

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 $\pm 20V$ 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 perfor-

mance 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 loop⁶ (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

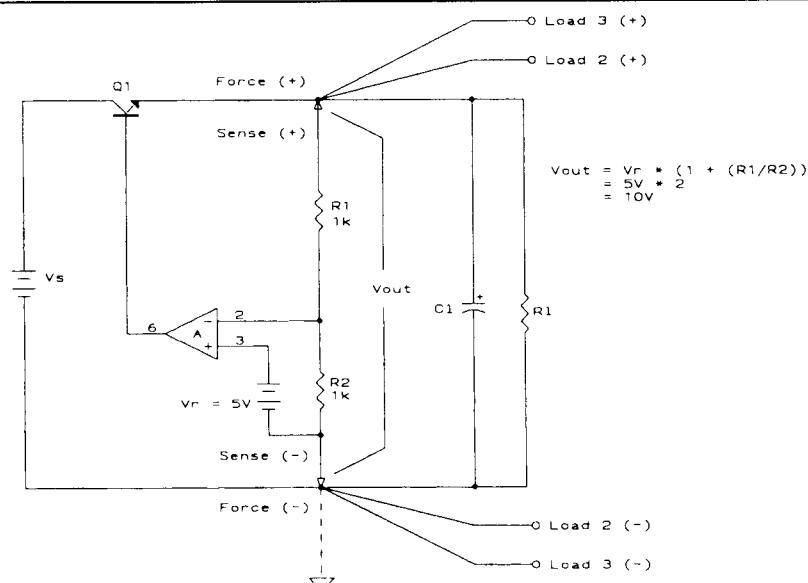


FIGURE 1: Series voltage regulator.

above the $\approx 100\text{mA}$ 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 $1\text{k}\Omega$ 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 ($<1\text{V}$). 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

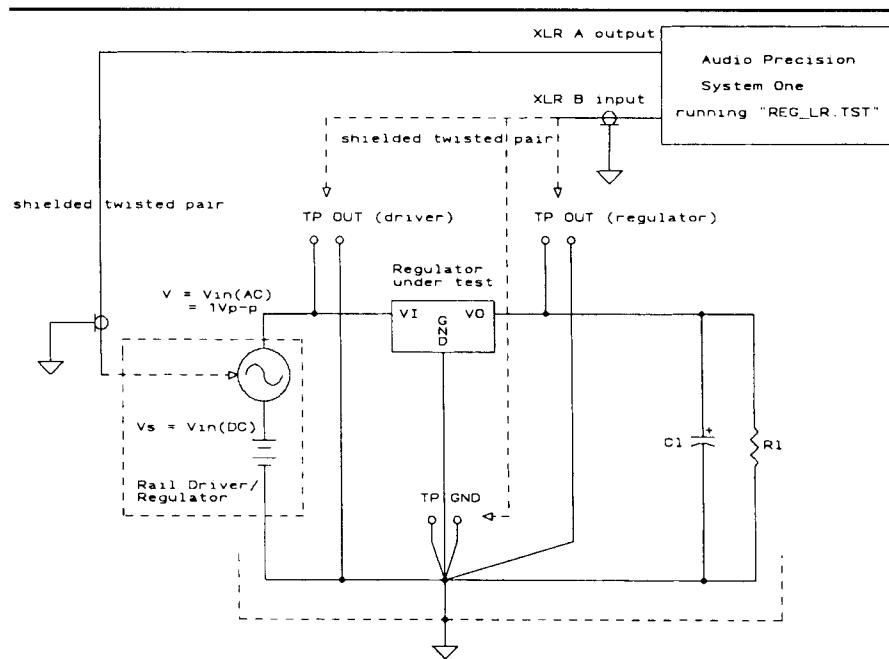


FIGURE 2a: Line rejection test.

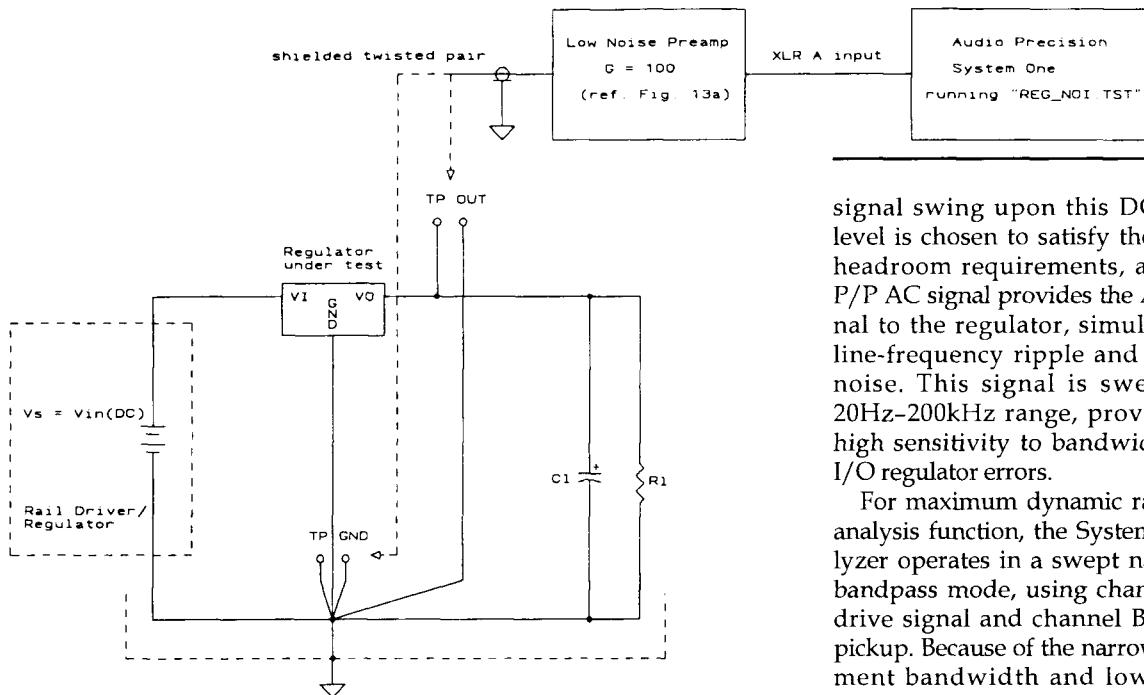


FIGURE 2b: Noise test.

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 fre-

quency. 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-

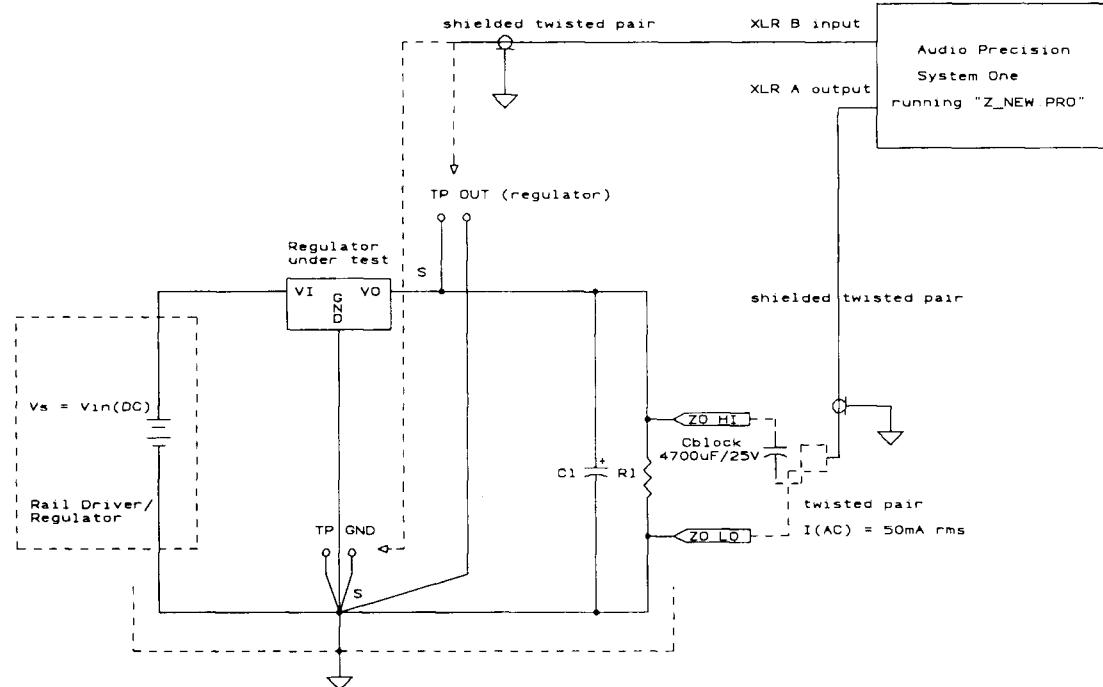


FIGURE 2c: Output impedance test.

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 $C_L = 100\mu F$ and $R_L = 100\Omega$, 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, narrowband bandpass mode.

A given noise test provides a 100 \times scaled display of regulator output voltage noise as a function of frequency. A

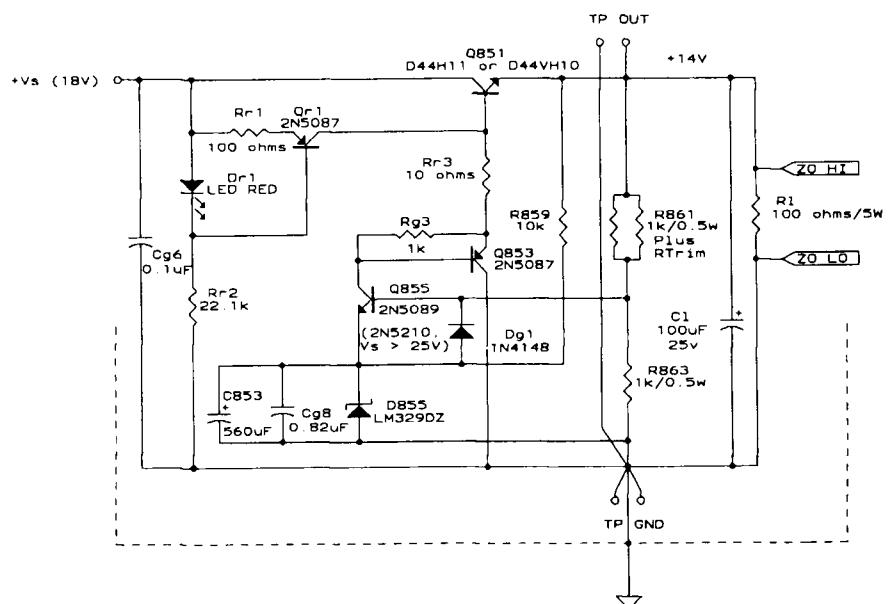


FIGURE 3: POOGIE 5.51 positive regulator.

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 $2.6nV/\sqrt{Hz}$, so this only becomes an issue for regulators with noise levels of about $7nV/\sqrt{Hz}$ 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 $4.700\mu F$ 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 $10\mu\Omega$ can be resolved at low frequencies, and $1m\Omega$ 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.

POOGIE 5.51 Regulators

The discrete regulator of Fig. 3 is an enhanced version of the POOGIE 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.

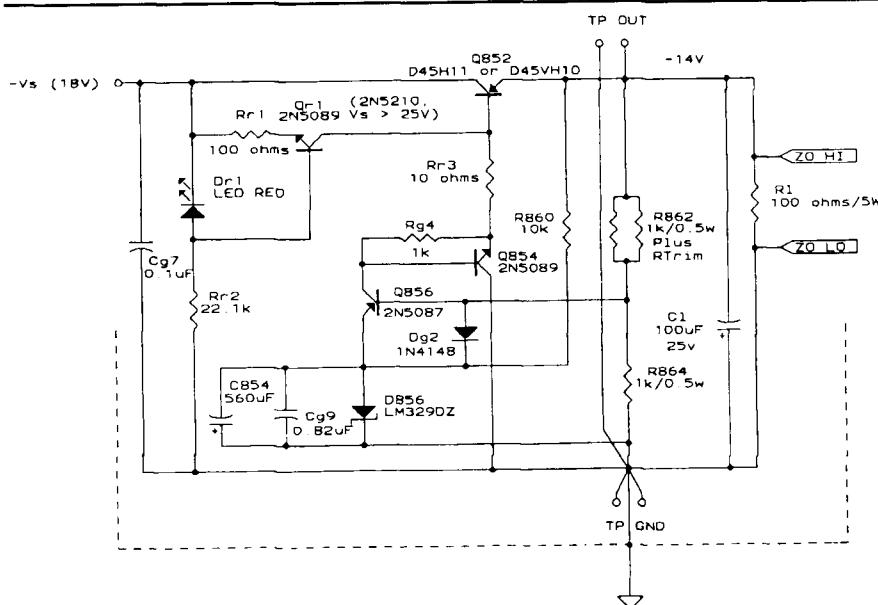


FIGURE 4: POOGIE 5.51 negative regulator.

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 POOGIE 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 POOGIE-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 POOGIE 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 POOGIE 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 pro-

vided 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 $C_l = 100\mu F$ and $R_l = 100\Omega$. The adjustable three-terminal regulators (types LM317, LT1085,

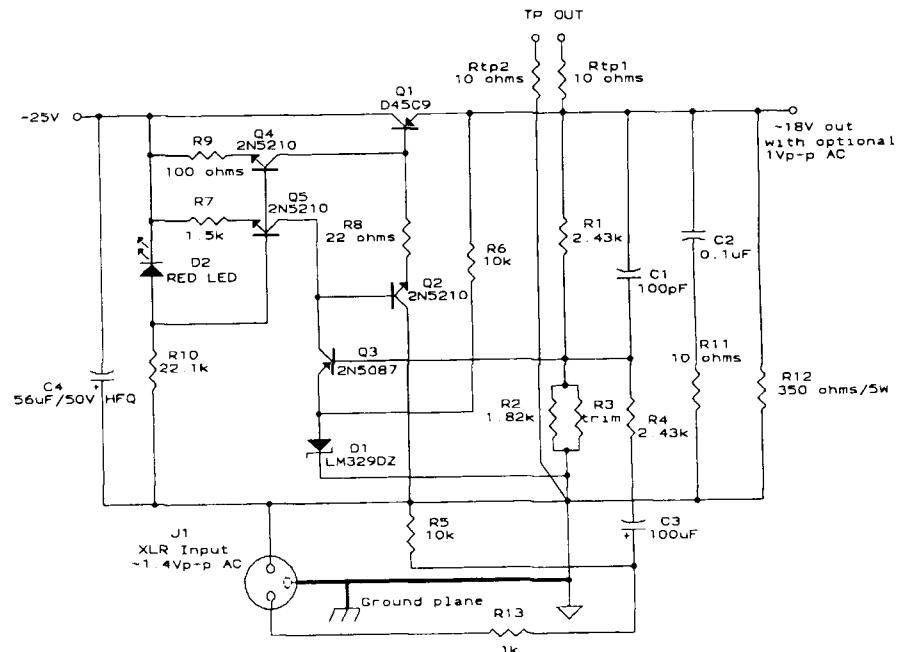


FIGURE 5: Negative rail driver/regulator.

LM337, and LT1033) are connected with an OUT-ADJUST pin resistance of $1\text{k}\Omega$ and an ADJUST-GND pin resistance of $10\text{k}\Omega$, which programs them to a nominal 14.2V. These also used a C_{ADJ} bypass capacitance of $100\mu\text{F}$, plus the loading of $C_1 = 100\mu\text{F}$ and $R_1 = 100\Omega$. 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 $10\text{k}\Omega$ resistor than with a $1\text{k}\Omega$ 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 $\pm 25\text{V}$ DC source, these drivers provide a regulator signal source environment which allows a fixed DC input level of $\pm 18\text{V}$, upon which you can optionally superimpose a 1V P/P AC test signal. In the line rejection tests, the $\pm 18\text{V}$ 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 $0.1\mu\text{F}$ input bypass capacitor for stability.

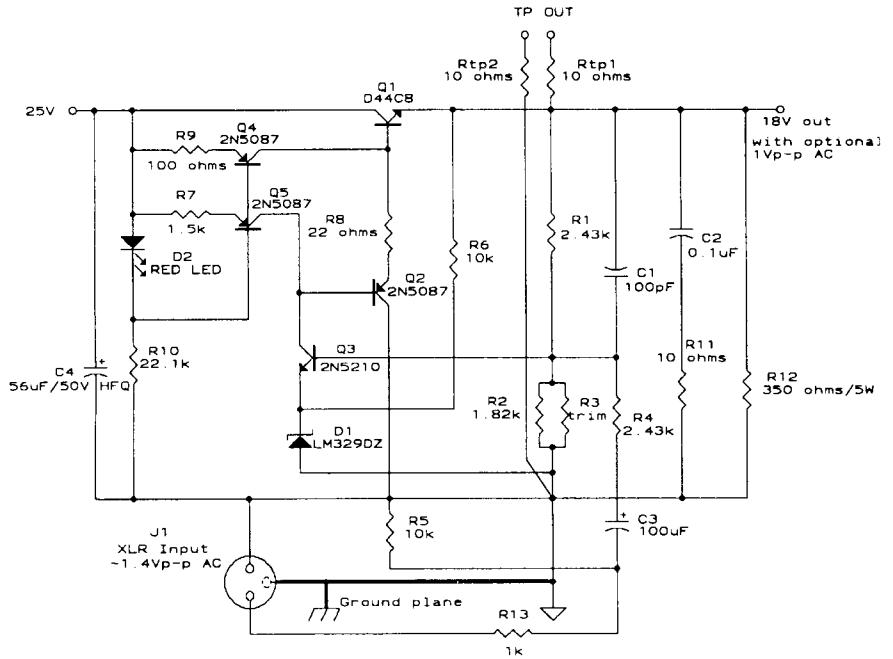


FIGURE 6: Positive rail driver/regulator.

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

The voltage slew rate to maintain a flat-frequency response into the $0.1\mu\text{F}$

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 POOGIE 5.51 plus/minus regulators, the

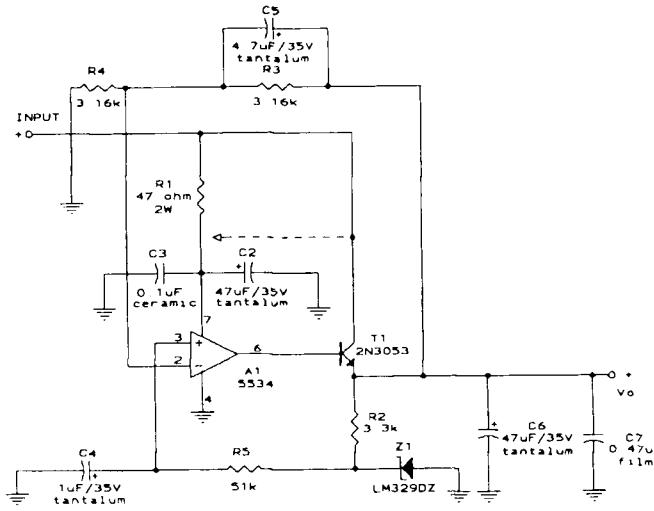


FIGURE 7a: Positive output Sulzer regulator.

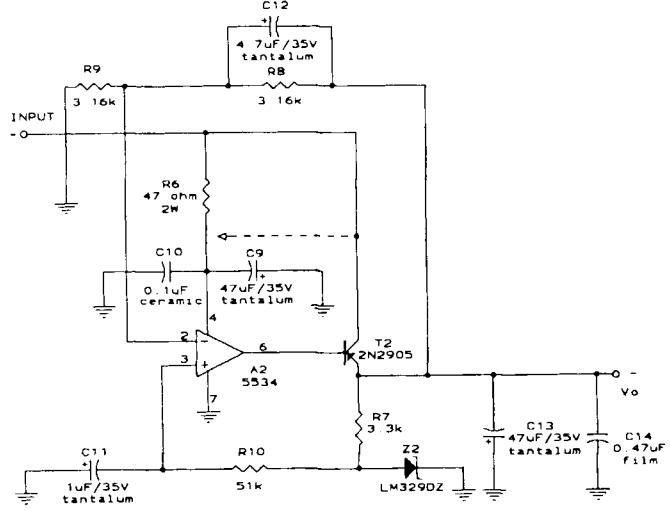


FIGURE 7b: Negative output Sulzer regulator.

output divider ratio and a reference voltage of $\geq 7.5V$ (6.9V plus 1V_{be}) 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 $18 \pm 0.1V$ 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 POOGIE 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_{AD1} 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 regulators³ 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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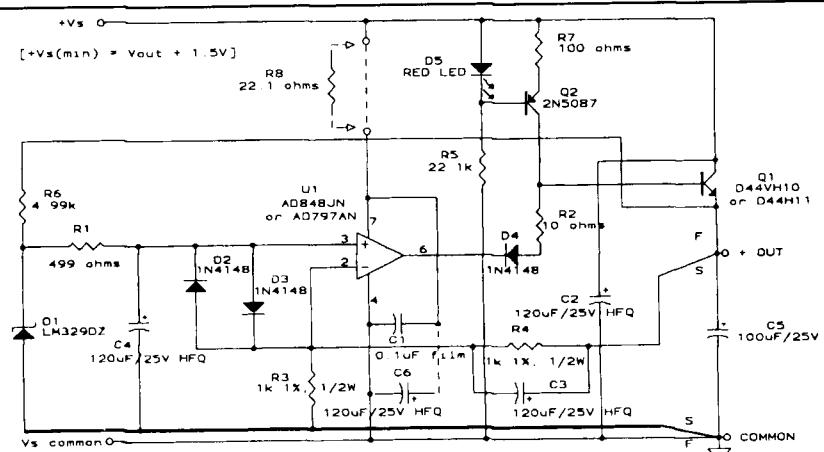


FIGURE 8a: Positive output, low-noise, low-dropout regulator. For this figure and Figs. 8b and 9, unless otherwise specified, all resistors are $1/4W$, 1% metal film. A heatsink is required on Q1 for Pd $> 0.5W$. In this figure and Fig. 8b only, minimize lead length from unregulated input and return. "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 al⁷). 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 $\pm 14V$.

Sulzer rated the original circuit at 100mA, so in this case I set load resistor R_l at 150 Ω (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 POOGIE 5 regulator,¹¹ and

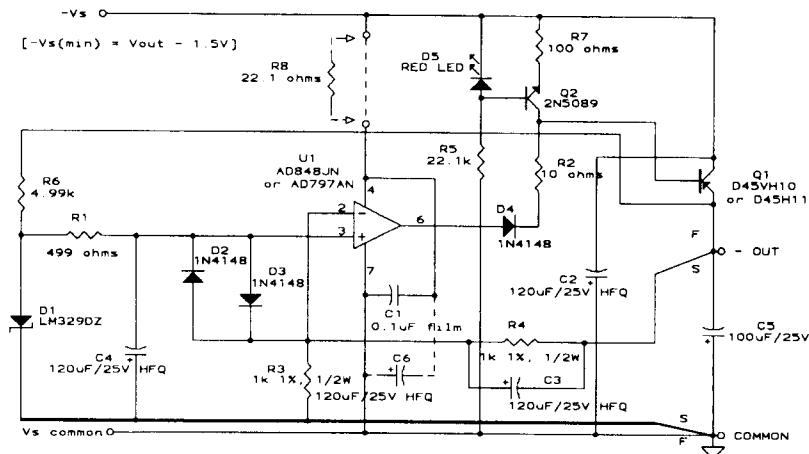


FIGURE 8b: Negative output, low-noise, low-dropout regulator.

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-

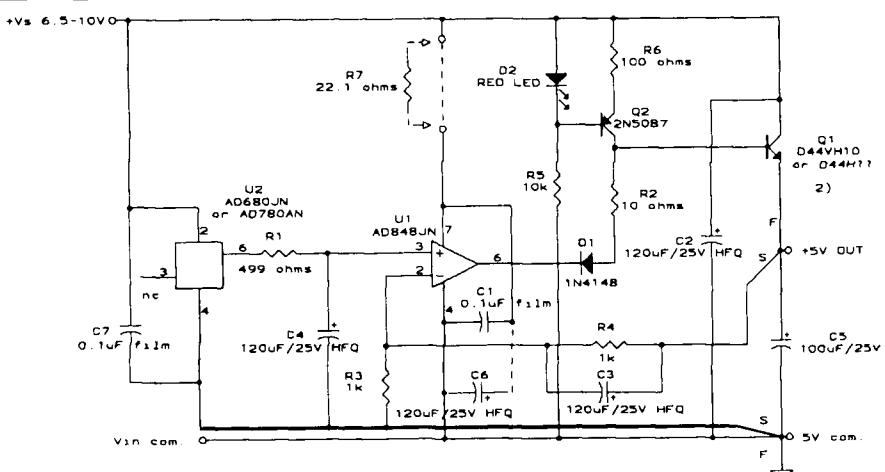


FIGURE 9: 5V, low-noise, low-dropout regulator. Minimize lead length from $+Vs$ and return.

um-level results for Z_O , LR, and noise takes both a skilled designer, plus lots of parts. For example, the POOGIE 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 POOGIE 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 500 Ω 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 $1\text{nV}/\sqrt{\text{Hz}}$ 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 120 $\mu\text{F}/25\text{V}$ -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 $\approx 14V$, 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. □

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 $C_1 = 100\mu F$ and $I_1 = 150mA$.

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.

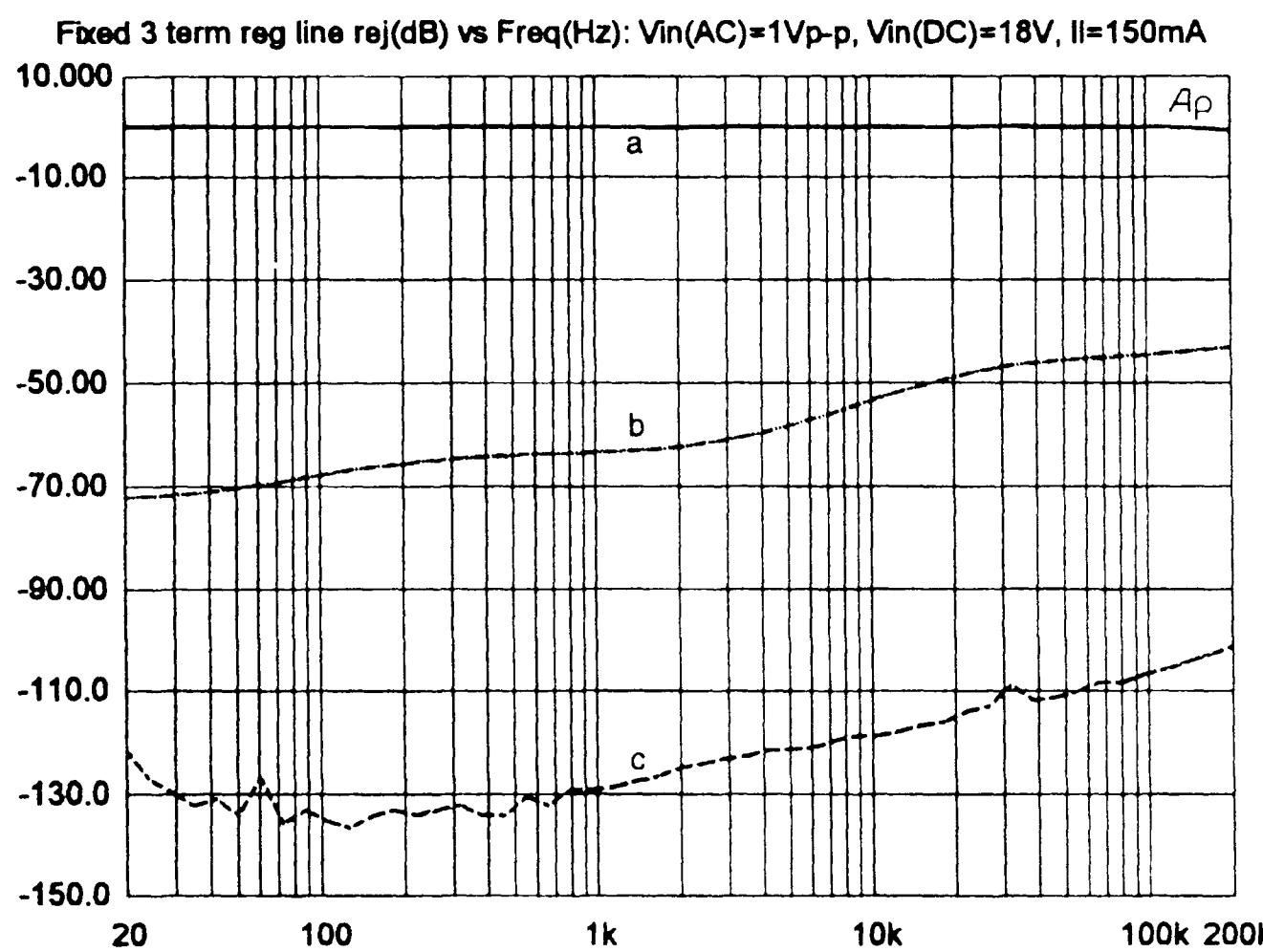


FIGURE 10: Fixed three-terminal 7815 regulator LR performance, a) $V_{IN} = 0dB$; b) V_{OUT} ; c) GND reference.

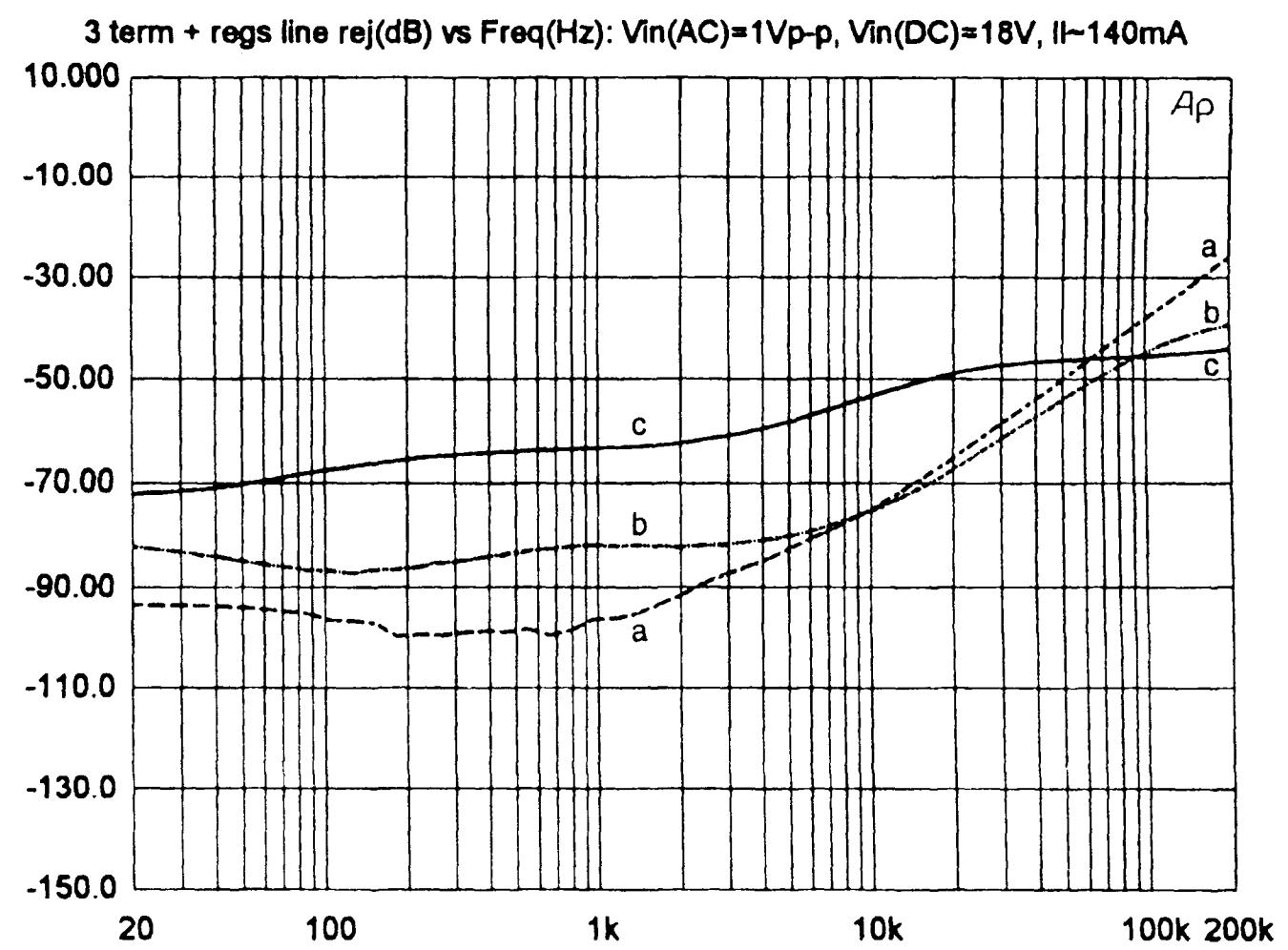


FIGURE 11a: Three-terminal positive regulator LR, a) 1085; b) 317; c) 7815.

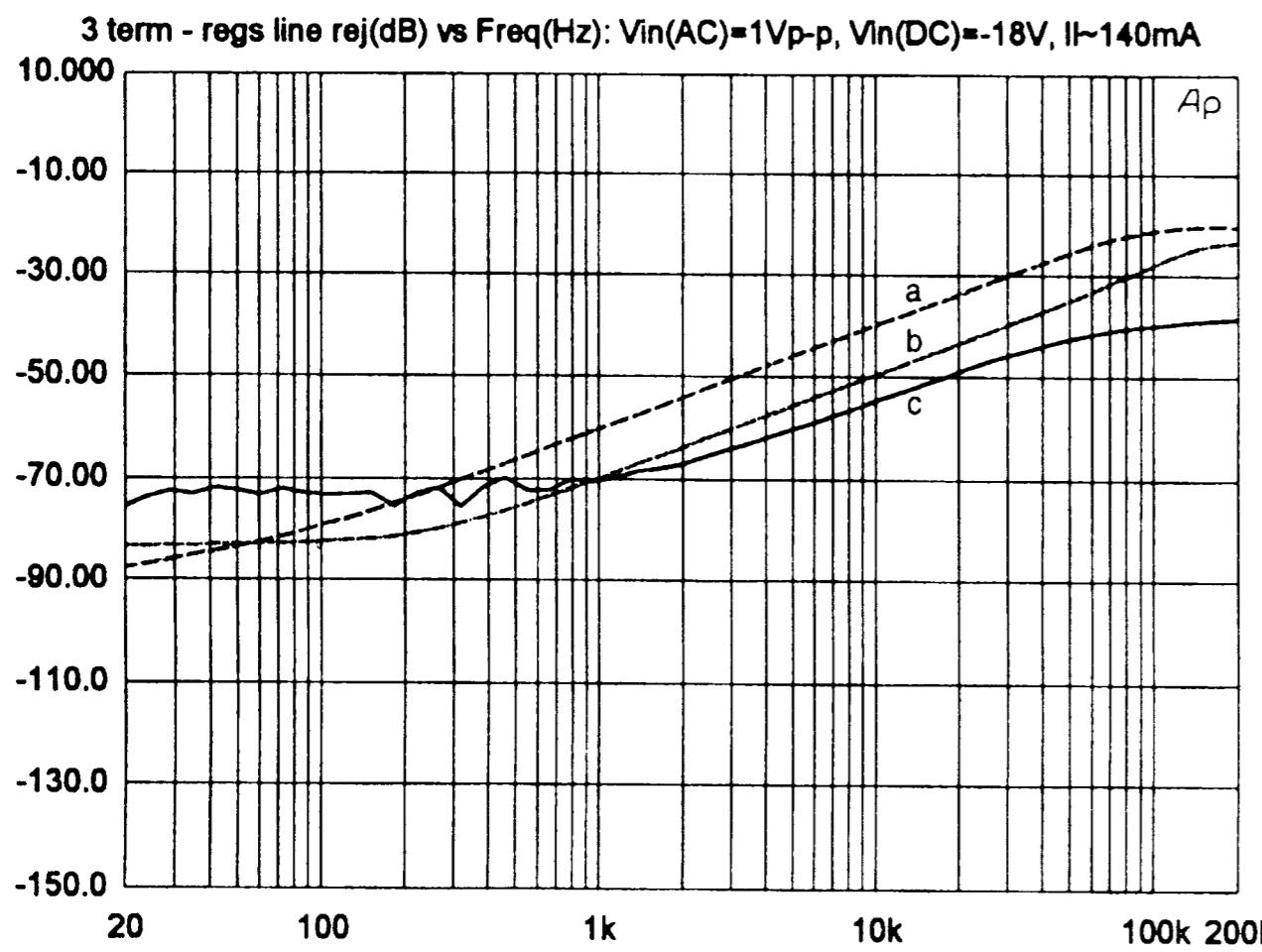


FIGURE 11b: Three-terminal negative regulator LR, a) 1033; b) 337; c) 7915.

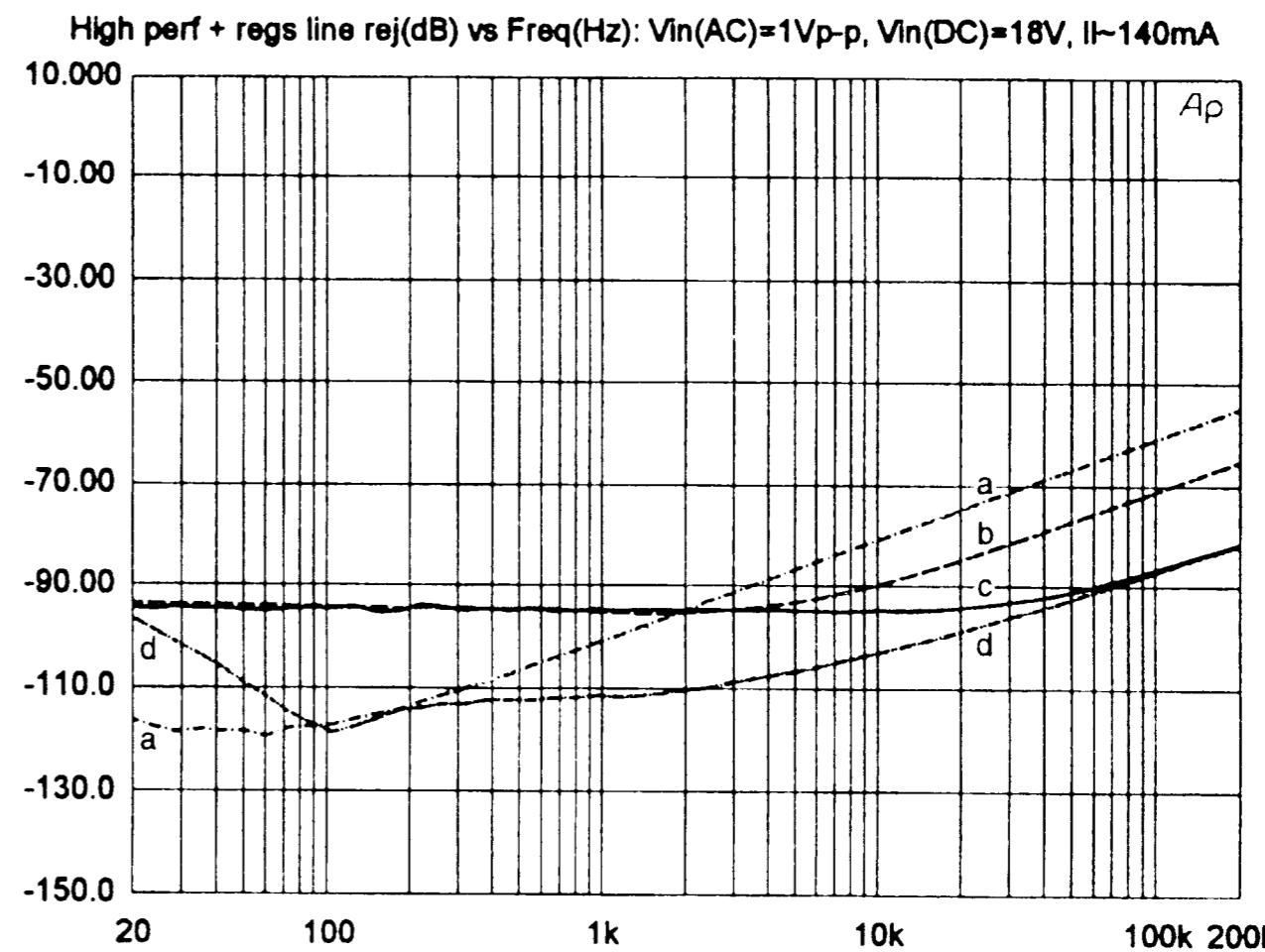


FIGURE 12a: High-performance positive regulator LR, a) AD797 (Fig. 8a); b) AD848 (Fig. 8a); c) POOGIE 5.51 (Fig. 3); d) Sulzer circuit (Fig. 7a).

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 POOGIE 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

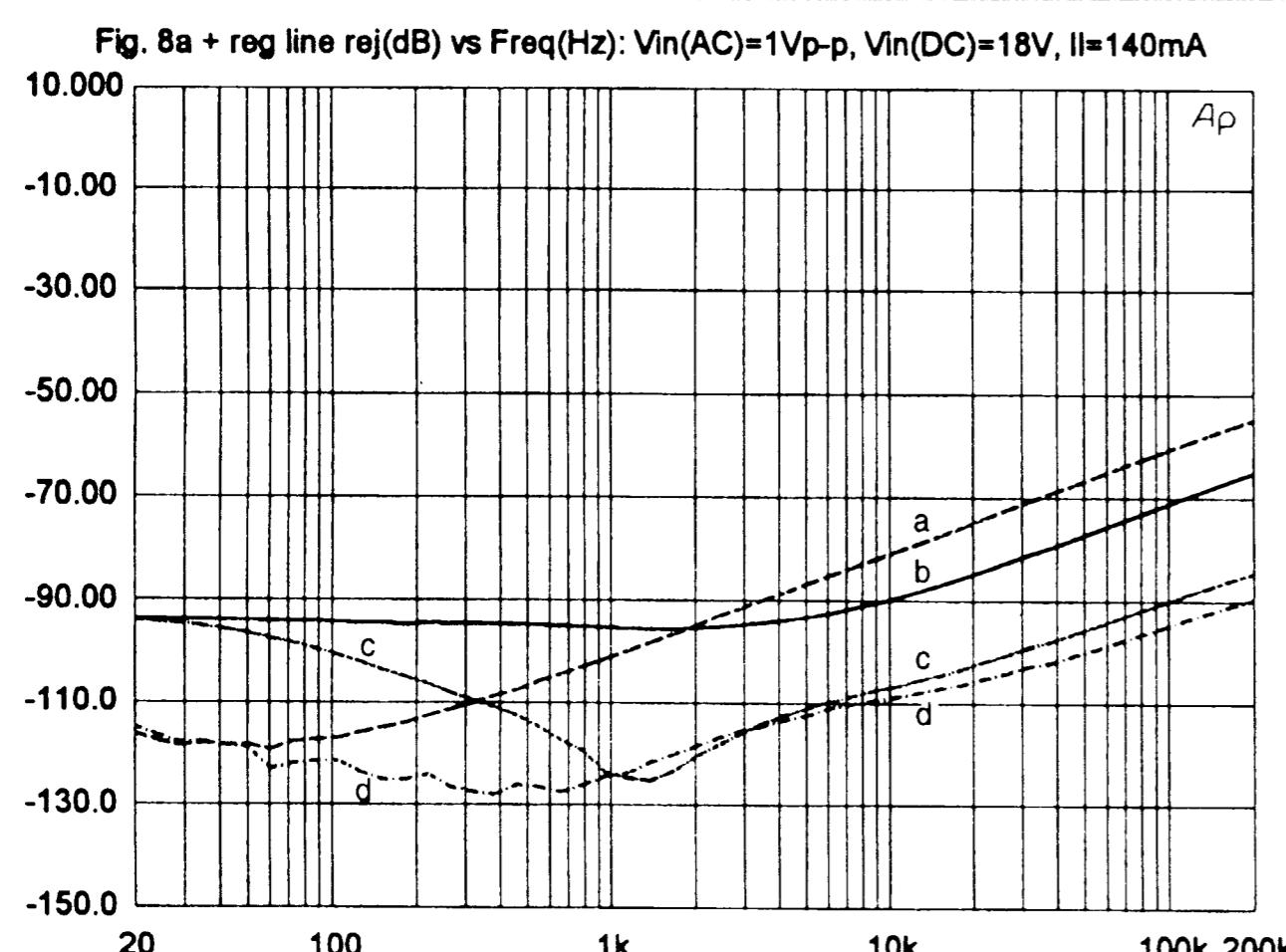


FIGURE 12b: Decoupling effects on LR (Fig. 8a), a) AD797; b) AD848; c) AD848 with decoupling; d) AD797 with decoupling.

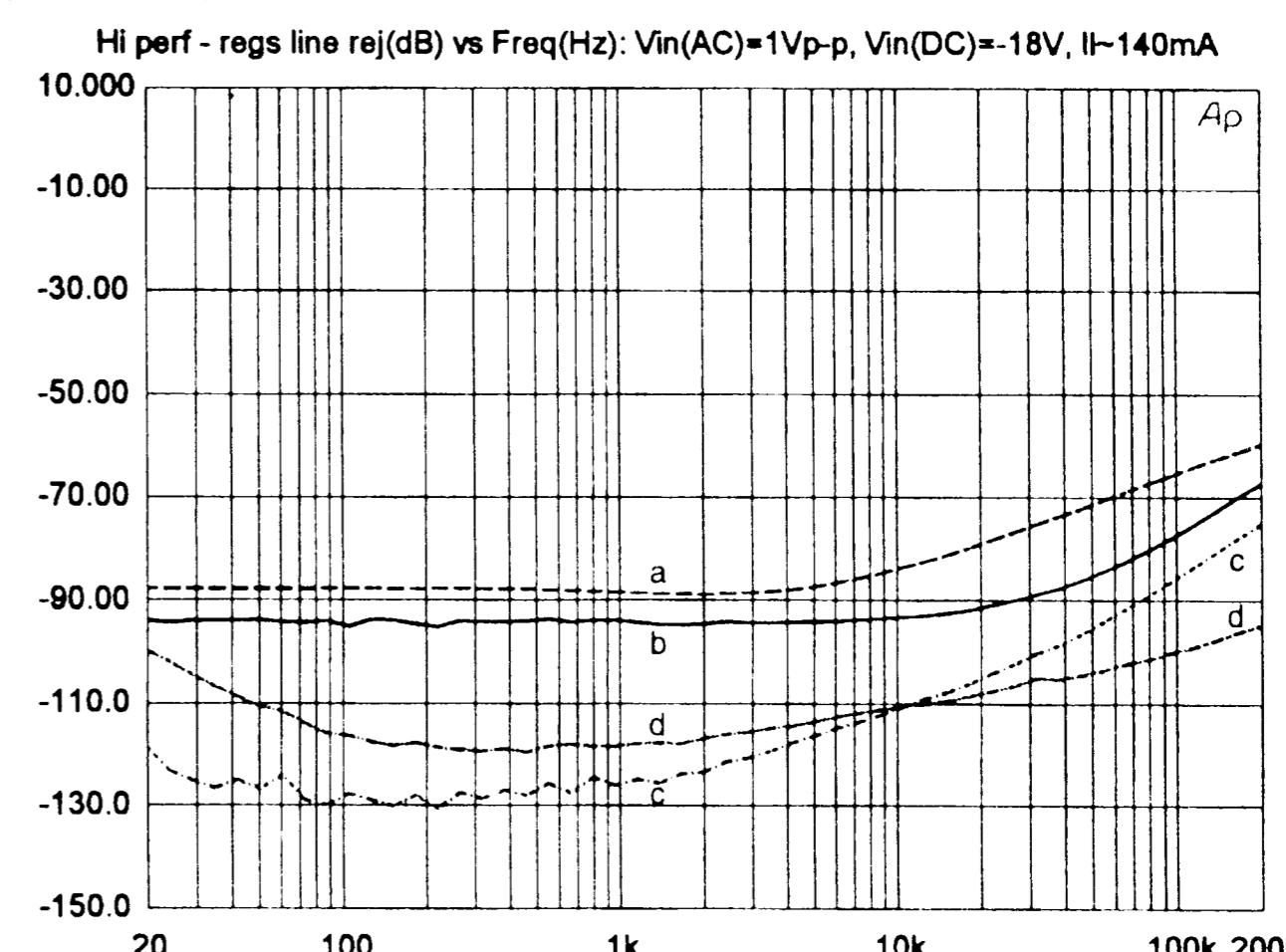


FIGURE 12c: High-performance negative regulator LR, a) AD848 (Fig. 8b); b) POOGIE 5.51 (Fig. 4); c) AD797 (Fig. 8b); d) Sulzer circuit (Fig. 7b).

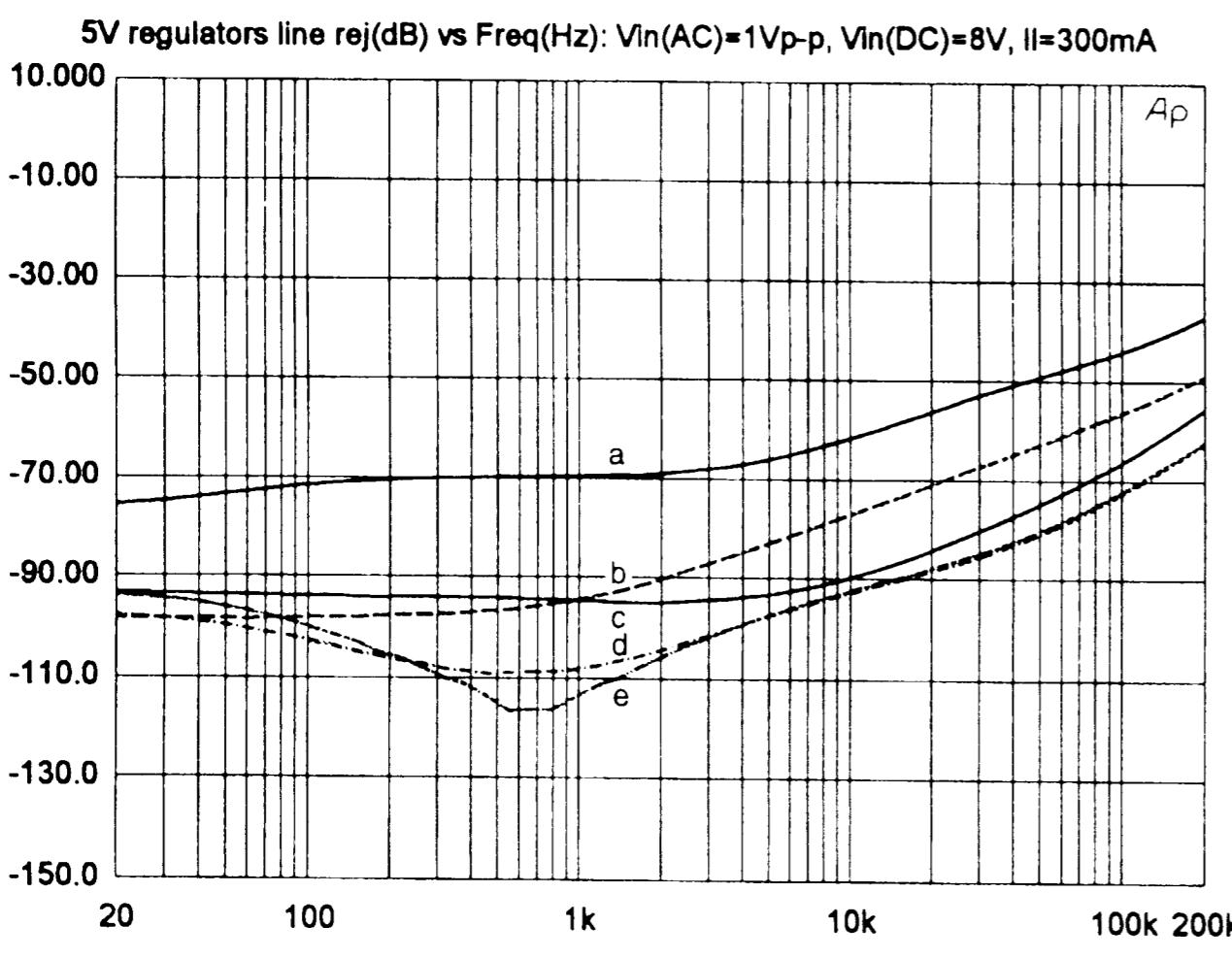


FIGURE 12d: 5V regulator LR, a) 7805CT; b) AD797 (Fig. 9); c) AD848 (Fig. 9); d) AD797 (Fig. 9, bypassed); e) AD848 (Fig. 9, bypassed).

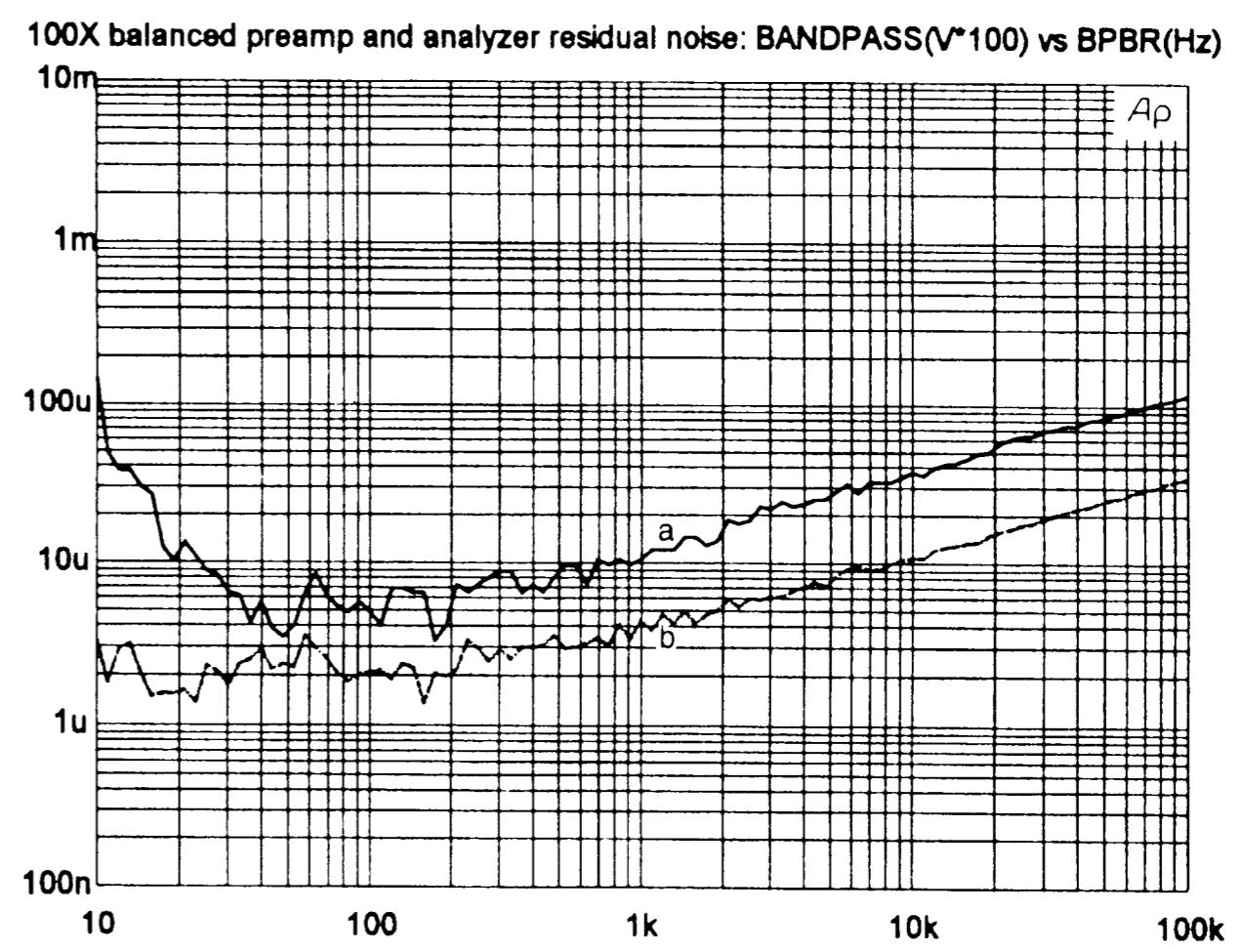


FIGURE 13b: Residual noise results, a) System One $\times 100$; b) preamp.

(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 POOGIE 5.51 regulator (b) are also tested.

Although the POOGIE 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-

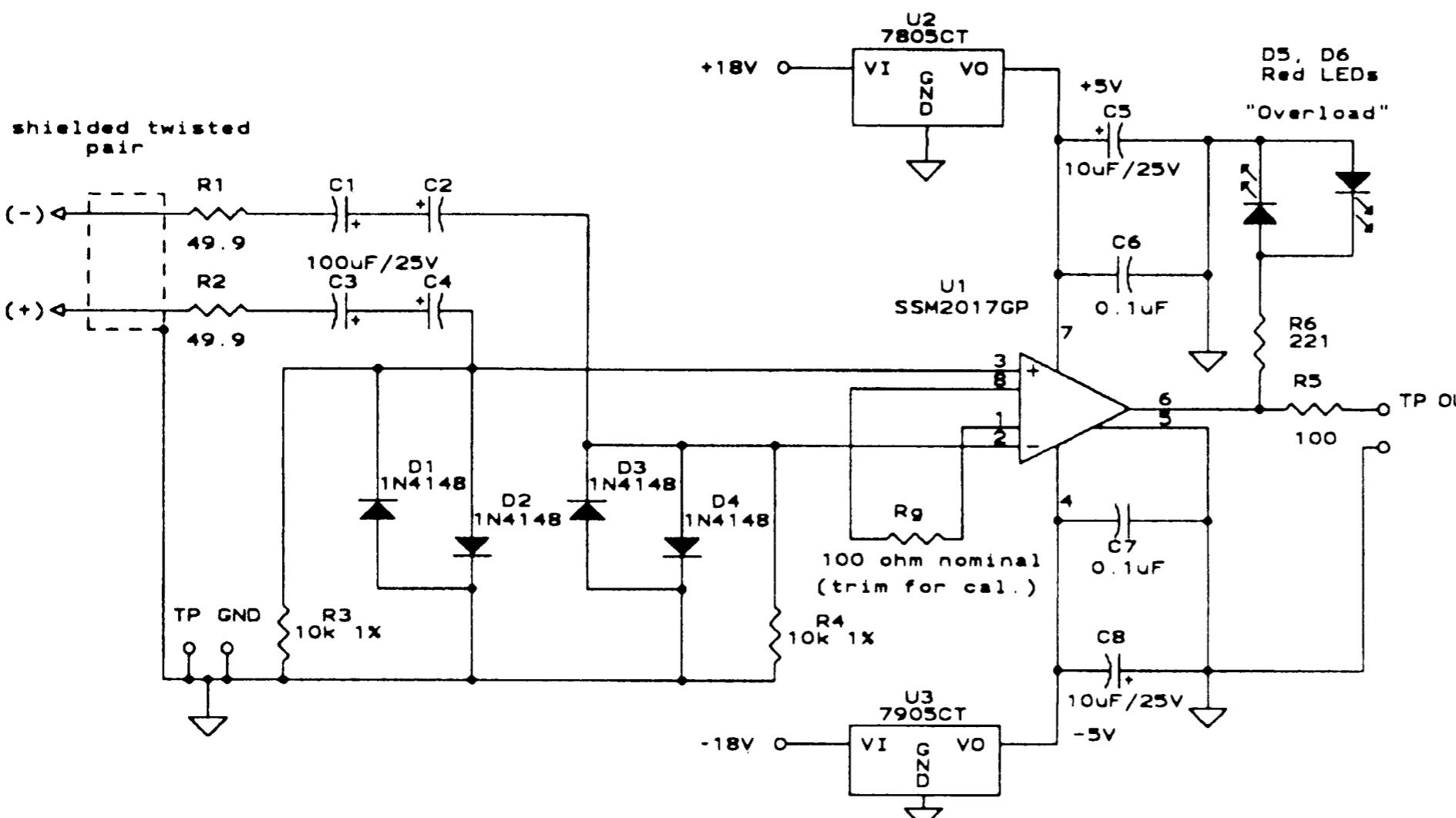


FIGURE 13a: Low-noise 100X balanced input preamp.

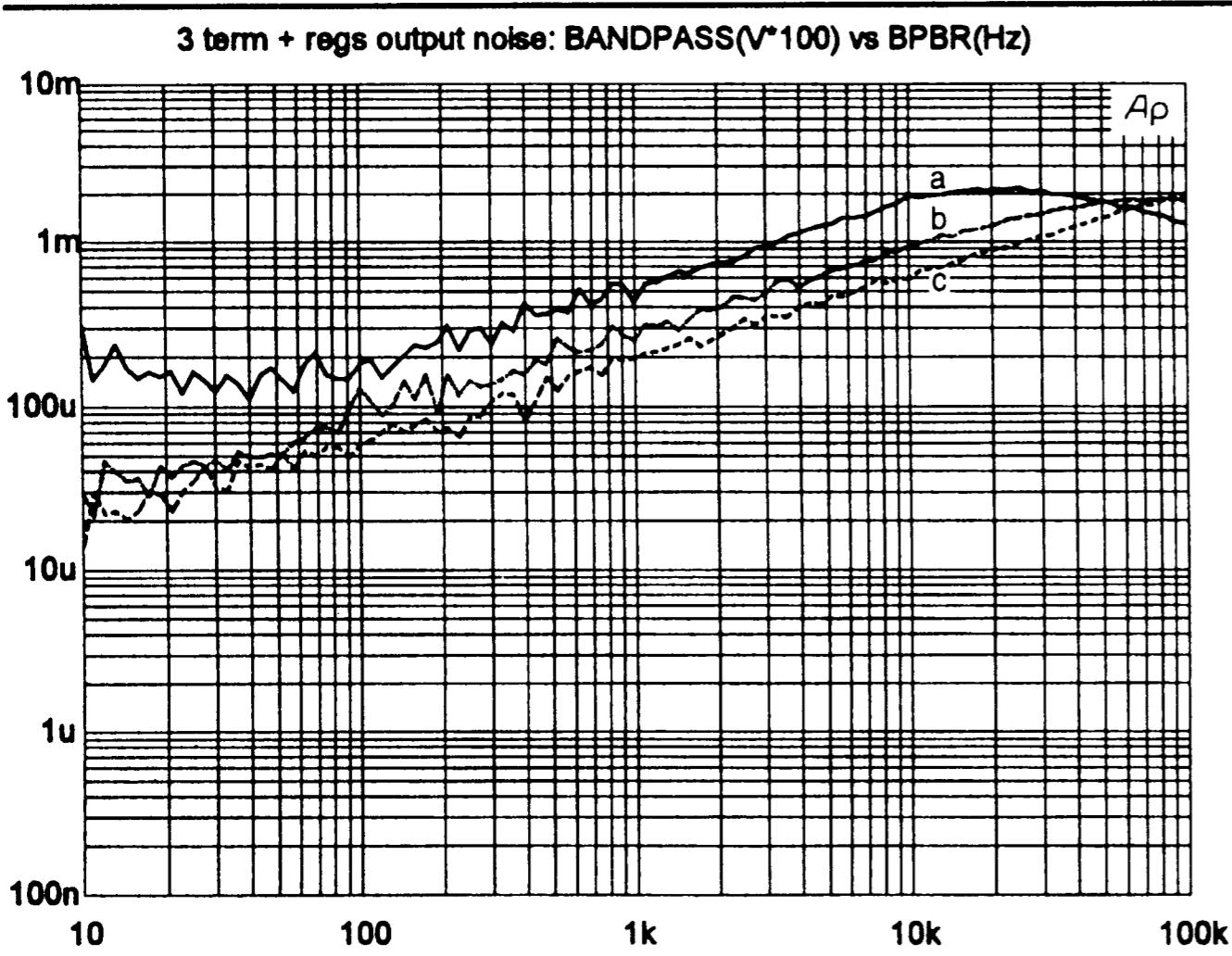


FIGURE 14a: Noise measurements for three-terminal positive regulators, a) 7815; b) 317; c) 1085.

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 $10\text{nV}/\sqrt{\text{Hz}}$ or less, particularly at low frequencies. A gain of 100 balanced input preamp was developed specifically for this purpose (*Fig. 13a*).

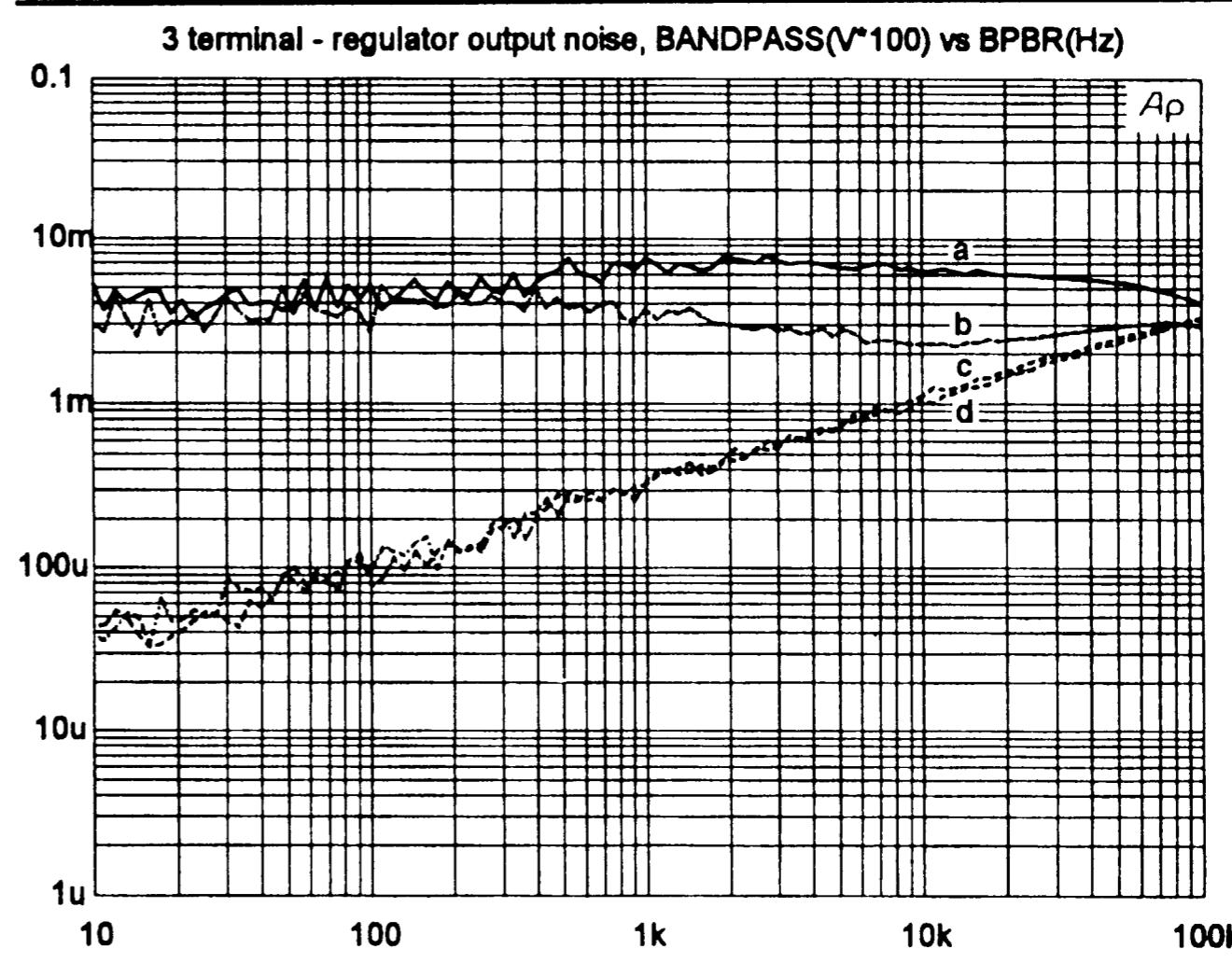


FIGURE 14b: Three-terminal negative regulator noise: 7915 samples—8818 (a) and 46AL (b); c) 337; d) 1033.

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 $10\text{nV}/\sqrt{\text{Hz}}$ might be measured at 1kHz, $14\text{nV}/\sqrt{\text{Hz}}$ at 2kHz, and so forth. Some subsequent noise test examples illustrate this phenomenon.

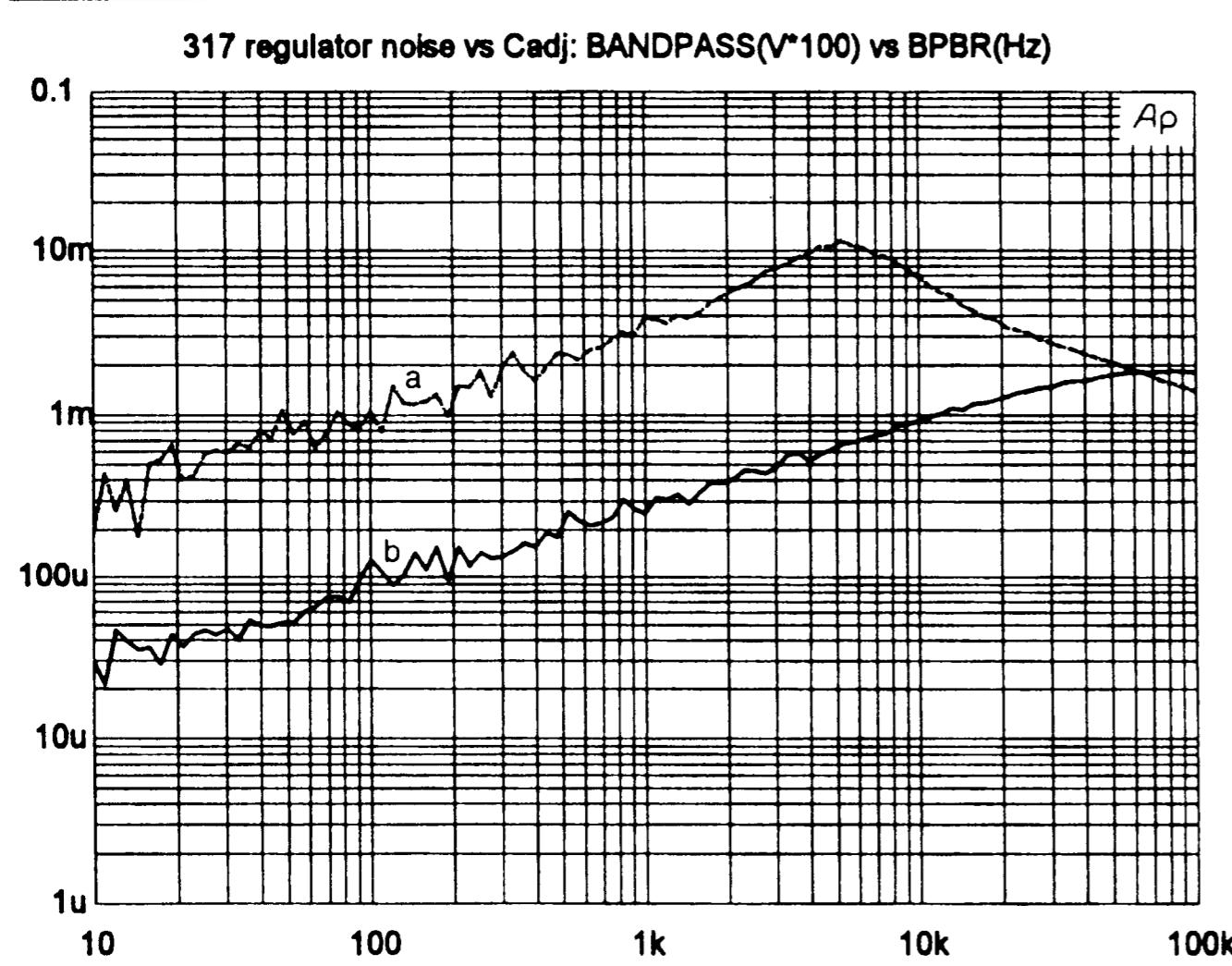


FIGURE 15: 317 regulator noise vs C_{ADJ} , a) $C_{ADJ} = 0$; b) $C_{ADJ} = 100\mu F$.

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 $100\sqrt{0.2316 \cdot F}$. (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 $\approx 1,518$, 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 $4\mu V/1,518 \approx 2.6nV/\sqrt{Hz}$.

For the analyzer-alone residual noise displayed in the upper trace (a), the as-measured noise data was scaled by exactly 100x, 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 2-3x at 1kHz, and by higher factors at very low frequencies where the analyzer input noise rises.

Preamplifier 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 1,000x. Here, at a gain of 100x, 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

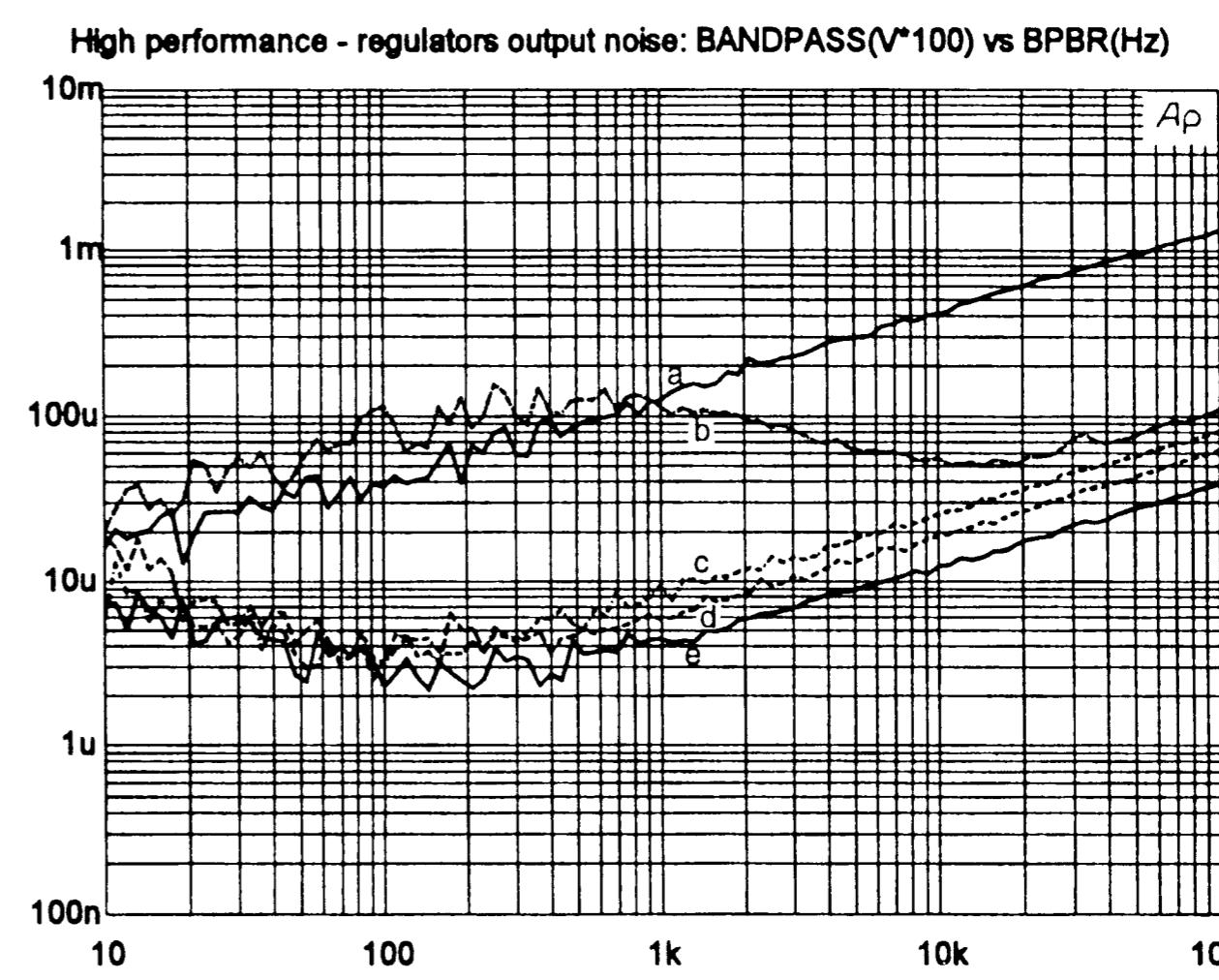


FIGURE 16a: High-performance negative regulator noise, a) un bypassed 329; b) POOGIE 5.51 (Fig. 4); c) AD848 (Fig. 8b); d) Sulzer (Fig. 7b); e) AD797 (Fig. 8b).

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 $\pm 15V$ DC levels. The "Overload" back-to-back LEDs across the output light up when the DC output swings beyond $\pm 1.5V$, 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 $\pm 0.5mV$.

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 12x 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

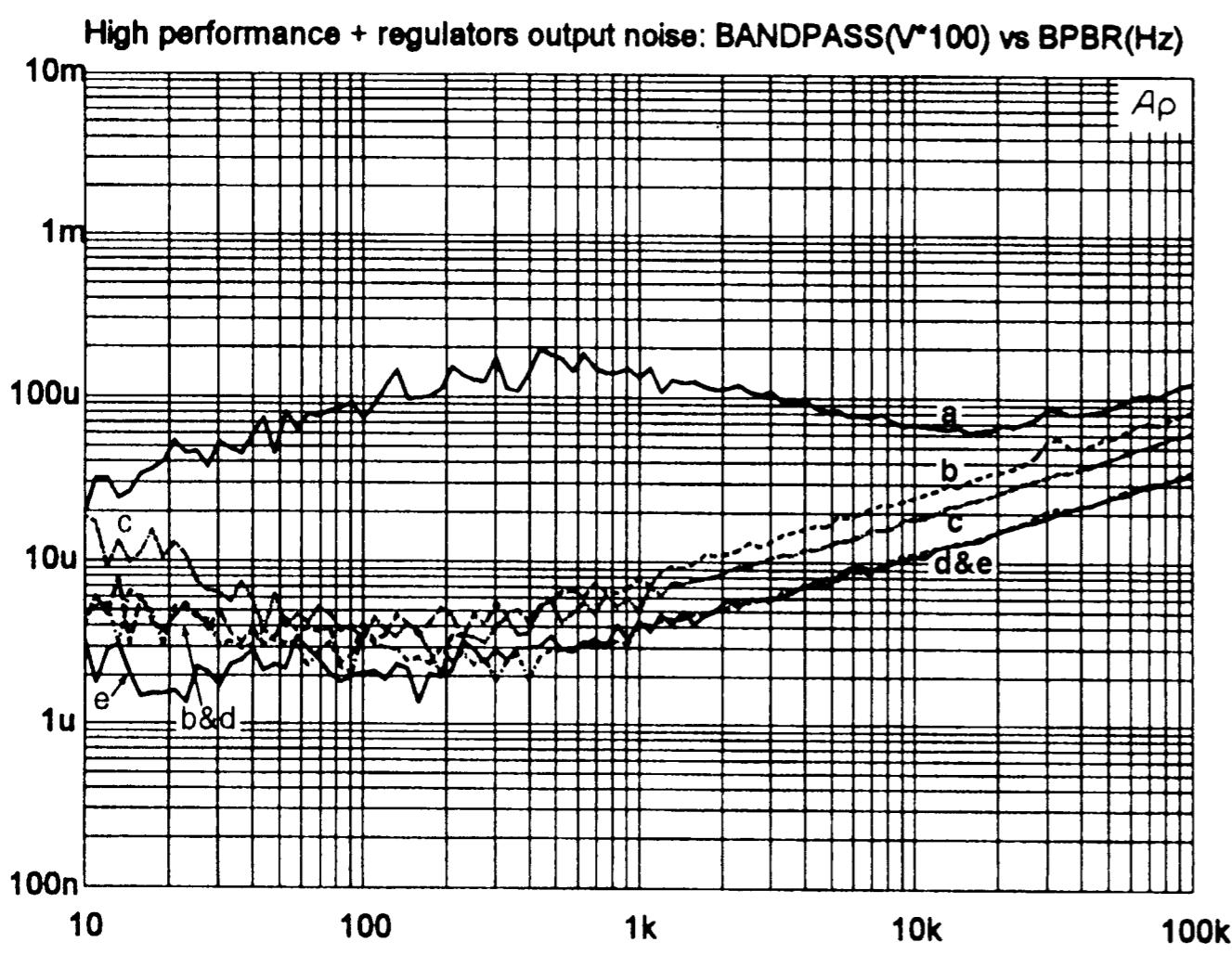


FIGURE 16b: High-performance positive regulator noise, a) POOGE 5.51 (Fig. 3); b) AD848 (Fig. 8a); c) Sulzer (Fig. 7a); d) AD797 (Fig. 8a); e) residual.

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 $210\text{nV}/\sqrt{\text{Hz}}$, while the two fixed output 7915 samples (a and b) show uncomfortably high 1kHz noise levels of about 4,600 and

$2,300\text{nV}/\sqrt{\text{Hz}}$. 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 $1\mu\text{V}$ to 0.1V .

Figure 15 illustrates that adjustable

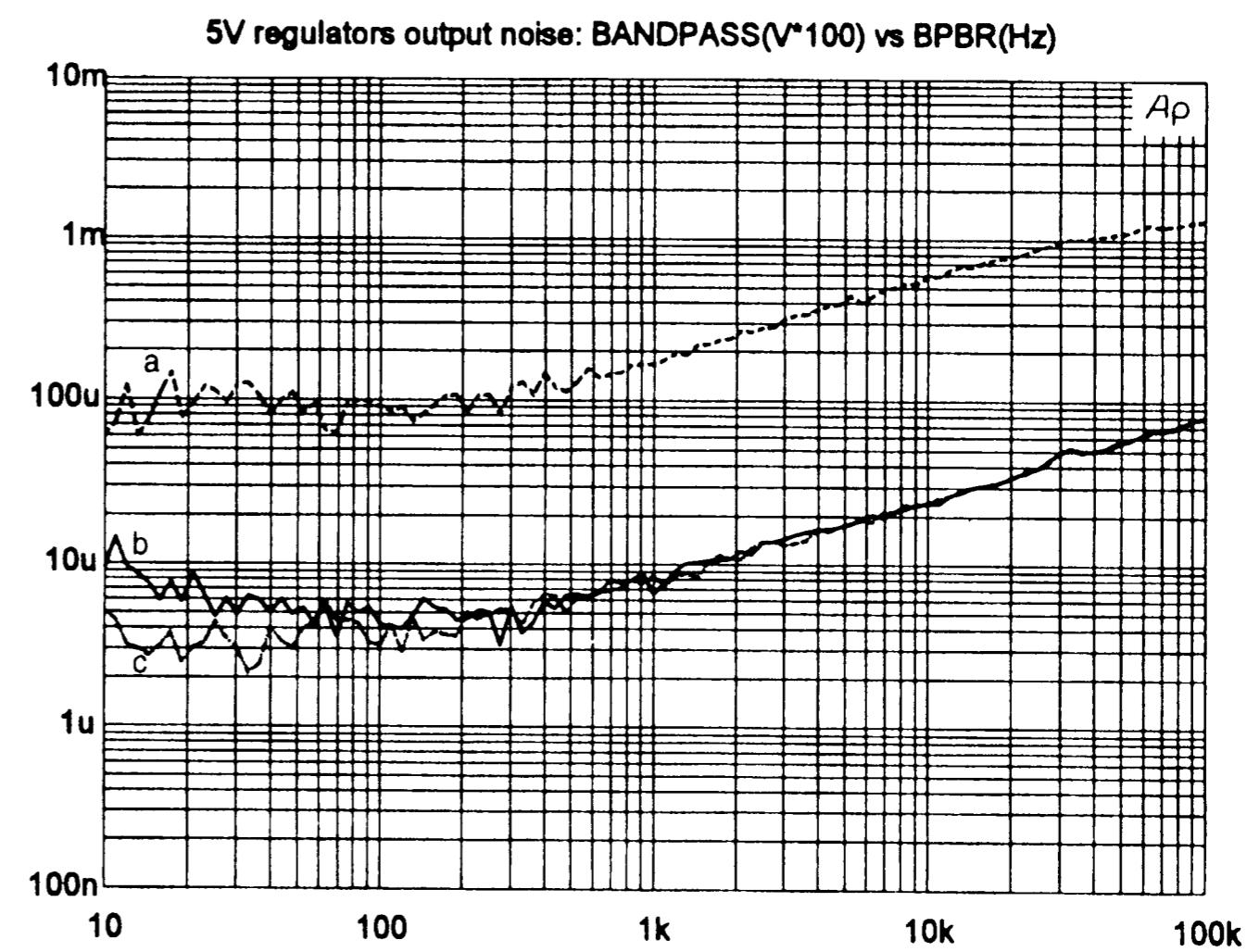


FIGURE 16c: 5V regulator noise, a) 7805CT; b) AD680 (Fig. 9); c) AD780 (Fig. 9).

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 $100\mu\text{F}$ C_{ADJ} capacitor connected. The 1kHz noise levels are about 180 and $2,300\text{nV}/\sqrt{\text{Hz}}$, respectively, roughly corresponding to the $\approx 12\times$ AC gain dif-

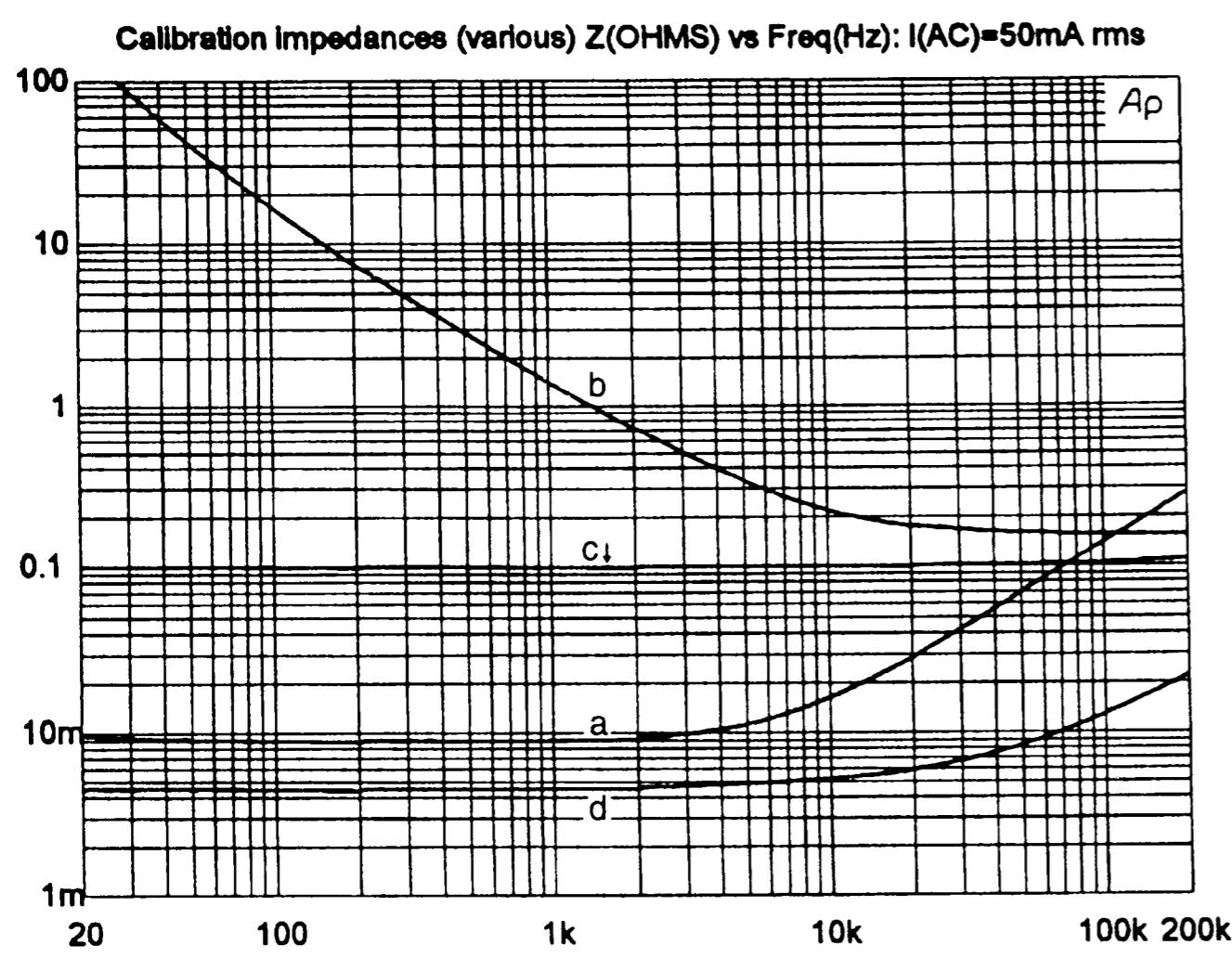


FIGURE 17: Z measurement calibration impedances, a) 12" loop of #18AWG spaced 1"; b) 120 μ F/25V HFQ; c) 0.1 Ω \pm 1%; d) ZEROHM jumper.

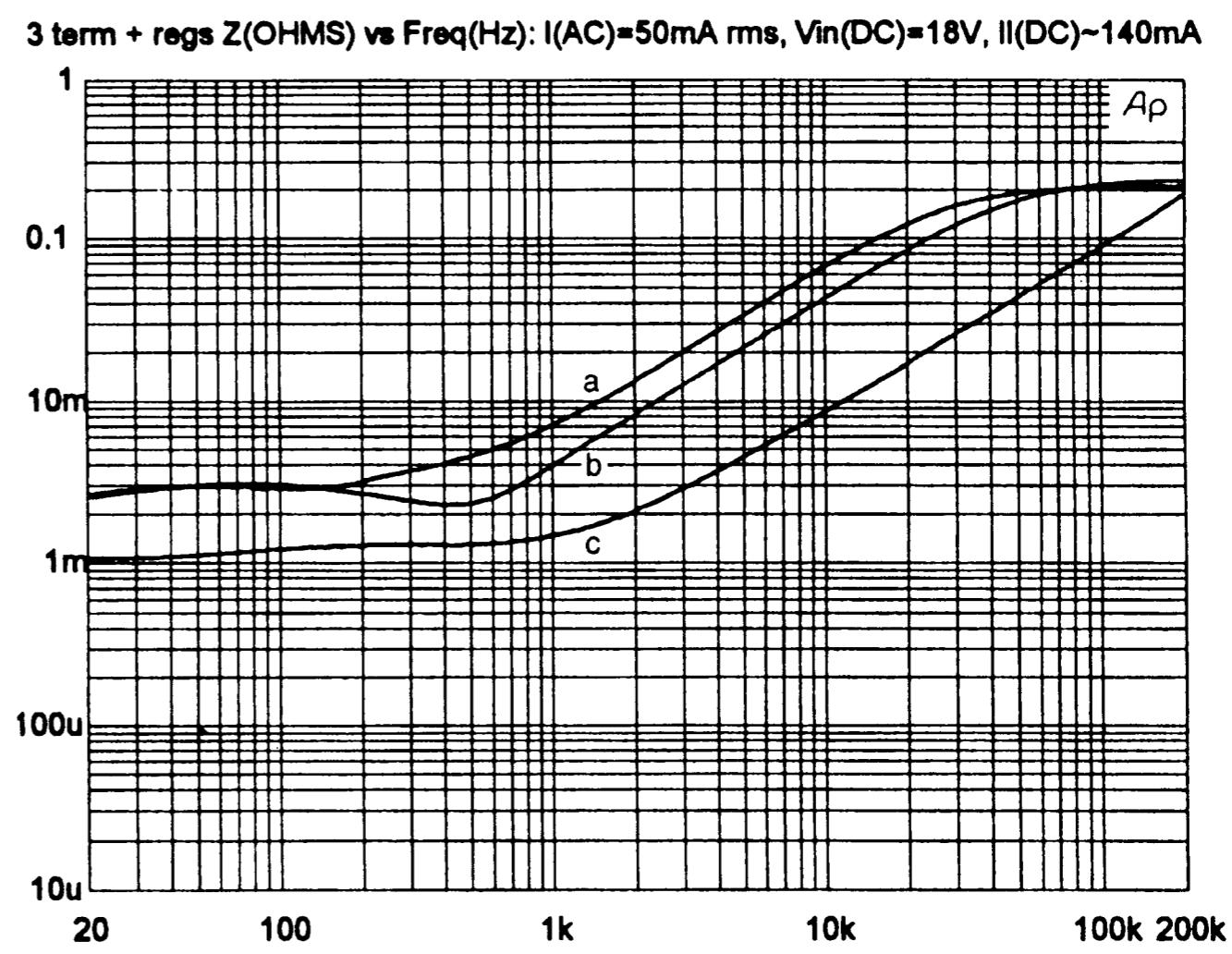


FIGURE 18a: Three-terminal positive regulator Z_O performance, a) 7815; b) 317; c) 1085.

ference mentioned above.

With C_{ADJ} , the regulator's AC gain is reduced essentially to unity, and the noise at the output consists of the basic bandgap reference cell within the device. This data amplifies the point that among the various three-terminal regulators the adjustable units offer best performance when the adjustment pin is bypassed. Note that the scale of this plot also ranges from 1 μ V to 0.1V.

High-Performance Topologies and Noise

A plot of high-performance negative regulator noise is shown in Fig. 16a (note scale change). Before comparing the various regulators, it is useful to explore how topology affects noise results. In this case the topology involves the specifics of how you choose the reference device and interface to the amplifier. These higher-performance regulators use as their basic reference the industry-standard 329 diode, a low-noise, *buried zener* IC, which offers the lowest self-generated noise among the various reference types.⁶ For the typical 6–7V breakdown voltage, the noise is on the order of 100nV/ \sqrt{Hz} .

This may sound like only a factor of two or so below popular 2.5V bandgap-based IC references such as the AD680, but as a percentage of full scale, the buried zener noise is closer to 6–7 \times lower and is thus an optimum choice from a noise standpoint. Further, by carefully controlling filtering and AC scaling within the regulator circuit, you can suppress the noise of the reference diode to negligible proportions.

Figure 16a shows the noise of an

unbypassed 329 diode as operated within the Fig. 4 POOGIE 5.51 regulator (a), which is roughly flat with frequency (considering the 3dB/octave upward slope). The 1kHz noise is just about 100nV/ \sqrt{Hz} , which is typical. Bypassing can reduce the effective diode noise as it appears at a regulator's output, but the degree to which this may be possible depends upon the surrounding circuit impedances.

Since the basic 329 dynamic impedance is less than 1 Ω , bypassing it directly requires rather heroic capacitance values, and even then it will only be effective at high frequencies. In Fig. 4 for example, the reference voltage appears at the error amplifier emitter, AC bypassed by the 560 μ F capacitance with a Z of $\approx 0.25\Omega$ at 1kHz. As a result, bypassing reference noise in this configuration can only be effective above 1kHz, and limited in degree.

The noise output of the Fig. 4 POOGIE 5.51 regulator (b), coincidentally, is also about 100nV/ \sqrt{Hz} at 1kHz. While it may seem odd that the regulator shows a noise level the same as its (unbypassed) internal reference at 1kHz, in this case it is because of the filtering effects. Above 1kHz, the regulator noise output drops considerably due to filtering, while below, the noise rises above that of the reference, as you would expect. While the POOGIE 5.51 design is quite effective for its simplicity, ultra-low noise is not its strongest suit. Nevertheless, the noise is still more than competitive with three-terminal regulators, particularly at high frequencies.

The high-performance op-amp-based

negative regulators (c, d, e) reveal sequentially decreasing output noise. All of the op amp regulator designs can more effectively low-pass-filter the reference noise, because they use additional series resistance between the reference and the amplifier.

This noise filter, for example R1-C4 of Fig. 8b (or 8a), allows you to easily lower the filter corner frequency, in this case to 2.6Hz. It dramatically affects regulator noise, which you can appreciate by comparing any of the three op amp regulator curves to that of the unbypassed 329 curve. For frequencies below 100Hz, the low-pass filter is less effective, which shows up as the rise in low-frequency noise for all these regulators.

Other Performance Factors

Another topological detail which minimizes output noise is the connection of an AC feedback capacitor, such as C3 of Fig. 8a or 8b. This capacitor, used in the original Sulzer regulator to lower output impedance, also decidedly affects noise. Without it, the op amp amplifies its own noise and that of the filtered reference by the output gain factor (V_{OUT} equation of Fig. 1). With it, the op amp amplifies these noises by a factor of unity. For the 5–20V output range regulators, the net difference with/without this capacitor can be 6dB or more, which is important for high performance.

At frequencies of 1kHz or more, the differences in noise generally follow the basic voltage noise characteristic of the op amp under test. For example, the specified voltage noise of the AD848, 5534, and AD797 follows a decreasing

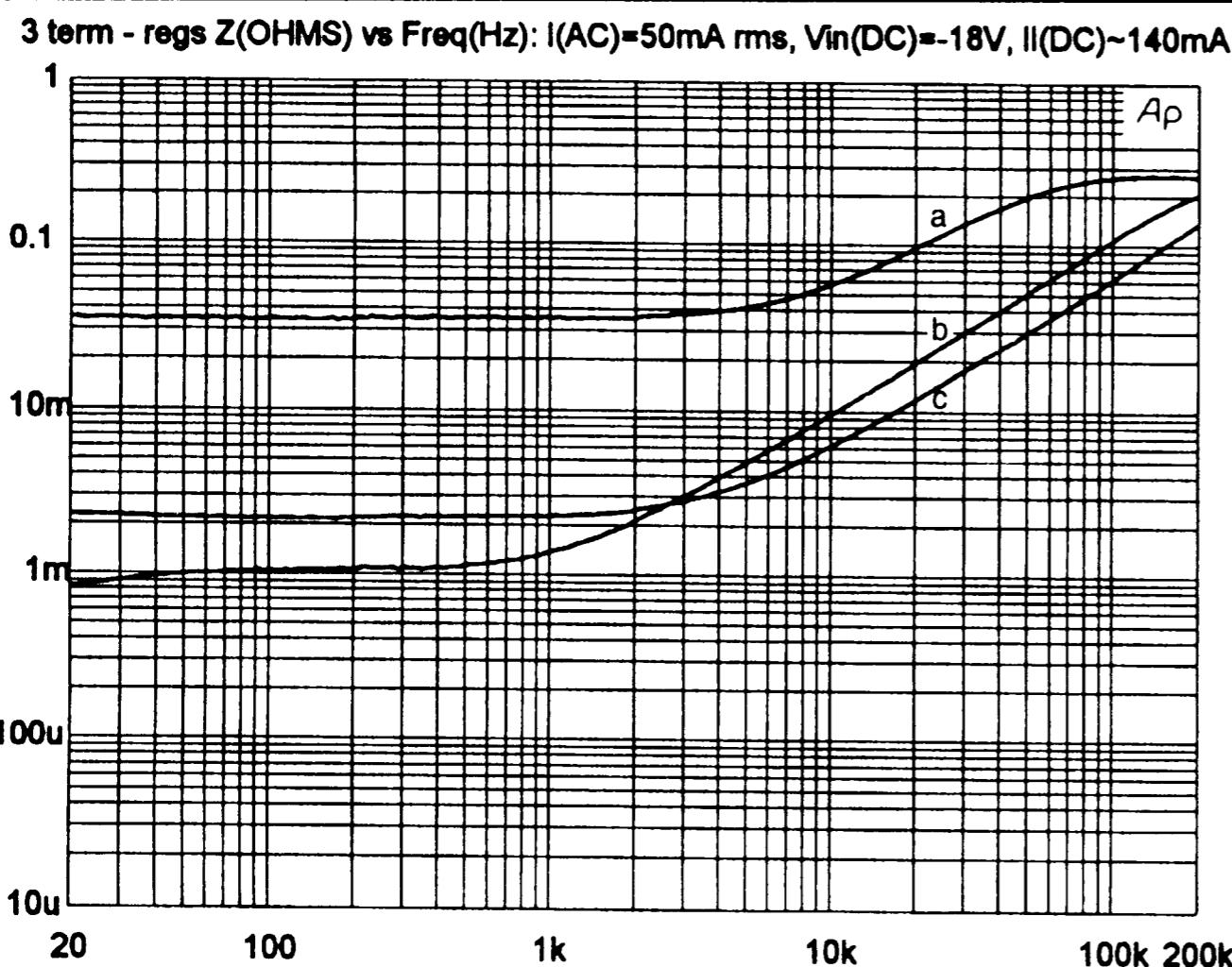


FIGURE 18b: Three-terminal negative regulator Z_O , a) 7915; b) 1033; c) 337.

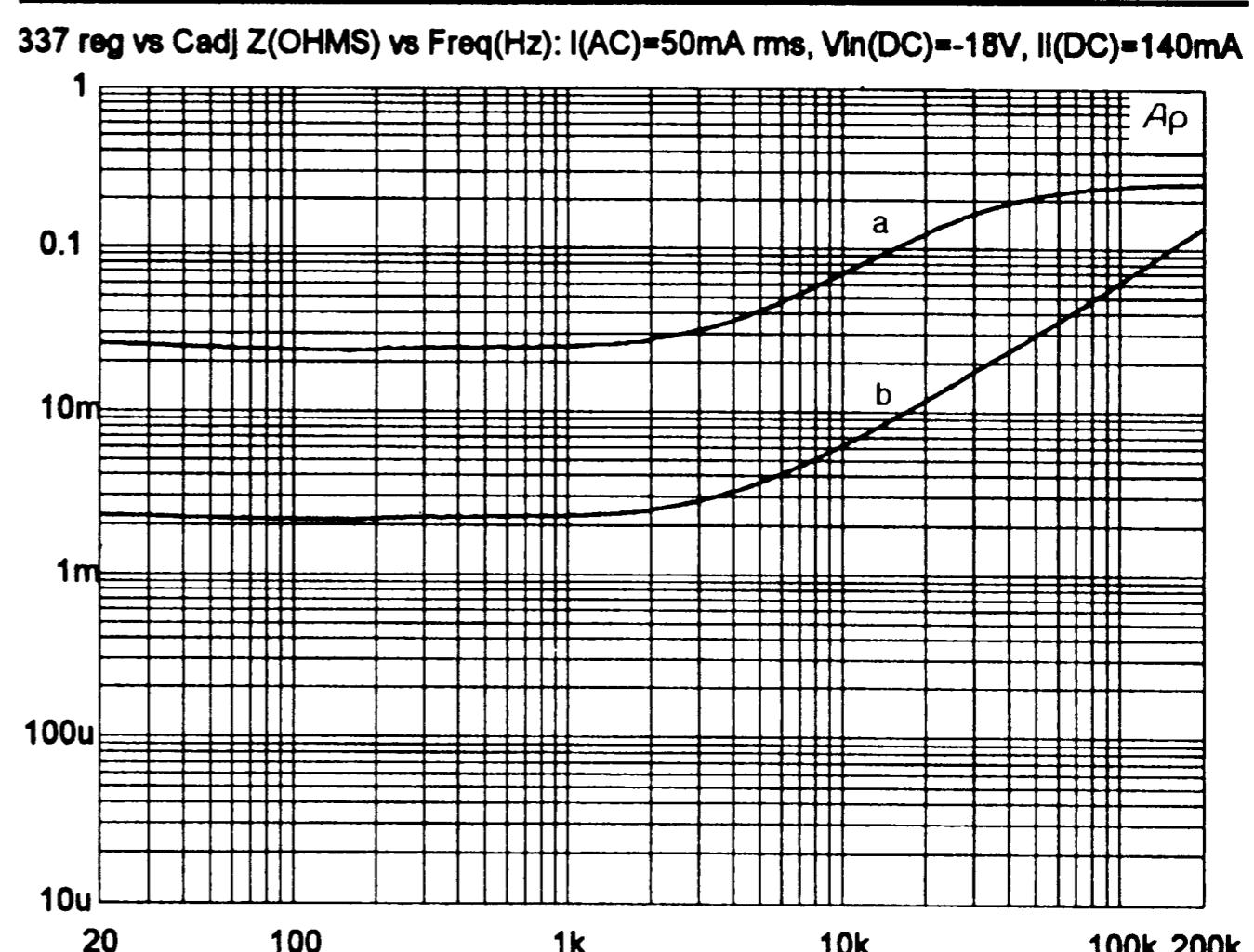


FIGURE 19: 337 regulator Z_O vs C_{ADJ} ; a) $C_{ADJ} = 0$; b) $C_{ADJ} = 100\mu F$.

sequence, which is also depicted by the three noise curves, representing the noise outputs of Fig. 8b with the AD848 (c) and AD797 (e), and Fig. 7b with the 5534 (d).

However, since the 1kHz noise level for any of these regulator designs is below $10nV/\sqrt{Hz}$, it begins to push the resolving power of the noise test preamp (Fig. 13a). As a result, the actual noise levels, which appear to be about 6, 4, and $2.8nV/\sqrt{Hz}$, respectively, are even lower. This is due to the $\approx 2.6nV/\sqrt{Hz}$ preamp noise, which sets a lower limit to noise resolution. However, their *relative* order is valid as shown.

Figure 16b shows a plot of high-performance positive regulator noise, which generally follows the pattern established for the negative regulators of Fig. 16a, with some variations in the POOGIE 5.51 regulator of Fig. 3 (a) and the Sulzer regulator of Fig. 7a (c). The two variations of the Fig. 8a circuit with the AD848 and AD797 (b and d) are a bit better than the Sulzer circuit of Fig. 7a below 100Hz.

In the sub-100Hz range, however, final noise performance of these op amp regulators can be a function of the filtering as well as the op amp (and the setup, of course). The Fig. 8a circuit at these frequencies is more limited by the filter than by the op amp, since the AD848 and AD797 curves tend to merge, even though the two devices have a roughly 5/1 different voltage noise.

Finally, a trace representing the setup residual noise (e) is added to this plot, which closely parallels the AD797's curve above 1kHz. This indicates that for the AD797 at least, the true regulator

noise is less than this setup can resolve at the high frequencies.

5V Regulator

As I mentioned earlier, the reference device itself can limit noise, even in instances of low-pass filtering. While buried zener reference types offer the lowest inherent noise, they are not always available to an application, since they must operate from at least $\geq 7V$. For a 5V regulator with low dropout, a low-voltage bandgap reference is therefore your only choice. The logic regulator of Fig. 9 has evolved from the regulator of reference 15 and operates with a three-terminal 2.5V bandgap reference at U2, either an AD780 or an AD680. The AD780 has the lower output noise, $\approx 100nV/\sqrt{Hz}$ versus $\approx 250nV/\sqrt{Hz}$.

You can use either IC in U2 position of the Fig. 9 circuit, and Fig. 16c shows the noise output from this circuit as a function of the reference IC in (b) and (c). For this test the op amp is the AD848, and for frequencies above 1kHz, the noise is comparable to that of the AD848 in the regulators of Figs. 8a and 8b, independent of the reference. While below 100Hz the AD780 is the quieter of the two U2 devices, it is unlikely that low-level variations of 20Hz noise in a logic regulator will have serious audible consequences.

Figure 16c also shows the noise of the 7805CT 5V three-terminal regulator (a). The 1kHz measurement of the Fig. 9 circuit using the AD848 op amp is about 20x better than the $\approx 120nV/\sqrt{Hz}$ 1kHz reading of the 7805. When the noise of the measurement preamp is accounted for, the 1kHz of the Fig. 9 circuit is more

like $\approx 5nV/\sqrt{Hz}$ with the AD848, and optionally lower yet with the AD797.

Z_O Tests

Output impedance tests for regulators are among the more important and identifiable. In general, the lower the Z_O the higher the regulator's quality. These

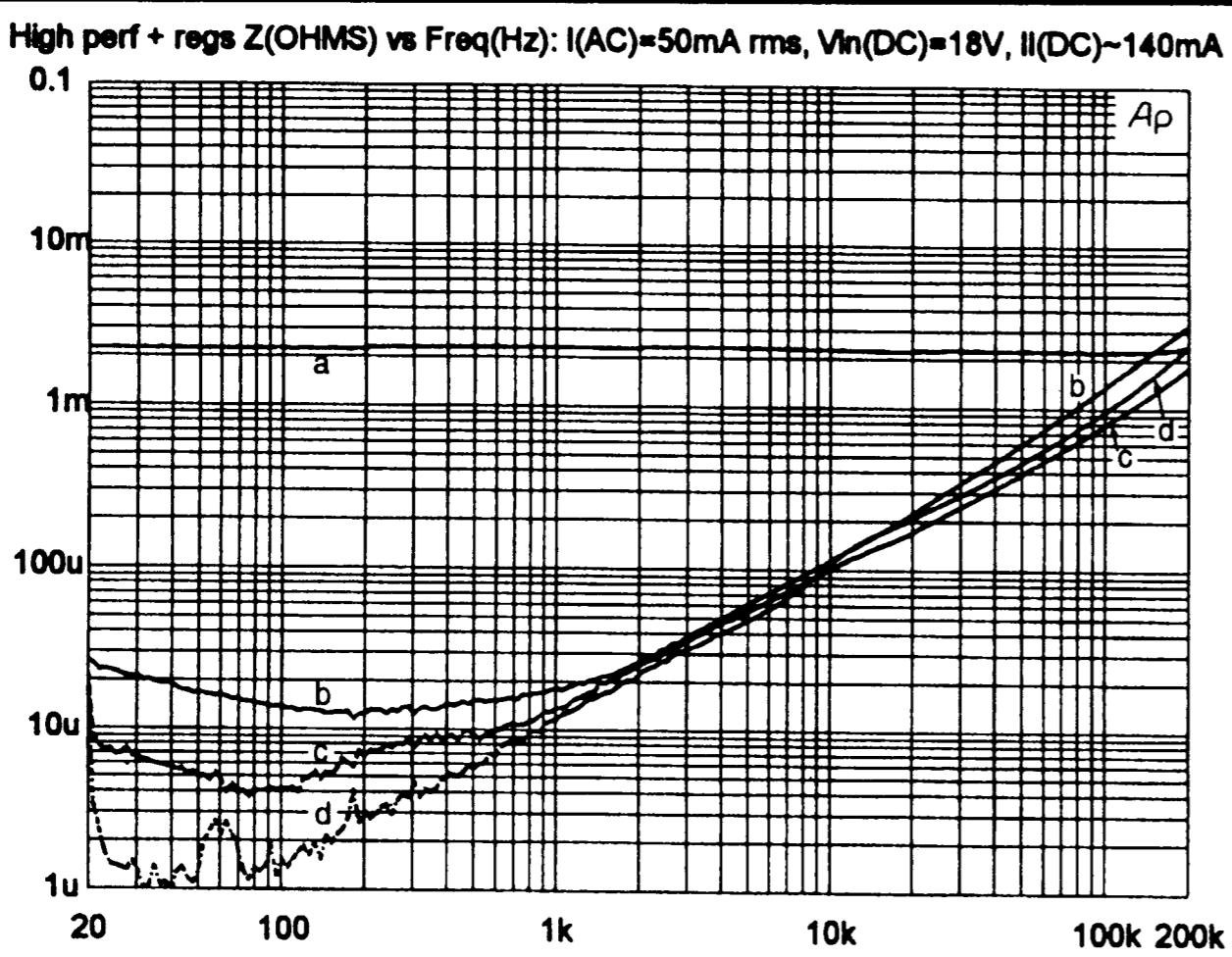


FIGURE 20a: High-performance positive regulator Z_O , a) POOGIE 5.51 (Fig. 3); b) Sulzer (Fig. 7a); c) AD848 (Fig. 8a); d) AD797 (Fig. 8a).

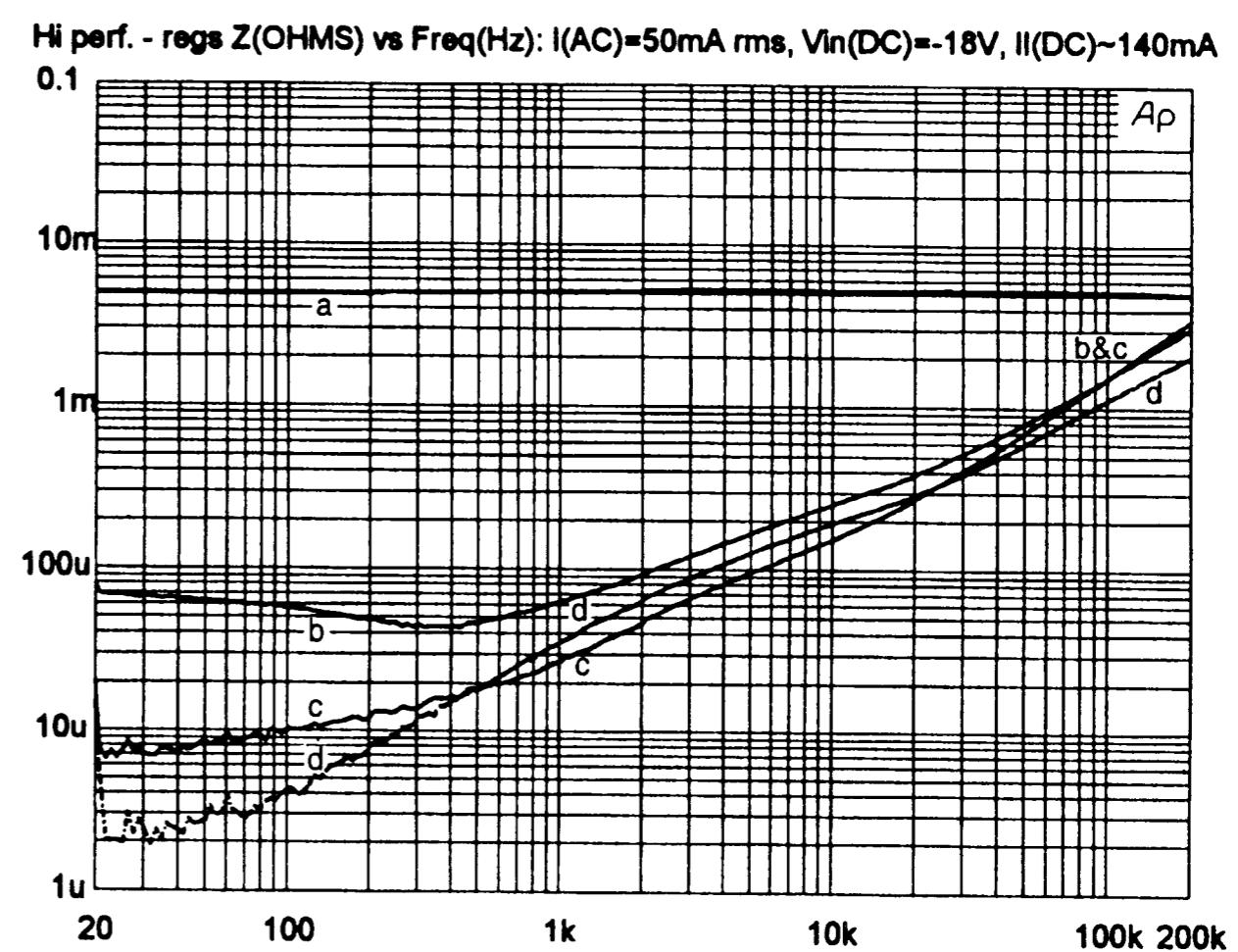


FIGURE 20b: High-performance negative regulator Z_O , a) POOGIE 5.51 (Fig. 4); b) AD848 (Fig. 8b); c) Sulzer (Fig. 7b); d) AD797 (Fig. 8b).

tests generated Z_O data for easy comparison by regulator type.

Considerable effort was spent developing the test techniques and hardware to allow high-resolution Z_O measurement over a wide range of frequencies. In the interest of setup calibration and an overall Z_O sanity check, various component impedances were plotted (Fig. 17) and illustrate the sensitivity of the test setup over the working frequency range of 20Hz–200kHz.

First, a $0.1\Omega \pm 1\%$ resistor is shown in the flat trace (c), which demonstrates just a hint of inductance towards 200kHz. A $120\mu\text{F}/25\text{V}$ -type HFQ capacitor (b) follows a capacitor-like Z fall until about 10kHz, where it then levels off, exhibiting an ESR-dominated Z of about 0.15Ω at 100kHz.

ACKNOWLEDGMENTS

Along the course of this project I was fortunate to have help in various forms from numerous parties, all much appreciated and gratefully acknowledged. Among those who replicated one or more of the regulator circuits were Paul Kelly, Rick Miller, Glenn Moore, Courtenay Osborne, Richard Pell, Phil Schniter, Don Spangler, Karl Youtsey, and Hank Zumbahlen. Hampton Childress's work on the POOGIE 5 project and its regulator was also a meaningful early contribution. I received helpful comments on the manuscript and circuits from Erno Borbely, Hampton Childress, Clarke Greene, Bruce Hofer, Kal Rubinson, Phil Schniter, Scott Wurcer, and Karl Youtsey. For their detailed comments which improved the manuscript, special thanks to Jan Didden, Gary Galo, and Rick Miller.

I could not contact Mike Sulzer during the article's preparation, but I acknowledge his pioneering regulator work. Rick Miller worked very closely with me on many phases of these as well as earlier discrete regulator developments, and also identified the high current D4X series transistors first suggested in reference 9 and used here. I especially appreciate his many helpful comments and audio enthusiasm. Finally, it has been a pleasure working with Jan Didden and Gary Galo on this article series, and I hope the implementation of its concepts will also be a pleasure of yours.

The remaining two curves quite graphically illustrate the practicalities of low- Z wiring. The trace for the 12" loop of #18AWG hookup wire spaced apart 1" (a) shows a Z of about $10\text{m}\Omega$ below 1kHz, rising to about $150\text{m}\Omega$ at 100kHz. The fourth trace (d) represents a minimum-length ZEROHM jumper,¹⁶ connected between the analyzer sensing points with a minimum lead length. As this data shows, even this *single* minimal-length conductor has about $4\text{--}5\text{m}\Omega$ of DCR, and begins to show inductive effects above 10kHz.

This plot illustrates that, even within the audio band, just relatively short conductors of ordinary size show both resistive and inductive components. In other words, don't consider wire as simply DC resistance alone when placed in the context of practical regulation. For regulation impedances below a few milliohms, the distributed wiring can dominate impedance very quickly. This underscores the virtues of the star power distribution scheme outlined in Fig. 1, but more so, the value of remote sensing where applicable (a topic discussed in the next part of this series).

Low Impedance

In the following regulator Z_O data, most measurements fall below $100\text{m}\Omega$, and for display purposes the vertical scaling is 5 decades, with consistent scaling where possible (for ease of comparison). In the three-terminal positive regulator group's Z_O performance (Fig. 18a), an option for 4-wire load connection is not possible, so you should use short heavy leads for good regulation, particularly from the regulator's V_{OUT} pin to the load. In these tests, the measurement connects directly to the IC's output pin.

All three ICs show a reasonably low $1\text{--}3\text{m}\Omega$ Z_O at low frequencies, but impedance starts to rise around 1kHz. The broad flattening off in Z_O around 100kHz is likely due to the $100\mu\text{F}$ output capacitor.

In the three-terminal negative regulator group's Z_O performance (Fig. 18b), the two adjustable types (b) and (c) perform far superior to the 7915 fixed regulator (a) at all frequencies, achieving $1\text{--}2\text{m}\Omega$ at low frequencies. However, they (like the positive three-terminal types) also show a rise in impedance above 1kHz.

As with noise and LR, you can optimize Z_O performance in three-terminal regulators by adjust pin bypassing. Figure 19 shows Z_O data for the 337 adjustable regulator with/without a $100\mu\text{F}$ adjustment bypass capacitor. Without this measure, the device Z_O is more than an order of magnitude higher at low frequencies for these test conditions (a).

In the plot of high-performance positive regulator Z_O data (Fig. 20a), note the vertical scale shift, which now ranges $1\mu\Omega\text{--}0.1\Omega$. You'll also see that the high-performance regulators, in particular the op-amp-based designs, can lower output impedance well below $1\text{m}\Omega$, or orders of magnitude below ordinary wire limited Z (Fig. 17).

This data, taken on carefully constructed layouts with measurements at the sensing divider, may appear overly optimistic, or even nonrelevant from a pessimistic viewpoint. But, the high sensitivity of the bandpass measurement Z_O technique allows such extremely low impedances to be resolved, sometimes revealing subtle differences between op amps applied in

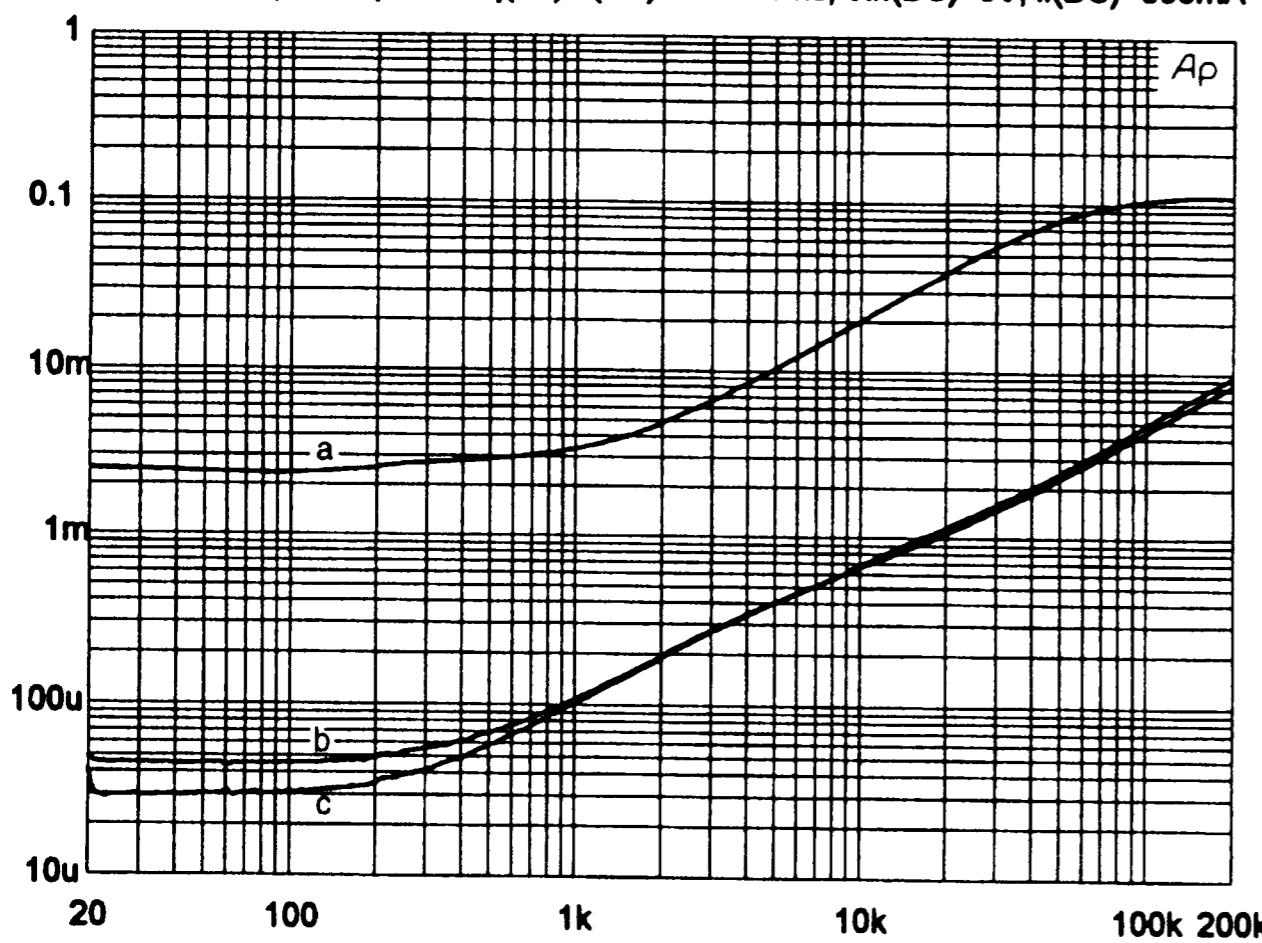


FIGURE 20c: 5V regulator Z_O , a) 7805; b) AD848 (Fig. 9); c) AD797 (Fig. 9).

the same circuit. More important, however, is the factor that the lower impedances *are* relevant, from a power distribution *quality* point-of-view.

Here quality implies both the magnitude of impedance, plus how well it is maintained over frequency. The lower the regulation impedance is at the sense point, the better a regulator can effectively short out current-proportional crosstalk from multiple load stages. This justification assumes that the power system has been wired with a star distribution scheme (Fig. 1). On the other hand, "daisy chain" power distribution may defeat virtually everything gained from high-quality regulation, by allowing the inevitable finite impedances of the power distribution system to build up crosstalk-related voltage drops at the power tap points.

More Data Analysis

It is interesting to note that the POOGIE 5.51 regulator of Fig. 3 (a) has a virtually flat $\approx 2\text{m}\Omega$ Z_O over the entire range, indicating its wide bandwidth. This modest regulator's Z_O is equal or superior to a ZEROHM jumper, suggesting that even a simple circuit can benefit from the star power distribution scheme.

With the Figs. 7a and 8a op-amp-based regulators, the low-frequency Z_O can go as low as $10\mu\Omega$ or less, depending on the specific device and/or circuit. The extremely high gain AD797 operating in the Fig. 8a circuit can produce a low-frequency equivalent Z_O of close to $1\mu\Omega$ (d), which, of course, in practice will be quite environment-dependent. The AD848 in the same circuit does nearly as well (c), with the Sulzer circuit of Fig. 7a not far behind (b).

As a practical matter, you should

TABLE 1

PROGRAMMING RESISTOR VALUES OF FIGURES 8A AND 8B FOR VARIOUS OUTPUT VOLTAGES

V _{OUT} (V, nominal)	R _{4/R₃} (ideal)	R ₁ (Ω)	R ₃₍₁₎ (Ω)	R ₄₍₁₎ (Ω)	R ₆ (Ω)	V _{OUT(2)} (V, actual)
10	0.449	499	1.58k	715	2.21k	10.022
12	0.739	499	1.18k	866	3.65k	11.964
13.8 (as shown)	1	499	1k	1k	4.99k	13.800
15	1.174	499	931	1.1k	5.90k	15.053
16	1.319	499	866	1.15k	6.49k	16.063
18	1.609	499	806	1.3k	7.87k	18.029

(1) Closest 1% values.

(2) Assuming 6.9V reference.

carefully consider impedances below about $100\mu\Omega$, making comparisons only under

the most controlled conditions, i.e., changing only an op amp in an otherwise identical circuit. With regard to this data, for example, you can directly compare AD848 and AD797, since they both have been operated within the same circuit. Comparison with the data of the Sulzer Fig. 7a circuit is somewhat less precise, since it is physically and electrically different.

Under the stated measurement conditions, all of the op-amp-based regulators can achieve high-frequency Z_O performance on the order of $1\text{m}\Omega$ at 100kHz, a feat which is by no means trivial. Key to this is a tight, compact, and low-inductance layout.

As with their positive counterparts, negative regulators (Fig. 20b) can also achieve output impedances well below $1\text{m}\Omega$. Also as before, the POOGIE 5.51 regulator of Fig. 4 is quite flat in impedance (a), at about $5\text{m}\Omega$ (albeit higher than the positive version, for reasons not obvious).

In the op-amp-based negative regulators, all circuits do very well at low frequencies, to below $100\mu\Omega$ for frequencies below 1kHz, with the AD797 in the Fig. 8b circuit the lowest of all, again

pushing $1\mu\Omega$ (d). At high frequencies all three op amp regulators are again close to $1\text{m}\Omega$ at 100kHz (b, c, d).

A plot of 5V logic regulator Z_O data (Fig. 20c) includes performance for the AD797 and AD848 in the Fig. 9 circuit, as well as that of a 7805CT for reference. Note the vertical scale shift, which now ranges from $10\mu\Omega$ to 1Ω .

The op-amp-based 5V logic regulators in this circuit do not perform quite as well as in the Fig. 8a circuit. In the audio band the AD848 and AD797 traces (b and c) achieve a Z_O of $1\text{m}\Omega$ or less, but at 100kHz they rise to 4–5m Ω .

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The standard 7805CT TO-220 regulator (a) achieves a low-frequency impedance of about $2\text{-}3\text{m}\Omega$, but at higher frequencies the Z_O rises to about 0.1Ω , and at all frequencies the 7805 Z_O is an order of magnitude or more higher than that of either Fig. 9 regulator variation.

Field Tests

Many people acknowledged below replicated the new $\pm 14\text{V}$ or $+5\text{V}$ logic regulator in one form or another, within various items of equipment. While in general these efforts were successful, an earlier circuit using a custom-compensated AD829 op amp exhibited some instances of instability. This factor was largely why I changed to the AD848 late in the project cycle. The AD848 is an uncompensated device, and, importantly, one which has shown none of the sometimes maverick tendencies of the AD829, either in the various equipment installations or in bench testing.

My experiences with the latest AD848-based Figs. 8a and 8b circuits have been positive, in both a preamp and DAC960 D/A converter, plus logic supply AD797 and AD848 circuits, also successfully applied in this same D/A. Earlier versions of the final Fig. 9 circuit

using an AD829 were used in a logic application with the AD1890 evaluation board,¹⁷ and in a modification to an Audio Alchemy D/A converter.¹⁸ In addition, Hank Zumbahlen has used a version of the Fig. 8a/8b regulators in a D/A design for an upcoming application note.

"Successful" here means that the circuits installed without mishap and worked according to theory—in other words, they behaved as they should. My prototype tests all were "kludge-card" versions implemented on sections of Radio Shack general-purpose PC board #276-168 (one of these cards can support a pair of the regulators). I recommend the Jan Didden PC board design as the best implementation choice, since it has an integral ground plane and a good layout. See Jan's article (Part 3) and Gary Galo's follow-up on its practical installation and operation (Part 4).

Summary, Sundries, and Hints

Summarizing an article of this scope is a difficult task, but various highlights do emerge. One is the fact that, not unexpectedly, the more complex regulators of Figs. 8a, 8b, and 9 yield the highest overall performance. This article includes the

performance data for all the various regulators, which allows ready comparisons of relative quality for the three measurement dimensions (along with a fourth already mentioned, complexity).

This information allows the designer to choose a circuit type appropriate to a given quality level. I hope to leave the reader with the impression that these more complex circuits deliver performance worth their price and the effort of implementation, as they certainly do. Some users have already reported so, as noted above.^{17,18}

Proponents often prefer *shunt regulators* for audio applications, and might ask why these new regulators don't operate in a shunt mode. One reason given for using a shunt regulator is that it can both source and sink output current, whereas a series regulator (such as these) can only source current. Without getting into the pros and cons of the (major) efficiency and LR differences between series and shunt designs, it is true that a series type regulator can be optionally operated with a high bleed current.

Dick Marsh covered this factor and other worthwhile points on system aspects of power supplies.¹⁹ This rather

easily implemented trick then allows the series regulator to both source *and* sink output current, with a sink capability up to the level of the bleed current. The test condition examples discussed here used a relatively high bleeder current of 100–140mA, but obviously you can use lower levels.

- A general point to be made is that Figs. 8a and 8b are *optimized* circuits, and as such, you should think long and hard before attempting to “tweak” them—they’ve already been tweaked! Don’t change the op amp from an AD848 or AD797, unless you don’t care about retaining all aspects of performance (if you don’t care about full performance, why build it at all?). Besides the degradation likely with other devices, there is also the real possibility of oscillation. The passive parts are much less critical, but again, why run a risk of degradation?

The selection between the AD848 and AD797 is best made in terms of polarity, dependent upon the specific application. While a good all-around choice, the AD848 has lower LR at low frequencies, particularly for negative operation. The AD797 is also preferable for lowest noise and highest DC stability (see below,

although the latter of these points may be moot for audio uses).

- A subtlety which may or may not be obvious is that these regulator circuits operate the op amp in a *single supply mode*. By definition, this eliminates noise from the opposite polarity unregulated supply, and also aids in start-up to the desired output state. For these reasons, I do not recommend that you operate the Fig. 8 circuits from dual unregulated supplies.

- The low dropout feature of the new circuits can be optionally disabled, by deleting the current source components and the diode in series with R2 (for stability, R2 should be retained in any case). However, you should carefully consider before you exercise this option. It simplifies the circuit, but at the expense of higher dropout, obviously. A more subtle point of this mode of operation is that it might sacrifice the robust startup characteristic provided by the current source (which tends to automatically force the circuit to the desired output state).

- A major caveat is to watch the capacitive bypassing details at the regulator output. Note that in Figs. 8a, 8b, and 9 there is no output film capacitor shown, which is done purposely. High-Q (low

ESL/ESR) capacitors such as stacked films or ceramics should *not* be connected directly across the regulator output! This caveat is nonnegotiable because of the wide regulator bandwidth, which reacts negatively with the rapid phase changes produced by a high-Q cap load, and as a result oscillates merrily. Film capacitor types, if used, are better applied right at the load decoupled by low DCR chokes or ferrite beads, or even just normal wiring of a few inches or more. The electrolytic bypass C5 is desirable, but its value is not highly critical.

- DC stability of the Figs. 8a and 8b circuits can be made quite high if desired, even though this is an unlikely audio priority. Using an AD797 amplifier with R1 and $R3 \parallel R4$ matched resistances, drift can approach that of the reference device used (many others offer lower TC than the 329DZ’s 100ppm).⁶ This also assumes due attention to operating C4 in a low leakage condition.¹⁴ The $\frac{1}{2}W$ resistor types specified for R3 and R4 also aid DC stability.

- Master/slave (same polarity) regulators can be operated, using the common reference of the “master” stage. In such a setup, a single reference source feeds the master regulator R1-C4 filter and error

amp/pass device, plus additional slave RC filter networks and error amp/pass devices. This allows the additional slave regulators to be added without the full complexity of 100% parts duplication, at little or no performance trade-off. To ensure proper startup, the reference current of D1 should be increased, by lowering R6.

- Preregulation is possible in principle, but I didn't explore it in this article. My goal was to achieve as high a LR and low a dropout voltage as possible, within a single relatively simple stage. With typical broadband LR of 80dB or more and dropout voltage of 1.2V, the goal was met. I will be interested to hear reader enhancements which extend these performance areas.
- Setting up other output voltages of the general-purpose Figs. 8a or 8b regulators is simply a matter of selecting R3/R4 value(s) to provide a V_{OUT} level different than the 13.8V shown. *Table 1* simplifies this process for 10–18V outputs, and is used as follows:

With the Fig. 8a (8b) circuit's reference voltage V_R set at 6.9V by D1, the output voltage V_{OUT} is generally:

$$V_{out} = V_R \left(1 + \frac{R_4}{R_3}\right)$$

The ratio of $1 + R_4/R_3$ scales the 6.9V up to the output level, with R1 nominally equal to $R_3 \parallel R_4$. In this table, R1 is kept at 499Ω for simplicity's sake, and the R3-R4 values are adjusted to provide both the desired ratio shown in column 2 and the nominal 500Ω equivalence. The suggested R3-R4 values of columns 4 and 5 for a range of popular voltages are preferred 1% value selections, with the key word here being "suggested." Note that since the 329 diode's nominal 6.9V output can vary $\pm 5\%$, exact output voltages should not be expected (without trimming). However, this should really not be necessary, as absolute accuracy is probably not required. This table works for either the Fig. 8a or the 8b circuit, as long as you use a 6.9V reference voltage.

- Although performance for these circuits should be generally repeatable for similar conditions in relative terms, *the data here is based on a small number of samples*. You should expect some deviation due to sample-sample variation, which is not shown for the broader spectrum of circuits tested. Statistical sample studies for a given circuit, while quite worthwhile in their own right, are unfortunately beyond the scope of this article.

Beyond the different regulator topologies, there are also the broader system level points of remote sensing and star power distribution. Optimum use of wiring and layout methods can make application of the new regulators most effective, so Parts 3 and 4 cover this in more detail. General background reading on power supply systems and regulators of various types can only enhance your overall understanding of the circuits and their operation.^{19–21}

However, bear in mind that, even though sophisticated, these new regulators are simply one element of an overall AC \Rightarrow DC power conversion system. Optimization of the raw DC supply towards RFI/EMI suppression and other factors is equally critical to attain the highest performance levels.

A final caveat for the reader is not to overinterpret the absolute terms of performance data. Much of the regulator Z_O performance lies in implementation details, which is under your control. Take the time to do a good job with parts selection, the layout of your circuit, and wiring it into the system. You will be amply rewarded! □

STAY TUNED

I JUST FINISHED part 1 of "Regulators for High-Performance Audio" and wonder whether future articles will deal with higher power applications (i.e., power amps)? While not technically sophisticated enough to comment on the details of your article, I thank you for your contributions.

PAUL ERICKSON
Paul1@wizard.ucs.sfu.ca

Walt Jung responds:

I do not know of any plans for a higher voltage/current "power amp" article on regulators. I agree with you that it would be welcome, and hope someone will step forward to try it. I'd be happy to provide them with the Audio Precision software plus suggestions towards its use. Testing themes of the current series would also be applicable towards higher powered regulators, just as they would towards tube type regulators.

BREADBOARD SEARCH

I AM LOOKING for a source for the Ivanboard breadboards which Walt Jung refers to in his "Regulators for High-Performance Audio" article (TAA 1/95, p. 14).

MICHAEL J. AARVOLD
Simi Valley, CA 93065

Contact Mike Chittenden, 900 W. Grove Pky. #1101, Tempe, AZ 85283, (602) 345-8870.—Ed.

save one voltage reference, and common mode operation ought to be improved.

Also, when using multiple regulators, such as one pair for each audio channel, is it a good idea to reuse the voltage reference and have it supply multiple regulators? That would also decrease regulator complexity (and cost).

Well, time to go parts hunting for the components you use in the article (Sweden is a black hole when it comes to advanced parts for audio amateurs).

ERLAND UNRUH
<Erland.Unruh@malmö.trab.se>

Walt Jung responds:

I know very little about your "X-DAC" or the CS4328. I do know about the AD811, and I can assure you that it simply loves these new regulators. I've built them up in about three different forms, within my preamp as well as a modified (POOGE 5.5 and then some) DAC960. These all use AD811s as line drivers, and in each case the improvement over the previous level of power supply was well worth the upgrade effort and cost.

I am not so sure about the common mode factor, in terms of being a critical issue. As published, the Fig. 8a and 8b regulators are optimized for single supply operation, which inherently eliminates noise injection from the opposite unregulated rail. Note that a tracking type regulator as you describe does not have this advantage, so this factor is a basic weak point of that approach.

You could, however, work around this issue by using the regulated opposite output as the supply feed for the error amp. This could be somewhat complicated by startup problems, which you may appreciate. I did have a tracking type regulator in my preamp for many years, and when it was abandoned for one of these new dual independent regulators, there was a net improvement. This may have been due to the relative sophistication of the older tracking regulator (similar to the POOGE 5.51 circuit) rather than the two different topologies themselves. In any event (in my opinion at least) you won't be wasting your time to try the new regulator.

Yes, you can operate multiple, same polarity regulators from a common reference, in a master/slave fashion. One example of this was the twin +5V supplies used in the AD1891 evaluation board upgrade described by Gary Galo ("Ask TAA: Analog Devices Chips/Super Regulators," TAA 4/94, p. 43). In the context of the TAA 1/95 article, this would be equivalent to using Fig. 9 with a common AD680 reference but separate op

SUPER REGULATORS

THANK YOU FOR a very interesting article about regulators in *TAA 1/95* ("Regulators for High-Performance Audio, Part 1," p. 8). I have also enjoyed earlier POOGE articles and some other articles. I'm currently working on a regulator mod for my X-DAC 3.0 from Audio Crafters Guild. My main interest is in regulators for the AD811-based output stages and the analog supplies for the CS4328 DAC circuit.

All these three "components" use \pm supplies and can (very probably) benefit from good "common mode" supplies. Their worst case supply is when one rail is stable and the other one moves.

When supplies have such high quality as the ones you are writing about, how critical is common mode? Is it still a good idea to somehow make the regulators work together?

I have seen some different solutions for common mode regulators, but none have approached the quality of those in your article. They were built around NE5532 op amps and lacked very many discrete components.

Even so, the design in itself might save some components. If, for example, you require a positive regulator, you could perhaps design the negative regulator with the goal to "keep the negative voltage at the same distance from ground as the positive voltage is from the ground." That is, the ground becomes the negative reference point and the midpoint of the \pm output becomes the measure point. That would

amps and pass transistors for the analog and digital 5V outputs.

I have not done this as yet for the Fig. 8a or 8b circuit, but this would be useful for left and right channels, as one example. You could even have different voltage outputs from the various stages, since they all will have separate feedback networks. However, use separate RC filtering for each slave, and it might be a good idea to bias the 329 heavier in current, to avoid any startup issues.

I'll be interested to hear of your experiences with the circuit(s) and how they work for your application. □

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

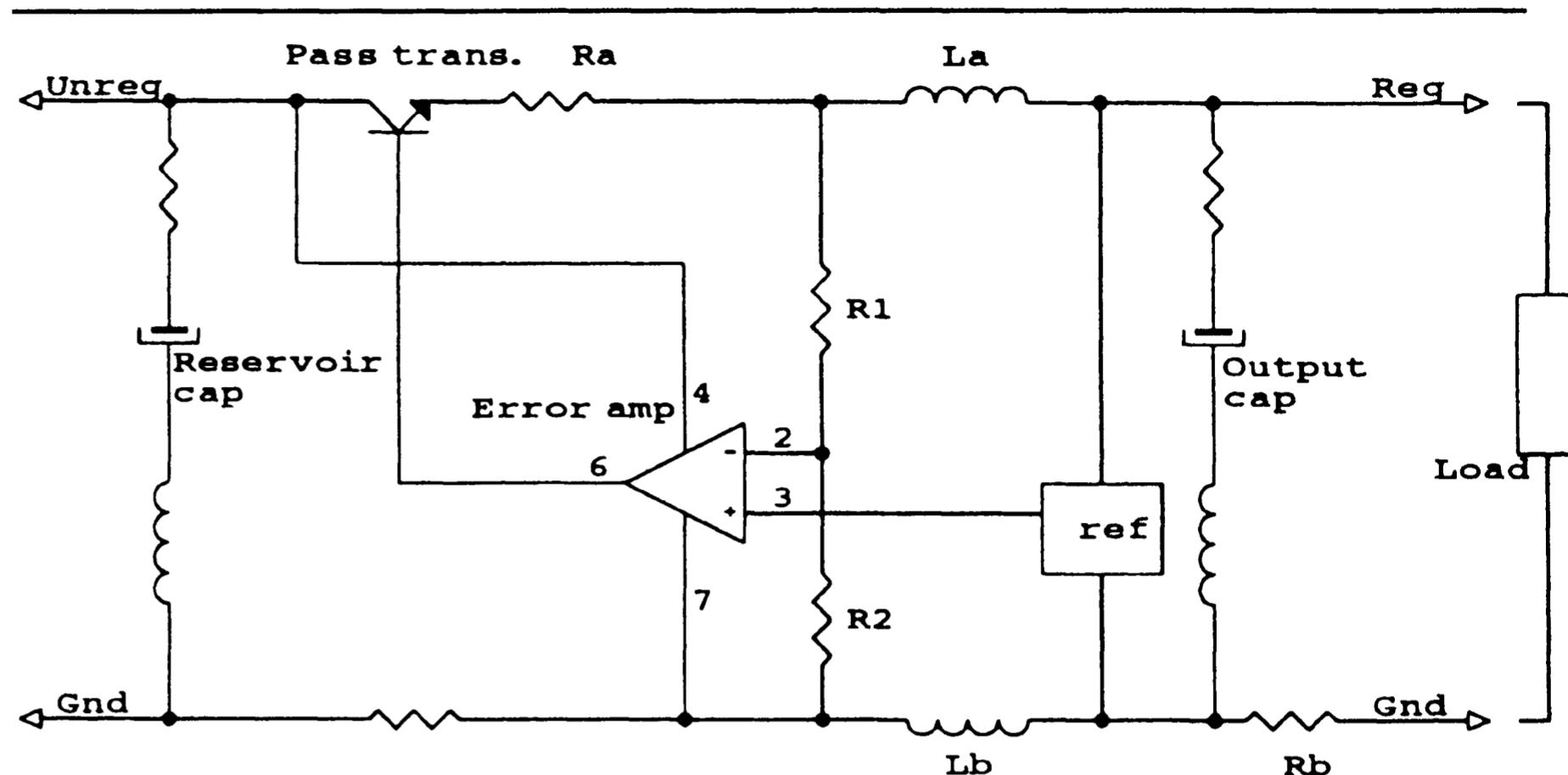


FIGURE 1: Real-world circuit, with parasitic components.

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.

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-

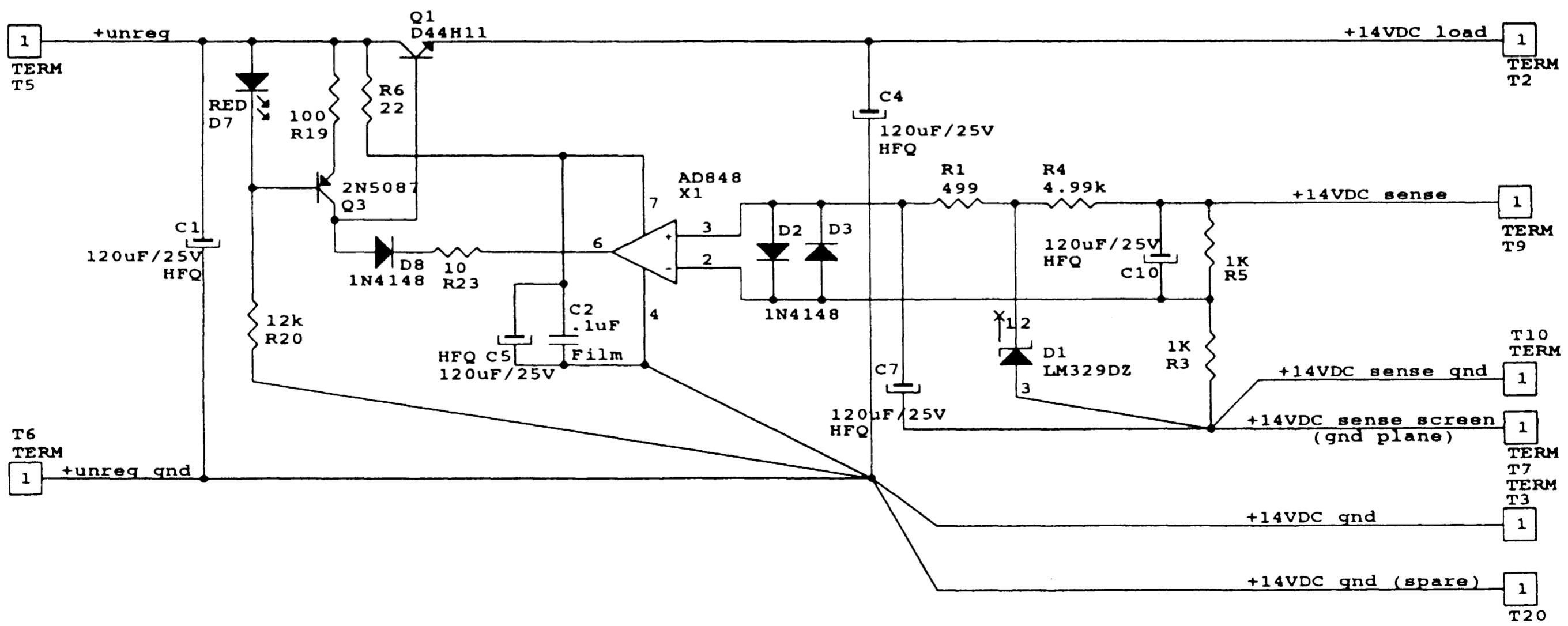


FIGURE 2: Positive regulator schematic diagram.

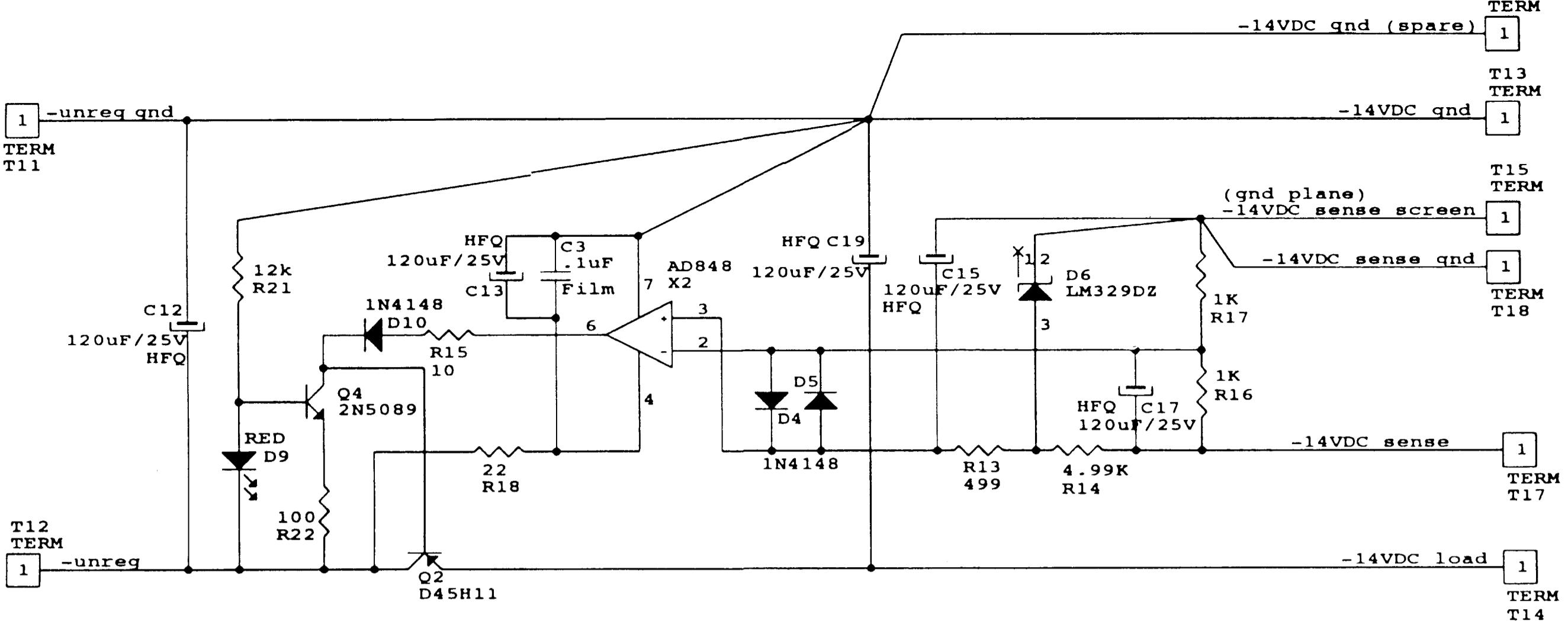


FIGURE 3: Negative regulator schematic diagram.

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 pre-

amp, 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

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).

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).

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

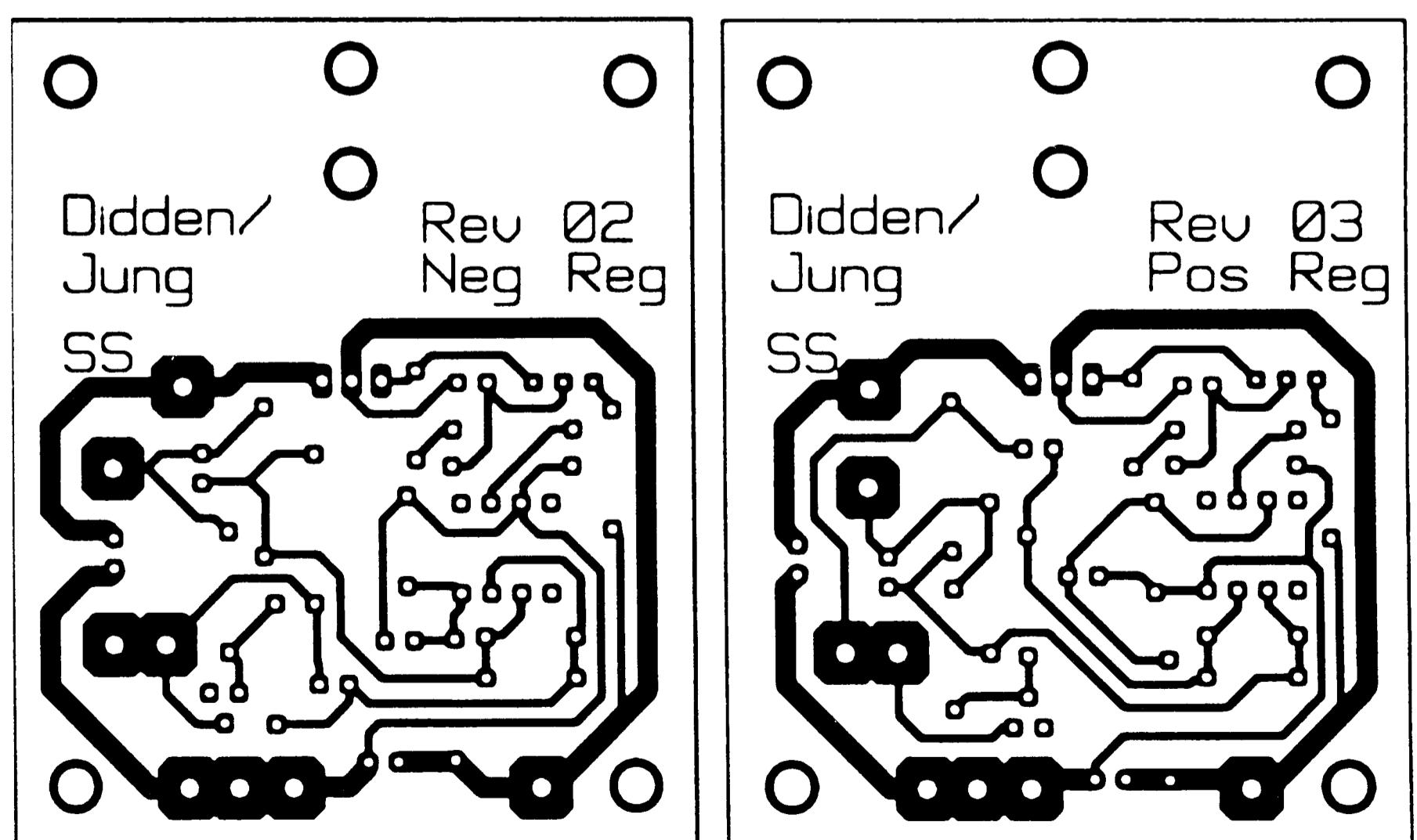


FIGURE 4: The board layout holds one positive and one negative regulator.

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 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.

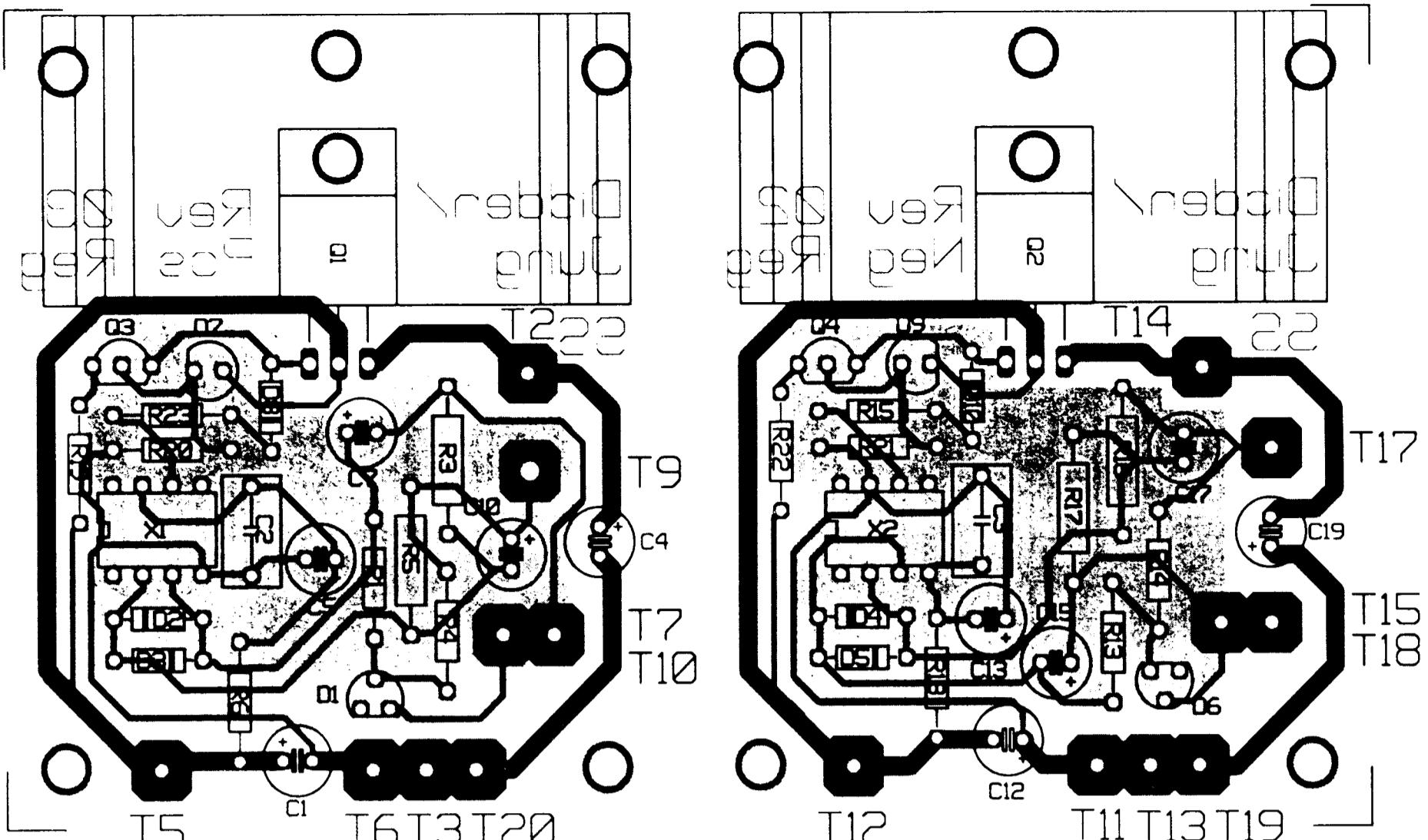


FIGURE 5: Stuffing guide for the combined positive/negative regulator board.

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.

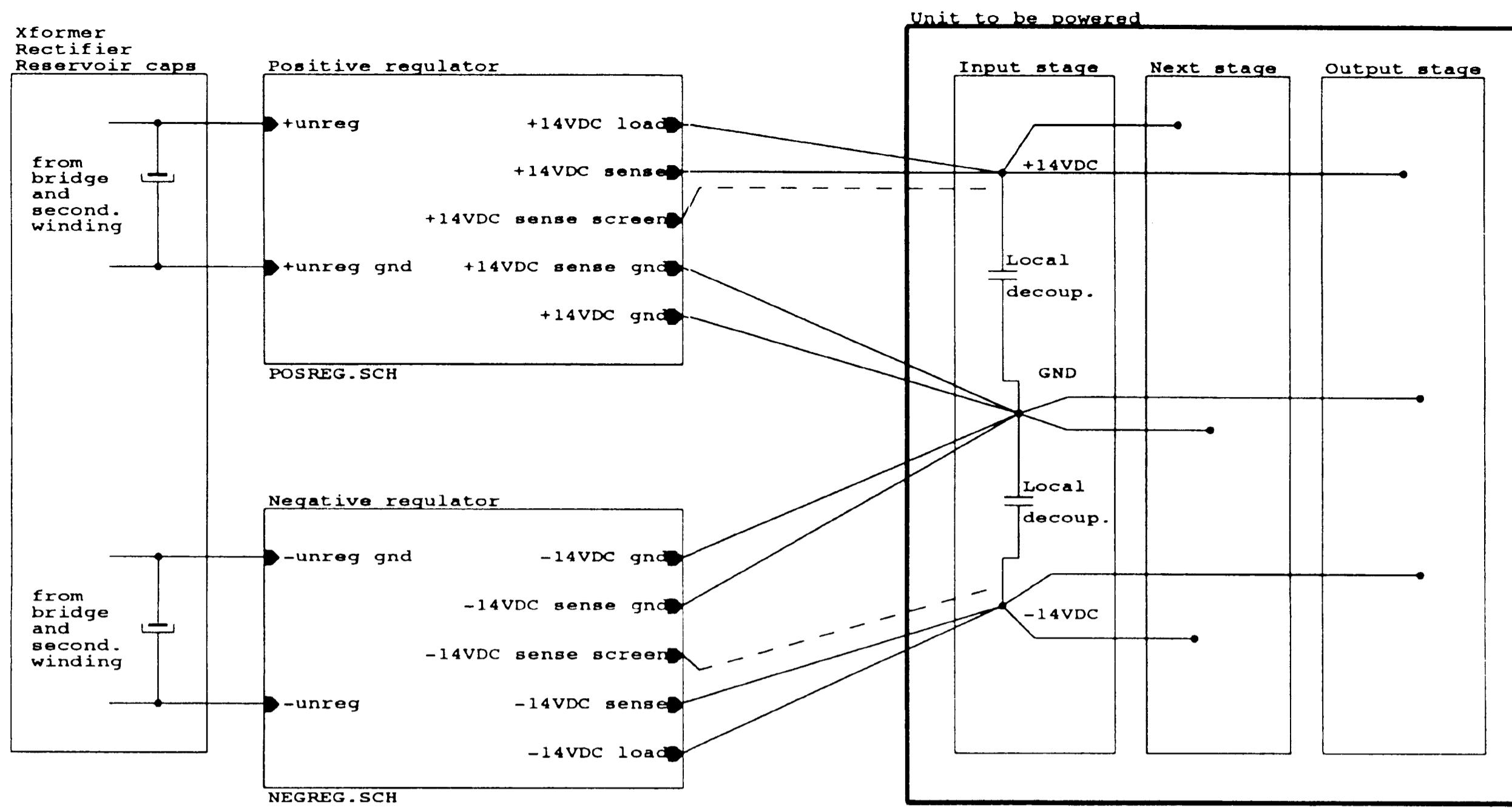
Now consider the op amp. Generally, the current that the op amp draws from its supply varies with the frequency and level of the output voltage and current that the op amp must deliver. Part of these current variations are short-circuited by the RC network's capacitor, but some current variations will 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



Note 1:
Twist load connection and supply input lines

Note 2:
Use balanced screened cable for sense lines
Connect screen to PCB sense screen pad only

FIGURE 6: Generic connection diagram for one channel.

ratings of $\pm 18V$, and operation within a volt of the recommended $\pm 15V$ 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 ($2 \times 15V$ AC for a standard 15V DC supply). You could also use a transformer with four sec-

ondaries. 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.

Old Colony Sound Lab often makes available printed circuit boards and kits in support of magazine projects if there is sufficient interest on the part of readers. To indicate your nonbinding vote for the availability of PCBs or complete kits of parts for this project, including PCBs, contact Old Colony, PO Box 243, Peterborough, NH 03458, (603) 924-6371, FAX (603) 924-9467. Availability decisions are usually made in about three months—watch Old Colony ads for further developments!

Positive reg. Z_o , remote load, local (top) and remote (bottom) sense.

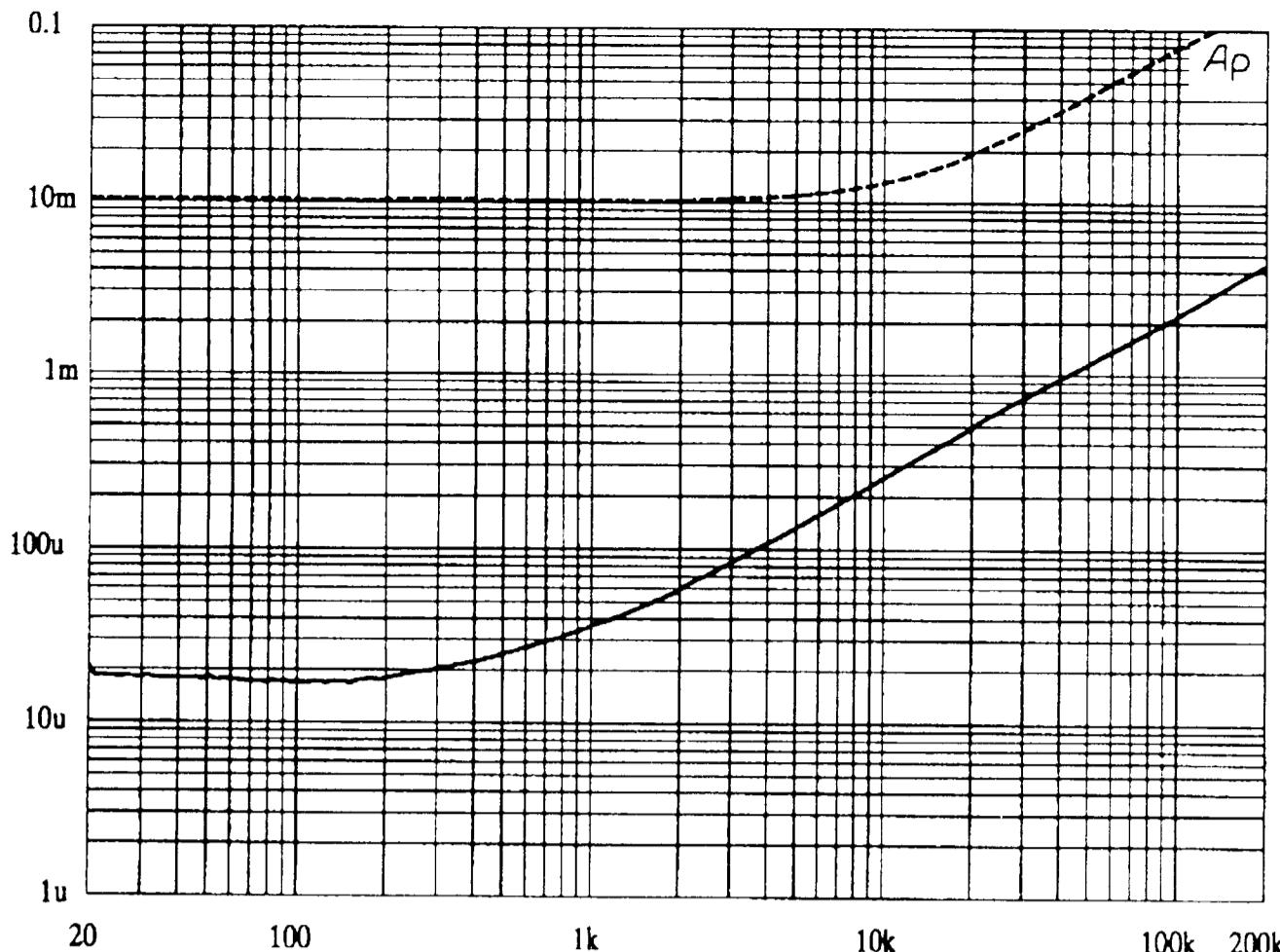


FIGURE 7: Positive regulator output impedance (Z_o) plots.

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 $1M\Omega$ 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

Negative reg. Z_o , remote load, local (top) and remote (bottom) sense.

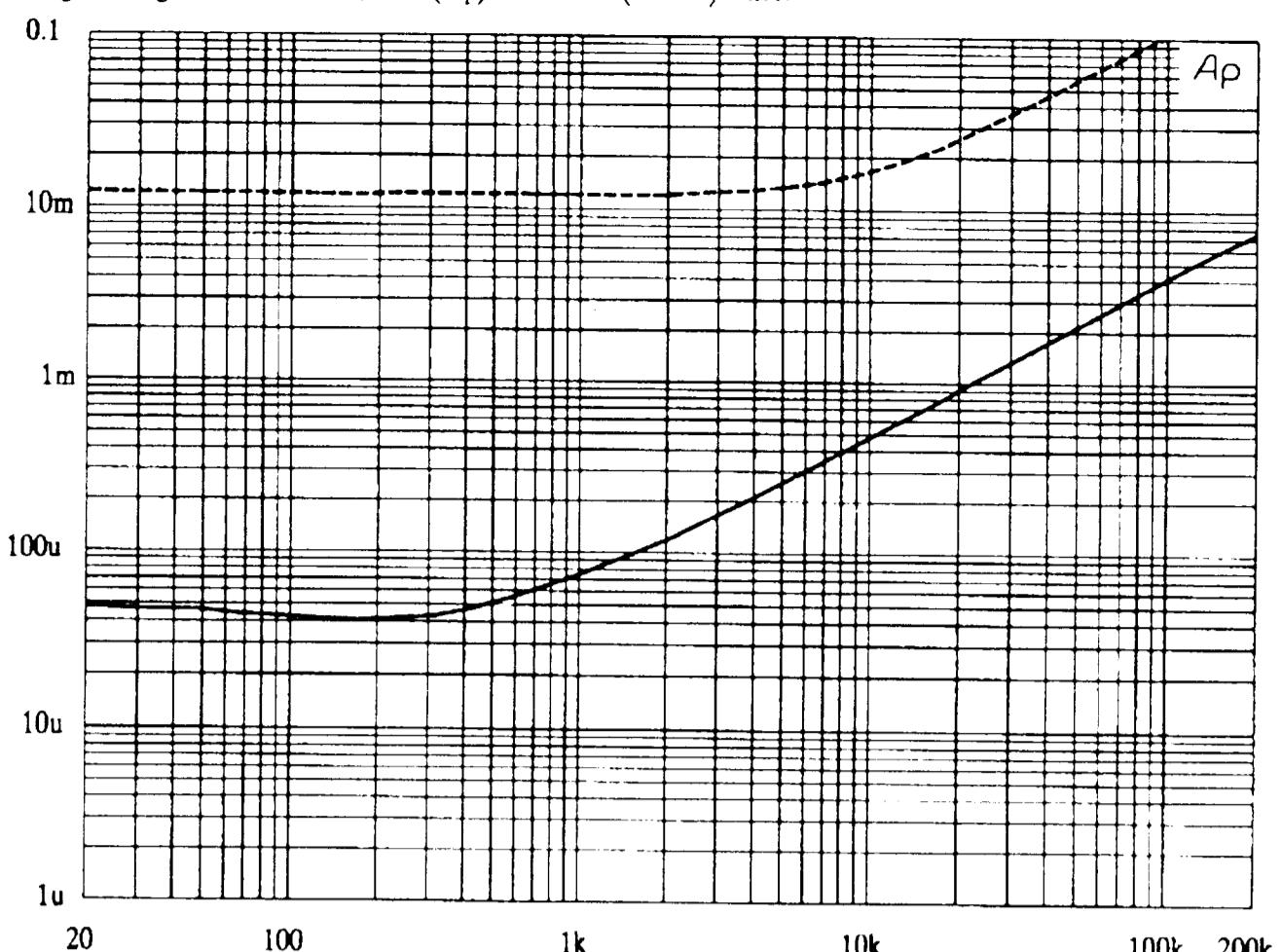


FIGURE 8: Negative regulator output impedance (Z_o) plots.

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

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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 $\pm 14V$ 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 consider-

ably 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 regula-

tors, 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)

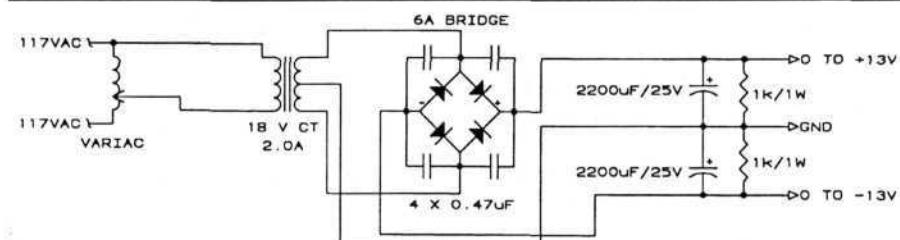


FIGURE 1: Raw supply for bench testing regulators prior to installation. The Variac is recommended to avoid damage if the circuit is not operating properly.

maximum sensitivity and minimum time base. On my scope these are 5mV and 0.2 μ s 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-

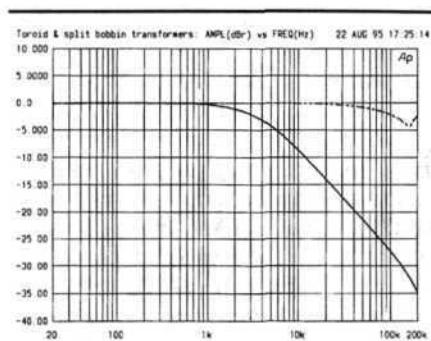


FIGURE 2: Bandwidth measurements on conventional and toroidal power transformers: dashed trace (top) is an Avel-Lindberg D-3022 toroid, flat to nearly 200kHz; solid trace (bottom) is a Magnetek FD7-36, nearly 35dB down at this frequency. (Courtesy of Rick Miller.)

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 $\pm 14V$ 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 0.47 μ F 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

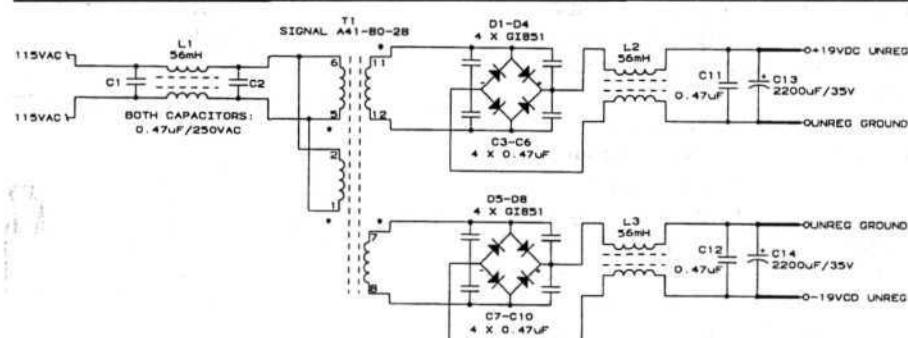


FIGURE 3: Raw supply suitable for use with high-performance regulators. Common mode chokes are used for AC line filtering and DC filtering after the rectifier bridges.

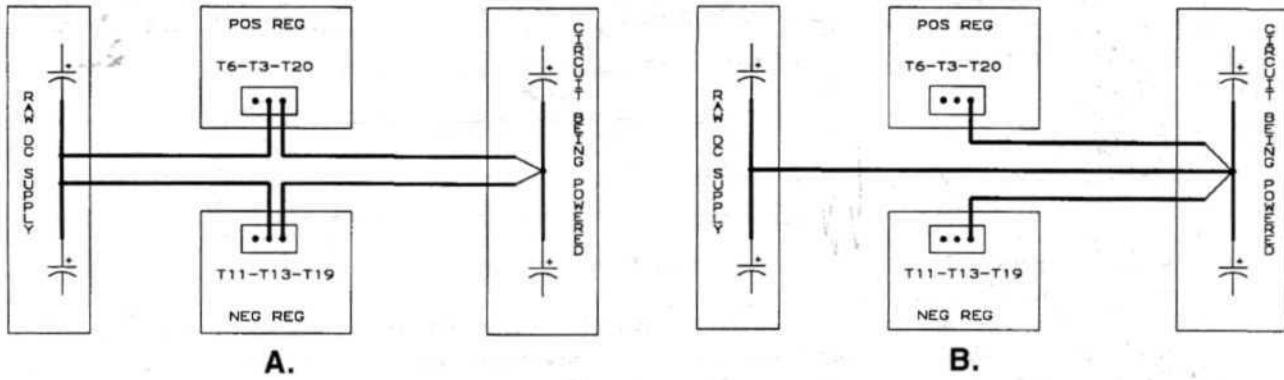


FIGURE 4: Two methods for connecting the regulators in equipment with a common raw DC ground. A: Incorrect method; B: correct method, recommended to avoid a ground loop.

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

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 $120\mu\text{F}$ 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

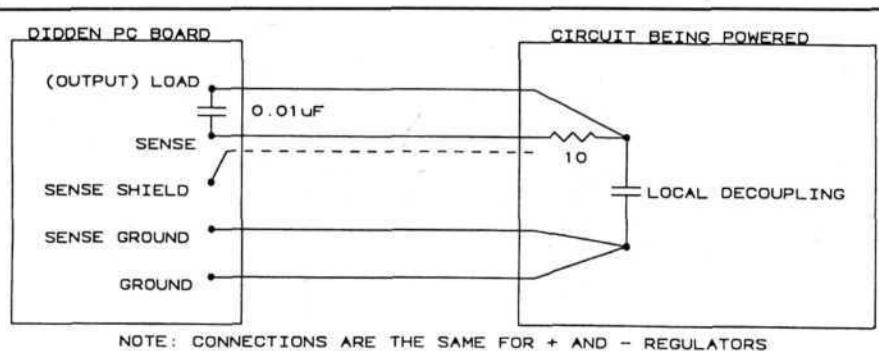


FIGURE 5: Remote sense decoupling scheme. Connections and parts values are the same for positive and negative regulators.

load-dependent; therefore, no ripple currents will develop across the 22.1Ω resistor. I have heard $\pm 14\text{V}$ 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 indicates

$\pm 5\text{V}$ 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 $\pm 5\text{V}$ 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 $\pm 5\text{V}$ 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 $+5\text{V}$ 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 $\pm 5\text{V}$ 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 $0.01\mu\text{F}$ 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 , $\frac{1}{4}\text{W}$ 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 millamps. 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.

Other Voltages

These regulators can be adjusted for other output voltages, though $\pm 14V$ and $\pm 5V$ 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 $\pm 14V$ 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 $100\mu F$ or so is overkill, and promotes instability. If you use film caps at the load, don't make them larger than $1\mu F$ 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 $0.015\mu F$, 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.

TABLE 3

PARTS LIST FOR MISCELLANEOUS REGULATOR CHANGES

REMOTE SENSE DECOUPLING

0.01 μF /50V	Panasonic V-series capacitor (Digi-Key P4513-ND)
10 Ω 1/4W, 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, 1/4W, 1%	Yageo metal film resistor (Digi-Key 10.0KX-BK-ND or Roederstein MK2)

$\pm 5V$ VERSIONS

2.49k, 1/4W, 1%	National Semiconductor LM336BZ-2.5 (Digi-Key LM336BZ-2.5-ND)
4.99k, 1/4W, 1%	Yageo metal film resistor (Digi-Key 2.49KX-BK-ND or Roederstein MK2)
	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 120 μF /25V HFO capacitors (Digi-Key P5698-ND)
 0.1 μF /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)

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 $\pm 14V$ 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 $\pm 5V$ 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 $\pm 5V$ 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 cellos, 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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7. *Noise Suppression Reference Manual*. Topaz Electronics, 1660 Scenic Ave., Costa Mesa, CA 92626, (714) 557-1636.
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