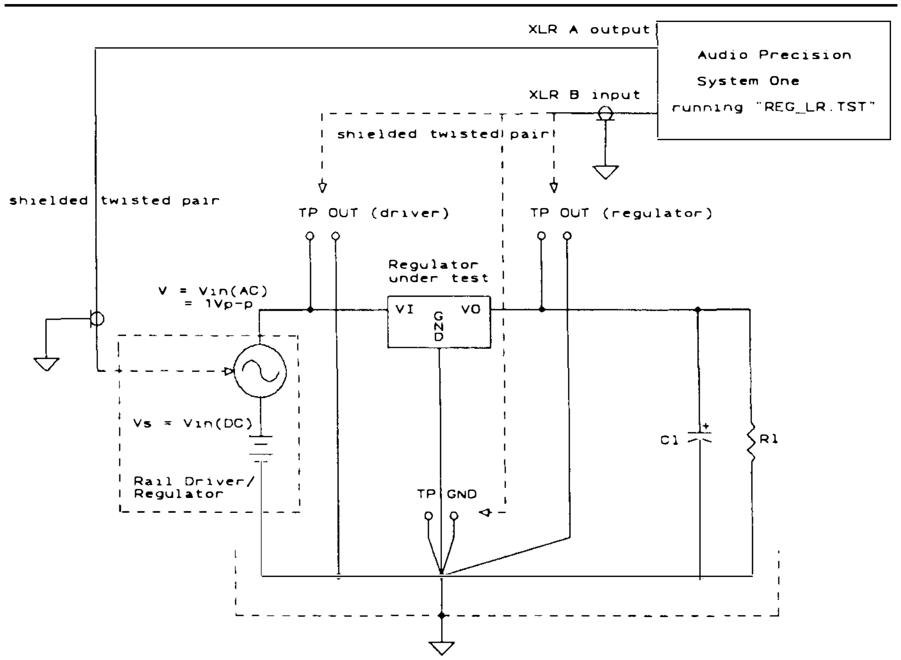

FIGURE 2a: Line rejection test.

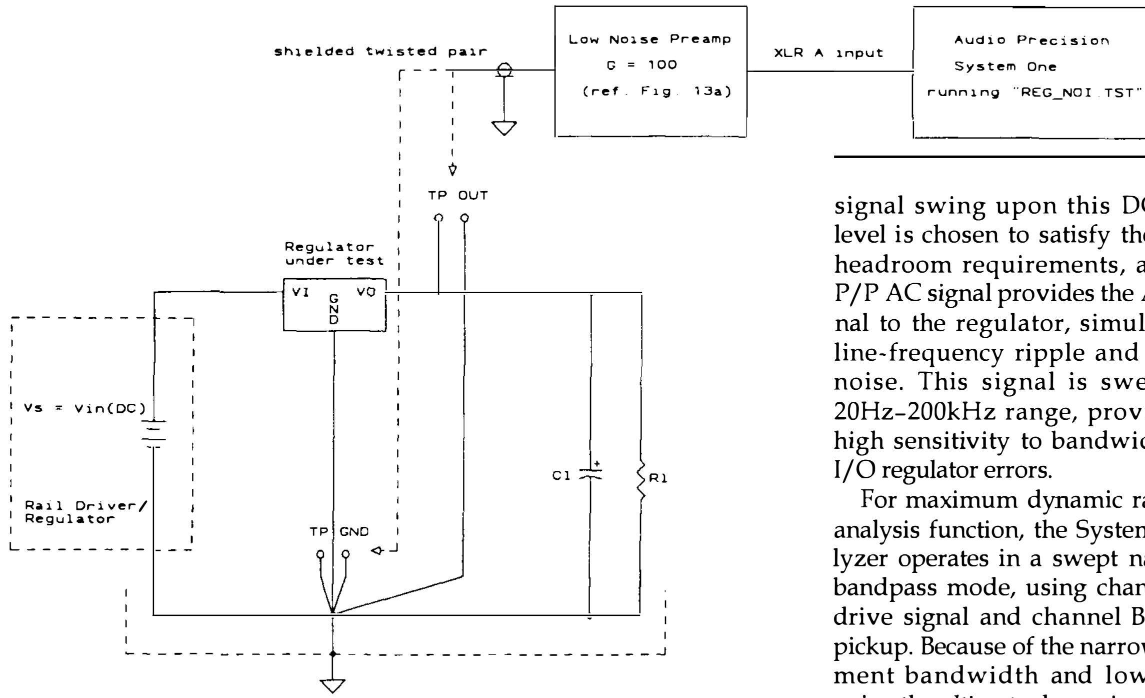

FIGURE 2b: Noise test.

sensitive high-resolution test instrument, the Audio Precision System One (PO Box 2209, Beaverton, OR 97075-3070, 503- 627-0832).

Line Rejection Tests

Figure 2a shows the setup for exercising a regulator for line rejection versus frequency. The DC input to the regulator, V_S, is taken from a positive or negative rail *driver/regulator* stage. This stage functions as a low-impedance source for both AC and DC and produces a nominal regulated DC voltage of 18V, with polarity suitable to the regulator under test.

You can superimpose a 1V P/P AC signal swing upon this DC. The DC level is chosen to satisfy the regulator headroom requirements, and the 1V P/P AC signal provides the AC test signal to the regulator, simulating both line-frequency ripple and wideband noise. This signal is swept over a 20Hz-200kHz range, providing very high sensitivity to bandwidth-related I/O regulator errors.

For maximum dynamic range in the analysis function, the System One analyzer operates in a swept narrowband bandpass mode, using channel A as a drive signal and channel B for signal pickup. Because of the narrow measurement bandwidth and low analyzer noise, the ultimate dynamic range of this test approaches 140dB. Consequently, very careful screening and guarding of the test setup is necessary to minimize contamination of the output signal and to take full advantage of the test system's dynamic range.

If you eschew such precautions, then the very best regulator in LR terms may be indistinguishable from those more pedestrian. More importantly, you will be unable to fully optimize a high-per-

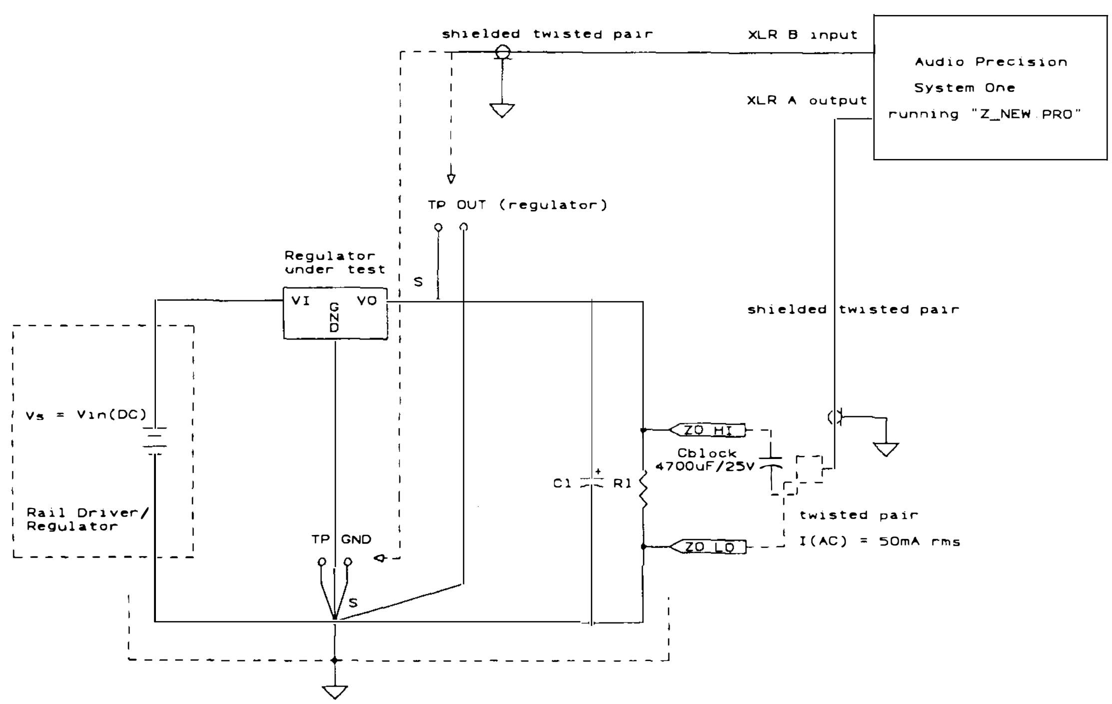

FIGURE 2c: Output impedance test.

formance design, unless the test setup is sufficiently sensitive.

In these tests, a grounded guard trace was used on three sides around all regulator circuits, which were built over a heavy ground plane. I also used shielded twisted pair construction on the analyzer I/O cabling, with plug-in tips for quick access to various sensitive circuit nodes. The TP OUT driver output two-terminal test jack provides a 0dB, 1V P/P input reference level, which calibrates the analyzer for full scale. All measured regulator output signals are referred to this 0dB level as it drives the input of the regulator under test.

The AC level measured at the TP OUT regulator jack represents the degree of isolation provided by the regulator, and is easily scaled and displayed in decibel versus frequency by the test analyzer. The TP GND jack provides a grounded-input, minimum signal reference for the analyzer.

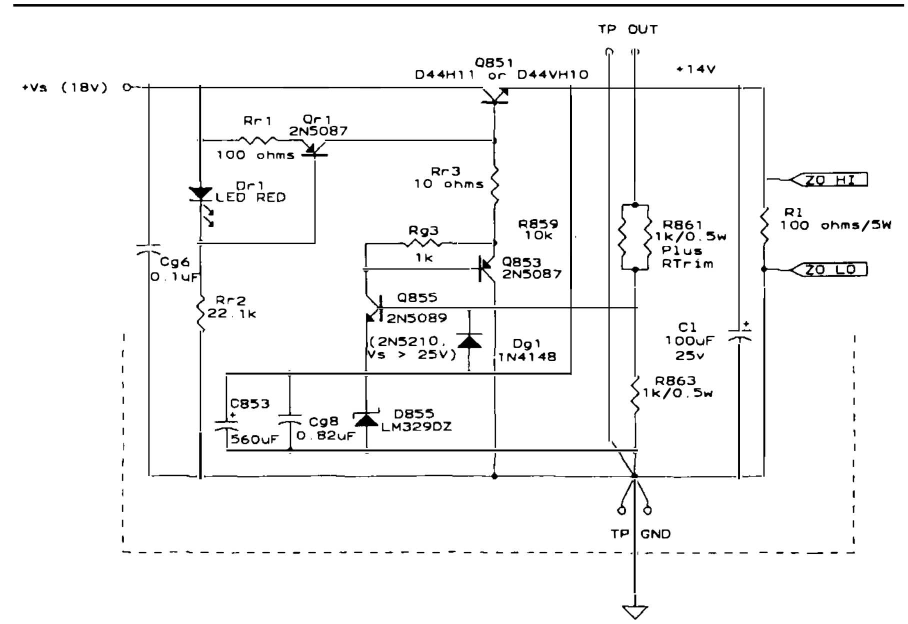

FIGURE 3: POOGE 5.51 positive regulator.

To allow relatively easy comparison among various regulator types, all regulators under test used setups similar to this, regardless of their circuit topology. Thus with similar loading and input conditions, direct comparisons of measured performance between different circuits is possible. This test and the others operate with a common loading for all regulators of C_l = 100μF and R_l = 100Ω, except as otherwise noted.

With both positive and negative regulator circuits available, both positive and negative rail drivers are required to perform the LR tests quickly and accurately. I describe these in more detail later.

Noise Tests

Figure 2b shows the setup for testing regulator output noise. No AC input signal is provided, and the appropriate DC voltage V_S is supplied by the rail driver/regulator. For most of the circuits tested, this is also 18V, with polarity as appropriate. The only output signal is the AC noise level, as measured at the TP OUT test point of the regulator under test.

Because many regulators have very low noise levels, a preamplifier circuit is required to raise the AC output signal to a level where it can be readily measured by the analyzer. For this a balanced input, gain of 100 circuit was used, coupled to the regulator under test by shielded twisted pair cabling. This preamp's output is coupled into the analyzer, which operates in a swept, narrow-band bandpass mode.

A given noise test provides a 100× scaled display of regulator output voltage noise as a function of frequency. A fundamental limitation of this setup is that the measurement preamp noise cannot be distinguished from the regulator noise, when they are of the same order. The preamp circuit's noise is about 2.6nV/√Hz, so this only becomes an issue for regulators with noise levels of about 7nV/√Hz or less. I describe the low-noise preamp circuitry in greater detail later.

Output Impedance Tests

Figure 2c shows the setup for testing regulator output impedance (Z_O). As with the noise test, an appropriate rail driver/regulator supplies an appropriate DC voltage V_S. To measure output impedance as a function of frequency, the analyzer is programmed to produce a constant 2.5V RMS behind a 50Ω resistance. This results in an AC current flow or I (AC) of 50mA RMS for load impedances low with respect to 50Ω.

An appropriately polarized 4,700μF DC blocking capacitor couples the AC test current directly across the load resistance of the regulator under test, through twisted pair wiring. This completely balanced signal transmission method was found necessary for the very highest-resolution measurements, where the equivalent amplifier input voltages to the analyzer are around the microvolt level in the highest-performance circuits.

In this test, the regulator is called upon to absorb the test signal AC current to maintain the output voltage at the DC design level. Since an AC signal is bipolar in nature, the regulator can only totally absorb this signal if it is prebiased to a DC load current higher than that of the highest AC signal peak (about 70mA). For these tests, all regulators were operated with DC loads of 100mA or more.

Of the three tests, this one is the most difficult in terms of wide dynamic range implementation. The location of the TP OUT test points must be accurately fixed at the true physical/electrical sensing points of the circuit (*Fig. 1*). No load currents must be allowed to flow in the wires to the test points; otherwise, the advantages of this four-wire sensing will be lost. This point is amplified by the "S" notations on the *Fig. 2c* diagram.

Overall sensitivity of this test setup is such that equivalent impedances of less than 10μΩ can be resolved at low frequencies, and 1mΩ or less at 100kHz. The respective figures list the test software, which is available by simply sending me a formatted 3.5" MS-DOS disk, *with a mailer including return postage.*

POOGE 5.51 Regulators

The discrete regulator of *Fig. 3* is an enhanced version of the POOGE 5 regulator described in Part 1 of Gary Galo's article.[11] For this article, it provides an example of a medium-to-high-performance discrete circuit regulator, and aside from the present comparative discussions, is practical and useful as shown. Those using this regulator must add some protection against overcurrent (this also applies to all others, with the exception of the internally protected IC types). Since there is no internal current limiting, a simple series fuse on the unregulated input side is sufficient protection.

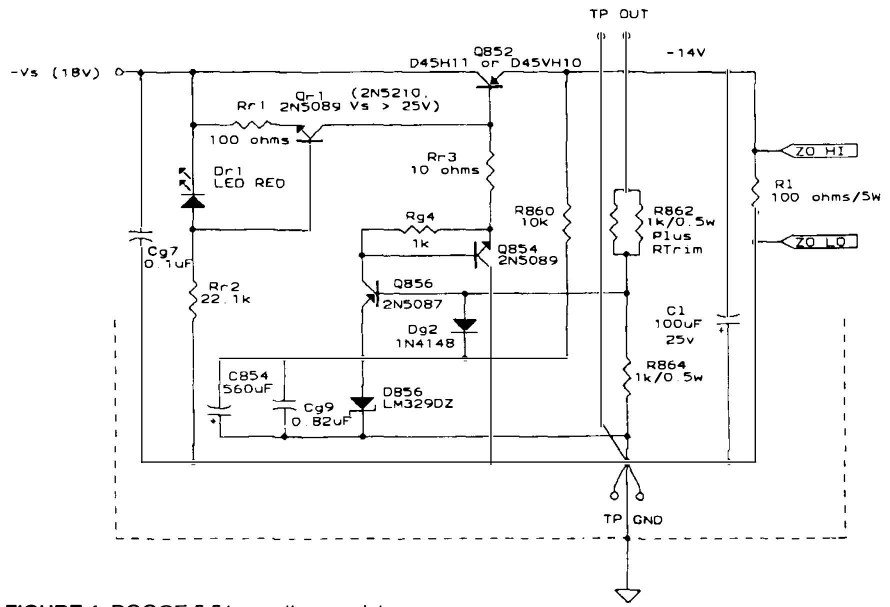

FIGURE 4: POOGE 5.51 negative regulator.

Close comparison of this schematic with the original version shows two main differences. One is Q851, either a D44H11 or D44VH10, both improved pass transistors as described in POOGE 5.5.[8] The other is a refined current source drive for Q851, Qr1, and the associated parts (such as Rr1).

The new current source functionally replaces the selected 2N5458 JFET in the original version, and allows improved performance with lower I/O voltages, i.e., lower *dropout*. While the simpler POOGE-5-style JFET current source works fine with medium-to-high I/O voltages, it requires 4–5V of bias to achieve highest LR. Since the combination of these two changes enhances the performance, I designated these new circuit versions as POOGE 5.51.

The LED-biased bipolar transistor source works well down to 1.5V of dropout voltage. This current source's output is set by Rr1 at approximately 10mA, which allows regulator output currents of up to 500mA, with a Q851 β of 50. Resistor Rr3 provides additional stability for Q851 in the presence of capacitive loads. When connected as shown, Rr3 has no negative effect on overall dropout voltage.

The negative version of the new regulator (*Fig. 4*) works identically in concept to *Fig. 3*, with, of course, the obvious polarity reversals and transistor complements. In both forms of the circuit, use the 2N5210 in place of the 2N5089 when input voltages are 25V or greater. For these tests, I set up this regulator (and most of the others, except as noted) with an output of 14V (trimmed just as described in Gary's POOGE 5.5 article) and a 140mA load, R_l.

Circuit Layout

As noted above, I built this circuit (and the others) in a small "cell" area surrounded on three sides by a grounded guard trace (#16 gauge, dotted lines in the schematic). In addition, a double-sided circuit board with a ground plane was used. (Breadboards for these tests are "IVANBOARD," an 8.5" × 11" RF design breadboard using a 0.1" grid surface mount pattern over a 2 oz copper ground plane.)

The standard dual test points are provided for ease-of-measurement via plug-in twisted pair cable connectors. One is at the TP GND reference point, provided at the circuit's common point, the common physical/electrical junction of reference diode D855 and divider resistor R863. The other is the TP OUT sense point, electrically connected between the above described TP GND point and the output node connection of divider sense resistor R861.

In the test strategy, I included a given test cell's TP OUT connection for all three tests, and used TP GND as a dynamic range and S/N reference check. For the Z_O tests, ZO HI and ZO LO provide access for the 50mA AC current test signal connections directly to R_l. I used similar test points and connections in all regulator test circuits.

Since the three-terminal regulator types tested are much simpler in their application, they are not shown in schematic detail. All types tested were by the original manufacturers, using the TO-220 package and carrying either a 1.5 or 3A rating. Note that many different versions of these regulators are available, some with much lower maximum current ratings. I did not test these types, but anticipate that those with lower current ratings (and associated higher output impedance) will not likely exceed the performance of the 1.5/3A versions.

The fixed 15V types (LM7815 and LM7915) are connected in their standard mode, with output loading of C_l = 100µF and R_l = 100Ω. The adjustable three-terminal regulators (types LM317, LT1085,

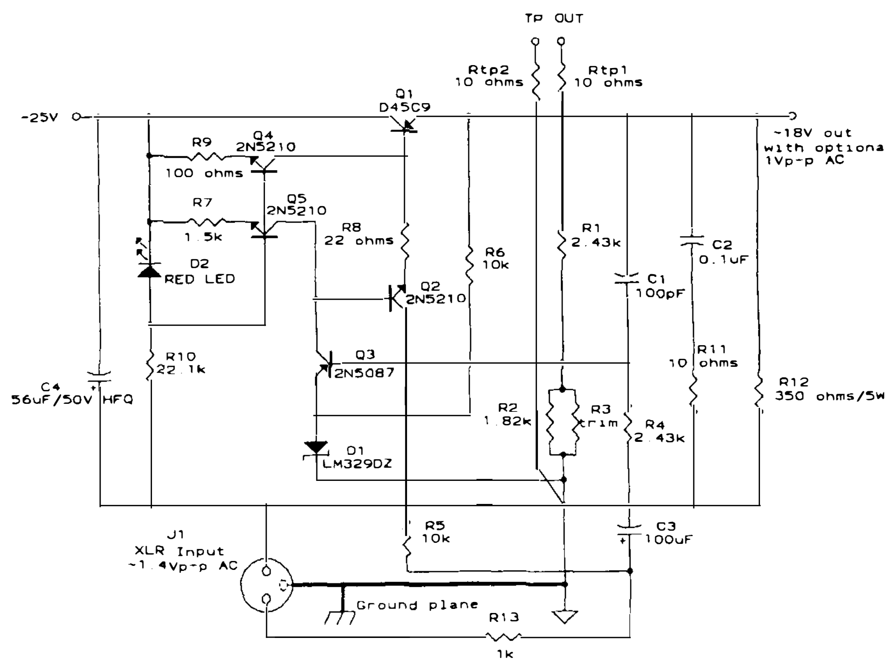

FIGURE 5: Negative rail driver/regulator.

LM337, and LT1033) are connected with an OUT-ADJUST pin resistance of 1kΩ and an ADJUST-GND pin resistance of 10kΩ, which programs them to a nominal 14.2V. These also used a C_{ADJ} bypass capacitance of 100µF, plus the loading of C_l = 100µF and R_l = 100Ω. It is worth noting that this relatively high divider impedance works to advantage for audio applications, since a given size C_{ADJ} capacitor is more effective across a 10kΩ resistor than with a 1kΩ value.[12]

Rail Driver/Regulators

For a controlled AC source test environment, the discrete regulators of *Figs. 3* and *4* are adapted to operate as low-impedance rail drivers. From an unregulated ±25V DC source, these drivers provide a regulator signal source environment which allows a fixed DC input level of ±18V, upon which you can optionally superimpose a 1V P/P AC test signal. In the line rejection tests, the ±18V output rails carry this AC signal, which is swept from 20Hz–200kHz.

This requirement demands a small-scale power amplifier, due to the fact that the various test regulators in many cases require a minimum value 0.1µF input bypass capacitor for stability.

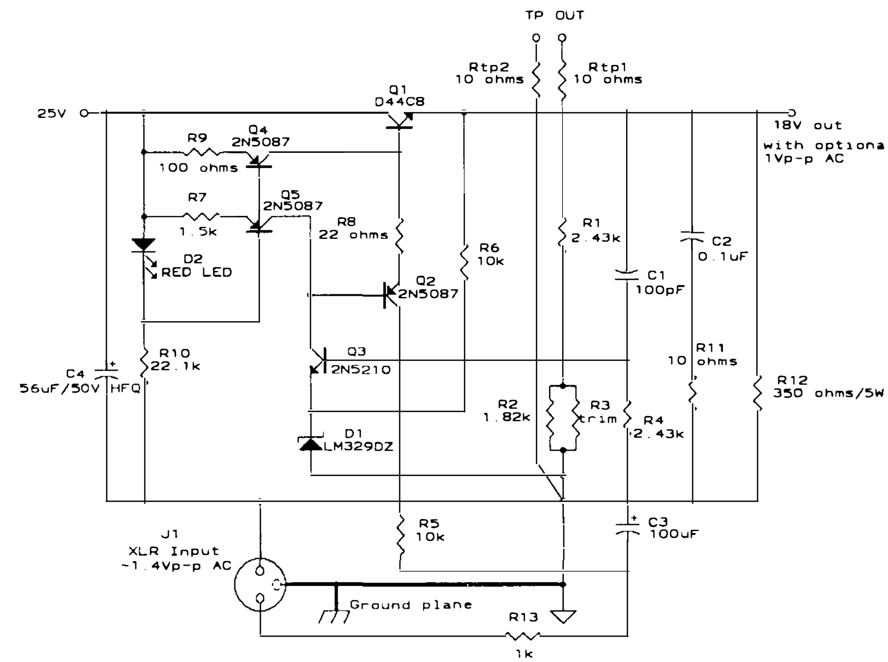

FIGURE 6: Positive rail driver/regulator.

Driver circuits suitable for negative and positive 18V plus AC outputs are shown in *Figs. 5* and *6*, respectively.

The voltage slewing to maintain a flat-frequency response into the 0.1µF load capacitance of a test regulator requires a substantial standing current in Q1, part of which is provided by the brute force load R12. As with the POOGE 5.51 plus/minus regulators, the

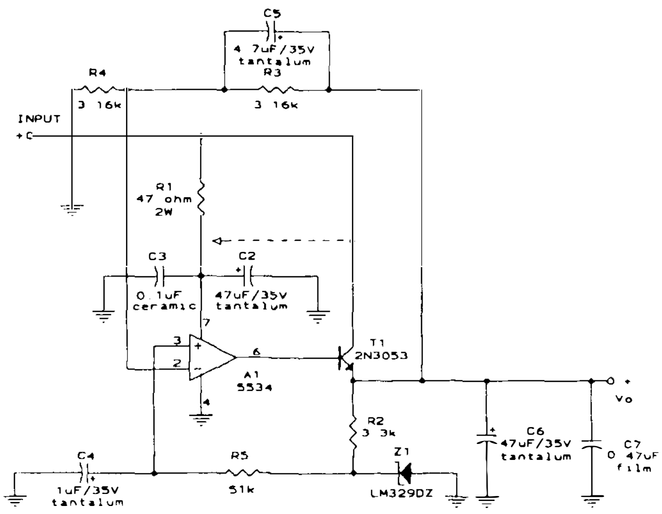

FIGURE 7a: Positive output Sulzer regulator.

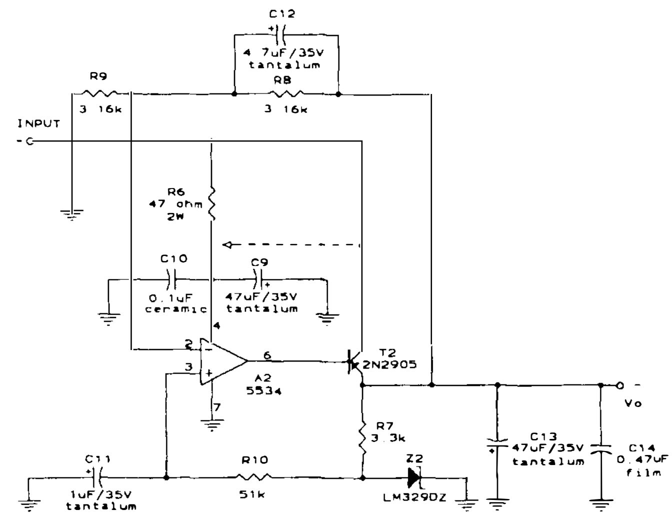

FIGURE 7b: Negative output Sulzer regulator.

output divider ratio and a reference voltage of ≅7.5V (6.9V plus 1Vbe) set output voltage. In this case, the bottom resistance R2 is trimmed by shunt R3 to eliminate interaction with the AC gain. R3 is trimmed for an output of 18 ±0.1V with the driver/regulator loaded. Loads of up to 500mA are possible. AC test signals at input J1 are passed to the output bus with less than unity scaling, since R1 < (R4 + R13).

In these drivers, the TP OUT test plug is used to verify the 0dB, 1V P/P 20Hz–200kHz AC swept reference signal as it is applied to the specific regulator under test for LR. The Audio Precision System One's function key F4 provides calibration to a specific measured output level for the 0dB amplitude reference for these test conditions. This calibration feature allows a measurement's 0dB reference to vary somewhat in absolute terms about the same nominal level (1V P/P or other), but still references all subsequent readings to this level, which is an extremely useful trick.

Another difference of this circuit in relation to the POOGE 5.51 regulators is the LED/bipolar transistor current source load Q5, which allows higher amplifier gain. This provides the driver with an output impedance measured at TP OUT of less than 5mΩ below 100kHz.

Some initial positive rail regulator testing was accomplished using a 317-type regulator as an 18V regulator/driver, with the AC test signal coupled into the normally grounded C_{ADJ} capacitor. While much simpler than the discrete circuit driver of *Fig. 6*, this setup was not completely satisfactory, since spurious resonances occurred at certain frequencies.

Sulzer Regulators

The original Sulzer regulators[3] for positive/negative supplies are shown in *Figs. 7a* and *7b*, respectively. These are reproduced almost identically to the original versions, with a couple of small but important exceptions. One change is the connection of the pass transistor's collector, which is taken directly back to the input, as opposed to passing the output current through R1 (or R6). This step allows more effective decoupling of the op amp rail, and will increase available

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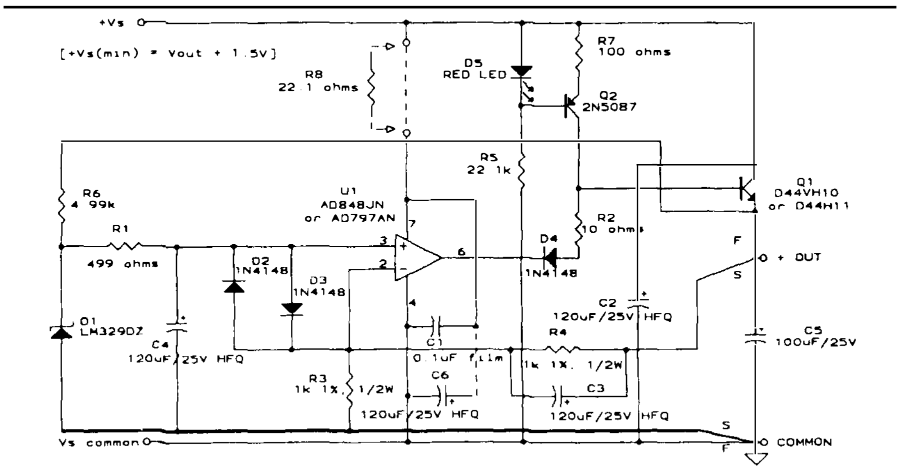

FIGURE 8a: Positive output, low-noise, low-dropout regulator. For this figure and *Figs. 8b* and *9*, unless otherwise specified, all resistors are ¼W, 1% metal film. A heatsink is required on Q1 for Pd >0.5W. In this figure and *Fig. 8b* only, minimize lead length from unregulated input and return. "S" indicates Sense lead; connect at point-of-load for best regulation

headroom. It was also advocated by Sulzer in the "revisit."[4]

That article also suggested the use of "high quality zener, composed of an integrated circuit," which was attributed to Joe Curcio. Presumably, this was the lower noise LM329 IC zener as shown here and used in these tests (the LM329 was used also by Breakall, et al[7]). The only other variations are minor ones in resistance values, based on available values (R5/R10 and the output divider). With equal divider resistors, the circuits produce about ±14V.

Sulzer rated the original circuit at 100mA, so in this case I set load resistor R_l at 150Ω (not shown). I used no additional C_l, beyond the values for C6 and C13, and built up the test circuit in the manner described above, being careful to locate the TP OUT test points appropriately at the sense points, to connect ZO HI and ZO LO to R_l and control (well) the high-current paths and grounding.

Low-Noise, Low-Dropout Regulators

Over the last two years, I've been working on a new family of high-performance, op-amp-based regulators. This work started shortly after the publication of the POOGE 5 regulator,[11] and sought to examine pass devices, amplifiers, and overall topologies to improve their performance. My goals were to extend a design allowing use at low output voltages such as 5V and to enhance dropout performance so DC regulation could be maintained down to V_{OUT} +1.5V (1.5V dropout). In addition to achieving Z_O performance similar to the already excellent Sulzer regulator, I wished to push LR and noise performance to levels as high as possible.

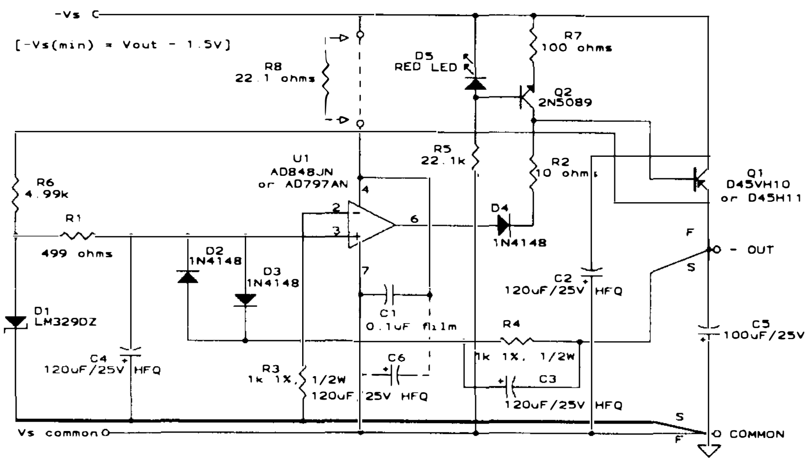

FIGURE 8b: Negative output, low-noise, low-dropout regulator.

In general all of these goals were met, but the route to the final result has been quite an adventure. The testing has been among the more fascinating parts of the development, and, as I'm sure most will agree, it is full of surprising results. This will not be at all obvious just by examining the circuits, but it may become clearer as I explain the various test results.

Some audio designers simply prefer discrete circuits for regulation. Even given this preference, however, premi-

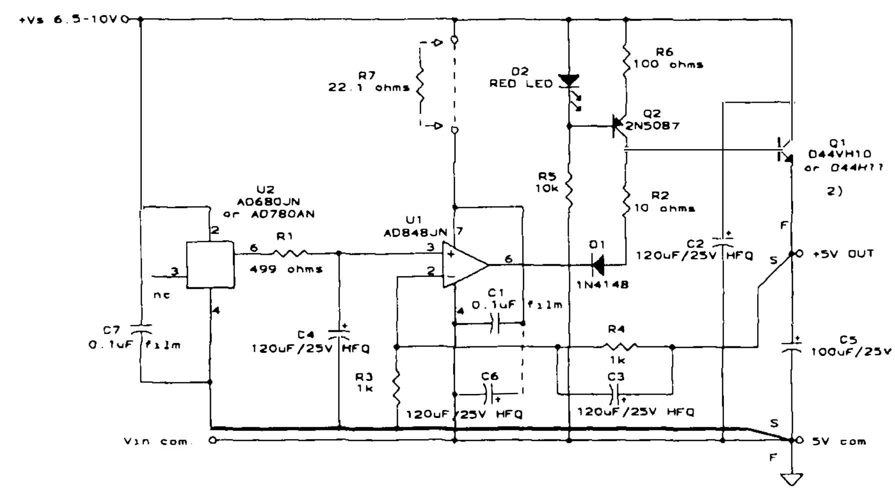

FIGURE 9: 5V, low-noise, low-dropout regulator. Minimize lead length from +Vs and return.

um-level results for Z_O, LR, and noise takes both a skilled designer, plus lots of parts. For example, the POOGE 5.51 circuits illustrate this problem. The performance is excellent for their relative simplicity, but if you wish to lower output noise, for example, you will not accomplish this simply.

You'll need additional parts to implement zener noise filtering and buffering, which may have a negative effect on Z_O performance. It also becomes extremely difficult to get a comparable level of all-around performance working at a 5V output level, since the reference voltage needs to be <5V, a level which typically is more noisy because of the physical limitations of available bandgap-based voltage references.[6]

This leaves the topology of choice a low-noise, wideband differential input amplifier which operates down to 8–9V (or less), can be configured for low dropout, and has high inherent LR properties. This is a tall order for a simple discrete amplifier, and most would agree that there might be quite an increase in parts count over *Fig. 3* or *4*, just to implement such an amplifier. At this point, a very carefully selected op amp sounds much more efficient. Unfortunately, the 5534 of the Sulzer design does not operate well at low voltages, so this leaves newer devices as potential candidates.

The circuits of *Figs. 8a* and *8b* represent the new positive/negative design solutions, the result of more than a few iterations with amplifiers and related components. Superficially, these designs look much like the Sulzer circuit, but with both major and minor differences. The circuit discussion that follows is in terms of the positive regulator of *Fig. 8a*, but the negative form works in a similar fashion, given polarity normalization.

The Circuit

A major departure from a straightforward U1/Q1 drive topology is the use of a current source drive for Q1's base, composed of Q2 and the associated parts. As with the POOGE 5.51 circuits, the LED-biased current source allows Q1 to operate down to 1.5V or less of differential, lending the circuit low dropout features. The potential for a limited swing of the op amp to affect dropout is lessened by operating it in a current sinking mode, which is enabled via D4.

Typically, this circuit achieves dropouts of 1.2V with several hundred milliamp outputs, making it suitable even for logic supplies. Note that high regulator dropout is a serious issue for a logic regulator, where a 3V dropout can increase the power dissipated in Q1 to an intolerable level.

A second important change in the new regulator is the addition of amplifier input clamp diodes, D2-D3. These normally zero-biased diodes protect the op amp input stage, in the presence of ON or OFF transient voltage differentials greater than 5V. Such a large voltage can harm an unprotected input stage by breaking down either transistor's E-B junction.[13]

With a 7V reference voltage, the possibility that U1 can be damaged (or subtly degraded for noise) exists, if the input stage is allowed to break down differentially. The clamp diodes prevent this from happening, and should be used in this circuit for cases where the op amp does *not* have such diodes internally. The AD848 doesn't, so use D2-D3 as an ounce of prevention.

With lower-level reference voltages (such as 2.5V), you can eliminate the clamping diodes in many instances. As a general rule for most unprotected op amps, the worst-case differential transient error should be maintained <5V for safety.

While the presentation of the reference voltage to the op amp is similar to the Sulzer configuration, as is the feedback network, the general impedances are lowered and made symmetrical for both AC and DC. The matched 500Ω DC source resistances enhance overall DC stability at little or no cost, and the 100% AC feedback around R4 lowers both output impedance as well as noise.

This latter technique, one hallmark of the Sulzer configuration, is a significant key to achieving the highest possible performance with a given op amp. It allows the net regulator output noise to approach that of the op amp itself, plus the filtered noise level of the reference input at Pin 3. With an ultra-low-noise op amp such as the AD797, 1kHz output noise levels approaching 1nV/√Hz are possible.[14] To realize the highest possible attenuation in the single-section reference noise filter, a low-ESR capacitor is used for C4, a 120μF/25V-type HFQ. A relatively high voltage rating also helps lower leakage, as well as ESR.

Network Advantages

With wide-bandwidth op amps for U1, supply bypassing is critical for stability. The small RF-quality film bypass C1 is located close to the device pins, and is mandatory. The minimal operating hookup consists of just C1 at U1, along with C2 close to the collector contact of Q1. Optionally, you can use extra noise filtering via R8 and C6 to increase the high-frequency supply rejection of U1.

The net advantage of using this network depends upon several factors. One is the specific part used for U1; another is whether a positive or negative output is being implemented (since the plus/minus supply rejection of many op amps differs). As shown, the corner frequency is about 60Hz, and while not absolutely necessary, a low-ESR HFQ type for C6 allows greater HF noise rejection working against the relatively low value of R8. Finally, this network increases the DC/LF supply impedance seen by the op amp, and you should apply it very carefully to op amps with less than 100dB of supply rejection (such as the AD848). Needless to say, substitutions of op amps in these circuits are strongly discouraged.

To set up this regulator for voltages other than the nominal $2 \times V_r$ or $\cong$14V, change the R3-R4 resistors as shown generally in *Fig. 1*, keeping in mind the 6.9V reference voltage used. You can expect some trim when the loose tolerance "DZ" version of the industry standard 329 diode is used. However, the exact DC output voltage is not likely to be critical, except as it may affect dropout with a marginally low raw-DC supply. For test purposes, I loaded this circuit with the standard loading of $C_l = 100\mu F$ and $R_l = 100\Omega$ (not shown in these figures) and was careful, as with the Sulzer circuits, in the physical wiring/layout.

Closely related to the *Fig. 8a* circuit is the 5V regulator shown in *Fig. 9*, which evolved from reference 15. It operates with a lower voltage three-terminal 2.5V bandgap reference at U2, an AD680, or an AD780, but is otherwise similar to the *Fig. 8a* positive regulator. It is tested at a 300mA current level with V_{IN} = 8V, as suitable to logic systems.

In Part 2 we will more closely examine the three test setups used, in conjunction with the Audio Precision System One, to determine regulator performance. The differences between regulators easily stands out, making an optimum performance choice easy. □

Part 1

REGULATORS FOR HIGH-PERFORMANCE AUDIO

By Walt Jung

Power-supply regulation is an important issue when highest audio system performance is your goal. While this may seem obvious, power-supply limitations can be quite pervasive. Power supplies can, to some degree, affect virtually all audio stage performance, since supply rails are a hidden source of crosstalk and noise coupling.

For example, at very low frequencies (<100Hz), op amps have very high power-supply rejection. However, this degrades in the audio range and may also be different for the plus/minus supplies. In another case, wideband IC video amplifiers and some discrete circuits generally have supply rejections which are flat to much higher frequencies, but are also lower—as much as 60–70dB—at the start. Given that they will inevitably have finite power-supply rejection, all amplifying stages can thus benefit from better power supplies.

This article focuses on relatively low-level power supplies, those suitable for analog preamp and line-level stages, digital logic systems, and other +5 to ±20V and less applications with current drains of 300mA or less. It emphasizes extracting the highest performance from both the power-supply regulation stage and its application within a system. The article's scope includes detailed design, testing, and performance information on plus/minus output regulators, from simple fixed and adjustable three-terminal types up through more sophisticated regulators, both discrete and IC-based.

I discuss comparative test performance data for *line rejection* or *LR, noise,* and *output impedance* or Z_O for a wide variety of regulators, but do not include more basic design information on the raw DC supplies to feed these regulators. The article provides an overview perspective of power regulation for audio uses.

Background

This magazine and others have already focused serious attention on power-supply regulation.[1-5] One of the more highly developed regulator models in use since the '60s features a buffered op amp with a self-regulating reference diode in a DC control loop[6] (see references therein, plus those of Chapter 4 of reference 2). However, for audio-oriented use, a more notable variation on this theme is the original Sulzer regulator,[3] and its subsequent relatives.[4,5] These regulator topologies are superb performers, but at the expense of greater complexity as compared with the simpler three-terminal types. It's still unclear, however, how distinct regulator types are differentiated in terms of various performance aspects.

Output impedance tests have covered a wide range of regulators, and generally yield useful comparisons.[7] However, while undoubtedly important, Z_O testing alone simply does not reveal the entire story of regulator performance. With the advent of digital technology and increased awareness of RFI, power regulator LR versus frequency becomes increasingly important and is one additional performance parameter worthy of more complete assessment. Similarly, regulator wideband noise is another useful performance-quality indicator.

With a good working knowledge of these parameters, you can choose a regulator type most appropriate to your desired quality level. This article addresses how to measure these performance factors in general, using sensitive tests with high-performance lab equipment. The tests are useful for both evaluating standard circuits and optimizing newer designs.

While the Sulzer regulator topology has become a standard for high-performance use, newer op amp devices introduced since 1980 offer potential for even further improvements in certain areas. In addition, very careful selection of the pass transistor allows currents well

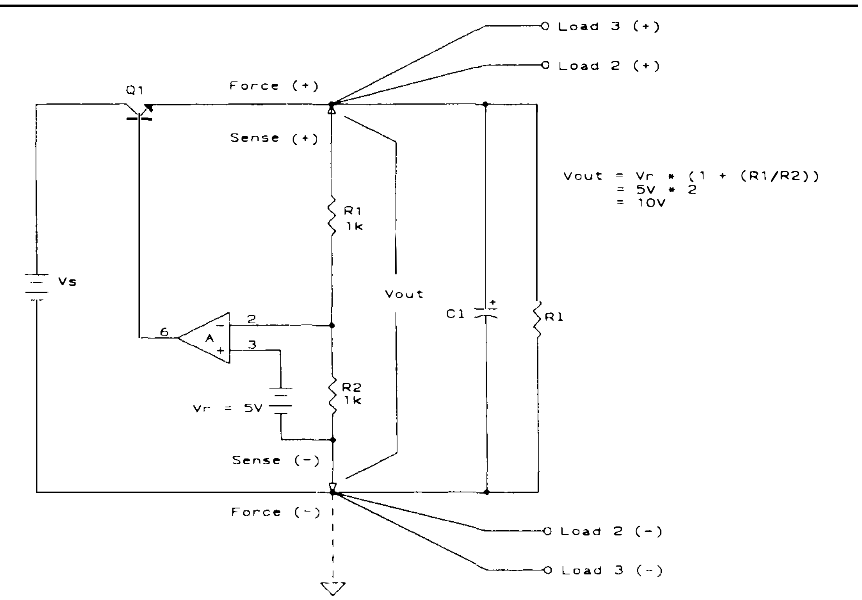

FIGURE 1: Series voltage regulator.

ABOUT THE AUTHOR

Walt Jung has been writing about electronics and audio topics for more than 25 years, in this and other publications. Among his more notable *TAA* articles have been the 1977 SID series and some of the POOGE series. As a practicing applications engineer and audiophile, he finds great reward in following audio circuits through the development, building, and bench testing phases to a final working form which serves musical signals. As a Corporate Staff Applications Engineer for Analog Devices, his duties involve technical articles and application notes, as well as seminar preparation and presentation. He has been interested in high-performance regulators for some time, and would appreciate feedback on this article by FAX at (410) 692-2158, or via E-mail, as Walter.Jung@analog.com.

above the ≅100mA of the original design.[5,8] And, lower-voltage headroom regulator designs allow you to employ power-efficient low-voltage regulators within logic systems, where they have a positive impact on jitter and phase noise performance. Finally, in conjunction with cleaner and more noise-free regulation, quieter raw DC supplies also help improve overall system performance by lessening the burden on the regulator for noise rejection.[8,9]

Regulation Basics

A brief review of regulation fundamentals as applicable to audio power supplies is helpful before discussing circuits and their testing. *Figure 1* shows a very general schematic of a *series*-type positive output voltage regulator, so-called because the NPN control transistor Q1 is in series with the load. To obtain a negative voltage, simply reverse V_S and V_r and use a PNP device for Q1.

This circuit produces a stable regulated output voltage V_{OUT}, which is programmed by resistors R1 and R2, given a stable reference voltage V_r. *Figure 1* assumes a V_r of 5V, so 1kΩ R1-R2 values result in a 10V output. The high-gain amplifier A (either an op amp or a discrete circuit) compares the fixed reference V_r against a sample of the output, as selected by the divider resistors. With sufficiently high gain in A, the control loop adjusts the conduction of Q1 so the output remains stable, relatively independent of both load current and variations in V_S.

The key to high performance is the closed feedback loop around Q1, which automatically drives the base to a level that maintains the lowest possible DC and AC errors at the output. This is in distinct contrast to the simpler "emitter-follower"-type power supply (which is technically not a true regulator, since there is no overall feedback and closed loop control). Such circuits simply drive a pass device such as Q1 from a stabilized DC source such as a zener, and the load-dependent V_{BE} variations of Q1 appear at the output.

As you might expect, various *Fig. 1* circuit subtleties impact audio performance. First and foremost, you should note that this regulator only maintains the output voltage constant between the two points sensed by the divider. In this case, the applicable nodes are labeled "Sense," further denoted by their arrowhead connection to the positive and negative rails (amplified by the label "V_{OUT}"). *Note that connections of loads to any other point than these will result in some reduction of performance.* Thus the "star" power connections are preferable for additional loads 2 and 3 (or more).

Regardless of the relative regulation quality, the best performance will always be obtained with the power distribution feeds taken from the sense points. By the same token, the most accurate performance measurement can only be made at these same sense points. At other points along the connecting wires, load-current-proportional voltage errors will occur. The driving connections to the sense nodes, labeled "Force (+)" and "Force (-)," connect to the control transistor Q1 and power supply return, respectively.

In terms of regulator performance, the major specifications mentioned above—LR, noise, and Z_O—are further qualified by how they vary with frequency. In addition, it makes sense to optimize the circuit for a minimum of operating headroom, so you can maintain a useful output for values of the unregulated input V_S just a few volts greater (or less) than V_{OUT}. The specifications for Q1 determine the maximum output current, maximum input voltage, and, to some degree, the operating bandwidth.

Devices most useful here are common-collector connected bipolar power types, which have the lowest output impedance before feedback and a reasonably low-voltage headroom (<1V). While MOSFETs might seem attractive on the surface, their very high voltage thresholds severely limit attempts for low dropout designs, and specified low ON impedances are available only with ampere level conduction, not at the 100–200mA level of these audio regulators. By contrast, an emitter follower's (open loop) output impedance is roughly 1Ω at a current level of only 25mA.

Assessing Performance

Most other regulation performance attributes are determined by the design details of amplifier A. This circuit can be either discrete transistor or op-amp-based, and I've included developed examples of both. Or, the entire regulator function can also be completely integrated, such as in the popular three-terminal IC regulators.

Obviously, a complete regulator in a three-pin package is both space efficient and economical, making such designs attractive where simplicity is paramount. However, like most other things in life, with voltage regulators you get what you pay for (in complexity terms). We'll discover that the more complex regulator circuits definitely offer the highest possible performance on an absolute basis.

To assess regulator performance as applicable to audio use, three test setups are required to exercise a regulator in terms of LR, noise, and Z_O characteristics. These tests are shown in block diagram form in *Figs. 2a–2c*, and are described briefly below. All three tests are derived from my earlier test series developed for reference ICs,[6] which, in turn, is partially based on the impedance measurement procedures of Brimacombe.[10] The common theme of these tests is the use of a

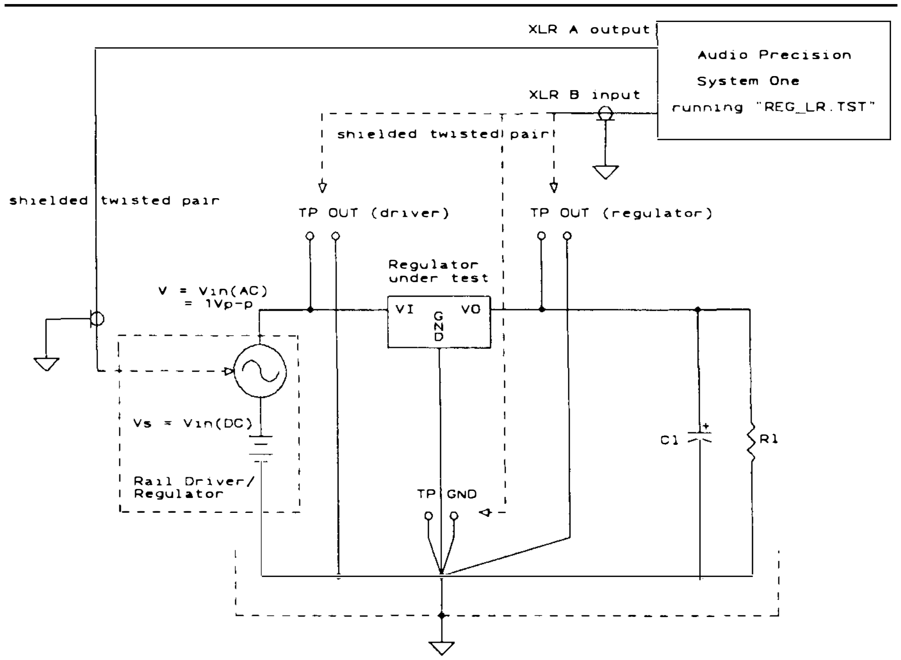

FIGURE 2a: Line rejection test.

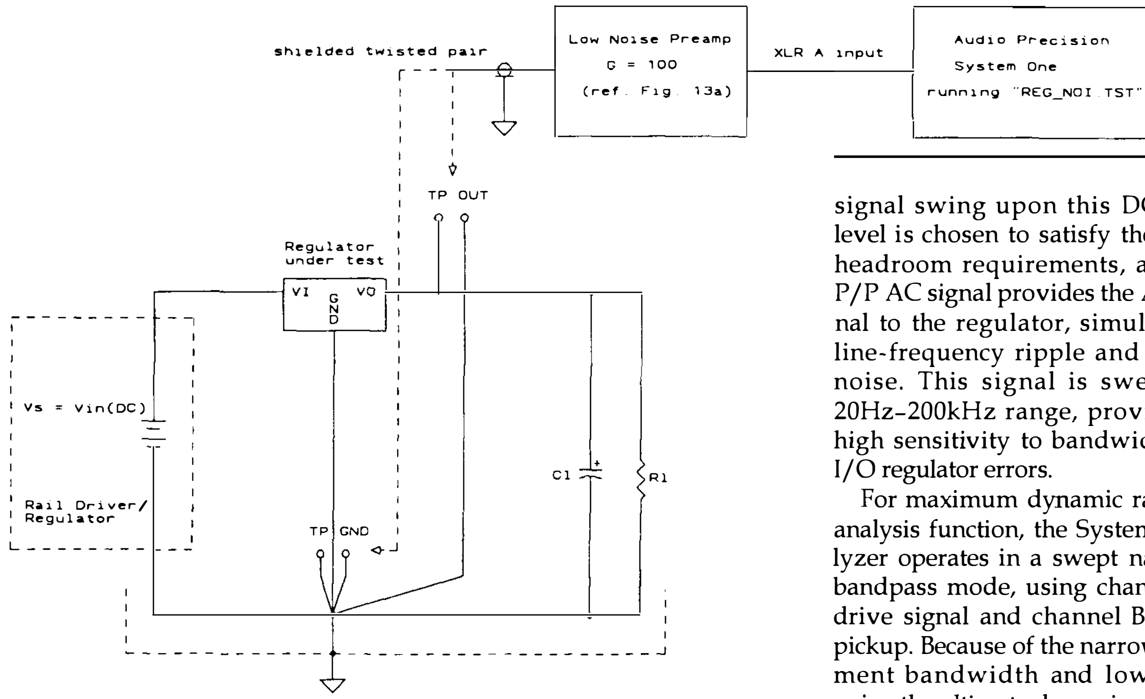

FIGURE 2b: Noise test.

sensitive high-resolution test instrument, the Audio Precision System One (PO Box 2209, Beaverton, OR 97075-3070, 503- 627-0832).

Line Rejection Tests

Figure 2a shows the setup for exercising a regulator for line rejection versus frequency. The DC input to the regulator, V_S, is taken from a positive or negative rail *driver/regulator* stage. This stage functions as a low-impedance source for both AC and DC and produces a nominal regulated DC voltage of 18V, with polarity suitable to the regulator under test.

You can superimpose a 1V P/P AC signal swing upon this DC. The DC level is chosen to satisfy the regulator headroom requirements, and the 1V P/P AC signal provides the AC test signal to the regulator, simulating both line-frequency ripple and wideband noise. This signal is swept over a 20Hz-200kHz range, providing very high sensitivity to bandwidth-related I/O regulator errors.

For maximum dynamic range in the analysis function, the System One analyzer operates in a swept narrowband bandpass mode, using channel A as a drive signal and channel B for signal pickup. Because of the narrow measurement bandwidth and low analyzer noise, the ultimate dynamic range of this test approaches 140dB. Consequently, very careful screening and guarding of the test setup is necessary to minimize contamination of the output signal and to take full advantage of the test system's dynamic range.

If you eschew such precautions, then the very best regulator in LR terms may be indistinguishable from those more pedestrian. More importantly, you will be unable to fully optimize a high-per-

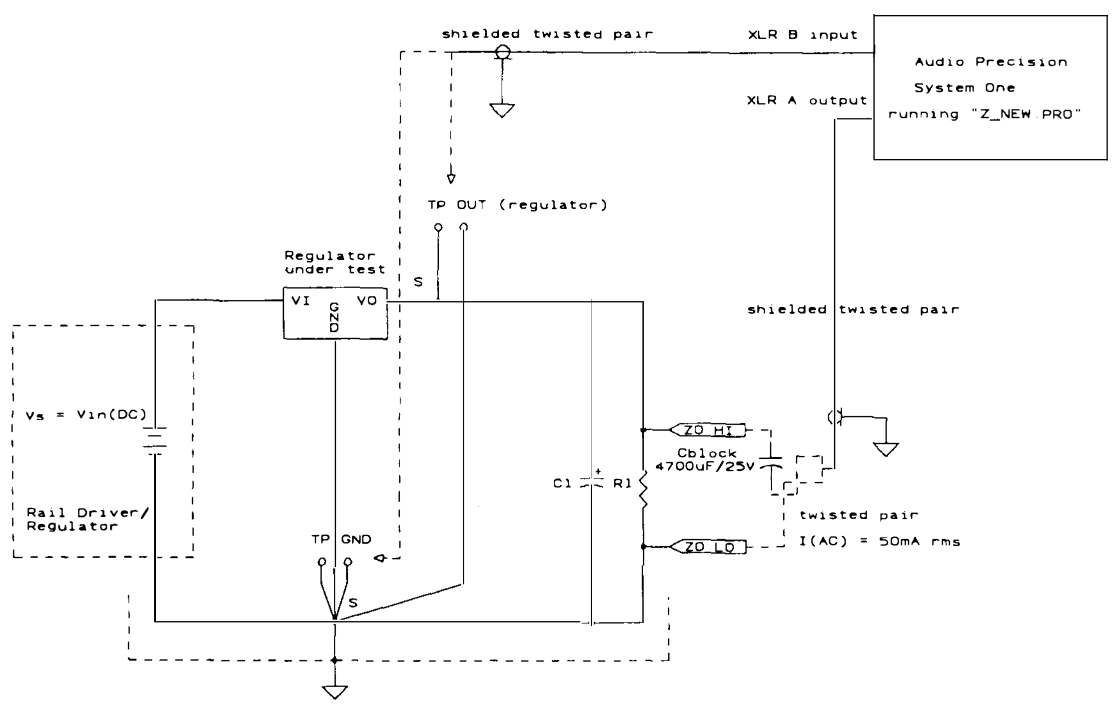

FIGURE 2c: Output impedance test.

formance design, unless the test setup is sufficiently sensitive.

In these tests, a grounded guard trace was used on three sides around all regulator circuits, which were built over a heavy ground plane. I also used shielded twisted pair construction on the analyzer I/O cabling, with plug-in tips for quick access to various sensitive circuit nodes. The TP OUT driver output two-terminal test jack provides a 0dB, 1V P/P input reference level, which calibrates the analyzer for full scale. All measured regulator output signals are referred to this 0dB level as it drives the input of the regulator under test.

The AC level measured at the TP OUT regulator jack represents the degree of isolation provided by the regulator, and is easily scaled and displayed in decibel versus frequency by the test analyzer. The TP GND jack provides a grounded-input, minimum signal reference for the analyzer.

To allow relatively easy comparison among various regulator types, all regulators under test used setups similar to this, regardless of their circuit topology. Thus with similar loading and input conditions, direct comparisons of measured performance between different circuits is possible. This test and the others operate with a common loading for all regulators of C_l = 100µF and R_l = 100Ω, except as otherwise noted.

With both positive and negative regulator circuits available, both positive and negative rail drivers are required to perform the LR tests quickly and accurately. I describe these in more detail later.

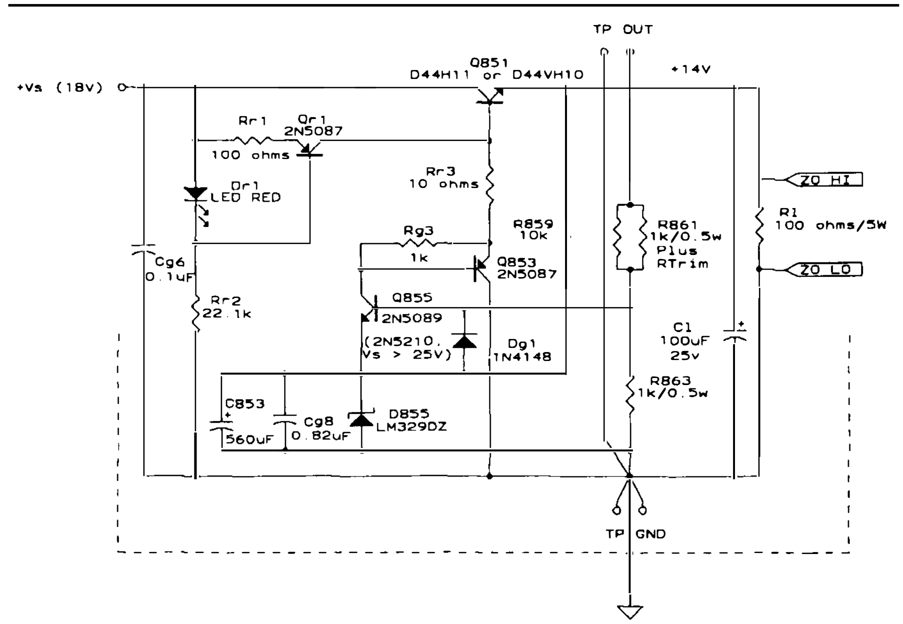

FIGURE 3: POOGE 5.51 positive regulator.

Noise Tests

Figure 2b shows the setup for testing regulator output noise. No AC input signal is provided, and the appropriate DC voltage V_S is supplied by the rail driver/regulator. For most of the circuits tested, this is also 18V, with polarity as appropriate. The only output signal is the AC noise level, as measured at the TP OUT test point of the regulator under test.

Because many regulators have very low noise levels, a preamplifier circuit is required to raise the AC output signal to a level where it can be readily measured by the analyzer. For this a balanced input, gain of 100 circuit was used, coupled to the regulator under test by shielded twisted pair cabling. This preamp's output is coupled into the analyzer, which operates in a swept, narrowband bandpass mode.

A given noise test provides a 100× scaled display of regulator output voltage noise as a function of frequency. A fundamental limitation of this setup is that the measurement preamp noise cannot be distinguished from the regulator noise, when they are of the same order. The preamp circuit's noise is about 2.6nV/√Hz, so this only becomes an issue for regulators with noise levels of about 7nV/√Hz or less. I describe the low-noise preamp circuitry in greater detail later.

Output Impedance Tests

Figure 2c shows the setup for testing regulator output impedance (Z_O). As with the noise test, an appropriate rail driver/regulator supplies an appropriate DC voltage V_S. To measure output impedance as a function of frequency, the analyzer is programmed to produce a constant 2.5V RMS behind a 50Ω resistance. This results in an AC current flow or I (AC) of 50mA RMS for load impedances low with respect to 50Ω.

An appropriately polarized 4,700µF DC blocking capacitor couples the AC test current directly across the load resistance of the regulator under test, through twisted pair wiring. This completely balanced signal transmission method was found necessary for the very highest-resolution measurements, where the equivalent amplifier input voltages to the analyzer are around the microvolt level in the highest-performance circuits.

In this test, the regulator is called upon to absorb the test signal AC current to maintain the output voltage at the DC design level. Since an AC signal is bipolar in nature, the regulator can only totally absorb this signal if it is prebiased to a DC load current higher than that of the highest AC signal peak (about 70mA). For these tests, all regulators were operated with DC loads of 100mA or more.

Of the three tests, this one is the most difficult in terms of wide dynamic range implementation. The location of the TP OUT test points must be accurately fixed at the true physical/electrical sensing points of the circuit (*Fig. 1*). No load currents must be allowed to flow in the wires to the test points; otherwise, the advantages of this four-wire sensing will be lost. This point is amplified by the "S" notations on the *Fig. 2c* diagram.

Overall sensitivity of this test setup is such that equivalent impedances of less than 10µΩ can be resolved at low frequencies, and 1mΩ or less at 100kHz. The respective figures list the test software, which is available by simply sending me a formatted 3.5" MS-DOS disk, *with a mailer including return postage.*

POOGE 5.51 Regulators

The discrete regulator of *Fig. 3* is an enhanced version of the POOGE 5 regulator described in Part 1 of Gary Galo's article.[11] For this article, it provides an example of a medium-to-high-performance discrete circuit regulator, and aside from the present comparative discussions, is practical and useful as shown. Those using this regulator must add some protection against overcurrent (this also applies to all others, with the exception of the internally protected IC types). Since there is no internal current limiting, a simple series fuse on the unregulated input side is sufficient protection.

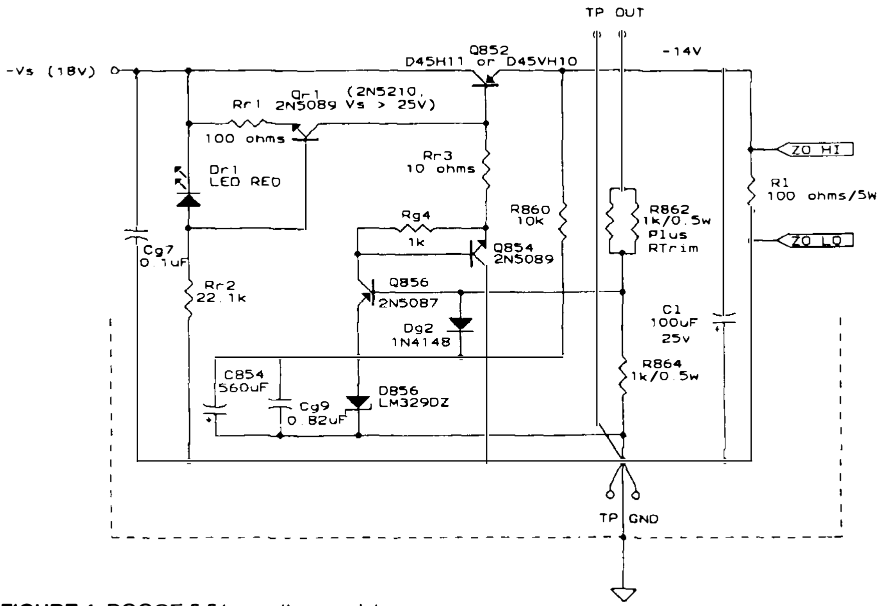

FIGURE 4: POOGE 5.51 negative regulator.

Close comparison of this schematic with the original version shows two main differences. One is Q851, either a D44H11 or D44VH10, both improved pass transistors as described in POOGE 5.5.[8] The other is a refined current source drive for Q851, Qr1, and the associated parts (such as Rr1).

The new current source functionally replaces the selected 2N5458 JFET in the original version, and allows improved performance with lower I/O voltages, i.e., lower *dropout*. While the simpler POOGE-5-style JFET current source works fine with medium-to-high I/O voltages, it requires 4–5V of bias to achieve highest LR. Since the combination of these two changes enhances the performance, I designated these new circuit versions as POOGE 5.51.

The LED-biased bipolar transistor source works well down to 1.5V of dropout voltage. This current source's output is set by Rr1 at approximately 10mA, which allows regulator output currents of up to 500mA, with a Q851 β of 50. Resistor Rr3 provides additional stability for Q851 in the presence of capacitive loads. When connected as shown, Rr3 has no negative effect on overall dropout voltage.

The negative version of the new regulator (*Fig. 4*) works identically in concept to *Fig. 3*, with, of course, the obvious polarity reversals and transistor complements. In both forms of the circuit, use the 2N5210 in place of the 2N5089 when input voltages are 25V or greater. For these tests, I set up this regulator (and most of the others, except as noted) with an output of 14V (trimmed just as described in Gary's POOGE 5.5 article) and a 140mA load, R_l.

Circuit Layout

As noted above, I built this circuit (and the others) in a small "cell" area surrounded on three sides by a grounded guard trace (#16 gauge, dotted lines in the schematic). In addition, a double-sided circuit board with a ground plane was used. (Breadboards for these tests are "IVANBOARD," an 8.5" × 11" RF design breadboard using a 0.1" grid surface mount pattern over a 2 oz copper ground plane.)

The standard dual test points are provided for ease-of-measurement via plug-in twisted pair cable connectors. One is at the TP GND reference point, provided at the circuit's common point, the common physical/electrical junction of reference diode D855 and divider resistor R863. The other is the TP OUT sense point, electrically connected between the above described TP GND point and the output node connection of divider sense resistor R861.

In the test strategy, I included a given test cell's TP OUT connection for all three tests, and used TP GND as a dynamic range and S/N reference check. For the Z_O tests, ZO HI and ZO LO provide access for the 50mA AC current test signal connections directly to R_l. I used similar test points and connections in all regulator test circuits.

Since the three-terminal regulator types tested are much simpler in their application, they are not shown in schematic detail. All types tested were by the original manufacturers, using the TO-220 package and carrying either a 1.5 or 3A rating. Note that many different versions of these regulators are available, some with much lower maximum current ratings. I did not test these types, but anticipate that those with lower current ratings (and associated higher output impedance) will not likely exceed the performance of the 1.5/3A versions.

The fixed 15V types (LM7815 and LM7915) are connected in their standard mode, with output loading of C_l = 100µF and R_l = 100Ω. The adjustable three-terminal regulators (types LM317, LT1085,

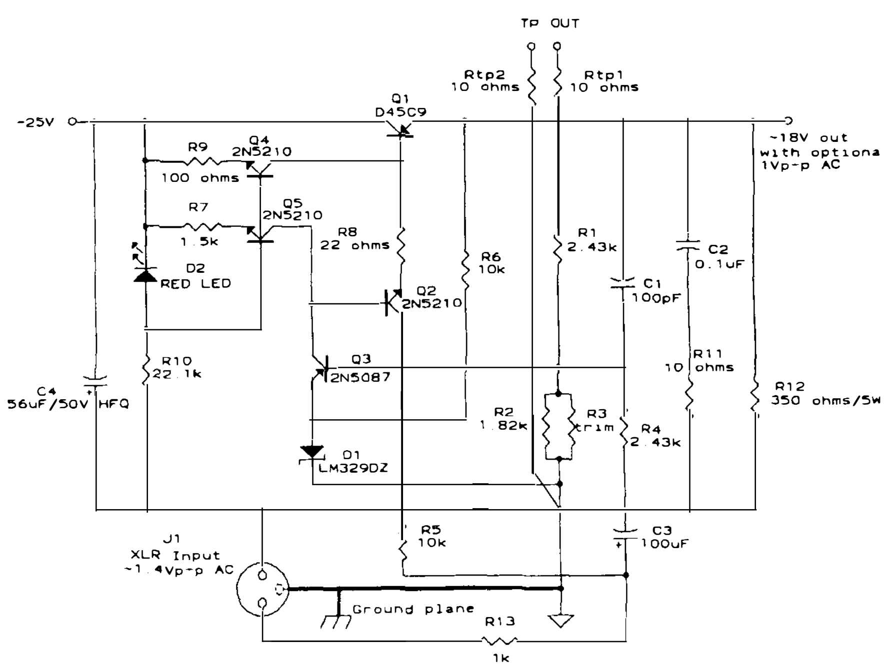

FIGURE 5: Negative rail driver/regulator.

LM337, and LT1033) are connected with an OUT-ADJUST pin resistance of 1kΩ and an ADJUST-GND pin resistance of 10kΩ, which programs them to a nominal 14.2V. These also used a C_{ADJ} bypass capacitance of 100µF, plus the loading of C_l = 100µF and R_l = 100Ω. It is worth noting that this relatively high divider impedance works to advantage for audio applications, since a given size C_{ADJ} capacitor is more effective across a 10kΩ resistor than with a 1kΩ value.[12]

Rail Driver/Regulators

For a controlled AC source test environment, the discrete regulators of *Figs. 3* and *4* are adapted to operate as low-impedance rail drivers. From an unregulated ±25V DC source, these drivers provide a regulator signal source environment which allows a fixed DC input level of ±18V, upon which you can optionally superimpose a 1V P/P AC test signal. In the line rejection tests, the ±18V output rails carry this AC signal, which is swept from 20Hz–200kHz.

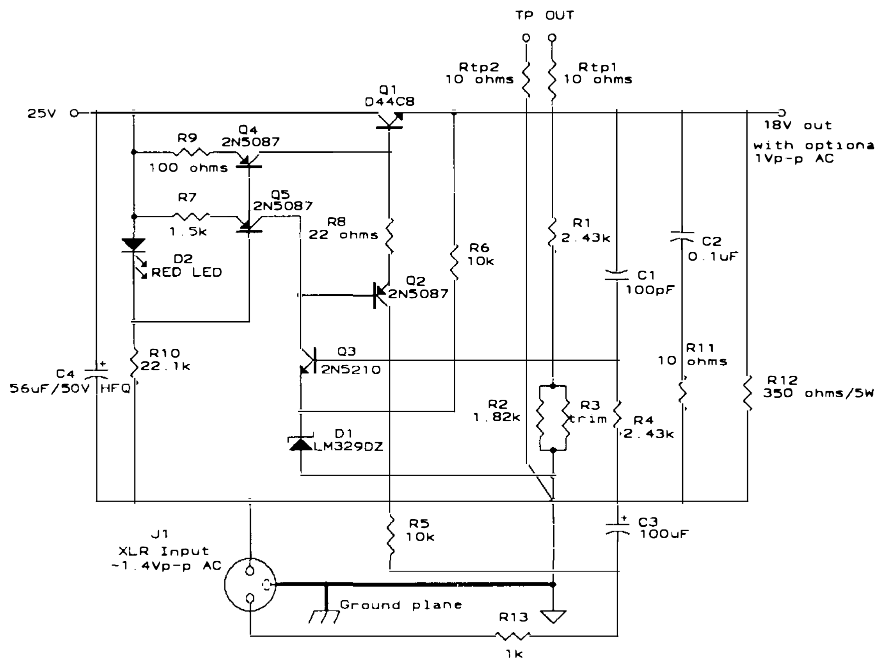

FIGURE 6: Positive rail driver/regulator.

This requirement demands a small-scale power amplifier, due to the fact that the various test regulators in many cases require a minimum value 0.1µF input bypass capacitor for stability. Driver circuits suitable for negative and positive 18V plus AC outputs are shown in *Figs. 5* and *6*, respectively.

The voltage slewing to maintain a flat-frequency response into the 0.1µF load capacitance of a test regulator requires a substantial standing current in Q1, part of which is provided by the brute force load R12. As with the POOGE 5.51 plus/minus regulators, the

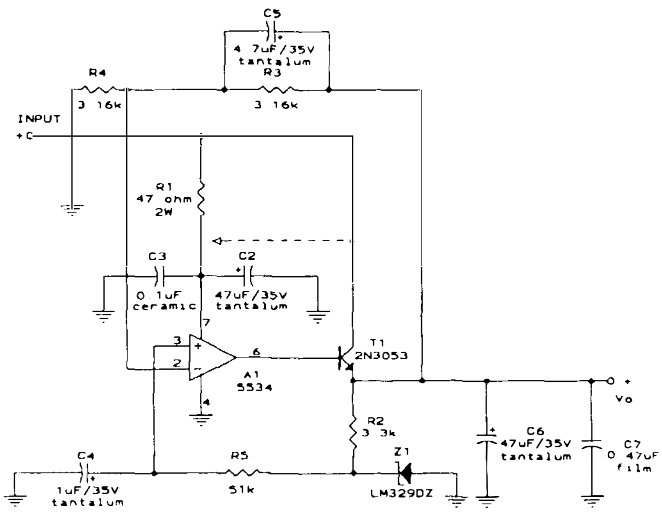

FIGURE 7a: Positive output Sulzer regulator.

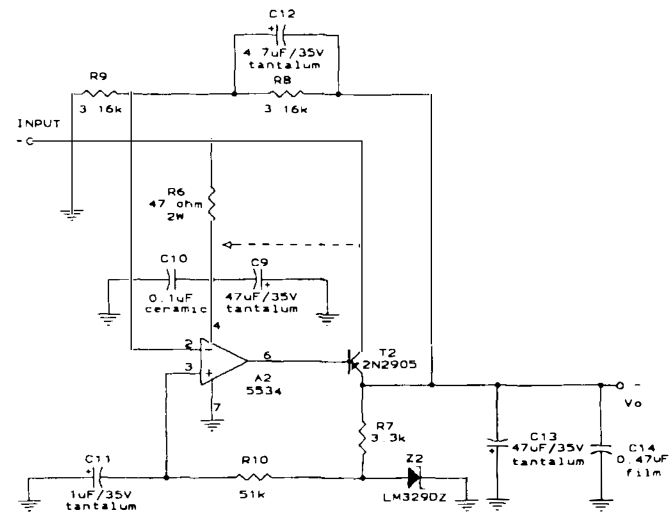

FIGURE 7b: Negative output Sulzer regulator.

output divider ratio and a reference voltage of ≅7.5V (6.9V plus 1Vbe) set output voltage. In this case, the bottom resistance R2 is trimmed by shunt R3 to eliminate interaction with the AC gain. R3 is trimmed for an output of 18 ±0.1V with the driver/regulator loaded. Loads of up to 500mA are possible. AC test signals at input J1 are passed to the output bus with less than unity scaling, since R1 < (R4 + R13).

In these drivers, the TP OUT test plug is used to verify the 0dB, 1V P/P 20Hz–200kHz AC swept reference signal as it is applied to the specific regulator under test for LR. The Audio Precision System One's function key F4 provides calibration to a specific measured output level for the 0dB amplitude reference for these test conditions. This calibration feature allows a measurement's 0dB reference to vary somewhat in absolute terms about the same nominal level (1V P/P or other), but still references all subsequent readings to this level, which is an extremely useful trick.

Another difference of this circuit in relation to the POOGE 5.51 regulators is the LED/bipolar transistor current source load Q5, which allows higher amplifier gain. This provides the driver with an output impedance measured at TP OUT of less than 5mΩ below 100kHz.

Some initial positive rail regulator testing was accomplished using a 317-type regulator as an 18V regulator/driver, with the AC test signal coupled into the normally grounded C_{ADJ} capacitor. While much simpler than the discrete circuit driver of *Fig. 6*, this setup was not completely satisfactory, since spurious resonances occurred at certain frequencies.

Sulzer Regulators

The original Sulzer regulators[3] for positive/negative supplies are shown in *Figs. 7a* and *7b*, respectively. These are reproduced almost identically to the original versions, with a couple of small but important exceptions. One change is the connection of the pass transistor's collector, which is taken directly back to the input, as opposed to passing the output current through R1 (or R6). This step allows more effective decoupling of the op amp rail, and will increase available

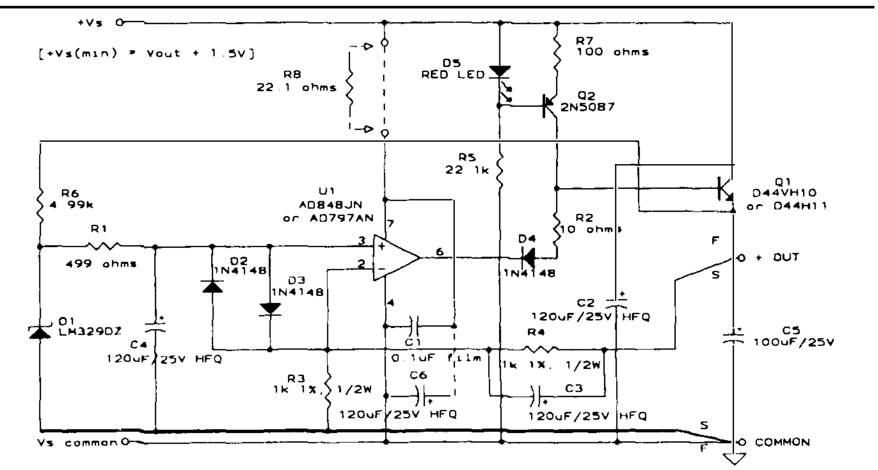

FIGURE 8a: Positive output, low-noise, low-dropout regulator. For this figure and *Figs. 8b* and *9*, unless otherwise specified, all resistors are ¼W, 1% metal film. A heatsink is required on Q1 for Pd >0.5W. In this figure and *Fig. 8b* only, minimize lead length from unregulated input and return. "S" indicates Sense lead; connect at point-of-load for best regulation

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headroom. It was also advocated by Sulzer in the "revisit."[4]

That article also suggested the use of "high quality zener, composed of an integrated circuit," which was attributed to Joe Curcio. Presumably, this was the lower noise LM329 IC zener as shown here and used in these tests (the LM329 was used also by Breakall, et al[7]). The only other variations are minor ones in resistance values, based on available values (R5/R10 and the output divider). With equal divider resistors, the circuits produce about ±14V.

Sulzer rated the original circuit at 100mA, so in this case I set load resistor R_l at 150Ω (not shown). I used no additional C_l, beyond the values for C6 and C13, and built up the test circuit in the manner described above, being careful to locate the TP OUT test points appropriately at the sense points, to connect ZO HI and ZO LO to R_l and control (well) the high-current paths and grounding.

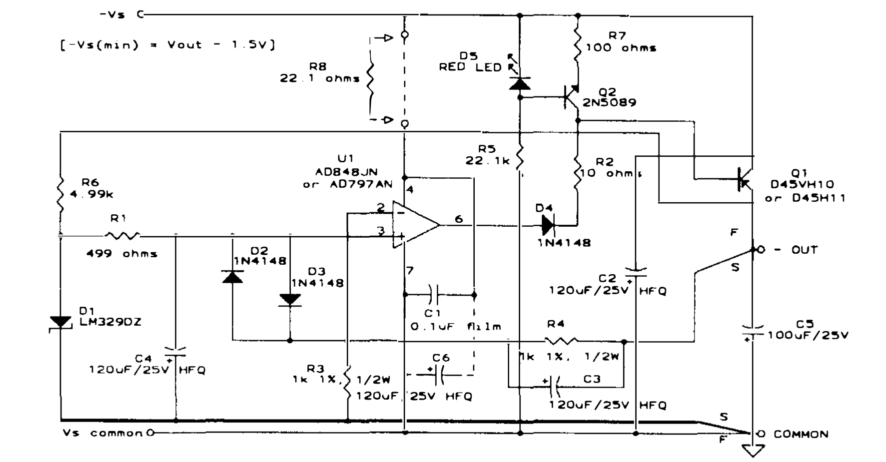

FIGURE 8b: Negative output, low-noise, low-dropout regulator.

Low-Noise, Low-Dropout Regulators

Over the last two years, I've been working on a new family of high-performance, op-amp-based regulators. This work started shortly after the publication of the POOGE 5 regulator,[11] and sought to examine pass devices, amplifiers, and overall topologies to improve their performance. My goals were to extend a design allowing use at low output voltages such as 5V and to enhance dropout performance so DC regulation could be maintained down to V_{OUT} +1.5V (1.5V dropout). In addition to achieving Z_O performance similar to the already excellent Sulzer regulator, I wished to push LR and noise performance to levels as high as possible.

In general all of these goals were met, but the route to the final result has been quite an adventure. The testing has been among the more fascinating parts of the development, and, as I'm sure most will agree, it is full of surprising results. This will not be at all obvious just by examining the circuits, but it may become clearer as I explain the various test results.

Some audio designers simply prefer discrete circuits for regulation. Even given this preference, however, premi-

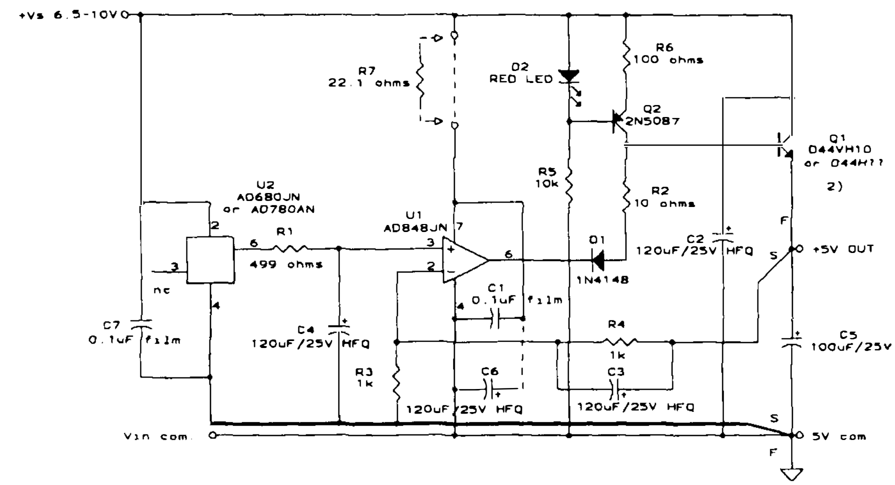

FIGURE 9: 5V, low-noise, low-dropout regulator. Minimize lead length from +Vs and return.

um-level results for Z_O, LR, and noise takes both a skilled designer, plus lots of parts. For example, the POOGE 5.51 circuits illustrate this problem. The performance is excellent for their relative simplicity, but if you wish to lower output noise, for example, you will not accomplish this simply.

You'll need additional parts to implement zener noise filtering and buffering, which may have a negative effect on Z_O performance. It also becomes extremely difficult to get a comparable level of all-around performance working at a 5V output level, since the reference voltage needs to be <5V, a level which typically is more noisy because of the physical limitations of available bandgap-based voltage references.[6]

This leaves the topology of choice a low-noise, wideband differential input amplifier which operates down to 8–9V (or less), can be configured for low dropout, and has high inherent LR properties. This is a tall order for a simple discrete amplifier, and most would agree that there might be quite an increase in parts count over *Fig. 3* or 4, just to implement such an amplifier. At this point, a very carefully selected op amp sounds much more efficient. Unfortunately, the 5534 of the Sulzer design does not operate well at low voltages, so this leaves newer devices as potential candidates.

The circuits of *Figs. 8a* and *8b* represent the new positive/negative design solutions, the result of more than a few iterations with amplifiers and related components. Superficially, these designs look much like the Sulzer circuit, but with both major and minor differences. The circuit discussion that follows is in terms of the positive regulator of *Fig. 8a*, but the negative form works in a similar fashion, given polarity normalization.

The Circuit

A major departure from a straightforward U1/Q1 drive topology is the use of a current source drive for Q1's base, composed of Q2 and the associated parts. As with the POOGE 5.51 circuits, the LED-biased current source allows Q1 to operate down to 1.5V or less of differential, lending the circuit low dropout features.

The potential for a limited swing of the op amp to affect dropout is lessened by operating it in a current sinking mode, which is enabled via D4.

Typically, this circuit achieves dropouts of 1.2V with several hundred milliamp outputs, making it suitable even for logic supplies. Note that high regulator dropout is a serious issue for a logic regulator, where a 3V dropout can increase the power dissipated in Q1 to an intolerable level.

A second important change in the new regulator is the addition of amplifier input clamp diodes, D2-D3. These normally zero-biased diodes protect the op amp input stage, in the presence of ON or OFF transient voltage differentials greater than 5V. Such a large voltage can harm an unprotected input stage by breaking down either transistor's E-B junction.[13]

With a 7V reference voltage, the possibility that U1 can be damaged (or subtly degraded for noise) exists, if the input stage is allowed to break down differentially. The clamp diodes prevent this from happening, and should be used in this circuit for cases where the op amp does *not* have such diodes internally. The AD848 doesn't, so use D2-D3 as an ounce of prevention.

With lower-level reference voltages (such as 2.5V), you can eliminate the clamping diodes in many instances. As a general rule for most unprotected op amps, the worst-case differential transient error should be maintained <5V for safety.

While the presentation of the reference voltage to the op amp is similar to the Sulzer configuration, as is the feedback network, the general impedances are lowered and made symmetrical for both AC and DC. The matched 500Ω DC source resistances enhance overall DC stability at little or no cost, and the 100% AC feedback around R4 lowers both output impedance as well as noise.

This latter technique, one hallmark of the Sulzer configuration, is a significant key to achieving the highest possible performance with a given op amp. It allows the net regulator output noise to approach that of the op amp itself, plus the filtered noise level of the reference input at Pin 3. With an ultra-low-noise op amp such as the AD797, 1kHz output noise levels approaching 1nV/√Hz are possible.[14] To realize the highest possible attenuation in the single-section reference noise filter, a low-ESR capacitor is used for C4, a 120μF/25V-type HFQ. A relatively high voltage rating also helps lower leakage, as well as ESR.

Network Advantages

With wide-bandwidth op amps for U1, supply bypassing is critical for stability. The small RF-quality film bypass C1 is located close to the device pins, and is mandatory. The minimal operating hookup consists of just C1 at U1, along with C2 close to the collector contact of Q1. Optionally, you can use extra noise filtering via R8 and C6 to increase the high-frequency supply rejection of U1.

The net advantage of using this network depends upon several factors. One is the specific part used for U1; another is whether a positive or negative output is being implemented (since the plus/minus supply rejection of many op amps differs). As shown, the corner frequency is about 60Hz, and while not absolutely necessary, a low-ESR HFQ type for C6 allows greater HF noise rejection working against the relatively low value of R8. Finally, this network increases the DC/LF supply impedance seen by the op amp, and you should apply it very carefully to op amps with less than 100dB of supply rejection (such as the AD848). Needless to say, substitutions of op amps in these circuits are strongly discouraged.

To set up this regulator for voltages other than the nominal $2 \times V_r$ or $\cong$14V, change the R3-R4 resistors as shown generally in *Fig. 1*, keeping in mind the 6.9V reference voltage used. You can expect some trim when the loose tolerance "DZ" version of the industry standard 329 diode is used. However, the exact DC output voltage is not likely to be critical, except as it may affect dropout with a marginally low raw-DC supply. For test purposes, I loaded this circuit with the standard loading of C_l = 100μF and R_l = 100Ω (not shown in these figures) and was careful, as with the Sulzer circuits, in the physical wiring/layout.

Closely related to the *Fig. 8a* circuit is the 5V regulator shown in *Fig. 9*, which evolved from reference 15. It operates with a lower voltage three-terminal 2.5V bandgap reference at U2, an AD680, or an AD780, but is otherwise similar to the *Fig. 8a* positive regulator. It is tested at a 300mA current level with V_{IN} = 8V, as suitable to logic systems.

In Part 2 we will more closely examine the three test setups used, in conjunction with the Audio Precision System One, to determine regulator performance. The differences between regulators easily stands out, making an optimum performance choice easy. □

Part 2

REGULATORS FOR HIGH-PERFORMANCE AUDIO

By Walt Jung

The actual testing of various regulators for this article was divided into three major tests for LR (line rejection), noise, and Z_O (output impedance), which are further divided by the various regulator topologies, and then into positive and negative forms. The aggregate regulator count was 13 separate circuits under test, described in the following sections, and broadly organized into the separate LR, noise, and Z_O test series.

Line Rejection Tests

The basics of the LR tests, summarized in Part 1 (Fig. 2a), can be illustrated with a simple example using the 7815 three-terminal regulator device. Powered up as shown in Fig. 2a, it uses an 18V DC source and standard 1V P/P AC input signal as provided by the positive rail driver/regulator (Fig. 6). The loading is C1 = 100µF and I1 = 150mA.

Figure 10 is a three-trace plot resulting from the LR tests on the 7815. It uses a horizontal frequency scale of 20–200kHz and a vertical scale of −150 to +10dB. This general procedure similarly applies to other regulators. For test calibration, the TP OUT driver signal is monitored with the analyzer B input, and is set to 0dB while running "REG-LR.TST" (using the analyzer's F4 function key, as explained in Part 1). With a given regulator, the analyzer input is thus referenced to 0dB = 1V P/P, for a reference V_{IN} frequency sweep trace (a).

Next, we run a second zero signal GND reference trace, with the analyzer input connected to the regulator TP GND test point. *Figure 10* shows this as the lower (varying) trace (c), which ranges from an isolation of about 105dB at 100kHz, down to the 130dB range below 1kHz. This trace represents the test setup noise floor, and thus is the maximum possible LR measurement for a specific test cell location (there is some variation across the 13 sites of the test breadboard, mostly at the higher frequencies). The (a) and (c) traces are not repeated in subsequent LR plots, but were recorded for reference purposes.

The V_{OUT} trace (b) shows the performance of a given regulator measured at the regulator TP OUT test point. In the 7815 example, the regulator isolation is more than 70dB at very low frequencies, and decreases to about 45dB at 100kHz. The object, of course, is to maximize regulator LR across this range of frequencies. The subsequent plots show the V_{OUT} (only) traces of various regulators, grouped by their respective families. The V_{IN} and GND reference traces are not shown but may be assumed comparable to *Fig. 10*.

The three-terminal *positive* regulator group's LR in *Fig. 11a* includes the 7815 *fixed* regulator (c), plus the 317 (b) and 1085 (a) *adjustable* regulators, connected as noted in Part 1. These adjustable versions are appreciably better than the 7815, except at the extreme upper frequencies. The 80–90dB isolation at low frequencies is possible due to the bypassing of the device adjust pin, which effectively lowers the AC noise gain of the IC, thereby increasing LR. Lacking the option of this AC bypass capability, fixed voltage three-terminal regulators such as the 7815 generally have worse LR than their adjustable counterparts.

The three-terminal *negative* regulator group LR in *Fig. 11b* includes the 7915 fixed regulator (c) and the 337 (b) and 1033 (a) adjustable regulators, connected as in Part 1. Here the adjustable regulators are still better than the fixed voltage 7915 at the very low frequencies, but this reverses above 1kHz, with the 7915 appreciably better at 100kHz.

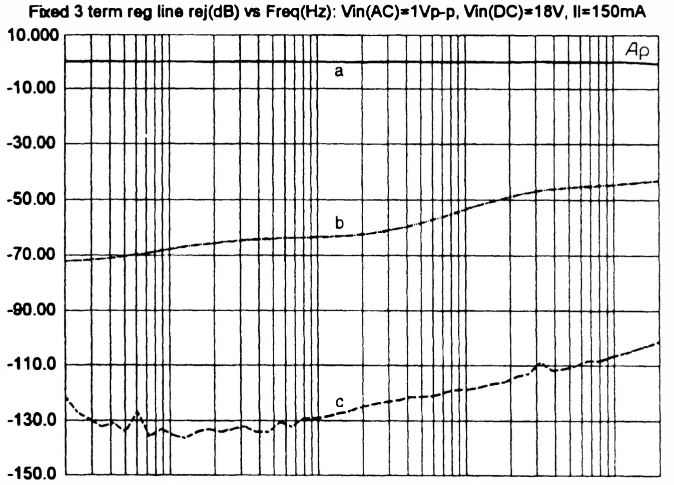

FIGURE 10: Fixed three-terminal 7815 regulator LR performance, a) V_{IN} = 0dB; b) V_{OUT}; c) GND reference.

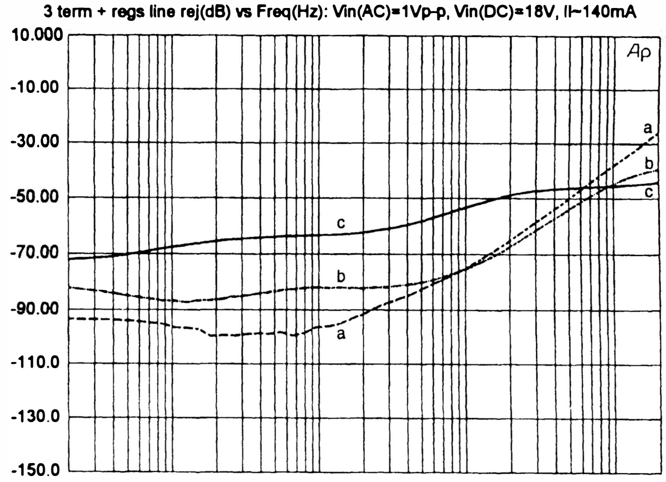

FIGURE 11a: Three-terminal positive regulator LR, a) 1085; b) 317; c) 7815.

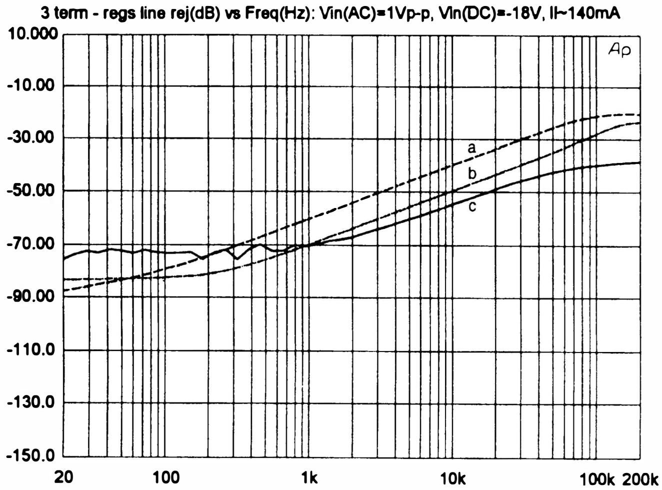

FIGURE 11b: Three-terminal negative regulator LR, a) 1033; b) 337; c) 7915.

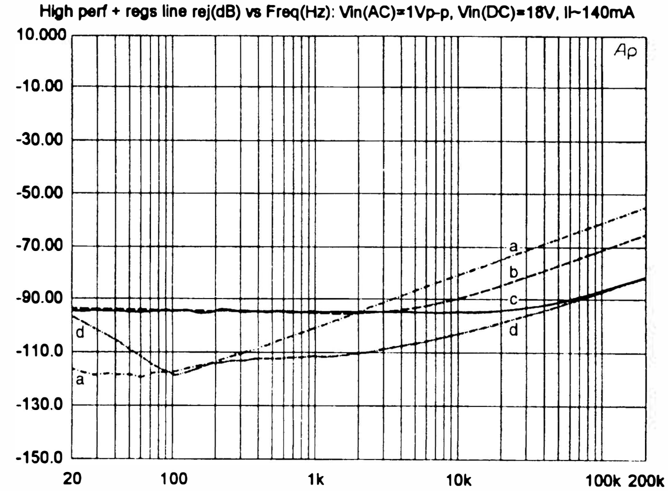

FIGURE 12a: High-performance positive regulator LR, a) AD797 (Fig. 8a); b) AD848 (Fig. 8a); c) POOGE 5.51 (Fig. 3); d) Sulzer circuit (Fig. 7a).

High Performers

Test results for the high-performance positive and negative regulator circuits (*Figs. 12a* through *12c*), as a group, show much higher LR, and also maintain this high rejection to higher frequencies than do the three-terminal types. Within the 20Hz–20kHz audio band, results of 90dB or greater are common in the high-performance group, with some circuits achieving isolation of 70dB or more at 100kHz. The very best regulators achieve 90dB or more at all frequencies, increasing to in excess of 100dB within the audio band.

In *Fig. 12a*, the Fig. 3 POOGE 5.51 regulator (c) has a remarkably flat LR of better than 90dB, up to about 100kHz. Two examples of the Fig. 8a circuit's performance are shown, using the AD797 (a) and AD848 (b) op amps. The AD848 attains about 90dB or more within the audio band, decreasing to 70dB at 100kHz. The AD797 achieves about 120dB at very low frequencies, decreasing to about 60dB at 100kHz. For these comparative conditions, the Sulzer circuit of Fig. 7a (d) achieves the best characteristics above 100Hz, decreasing to just under 90dB at 100kHz.

One reason for the excellent wideband LR of the Fig. 7a circuit is that the supply line to the op amp is passively decoupled, which enhances the op amp's inherent noise rejection above the R1-C2 corner frequency. In the data of *Fig. 12a*, the AD848 and AD797 op amps were *not* decoupled. When the optional Fig. 8a R8-C6 network is not used, the LR simply reflects that of the basic op amp.

The performance of the Fig. 8a circuit improves dramatically using both of these op amps with the network (*Fig. 12b*). The a and b curves reflect the same LR data for this AD797 and AD848 as displayed in *Fig. 12a* (that is, no decoupling). However, with the 22Ω/120μF filter active, the LR of the Fig. 8a regulator using either device increases to more than 120dB at 1kHz and 90dB at 100kHz, as shown in the two lower curves (c and d). The AD797's LR at low frequencies is superior, and approaches the noise floor.

Therefore, for the circuit of Fig. 8a (and also for its Fig. 8b complement), the best wideband LR performance is achieved with the optional noise filter in the op amp supply line. Because of the current drive mode used with the op amp, this supply line filter has a minimal effect on overall regulator dropout. In sum, there are two caveats to using this filter: the aforementioned (Part 1) potential degradation in LF response

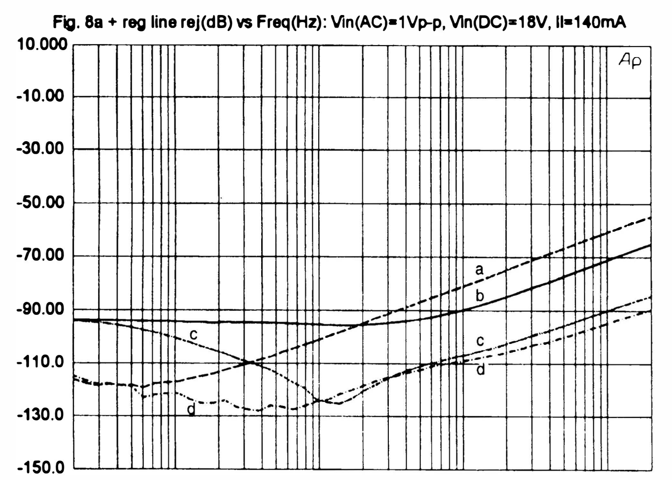

FIGURE 12b: Decoupling effects on LR (Fig. 8a), a) AD797; b) AD848; c) AD848 with decoupling; d) AD797 with decoupling.

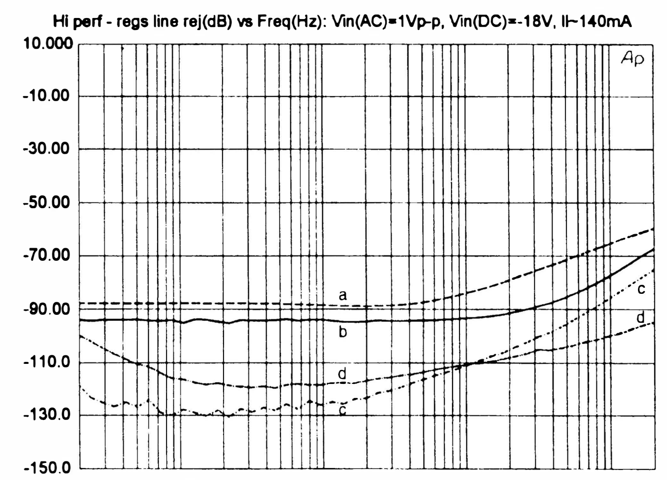

FIGURE 12c: High-performance negative regulator LR, a) AD848 (Fig. 8b); b) POOGE 5.51 (Fig. 4); c) AD797 (Fig 8b); d) Sulzer circuit (Fig. 7b).

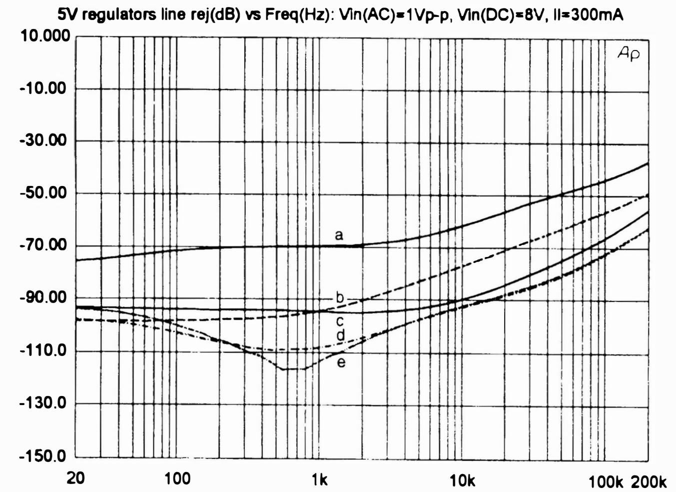

FIGURE 12d: 5V regulator LR, a) 7805CT; b) AD797 (Fig 9); c) AD848 (Fig. 9); d) AD797 (Fig. 9, bypassed); e) AD848 (Fig. 9, bypassed).

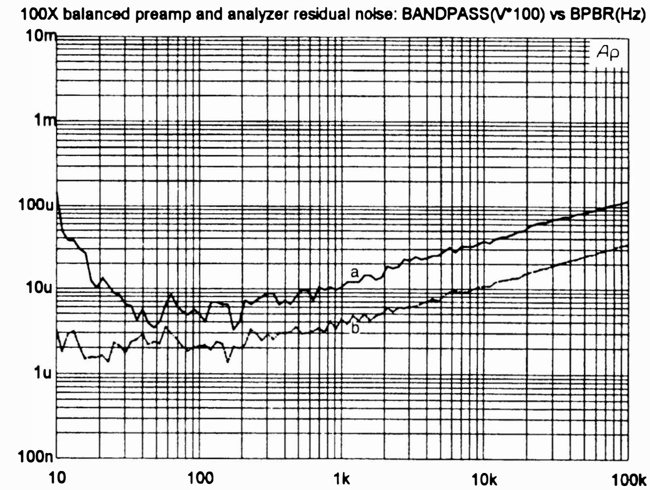

FIGURE 13b: Residual noise results, a) System One ×100; b) preamp.

(due to the supply Z increase) and the dropout voltage increase (a few hundred millivolts).

Negative Regulators

Op amps typically differ in their ability to reject power supply noise between the positive and negative supply rails. In these op amp regulators, just one supply line is relevant, with the other supply terminal grounded. As a result, the LR performance of otherwise mirror-imaged regulator circuits can and will vary substantially (for example, the Fig. 8a and 8b circuits).

Figure 12c shows LR results for the negative high-performance regulator group, with conditions generally similar to those of *Fig. 12a*. The AD848 (a) and AD797 (c) are operated within the Fig. 8b circuit *without* decoupling, while the Fig. 7b Sulzer regulator (d) and the Fig. 4 POOGE 5.51 regulator (b) are also tested.

Although the POOGE 5.51 negative regulator still shows a quite high and flat LR, all three op-amp-based negative regulators differ in their LR characteristics, vis-à-vis the data of *Fig. 12a*. This different op amp plus/minus supply rejection works to strong advantage for one form of the Fig. 8b negative regulator, i.e., the AD797. The Sulzer Fig. 7b circuit performance changes somewhat as compared with that of Fig. 7a, but overall it is still excellent.

The AD848 in the Fig. 8b circuit without decoupling does not perform as well as the AD848 positive counterpart of Fig. 8a, when similarly configured. However, the decoupling option does improve the LR substantially (data not shown). Where maximum LR is critical, the AD797 in the Fig. 8b circuit offers highest performance in the audio band.

The results of several 5V logic regulators for LR (*Fig. 12d*) illustrate curves for the AD797 and AD848 in the Fig. 9 circuit under two conditions, plus data for the standard 5V logic regulator, the 7805CT. The test conditions include a DC input voltage of 8V and a load current of 300mA, with the standard 1V P/P AC signal.

The 7805's LR (a) for this test is 55dB or better within the audio band, decreasing to about 45dB at 100kHz. This roughly compares to the 7815. The op-amp-based 5V regulators achieve better LR results, as you would expect, but not as good as the lower-current counterparts of Fig. 8a.

As with their operation unbypassed within the Fig. 8a circuit, the AD797 (b) and AD848 (c) in this 5V circuit show a LR of 70 or better within the audio band, with the AD848 appreciably better at high frequencies. The additional decoupling (d) and (e) improves LR above 100Hz for both devices, but not to quite the same degree as with the Fig. 8a circuit. Bypassing increases LR to 90dB in the audio band. Because of the greater potential for noise in a logic regulator, some type of raw supply filtering can be quite useful.

Noise Tests

Noise testing generally follows the scheme as applied to Fig. 2b. One of the keys to highly sensitive yet uncontaminated noise measurements is using low-

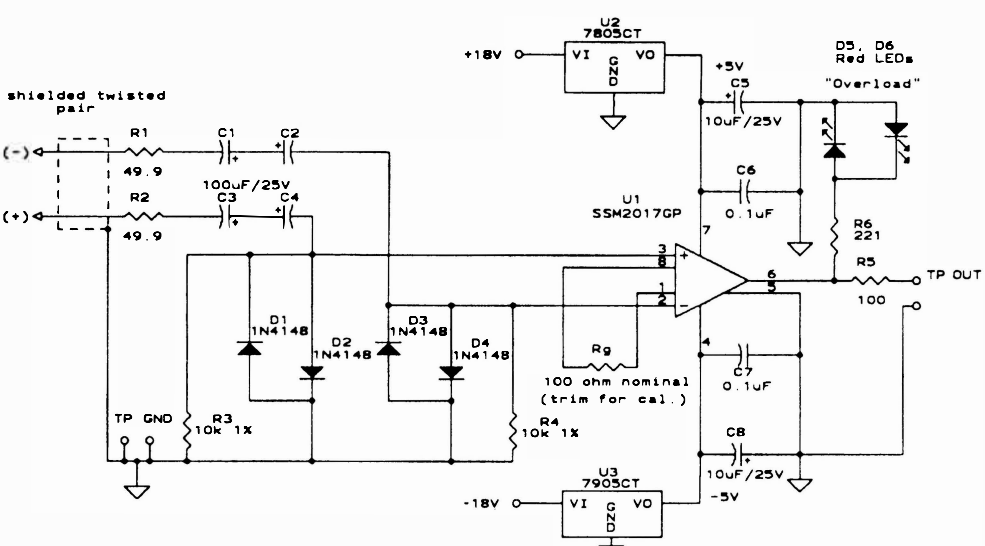

FIGURE 13a: Low-noise 100X balanced input preamp.

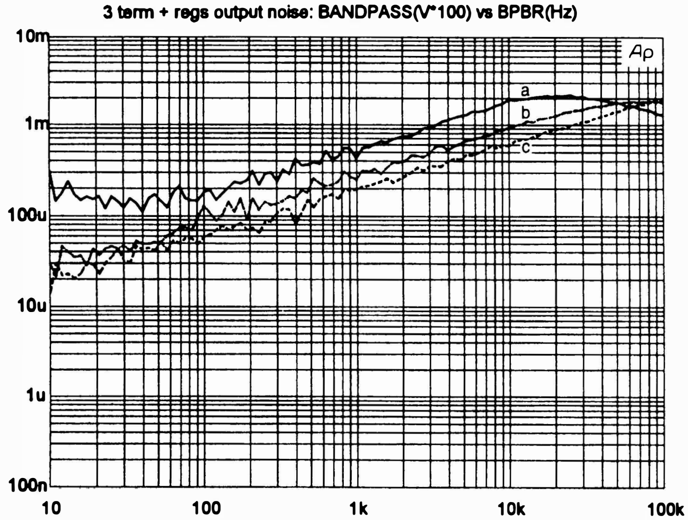

FIGURE 14a: Noise measurements for three-terminal positive regulators, a) 7815; b) 317; c) 1085.

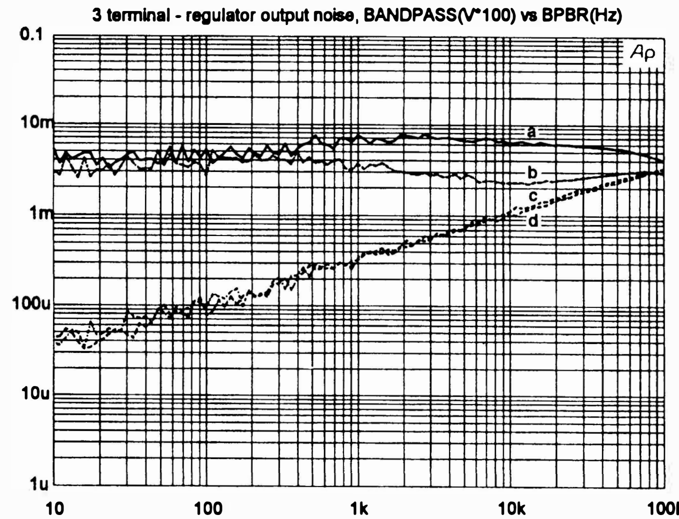

FIGURE 14b: Three-terminal negative regulator noise: 7915 samples—8818 (a) and 46AL (b); c) 337; d) 1033.

noise balanced preamplification. While the Audio Precision System One is a balanced input instrument, the sensitivity is not optimum for the lowest level measurements. These limits are pushed as the measured noise approaches noise densities of 10nV/√Hz or less, particularly at low frequencies. A gain of 100 balanced input preamp was developed specifically for this purpose (*Fig. 13a*).

The lower trace of *Fig. 13b* (b) shows the residual noise of this preamp (measured by the System One analyzer running "REG-NOI.TST"). In this test the analyzer's tracking bandpass filter sweeps over a range of frequencies, in this case 10Hz–100kHz, measuring the output in a narrow bandpass range. The filter for this function is a constant-Q type; thus, as the center frequency increases, the filter bandwidth increases.

As a result, a noise output which is spectrally flat measured through such a filter will appear to rise as frequency ascends, with a 3dB/octave slope. For example, for such a source, a noise density of 10nV/√Hz might be measured at 1kHz, 14nV/√Hz at 2kHz, and so forth. Some subsequent noise test examples illustrate this phenomenon.

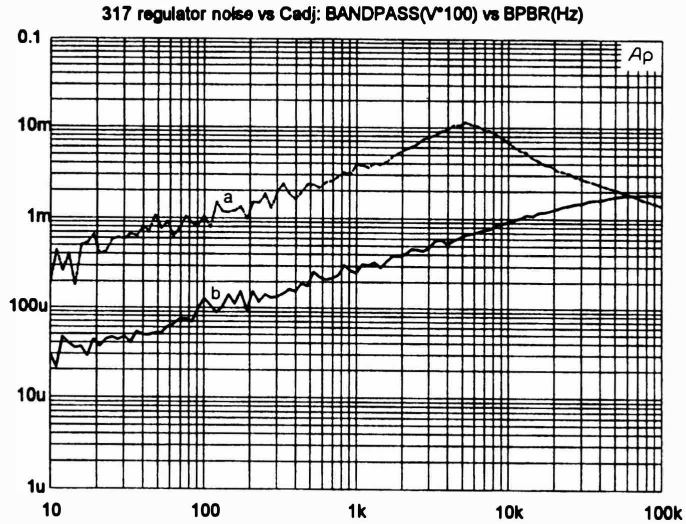

FIGURE 15: 317 regulator noise vs C_{ADJ}, a) $C_{ADJ} = 0$; b) $C_{ADJ} = 100\mu F$.

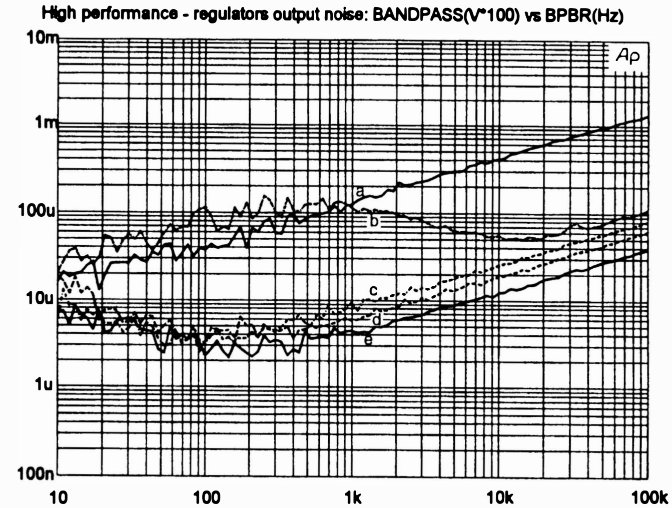

FIGURE 16a: High-performance negative regulator noise, a) unbypassed 329; b) POOGE 5.51 (Fig. 4); c) AD848 (Fig. 8b); d) Sulzer (Fig. 7b); e) AD797 (Fig. 8b).

In the *Fig. 13b* residual noise sweep including the X100 preamp (b), you can determine the equivalent input voltage noise density at a given frequency "F" by dividing the measured voltage by a factor of $100^*\sqrt{0.2316^*F}$. (The factor of 0.2316 is based upon the bandpass filter noise bandwidth, which is a third-octave type.) You can simplify this divisor to approximately $48^*\sqrt{F}$. At 1kHz for example, the division factor is ≅1,518, at 100Hz it is 480, and at 10kHz it is 4,800. For the X100 preamp residual, the input referred noise at 1kHz is measured at $4\mu V/1{,}518 \cong 2.6nV/\sqrt{Hz}$.

For the analyzer-alone residual noise displayed in the upper trace (a), the as-measured noise data was scaled by exactly 100×, to keep the two traces on a similar display scale and allow direct comparison. As you can note, operating the preamp before the analyzer lowers the effective noise by a factor of 2–3× at 1kHz, and by higher factors at very low frequencies where the analyzer input noise rises.

Preamp Circuit

While it would be desirable to lower the preamp noise further, increasing the gain of the circuit raises the risk of dynamic range and saturation problems. The preamp IC, an SSM2017, is capable of noise levels down to $1nV/\sqrt{Hz}$ operating at gains of 1,000×. Here, at a gain of 100×, the typical equivalent SSM2017 input noise is on the order of $2nV\ \sqrt{Hz}$, to which is added the noise of the input protection components. These nonideal (but necessary) parts raise the net overall noise to the measured level.

The preamp's noise rises gradually at low frequencies, as evident by the leveling of its trace. Above 1kHz the noise follows a roughly 3dB/octave rise, indicating the preamp noise is relatively flat at higher frequencies.

The *Fig. 13a* circuit includes back-to-back diodes across each of the two balanced inputs, which are necessary to prevent destruction of the SSM2017, as the inputs are connected directly to 15V outputs. R1-R2 limit the charging current to bipolar capacitors C1-C2 and C3-C4, as well as the clamp diodes, which allows safe direct jacking of the input to ±15V DC levels. The "Overload" back-to-back LEDs across the output light up when the DC output swings beyond ±1.5V, indicating possible nonlinear operation during the transient period as the input caps charge. When these LEDs extinguish, the output from U1 is in a linear range, so you can measure for noise.

You can facilitate accurate measurements by applying a known 1mV RMS level to the input of the preamp circuit, while monitoring the output for 100mV RMS. Calibration for various losses within the circuit is accomplished by trimming R_G, the SSM2017's gain set resistor, for a measured output of 100 ±0.5mV.

In measuring noise on regulators, the analyzer input is connected to the preamp TP OUT points. The regulator under test is probed directly with the shielded twisted pair input cable, which is connected to the regulator's TP OUT points (Fig. 2b). A noise analysis frequency sweep then produces a display similar to those of *Fig. 13b*, but in most cases higher in level, and often with one or more regions of complex frequency dependence. As you can see from the multiple decade log vertical scale of *Fig. 13b*, there is room for additional curves. The vertical scaling of 100nV to 10mV is maintained as consistently as possible in the following noise plots for ease of comparison between regulators.

Output Noise

In the noise results for the three-terminal positive regulator group (*Fig. 14a*), most units operate on an internal bandgap voltage reference, which is scaled upward from about 1.2V to the final output level. This amounts to a gain factor of about 12× in a 15V regulator. Accomplishing this scaling for DC only is sometimes difficult, that is, to *not* raise the AC noise components by the same factor.

As a result, three-terminal regulators can be quite noisy, especially if concern isn't taken in their selection and application. Among the various three-terminal regulators, lowest noise is generally realized with the adjustable types, *with the adjust pin bypassed* (as in these tests). The data below aptly demonstrates this point.

Using the 1kHz division factor of 1,518 with the positive regulator data of *Fig. 14a*, the measured noise is highest with the 7815 (a), at about $350nV/\sqrt{Hz}$ at 1kHz, while the 317 (b) and 1085 (c) measure about 180 and $130nV/\sqrt{Hz}$ at 1kHz, respectively. You'll notice some evidence of bandwidth limiting in the 7815, as the noise level falls off above 10kHz. This device's low-frequency noise also rises compared to the adjustable types, both of which maintain

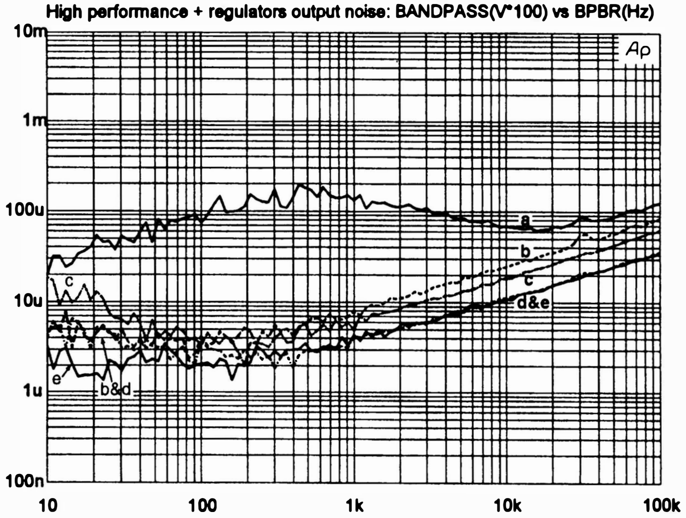

FIGURE 16b: High-performance positive regulator noise, a) POOGE 5.51 (Fig. 3); b) AD848 (Fig. 8a); c) Sulzer (Fig. 7a); d) AD797 (Fig. 8a); e) residual.

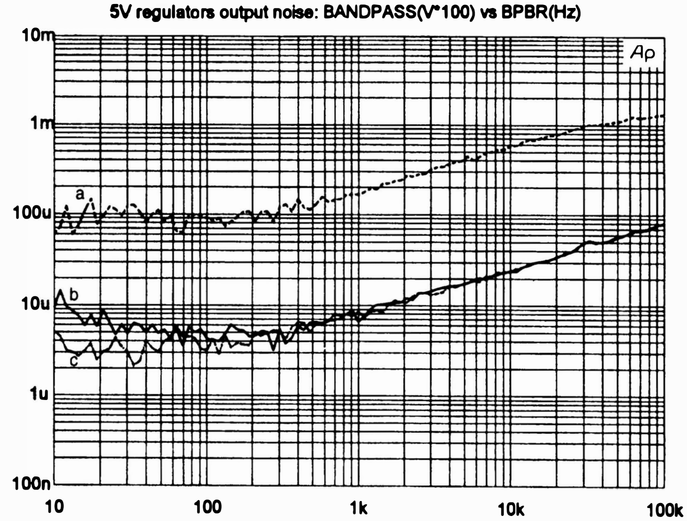

FIGURE 16c: 5V regulator noise, a) 7805CT; b) AD680 (Fig. 9); c) AD780 (Fig. 9).

a relatively constant 3dB/octave slope with frequency.

In the noise results of the negative group (*Fig. 14b*), the adjustable 337 (c) and 1033 (d) units have a similar 1kHz noise level of about 210nV/√Hz, while the two fixed output 7915 samples (a and b) show uncomfortably high 1kHz noise levels of about 4,600 and 2,300nV/√Hz. The two samples have different date codes, since the older initially measured unit (a) seemed to be unusually high. Noise in the (b) range for several 46AL date code samples seems to be typical, as does the 1/F noise characteristic. *Note that the scale of this plot ranges from 1µV to 0.1V.*

Figure 15 illustrates that adjustable three-terminal regulators with the adjustment pin bypassed offer lowest noise performance. These two plots measure the same 317 test regulator, both without (a) and with (b) the 100µF C_{ADJ} capacitor connected. The 1kHz noise levels are about 180 and 2,300nV/√Hz, respectively, roughly corresponding to the ≅12× AC gain dif-

Part 3

REGULATORS FOR HIGH-PERFORMANCE AUDIO

By Jan Didden

Walt Jung's articles on high-performance regulators for audio (*TAA* 1/95, p. 8, 2/95, p. 20) clearly show that the state of the art in this field is very high-performance indeed. His new regulators approach the ideal—a DC voltage source with zero impedance for AC signals, which is beneficial for the total audio system. After all, these systems are normally designed assuming that the power supply acts as such an ideal voltage source.

But in practice, power supplies deliver a DC voltage contaminated with noise, mains ripple, and signal-voltage residues, which are caused by the frequency- and level-dependent currents produced by the supply. The varying currents are delivered through the supply's output impedance Z_O, and current times Z equals volts.

Of course, various smart-circuit topologies can make the amplifier stages relatively insensitive to supply variations, but there will inevitably be some effect. Nonideal supply rails not only are detrimental to a stage's performance, but also are responsible for mutual interference between stages in a channel, or between channels in a system. This occurs because the contaminations caused by one stage or channel are coupled to another part of the circuit through the supply lines (unless each stage has a completely separate supply, which I haven't seen yet).

Again, you can take steps to limit this, but it is always better to avoid it in the first place. This article attempts to preserve as much of the excellent performance as possible in a real-world application, and is based on the high-performance regulators in Figs. 8a and 8b of the referenced articles.

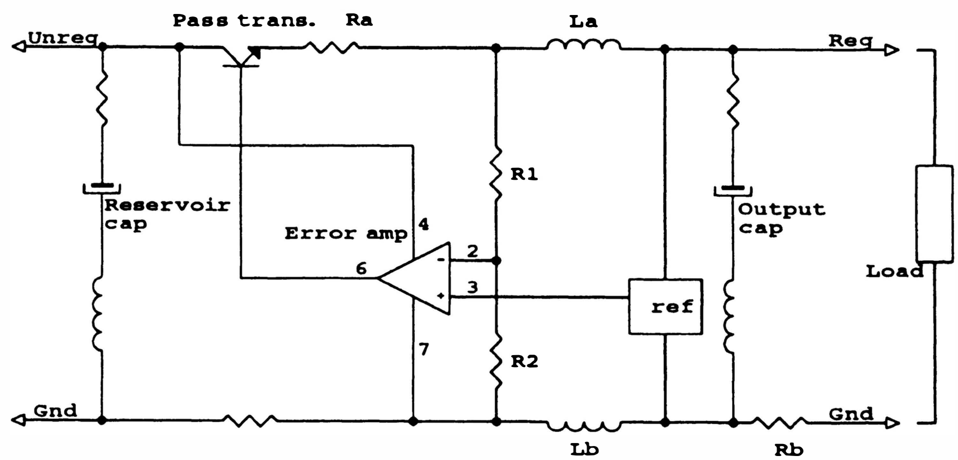

FIGURE 1: Real-world circuit, with parasitic components.

The Map Is Not the World

The key to successfully building the regulators is, surprisingly, a bit philosophical. We are so accustomed to viewing schematics as accurate depictions of the circuits that we rarely realize there are many components not shown in the diagram. But they are there, and the more you tune the cir-

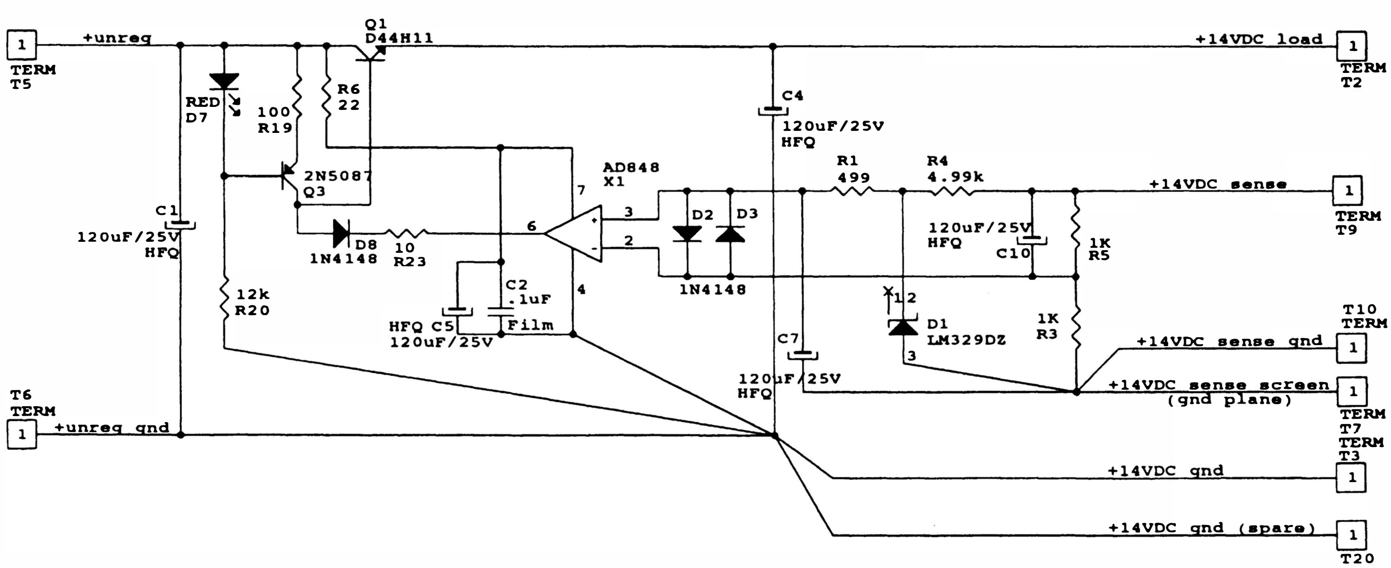

FIGURE 2: Positive regulator schematic diagram.

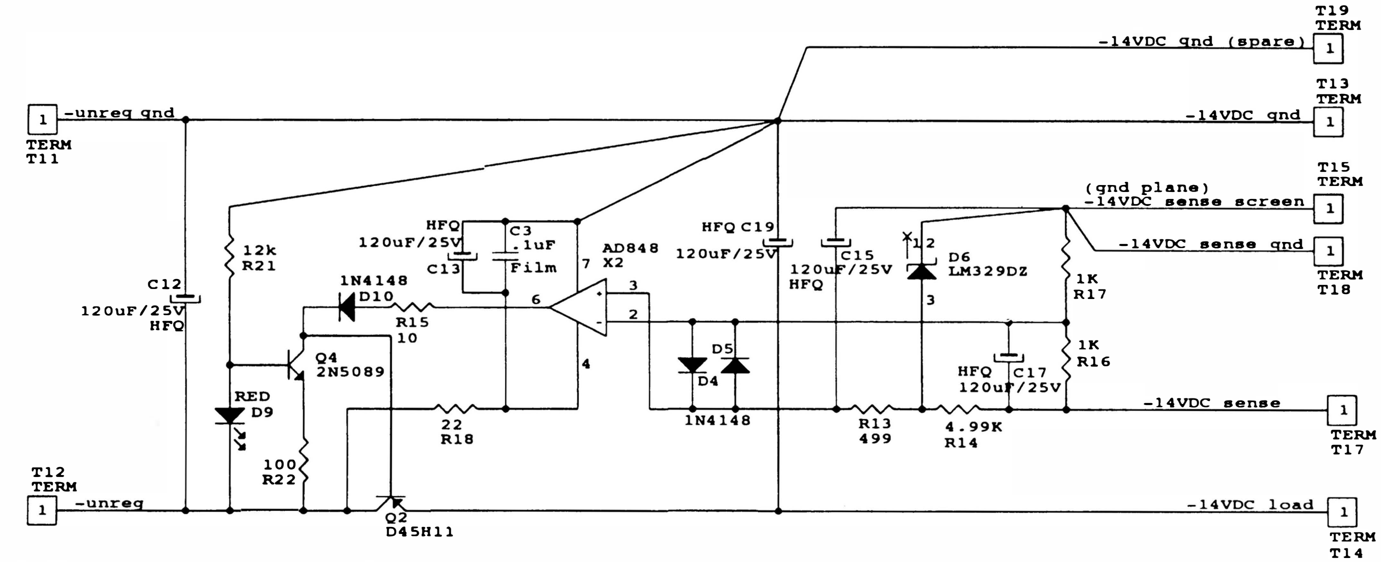

FIGURE 3: Negative regulator schematic diagram.

cuit, the more they become limiting factors.

Consider, for example, the rudimentary circuit of *Fig. 1*. Without going into details, let's just look at the input signals to the op amp. Here, you're looking for the reference at the noninverting input, and a sample of the output voltage at the inverting input. As you know, the op amp drives the series transistor to make the two inputs equal. You need the output voltage *at the load* to be clean and stable.

The load current runs through La, Lb, Ra, and Rb, which you won't find in the schematics, because they are the resistance and inductance of wiring and PC board tracks. The voltage at the inverting input depends on load-current amplitude and frequency. The reference voltage, which should really be developed relative to the common ground point, also depends on load current and frequency because of Lb and Rb.

So now you have a circuit in which the op amp works hard to accurately reproduce load- and frequency-dependent AC components at the supply output. Static DC errors normally are not problematic in audio; a nominal 15V DC supply works perfectly at 14.5V or 15.5V. The AC components cause the detrimental effects of supply ripple, which is really a manifestation of supply voltage varying with frequency and load current.

Unless you take adequate measures in the physical design, you cannot realize the superior regulator performance *at the load*, that is, at the circuit to be powered. If you need convincing, reread the sections on the measurement setups in the referenced articles.

The Way Ahead

So, what can you do? First, of course, use the best components you can afford. A capacitor should be a capacitor, not a series circuit that resonates at some inconvenient frequency. For this supply, Panasonic HFQs are the best available, and are prescribed for all electrolytics.

Second, avoid stray capacitances within the regulator circuit by using a sensible layout and screening where possible. Third, use short, thick, twisted wires for all external power connections to the raw supply and the load. This minimizes AC frequency- and load-dependent errors, as well as DC errors, caused by parasitic impedances.

All this can be quite effective, but the third measure is the most difficult. Remember that these regulators can provide output impedances lower than a standard "0Ω" jumper, coming very close to an AC short circuit. So, if you must use a few inches of wire to connect the regulator to, say, a preamp, you inevitably increase the output impedance Z_O at the load, which is where it matters. There's no use having zero ripple at the regulator board; you need it at the load.

Fortunately, you can take care of that, too, with a technique called Remote Sensing (RS). To clarify this and to focus on the physical layout of the boards, look at the familiar circuit diagrams. *Figure 2* shows the positive voltage regulator, and *Fig. 3* shows the negative voltage regulator. The specific RS provisions in *Fig. 2* are immediately apparent.

The circuit has four outputs: the

ABOUT THE AUTHOR

Johannes (Jan) Didden made his career in the Royal Netherlands Air Force, working in areas such as Air Defense, Software Development, Quality Assurance, and Communications. Several NATO assignments brought him to remote places such as Nashua, NH, and Torrance, CA. He has been building audio systems as long as he can remember, and has had several projects published in *TAA*. His other hobby is driving his Land Rover with his wife through famous tourist attractions such as the Libyan Desert and the interior of Iceland. He welcomes reactions to this article by FAX (+31 1680 23895) or by E-mail (didden@shape.nato.int).

TABLE 1

COMBINED POSITIVE/NEGATIVE REGULATOR BOARD PARTS LIST

REFERENCE	PART
C1, 4, 5, 7, 10, 12, 13, 15, 17, 19	120μF/25V Panasonic HFQ (Digi-Key P5698-ND)
C2, 3	0.1μF/50V film
D1, 6	LM329DZ National Semiconductor
D2–5, 8, 10	1N4148
D7, 9	Red LED
Q1	D44H11 Harris or Motorola
Q2	D45H11 Harris or Motorola
Q3	2N5087 (or equivalent)
Q4	2N5089 (or equivalent)
R1, 13	499 0.25W film
R3, 5, 16, 17	1k 0.5W film
R4, 14	4.99k 0.25W film
R6, 18	22 0.25W film
R23, 15	10 0.25W film
R19, 22	100 0.25W film
R20, 21	12k 0.25W film
X1, 2	AD848JN or AD797JN, ADI

MISCELLANEOUS

Heatsink (Digi-Key HS112-ND or equivalent), circuit board(s), mounting hardware, transformer(s), rectifiers (see text).

load connections indicated by "+14V DC load" and "+14V DC gnd" and a pair of separate connections for the sense points, called "+14V DC sense" and "+14V DC sense gnd." The aim is to take the sense connections "as close as possible to the load," which may require some pondering, and I'll examine this later.

So, now we feed the regulator feedback circuitry (R5, R3 in *Fig. 2*) with the actual output voltage at the load. Furthermore, the reference voltage is developed from the same points as well. The currents through the sense wires are very small and constant, and thus do not affect the performance. You have now all but eliminated the influence of the connecting wiring on both the feedback and the reference.

The supply current for the op amp and the current source around Q3, however, vary with frequency and load current. After all, these currents vary with the pass transistor's base current demand, and are thus clearly load- and frequency-dependent. Therefore, these circuit points are not returned to the clean sense points, but to a star ground as a "next best" alternative. Parts 1 and 2 already addressed the rest of the circuitry, so I won't go into further details. The negative version (*Fig. 3*) is, of course, just the twin of *Fig. 2*.

Getting Physical

Figure 4 shows the actual solder-side board layout. One board holds one positive and one negative regulator. The circuits are completely separate, so you can, if you wish, cut the two halves apart for your specific mounting/space requirements.

When the load exceeds 50mA or so, you should mount the pass transistor on a heatsink. As you can see, the board layout has provisions for that. The outline on the stuffing guide (*Fig. 5*) is a heatsink from Fischer in Switzerland; I couldn't find an exact US replacement type. However, you can use the one in the parts list (*Table 1*), available from Digi-Key. Again, you have the flexibility to cut off those areas of the boards. You can then mount the pass transistor on the enclosure wall or other heatsink, positioned perpendicular to the board. I tried to keep this layout as flexible as possible, as many readers might wish to use the new supplies in existing equipment.

The board also has a screened area on the component side, which extends underneath the sensitive control circuitry (*Fig. 5*). It purposely does not cover the higher-current areas to avoid possible capacitive coupling of ripple currents into the control section. In addition, you'll get best results if you mount the board closely over a metal sheet or enclosure wall to provide further screening. Such practice results in measurably lower noise and lower Z_O.

Use 18 AWG or heavier wiring for the raw supply input and load connections. Twist the hot wiring with the corresponding ground return line and route them away from the board and active circuitry as much as possible. Use a balanced screened cable for the sense connections, and connect the screen to the provided pad only at the board. The sense lines carry very little

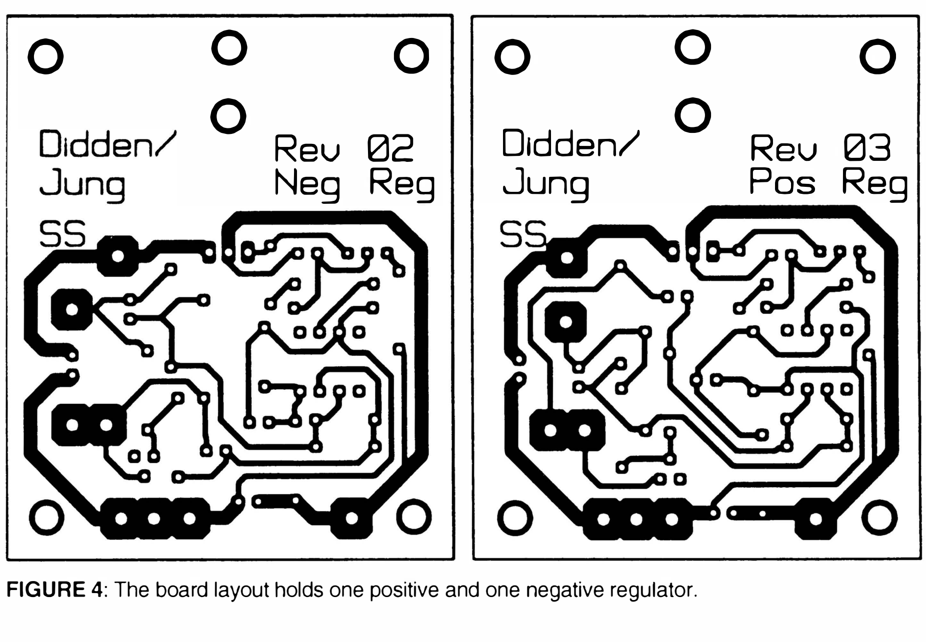

FIGURE 4: The board layout holds one positive and one negative regulator.

current (roughly 8.5mA), so a good-quality stereo lead should be adequate.

The two "sense gnd" and "sense screen" pads on the proposed boards (T7 and T10) are plated through, which will automatically connect the screened board area to this point as well. Of course, you should make every effort to keep the wiring as short as possible. These requirements are conflicting, so spend some time figuring out the best layout.

Stuffing Techniques

Stuffing the boards should not pose any problems. They are quite compact, but careful soldering will ensure that they will work straightaway. The easiest way is to start with the low-profile components (resistors and diodes), then the film caps, transistors, LED, and finally the electrolytics.

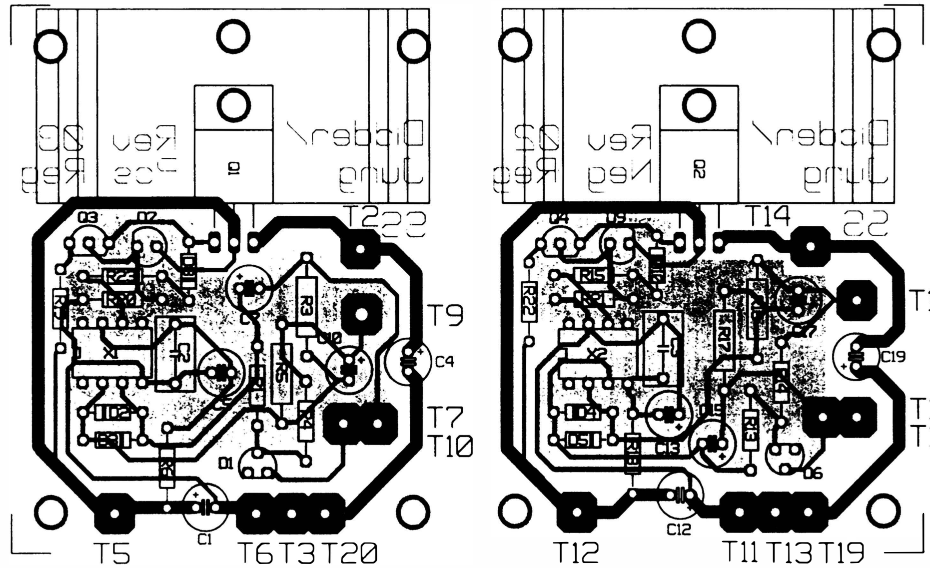

FIGURE 5: Stuffing guide for the combined positive/negative regulator board.

Figure 5 shows the stuffing guide and *Table 1* the parts list. Select the parts with care, because they and the layout determine the ultimate performance. I've already mentioned using HFQs for all electrolytics. The specified op amp and the reference diode are crucial to the wide bandwidth and the low noise. You could use an AD797 instead of an AD848, but limit yourself to these two types.

The best way to mount the op amps is to solder them directly to the board. This gives you the fewest problems with stray capacitances and intermittent contacts. Many readers might prefer to use sockets for easy swapping in case of problems, but the op amps are virtually indestructible in this application, and if the boards don't work, it is probably due to a wrong component value or polarity. I strongly recommend direct soldering.

The other parts should be as good as those in a high-quality preamp. Use good film caps and low-noise metal film resistors throughout. The board will hold 0.5W resistors for the feedback dividers R3, R5 (positive regulator) and R16, R17 (negative regulator). Their dissipation is much lower, but the physically larger resistors ensure very low temperature drift and high reliability. All other resistors can be 0.25W types that offer adequate overcapacity.

The output voltage is set by the ratio of the two feedback resistors. You can adapt it for other voltages, as explained in the referenced articles. These voltages are normally not critical for powering nominally 15V DC circuits. You should, of course, stay below the maximum ratings. Most op amps used in audio have maximum

Supply Decoupling: A "Yes, But" Story

The proposed regulators use an extra RC decoupling for the op amp supply. (In the positive regulator, *Fig.* 2, these are R6 and C5, C2.) Such a network, often seen in low-level circuitry, has been the source of some controversy. I have done some research into the effects and offer my findings here, for whatever it's worth.

The network has two effects, working against each other. Whether the result is positive or negative therefore depends on the relative magnitude of the effects.

Let's first look at the raw supply. As discussed in the article, the raw supply line has AC components from the mains, and from the frequency- and level-dependent load current. (Because the raw supply has an internal impedance, the varying load current results in a varying ripple component.) We need to keep these unwanted artifacts out of the op amp supply. We can attenuate them greatly with the RC network mentioned before, which looks quite attractive.

Now consider the op amp. Generally, the current that the op amp draws from its supply varies with the frequency and level of the output voltage and current that the op amp must deliver. Part of these current variations are short-circuited by the RC network's capacitor, but some current variations will result through the resistor and thus cause a ripple voltage across this resistor. The ripple voltage that finally appears on the op amp's supply pin is the sum of this ripple, plus the one on the raw supply line.

Many factors determine the net result. The two ripple voltages could be equal in magnitude but opposite in phase and cancel each other, which makes the RC network extremely useful. On the other hand, they could reinforce each other, in which case you would be better off without it.

In considering the positive regulator (the following holds equally well for the negative supply), the base current for the pass transistor is furnished by the current source. This current is way too high, and the op amp must siphon off the excess. By taking away just enough to maintain the output voltage at the set value, regulation is obtained. The op amp output current flows from the current source *into* the output pin to ground; no current is drawn from the op-amp positive supply point. (The only current the op amp takes from its supply is for internal biasing, which basically is for constant-current sources.)

So, with the above in mind, I conclude that there is no frequency- or load-dependent op-amp supply current and no ripple component developed across the resistor of the RC network. My measurements confirm that, with the network, the supply-output ripple is slightly lower. Which is why I recommend its use here.—JD

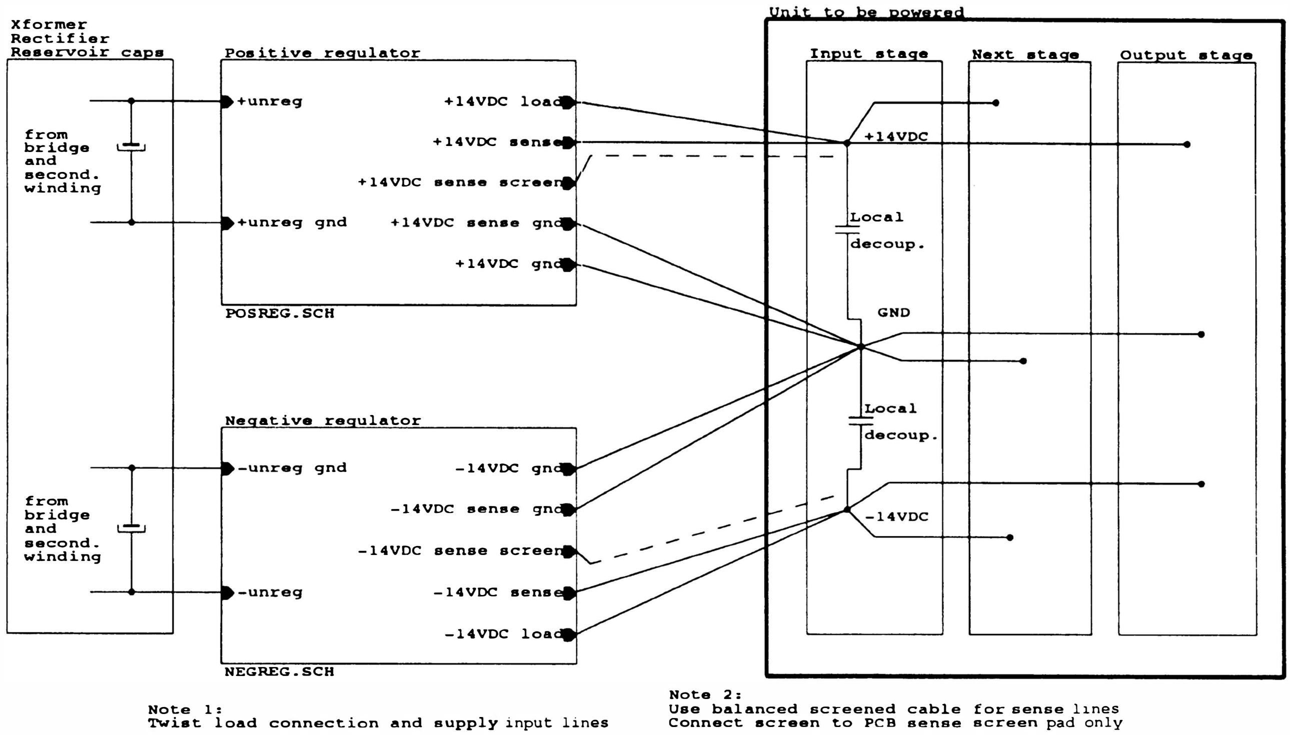

FIGURE 6: Generic connection diagram for one channel.

ratings of ±18V, and operation within a volt of the recommended ±15V DC is OK. A higher voltage causes higher dissipation, but not higher fidelity.

Make sure that the input supplies deliver at least 2V above the set output level under all load conditions. Higher input voltages don't improve the performance, but they do increase dissipation, which is undesirable. Discrete circuits have more variation in operating voltage. These boards can supply up to about 26V DC, with a raw input of 30V DC, limited by the AD848 operating maximum. The board implements the extra filter for the op-amp supply, and I recommend its use. The extra cost is small, and it provides additional attenuation of any "dirt" on the raw supply (see "Supply Decoupling" sidebar).

Getting Connected

Now I'll return to the issue of connecting the load. Your circuit may consist of several stages, and there is probably only a single connection for the supply. This may not be optimal for your purpose. However, if you take the time to examine the layout, you can locate the most sensitive stage, which normally is the input stage. This is where the supply voltage should be sensed. Toward the circuit's output stage(s), load current generally increases, and with it the ripple voltages.

Connecting the supply directly to the input stage has two important advantages. First, it ensures that the most sensitive stage gets the cleanest power available. Second, because of the extremely low supply-internal impedance (Z_O), any ripple voltages generated elsewhere are effectively short-circuited at the input stage. You should not underestimate this last effect. It prevents ripple voltage and signal residue from one channel's supply lines from entering the other channel.

The flip side of this is that you need to use a separate pair of supply boards for the left and right channels in the system. Only this will give you the cleanest supply at each channel and minimum interchannel crosstalk. You can realize the highest performance only at a single point: where the sense lines are connected.

Normally, a decoupling capacitor will be very close to the input stage, which is a good point to connect the sense and power lines. Connect the other stages of the particular circuitry you need to power to these points as well. *Figure 6* shows a generic example for one channel.

A few words on the raw supply are also in order. Ideally, you should use one transformer per channel, each transformer having two separate secondary windings (2 × 15V AC for a standard 15V DC supply). You could also use a transformer with four secondaries. Each secondary then would feed a diode bridge and a reservoir cap (*Fig. 6*). Combining regulator boards or transformer secondaries, or using center-tapped windings will be less effective, generally leading to ground loops.

What's in It for Me?

How does this compare to the best theoretical performance? Parts 1 and 2 addressed three significant performance indicators. The input-line rejection is not really influenced by the output/sense configuration you choose, but it depends mostly on the circuit topology itself. Therefore, my application duplicates the performance for this area.

A similar argument can be made for the output noise level. The circuit topology, specifically the selection of op amp and reference, and the reference filter, determines the noise performance. Again, this implementation essentially is equivalent to the "laboratory" results.

Large differences can occur in the output impedance Z_O. This is clearly illustrated by the plots I made with my Audio Precision system (*Figs. 7* and *8*). I used the same measurement technique (with the same software) that Walt Jung used for his plots.

In each plot, the graphs represent the Z_O as measured at the load. For the top graphs, the sense lines are connected at the board to the terminals of the R3, R5/R16, R17 feedback divider. This is the way you would normally connect a supply. In this case, the at-the-

ACKNOWLEDGMENTS

I am grateful to Walt Jung for involving me in this project. Most people do not realize how much time and effort are absorbed by such undertakings, and this was no exception. But working with like-minded people on projects of interest is quite enjoyable, and you always learn from it. Also, I am indebted to Gary Galo for his review of my manuscript and for testing my prototype boards. Finally, I recommend reviewing the references, which I won't repeat here, at the ends of Parts 1 and 2. They apply to Part 3 as well.

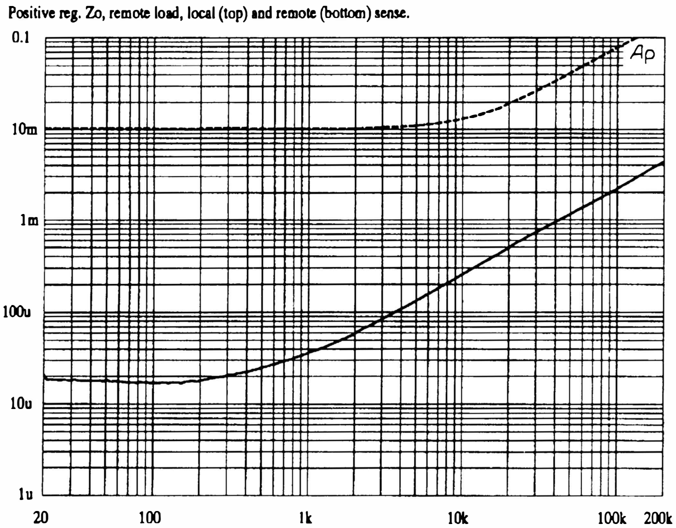

FIGURE 7: Positive regulator output impedance (Z_O) plots.

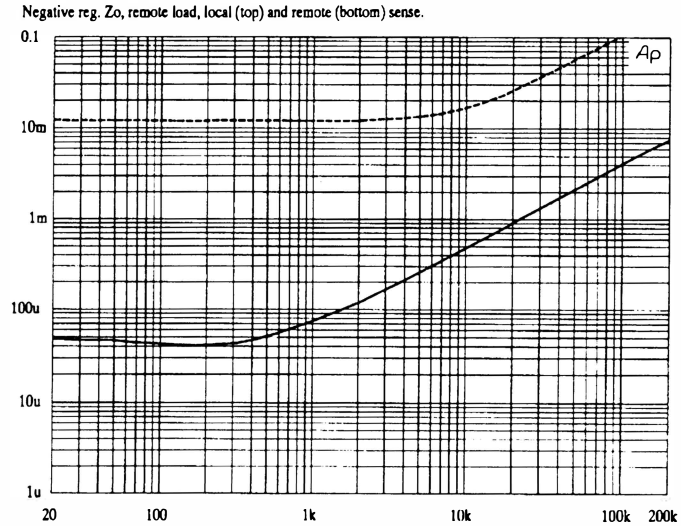

FIGURE 8: Negative regulator output impedance (Z_O) plots.

load Z_O is several orders of magnitude worse (higher) than the lab results.

However, when you connect the sense lines directly to the load point, you notice a vast improvement (the lower graph). It is not exactly equal to the lab curves, but it comes quite close. For both regulators, Z_O is now below 1MΩ up to 20kHz, where moving the actual sense points fractions of an inch on the load connections produces a readily measurable difference. These measurements are made with AD484 op amps in the prototypes. Using AD797 types gave similar or slightly better results.

But the measurements are only part of the story. At the end, we need an improvement in the sound of our systems. (See Gary Galo's article in the next issue on listening tests with the supplies.)

These regulators are as good as present state-of-the-art components permit. The limiting factor is the environment in which they are used, the connections to the load, and the lead lengths. With the remote-sensing setup, it is unlikely that significant improvements are possible in this "POOGE" manner.

The only improvement I can think of at this time is to design your circuits from the ground up with integral supply regulators for each stage and each polarity. Human nature being what it is, someone will probably eventually try that. Powering your circuits with these new regulators gives you an immediately clear improvement that will be very hard to surpass in the years to come. This is another step toward that elusive ideal, and it puts the bar another notch higher. □

Part 4

REGULATORS FOR HIGH-PERFORMANCE AUDIO:

Real-World Implementations and Sonic Evaluations

By Gary A. Galo
Contributing Editor

In the first three issues of 1995, *TAA* readers were given a wealth of information on state-of-the-art regulators for low-level audio applications. Now that you've seen Walt Jung's circuits (1/95, p. 8), analyzed his measurement data (2/95, p. 20), and built Jan Didden's clean, easy-to-assemble printed circuit board (3/95, p. 20), you're ready to drop a few of these regulators into real-world projects. Perhaps you've been reluctant to build them until a report on how they affect the sound was published. This, and more, is what Part 4 is all about.

Over the past several months I've amassed a great deal of experience building and implementing these regulators using Jan's PC boards. I have installed and evaluated ±14V versions in an extensively modified Adcom GFP-565 preamplifier, along with +5V, -5V, and -15V versions in a Philips DAC960 digital-to-analog converter modified to Pooge 5.5 standards.[1,2] From reading Parts 1–3, you know that these fast, ultra-wideband regulators require careful attention to parts selection and layout in order to perform to their potential.

As Jan's article so aptly illustrates, there are more "components" in a circuit than those shown in the schematic diagram, and much more to these regulators than input, ground, and output. Jan's Fig. 1 shows the other components that exist within a real-world regulator circuit, and even more phantom inductances, capacitances, and resistances can enter the picture once the regulator is installed in a real-world device. With a feature as sophisticated as remote sensing, the assembled boards can't be considered drop-in replacements for the old 7815 and 7915 three-terminal types. Connecting these regulators to real electronic circuits requires considerably more effort—and expertise—on the part of the builder.

Test Gear Necessities

At several points throughout the previous articles, Walt and Jan have mentioned the potential for oscillation of these regulators. With such sophisticated circuitry, we can no longer rely only on a digital voltmeter for regulator testing (though we certainly *do* need one). A wideband oscilloscope, 20MHz minimum, is an absolute must for verifying their proper operation. I know many of you will be tempted to build and install these devices without a scope, *but you must avoid this temptation at all costs.*

Jan's excellent PC board is extremely simple to assemble. Thanks to a clean layout and clear instructions, it resembles a "One Evening" novice Heathkit project, but its simplicity is incredibly deceiving. Implementing these regulators in an overall system design is no less than an advanced project. Whether you use the regulators as part of a new design or modify existing equipment, there's no substitute for proper test equipment.

A few of my experiences may help emphasize this point. In issue 2/86 I reviewed Phoenix Systems' P-94-SR Parametric Equalizer Kit.[3] This was a Heathkit-style project, in which builders assembled the factory-made printed circuit boards and mounted everything in factory-supplied cases. Since this was a finished kit design, you would expect it to operate perfectly when it was completed. Mine didn't.

Ever since I began working with fast, wideband op amps in the late 1970s, I have always checked power supply rails for oscillations. I found a problem when I checked these. Far from sophisticated, wideband regulators, this equalizer used the 7815/7915 pair. I fixed the problem by adding more local supply bypassing near the TL-074 quad op amps. Even foolproof, easy-to-use, three-terminal regulators can oscillate under the right (wrong?) conditions. You'll never know unless you check them with a scope.

A problem I had with a Nelson Pass A-40 power amplifier is also worth relating. Pass did everything he could to make its construction possible for the novice, even if no test equipment was available. If you didn't have a voltmeter to check DC offset, you could put a 470Ω, 1W resistor across the speaker terminals. No heat = safe offset.[4] When I measured harmonic distortion, one channel was excessively high above 12W output; below 10W the amplifier measured just as Pass specified.

The problem was amplifier oscillation caused by insufficient local supply bypassing. The Old Colony kit contained much larger heatsinks than those used by Pass in his prototype, so the leads to the output transistors were rather long. Adding bypassing from the output transistor cases to local ground solved the problem, and the amplifier sounded excellent.[5] Without the proper test equipment, I would never have known the amplifier was malfunctioning, and would probably have blamed the bad sound on the design. The design was not the problem, however; my particular layout created a situation the designer did not foresee.

Since Walt, Jan, and I are well aware of the potential for oscillation, it would be irresponsible of us to avoid stressing the test equipment issue. Every hobby requires an investment in the right tools and equipment, and good test gear is more affordable than ever. MCM Electronics' Tenma line includes a dual-trace "Trainer" scope with a 20MHz bandwidth (#72-905) for $335. They also have a complete line of digital voltmeters.

Heatsinks

In Part 3 Jan makes some suggestions for heatsinking the pass transistors, Q1 and Q2 (Q1 in Walt's diagrams). Jan uses a TO-220 heatsink made in Switzerland by Fischer Electronic. Since there isn't an exact replacement in the US, he recommends Aavid's HS-112. When I installed the ±14V regulators in my modified Adcom 565 preamp, I found the HS-112 inadequate for this load, which is approximately 100mA.

There's a fairly easy solution to this problem. The HS-112 has three fins per side. Aavid also makes the HS-114, with six fins per side. For even more heat dissipation, Aavid's HS-113 "booster" is used in conjunction with either the HS-112 or HS-114. It mounts on the top of the transistor, so heat can be dissipated from both sides of the metal tab. I recommend this combination for higher current applications.

The 1.5-inch-long HS-114 overhangs Jan's board, which may not work in some physical layouts. I find that it's very easy to cut off one or two pair of fins with a band saw or hacksaw to custom fit the heatsink to my own requirements. Be careful not to bend or twist the heatsink when you make the cut.

If you use a hacksaw, I recommend clamping the heatsink to a wood block with a #6 sheet metal screw and flat washer. Use one of the two holes in the heatsink, and tighten the screw just enough to hold the heatsink in place: you don't want to bend it. Then make the cut with a sharp hacksaw blade on the opposite end of the heatsink. When finished, remove any burrs or rough edges with a small file, and be sure the heatsink is completely free of metal filings. This is really a simple process and takes only a few minutes.

For the Adcom 565 preamp regulators, I trimmed HS-114s to five pair of fins. Four pair works fine for the 5V regulators I built. In both cases I used trimmed HS-114s along with the HS-113 booster. Your exact current requirements will determine just how much heatsinking is needed. With some ingenuity, you can easily fabricate parts from readily available sources. For example, Digi-Key stocks all three Aavid products and ships the same day.

If you ensure that there won't be any electrical contact between the heatsinks and any other point, particularly ground, you shouldn't need to insulate the pass transistors. This shouldn't be a problem in most cases, but make sure the ground plane on the PC board doesn't touch the heatsink. They were perilously close on the board samples I got from Jan, so I had to trim the positive ground plane. A single-edge razor blade works fine for this.

Use the insulators if you have any doubts. (You'll get slightly better heat transfer without them, however.) Jan suggests placing a plastic spacer between heatsink and board to solve this problem. This also isolates the board and other components from the heat, which is worth considering in high-current situations. Always use white silicone thermal compound, such as GC Electronics 8109 (available from Mouser or Newark). Metallic oxides contained in the white compounds facilitate heat transfer.

Preliminary Tests

It is worthwhile to bench test your regulators prior to installing them, as a malfunctioning device is much easier to troubleshoot before it is buried in a chassis. The raw supply I built for this purpose is shown in *Fig. 1*, with a parts list in *Table 1*.

This ±13V supply can be used for testing 5V and 14V regulators. To test the latter, use the full 26V available between the positive and negative rails; 5V regulators can be fed from the 13V rails. When conducting your tests, be careful to observe correct input polarity: you can damage the op amp and transistors with reversed polarity. The 1k bleeder resistors are overrated at 1W, but they stay cool, and the stiff leads make solid connection points for clip leads.

For bench testing, omit the remote sensing by jumpering load output to sense, and load ground to sense ground. You should also put a resistor across the output to pull 75–100mA from the regulators. Use 150Ω, 1W for the 14V versions, and 50Ω, 1W for the 5V ones. The load resistor values can be tailored to match the current drain in your specific application.

I find a Variac (VARIable AC transformer), shown in *Fig. 1*, extremely useful for testing. If you put a digital voltmeter across the regulator outputs, you can slowly increase the Variac's AC output while monitoring the regulator's DC output. When testing a positive regulator, be sure the output is actually going positive as you begin turning the Variac. If it isn't, power down and find the problem. The output of a negative regulator should begin to go negative as soon as AC is applied.

It's easy to reverse polarity on a bench setup connected with clip leads, but if you make a mistake the Variac avoids potential disasters. It also allows you to adjust the DC input to the regulator, duplicating the voltages which will be present in your equipment.

Once you have verified correct DC performance, check the regulator for oscillation. Set your oscilloscope for

TABLE 1

RAW DC TEST SUPPLY PARTS LIST

TRANSFORMER	
18V CT, 2.0A	(Mouser 41FJ020)
BRIDGE	
6A, 100PIV	(Digi-Key PB61-ND)
CAPACITORS	
0.47µF/100V	Panasonic V-series (Digi-Key P4733-ND)
2,200µF/25V	Panasonic HFQ (Digi-Key P5716-ND)
RESISTORS	
1k/1W	Yageo metal oxide (Digi-Key P1.0KW-1BK-ND)
VARIAC	
Tenma 10A	Variable Autotransformer (MCM 72-110)

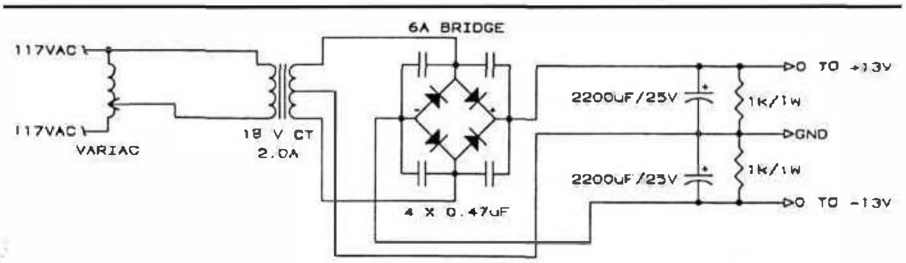

FIGURE 1: Raw supply for bench testing regulators prior to installation. The Variac is recommended to avoid damage if the circuit is not operating properly.

maximum sensitivity and minimum time base. On my scope these are 5mV and 0.2µs per division. Always set the scope for *AC coupling* in these tests. Also, if your scope's input coupling selector has a ground position, *make sure it is not in this position*. Check that no oscillations are present after the regulator has reached its rated DC output voltage.

You can check dropout voltage by connecting your digital voltmeter from the input to the output of the regulator. Set your scope to 20–50mV/division sensitivity and 2.0ms/division time base, and connect it to the regulator output. As you decrease the Variac output from 117V AC, you'll see the voltage difference between the regulator input and output decrease proportionally.

The scope trace will remain a straight line until the regulator "drops out" of regulation. At this point, 120Hz sawtooth ripple will appear on the scope, and will quickly rise in level as the input voltage is dropped further. The dropout voltage is the input/output voltage differential at the point where the ripple just barely appears. With a 100mA load, dropout can be anywhere from 1V to 1.8V in a properly functioning regulator. (There's more on the dropout issue later in this article.)

Transformers and Raw Supplies

Whether you use these regulators in a new design or modify existing equipment, you must make some decisions regarding the raw, unregulated portion of your power supply. In the past I have used toroidal transformers for practically all of my audio projects. These devices are extremely efficient, and, since they concentrate the magnetic field in the core, radiate a low hum field. Rick Miller, author of the sidebar on rectifier diode noise which accompanies my Pooge 5.5 article,[6] has been measuring power transformer bandwidths. He concludes that we are barking up the wrong tree with power toroids.

As it turns out, toroidal transformers are wideband devices which are extremely effective at transferring power line noise to equipment. *Figure* 2 (prepared by Rick on an Audio Precision System 1) is a comparative frequency response plot of the two transformers. It illustrates the problem quite dramatically. The top, dashed trace is an Avel-Lindberg D-

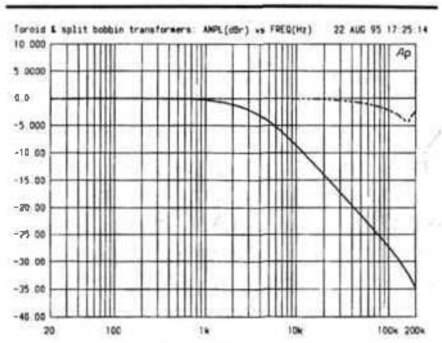

FIGURE 2: Bandwidth measurements on conventional and toroidal power transformers: dashed trace (top) is an Avel-Lindberg D-3022 toroid, flat to nearly 200kHz; solid trace (bottom) is a Magnetek FD7-36, nearly 35dB down at this frequency. (Courtesy of Rick Miller).

3022 toroidal transformer, which is nearly flat to 200kHz. The bottom, solid trace is a Magnetek FD7-36, a split-bobbin design which is part of their Quick Pack series. The Magnetek is nearly 35dB down at 200kHz, with a –3dB point around 4kHz, while the Avel-Lindberg's –3dB point is well above 100kHz. I don't mean to single out Avel-Lindberg in this example; toroidal transformers from other sources have similar characteristics.

Rick's measurements show that Signal Transformers' A-41 series offers even more effective high-frequency noise attenuation. These dual-bobbin designs have two independent primary and secondary bobbins. This greatly reduces the capacitive coupling between them, which is extremely important for attenuation of common mode noise. Split- and dual-bobbin construction eliminates the need for expensive—and not nearly as effective—electrostatic shielding.

For more information on the effects of transformer construction on noise transfer, Rick suggests Topaz Electronics' *Noise Suppression Reference Manual*.[7] It makes two important points: "physical separation of coils placed side by side on separate legs of the magnetic core of a transformer will provide far less capacitive coupling than coils wound directly over one another"; and related to electrostatic shielding, "capacitance around the Faraday shield will still couple enough noise from the primary to secondary to cause problems in sensitive equipment."

The capacitive coupling between transformer windings is inversely proportional to the transformer's hi-pot rating. ("Hi-pot" is short for *high potential*, the point at which the dielectric material—in this case the enamel insulation on the transformer wire and the bobbin itself—breaks down.) The higher the rating, the lower the coupling. Magnetek Quick-Pack transformers have a hi-pot rating of 2.5kV RMS; Signal's A-41 series is rated at 4kV RMS. (Magnetek transformers are available from Mouser; Signal sells factory-direct.)

Beyond the selection of the power transformer, we now recommend raw supplies even more sophisticated than those of Pooge 5.5. A raw supply for ±14V supplies is shown in *Fig.* 3, with a parts list in *Table* 2. A unique feature is the common mode chokes on the DC side of the rectifier bridges, another Rick Miller innovation. These chokes are 56mH Panasonic types, carried by Digi-Key. The 0.47µF capacitors on the input line filter are special 250V AC Panasonic Interference Suppression caps.

Further sonic improvements are noticeable when common mode chokes are used between the rectifier diodes and the input filter capacitors, as in *Fig.* 3. Note the absence of large film capacitors directly across the transformer secondaries. They are unnecessary with the common mode chokes, and can

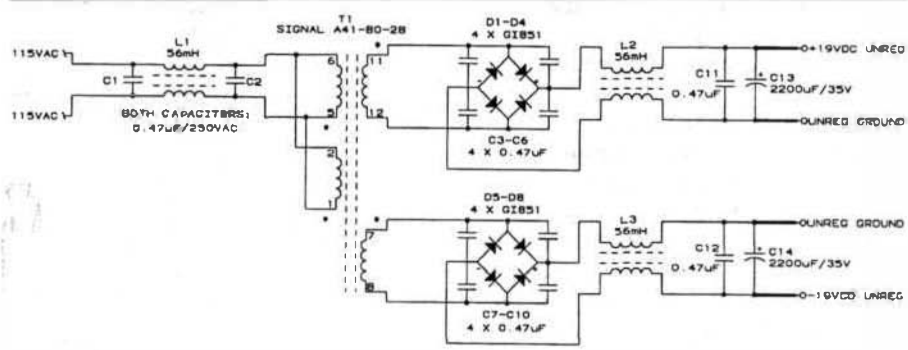

FIGURE 3: Raw supply suitable for use with high-performance regulators. Common mode chokes are used for AC line filtering and DC filtering after the rectifier bridges.

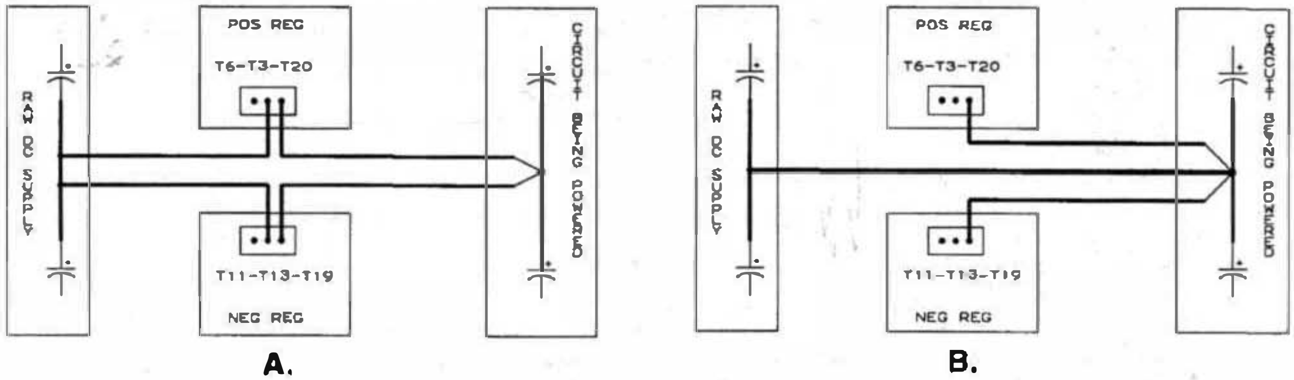

FIGURE 4: Two methods for connecting the regulators in equipment with a common raw DC ground. A: Incorrect method; B: correct method, recommended to avoid a ground loop.

cause high-Q resonance problems with these transformers. As outstanding as these regulators are in terms of line rejection, effective power line filtering and low-noise rectifier diodes are still sonically beneficial.

In Part 3 Jan shows the best method for connecting these regulators to the raw supplies and the powered circuitry. *Figure 3* is consistent with his recommendations, since it has separate bridge rectifiers for the positive and negative raw supplies; the unregulated supplies do *not* share a common ground. If you are working with existing equipment as part of a modification project, you may be forced to use a raw supply with a single rectifier bridge and a common ground.

Figure 4 shows two options for connecting these regulators in such cases. "A" is *not* recommended, since there are two ground paths between the raw supply and the powered circuitry. The correct method is shown in "B," where one ground lead is run from each regulator board to load ground, or the existing equipment's main

TABLE 2

HIGH-PERFORMANCE, DUAL-POLARITY RAW SUPPLY PARTS LIST

L1, L2, L3	Panasonic 56mH common mode line filter (Digi-Key PLK1017-ND)
C1, 2	Panasonic 0.47µF/250V AC interference suppression (Digi-Key P4614-ND)
C3-12	Panasonic 0.47µF/100V V-series (Digi-Key P4733-ND)
C13, 14	Panasonic 2,200µF/35V HFQ (Digi-Key P5751-ND)
D1-8	General Instrument GI-851 (Digi-Key GI851CT-ND)
T1	Signal Transformer A41-80-28

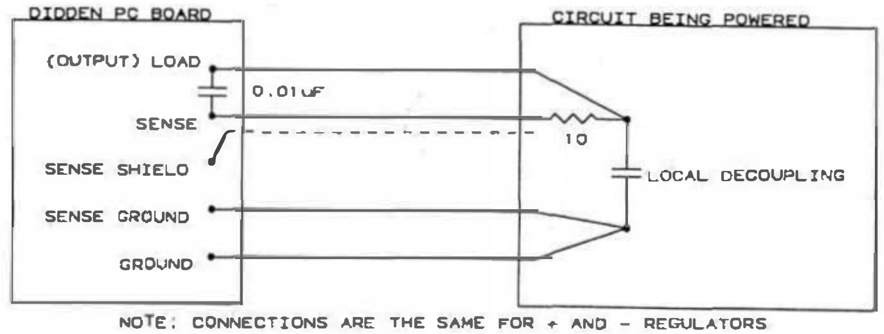

FIGURE 5: Remote sense decoupling scheme. Connections and parts values are the same for positive and negative regulators.

power supply ground bus. I used this method for the Adcom GFP-565 preamp, since it eliminates the possibility of a ground loop.

Op Amps and Decoupling

Walt offers builders a choice of two op amps: Analog Devices' AD848 and AD797. The graphs in Part 2 show the latter to be superior in virtually all aspects of performance. You may wonder just how audible these effects are, but in my listening tests I also found the 797 to be the superior sonic performer. (I elaborate on this later in the article.)

Remember that the op amp is powered from an *unregulated* supply. So regardless of which one you choose, its own power supply can affect the overall regulator performance. In Part 1 Walt provides an optional low-pass decoupling, consisting of a 22.1Ω resistor in series with the op amp supply, and a 120μF electrolytic capacitor added for local bypassing. This R/C combination produces a first-order, low-pass filter with a corner frequency of 60Hz.

In Part 3 Jan discusses the pluses and minuses of powering the op amp through a series resistor. He concludes that, in this application, op amp supply current is neither frequency- nor load-dependent; therefore, no ripple currents will develop across the 22.1Ω resistor. I have heard ±14V regulators both with and without decoupling, and the former is sonically superior. Walt's Fig. 12b shows a dramatic improvement in line rejection with it, with the AD797 performing substantially better than the AD848 at low frequencies. Jan's board accommodates the decoupling, and I highly recommend its use.

Oscillations and Decoupling

The remote sensing capability is a highlight of these regulators, but the potential for oscillation in the megahertz region exists in any wideband op amp. The AD797 is particularly susceptible. Supply oscillation is governed by any number of layout-related issues, including length and inductance of the remote sensing wiring, shielding characteristics, and the amount of low-ESR local supply bypassing. *There's no way to predict whether the supply will oscillate—each implementation must be tested.*

Once the regulator has been installed, connect your scope probe between load and load ground on the regulator PC board. Use the same oscilloscope settings as previously noted in the preliminary tests. Rotate the triggering until a stable trace appears. A virtual straight line indi-

±5V Supplies Using the Didden PCB

By Walt Jung

The original 5V regulator published as Fig. 9 in Part 1 is a good positive device, capable of very low noise when used with the AD780. Unfortunately, no board design has specifically addressed this circuit for 5V use. In addition, this type of design, which is based on a three-terminal IC reference, cannot be "flipped" to provide negative or –5V outputs. Many modern designs *do* need high-quality ±5V sources, for example, the Youtsey et al mods to the DAC-in-the-Box (*TAA* 4/94, p. 8).

Fortunately, some fairly simple adaptations to the original article's Fig. 8a or 8b can provide the desired functionality to achieve ±5V operation. What is even more fortuitous is that these changes can be readily implemented by simple part substitutions on the Jan Didden PCB design in Part 3.

This adaptation implements the lower output voltage (+) or (–) 5V versions by using a 2.5V reference diode in place of the original 6.9V LM329. Specifically, the industry standard LM336, a two-terminal 2.5V reference IC, allows this. It can simply be substituted in the same footprint as the LM329 on Jan's board. As long as the surrounding support circuitry is fully compatible with the lower voltage operation, this can implement a very high quality +5V or –5V regulator, with the same ease of construction as noted by Gary.

The specific changes to the positive/negative regulator circuits which accomplish these goals are listed here. Note: In adapting the original circuits to 5V output, change *only* the following items; leave all other details as originally published. (Complete part numbers and order information appear in *Table 3.*) In each step, the first reference designation pertains to Part 3, Fig. 2 (positive regulator); the designations in parentheses refer to Part 3, Fig. 3 (negative regulator); the original reference designations to Figs. 8a and 8b from Part 1 are in brackets.

1. Change D1 (D6) [D1] to an LM336 2.5V TO-92 diode type, polarizing it as shown in the original schematic. Special note: Do *not* connect the adjust pin.

2. Change R4 (R14) [R6] to a 2.49k 1% metal film type.

3. Change R20 (R21) [R5] to a 4.99k 1% metal film type.

4. Change X1 (X2) [U1] to an AD797. This is optional in terms of basic functionality, and the circuit also works with the AD848. Do *not* substitute other op amps, as the input CM range must be compatible with 2.5V operation!

5. Delete the D2 and D3 (D4 and D5) [D2 and D3] 1N4148 diodes when/if using the AD797. Retain them if using the AD848.

6. Low-dropout operation is highly recommended for the ±5V regulators, and should be implemented *using all three of the steps outlined above.*

These changes affect DC operation for the most part, so AC performance can generally be expected to be consistent with what has already been published for the original Fig. 9 circuit.

cates the supply is working correctly; an oscillating one won't be subtle. If the regulator is powering digital circuitry, random digital hash on the supply lines is normal. An oscillation will be repetitive, at a specific frequency. A scope with good triggering should enable you to get a firm sync on any oscillation which may be present.

Don't be surprised to see some very low level ripples toward the left edge of the screen, even if the supply is working properly. The probable cause is the test setup's ground lead inductance. To verify, connect the scope probe to the chassis ground near the same point as the ground lead. You should see the same low-level ripples as before, particularly on a wideband scope, but you now know that this is a function of the test setup rather than the regulator.

While I found that 797-based regulators oscillate in some cases and not in others, I was determined to make this op amp work properly. Walt and I agreed on the need to ensure that the 797 could be reliably implemented—*with remote sensing*—in a variety of layouts. To solve the oscillation problem, Walt devised an implementation for decoupling the remote sensing lines at very high frequencies (*Fig.* 5). The polarity of the regulator isn't defined, since both positive and negative use the same decoupling topology. The local AC bypass removes the remote sense feedback path, and its associated phase shift, at very high frequencies.

First, solder a 0.01µF Panasonic V-series stacked film capacitor between the load (output) and sense pads on the PC board. You can solder this small cap to the board's foil side. Use insulating sleeving, particularly on the negative board where the cap must jump over a PC trace. Next, insert a 10Ω, ¼W resistor in series with the remote sense line *at the load*. This R/C network results in a –3dB point of 1.6MHz; the regulator still qualifies as a wideband audio device, but the chance of oscillation is greatly minimized.

With remote sensing, I recommend this decoupling regardless of which op amp you choose. Since no one can predict the effect of every possible layout and implementation, you must still check the supplies with a scope. The 10Ω resistor actually changes the DC gain of the op amp and raises the output voltage. The change is very slight, though ~10Ω is the tolerance of the 1k feedback resistor. Even with worst-case tolerances, a 5V regulator is well within safe operating limits for logic circuits.

A final note on oscillation: even regulators built with the AD848 can oscillate if a low-Z film cap is placed directly across the regulator output (as noted in Part 2, p. 34). *It is very important to build the PC board as specified!* Don't be tempted to add any film bypassing to the input or output. A low-Z film cap across the output electrolytic will virtually *guarantee* regulator oscillation. If the regulator is within 2" of the powered circuitry, don't use any local film bypassing, either. Local film bypass capacitors should be at least 3–4" from the PC board. In difficult circumstances a ferrite bead between the regulator output and the load can be helpful. Jan's sidebar offers some helpful suggestions for oscillation problems.

Dropout Warnings

When I first tested regulators built with the AD797, I found that the dropout voltages were not as low as those noted by Walt (his measurements were based on the AD848).

With a 100mA load, my positive regulators measure as high as 2V, whereas 1.5V or less is typical of the AD848, even with loads of several hundred milliamps. The 797 requires more input headroom than the 848, primarily because of differences in its output stage design.

The op amp in Walt's Fig. 8a must bias up to nearly the same DC potential as the output DC voltage, since the V_{BE} of Q1 and the V_F of D4 essentially cancel. As the input voltage V_S is lowered, the output swing limitation of U1 can limit the regulator dropout if the op amp voltage limit with respect to its supply rail is significantly higher than 1V (the dropout of the current source, Q2, D5, and R7). The 797's output can't swing as close to its rail voltage as the 848, which results in higher dropout voltage for the regulator.

The following enhancements are suggested to improve the dropout voltage with both the AD797 and the AD848 op amps. (Part numbers are listed in *Table 3*.) These changes are listed in order of decreasing sensitivity. In each step, the first reference designation pertains to Part 3, Fig. 2 (positive regulator); the second designation in parentheses refers to Part 3, Fig. 3 (negative regulator); the original reference designation to Figs. 8a and 8b, from Part 1, appears in brackets.

1. Change D7 (D9) [D4] to a Panasonic 30mA green LED. This enhances the output swing of the op amp. Don't be tempted to substitute another LED: the 2V drop across the specified part serves as a level shift, and is critical.

2. Lower R20 (R21) [R5] to 10k or 12k. This yields a small improvement, typically 0.05–0.1V. Note that Jan has already made this change in Part 3. Don't worry about this unless you built a regulator from Walt's Figs. 8a or 8b, and low dropout is an issue in your application.

3. Change current source Q3 (Q4) [Q2] to a device with lower V_{SAT}. This improves dropout by about the same amount as the resistor change in step 2. Recommended transistors are PN2907A (PNP–Q3) for positive regulators and PN2222A (NPN–Q4) for negative.

With all three changes, the dropout voltage will be close to 1V, but the first two steps will get you most of the way. You don't need to change Q3 and Q4 unless you absolutely must squeak out another 0.1V or so. Low dropout won't be important if your raw input rail voltages are high enough. The flip side of this coin is heat dissipation. One of the advantages of low-dropout regulators is they enable you to use lower raw DC voltages. Most of the pass transistor's heat is produced supplying current to the load, rather than dropping voltage. Remember that for a given current drain from the regulated output, the heat will increase as the raw DC voltage is raised.

TABLE 3

PARTS LIST FOR MISCELLANEOUS REGULATOR CHANGES

REMOTE SENSE DECOUPLING	
0.01µF/50V	Panasonic V-series capacitor (Digi-Key P4513-ND)
10Ω/¼W, 1%	Yageo metal film resistor (Digi-Key 10.0X-BK-ND or Roederstein MK2)
LOW-DROPOUT MODIFICATION	
30mA	Panasonic green LED (Digi-Key P309-ND)
	PN2907A transistor, TO-92 case (Digi-Key PN2907A-ND)
	PN2222A transistor, TO-92 case (Digi-Key PN2222A-ND)
10k, ¼W, 1%	Yageo metal film resistor (Digi-Key 10.0KX-BK-ND or Roederstein MK2)
±5V VERSIONS	
	National Semiconductor LM336BZ-2.5 (Digi-Key LM336BZ-2.5-ND)
2.49k, ¼W, 1%	Yageo metal film resistor (Digi-Key 2.49KX-BK-ND or Roederstein MK2)
4.99k, ¼W, 1%	Yageo metal film resistor (Digi-Key 4.99KX-BK-ND or Roederstein MK2)

SOURCES FOR OTHER REGULATOR PARTS

Analog Devices AD797AN and AD848JN (Newark Electronics, both items listed in current catalog #114)
Panasonic 120µF/25V HFQ capacitors (Digi-Key P5698-ND)
0.1µF/50V Panasonic V-series capacitors (Digi-Key P4525-ND)
LM329 Reference (Digi-Key LM329DZ-ND)
1N4148 Diodes (Digi-Key 1N4148-ND)
2N5087 Transistor (Digi-Key 2N5087-ND)
2N5089 Transistor (Digi-Key 2N5089-ND)
D44H11 Transistor (Mouser 570-D44H11)
D45H11 Transistor (Mouser 570-D45H11)
Roederstein MK3 series, 1k, 0.5W (Michael Percy)

Other Voltages

These regulators can be adjusted for other output voltages, though ±14V and ±5V should cover most audio applications. Table 1 in Part 2 supplies alternate resistor values for voltages from 10V to 18V.

Some of you may be tempted to raise the rail voltages beyond ±14V for preamp power supplies. With the gain

So It Oscillates—Now What?

By Jan Didden

As Gary explains, you should definitely check your supplies for oscillations. They are not inherently unstable, but with so many variables there is always a chance. Use the following checklist to systematically review and remove the possible causes.

1. Assuming you use the proposed PCB, did you build it as described in Parts 3 and 4? No extra film caps should be placed at the regulated output!

2. Have you limited the lead lengths as much as possible? Preferably, they should be no longer than 4 or 5". Be sure to twist the raw supply lines and the regulated load connections, *but not to each other*. Do not connect the sense shield to the ground point at the load, only at the PCB.

3. Mount the board(s) over a metal enclosure wall or partition, as close as possible. This will decrease any oscillatory tendencies, and also improve the noise figure.

4. Check the circuit to be powered for excessive decoupling capacity. With these very low impedance regulators, more than 100µF or so is overkill, and promotes instability. If you use film caps at the load, don't make them larger than 1µF or so. I know it goes against the grain to actually remove a film bypass cap. Although they are a solution to many problems, in this application they can actually cause problems.

5. The remote sense decoupling filter should take care of any remaining oscillations, but in persistent cases you can increase the resistor value to 22Ω and the capacitor to 0.015µF, with negligible impact on performance.

6. The AD848 in this application is a bit more stable than the AD797, so if all else fails this could be a solution. As Gary notes, this is not as good sonically.

7. Finally, don't get discouraged. We have built many of these regulators, and every one could be persuaded to work as advertised.

determining resistors set at 1k each, the op amp has a voltage gain of 2. When the reference is the 6.9V LM329, this actually produces 13.8V. The 10Ω resistor in the remote sense decoupling brings it up to around 13.9V.

In most cases, there's no need to operate preamp power supplies at higher rail voltages. Even with these values, my modified Adcom GFP-565 outputs close to 8V RMS before clipping. Since most power amps clip with 2.5V input, it is pointless to raise the preamp rail voltages. Contrary to what some believe, higher rail voltages won't give you more "headroom" if the next device (i.e., your power amp) can't handle the input signal.

Listening Evaluations

A set of ±14V regulators has been installed in my extensively modified Adcom GFP-565 preamp for over six months. Throughout most of the modification process (subject of a future article), I have used a second, unmodified GFP-565 for comparison. Installing the high-performance regulators in the 565 affected nearly every aspect of performance, with the most striking improvement in the area of dynamics. I was amazed to find that, even though the line stage gains in both units were identical, the modified 565 actually played louder than the stock preamp. This may seem strange at first, but there is a logical explanation.

The new power supplies offer a sense of unrestricted dynamics; full orchestral crescendos are rendered with a sometimes overwhelming impact. Subjectively, the original supply regulators compress orchestral *tutti* passages, whereas the modified preamp releases them with full force. I was repeatedly lowering the volume relative to the stock preamp to achieve the same subjective playback levels. This was frustrating, since I was now playing most of my reference CDs with the volume around 9:00, leaving little room for adjustment. Walt and I have actually lowered the voltage gains in our line stages from 11 (a 10k/1k feedback combination) to 5 (4k and 1k) to allow a greater range of volume control adjustment.

There's more to the dynamic improvements than sheer volume, however. With the new regulators, the preamp sounds effortless no matter how taxing the source material. Even with the most demanding recordings, it remains clean and detailed, free of harshness or edge.

These super-quiet regulators also lower the subjective noise floor: low-level dynamics aren't artificially elevated, they simply descend effortlessly. In "Siegfried's Funeral March" from Wagner's *Götterdämmerung* (Solti, London CD 414-115-2), it is easy to miscalculate the dynamic contrasts. If you adjust the volume so the soft timpani notes (CD 3, Track 7) are at a comfortable level, the *fortissimo* at the climax can be unbearably loud. With the climaxes set at a realistic level, the

ACKNOWLEDGMENTS

Many thanks to Walt Jung and Jan Didden for their excellent and hard work throughout this project. Their involvement did not end with the publications of Parts 1, 2, and 3, but continued until the completion of this article, including (but hardly limited to) proofreading of this manuscript. Special thanks to Rick Miller for providing the transformer measurements, and considerable information on related issues, and to Walt for preparing the laser printout from Rick's .HGL file. Rick's important work on common mode chokes in raw DC supplies is also greatly appreciated.

entire dynamic range sounds much closer to a real concert hall experience.

The new analog regulators also increase soundstage size, both left-to-right and front-to-back. Prior to installing them, I had difficulty separating the bass drum from the timpani in "The Hut on Fowl's Legs" from Mussorgsky/Ravel's *Pictures at an Exhibition* (Reiner, RCA Victor Living Stereo CD 61958-2, Track 14), not in terms of timbre but of localization. The placement of these instruments is now reproduced with pinpoint accuracy, and considerably deeper in the soundstage than before. Inner detail and articulation are also improved.

To compare the AD848 and AD797 op amps, I soldered machined-pin, gold-plated sockets to the ±14V regulator board. I then soldered the op amps to gold-plated headers (with Caig ProGold contact conditioner on the header pins) for a gold-on-gold contact.

The 797 reveals greater inner detail from recordings than the 848. Its sonic presentation is also a bit more "laid back," more natural and musically convincing. The soundstage is not only deeper, it seems to have been moved back slightly. The 848's presentation is closer and more "forward." The line rejection measurements bear out these differences, and it's not surprising that the results of improved line rejection are similar to those of better power line filtering.

DAC Regulators

I installed and evaluated the new digital regulators in the DAC960 in three phases: a +5V regulator for the demodulator board; +5 and −5V regulators for the TDA1541A DAC chip; and a dedicated −15V supply for the TDA1541A. All digital regulators use the AD797. Instructions for this process are beyond the scope of this article. If there is sufficient interest, I'll prepare an "Ask *TAA*" column on the subject. Write to me (c/o *TAA*) if you would like to see this published.

Based on my experience upgrading the original digital supplies, I expected similar improvements from these changes, and the DAC regulators would offer "more of the same." My assumptions were wrong. Each of the upgrades produced different results. All listening evaluations were conducted with Analog Devices AD1890 evaluation board for jitter suppression.[8]

The real surprise is the demodulator regulator, which yields an improvement in dynamics and bass similar to the analog regulators in the 565 preamp. The effect on weight and impact in the bass region is like getting better subwoofers or a new power amp. The bass drum in Reiner's *Pictures* is deeper and more powerful. In Ernest Ansermet's recording of Ravel's *Alborada del gracioso* (London CD 433-717-2), the bass drum, while always quite impressive, is now even heftier than I had realized.

I was completely surprised at the ability of a digital regulator to make such a striking improvement in the bass. This is undoubtedly jitter-related, since the demodulator board's input switching circuitry and input receiver should have a significant effect on jitter performance.

The Ansermet recording is also incredibly clean and well-defined, not just in the bass but across the entire spectrum. The ±5V supplies for the 1541A DAC result in improved articulation, detail, and the sense of air and space around the instruments. Ravel's colorful orchestrations are reproduced with an openness and transparency which are uncanny, and the delicacy of his scoring is far more evident here than in my British-pressed London Treasury Series LP, which sounds dull and lifeless by comparison (most of London's "Made in England" Richmond and Treasury Series LPs were extremely good; when they began pressing these LPs in the US in the mid-1970s, the sound quality became abysmal).

Track 2 of Reiner's *Pictures*, which I regularly use as reference material, has a series of four string *glissandi* bowed close to the fingerboard. Prior to installing the ±5V DAC regulators, the effect sounded merely like fingers sliding up and down the strings. Now I can clearly hear the subtle articulations of the bow. Such a soundstage subtlety often goes unnoticed. The four *glissandi* begin with the violas, move to the celli, then the second and first violins. On a less refined D/A converter, there appears to be a general movement from right to left. With these regulators, the exact placement is clear: right, far right, left, far left. Musical subtleties, carefully notated by Ravel and superbly executed by the Chicago Symphony, are revealed in all their glory by the DAC960.

I have noticed an amazing number of tape edits on CDs made from analog sources, which, prior to installing the DAC regulators, had escaped my attention. One example is the EMI reissue of Boris Christoff's 1962 stereo remake of Mussorgsky's *Boris Godunov* (André Cluytens, CDS7 47933-8), during Boris' Monologue in Act II (CD 2, Track 6). Another occurs in the Solti *Götterdämmerung*, in the final scene in Act II (CD 2, Track 12). Quite a number of tape edits are audible on the Decca/London operas produced by John Culshaw. Some of these edits were not obvious until I replaced the DAC regulators. In fact, the DAC960's ability to resolve minute details is so great that I sometimes hear off-axis microphone colorations as the singers move around the soundstage.

Icing on the Cake

The sonic effect of the new digital regulators is nothing short of dramatic, every bit as important as the analog regulators in my preamp. After spending nearly a week with the DAC960 in this state, I decided to give the TDA1541A a dedicated −15V supply. While this effect was more subtle, it was nonetheless worthwhile. Low-level resolution and detail were enhanced a bit further, with the last ounce of performance squeaked from the 1541A.

The −15V supply is critical, since it is the voltage source for the DAC's current outputs. If you check this sup-

SOURCES

Digi-Key Corp.
701 Brooks Ave. S., PO Box 677
Thief River Falls, MN 56701-0677
(800) 344-4539, FAX (218) 681-3380

MCM Electronics
650 Congress Park Dr.
Centerville, OH 45459-4072
(513) 434-0031, FAX (513) 434-6959

Michael Percy Audio Products
Box 526, 170 Highland
Inverness, CA 94937
(415) 669-7181, FAX (415) 669-7558

Mouser Electronics
958 N. Main St.
Mansfield, TX 76063
(800) 346-6873

Newark Electronics
(312) 784-5100
(Call for branch nearest you)

Signal Transformer
500 Bayview Ave.
Inwood, NY 11696
(516) 239-5777, FAX (516) 239-7208

ply line with a scope, you'll see some low-level digital hash. This will also appear on the ±15V rails to the analog circuitry if they share a common supply. Part of the perceived improvement may be due to removal of noise from the analog supply rail.

Some of you may wonder why I continue to modify a digital-to-analog converter that uses an "obsolete" digital chip set. Some truly wonderful digital filters and D/A converter chips are now available, with 20-bit resolution, 8x oversampling, and such. Several manufacturers produce D/A converters equipped with HDCD decoding capability.

Having upgraded the regulators in the DAC960 and heard their sonic effects, I honestly believe no one realized the full potential of the Philips 16-bit chip set while it was in production—least of all Philips. I recently auditioned an $800 HDCD D/A converter made by a leading American manufacturer which was inferior to my DAC960 in every respect, even before the high-performance regulators were installed. It used three-terminal adjustable regulators for the analog circuitry and 7805 types for the digital circuits.

There is a lesson to be learned from all of this: digital circuitry, no matter how sophisticated, will never perform to its potential with cheap, low-end power supply regulators. Manufacturers of both digital recording and playback equipment need to seriously reconsider the criticalness of power supply regulation to the performance of high-end digital circuitry. If you own a DAC960 modified to Pooge 5.5 standards, I believe the digital supply upgrades are well worth installing. You may find that the DAC960—once again—outperforms many expensive products with more sophisticated digital circuitry.

Conclusions

A great deal of discussion about the virtues of shunt regulators has occurred in the audio press over the past few months. Since these devices require a series resistor terminated by a shunt capacitor, they automatically provide low-pass filtering of power line and rectifier noise. As important as this may be, several other critical performance areas must be addressed. Before choosing shunt regulators for your next project, I suggest verifying their measured performance in *every* area, as discussed in Part 2 of this series.

Walt Jung's high-performance power supply regulators set new standards of performance in both analog and digital applications. They require a great deal from the builder, yet I have found the sonic improvements worth every hour spent on this project. I hope that you agree. The cumulative improvement in my audio system has truly been a revelation. ■

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4. N. Pass, "Pass/A40 Power Amplifier," *TAA* (4/78): 4.

5. G. Galo, "Distributor Delight," *TAA* (3/86): 52.

6. R. Miller, "Measured RFI Differences Between Rectifier Diodes In Simple Capacitor-Input Power Supplies," *TAA* (1/94): 26.

7. *Noise Suppression Reference Manual*. Topaz Electronics, 1660 Scenic Ave., Costa Mesa, CA 92626, (714) 557-1636.

8. G. Galo, "Ask *TAA*," *TAA* (4/94): 41.

The FET Constant-Current Source/Limiter

Introduction

The combination of low associated operating voltage and high output impedance makes the FET attractive as a constant-current source. An adjustable-current source (Figure 1) may be built with a FET, a variable resistor, and a small battery. For optimum thermal stability, the FET should be biased near the zero temperature coefficient point.

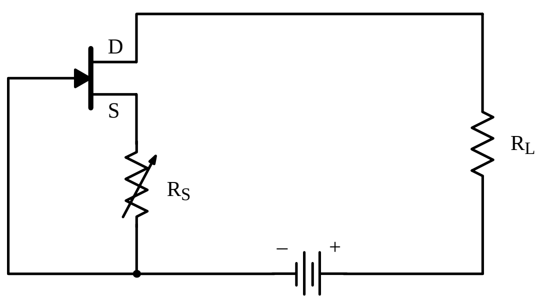

Figure 1. Field-Effect Transistor Current Source

NO TAG

Whenever the FET is operated in the current saturated region, its output conductance is very low. This occurs whenever the drain-source voltage V_{DS} is at least 50% greater than the cut-off voltage $V_{GS(off)}$. The FET may be biased to operate as a constant-current source at any current below its saturation current I_{DSS}.

Basic Source Biasing

For a given device where I_{DSS} and $V_{GS(off)}$ are known, the approximate V_{GS} required for a given I_D is

$$V_{GS} = V_{GS(off)} \left[1 - \left(\frac{I_D}{I_{DSS}} \right)^{1/k} \right] \tag{1}$$

where k can vary from 1.8 to 2.0, depending on device geometry. If K = 2.0, the series resistor R_S required between source and gate is

$$R_S = \frac{V_{GS}}{I_D} \quad \text{or} \quad R_S = \frac{V_{GS(off)}}{I_D} \left(1 - \sqrt{\frac{I_D}{I_{DSS}}} \right) \tag{2}$$

A change in supply voltage or a change in load impedance, will change I_D by only a small factor because of the low output conductance g_{oss}.

$$\Delta I_D = (\Delta V_{DS})(g_{oss}) \tag{3}$$

The value of g_{oss} is an important consideration in the accuracy of a constant-current source where the supply voltage may vary. As g_{oss} may range from less than 1 μS to more than 50 μS according to the FET type, the dynamic impedance can be greater than 1 MΩ to less than 20 kΩ. This corresponds to a current stability range of 1 μA to 50 μA per volt. The value of g_{oss} also depends on the operating point. Output conductance g_{oss} decrease approximately linearly with I_D. The relationship is

$$\frac{I_D}{I_{DSS}} = \frac{g_{oss}}{g'_{oss}} \tag{4}$$

$$\text{where } g_{oss} = g'_{oss} \tag{5}$$

$$\text{when } V_{GS} = 0 \tag{6}$$

So as $V_{GS} \rightarrow V_{GS(off)}$, $g_{oss} \rightarrow$ Zero. For best regulation, I_D must be considerably less than I_{DSS}.

Cascading for Low g_{oss}

It is possible to achieve much lower g_{oss} per unit I_D by cascading two FETs, as shown in Figure 2.

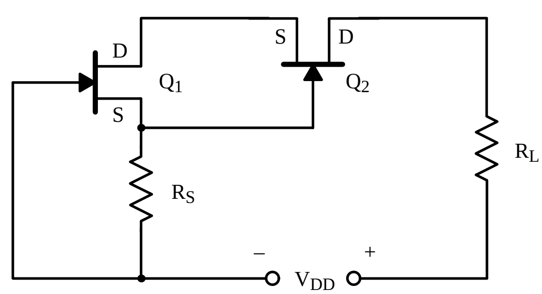

Figure 2. Cascade FET Current Source

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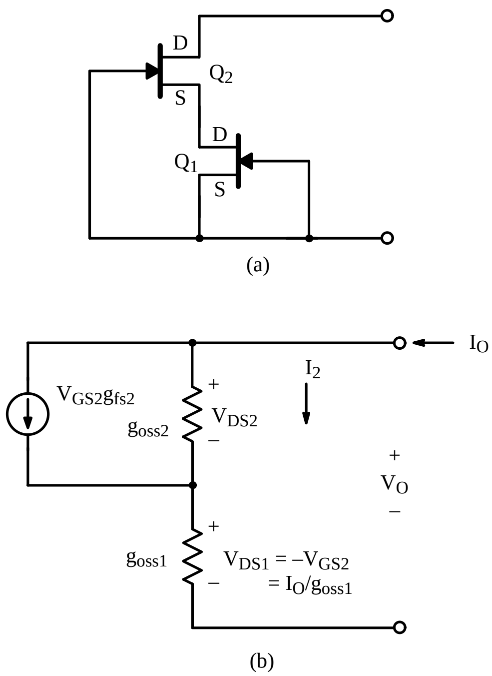


Figure 3. Cascade FET $V_{GS1} = 0$

Now, I_D is regulated by Q_1 and $V_{DS1} = -V_{GS2}$. The dc value of I_D is controlled by R_S and Q_1. However, Q_1 and Q_2 both affect current stability. The circuit output conductance is derived as follows:

If $g_{oss1} = g_{oss2}$ (7)

$$g_o = \frac{g_{oss}}{2 + \frac{g_{fs}}{g_{oss}}} \tag{8}$$

when RS $\neq$ 0 as in Figure 2

$$g_o \approx \frac{g_{oss}{}^2}{g_{fs}(1 + R_S g_{fs})} \tag{9}$$

In either case ($R_S = 0$ or $R_S \neq 0$), the circuit output conductance is considerably lower than the g_{oss} of a single FET.

In designing any cascaded FET current source, both FETs must be operated with adequate drain-gate voltage, V_{DG}. That is,

$V_{DG} > V_{GS(off)}$, preferably $V_{DG} > 2V_{GS(off)}$ (10)

If $V_{DG} < 2\ V_{GS(off)}$, the g_{oss} will be significantly increased, and circuit g_o will deteriorate. For example: A JFET may have a typical $g_{oss} = 4\ \mu S$ at $V_{DS} = 20$ V and $V_{GS} = 0$. At $V_{DS} \sim -V_{GS(off)} = 2$ V, $g_{oss} \sim 100\ \mu S$.

The best FETs for current sources are those having long gates and consequently very low g_{oss}. The Siliconix 2N4340, J202, and SST202 exhibit typical $g_{oss} = 2\ \mu S$ at $V_{DS} = 20$ V. These devices in the circuit of Figure 4 will provide a current source adjustable from 5 µA to 0.8 mA with internal impedance greater than 2 MΩ at 0.2 mA. Other Siliconix part types such as the 2N4392, J112, and SST112 can provide 10 mA or higher current.

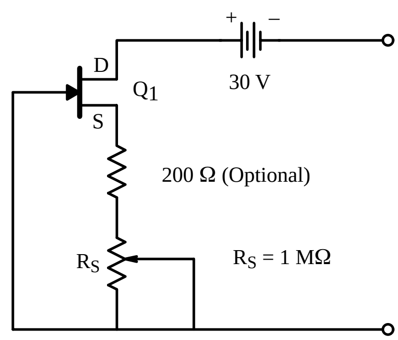


Figure 4. Adjustable Current Source $R_S = 1$ MΩ

Instead of the adjustable resistor, the JFETs can be put in I_{DSS} range groupings with an appropriate R_S resistor selected for each group. This method is common in high volume applications.

The cascade circuit of Figure 5 provides a current adjustable from 2 µA to 0.8 mA with internal resistance greater than 10 MΩ.

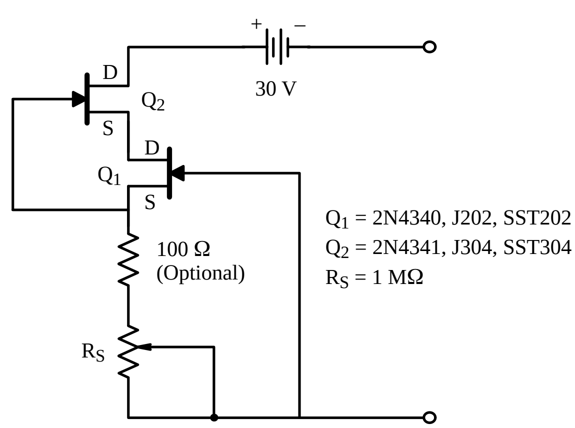


Figure 5. Cascade FET Current Source

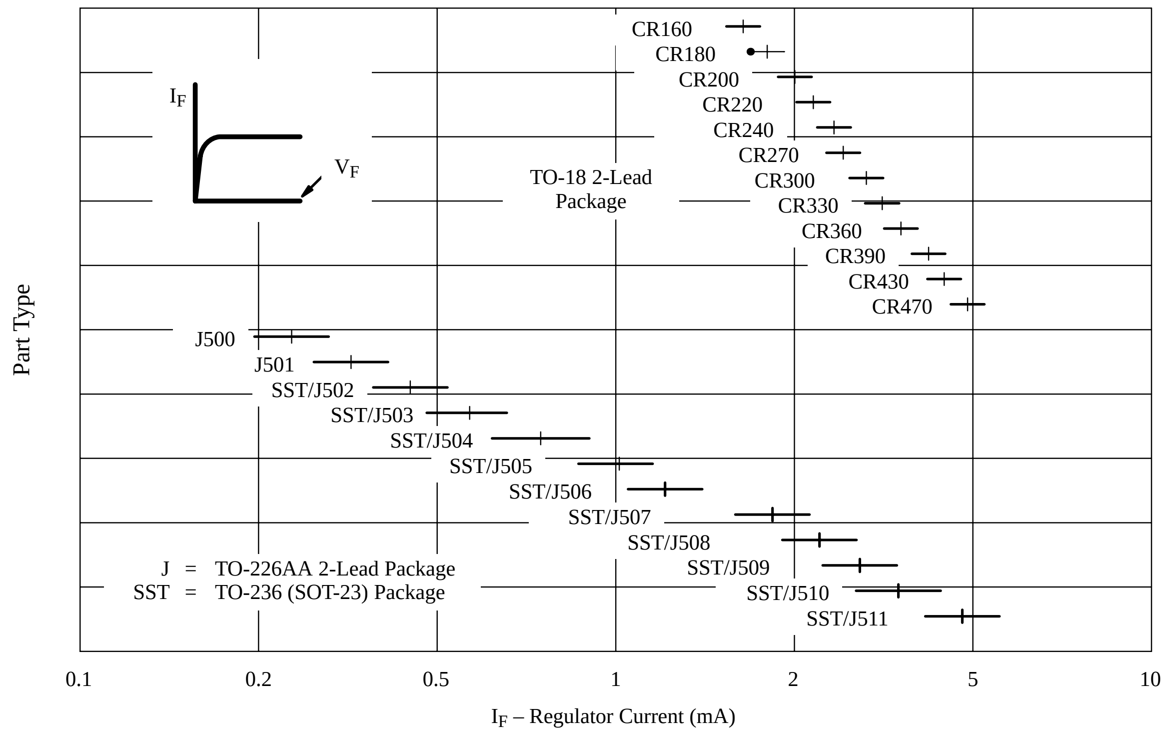


Figure 6. Standard Series Current Regulator Range

Standard Two-Leaded Devices

Siliconix offers a special series of two-leaded JFETs with a resistor fabricated on the device, thus creating a ±10% current range. Devices are available in ranges from 1.6 mA (CR160) to 4.7 mA (CR470).

For designs requiring a ±20% current range, Siliconix offers devices rated from 0.24 mA typical (J500) through 4.7 mA typical (J511) in a two-leaded TO-226A (TO-92) package. The SST502 series is available in surface mount TO-236 (SOT-23).

Each of these two-leaded devices can be used to replace several typical components.

Figure 6 shows the current ranges of these two device series. Further information is contained in the individual data sheets appearing elsewhere in this data book or from Siliconix FaxBack.

The CR160 series features guaranteed peak operating voltage minimum of 100 V with a typical of 180 V. The J500 series features 50 V minimum with a typical of 100 V. The lower current devices in both series provide excellent current regulation down to as little as 1 V.

Bias Resistor Selection

All industry JFET part types exhibit a significant variation in I_{DSS} and $V_{GS(off)}$ on min/max specifications and device-to-device variations.

Using the simple source biasing current source as illustrated in Figure 1, the designer can graphically calculate the R_S which best fits the desired drain current I_D. Figure 7 plotting I_D versus V_{GS} over the military temperature range shows the resulting I_D for different values of R_S.

The R_S lines are constructed by drawing the slope of the R_S desired value starting at the origin, eg. R_S = 2 k slope. Find a convenient point on the X – Y axis to mark a

$\frac{V_{GS}}{I_D}$ of 2 kΩ such as $V_{GS} = –1.5$ V and $I_D = 0.75$ mA.

Then, draw a straight line from this point to the origin. The intersection of this R_S line and the device I_D versus V_{GS} will be the operating I_D. In this example, the resulting $I_D = 0.35$ mA at $T_J = 25°C$. The intercepts of the $T_J = –55°C$ and 125°C show the minimal variation with temperature.

Also note that JFETs have a I_D current where there is no change with temperature variation. To achieve this $0T_C$, the $–V_{GS}$ voltage (I_D x R_S) is approximately:

$$V_{GS(0TC)} \simeq V_{GS(off)} - 0.65\text{ V} \qquad (11)$$

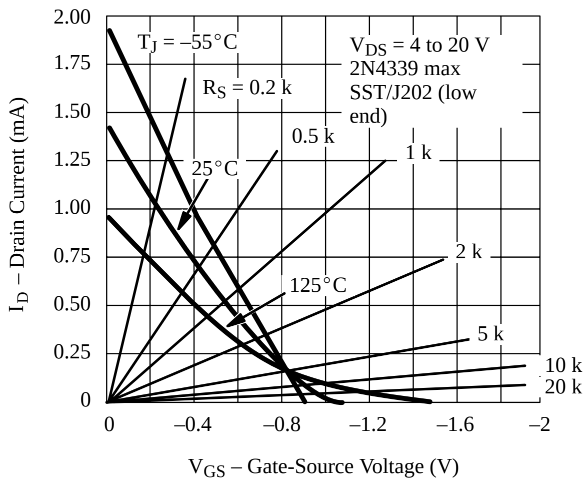


Figure 7. JFET Typical Transfer Characteristic

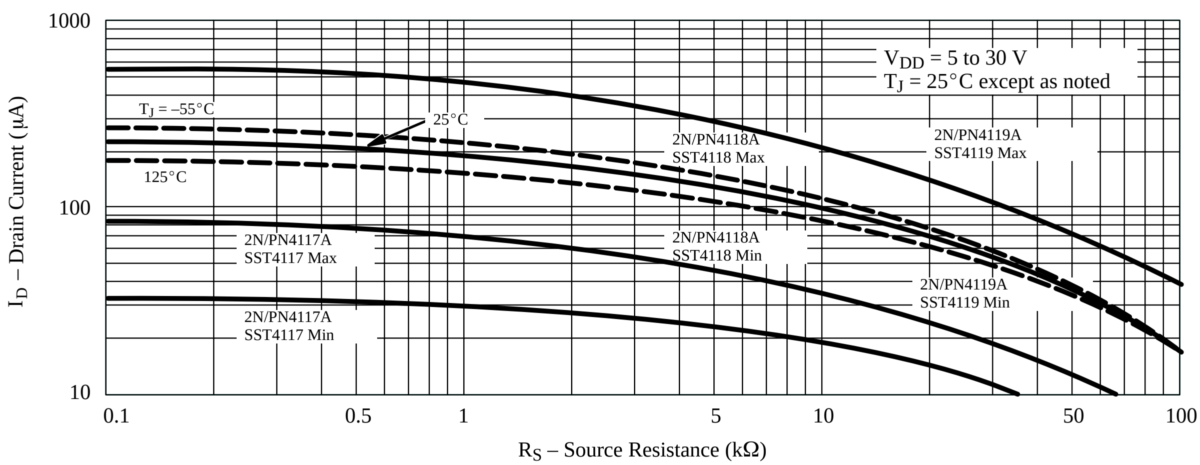


Figure 8. Source Biased Drain-Current vs. Source Resistance

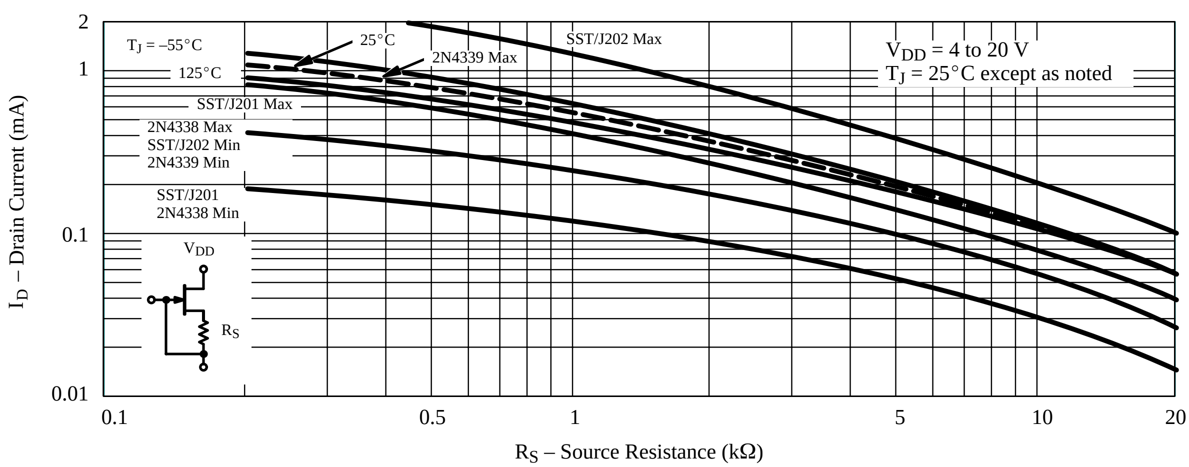


Figure 9. JFET Source Biased Drain-Current vs. Source Resistance

Choosing the Correct JFET for Source Biasing

Each of the Siliconix device data sheets include typical transfer curves that can be used as illustrated in Figure 7.

Several popular devices are ideal for source biased current sources covering a few μAs to 20 mA. To aid the designer, the devices in Table 1 have been plotted to show the drain current, I_D, versus the source resistance, R_S, in Figures 8, 9, and 10. Most plots include the likely worst case I_D variations for a particular R_S. For tighter current control, the JFET production lot can be divided into ranges with an appropriate resistor selection for each range.

Table 1: Source Biasing Device Recommendations

Practical Current Range I_D (mA)	Through-Hole Plastic Device	Surface Mount Device	Metal Can Device
0.01 – 0.02	PN4117A	SST4117	2N4117A
0.01 – 0.04	PN4118A	SST4118	2N4118A
0.02 – 0.1	PN4119A	SST4119	2N4119A
0.01 – 0.1	J201	SST201	2N4338
0.02 – 0.3	J202	SST202	2N4339
0.1 – 2	J113	SST113	2N4393
0.2 – 10	J112	SST112	2N4392

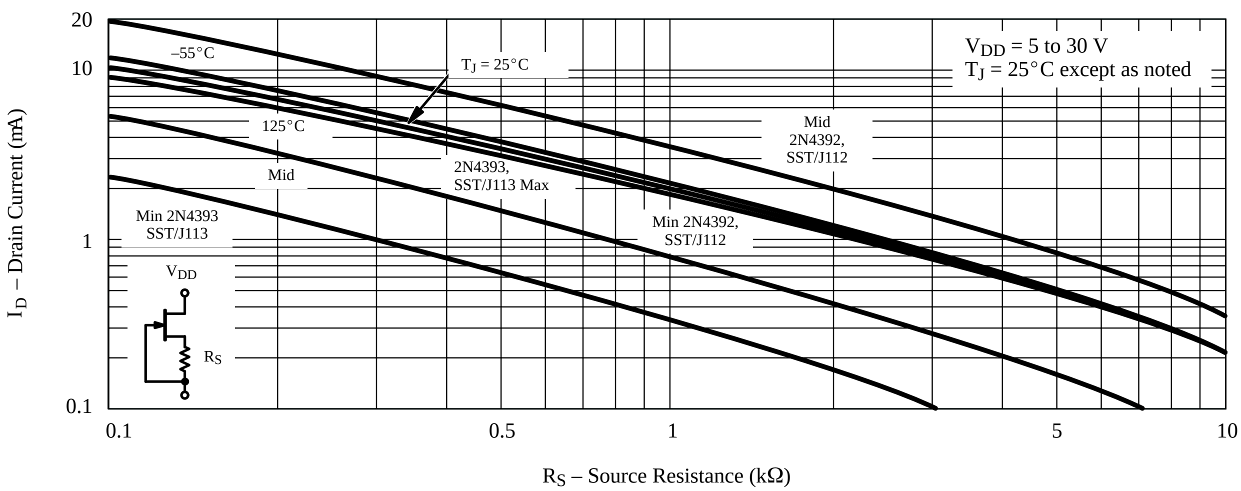


Figure 10. JFET Source Biased Drain-Current vs. Source Resistance

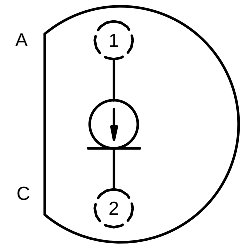
Current Regulator Diodes

J500	J503	J506	J509
J501	J504	J507	J510
J502	J505	J508	J511

PRODUCT SUMMARY

Part Number	Typ I_F (mA)	P_{OV} (V)		Part Number	Typ I_F (mA)	P_{OV} (V)
J500	0.24	50		J506	1.40	50
J501	0.33	50		J507	1.80	50
J502	0.43	50		J508	2.40	50
J503	0.56	50		J509	3.00	50
J504	0.75	50		J510	3.60	50
J505	1.00	50		J511	4.70	50

TO-226AA (TO-92) Modified



Top View

FEATURES

- Two-Lead PlasticPackage
- Guaranteed ±20% Tolerance
- Operation from 1 V (J500–J503) to 50 V
- Excellent Temperature Stability

BENEFITS

- Simple Series Circuitry, No Separate Voltage Source
- Tight Guaranteed Circuit Performance
- Excellent Performance in Low-Voltage/Battery Circuits and High-Voltage Spike Protection
- High Circuit Stability vs. Temperature

APPLICATIONS

- Constant-Current Supply
- Current-Limiting
- Timing Circuits

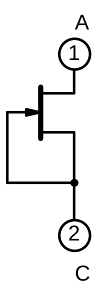
DESCRIPTION

The J500 series is a family of ±20% range current regulators designed for demanding applications in test equipment and instrumentation. These devices utilize the JFET techniques to produce a single two-leaded device which is extremely simple to operate.

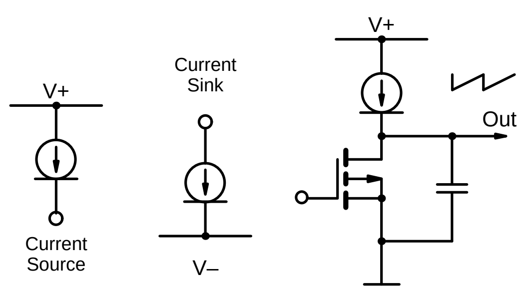
With nominal current ranges from 0.24 mA to 4.7 mA, the J500 series will meet a wide array of design requirements.

The low-cost TO-226A package ensures a cost-effective design solution.

SCHEMATIC DIAGRAM



APPLICATIONS



Linear Ramp Generator

Applications information may be obtained via FaxBack, request document #70596.

ABSOLUTE MAXIMUM RATINGS

Peak Operating Voltage . . . . 50 V

Reverse Current . . . . 50 mA

Storage Temperature . . . . –55 to 150°C

Power Dissipation[a] . . . . 350 mW

Notes:
a. Derate 2.8 mW/°C above 25°C

SPECIFICATIONS[a]

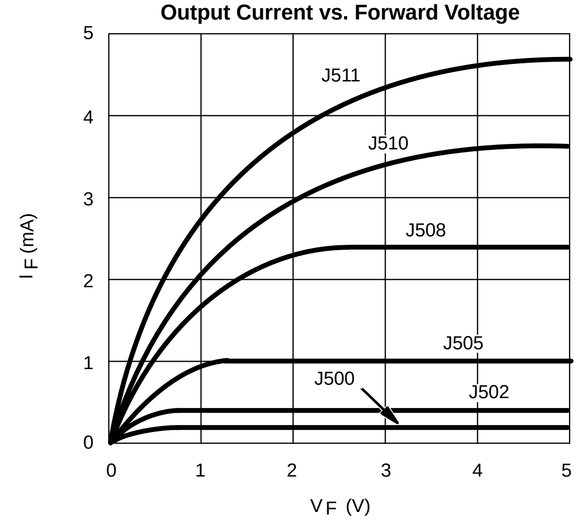
Parameter	Symbol	Test Conditions	Limits Min	Limits Typ NO TAG	Limits Max	Unit
Peak Operating Voltage	P_{OV}	$I_F = 1.1\ I_{F(max)}$ NO TAG	50	95		V
Reverse Voltage	V_R	$I_R = 1$ mA		0.8		
Capacitance	C_F	$V_F = 25$ V, f = 1 MHz		2.2		pF

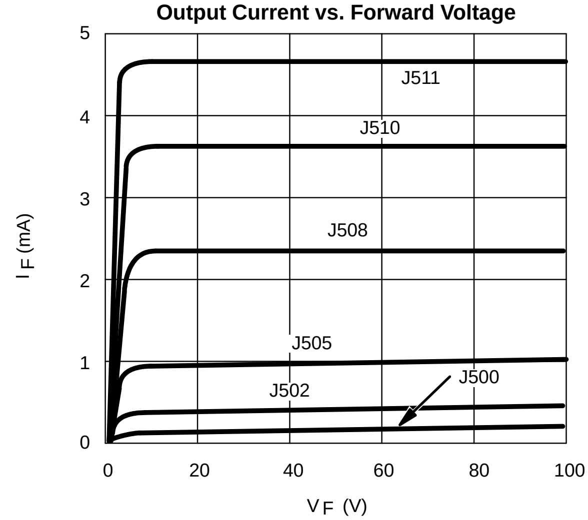
	Regulator Current[d] (I_F)			Dynamic Impedance[e] (Z_d)		Knee Impedance (Z_k)	Limiting Voltage[f] (V_L)		Temperature Coefficient (θ_1)
	$V_F = 25$ V			$V_F = 25$ V		$V_F = 6$ V	$I_F = 0.8\ I_{F(min)}$		$V_F = 25$ V $0°C \le T_A \le 100°C$
	mA			MΩ		MΩ	V		%/°C
Part Number	Min	Nom	Max	Min	Typ[b]	Typ[b]	Max	Typ[b]	Typ[b]
J500	0.192	0.24	0.288	4.00	15	2.50	1.2	0.4	0.95%
J501	0.264	0.33	0.396	2.20	10	1.60	1.3	0.5	0.81%
J502	0.344	0.43	0.516	1.50	7	1.10	1.5	0.6	0.70%
J503	0.448	0.56	0.672	1.20	5	0.80	1.7	0.7	0.58%
J504	0.600	0.75	0.900	0.80	3.5	0.55	1.9	0.8	0.46%
J505	0.800	1.00	1.200	0.50	2	0.40	2.1	0.9	0.33%
J506	1.120	1.40	1.680	0.33	1.5	0.25	2.5	1.1	0.19%
J507	1.440	1.80	2.160	0.20	1	0.19	2.8	1.3	0.08%
J508	1.900	2.40	2.900	0.20	0.7	0.13	3.1	1.5	–0.05%
J509	2.400	3.00	3.600	0.15	0.5	0.09	3.5	1.7	–0.14%
J510	2.900	3.60	4.300	0.15	0.4	0.07	3.9	1.9	–0.22%
J511	3.800	4.70	5.600	0.12	0.3	0.05	4.2	2.1	–0.34%

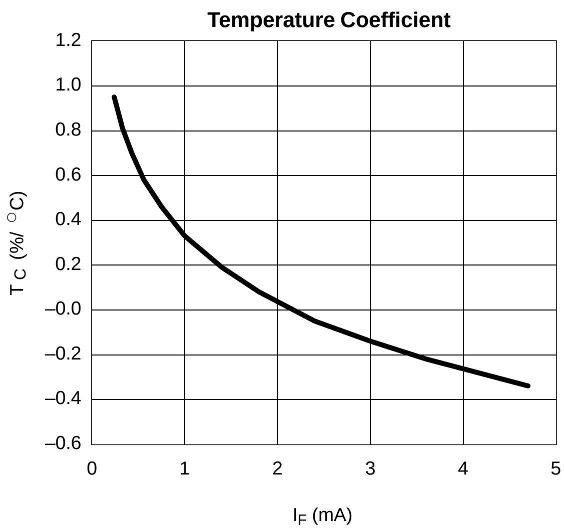
Notes:
a. $T_A = 25°C$ unless otherwise noted.
b. Typical values are for DESIGN AID ONLY, not guaranteed nor subject to production testing.
c. Max V_F where $I_F = 1.1\ I_{F(max)}$ is guaranteed.
d. Pulse test—steady state currents may vary.
e. Pulse test—steady state impedances may vary.
f. Min V_F required to insure $I_F = 0.8\ I_{F(min)}$.

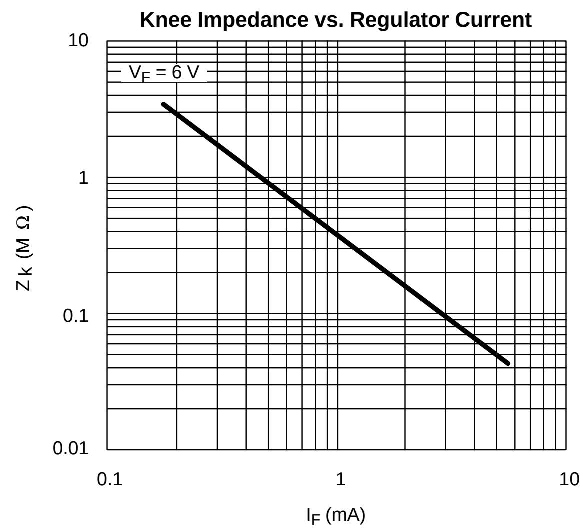
NCL

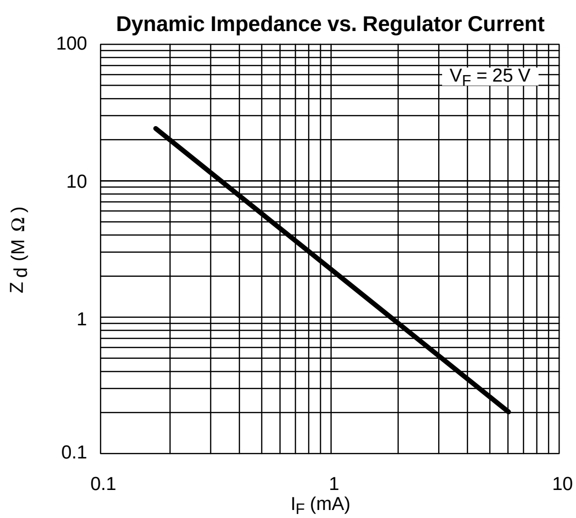
TYPICAL CHARACTERISTICS

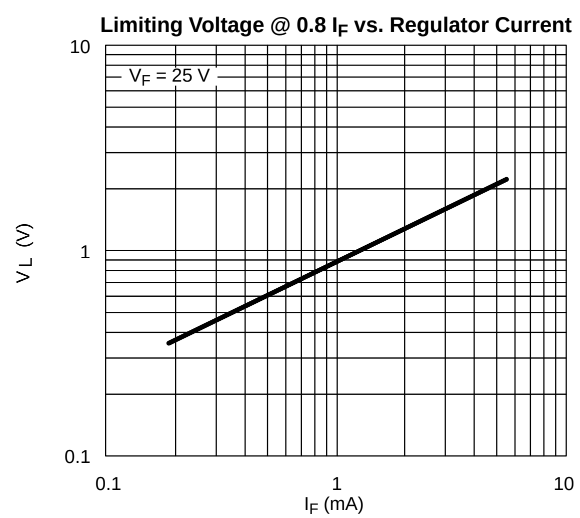




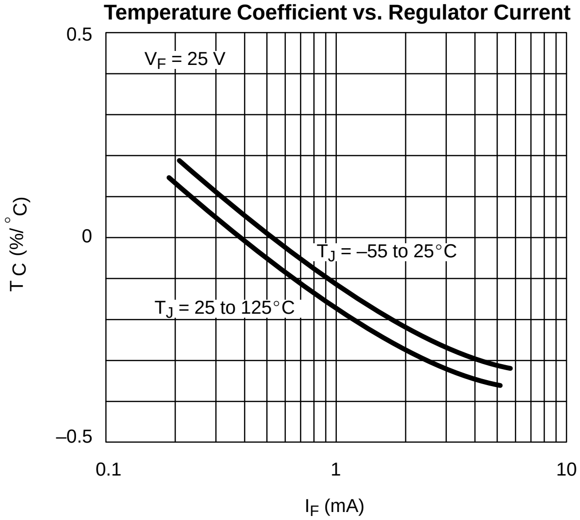


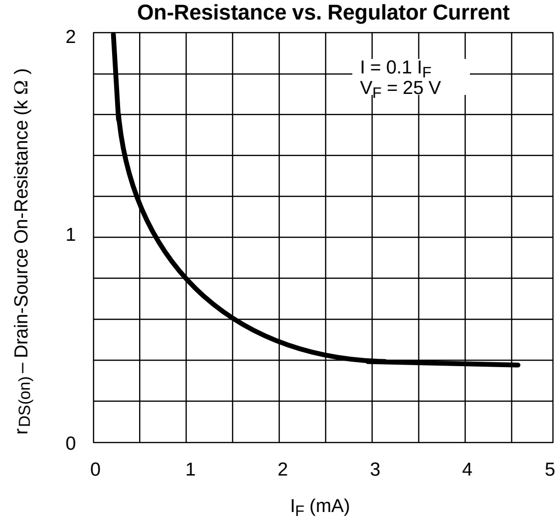


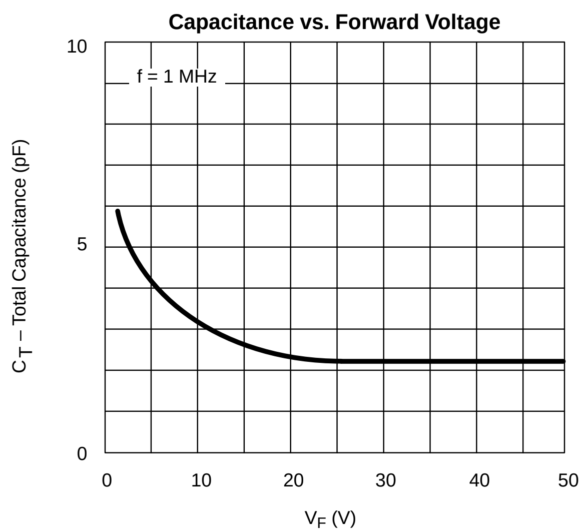




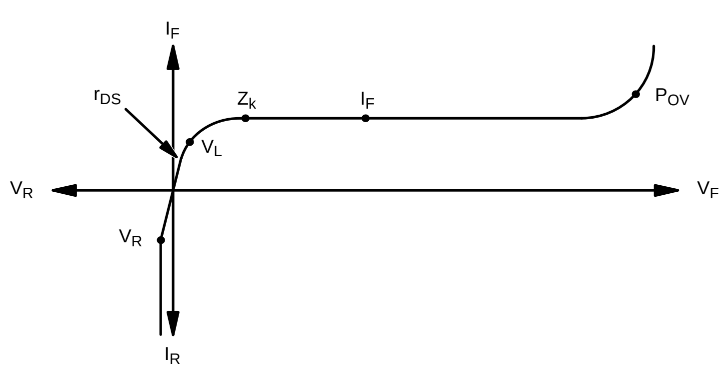
TYPICAL CHARACTERISTICS







CURRENT REGULATOR DIODE V-1 CHARACTERISTIC



Getting the Most from IC Voltage References

A brief guide for users

by Walt Jung

As resolution and accuracy requirements of modern systems rise to 12 bits and beyond, the selection, specification, and application of voltage references becomes a key factor in system design. This article, devoted to designing with IC references, starts with the basic features of a good reference, discusses reference performance parameters, and concludes with examples of IC reference applications in high-performance circuits.

REFERENCE BASICS

Figure 1 depicts an ideal 1-volt reference source. This source is ideal in the sense that the 1.000000-V output is independent of time, temperature, and other environmental factors. Furthermore, neither connection polarity nor loading affects the voltage it delivers to the load. In this ideal world, connection of load, R_L, with either polarity, produces a constant voltage at the load, equal in magnitude to the original source, V_R.

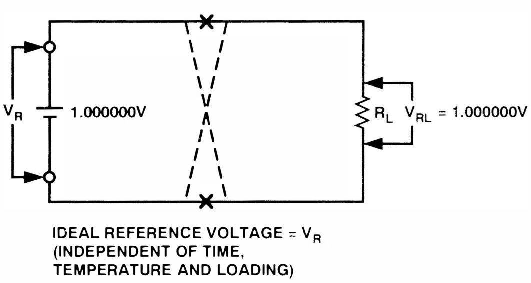


Figure 1. Ideal reference-voltage source.

In the real world, all the factors assumed ideal can and will vary. Major sources of error in reference voltage include initial calibration tolerance, output voltage drift with temperature and time, loading effects (characterized by output and wiring impedances), and noise components (both intrinsic and supply-related).

A variety of ways exist to produce a relatively stable voltage, including chemical, solar and low-temperature quantum devices. As a starting point, we will restrict our discussions to reference sources derived from system power, and in particular to reference circuits powered from positive or negative 3-30-volt dc supplies.

Most commonly, standard reference ICs are available in *three-terminal* form (V_{IN}, Common, V_{OUT}), with positive polarity. *Two-terminal* (diode-like) references, while more flexible regarding polarity, are restrictive as to loading. The constraints often complicate reference designs, making choices difficult (but inviting ingenuity).

Some basic two-terminal references are shown in Figure 2. In (a), a current-driven forward-biased diode (or diode-connected transistor) produces a voltage, V_F, approximately proportional to the logarithm of current and hence relatively insensitive to small changes in current. While its junction drop is somewhat independent of the raw voltage supply, it has numerous deficiencies as a reference. Among them are a significant temperature coefficient (TCV) of about −0.3%/°C, some sensitivity to loading, and a rather

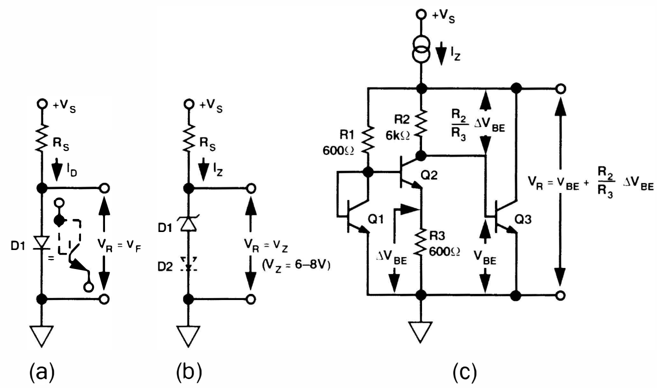


Figure 2. Basic semiconductor reference circuits. a. Simple forward-biased diode. b. Zener (avalanche) diode reference. c. Basic bandgap reference.

limited output voltage, about 600 mV; multiples of this voltage can be obtained by using series-connected junctions.

By contrast, this simple reference (as well as other shunt-type two-terminal regulators) has a basic advantage—that polarity can be readily inverted by reversing connections and drive current. However, a basic limitation of all shunt regulators is that driving current, I_D, does not decrease substantially when load current decreases.

In (b), by driving an appropriately selected reverse-breakdown diode at a current exceeding a threshold, an appreciably higher reference voltage can be realized. Although such reference diodes are almost universally referred to as "Zener" diodes, true Zener breakdown occurs below 5 V, while avalanche breakdown occurs at higher potentials.[1] With D1 chosen to have breakdown voltage in the 5-to-8-V range, its net positive TC adds to the negative TC of a series forward diode (D2), yielding a net TC of 100 ppm/°C or less with proper current bias.[2] In the past, carefully chosen diodes were combined to form single-package "zero-TC Zener references", such as the 1N821-1N829 series. Designs based on this concept were the basis for early hybrid IC references, which are in fact still sold today (more below).

While low TC can be realized in 2(b), the circuit has limitations on direct use of its output: First, the choice of voltage with high-accuracy diodes is limited, because the best TC combinations occur at specific voltages, such as the 1N829's 6.2 V. In addition, the range of load currents is limited, since the diode current must be carefully controlled for best TC. And, unlike a fundamentally low-voltage (<2-V) reference such as 2(a), Zener-diode-based references must of necessity be driven from voltage sources appreciably greater than 6 V, so this precludes their operation from 5-V system supplies. References based on old-style low-TC avalanche diodes tend to be noisy, an inherent property of the surface-breakdown mechanism. This noise is lower with monolithic buried-Zener types (more below).

The development of low-voltage (<5-V) reference circuits based on the bandgap* voltage of silicon led to the introduction of ICs that could provide good TC performance operating on low voltage supplies.[3] A bandgap reference develops an internal voltage proportional to absolute temperature (PTAT) to null out the temperature variation of a junction voltage, which has a negative TC (complementary to absolute temperature—CTAT). A basic

*The bandgap is the energy difference between the bottom of the conduction band and the top of the valence band. For references using silicon transistors the corresponding voltage, extrapolated to T = 0 K, is about 1.21 V—but is dependent on process and detailed curvature-compensation circuitry.

bandgap-based reference cell, driven by a constant current, is shown in Figure 2 (c). This circuit is also called a "ΔV_{BE}" reference, because of the correction voltage across R2. This voltage, based on the V_{BE} difference produced by differing current densities between matched transistors, Q1-Q2, is developed by a current resulting from ΔV_{BE} across R3 and transduced to voltage by R_2. It is summed with the V_{BE} of Q3 to produce V_R.

The bandgap technique is attractive in low-voltage IC designs because it is relatively simple and avoids noisy Zeners. It is used both for stand-alone IC references and as an internal reference within linear ICs. Buffered forms of 2-terminal 1.2-V reference ICs employing the bandgap concept provide additional current gain for stable, accurate operation over wide current ranges. Among them is the AD589, a synthesized 1.235-V "diode" with a 0.6-Ω dynamic impedance, a 50 µA to 5 mA operating current range, and TC grades ranging from 10 to 100 ppm/°C.

The basic designs shown in Figure 2 are sensitive to loading and require stable current drive. They generally need scaling of the output to more-useful levels, e.g., 2.5 V, 5 V, etc. For most applications, a buffer amplifier is used; besides driving loads, it provides voltage scaling to more useful levels.

An improved bandgap circuit (Figure 3), the "Brokaw cell", addresses these issues.[4, 5] This circuit is used in the AD580, the first precision bandgap-based IC reference. Still in production after 20 years, it is the first of a family of reference devices, such as the AD581 and AD584; the circuit also provides the internal reference in many Analog Devices ADCs and DACs.

At the heart of the AD580 are two transistors, Q2 and Q1, with equal collector currents and 8:1-scaled emitter areas (resulting in a 1:8 current-density ratio). The currents are maintained equal by matched load resistors and overall feedback voltage from the output amplifier (which also provides buffering), applied to the transistor bases. In this closed loop, the difference in the V_{BE}s (i.e., ΔV_{BE}) appears across R2, and a current equal to $2\Delta V_{BE}/R_2$ flows through R1, producing a PTAT voltage, V_1:

$$V_1 = 2\frac{R_1}{R_2}\Delta V_{BE}$$

V_1 appears in series with V_{BE}, thus a constant voltage, V_Z (about 1.205 V)—appropriately compensated for the variation of V_{BE} with temperature—appears between the bases and common.

The feedback attenuator, R4 and R5 (laser trimmed) permits the actual voltage appearing at V_{OUT} to be scaled higher, 2.5 V in the case of the AD580. In principle, this voltage can be raised to any practical level; for example, the selectable AD584 provides taps for 2.5, 5, 7.5, and 10-V operation.[6]

In practical applications, the amplifier is an invaluable feature. Besides its central role in optimizing the basic bandgap cell's performance, it also provides scaling and low output impedance. The AD580, operating from supplies of 4.5-30 V, outputs 2.5 V at up to 10 mA, a useful feature for a variety of circuits.[7] It is available in tolerances as low as 10 mV, with TCs as low as 10 ppm/°C (Table 1).

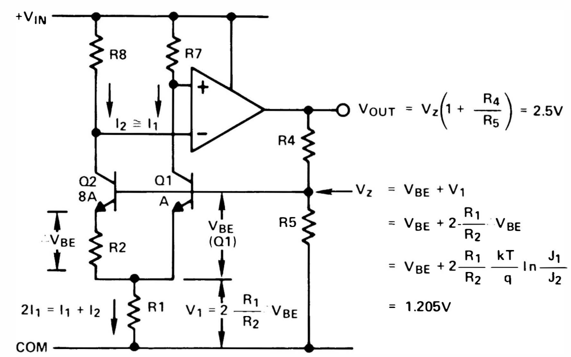


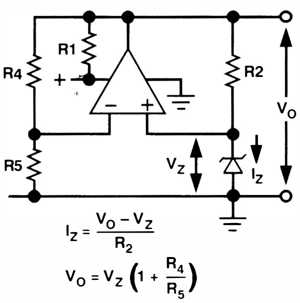
Figure 3. Functional diagram, AD580 precision bandgap reference

Table 1. Fixed Positive Output Three Terminal Monolithic IC References

Device	Type[1] (B, Z)	V_{OUT} (V)	Tolerance (+mV max)	Drift (ppm/°C, max)	$+V_S$ (V)	Load Sensitivity[2]	Sensitivity Line (µV/V, max)	Noise[12]	R_{TRIM} (Ω) ±Range (V)	I_Q (mA, typ)	PTAT Output (mV/°C)	Comment
AD780[8, 9, 11]	B	2.5	1–5	3–20	V_{OUT}+1.5 to 36	50	10	4 µV, 100	25k, 0.1	0.75	1.9	Precision 2.5 V
REF-43	B	2.5	15–50	10–25	4.5–40	50	5	220 (1 kHz)	10k, 0.095	0.45	1.9	Precision 2.5 V
AD680	B	2.5	5–10	20–30	4.5–36	100	40	250	NA	0.2	2.0	Low I_Q
REF-03	B	2.5	15	50	4.5–33	250	125	6 µV	10k, 0.15	1.0	2.1	Standard 2.5 V
AD580	B	2.5	10–75	10–85	4.5–30	1000	1–6 mV[7]	8 µV	NA	1.0	NA	3 Pin TO-52
AD1403	B	2.5	10–25	25–40	4.5–40	1000	3–4.5 mV[7]	8 µV	NA	1.2	NA	3 Pin Mini-DIP
AD586[11]	Z	5	2–20	2–25[4]	10.8–36	100–150	100	4 µV, 100	10k, +6%,–2%	2.0	NA	Precision 5 V
REF-195	B	5	2–10	5–10	5.1–15	20–40[13]	20–40[14]	50 µV	NA	30 µA[15]	NA	Note 10
REF-05	B	5	15–25	8.5–25[3]	8–33	500	500	10 µV	10k, 0.3	1.0	2.1	Note 5
REF-02	B	5	15–100	8.5–250	8–33	500–2500	500–2500	10 µV	10k, 0.3	1.0	2.1	Standard 5 V
AD587[11]	Z	10	5–10	5–20[4]	13.5–36	100	100	4 µV, 100	10k, +3%,–1%	2.0	NA	Precision 10 V
AD581[8]	B	10	5–30	5–30	13–30	500	200	40 µV	NA	0.75	NA	3 Pin TO-5
REF-10	B	10	30–50	8.5–25[3]	13–33	800–1000	1000	20 µV	10k, 0.3	1.0	NA	Note 6
REF-01	B	10	30–100	8.5–65	13–33	800–1000	1000–1500	20 µV	10k, 0.3	1.0	NA	Standard 10 V

NOTES
NA = not applicable for device in question.
[1]B = Bandgap, Z = Buried Zener.
[2]µV/mA, max, I_L = 0-10 mA, Sourcing.
[3]Long term stability 100 ppm (max.) per 1khours.
[4]Long term stability 15 ppm (typ.) per 1khours.
[5]Similar to REF-02 with long term drift specified.
[6]Similar to REF-01 with long term drift specified.
[7]Total over applicable supply range.
[8]Operates in two-terminal mode.
[9]2.5 V & 3 V output modes.
[10] Low I_Q, low dropout, shutdown pin.
[11]Optional noise reduction feature
[12]Typical, µV p-p, 0.1 to 10 Hz or nV/$\sqrt{\text{Hz}}$ at 100 Hz.
[13]I_L = 0–30 mA, $+V_S$ = 6.3–15 V.
[14]$+V_S$ = 5.1–15 V.
[15]5 µA standby.

Zener-based references also benefit from careful buffering; and overall accuracy and stability are improved by including the Zener in the buffer circuit's feedback loop. Figure 4 depicts the basic circuit architecture of the hybrid IC AD27xx series. [8, 9] These devices have long provided stable +10-V, −10-V and ±10-V sources with very tight tolerances and TCs, as low as ±1 mV and 1-2 ppm/°C, performance only recently achieved by monolithic devices.



DEVICE	OUTPUT CHARACTERISTICS
AD2700	+10V (±2.5-5mV) 3-10ppm/°C
AD2710	+10V (±1mV) 1-2ppm/°C
AD2701	−10V (±2.5-5mV) 3-10ppm/°C
AD2702	±10V (±2.5-5mV) 3-10ppm/°C
AD2712	±10V (±1mV) 1-2ppm/°C

Figure 4. Precision Zener buffering architecture with gain provides regulated diode current for optimum TC.

Circuits of this type can feed back the optimum value of regulated current from the output [$(V_O - V_Z)/R_2$] to operate a temperature-compensated diode at the op amp's + input. The Zener voltage is amplified using a pair of resistances whose ratio is laser-trimmed for accurate output voltage, [$(1 + R_4/R_5)V_Z$]. The op amp can drive load currents up to 10 mA with a typical output impedance of 50 µV/mA, and the inclusion of the diode in the feedback loop makes the device relatively insensitive to line-related errors, typically 125 µV/V for the AD2710. To further reduce the low calibration errors, a pair of fine-trim terminals is provided. The +10-V- output devices operate as shown, while the ±10-V AD2712 adds a precision inverter for the negative output.

IC REFERENCE SPECIFICATIONS

Monolithic IC references come in a variety of functional styles, dominated by three-terminal types with fixed positive output(s). The choice of bandgap or Zener technology determines the class of ultimate specifications and performance.

Figure 5 shows the standard basic pinout (input-2, output-6, ground-4) for + 2.5-, +5-, and +10-volt IC references in 8-pin cans and DIPs. Additional pins may be used for important housekeeping details, such as optional trimming (e.g., R_{TRIM} at pin 5) or providing a PTAT kelvin-scale thermometer output—an inherent bonus feature in bandgap devices ("V_{TEMP}" at pin 3). In general, all references should use an adjacent RF-quality input bypass capacitor, C_1, sometimes paralleled with larger C_2 for increased capacitance, to handle noisy sources and rapidly varying heavy loads. Some references may also allow (or require) an output bypass, C_{OUT} —or a noise-reduction capacitor connected to an internal point. Layouts should use a short, heavy (+) output conductor to minimize IR drops, while the (−) lead is less critical in this configuration, typically carrying ≤1 mA.

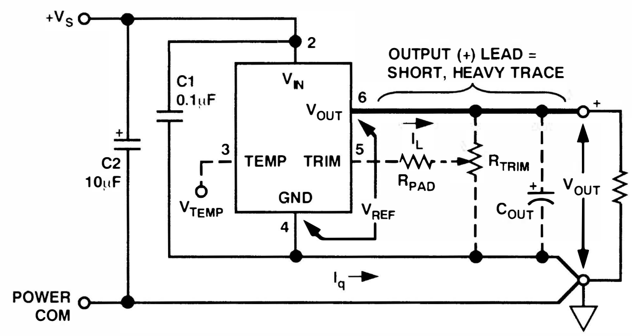


Figure 5. Standard positive-output three-terminal reference hookup (8 pin DIP pinout).

Table 1 is a summary of fixed-voltage three-terminal bandgap (B) and "buried"-Zener-based (Z) positive references. It compares the key device specs of output tolerance, drift, operating input supply range, sensitivity to load current (output impedance) and line voltage, noise, quiescent current, as well as trim range and PTAT thermometer availability. Comments and notes interpret and supplement the numerical specs. Table 2 compares references having selectable output voltage.

Table 2. Selectable Output Monolithic IC References

Device	Type[1] (B, Z)	V_{OUT} (V)	Tolerance (± mv max)	Drift (ppm/°C, max)	$+V_S$ (V)	Load Sensitivity (µV/mA, max, I_L=0–10 mA, Sourcing)	Sensitivity Line (µV/V, max)	Noise (typ, nV/√Hz or µV p-p 0.1-10 Hz)	Trim Range (±mV)	I_Q (mA, typ)	Tracking in Bipolar Mode (± mV, max)	Comment
AD584[8, 9]	B	2.5, 5, 7.5, 10	2.5–5, 7.5–30	5–30[5]	V_{OUT}+2.5 to 30 V	50[7]	Note 10	50 µV	Note 11	0.75	NA	Multiple positive output.
AD588[9]	Z	±5, +5/+10, −5/−10	1–5	1.5–6[3]	±13.5 to ±18	50	±200[6]	100	4	6	0.75	Precision programmable low noise + TC, Kelvin sensing buffer amplifiers.
AD688[9]	Z	±10	2–5	1.5–6[4]	±13.5 to ±18	40	±200[6]	140	5	9	1.5–3	Precision ±10 V, low noise + TC, Kelvin sensing buffer amplifiers.
REF08	Z	−10, −10.24	30–40, 40–60	50–100	−11.4 to −36	250	500	140	270	1.1	NA	Dual negative output.

NOTES
[1]B = Bandgap, Z = Buried Zener.
[2]NA = not applicable for device in question.
[3]Long term stability (ppm) 15 (typ), 25 (max) per 1khours.
[4]Long term stability (ppm) 15 (typ) per 1khours.
[5]Long term stability (pm) 25 (typ) per 1khours.
[6]T_{MIN} to t_{MAX}
[7]ppm/mA, 0-5 mA.
[8]Operates in two-terminal mode, 5 + 10 V.
[9]Optional noise reduction feature.
[10]0.002%/V, 15 to 30 V; 0.005%/V, V_{OUT} + 2.5 to 15 V
[11]Determined by user resistances.

All Analog Devices monolithic IC Zener references employ a subsurface breakdown technology, providing a salient improvement over the noise, drift, and reliability of surface-mode operated devices.[10] It was first applied in 1974, within the AD534 analog multiplier, [11] and later in DACs and other conversion products. The first stand-alone buried-Zener reference was the multiple-output AD588, a ±5-V, +10-V, −10-V precision unit, [12, 13] followed by the three-terminal +5-V AD586 and +10-V AD587, [14] and the negative-output REF08. Buried-Zener references offer the lowest drift, down to the 1-2-ppm/°C range (AD588 and AD586), and the lowest noise as a % of nominal output, 100 nV/√Hz or less at 5 or 10 V (AD586, -587, -588). The multiple-output AD588 and AD688 (±10 V) are listed in Table 2.

Tolerance: By choosing a unit specified for the required accuracy when possible, the user can avoid trimming (or gain scaling). This results in the best TC performance, since tight tolerances and low TCs usually go hand-in-hand. Tolerances as low as 0.04% can be achieved with the AD586, AD780, and REF195, while the AD588 goes as low as 0.01%. If trimming must be used, be sure to use the specific circuit recommended on the device data sheet, with no more range than necessary. For scaling beyond the recommended range, use a precision op amp and accurate-ratio, low-TC tracking thin film resistors.

Drift: The lowest-drift (long-term and temperature-related) references are monolithic buried-Zener and hybrid types using temperature-compensated Zeners. Maximum TCs as low as 1 ppm/°C are available with the AD2710 hybrids, and 1.5 ppm/°C with the AD588 and AD688. Close behind is the AD586, at 2 ppm/°C; and the best bandgap is the AD780, at 3 ppm/°C. Lowest maximum long term drift is 25 ppm/1000 hr, in the AD588.

Temperature drift can affect full-scale accuracy in systems using A/D and D/A converters, as indicated by Table 3. This table shows system resolution in bits (column 1), required drift rate for 1/2-LSB drift over a 100°C change (column 2), and the voltages corresponding to 1/2 LSB for this 100°C example for three reference voltages. Drift of <1.2 ppm/°C is required to maintain 1/2 LSB error at 12 bits, but lesser temperature spans will require less-stringent drifts.

The temperature drift of references is seldom monotonic; there may be several reversals over the rated temperature span. Modern practice is to measure output at several temperatures, so as to guarantee a maximum error band applicable to the temperature range. The rated drift in ppm/°C is defined as the slope of a diagonal drawn between opposite corners of a box that bounds the applicable temperature range and the allowable maximum change. For example, the 5 ppm/°C, 5-volt AD586L (25 μV/°C) has an allowable change of 1.75 mV over a 70°C range (this technique is discussed in greater detail on the AD586 data sheet).

Table 3. Reference Temperature Drift Requirements for Various System Accuracies (1/2 LSB Criteria, 100°C Span)

Bits	Required Drift, (ppm/°C)	1/2 LSB Weight (mV), Various FS Ranges 10 V	5 V	2.5 V
8	19.53	19.53	9.77	4.88
9	9.77	9.77	4.88	2.44
10	4.88	4.88	2.44	1.22
11	2.44	2.44	1.22	0.61
12	1.22	1.22	0.61	0.31
13	0.61	0.61	0.31	0.15
14	0.31	0.31	0.15	0.08
15	0.15	0.15	0.08	0.04
16	0.08	0.08	0.04	0.02

Supply Range: Reference ICs generally require a supply range from about 3 V above rated output, to 30 V or more, except for devices designed for low dropout, such as the REF195 and the AD780. At low currents, the REF195 can maintain 5-V output with input voltage as low as 5.1 V (0.1-V *dropout*).

Load Sensitivity: Load sensitivity, or output impedance, is usually specified in μV/mA of load current (or mΩ), for output source currents of 0-10 mA. A reasonable value at low frequencies is 100 mΩ or less (AD780, REF43, REF195), but without due care, external wiring drops can add a comparable amount of series impedance, producing additional error (see Figure 5). Errors depending on load current are minimized with short, heavy conductors on the (+) output, and return wires to reference and power common from the low end of the load. For the highest precision, buffer amplifiers and separate force-sense (or Kelvin) connections—like those provided in the AD588 & AD688—can guarantee a precise voltage at the point-of-loading. Several of the applications illustrate Kelvin sensing.

Figure 6 is a plot of *dynamic* output impedance as a function of frequency for three 2.5-V references. These data were collected with a high resolution test setup using an Audio Precision System One, with software adapted from "IMPD"*, modified for 4-terminal high-resolution bandpass-mode operation. Input to the device under test is +15 V dc, and the test signal is 0.83-mA rms swept at 20-200-kHz, superimposed on a dc load of 2 mA.

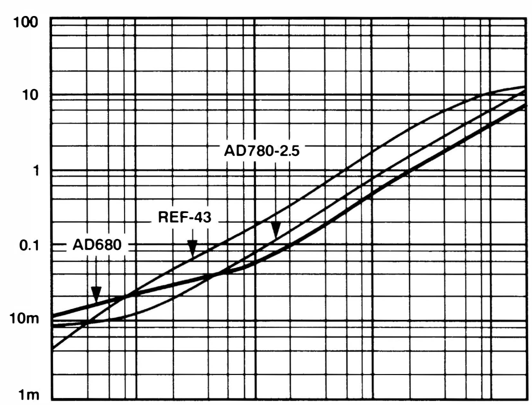


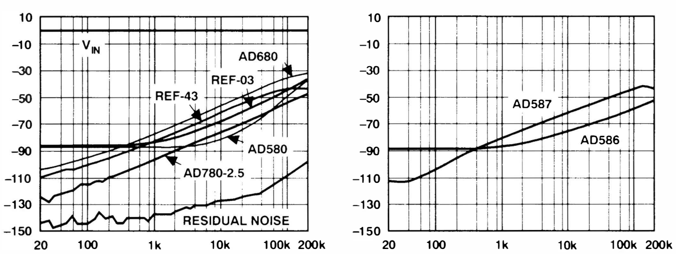
Figure 6. AD680, AD780, REF43; output impedance (ohms) vs. frequency (Hz); (V_{IN} (DC) = 15 V, I_I (DC) = 2 mA, I_I (AC) = 0.83 mA rms)

The plot compares their output impedance as a function of frequency. The characteristic 6-dB/octave rise above 100 Hz is essentially inductive (8-25 μH), while at low frequencies the impedance approaches or reaches a constant resistance in the vicinity of 10 mΩ for these devices. Some devices allow additional output load capacitance, which can be employed to further decrease output impedance at higher frequencies.

*Debi Brimacombe, "Generating Impedance vs. Frequency Plots With System One", AUDIO.TST, November 1992.

Line Sensitivity: Line sensitivity, the ratio of output change to a change of input, is less than 50 μV/V (−86 dB) in the REF43, REF195, AD680 and AD780. For dc and very low frequencies, such errors are easily masked by noise.

Plots of line rejection vs. frequency show susceptibility of a device to wideband noise on the input line (Figure 7). Data were collected with a high-resolution, screened and guarded test setup employing an Audio Precision System One analyzer operating in a bandpass-filtered crosstalk mode, for a dynamic range in excess of 130 dB. The device input is +15 V dc, and the output load is 1 mA. The test signal, superimposed on the input, is at 1 V rms, swept from 20 Hz to 200 kHz. For these plots, the V_{IN} 0-dB reference is 1 Vrms, and test-circuit residual noise below 1 kHz is ≅ −140 dB. Because of the bandpass nature of these measurements, in some instances they may not directly compare to results using wideband methods (which tend to become noise-limited at −90 to −100-dB levels).



(a) (b)

Figure 7. Various references, line rejection (dB) vs. frequency (Hz); (V_{IN} (DC) = 15 V, I_I (DC) = 2 mA, V_{IN} (AC) = 1 V rms). a. 2.5-V bandgap types. b. Buried Zener types.

Figure 7a shows line-rejection vs. frequency for a group of 2.5-V bandgaps. The recent AD680, AD780 and REF43 show rejection of 100 to 130 dB at low frequencies, decreasing to 40-55 dB at 100 kHz. The AD580 and REF03 have more-limited rejection (≅90 dB) below 1 kHz, but their behavior is comparable at the higher frequencies.

Line rejection for two 5- and 10-V buried-Zener devices (b) was measured in standard operation (no "noise reduction" used). The AD586 has greater rejection at high frequency; the AD587 at low frequency.

For references requiring greater line rejection, simple input filtering can be effective. A 100-Ω decoupling resistance with a 1-μF bypass filters frequencies above 1.6 kHz; the input DC headroom suffers only 0.2 V increase for 2-mA loading. This step is a wise precaution when references must derive their power from switch-mode power supplies. Additional output capacitance (where allowable) can also be helpful. Alternatively, line rejection can be increased with a preregulator, either a 78Lxx type regulator or stacked reference. [7]

Noise: Not all manufacturers specify the noise generated within a reference; when it is specified, there is little uniformity on how to measure it and present the data. For example, some devices are characterized for peak-to-peak noise in a 0.1-10-Hz bandwidth, others are specified in terms of rms for a specified bandwidth, and yet others in noise spectral density (nV/√Hz rms) at a given frequency. The most useful characterization would be a plot of noise spectral density over a range of frequencies, since it can be used for calculating any of the other specifications. Any noise fed through from the supply due to line sensitivity must be added (root-sum-of-squares) to the noise generated by the device.

Noise is an important characteristic in references because it limits accuracy and introduces uncertainty in high-resolution, wide-bandwidth systems. A noisy reference source used in a conversion system can result in reduced resolution. For low-frequency measurement systems, peak-to-peak specifications in the time domain are useful, because noise adds to the uncertainty of each unique data point. In higher-frequency systems, rms values for noise are more useful, because information usually has more redundancy, and signal-to-noise ratio, which compares their rms values, becomes a relevant criterion.

Gaussian noise is theoretically unbounded; for a given rms level, very large peak-to-peak values are possible, but their probability decreases very rapidly (for example, the probability of 14×rms peak-to-peak is only 2.6×10^{-12}). Conventionally, noise specs use a 6× ratio of p-p p/rms (0.27% probability of higher peaks).

For white noise (constant noise spectral density, e_n nV/√Hz), the rms value in a given bandwidth is the product of e_n and the square root of bandwidth, i.e., $e_n\sqrt{B}$, where B is a "brick-wall" noise bandwidth, $f_2 - f_1$. For converters, of resolution N bits, the target value of errors, 1/2 LSB, is $V_{REF}/2^{N+1}$. So, for 1/2-LSB rms white noise, the noise spectral density has to be

$$e_n \leq \frac{V_{REF}}{2^{N+1}\sqrt{B}}$$

and for 1/2-LSB peak-to-peak rms white noise, divide by 6:

$$e_n \leq \frac{V_{REF}}{6 \times 2^{N+1}\sqrt{B}}$$

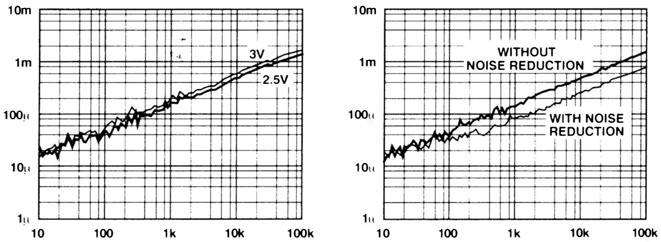
For a 10-V, 12 bit, 100-kHz system with an unfiltered reference, the p-p noise requirement is modest, 640 nV/√Hz. Table 4 provides the set of required values for resolutions from 12 to 16 bits using 10-, 5-, and 2.5-V references. Note that the required e_n decreases with increased resolution, decreased V_{REF}, and increased bandwidth. However, the user can control the bandwidth with filtering, to make a noisy reference more useful.

Table 4. Reference Noise Requirements for Various System Accuracies (1/2 LSB/100 kHz Criteria)

Bits	Noise Density (n/V/√Hz), Various FS Ranges		
	10 V	5 V	2.5 V
12	643	322	161
13	322	161	80
14	161	80	40
15	80	40	20
16	40	20	10

From Tables 1 and 2, bandgap and Zener references are available with noise densities from 100 nV/√Hz (usually specified at 100 Hz) and low-frequency noise from 4 μV p-p. Figure 8 illustrates the noise as a function of bandwidth and frequency for the AD780 (a) and the AD587 (b). These plots are taken with a swept bandpass filter with a gain of 100 and noise bandwidth, $B = f_2 - f_1 = 0.2316F$, in the vicinity of each frequency, F. Since the bandwidth is proportional to F and noise is proportional to $\sqrt{B}$, the plotted noise will rise at about 3 dB per octave for white noise (constant e_n). At a given frequency, "F", e_n can be calculated by dividing the reading by $100 \times \sqrt{(0.2316F)}$ (approximately $48\sqrt{F}$, about 1522 $\sqrt{\text{Hz}}$ at 1 kHz).

For example, the AD780 at 2.5 V (averaged plot through lower curve) reads about 160 μV at 1 kHz, whence the noise density is $160 \times 10^{-6}/1522 \cong 105$ nV/$\sqrt{}$Hz, as expected. The noise at 10 kHz is about $460 \times 10^{-6}/(48\sqrt{10000}) \cong 96$ nV/$\sqrt{}$Hz). Noise in the 3-V mode is proportionally higher. These curves reflect standard



(a) (b)

Figure 8. Various references, bandpass noise (V × 100) vs. bandpass frequency (Hz); (I_L (DC) = 2 mA). a. AD780, 2.5- and 3-V modes. b. AD587, with & without noise-reduction capacitor.

operation for the AD780; when the suggested noise-reduction capacitors are used, the noise at 1 kHz and 10 kHz is reduced to 53 and 32 nV/$\sqrt{}$Hz, respectively (not shown).

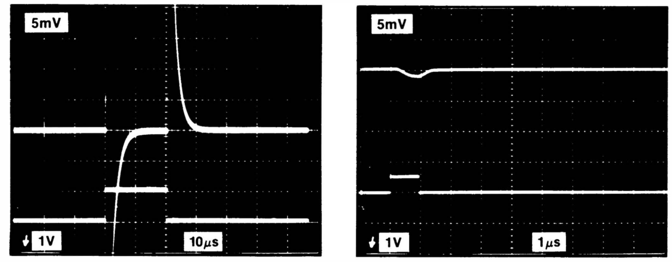
Data is plotted in Figure 8b for the AD587 in standard operation (upper), and with the recommended noise bypassing (lower). The noise is reduced by about 5 dB, from 200 Hz to 100 kHz; for example, at 1 kHz the reduction is from about 106 nV/$\sqrt{}$Hz to 59 nV/$\sqrt{}$Hz. Noise bypassing works similarly for AD586, AD588 and AD688. The capacitor used should be a low leakage type (e.g., compact stacked film) placed close to the pin.

A useful alternative with any reference is a dc-accurate post-filter stage. This involves a low-impedance single- or multiple-pole low-pass filter, buffered by a precision low-noise op amp; it passes the reference voltage while removing high frequency noise (an example is shown in detail below).

Reference pulse response: Often of concern in reference applications is the transient change of output voltage in the presence of stepped loads. Fast load changes of up to full-scale current perturb the output voltage, often beyond the rated error band. Key questions: how quickly does the output transient recover to within the rated accuracy band after a load change? Can anything be done to reduce the effect?

For example, Figure 9 shows the response of a REF43 IC to a 10-mA load step, for conditions of no output decoupling (a) and with the recommended decoupling network (b) [10 μF tantalum in parallel with 0.01 to 0.1 μF ceramic from V_{IN} and V_{OUT} to ground]. Without decoupling, the output has a large spike, settling to within ±2.5 mV in 3-4 μs and producing a further disturbance in circuits served by the reference. When decoupled, the output remains within the error band, changing only slightly.

For references with similar dynamics to the REF43, output decoupling is useful in maintaining control. However, additional output capacitance may or may not be allowable towards buffering a given reference type against transient loads, so specific data-sheet recommendations should be followed.



(a) (b)

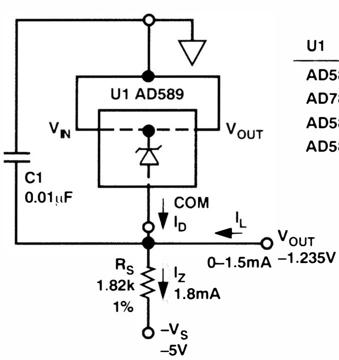
Figure 9. REF43 pulse response for 0-10-mA load change, with and without decoupling. a. Response with no decoupling. b. Response with output decoupling of 1 μF & 0.01 μF.

REFERENCE CIRCUIT APPLICATIONS

Shunt references: As noted earlier, shunt mode references can be used in either polarity, but they have the disadvantages of limited drive and relatively high output impedance; these tend to restrict them to applications with limited ranges of load variation.

Figure 10 illustrates a shunt-mode application, using two-terminal devices and three-terminal devices with V_{IN} and V_{OUT} jumpered for two-terminal operation. Here, an AD589 is used with a negative supply to provide a regulated negative reference. All shunt-operated references should be designed with careful attention to dc currents. R_S must be selected appropriately to maintain shunt current, I_D, in a limited range for any specified combination of load current, I_L, and supply voltage, V_S. With the AD589, R_S is chosen to allow 1.8 mA to flow with the magnitude of V_S 10% low. This will allow a 0 to 1.5-mA load current range, and the device will remain in a safe range with the load removed and $|V_S|$ 10% high. The AD589's typical R_Z of 0.6 Ω holds output changes <1 mV for a 1.5-mA I_L change.

A bypass capacitor is recommended to reduce high-frequency noise and ac impedance; the value shown in the figure can be increased for further impedance reduction. As the table shows, a wide range of voltages are available—with the same device in some cases (AD584, AD780); the AD780 provides the lowest output impedance.



U1	V_{OUT}(V)	I_D(mA)	R_Z(Ω)
AD589	1.235	0.05–5	0.6
AD780	2.5/3.0	1–10	<0.075
AD581	10	1–5	2
AD584	5/7.5/10	1–5	2

$$R_S = \frac{|V_S| - |V_{OUT}|}{I_Z}$$

(I_Z WITHIN I_D RANGE)

Figure 10. Shunt reference IC operation.

Negative references: Positive voltage references are the most widely produced and used, but negative references are often required. While this can be accomplished with a simple shunt reference, it is usually more desirable to operate negative references as current-buffered voltage sources, to avoid the shunt circuit's loading restrictions.

A simple technique is to cascade a stable positive voltage reference with a precision inverting scaling amplifier, using a high quality op amp, such as an AD707 or OP177, and a high-accuracy resistor pair. This approach is workable and straightforward but its cost/performance is limited by the tradeoff between resistor accuracy and the cost of precision resistors—which can greatly exceed the cost of the op amp.

A more direct approach for a negative reference is to use an IC specifically designed for such use, namely the REF08 (Table 2). The buried-Zener REF08 is designed as a 10-V three-terminal negative reference; it functions as a mirror image of 10-V positive references, such as the AD587. It is applied in simple fashion, furnishing a −10-V output with tolerances of ±30 mV or ±40 mV; a ±270-mV trim range is available. With pin 4 strapped to ground, it furnishes an alternative −10.24 V ±40 or ±60 mV, suitable for easy scaling in 10-mV/LSB 10-bit applications. The REF08 is available with TCs of 50 or 100 ppm/°C.

Another alternative is to "invert" positive IC references, a design approach valuable because of the wider array of high-performance references from which to choose (Table 1); and the elimination of resistors and their scaling/drift errors. Resistorless inverters basically enclose the IC reference within a precision op amp feedback loop, driving the common (or negative) terminal so as to maintain the normally positive output at ground; thus the reference IC's common terminal is driven at $-V_{REF}$.[15]

An example of this scheme is illustrated in Figure 11, using an AD780 (or other positive three-terminal IC) for U1 as the reference IC. Overall, this circuit supplies a stable $V_{OUT} = -V_{REF}$, where V_{REF} is the 2.5-V (or other) voltage at U1's output (with no load). Thus the circuit inverts a positive reference's output, without the expense or errors of a precision resistor pair. The full dc precision of the basic reference IC is easily maintained, due to the buffering by the op amp, U2, while ac performance can be optionally improved even further.

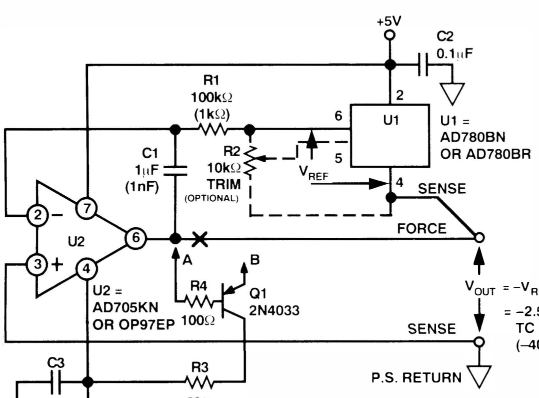


Figure 11. Buffered "inverted" negative reference.

DC accuracy is optimized largely by careful selection of the U1 device from Table 1, for initial tolerance, drift, etc. The drift characteristics of the op amp are usually less important, but may not be negligible. For example, a 1 μV/°C drift in U2 is equivalent to a negligible 0.4 ppm/°C drift in U1's 2.5 V. U2 should also have V_{OS} <100 μV, for negligible contribution to tolerance error. Given this drift/accuracy criterion, U2 should be a high DC precision type such an AD705KN or OP97EP when used with a low-drift IC such as the AD780 (or other). Current drain for the unloaded circuit is typically 1.2 mA.

U2 determines the output drive capability of the circuit. For lowest self-heating errors, dissipation in U2 should be minimized, with current outputs restricted to 10 mA or less (including the quiescent current of U1). Substantially higher currents of say 50 mA can be accommodated without side effects using a PNP booster transistor such as a 2N4033, inserted between points "A" and "B". Lower output impedance in wideband applications is available with the AD820 for U2, with some tradeoffs in dc accuracy and drift. Connecting the FORCE and SENSE leads as noted minimizes wiring-drop errors.

Filter R1-C1 sets the integrating time constant in U2 to promote stability and noise reduction. With the choice of larger values, the broadband noise of U1 is reduced to a minimum, and overall noise is close to the noise of U2; a typical measurement is <20 nV/$\sqrt{Hz}$ at 1 kHz. If attenuation of reference noise is not necessary, the smaller values (in parentheses) should be used. Using the devices in the figure, the (untrimmed) output is −2.5 V ±1.1 mV, with a TC of 3.4 ppm/°C. U1 and U2 are available in both SOIC and DIP packages.

Shunt mode references, as essentially floating ICs, are used for positive or negative outputs. When used with a buffer op amp, in an inverted configuration, their load-current restrictions are removed. Figure 12 is an example of a buffered inverted shunt-mode negative reference. This circuit is similar to Figure 11, but includes input resistor, R1, to supply bias current for the reference diode. Because the op amp's (+) input is grounded, the feedback loop holds the positive terminal of D1 at virtual ground. As a result, the amplifier output is driven at the reference voltage, which, in the case of the AD589, is −1.235 V. The diode requires only a small bias current; the available load current is then limited only by the output specification of U1 (15 mA minimum).

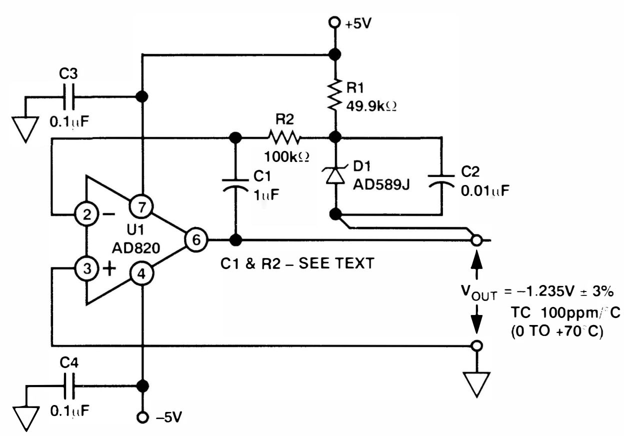


Figure 12. Buffered "inverted" low noise shunt reference.

This scheme can also be applied for the opposite output polarity (positive output), with D1 reversed and R1 returned to a negative voltage. For other supply voltages, R1 can be chosen for a standard value of D1 bias current, say 100 μA. Note that this principle applies for other voltages, using devices designed to operate in the shunt mode (see table of Figure 10). Kelvin sensing, used as shown, maintains high dc accuracy at the load.

As in the circuit of Figure 11, integrator time constant, R_2C_1, can be optionally chosen for noise filtering as shown; the noise is reduced to that of the op amp used. Using the AD820, the filtered noise measures ≅15 nV/√Hz at 1 kHz, and the circuit's output impedance is ≅0.05 Ω at 1 kHz. Without this filtering (much smaller time constant), noise output of the circuit is that of the AD589, or ≅200 nV/√Hz. In this simple circuit, the output accuracy is ±3%, that of D1. Trim is possible, using a resistive divider across D1, feeding R2, to adjust the output voltage.

Low-noise references for wide-dynamic-range converters: High-resolution converters, including Σ-Δ and (especially) high-speed types, benefit from the improved noise and load-capacitance tolerance of recently available references.

Figure 13 shows the AD780 used as a 3-V reference for the AD711x series of >20-bit Σ-Δ converters. The 3-V scaling (rather than 2.5) enhances the dynamic range of this and many other 5-V single-supply converters, while the ≅4-μV p-p noise (0.1-10 Hz) minimizes overall system noise.[16] In addition, the large decoupling capacitance at the converter's REF IN pin minimizes voltage errors due to transients. These same factors also enhance performance of wider bandwidth converters such as the 16 bit AD7884.[17]

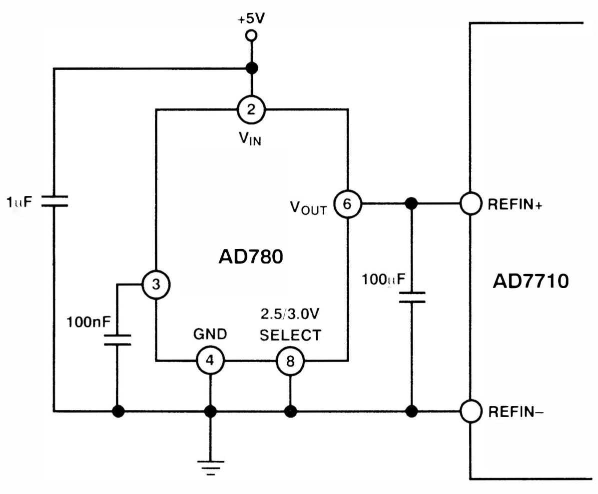


Figure 13. Precision 2.5/3-V reference for the AD7710-series high-resolution Σ-Δ ADCs.

Current-boosted 50-mA three-terminal reference: For highest dc accuracy, output current of reference ICs should be kept well within the 10-mA rating. Thus for loads of substantially more than about 5 mA, some form of booster should be considered. This can be accomplished with the addition of a PNP pass transistor, as shown in Figure 14.

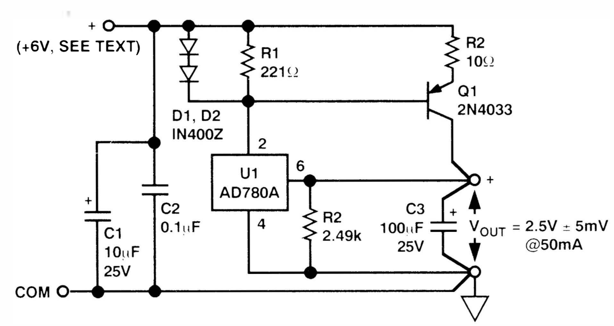


Figure 14. 50-mA current-boosted three-terminal reference with current limiting.

In this circuit the input current of U1 through R1 develops a base drive for Q1, whose collector provides the bulk of the load current. U1 never is called upon to furnish more than a few mA, so this minimizes internal temperature differential and drift. Short-circuit protection is provided by the diode clamps which limit drive to Q1, at about 80 mA of load current.

Other references and output voltages can be used for U1, but may require some R1 adjustment, dependent upon their I_q. The booster current-limiting configuration causes the dropout voltage of the circuit to increase, and operation from a +5-V supply may be marginal, especially when references having greater dropout voltage are used for U1.

Besides increasing output current, Q1 decreases output impedance; the loaded output impedance at 1 kHz is <10 mΩ, about 10-20 times better than for the AD780 alone. C_3 is high to minimize high-frequency output impedance (400 mΩ at 100 kHz), but smaller values can also be used. If an input voltage appreciably more than +6 V is used, heat sinking may be needed for Q1.

Bipolar-reference bridge driver: For optimum operation of a dc bridge, bipolar drive is useful; it virtually eliminates the output common mode component. Figure 15 shows an implementation using either an AD588 or AD688 buried-Zener reference with Kelvin sensing of the drive voltages at the bridge.

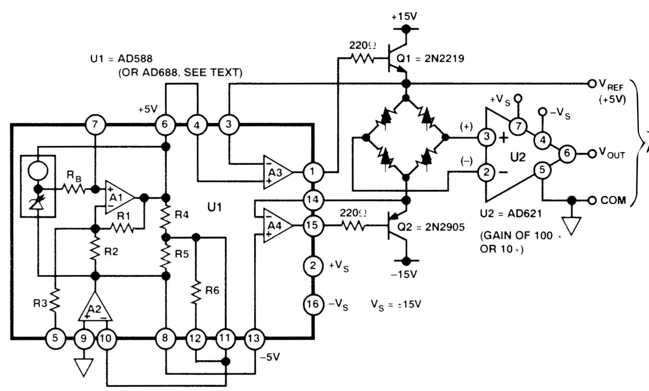


Figure 15. Bipolar (±5-V) bridge with precision in-amp.

The AD588 Zener cell and A1 produce a 10-V total potential, which can be shifted either above or below common, or split into ±5 V by R4 and R5, with ground enforced by A2, as shown here. The basic ±5-V signals then appear at pins 6 and 8; they drive Kelvin sensing follower-amplifiers A3 and A4. The amplifiers sense the voltage at each end of the bridge and drive those points through external current boost transistors Q1 and Q2, respectively, forcing the bridge end points to equal the ±5 V output of the reference, within the symmetry specifications of the AD588 (±1.5 mV). Metal-can transistors are used for Q1 & Q2, for best dissipation at the 30-mA drive level. For ±10-volt drive, the AD688 can also be used in a similar fashion.

The output sensing in-amp is the AD621, which provides tap-selectable gains of 10 and 100, with a gain tempco of 5 ppm/°C or less, lower in fact than can easily be done with readily available gain set resistors and a single ended amplifier. The scaled bridge output signal can drive an ADC; which can also use the +5-V bridge reference voltage as the conversion reference.

30-mA reference with shutdown: The REF195 bandgap reference is like the other references of Table 1; but it has a unique shutdown capability, which allows a precision 5-V output to be turned ON and OFF by a TTL/CMOS compatible digital input, as shown in Figure 16. It has a low dropout of 0.5 V at 10 mA and low current

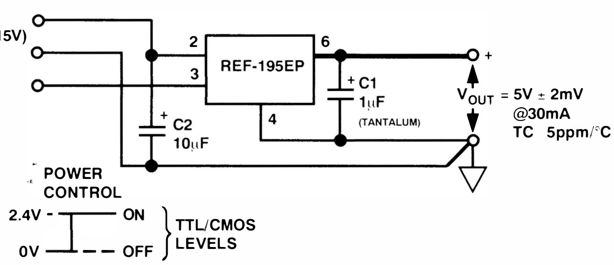


Figure 16. 30-mA reference with shutdown.

drain for both quiescent and shutdown states, 45 and 15 μA (max), respectively. For inputs in the range of 6.3 to 15 V, the REF195EP shown can furnish 5 V (±2 mV) for loads of up to 30 mA, with a 5-ppm/°C max TC.

The shutdown pin (3) is controlled by TTL or CMOS logic levels, with a "1" (or $+V_S$) commanding the output ON, while a low input shuts it OFF. With a 1-μF load bypass capacitor (the minimum recommended), transition OFF-ON time is several hundred μs. This OFF-ON transition time will be load dependent, increasing for higher C_L values. ON-OFF transition timing is determined by the load current and C_L.

To maximize DC accuracy in this circuit, the output of U1 should be connected directly to the load with short heavy traces, to minimize IR drops. The common pin is less critical, due to the much smaller current returning to the device.

Low noise 2.5/5/10-V reference: As noted earlier, voltage reference noise can contribute to system error. But the output of a reference can be buffered and filtered to effectively lower wideband noise by an order of magnitude or more.[18] For example, the low-noise reference circuit of Figure 17, using simple filtering, combines good ac and dc performance. It comprises a reference, U1, and a low-noise, buffered output circuit. Final output noise is largely determined by U2, and can range from under 2 nV/√Hz, to 20 nV/√Hz or more at 1 kHz, depending on the device.

The basic reference voltage is set by U1, a 2.5, 5 or 10-V IC chosen from Table 1 for required accuracy and drift. This circuit uses an AD586MN, a 5-V (±2 mV) buried-Zener reference with a 2 ppm/°C drift and low 1/f noise. U1's stable 5-V output is applied to a R1-C1/C2 noise filter, using electrolytic capacitors for a low corner frequency. DC leakage errors are minimized by bootstrapping C1 so as to see only the small R2 dc drop as bias, effectively lowering leakage to negligible levels. The filter corner frequency is about 1.7 Hz, providing about 35 dB of attenuation at 100 Hz. Attenuation is modest below 10 Hz, so reference choice is still important to noise performance at low frequencies.

The filter's low-noise, dc-accurate output is buffered by a unity-gain buffer using an OP113EP low-dc-error, low-noise op amp. With less than ±150 μV of V_{OS} error and less than 1 μV/°C drift, the buffer's dc performance will not compromise accuracy/drift of Table 1 references. The OP113 has a typical current limit of 40 mA, more current than IC references usually provide.

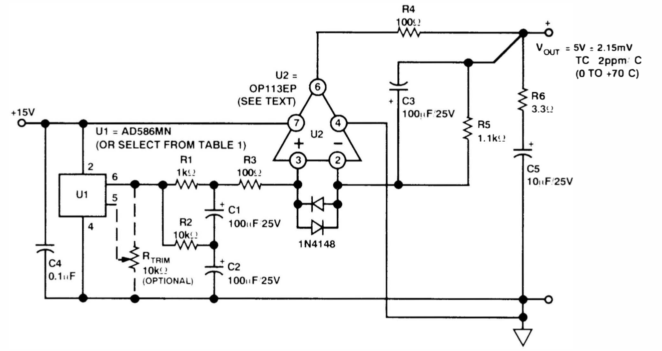


Figure 17: Low-noise 2.5/5/10-V reference.

While the single-supply OP113 is useful over the entire 2.5-10-V range, even lower-noise op amps are available for 5-10-V use. The AD797 has measured 1-kHz noise less than 2 nV/√Hz, compared to about 6 nV/√Hz for the OP113. Other 5-10-V range possibilities include the OP27 and OP176.

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