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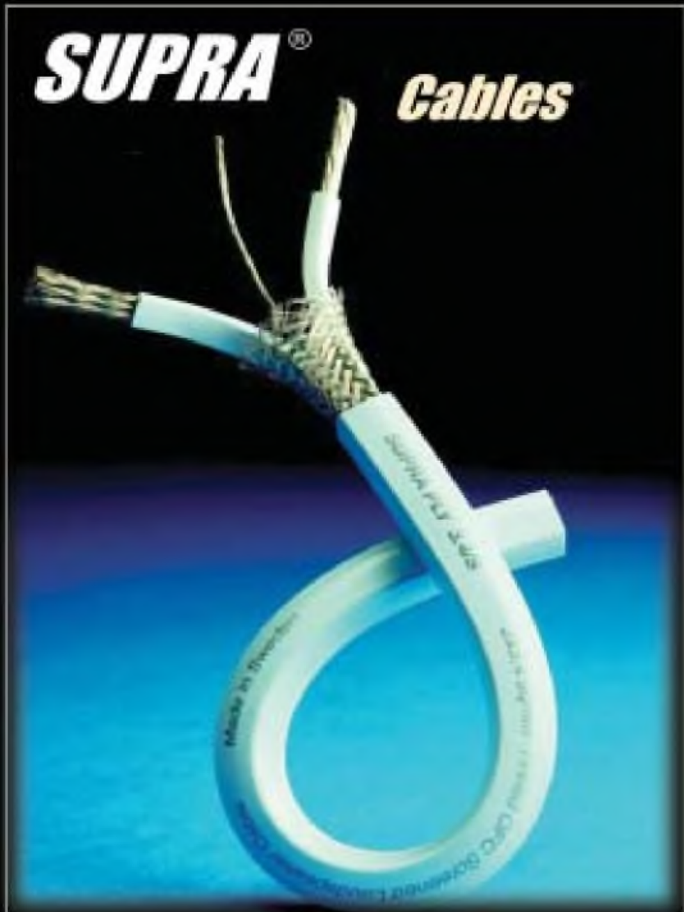
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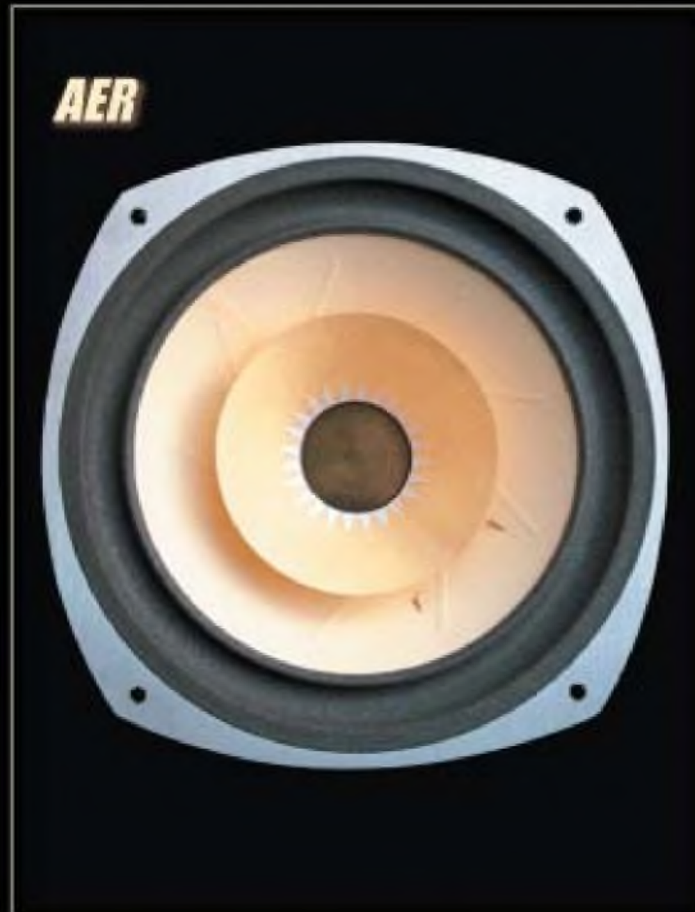
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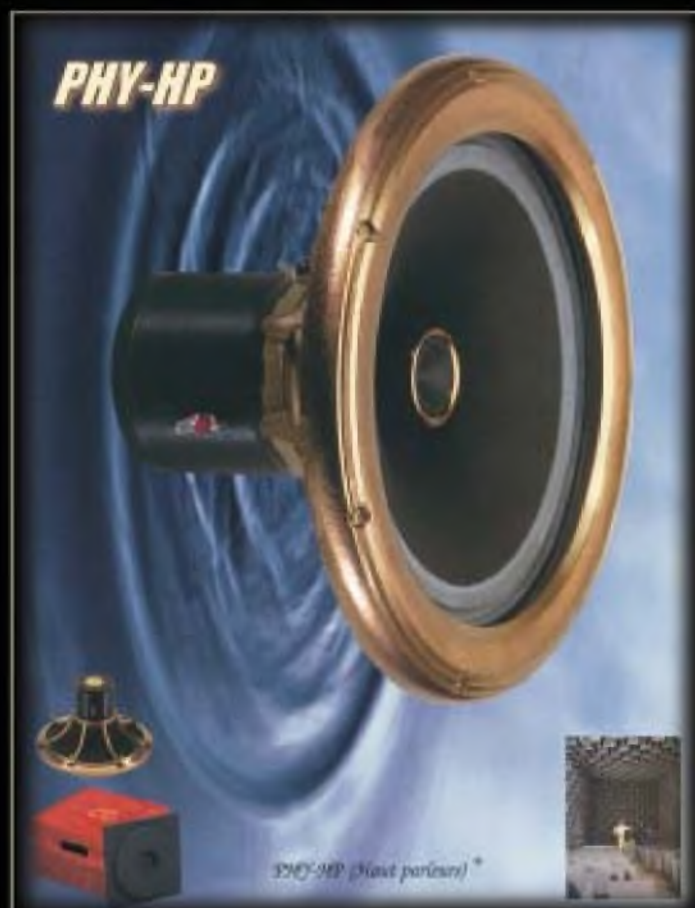
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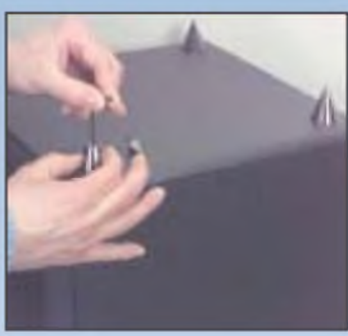
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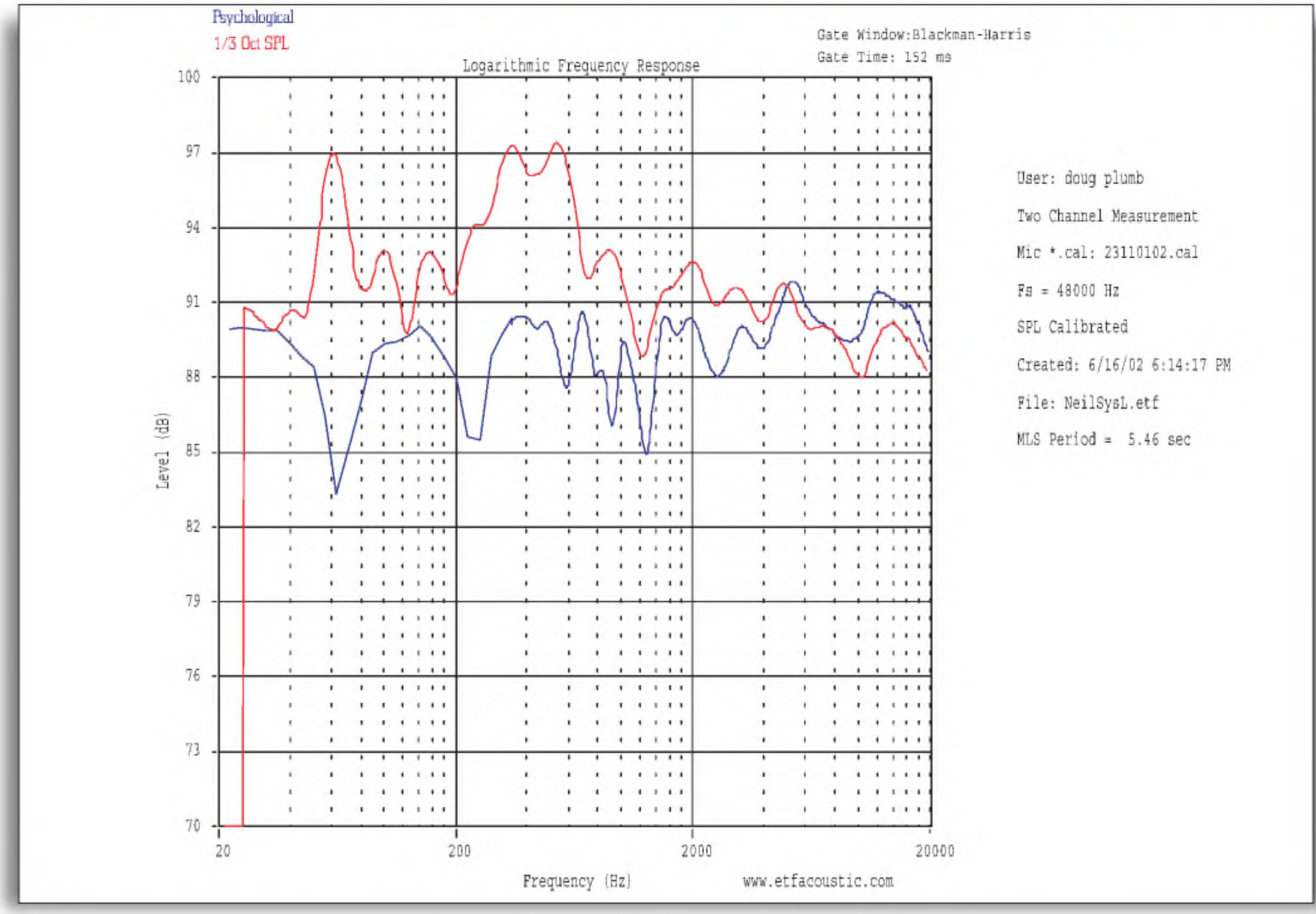
ETF5.X SOFTWARE FEATURES



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A 100W/Channel AB Differential Input Amplifier Part 2

Part two concludes our look at this 100W/channel amp, with information about the power supply, construction, and more.

By Norman Thagard

POWER SUPPLY

The voltage doubler/regulator schematic is shown as *Fig. 4*, while the raw power supply schematic is drawn as *Fig. 5*. The power supply of the prototype uses the same type of 200VA toroidal power transformers that were used in my 100W monoblocks. These are transformers that I picked up as surplus for $15 each. They have two independent 20V RMS windings rated at 5A. As in the monoblocks, two are used in series to provide approximately ±50V DC of unregulated, but filtered voltage.

As is my established practice with MOSFET output stages, the DIFF 100 uses a voltage doubler/regulator in order to provide the input stages with a regulated power supply whose voltage is about 5V higher than that for the output stage. This allows the output to nearly "swing the rails" and isolates the output supply from the input stages. Since the supply is bipolar with a center-tapped transformer secondary, it is possible to use full-wave voltage doublers.[10] However, I incorporated half-wave voltage doublers, with which I have long experience, into the DIFF 100.

Regulator outputs were measured at ±56.4V, but the value of R107 in the positive regulator differs from its counterpart, R108 of the negative regulator. R107/R109 on the upper and R108/R110 on the lower half of the regulator circuit set the positive and negative regulator output voltages, respectively. The greater base current draw of the pnp MPSU60–used as Q104 as opposed to the npn MPSU10 used as Q103–would otherwise cause Vreg(–) to be a couple of volts higher than Vreg(+).

VR101 and VR102 are zener-connected Motorola pnp transistors. Since currently sold units may not give the desired 7V reference when so used, it is perfectly acceptable to substitute 6.8V zener diodes. Check the regulated output voltages to be sure that they remain around 55V magnitude.

Little further elaboration on the operating principles of the voltage doubler/regulator is necessary, because it is almost identical to that described in a previous article.[11] However, please note that the voltage doublers reference one end of C103 and C104 to the positive and negative unregulated rails, respectively. In other amplifiers, I have referenced these capacitors to ground.

The advantage to the approach here is that the required voltage rating of these capacitors is reduced by almost half. The disadvantage is that any ripple on the unregulated rails is applied to the input of the voltage regulators that immediately follow the doubler circuitry. Since I pre-regulate doubler output with zener shunt regulation, this should have little impact.

If you desire the more conventional technique, *do not* connect either the Vunreg(+) or the Vunreg(–) outputs from the unregulated DC rails to the voltage doubler/regulator PC board. Instead, interconnect the Vunreg(+), Vunreg(–), and star ground traces on the input side of that board together with jumpers. Star ground itself is the midpoint of a heavy brass bus bar that spans the common terminals of the unregulated power-supply filter capacitors.

Of course, C103 and C104 should be 150V DC-rated units if this modification is made. Otherwise, the voltage rating of C103 and C104 may be as low as 63V. A rating of 80V is shown on the schematic and in the parts list because similarly rated components were incorporated into the prototype due to availability.

If you take this option (and even if you don't), you may omit the pre-regulator. This will increase power dissipation in pass transistors Q5 and Q6, so larger heatsinks may be required for these devices. These power BJTs could dissipate several watts in the absence of pre-regulation.

PARTS AVAILABILITY

It is unusual when no suitable alternative to a discontinued component is available, but can occur. The DIYer may experience some difficulty obtaining some existing parts. Although I attempt to use readily available components, other considerations sometimes supersede this priority.

An example of this would be the

PHOTO 1: Amp front.

high-g_m JFETs used in my phono preamp.[12] In that case, I knew of no alternative device that would be comparable. However, the fact that these devices allowed realization of a discrete topology otherwise unrealizable took precedence. I believed it possible that if the article attracted sufficient interest, *audioXpress* might step in and secure the transistors for sale to builders.

Also, in that particular case, I knew that those JFETs were readily available in small quantities by purchase over the Internet from Erno Borbely, albeit at a price significantly in excess of that obtained through quantity purchase. A reader has subsequently sent an e-mail indicating that Match-A-Knob of 170-30 Jamaica Avenue, Jamaica, NY 11432 (800-321-5662) has plenty of these devices in stock for about $1.25 each with a minimum order of $25.

Employment of replacement-series parts may not be a panacea for the builder. I found one of four channels of my phono preamp to be unusually noisy, and traced it to the negative voltage regulator, a device from one of the several popular replacement-series manufacturers. A quick look showed a second such part to be noisy as well. None of the several standard brand regulators that I subsequently purchased suffered from this defect.

While this could be a fluke, it is one reason I prefer standard brand components rather than replacement-series devices. On the other hand, some of my designs have specified replacement-series devices, so my reluctance is admittedly not absolute.

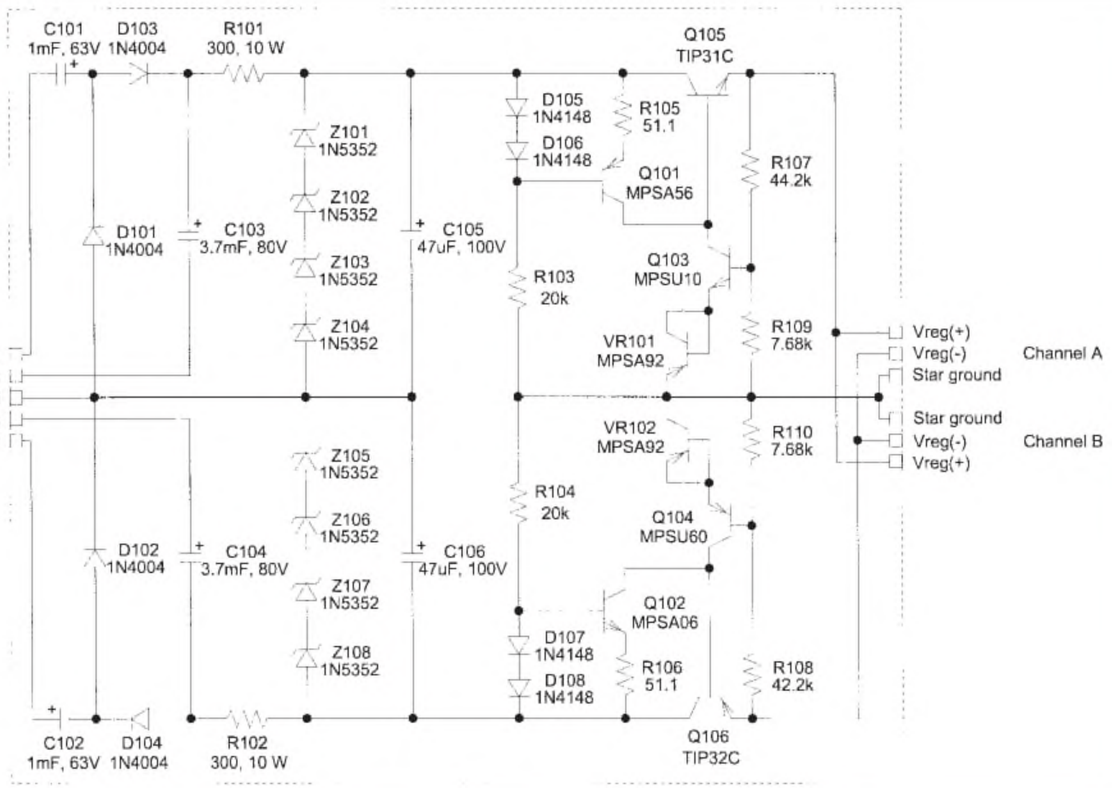
When I offer designs for publication, I am at least as much interested in the presentation of concepts as I am in the presentation of a DIY project. My hope is that a DIYer might gain sufficient knowledge of principles to modify my designs or even effect an entirely new design. It is very satisfying when former students of my "Feedback Amplifier Principles" course write or e-mail me to say that they've successfully built their own preamp or amp; some have even boasted that their design was better than anything that I presented. Surprisingly, I find that pleases rather than offends me. My further hope is that my articles have enough socially redeeming value to justify publication, even if the design is not practical for the average DIYer.

It would be dangerous to speak for the editor, but I suspect that the purpose of *audioXpress* is broader than presentation of a collection of DIY projects. My impression is that the magazine generally serves the interest of technically minded devotees of sound reproduction. There certainly have been a number of purely tutorial articles over the years, so I am satisfied if purported design articles adequately serve this function, even though centered on a design rendered unrealizable through component unavailability.

I do acknowledge that the author has an obligation to apprise the reader of any known availability problems. This is something that I have done in my articles. It should be understood, however, that the author doesn't always know that availability problems exist at the time of article submission and certainly cannot know whether manufacturers will discontinue components afterwards.

Bear in mind, too, that *audioXpress* is not a peer-reviewed journal. In fact,

FIGURE 4: Voltage doubler/regulator schematic.



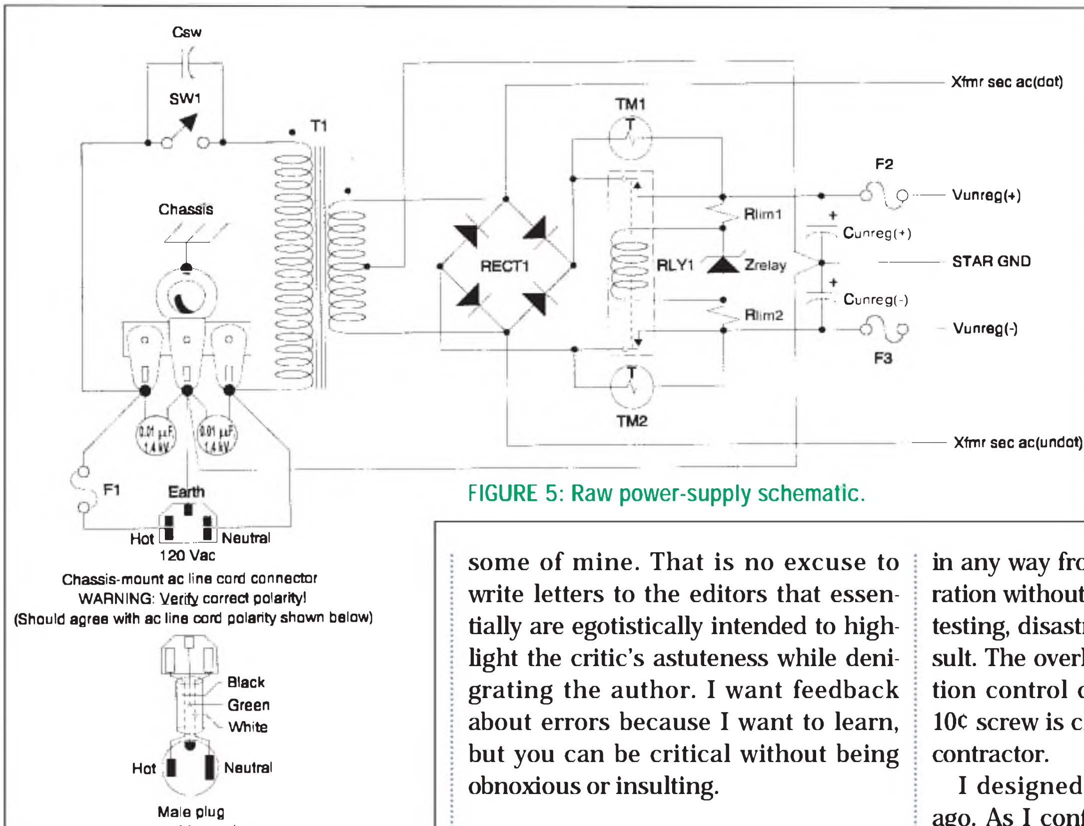


FIGURE 5: Raw power-supply schematic.

while authors are paid for accepted submissions, the pay does not justify the expense of even proofreaders, let alone independent technical review prior to publication. There may be mistakes in articles; there certainly have been in some of mine. That is no excuse to write letters to the editors that essentially are egotistically intended to highlight the critic's astuteness while denigrating the author. I want feedback about errors because I want to learn, but you can be critical without being obnoxious or insulting.

CONSTRUCTION DETAILS

You can infer many of the recommended construction details from the amplifier schematic (Fig. 1, Pt. 1), voltage doubler/regulator schematic (*Fig. 4*), and raw power-supply schematic (*Fig. 5*). In addition, the PC board patterns for the amplifier front-end and voltage regulator/doubler are provided as *Fig. 6* and 7, respectively.

I have developed a much more sympathetic attitude concerning quality assurance issues since I began writing articles. There is a reason why NASA human space flight costs are high. If every last thing is not well-documented and if the final product changes in any way from design or test configuration without appropriate testing or retesting, disastrous failure can be the result. The overhead of careful configuration control can be the reason that a 10¢ screw is charged at a $1 rate by the contractor.

I designed this amp several years ago. As I confirmed the actual component types and values, I realized that potential builders would need a "heads-up" since some of those component values and types would ideally be different. Also, I am more experienced now, and some of my design and construction preferences have changed.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These concerns about configuration control notwithstanding, there was an awful temptation to offer a modified design for this article that would differ from the as-built amplifier. Thus, Q1–Q10 are specified here as MPSA06/MPSA56 complementary devices. However, the actual amplifier is populated with MPSA05/MPSA55 BJTs.

There is no difference in the AC parameters, but the '06/'56 pairs have an 80V DC collector-base breakdown voltage, while the '05/'55 pairs are rated at only 60V DC. Given the voltages involved, the former are to be recommended for a new design, but again, they are not the devices actually in the prototype. Given that there have been no failures of any of these particular transistors, I am not inclined to retrofit my amplifier at this point.

Similarly, protective zener diodes Z3–Z6 are specified as 1N4739A units. The amplifier had 1N5242Bs in these positions. Again, no failures have occurred with the former devices, so I did not change the prototype per my current recommendation. The 1N4739A, whose Vz is 9.1V, affords better protection to its associated MOSFETs and is also a sturdier device due to its 1W power rating. Since, as previously mentioned, the 2SJ49/ 2SK134 Hitachi power MOSFETs can be damaged if gate-source voltage (Vgs) exceeds 14V, the 12V Vz of the 1N5242B added to the forward drop of a diode limit is too close to $(Vgs)_{max}$ for comfort.

Z5 and Z6 were added after the fact when MOSFET M5 failed.[13] Since the PC board pattern did not provide for installation of Z5/Z6, I soldered them to the foil side of the board. Per the cited reference, which was a letter to the editor from Nelson Pass, a signal diode such as a 1N914 or 1N4148 can be oriented in the same polarity as a zener and used in lieu of a zener. Unfortunately, I had made this modification before I saw the recommendation contained therein. As I recall from conversation with Nelson, the failure mode was discovered when a storage scope documented the condition that caused the failures.

Some components were selected simply because they were in my parts bin. For example, I used a TIP31C for Q11, but a more cost-effective choice is the TIP29, since the $(Vce)_{max}$ = 100V and $(Ic)_{max}$ = 3A ratings of the former are ridiculously in excess of the requirement.

The amplifier schematic (Fig. 1) depicts a single channel. Duplicate the circuitry for a stereo (two-channel) configuration. The exception is the balanced/unbalanced mode switch, which appears on the schematic as a SPST type. If you construct the DIFF 100 as a monoblock, this is the appropriate type. However, as a stereo unit, only a single DPST mode switch is required to accommodate both channels, unless you plan to operate one channel in balanced and the other in unbalanced mode.

The original schematics and PC board patterns were manually drawn. There lurks a danger that the newly produced electronic versions submitted to the publisher contain errors despite my careful attempts to verify exact equivalence. I made some minor changes to the patterns that should have no effect, but new boards were not etched, populated, and tested from the modified patterns. No doubt, letters will ensue quickly and will be welcomed if mistakes are found.

I have tried to provide sufficient detail in the schematics to make physical construction and interconnections obvious. I use quick disconnects (QDs) liberally in my prototypes and recommend them for beginners who are more likely to make a failure-producing mistake requiring board removal. Were I to construct this amplifier anew, I would omit them due to the proven reliability of the prototype. The PC board patterns

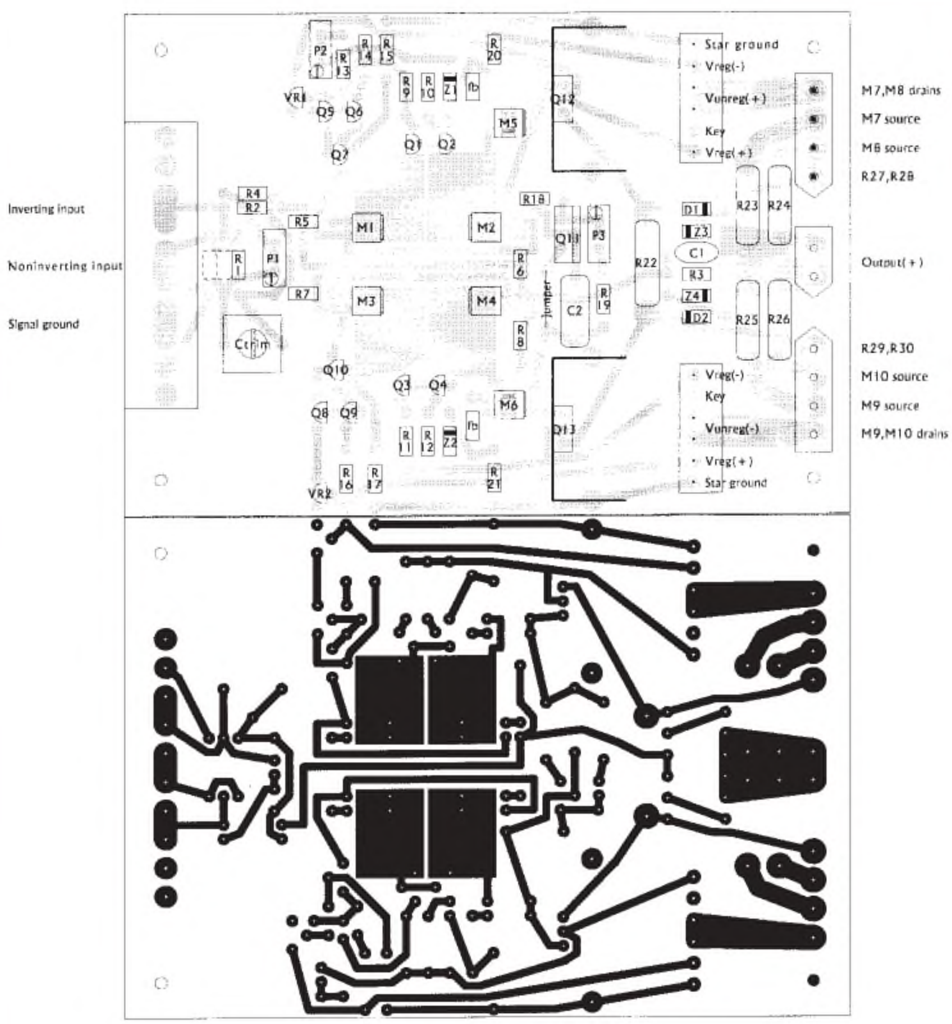

FIGURE 6: Amplifier front-end PC board pattern and component placement guide (75%).

reflect the particular QDs actually used right down to the mix of male/female headers that prevent inadvertent interchange of cables.

Transistors Q12/Q13 require heatsinks. Please note that the heatsinks are grounded by the PC board foil pattern, making it necessary to insulate the BJT collectors from the heatsink in the usual manner.

WARNINGS

Two warnings are in order. The first involves power-supply connections to the standard 3-wire, 120V AC-line cord. The power-supply schematic (*Fig. 5*) shows the connection via a standard chassis-mount male QD receptacle, which would receive 120V AC power through a standard detachable power cord. I actually use AC line filters that are integrated with a male, chassis-mount AC line connector. These filters may be labeled to indicate "hot" and "neutral" connector pins. The "earth" connection to chassis may be via a similarly labeled "earth" or "ground" connector, via a wire with green insulation that must be bolted to the chassis, or may automatically be accomplished through the metal screws that attach the filter/connector unit to the chassis.

The best approach for verification of proper connection of any chassis-mounted AC connector is accomplished with the AC line cord plugged into the connector but unconnected to the AC wall outlet. Use an ohmmeter to check continuity between the plug (AC wall outlet) end of the line cord and the three chassis-mounted connector pins in turn. The smaller of the two parallel spades plugs into the "hot" terminal of the AC wall outlet, the larger into the "neutral," and the circular middle prong contacts the "earth" terminal. If you use an integral filter, it should be rated at least 3A. You may omit the two 0.01µF, 1.4kV disc ceramic capacitors shown on the power-supply schematic if you use a filter.

Second, the 10Ω resistor shown connected between the outer (shield) connection of the RCA jack and chassis in Fig. 1 is prone to fail open. The placement of this resistor in this manner is very effective at eliminating a potential ground loop, but any voltage difference between ground potential of preamplifier and amplifier will cause current flow through this resistor. This scheme has been employed in all of my amplifier designs, and resistor failures have occurred in every one of them.

I am not the first to use this technique for ground loop prevention. Recently, I found that the signal ground to chassis (and hence to earth ground) resistor used in my Krell preamplifier measured 1.5kΩ instead of its marked value of 10Ω. I have not determined whether these resistors fail due to excessive current flow or voltage spikes, but I suspect the latter since measurements taken in normal circumstances have failed to show significant steady-state voltage across them.

In the A75, Nelson elected to isolate chassis from circuit ground at the power supply with a 5.1Ω, 1W resistor. This resistor was in parallel with two

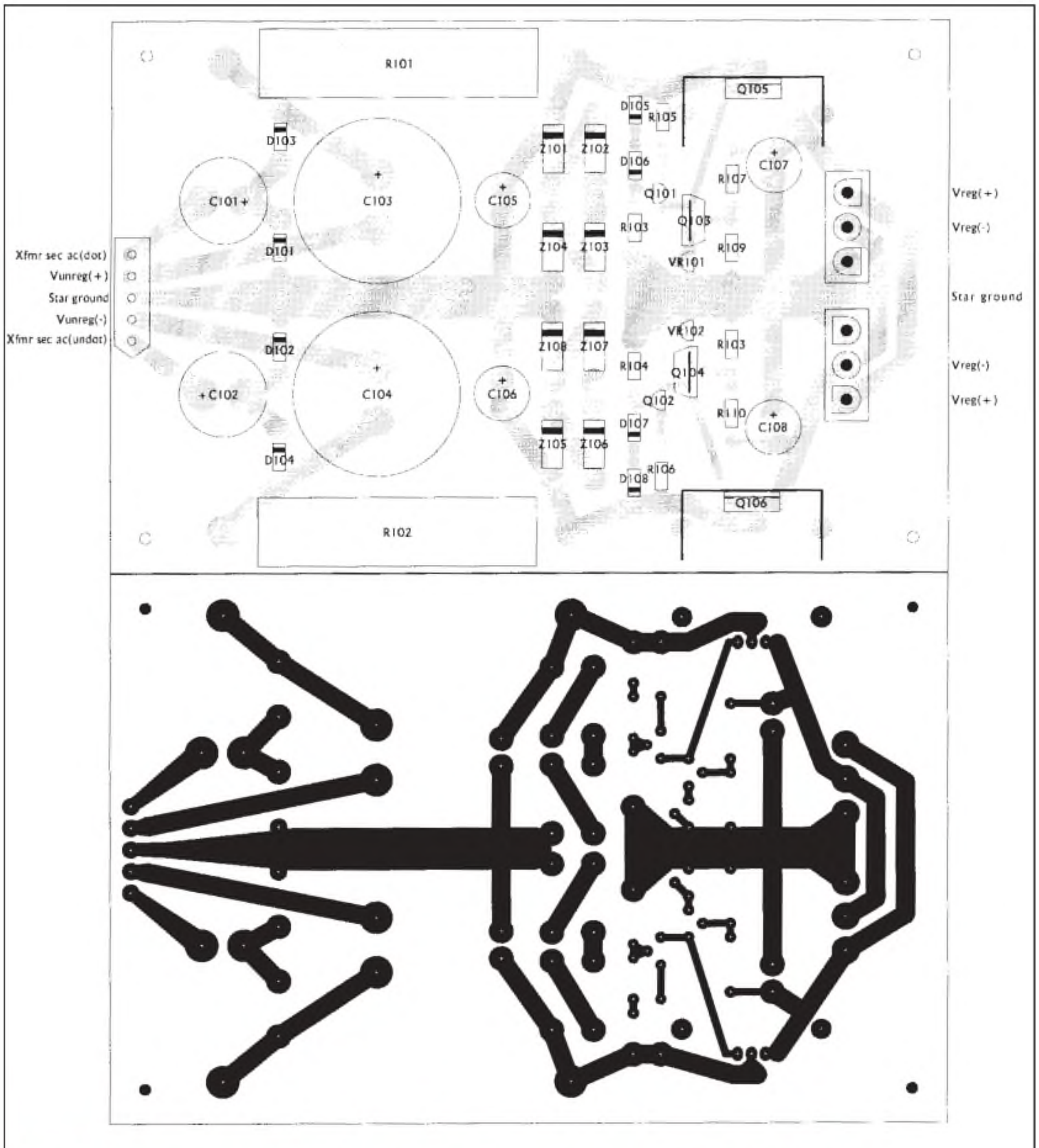

FIGURE 7: Voltage regulator/doubler PC board pattern and component placement guide (75%).

oppositely oriented 1N5401 diodes. This suggests that he was protecting against failure due to voltage spikes, although the diodes are rated at 3A continuous. You may similarly choose to parallel the 10Ω resistors with diodes as do (some of) the prominent commercial designers. In general, the higher the resistor's wattage rating, the higher is its maximum voltage rating.

INITIALIZATION

The initial operation requires two to four adjustments and should be made with no load attached. P1 sets the low-frequency common-mode rejection ratio (CMRR), and Ctrim sets the high-frequency CMRR. P2 is used to null any DC offset at the output. Finally, P3 sets the DC bias current of the power MOSFETs in the output stage.

The initial adjustment should be to P3, but if two voltmeters are available, you should use the second to monitor output offset. Before power-on, place P3 at its maximum resistance. Measuring voltage across any of the four resistors R23–R26 affects monitoring of output MOSFET source current. With a resistor value of 0.5Ω, current (mA) is twice the measured voltage (mV). Adjust for a reading of 50–60mV, which corresponds to a bias current of 100–120mA.

Next adjust P2 for 0V DC at the output. This voltage will vary over the short run, but should stabilize at no more than 60mV magnitude. Both P2 and P3 should be readjusted after at least 30 minutes of continuous operation to allow for the inevitable thermal effects.

Adjustments of P1 and Ctrim are not critical since, as I mentioned earlier, each can be replaced with a fixed value component. I used my oscilloscope to make these adjustments. With the balanced/unbalanced switch in the balanced position and driving both inverting and non-inverting inputs with the same 1kHz signal, I adjusted P1 for *minimum* output. Then, with 20kHz input, I adjusted Ctrim for *minimum* output. Using the highest driving signal

TABLE 2
POWER SUPPLY PARTS LIST

(ALL RESISTORS ARE 1%, ¼W METAL FILM UNLESS OTHERWISE SPECIFIED)

SCHEMATIC REFERENCE	PART #/VALUE	DESCRIPTION	MANUFACTURER
C101, C102	1m, 63V	electrolytic	
C103, C104	3.7m, 80V	electrolytic	
C105, C106	47µ, 100V	electrolytic	
Csw, AC hot, neutral to gnd (2)	10n, 1.4kV	disc ceramic	
Cunreg(+), Cunreg(-) (3)	10m, 63V	electrolytic	
R101, R102	300, 10W	wirewound	
R103, R104	20k		
R105, R106	51.1		
R107	44.2k		
R108	2.2		
R109, 110	7.68k		
Rlim1, Rlim2	470, 5W	wirewound	
Q101	MPSA56	pnp BJT	Motorola
Q102	MPSA06	npn BJT	Motorola
Q103	MPSU10	npn BJT	Motorola
Q104	MPSU60	pnp BJT	Motorola
Q105	TIP31C	npn power BJT	TI
Q106	TIP32C	pnp power BJT	TI
VR101, VR102	MPSA92	pnp BJT	Motorola
D101–D104	1N4004	1A, 400 PIV Diode	
D105–D108	1N4148	diode	
Z101–Z108	1N5352B	15V, 5W zener	
Zrelay	1N5359B	24V, 5W zener	
RECT1	600 PIV, 25A	bridge rectifier	
TM1, TM2	KC011L	NTC thermistor	Keystone
F1	3A slow-blow	line fuse	
F2, F3	3A fast-blow	DC rail fuse	
T1	80Vct, 5A	power transformer	
RLY1	76R4-24DC-SCO	DPDT relay	Sigma

Chassis-mount male AC power connector, 3-conductor AC line cord, SPST on/off switch rated 120V AC @ 10A, fuse holders, TO220 heatsinks, standoffs, enclosure

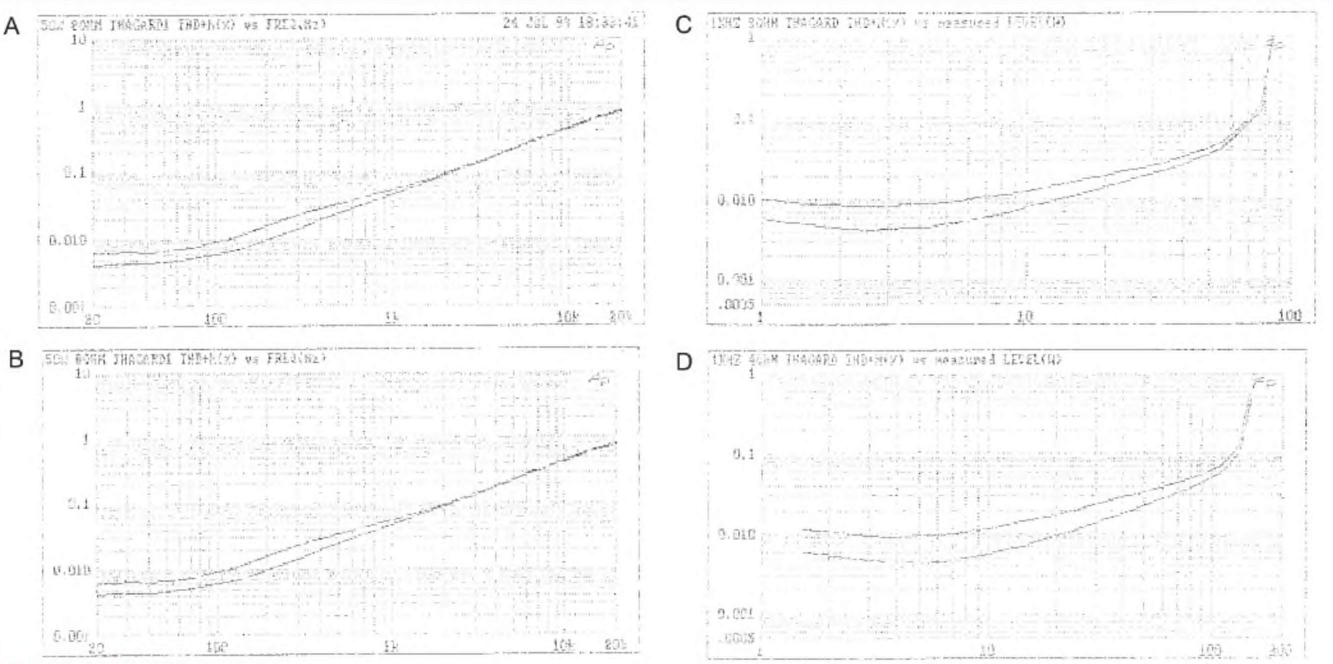

FIGURE 8A–D: THD curves.

not to exceed clipping or obvious distortion gives the maximum sensitivity since even in the 2mV/division scope vertical sensitivity position the trace can be nearly flattened.

I measured a high-frequency CMRR of over 60dB with these adjustments. The CMRR is just the ratio of differential-mode gain to common-mode gain. The differential-mode gain is about 20, and the common-mode gain will be the observed output voltage from the test divided by the input voltage that produced it. CM gain will be much less than 1, so CMRR should be much greater than 20 (26dB).

PERFORMANCE

The THD curves supplied (*Fig. 8, a–d*) were generated on PASS Lab's Audio Precision analyzer. They show that the prototype could not achieve more than about 85W into an 8Ω load. Since I can achieve slightly more than 100W at clipping on my test bench, I assume the published curves were for both channels driven simultaneously. The limitation is due to the power supply. Rail voltage drops rapidly as the output power increases.

Again, the power transformers were surplus units with only 10,000μF filter capacitors used. A common power supply powered both channels. A stiffer supply should achieve 100W/channel with little trouble, even with both channels driven simultaneously.

Note that at the 50W output level, distortion ramps up smoothly with increasing frequency to 0.75% at 20kHz. My goal is usually no more than 0.1% even at this frequency at or near full-rated output. This level of high-frequency distortion reflects the high impedance in the feedback network. Recall that this was a necessary condition to keep input impedance at a reasonably high level in balanced input mode. At lower output levels or frequencies, THD is quite low.

In no event should anything close to 50W at 20kHz be output in normal use. There is very little energy at the upper end of the audio band and high-frequency drivers; i.e., tweeters are seldom designed to accommodate more than a few watts without damage. In listening tests, highs are smooth. There is no audible concomitant to the THD. Indeed, it is not unusual for tube amplifiers to have such levels of THD at even lower frequencies.

It is interesting that a friend of mine who has been using a pair of my 100W class A monoblocks for ten years in a pretty good system thought that this was the best sounding of the four amplifiers with which I have been associated. This, despite the fact that this amp is class AB using a common power supply, whereas the others were operated class A and constructed as either monoblocks or as dual mono units.

I am currently working on a balanced mode preamp. Until then, I cannot comment on the sonic quality of this amplifier when driven by differential sources. With unbalanced inputs, the DIFF 120 is a solid performer, even in a system with a Krell preamplifier and Martin Logan ReQuest speakers. ❖

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12. Norman E. Thagard, "A Phono Pre-Preamplifier for the CD Era, Part 1" *audioXpress*, 1/01, p. 25.

13. Nelson Pass, "Letters," *TAA*, 2/93, p. 38.

Zen Variations Part 4: The Penultimate Zen

Much spit and polish has been applied to this Zen amp design to maximize its efficiency. **By Nelson Pass**

The Penultimate Zen is the sum of several incremental improvements to the original Zen amplifier of 1994 ("The Pass Zen Amplifier," *TAA* 2/94). Eight years just flies by, doesn't it? These improvements are contained here and in Parts 2 through 3 (July and Aug. '02) of the Zen Variations, and is likely the last version of this amp, although by no means the end of the variations on the theme of single-stage amplification.

In Part 2 we developed an improved active current source load for the single gain device, which is at the heart of the amplifier. Originally designed for the Aleph amplifier series, this current source doubles the output current and significantly lowers the distortion of the circuit over the original constant-current source. Because the original Zen amplifier is limited in both power and fidelity, this is a welcome improvement.

The original Zen amp is also limited in its rejection of power-supply noise, and benefits from having a quiet, stable power supply. In Part 3 we discussed some possibilities for power-supply regulation for this and other Zen amplifiers.

The final shortcoming to address in the Penultimate Zen is the low input impedance. Depending on the version and the desired gain and distortion figures, the original Zen amp has an input impedance that varies from 600Ω to a couple thousand ohms. For many audio sources, this input impedance is simply too low to give optimal performance. The ideal input impedance would be something up around 47kΩ.

We will address this fault here and now, and then go on to present a final circuit with a nice finished printed circuit board design.

UPPING THE INPUT IMPEDANCE

The issue of increasing the input impedance of the Zen amp is a delicate one. There is no way around the need to add circuitry to accomplish this, but in the spirit of the Zen amp, we wish to do it minimally.

The input junction of the gain MOSFET of the amplifier is operated at virtual ground, so that previous Zen amps have an input impedance that is the value of the input resistor between the input connector and the Gate of the MOSFET. You could simply increase the values of the input and feedback resistor, but then you bump into the nonideal nature of the MOSFET.

The Gate of the MOSFET has essentially an infinite input impedance at DC, but offers capacitance between the Gate and Source and the Gate and Drain pins. The amount of the capacitance is not so great and by itself does not impose serious limits on the resistor values. The problem is that this capacitance varies under differing voltage and current conditions, and is thus nonlinear in character, producing distortion at higher frequencies. You can clearly see this effect on the distortion versus frequency curves of the previous Zen amplifiers, where somewhere above the midrange the harmonic distortion begins increasing with frequency. The higher the resistance at the input, the earlier this effect starts.

To have a high input impedance while maintaining low distortion at high frequencies, the amp needs a buffer. The most obvious approach is to use an input follower (*Fig. 1*). Here a small N channel MOSFET is biased up to serve as a Source follower, presenting a very high input impedance, and a fairly low output impedance while following the input voltage. Since we use a small MOSFET for this purpose, the input capacitance is tiny enough not

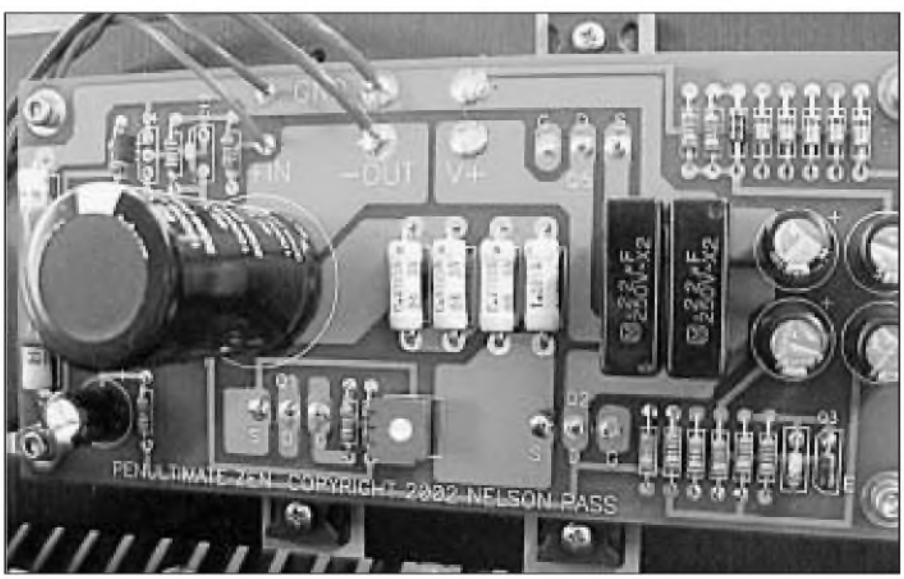


PHOTO 1: The loaded PC board.

to have a significant effect at audio frequencies.

There is another, more clever way to accomplish this follower (*Fig. 2*). Here we use a P channel MOSFET with the Drain connected to ground and the Source biased up by a current source, in this case just a resistor. There are several advantages to this approach.

First, the input can be direct coupled, because it operates at ground potential, eliminating the input capacitor and bias circuit.

Second, it is self-adjusting in the sense that the Source will operate at +Vgs, where Vgs is the Gate to Source voltage for the MOSFET, or about 3.5V. The amount of current the resistor supplies is not critical, and can be quite high if desired, as the MOSFET will only dissipate that current times the Vgs value. For the ZVP3310 you will be using, you could run it as high as 50mA, which is a lot.

Third, the P channel nonlinearity will tend to operate in opposition to the N channel distortion of the gain MOSFET, giving some distortion reduction due to cancellation. You can alter this cancellation a bit by adjusting the bias current and loading the P channel device.

Fourth, because the input system is at virtual ground, you can swing all the drive current you want without being concerned about the voltages across the input device, since it will be held very close to the constant Vgs of the P channel input. No Miller effect, hardly any gain or capacitance modulation, better performance.

Talk among yourselves. I'll give you a topic: Is this still considered a single-gain-stage circuit? If a differential pair or Darlington or a cascode is considered single-stage, would this be considered one also?

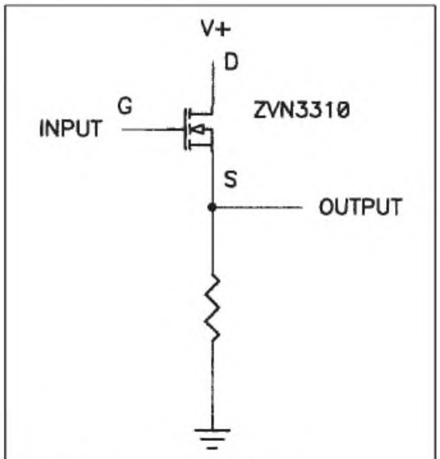


FIGURE 1: N channel follower. A-2185-1

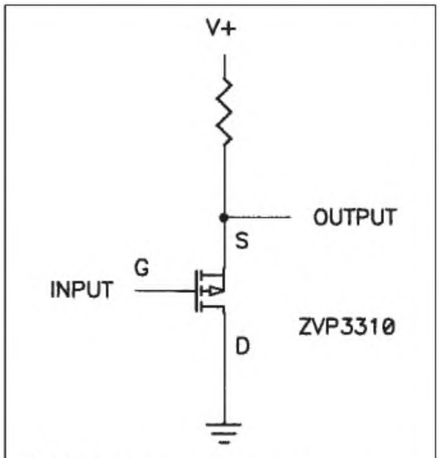


FIGURE 2: P channel follower. A-2185-2

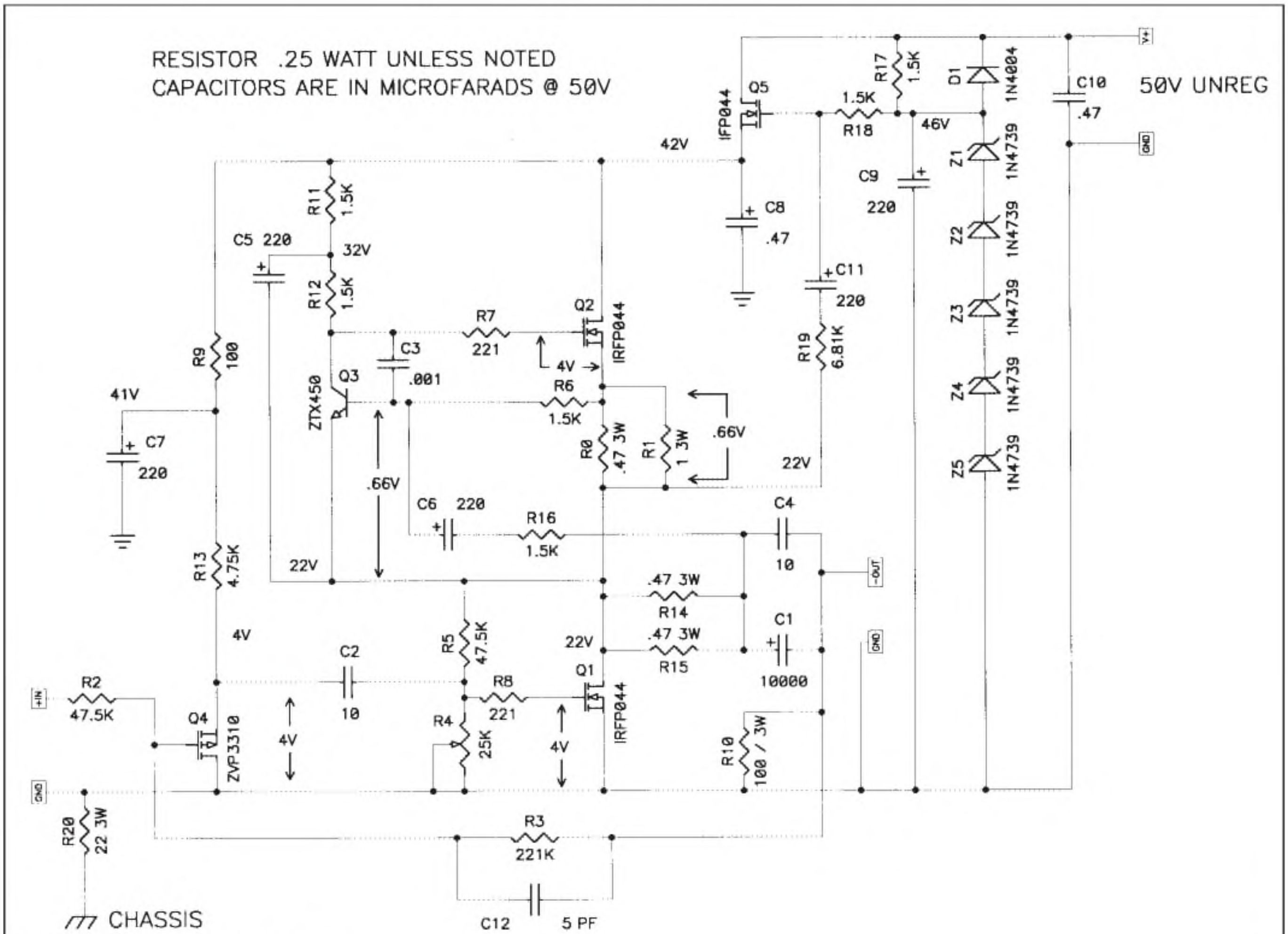


FIGURE 3: Zen amp with input buffer. A-2185-3

THE CIRCUIT

Figure 3 shows the circuit of *Fig. 2* incorporated into the Zen amp with the regulated power supply. Q4 is the added buffer transistor, and it is biased by R13. R9 and C7 filter the voltage provided to R13. The feedback loop of R2 and R3, which previously connected to the Gate of Q1, is now connected to the Gate of Q4 and have values two orders of magnitude higher, giving the amp an input impedance of 47kΩ.

This circuit was designed to idle at 2A, which is more than it actually needs to achieve the desired 25W into 8Ω figure. R0 and R1 together form a .33Ω resistor, and if you leave R1 out, the bias figure drops to 1.3A. With both R0 and R1 and a 50V rail, the amplifier will dissipate 100W, and with R1 removed, the circuit dissipates 67W at idle.

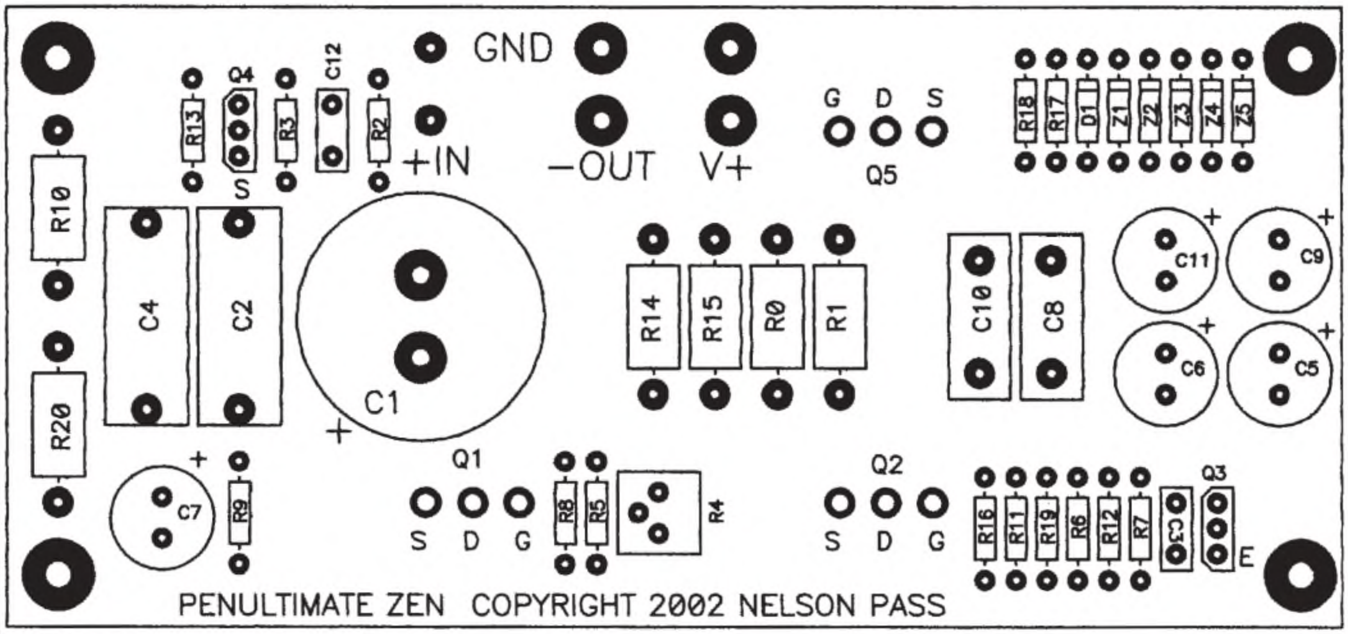


FIGURE 4: PC board component placement.

A-2185-4

Depending on your heatsinking and power needs, you can include R1 or not. If you are planning on driving 4Ω, you will definitely want it. If you leave R1 off, you should also change the value of R16 from 1.5kΩ to 1.0kΩ.

As with all the Zen amps, very few of the component values are critical; in fact, none of them are, so if you have resistors and capacitors that come close to these values, feel free to use them.

As discussed previously, many substitutions will work for the transistors. You can choose Q2 and Q5 from a wide array of IRF-type N channel devices as long as they have the power and current rating to do the job, which would be around 150W and 10+A. They do not have very much influence on the sound compared to Q1 and Q4.

In the case of Q1, the IRFP044 is the preferred part, but you can achieve almost as good performance from the IRFP140 and 240 parts, and pretty much anything else that is similar in character.

In the case of Q4, the Zetex ZVP3310 performed the best among the parts tried. The IRF9510, 9610, and similar parts worked, but have higher capacitance, and thus higher distortion at the top end with these input impedances. You can improve this by lowering the values of R2 and R3. If you can accept a 10k input impedance, you could consider using these alternate P channel MOSFETs.

While we're talking about the input, note that the input of this amplifier is not protected against high voltage transients. Input voltages in excess of 20V will have some chance of damaging the input MOSFET Q4. If this occurs, it should not damage other parts, so if you plan on being careless when connecting the amp, keep a couple of spare ZVP3310s lying around.

Q3 works well with the Zetex ZTX450 NPN transistor, but note that any decent NPN signal device will work here, as you are primarily using it for the .66V junction voltage to control the current through Q2. Like most of the parts here, it was chosen because it is available through Digi-Key.

C4 is simply a bypass capacitor for C1. You may omit it altogether or replace it with your favorite bypass part. C2 can be electrolytic or film (film being preferred) but should be at least 10µF to avoid low-frequency rolloff. If C2 is an electrolytic type, the polarity will not usually be important, as the DC values on either side are approximately the same. If you find that they are not, you will know where to point the positive side, won't you? For more commentary on the main audio circuit, refer to "Zen Variations, Part 2."

C8 and C10 are simply high-frequency bypass capacitors. Usually the circuit will work fine without them, but they provide a nice margin for stability. Similarly with C12, which trims the high-frequency response so as to avoid peaking at high frequencies. Digi-Key no longer stocks 5pF caps, but I have listed 10pF parts in the parts list. You can buy two and put them in series to form a 5pF capacitor.

Alternatively, you can twist two fine insulated wires (like those used in wire-wrap construction) together at a length of about 1.5″ and get about 5pF. This technique has the advantage that you can start out too long and trim it with a pair of scissors while watching the square wave of the amplifier.

Z1 through Z5 are 9.1V zener diodes, which produce a 45.5V reference for the regulator Q5. You can adjust these to different voltages or bypass any of them if you find yourself using a different rail voltage. You can leave them out altogether, in which case the regulated supply will be about 4V less than the unregulated supply, but with the AC ripple removed. This works just fine in practice, and is usually referred to as a capacitance multiplier. For more information on the other parts in the supply regulator, refer to "Zen Variations, Part 3."

CONSTRUCTION

Figure 4 shows the printed circuit board for this project. It measures about 6.5 × 3″ and holds the circuitry shown in *Fig. 3*. The board is available for sale from www.passdiy.com, and you can also download the Gerber files to get your own made.

Photo 1 shows the stuffed board, which does not include the components for an unregulated 50V supply, but *Fig. 5* shows a typical circuit for one channel and the grounding layout for a stereo

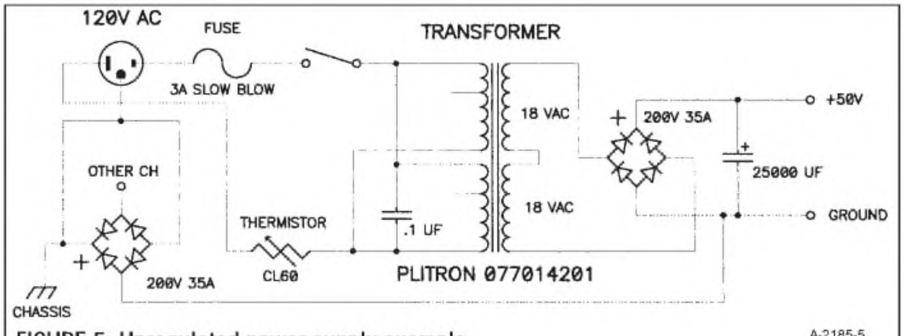


FIGURE 5: Unregulated power supply example.

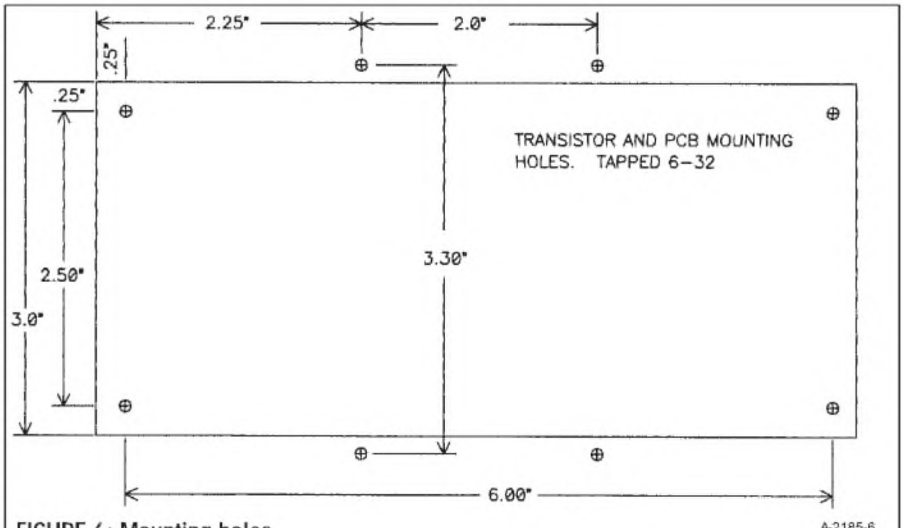


FIGURE 6: Mounting holes.

amplifier. The transformers are Plitron 077014201, with two 18V secondary windings and a 300VA wattage rating. While two transformers are recommended for two channels, we did build an amplifier using only one. It worked well enough, but ran hot and mechanically buzzed a bit at 200W draw.

Note in *Fig. 5* that we have chosen to isolate the two channels through a rectifier bridge to ground, with each channel's ground appearing on one of the AC legs of the bridge. If you use a single transformer supply for two channels, tie both of their grounds together and to both AC legs of the bridge.

If you construct the amplifier with two separate transformers and decide to operate the amplifier balanced or bridged, be certain to tie the ground speaker outputs together with a heavy piece of wire to provide a good low impedance ground connection between both channels, otherwise speaker current will attempt to flow through the input ground cables.

Figure 6 shows the mounting hole dimensions for the board and power transistors. These holes are tapped 6–32 onto a flat surface of a heatsink. You must insulate the power transistors from the heatsink using thermal insulators—either silicone pads or mica and thermal grease. Use a washer between the head of the screw and the case of the transistor to spread the force across the package, and don't tighten them very tight. The ground on each board connects to the heatsink through R20, providing a resistive connection to the chassis.

The heatsinks must dissipate up to 100W per channel, and need to do so with a temperature rise of about 30°C, which gives them a rating of about .3°C/W. If you use smaller sinks, then a lesser bias (no R1) and/or fan cooling will be necessary.

If you manage to cool the amplifier very well, you can consider whether you can get more power out of it. I would say the practical limit to this will be Q1, which is run at about 44W with the 2A bias. Personally, I hesitate to operate this device at greater than 50W, but you can crank the supply voltage and bias up higher if you want. With the IRFP044 I would not exceed 50V on the regulated supply. If you want to try higher, I suggest the IRF140 or 240 devices.

More bias is easily provided by replacing R1 with .47Ω, which will give you 2.8A, enough to probably get 50W into 4Ω and 100W balanced into 8Ω. Beyond this, you can expect some excitement. If you do replace R1 with .47Ω, I suggest R16 at 2kΩ or so.

Table 1 shows the parts list. You can

TABLE 1
ONE-CHANNEL PARTS LIST FROM DIGI-KEY

REFERENCE	DESCRIPTION	PART #
Q1, Q2, Q5	IRFP044 N channel power MOSFET	IRFP044N-ND
Q3	Zetex NPN ZTX450	ZTX450-ND
Q4	Zetex ZVP3310 P channel MOSFET	ZVP3310A-ND
D1	1N4004 diode	1N4004GICT-ND
Z1–Z5	1N4739 9.1V zener diode	1N4739ADICT-ND
C1	10,000µF @ 50V electrolytic	P6939-ND
C2, C4	10µF polyester film	EF1106-ND
C3	.001µF film	4773-ND
C5, C6, C7, C9, C11	220µF @ 50V electrolytic	P1246-ND
C8, C10	.47µF polyester film	E1474-ND
C12	5pF film	399-1806-1-ND (2 in series)
R0, R14, R15	.47Ω 3W metal film	P0.47W-3BK-ND
R1	1Ω 3W metal film	P1.0W-3BK-ND
R2, R5	47.5k .25W metal film	47.5KXBK-ND
R3	221k .25W metal film	221KXBK-ND
R4	25k trim potentiometer	3386P-253-ND
R6, R11, R12, R16, R17, R18	1.5k .25W metal film	1.50KXBK-ND
R7, R8	221 .25W metal film	221XBK-ND
R9	100 .25W metal film	100XBK-ND
R10	100Ω 3W metal film	P100W-3BK-ND
R13	4.75k .25W metal film	4.75KXBK-ND
R19	6.81k .25W metal film	6.81KXBK-ND
R20	22Ω 3W metal film	P22W-3BK-ND

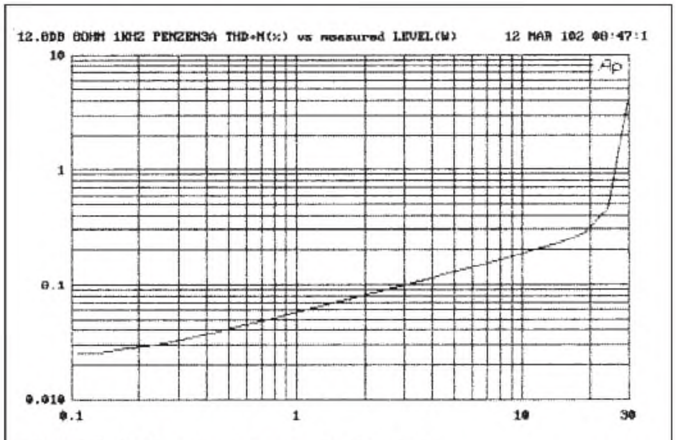


FIGURE 7: THD versus power with 2A bias. A-2185-7

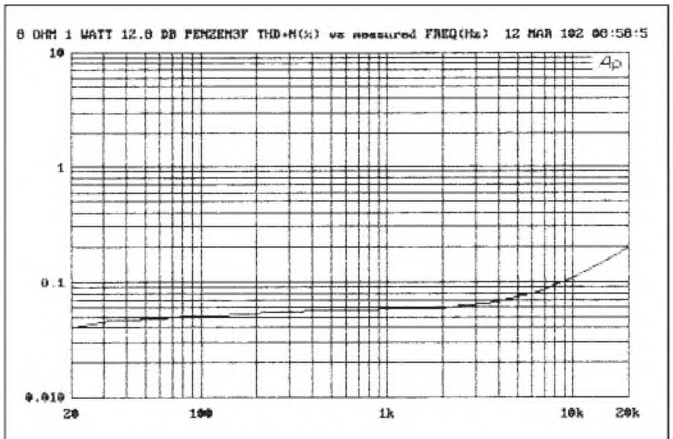


FIGURE 9: THD versus frequency at 1W. A-2185-9

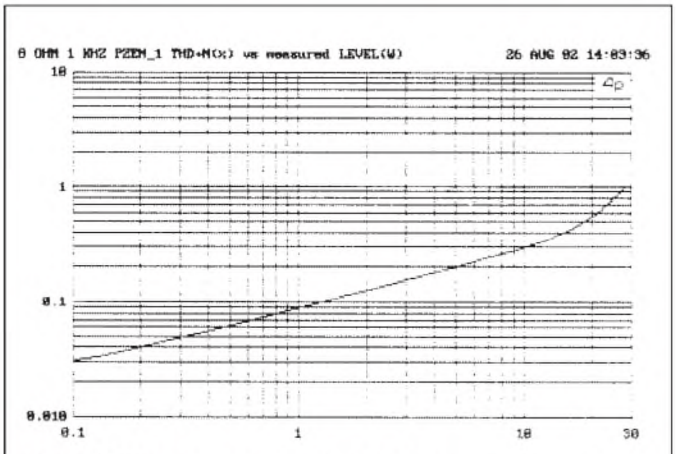


FIGURE 8: THD versus power with 1.3A bias. A-2185-8

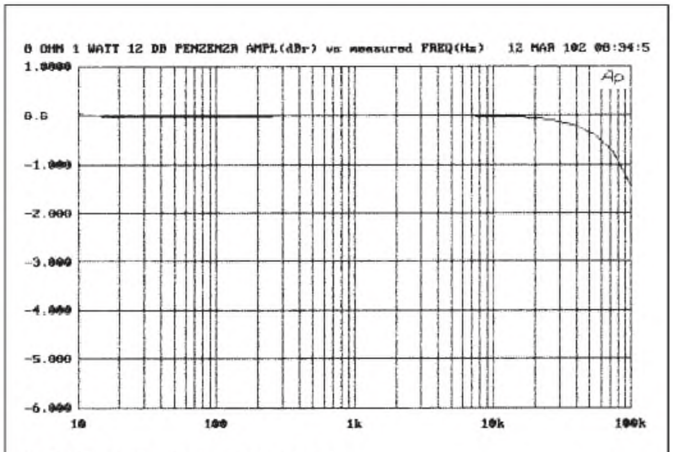


FIGURE 10: Frequency response. A-2185-10

find all the parts except the transformer and mechanical package at Digi-Key (www.digikey.com).

The amplifier is not very susceptible to noise pickup, and you can achieve good results enclosing it in a wooden chassis. If you do, be certain to heatsink the rectifier bridges, which ordinarily get their heatsinking from the chassis. Placing them on the main sinks is OK.

ADJUSTMENT

Assuming that you have triple-checked your assembly work, it is time to "light this candle." Do so one channel at a time connected to unregulated V+. Wear safety glasses. Fuse the transformer at the minimum value, about 1.5A. Set R4 at maximum value, which is counter-clockwise.

Bring the amplifier up slowly using a Variac® and watch the voltage across R0. If it does not exceed .7V or so, then continue to increase the Variac. Adjust R0 so that the Drain voltage of Q1 is half the regulated voltage plus about 2V.

If you get this far without problems, then check for the other voltages called out in the schematic. Readjust R4 after the amplifier has warmed up for a couple of minutes. Now do the same thing for the other channel. Watch where you stick those probes!

If the voltages all check out, then you have a 95% chance that everything is proper. If not, it is time to start doing your detective work. This is best done slowly. Wait till the next day. The vast majority of problems are board stuffing and wiring errors. Look for those at length before picking up a soldering iron. Think about it twice.

When you have the bugs out, run both channels for an hour and check the temperature of the heat every few minutes. Check for burning smells. Readjust R4 after an hour.

PERFORMANCE

In addition to having the high input impedance, this circuit also measures better than the previous editions. Due to the regulated power supply, the noise is about 35μV at the output, a really excellent figure.

The distortion plus noise versus power into 8Ω at 1kHz is shown in *Fig. 7* with the 2A bias. *Figure 8* shows the same curve at 1.3A bias with R1 removed and R16 at 1.0kΩ.

Figure 9 shows the distortion at 1W from 20 to 20kHz, which is dramatically better than any earlier Zen amp. *Figure 10* shows the frequency response, which is totally flat at 10Hz, and down about 2dB at 100kHz.

Like just about any other amplifier, the Penultimate Zen will drive 4Ω also, but at a higher distortion figure. If you include R1, you should get about 30W into 4Ω. If you decide to run a pair of the channels balanced (or bridged) with R1 installed, you can expect to see about 60W into 8Ω, and the performance is very good. As documented in Zen Variations 1 and 2, the distortion tends to drop due to cancellation of the second harmonic, which is the dominant distortion. This mode of operation is highly recommended if you need the most power possible into 8Ω.

CONCLUSION

Sonically, this is about as good as it's going to get for this topology, and I would match it up against an Aleph 3 without hesitation, and indeed the performance is quite similar. Its signature is relaxed and very easy to listen to, although it is not a powerhouse at the bottom end. It works best with 8Ω speakers, 90+dB sensitivity, which do not require a lot of control from the amplifier. In other words, a tube friendly loudspeaker.

Does it sound like a tube amp? It's a little bit of a cross between tubes and solid state, but it has its own character, the result of a very simple MOSFET circuit operated pure Class A.

So there you have it. The amplifier is called the Penultimate Zen because it represents how far we want to take this particular topology without getting medieval.

ACKNOWLEDGMENT

Thanks to Karen Douglass, Wayne Colburn, and Desmond Harrington for their help. ❖

An Against the Wall Speaker

Thin is in . . . for fashion models, laptops, and now speakers. This thin design suggests some interesting future applications. **By John Mattern**

I had three objectives in pursuing a thin speaker design. First, I believed that the effect of the wall on the diffraction step would be lowered by having the speaker hug the wall. In the limit as the thickness becomes very small, it satisfies the large flange assumption that is basic to the simulation software for both the cone and the port openings.

Second, I believe that there is a future in the thin LCD and plasma displays. They have a depth of about 3.5″ at present. A speaker that had a depth of the same amount would harmonize with such a monitor. This particular thin design also has the advantage of placing the drivers at a good height.

The final objective was to verify the latest version of my software that simulates the performance of a back-loaded speaker with fiberglass stuffing. The software has been modified to eliminate the major discrepancies noted in my prior articles ("Another Look at TL Design," *SB* 3/99, and "More on TL Speaker Design," *aX* Feb. '02). The latest simulation software is based on the acoustic treatment of the conical horn by Stewart and Lindsay.[1]

The Stewart and Lindsay equation for velocity potential in a conical pipe defines the forward and backward velocity potential along the length of the pipe as a function of the distance from the apex. However, the equation does not provide for losses. To allow for losses I combined a loss term with the Stewart and Lindsay phase term yielding e^(ikr-ar), where -ar is the loss term that I added ("a" is the loss exponent per meter and "r" is distance in meters). The loss, therefore, is introduced into Stewart and Lindsay's subsequent terms for pressure and velocity.

I used the same mathematics they used in the lossless case to develop acoustic pressure and velocity at any distance from the apex. This also required a major revision in my software. In the end this change allowed the cone and port outputs to cancel each other at

PHOTO 1: Bass speaker assembly with fiberglass installed before attaching front panels and top.

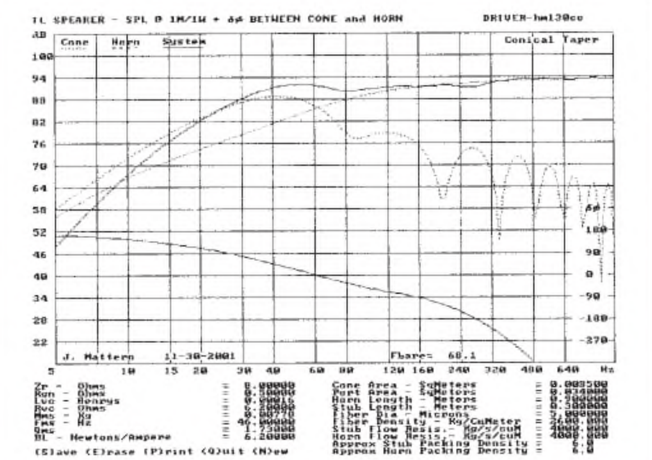


FIGURE 1: Bass speaker simulated response. B-1810-1

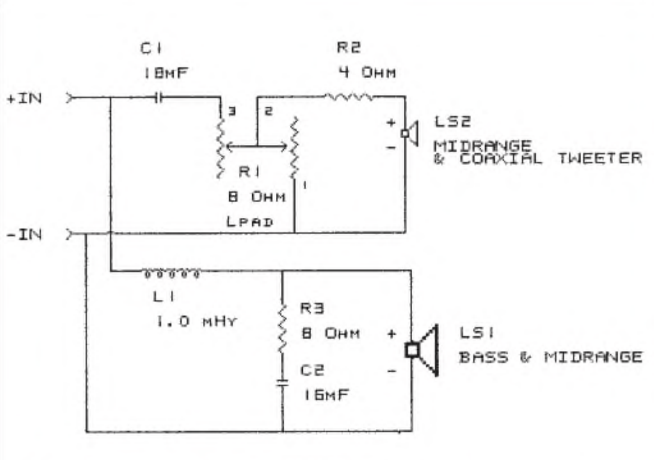


FIGURE 2: Speaker system electrical schematic. B-1810-2

very low frequencies, a necessary condition brought to my attention in correspondence with G.R. Koonce. My early software based on Olson's conical horn treatment[2] did not lead to this result. My effort to introduce losses into Olson's conical horn formula was apparently unsuccessful.

THE TAPERED TRANSMISSION LINE

The acoustic transmission line may be used to control the back wave from a bass driver in a multiple driver system. A tapered transmission line uses a changing cross-section to achieve a particular characteristic. If the taper is flared outward (the mouth is larger than the throat), then there will be gain that can provide more bass output. If the taper is in the other direction, it saves space and diminishes the radiation from the port (should there be one). Whether tapered or not, the line will be filled with a tangle of fibers that selectively absorb the back wave.

The action is largely mechanical. According to Bradbury[3], at high frequencies where the inertia of the fibers renders them stationary there is absorption of sound but little change in speed. At low frequencies where the inertia forces are small, the fiber tangle moves with the air causing a significant reduction in the speed of sound but with little attenuation. At some frequency there is a turning point causing a rolloff with increasing frequency similar to an L-R filter. But in this case the action is mechanical rather than electrical.

There are three factors that determine the filter action: fiber density, fiber diameter, and flow resistance. The flow resistance and packing density are interdependent. Increasing the packing density (compressing the fiber tangle) increases the flow resistance, and, conversely, specifying a higher flow resistance demands a higher packing density. A figure of merit for the fibers may be written as the fiber density divided by the fiber diameter. The greater this number the lower the turnover frequency and the greater the attenuation in the midbass region above the turnover frequency.

I have examined a number of materials, and of these the fiberglass tangle has the greatest figure of merit for use in the conical horn. Unfortunately, fiber-

TABLE 1
SPEAKER SYSTEM ELECTRICAL AND MECHANICAL PARTS LIST

PART	VALUE	SOURCE	DESCRIPTION	MFG PT#	SUPPLIER PT
C1	18.0mF	Parts Express—Solen	400V polypropylene		027-580
C2	16.0mF	Parts Express—Solen	400V polypropylene		027-578
R1	8.0Ω	Parts Express	50W Lpad		260-255
R2	4.0Ω	Parts Express	20W wirewound		017-4
R3	8.0Ω	Radio Shack	20W non inductive		271-120
L2	1.0mH	Parts Express	Air core .21Ω		266-350
LS1	8.0Ω	Parts Express	Carbon fiber midbass woofer	HT 130C0	296-063
LS2	4.0Ω	Boston Acoustics	Midbass and coaxial tweeter		CX3/FX3

QTY	DIMENSIONS	MATERIAL
1	12 × 4 × .5″	Fiberboard (CO base—with cutout for Lpad)
1	12 × 6 × .5″	Fiberboard (CO panel)
12	#6-32 × 1.25″	Brass screws (3 nuts/screw)

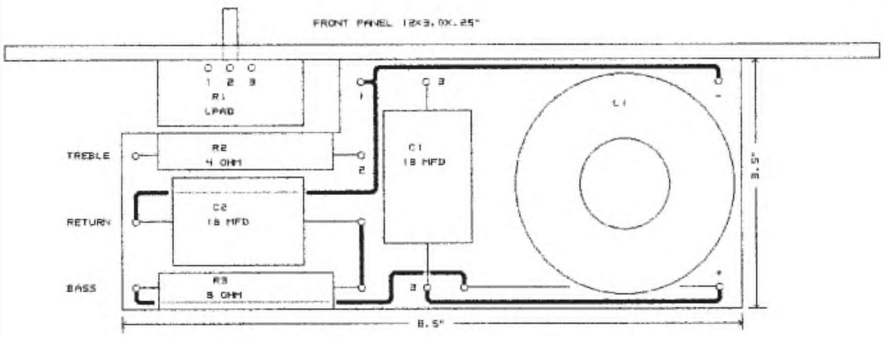


FIGURE 3: Crossover parts mounted on 3.5 × 8.5 × ⅝″ plywood board with 1 × 3″ cutout for Lpad.
B-1810-3

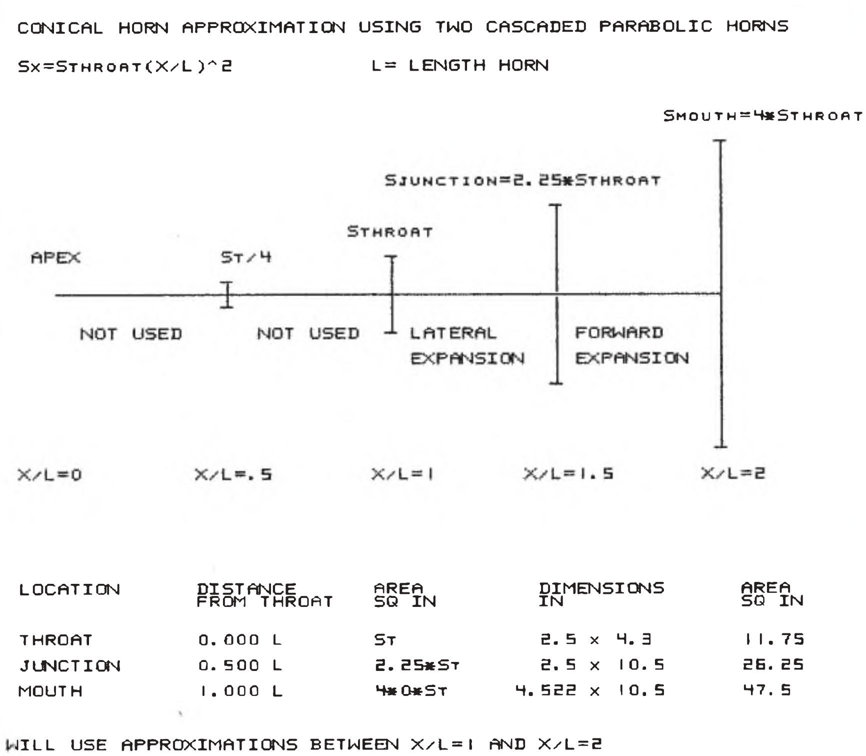


FIGURE 4: Calculation showing approximation to conical horn using two short cascaded parabolic horns.
B-1810-4

glass is the least predictable, because the fibers are anything but uniform. Bradbury gives the diameter as .005mm. The fiberglass tangles I have examined appear to have a large spread in fiber diameter as well as length. Acousta-stuf, in contrast, is quite uniform but has a lower figure of merit as a result of its larger diameter and light weight.

The design of a tapered transmission (or horn-loaded speaker, if you prefer) will benefit from simulation software that predicts the approximate results to be expected. The simulation software I am now using is the result of many years of casual study backed up by system construction for experimental verification. Only now has the agreement between theory and practice been reasonably good.

SIMULATION

Since the simulation precedes the experimental phase, I will start with it first. *Figure 1* shows the predicted performance for an optimized enclosure using the Audax HM130CO. This is an upscale 5.25″ mid woofer featuring a fully ventilated woven carbon fiber cone. This figure shows the cone response, the port response, the phase difference between the cone and port, and the vector sum of the two responses.

I obtained the driver parameters from the data sheets in the Solen catalog, and the fiberglass diameter from Bradbury's paper. I obtained the fiberglass density from *the ITT Engineers Handbook*.[4] It differs somewhat from that given by Bradbury.

The simulation suggests that, with a port-to-cone-area ratio of four or greater, the port response can look very much like that of a subwoofer. I have verified the predictions of *Fig. 1* using the enclosure parameters listed on the lower right side of the figure.

CROSSOVER

I chose to stay with the simple 6dB per octave crossover for both the bass and treble speakers. It served me well in the previous design and I could see no reason to change it. The requirement that the bass and treble speakers have substantial overlap is still satisfied, and the treble speaker, a nearly full-range coaxial unit, can handle what little bass is fed to it. I do intend to try the Infinity Kappa 3.5″ car speaker, which has even more bass capability than the Boston Acoustics FX3, an upgraded version of the CX3.

There were more pressing issues to resolve, so I simply will repeat the earlier crossover (CO) design. I omitted the inductance and resistance used in the original CO to correct for the diffraction step because they were not needed. This is not to say that there are no diffraction edge effects because there still are. But I will cover this later.

Table 1 provides the CO parts list. The electrical schematic for the CO is shown in *Fig. 2* and the mechanical arrangement is shown in *Fig. 3*. I did not enclose the CO this time but used a wide front panel to hide the parts and, for convenience, placed the assembly on top of the treble speaker.

I have reservations about any speaker that uses a bipolar electrolytic as part of the CO. Both the CX3 and FX3 use a 3.3mF electrolytic capacitor in series with the tweeter. I have successfully modified the wiring of both types so a quality external capacitor replaces the original internal one. *aX* willing, I will report the details in a follow-up article.

CONICAL HORN APPROXIMATION

The simulation assumes a conical horn connecting the back of the cone to the port. This, of course, implies a smooth

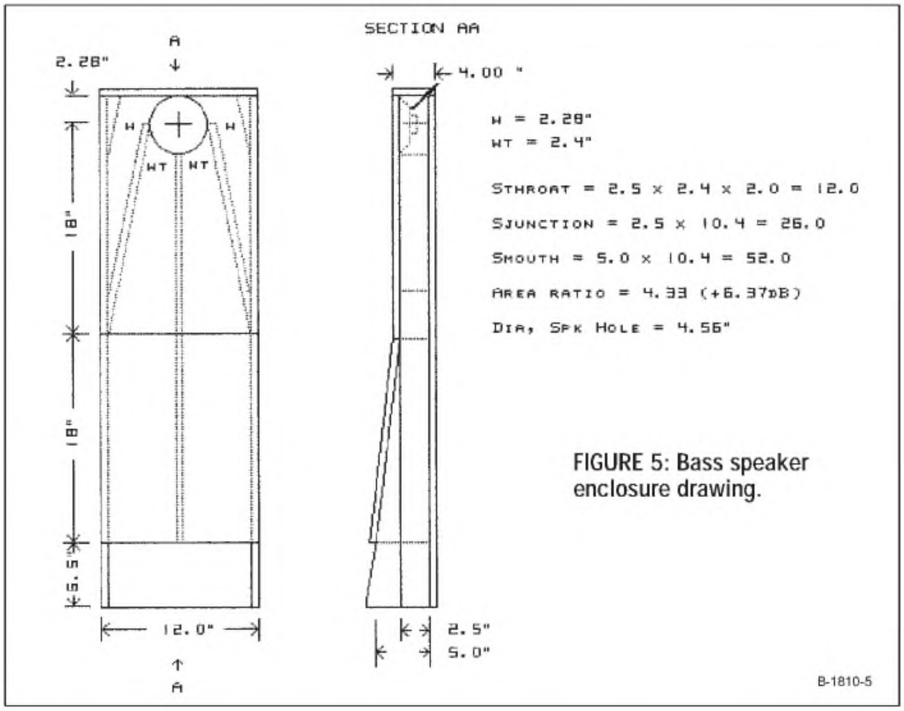


FIGURE 5: Bass speaker enclosure drawing.

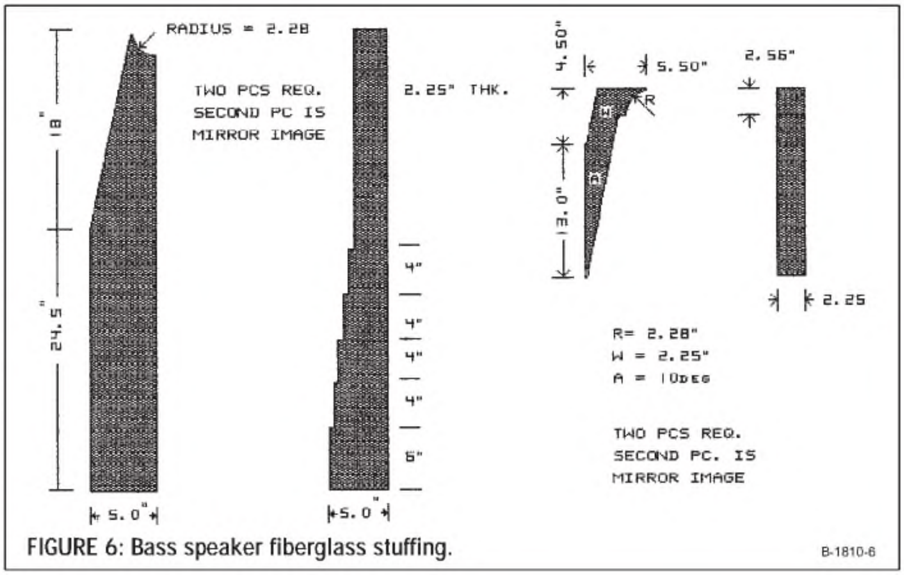


FIGURE 6: Bass speaker fiberglass stuffing.

change in dimensions in two directions, a difficult design to execute. For a 4:1 area ratio the difference between the conical horn and the parabolic horn is not great. The parabolic horn can be formed with a uniform change in one direction only.

Figure 4 shows how a conical horn can be approximated by cascading two parabolic horns to closely approximate a conical horn. Of course, were the area ratio much greater than four, this simple solution would not suffice. The example in the diagram approximates a conical horn between Sthroat and Smouth for an area ratio of four. This is the basis for the claim that the thin speaker is back loaded by a conical horn as assumed in the simulation.

BASS SPEAKER DESIGN

I took the bass speaker dimensions from the simulation plot of *Fig. 1.* I had difficulty meeting the requirement that the cone area, the stub area, and the horn mouth area be equal. The reason becomes apparent when you examine the speaker drawing of *Fig. 5.* Note that the position of the horn's mouth is not well defined. I did the best I could, and I think the discrepancy is less than 10%.

The front and back of this enclosure uses ¾″ particleboard shelving. It is called MDF but is more dense than what usually passes for MDF.

The outside pieces connecting the front and rear surfaces are 5⁄16″ thick hardwood. Their depth is 2.5″ in the top half and linearly increases from 2.5″ to 5.0″ in the bottom half. The internal pieces are only ½″ thick. The bottom halves are a composite of a straight piece and a wedge.

I use hardwood for two reasons. First, it is much stiffer than an equal sized piece of MDF. Second, the hardwood will hold the screws attaching the lower front panel more securely.

I thought it was a good plan to make the lower front panel removable. I glued the upper front half trapping the stuffing, and fastened the lower front half with screws on 3″ centers. Of course, you can glue the lower half once you have verified the performance of the stuffing. *Photo 1* shows a partially constructed bass speaker with the fiberglass stuffing installed. The final speaker uses stuffing surrounding the installed driver, not shown in the photo.

CONSTRUCTION OF THE BASS SPEAKER

The material list is provided in *Table 2.* Purchase shelving that is long enough so you won't need a splice in the back. After cutting it to length, you will need to prepare the side pieces and internal partitions. I used scraps from Home Depot's scrap box for these pieces. You may prefer to salvage your parts from oak flooring. After cutting the back to size, I penciled in the plan from *Fig. 5* to the back panel for use in locating the five internal parts.

PHOTO 2: Treble speaker assembly with fiberglass installed before attaching back panel.

I used a jig to cut the slopes for the internal hardwood pieces. The jig itself required work with a saw followed by a plane to render it accurate enough for the task. With a radial arm saw, you might clamp the work at an angle and then move the saw through the work.

I made sure that the hardwood pieces were true and then brushed on a uniform coating of carpenter's glue. After clamping them to the back I wiped the excess glue away leaving a fillet of glue on the inside joints. I clamped the work in three places while the glue dried.

The three internal pieces should touch the 4.56 diameter speaker circle. Being careful to assure an airtight joint, you glue the three internal pieces to the back. This is a good time to fabricate the corner pieces and attach them. Remember this is the back and the only hole you will make is through the top for the wire.

I then made the 24.5″ wedges that I glued to the bottom of the side pieces. I used a thinner ½″ piece for the inside brace. I used a jig to fashion the bottom half of the outside pieces as well as the inside brace. Some ingenuity is required to construct this speaker, but, believe me, it is far more easily built than the first TL.

You will want to cut the 4⁵⁄₁₆″ speaker mounting hole in the top front panel. I did this in two stages. First, I used a 4½″ hole saw and followed it with a drum sander to open the hole until the speaker fit.

I then used the router to cut a 45° chamfer on the back side, making the chamfer about ¼″ deep. You will probably do this before cutting the top front panel to size. You may wish to leave a sliver of material at the top of the speaker cutout. The speaker fits on the outside of the front panel.

BASS SPEAKER STUFFING AND FINAL CONSTRUCTION

The stuffing is a bit "iffy," because Radio Shack has discontinued their fiberglass blanket. I know of two other materials: the Home Depot ¾″ pipe insulation and their 3½″ home insulation. In the first instance, you will need to attach three layers together with a very thin layer of spray adhesive to form a

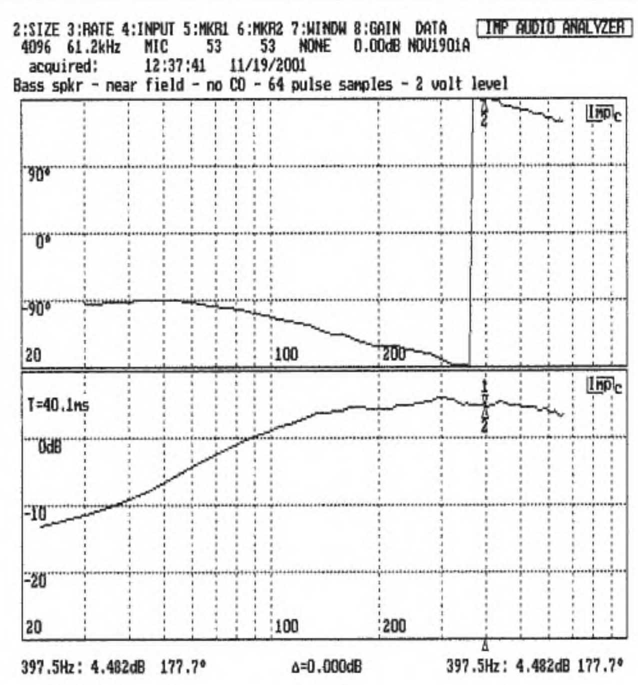


FIGURE 7: Bass speaker near-field cone response and phase characteristic without CO. B-1810-7

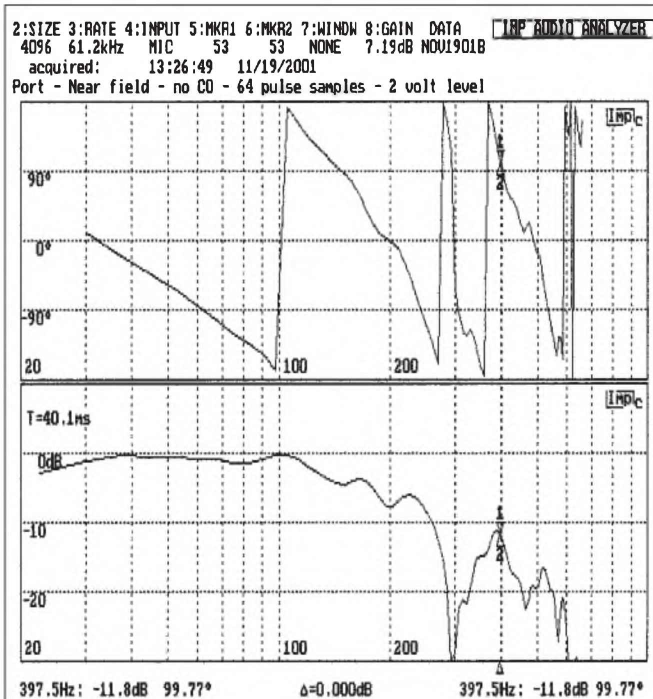


FIGURE 8: Bass speaker near-field port frequency response and phase characteristic without CO. B-1810-8

2.25″ laminate. In the second case, you will need to remove a 1¼″ layer from the insulating blanket, but getting a neat result may be difficult.

You may want two patterns for cutting the four pieces of stuffing. Plans for the patterns are shown in *Fig. 6*. The speaker here used the Radio Shack fiberglass, so the steps you take will be somewhat different.

First, I used the pattern to make the two long pieces, which extend to the floor. I made the side pieces in two steps. I fabricated the uniform 42.5 × 2.0″ piece and then glued it in place between the back and side pieces with a thin coat of carpenter's glue. Then I glued the top front panel in place with carpenter's glue with the chamfer on the inside, being careful to avoid air leaks.

Then I made the fiberglass "wedges" that fit under the lower front panel. I attached the fiberglass wedge with spray adhesive and attached the bottom front panel with screws on 3″ centers using closed cell weather strip to seal the joints. When you are using thinner material you may want to use more pieces, but shorter ones. The idea is to not quite fill the space between the front and back panels.

Finally, I glued the top on and drilled a slanted hole through it close to the edge for the speaker wire. This allows you to place the speaker against the wall because the treble speaker box is not as deep as the bass speaker. I did not allow for the room's base board in the drawing of *Fig. 5*, since I will use my speaker in an area without one.

I used Home Depot pipe insulation in the second unit. This fiberglass comes in sheets that are 16 × 48 × .75″. I cut out the pieces and then glued three pieces together with a very thin coat of spray adhesive, giving a nominal thickness of 2.25″. The three-piece laminate at the back extends to the floor. This insulation also comes in a narrow ½″ roll, which may be of some use.

MEASUREMENTS

I made all measurements using an early IMP audio analyzer on indefinite loan from G.R. Koonce. The test location was my one-car basement garage, with dimensions of roughly 8 × 10 × 24′. I placed the speaker against the inside of the closed door near its center. The side walls near the door were covered with 7″ thick fiberglass. I covered the floor near the speakers with 6″ of foam absorber. This was needed even when making near-field measurements to reduce nearby reflections.

TABLE 2
BASS ENCLOSURE MECHANICAL PARTS LIST

QTY	DIMENSIONS	MATERIAL
1	12 × 22 × ¾″	Fiberboard shelving (top front)
1	12 × 18¼ × ¾″	Fiberboard shelving (bottom front)
1	12 × 44¾ × ¾″	Fiberboard shelving (back)
2	44.8 × 2.5 × [illegible]/16″	Hardwood (side)
2	24.5 × 2.5/0 × [illegible]/16″	Hardwood (tapered side addition 7.91° slope)
1	33.7 × 2.5 × ½″	Hardwood (internal brace)
2	24.5 × 2.5/0 × ½″	Hardwood (tapered side addition 7.91° slope)
2	18 × 2.5 × ½″	Hardwood (partition with 10° taper)
2	4.25 × 2.5 × .8/0″	Wood (tapered corner piece with 10° taper)
1	12 × 4 × ¾″	Fiberboard shelving (top)

I set the analyzer window to 40ms to eliminate the rear wall reflection, which was very strong. With a steady 40Hz excitation I found the sound stronger in the center of the room than at the speaker cone! The 40ms window allowed the computation to go below 30Hz. I used the test probe to calibrate the system at the output of the DC-coupled amplifier that drove the speaker under test with a 2V test pulse.

BASS SPEAKER PERFORMANCE

Figure 7 shows the cone response of the Audax driver. The test was near-field with the speaker placed against the garage door. The cone response at 60Hz is down about 10dB. This is in good agreement with *Fig. 1*, which shows a 9dB loss. *Figure 8* shows the port response at 40Hz to be down about 5dB. This is in good agreement with the prediction in *Fig. 1*, which shows about a 6dB loss at 40Hz.

I measured the port pressure at the exit opening that is somewhat larger than the horn mouth, hence the 7.2dB area ratio gain correction rather than 6dB. The measurements pertain to the first speaker using the Radio Shack fiberglass stuffing. A comparison of *Figs. 7* and *8* shows the port to be in phase with the cone at 60Hz. This also is in good agreement with the prediction in *Fig. 1*. The second speaker using Home Depot pipe insulation stuffing provided similar results.

TREBLE SPEAKER DESIGN

The mechanical design for the treble speaker is shown in *Fig. 9*. This is a true transmission line with a reverse taper and no port. I omitted the port because I wanted eventually to try other 3.5″ car speakers and did not want to be tied to a length that depended on the treble unit's resonant frequency.

As you will see later, the sensitivity of the Boston FX3 is adequate and the rise in impedance at resonance is well enough controlled so that a simple 6dB rolloff works well. I feared that the rise in impedance at resonance might be a problem with a simple 6dB per octave bass rolloff. Certainly the electrical damping would be compromised by the high impedance of the series CO capacitor.

In fact, I built an active CO and compared the treble unit's performance with the two types of COs. The measured and audible differences were negligible so I kept the passive CO.

I had several objectives to meet with this design. First, I wanted a thin enclosure to match the bass enclosure in width. Second, I chose to locate the tre-

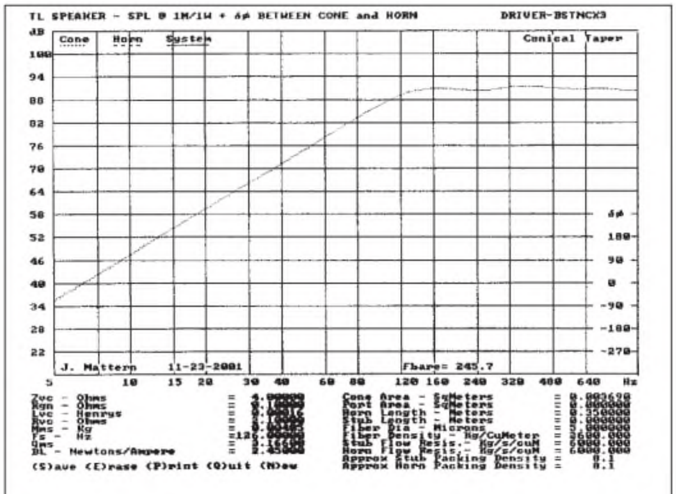


FIGURE 11: Treble speaker simulated response. B-1810-11

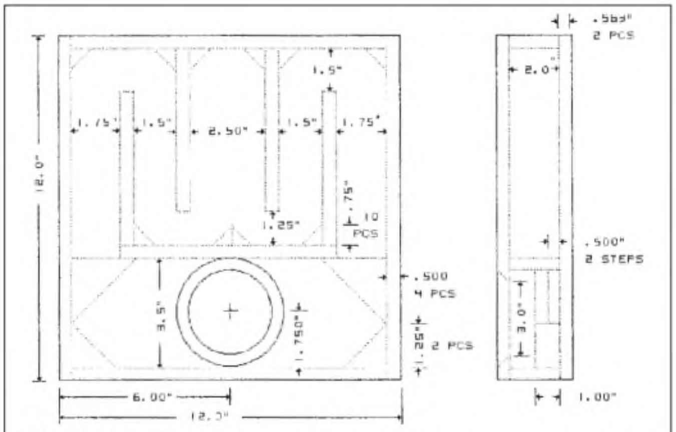


FIGURE 9: Treble speaker enclosure drawing. B-1810-9

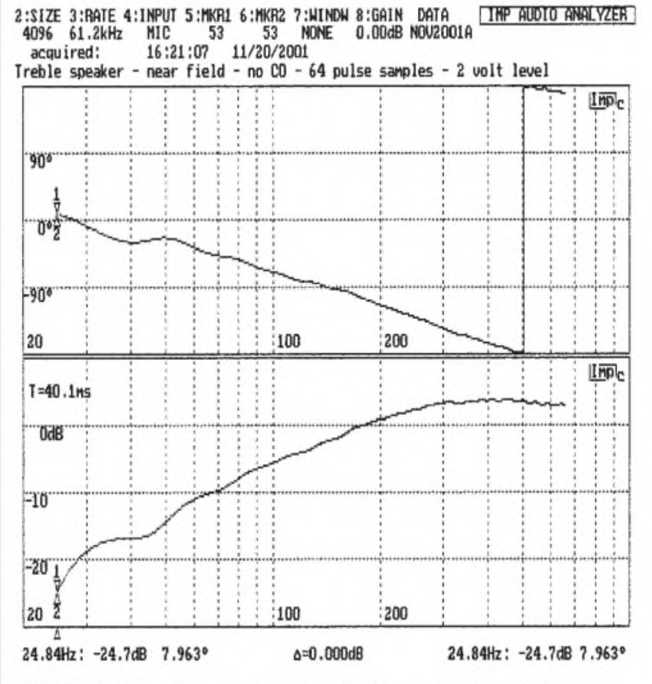
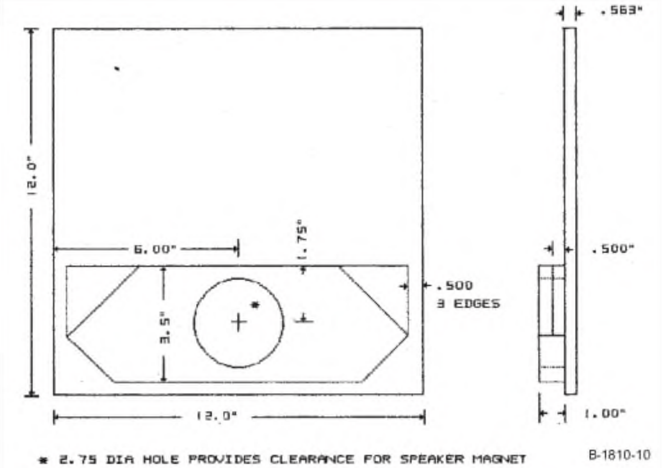


FIGURE 12: Treble speaker near-field response and phase without CO. B-1810-12



* 2.75 DIA HOLE PROVIDES CLEARANCE FOR SPEAKER MAGNET B-1810-10

FIGURE 10: Treble speaker enclosure back panel subassembly.

ble driver as close as possible to the bass driver. Third, I wanted a balanced acoustic load on the back side of the driver. And fourth, I wanted the initial transmission line area to equal the cone area.

The last item required a 1″ fill with a cutout for the magnet. The fill is attached to the back panel as shown in *Fig. 10*. The shallow 1″ TR feeds into a narrow TR having a depth of 2″. The transition is accomplished by a 45° step, with a depth of 1.5″. The remaining passages are narrower but deeper. Do not confuse this step with the 45° corner piece. There are two large corner pieces in the speaker section and ten other smaller corner pieces.

TABLE 3
TREBLE SPEAKER ENCLOSURE PARTS LIST

QTY	DIMENSIONS	MATERIAL
1	12 × 12 × 5/16″	MDF (front)
1	12 × 12 × 5/16″	MDF (back)
2	12 × 2 × ½″	Hardwood (sides)
2	11 × 2 × ½″	Hardwood (top and bottom)
1	7.5 × 2 × ½″	Hardwood (internal horizontal)
1	11.0 × 3.5 × ½″	Hardwood (internal filler—see *Fig. 9*)
2	5.5 × 2 × ½″	Hardwood (internal outside vertical)
2	5.75 × 2 × ½″	Hardwood (internal inside vertical)
2	1.25 × 2 × 1¼″	Wood (45° corner piece)
10	.75 × 2 × .75″	Wood (45° corner piece)
14	#6 × 1¼″	Steel wood screws

TABLE 4
TREBLE SPEAKER ENCLOSURE STUFFING—CORNERS PER *FIG. 9*

QTY	DIMENSIONS	MATERIAL
1	11 × 3.5 × 1″	Fiberglass blanket (2-.5″ layers)
2	7.5 × 1.75 × 2.0″	Fiberglass blanket (2-.75 + 1-.5″ layers)
2	7.0 × 1.5 × 2.0″	Fiberglass blanket (2-.75 + 1-.5″ layers)
1	7.0 × 2.5 × 2.0″	Fiberglass blanket (2-.75 + 1-.5″ layers)
1 pkg	.75 × 5/16″	Closed cell foam weatherstrip

TREBLE SPEAKER CONSTRUCTION

The front and rear panels are made from 5/16 high density fiberboard. I used ½″ fiberboard for the side pieces as well as the inside pieces, but I highly recommend using hardwood instead because the screws hold so much better. The gap between the front and rear panels is 2.0″. Details of the major parts are provided in *Tables 3* and *4*.

I used the same procedure to assemble the treble speaker as I did with the bass speaker; that is, I made a drawing on the inside of the front panel locating all the 2″ pieces and the driver cutout. The next step I should have done before but didn't—that is, make an identical drawing on the outside of the back panel as an aid in locating the screw holes. Since the speaker is mounted on the inside, it is desirable to have the back removable. I then cut the 3.0″ hole for the driver in the front panel and cut a 45° chamfer with my router on the outside of the panel. You will want to leave an ⅛″ section of wall without a chamfer.

I used the pencil layout on the front panel to place the parts, starting with the outside parts first. I glued all pieces including the ten small corner pieces and the two large corner pieces to the inside of the front panel a few pieces at a time because of the limited number of available clamps. The rear panel is attached with screws on 3″ centers with closed cell weather stripping providing the seal. *Photo 2* shows the entire assembly just before closing it. This enclosure is very sturdy and is virtually immune to speaker-induced vibrations, as is the bass speaker.

PHOTO 3: Unfinished speaker system with CO at top, set up against garage door for tests.

TREBLE SPEAKER PERFORMANCE

Figure 11 shows the predicted low frequency response using a modified simulation program that omits the parts that do not apply to a TL without a port. I obtained most speaker parameters using the IMP analyzer to measure the Thiele/Small parameters, which I converted to the alternate parameters, except for BL, which I am able to measure directly at home. *Figure 12* shows the measured low frequency response of the treble speaker. The two are in good agreement at 40Hz.

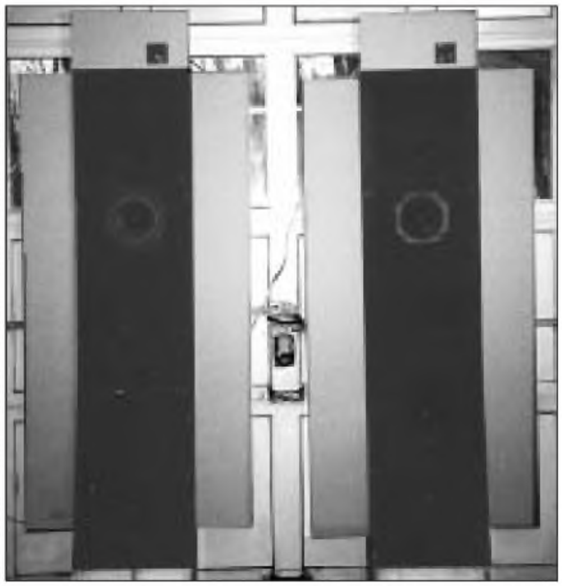


PHOTO 4: Stereo pair with diffraction-reducing side boards.

COMBINED BASS AND TREBLE PERFORMANCE

Figure 13 shows the magnitude and phase of the bass speaker with the CO in the path. *Figure 14* shows the magnitude and phase of the treble speaker with the CO in the path. I made both measurements near-field with the window set to 6ms. The measurement system gains were equal and the Lpad setting was maximum. The design crossover is 1250Hz.

The phase difference at this frequency was 90° as required. The dip at 900Hz appeared to be related to edge diffraction, since the size was reduced with a window smaller than 6ms. *Photo 3* shows the completed speaker against the inside of the garage door. I abandoned outdoor testing because of the noisy environment.

EDGE DIFFRACTION

The stereo speaker configuration as tested is shown in

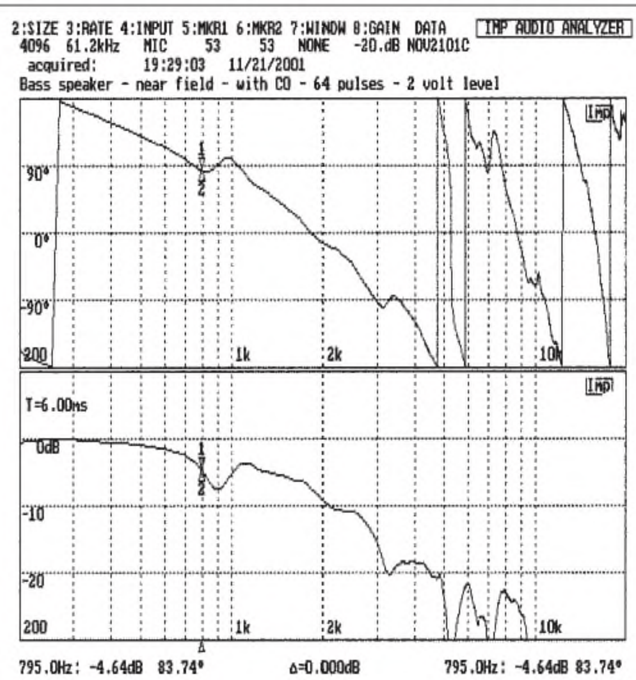


FIGURE 13: Bass speaker near-field response and phase with CO.

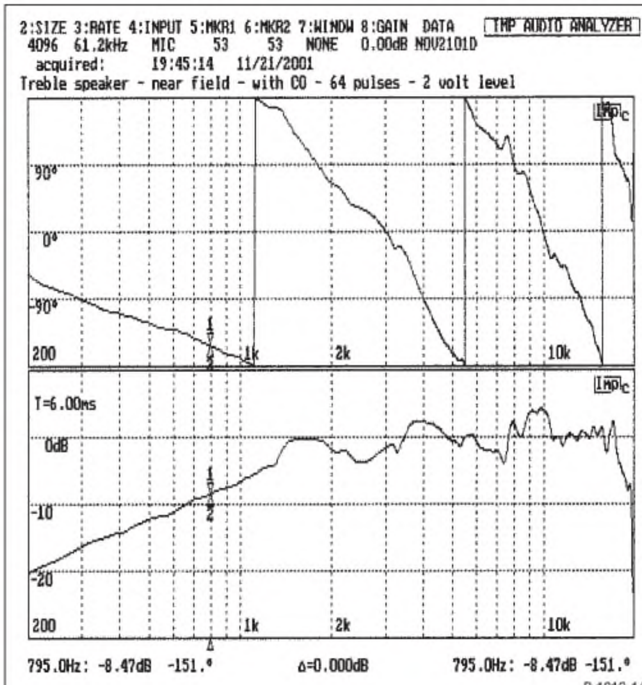


FIGURE 14: Treble speaker near-field response and phase with CO.

Photo 4. In Beranek's[5] "acoustic measurements" his illustration of the diffracted wave from the edges of the speaker cabinet gave me an idea for canceling the wave with a second diffracted wave delayed 180° beyond the first. The measurements did show partial cancellation. The first peak and first dip, near 1000Hz, were reduced to half their value without the canceling panels. Of course, this works only if the outside panel ends are tight against the wall.

The panels are 6″ in width and the front surface is set back 2″ from the front surface of the main body of the speaker. There are two edge pieces giving the pieces a shallow U-shaped cross-section. The center contains fiberglass to suppress resonance. The depth of the panel and sidepiece is 2″. I think sloping pieces might be more effective, but I was dissuaded by the greater construction difficulty.

SUBJECTIVE EVALUATION

Finally, I replaced the *SB* 3/99 speakers in the basement club room with the flat ones. I chose the following sources for the listening tests because I am familiar with the sound of a symphony orchestra and because it is difficult to reproduce.

1. The *Nutcracker* with the Kirov Ballet on a Philips DVD.
2. Beethoven adagio from piano concerto No. 5 on Silverline DVD Audio.
3. Murray Perahia's Mozart on Sony VHS.
4. Bruckner Symphony No. 8 by the Israel Philharmonic on Image DVD.
5. Excerpts from Tchaikovsky's Ballets on Silverline DVD Audio.

In comparing the new speaker with the old one, I noted that the biggest difference was the cleaner highs. I attribute this to the changes in the FX3 and possibly to reduced diffraction effects. The bass was warmer than in the garage, but in the clubroom a tad more powerful than the old speaker. I have wondered about the effect of the garage door on the sound! In the clubroom the speakers are up against a 12″ cement block wall. The new speaker's midrange definitely seemed more transparent.

In comparing the five sources, I would give the highest marks to the VHS and the lowest to the CD. The CD could not carry the peaks without audible distortion, particularly in the first ballet selection. All the DVDs produced fine sound even on peaks but lacked the polish of the VHS tape. ❖

ACKNOWLEDGMENT

I would like to thank G.R. Koonce for the loan of the IMP audio analyzer as well as his invaluable comments on my earlier simulation mistakes.

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Design an Active Transient-Perfect Second-Order Crossover

After constructing and testing an active circuit, this author shares his design rules for actively implementing transient-perfect, second-order crossovers. **By John Kreskovsky**

In my article, "A Transient-Perfect Second-Order Passive Crossover" (May '01 *aX*), I presented the basic theory behind the development of transient-perfect second-order crossovers using symmetric, overlapped, low Q, second-order filters with equalization based on a simple second-order resonant circuit. The crossover yielded a summed response that had flat amplitude and zero phase shift, and hence was transient perfect. The design of the filters and equalization was based on a series of computer-optimized results from which a set of design graphs were presented.

In this installment I have returned to the analysis of the transfer functions and developed a set of design rules (equations) for determining the required filter and equalizer parameters as a function of the crossover frequency and overlap between the high-pass and low-pass filters. I have also developed a spreadsheet-based CAD tool for the design of active transient-perfect second-order crossovers that you may download from the web.

ANALYSIS

I begin with the assumption that a filter can be constructed such that

$$T_{EQ}(s)\,(T_{HP}(s) + T_{LP}(s)) = 1 \qquad [1]$$

Here $T_{EQ}(s)$ is the transfer function of the equalization network, $T_{HP}(s)$ the transfer function of the high-pass filter, and $T_{LP}(s)$ the transfer function of the low-pass filter. I further assume that the transfer functions of high-pass and low-pass sections are those of symmetric, overlapped second-order filters. Defining *s* as

$$s = j\omega/\omega_o \qquad [2]$$

where ω_o is the crossover frequency, and an overlap parameter, γ, such that the corner frequencies of the high- and low-pass filters are given, respectively, as

$$\omega_{HP} = \omega_o/\gamma \qquad [3a]$$

and

$$\omega_{LP} = \gamma\omega_o \qquad [3b]$$

The transfer functions for the high-pass and low-pass filters can then be expressed as

$$T_{HP}(s) = \frac{(\gamma s)^2}{1 + s\gamma/Q_f + (\gamma s)^2}, \qquad [4]$$

and

$$T_{LP}(s) = \frac{1}{1 + s/(\gamma Q_f) + (s/\gamma)^2}, \qquad [5]$$

Q_f is the Q of the filter section in the usual sense. Please note that the definition of γ is different here from that which I used in the original paper. The present γ is equal to the square root of gamma from the original work.

The transfer function of the equalization network is given as

$$T_{EQ}(s) = \frac{1 + sG/Q_{eq} + s^2}{1 + s/Q_{eq} + s^2} \qquad [6]$$

In Equation 6 G is the gain at resonance and Q_{eq} is the Q of the resonant circuit.

Equations 4, 5, and 6 are then substituted into Equation 1. After some mathematical manipulation you end up with the rather ominous equation,

$$[(\gamma s)^2\,(1 + s/(\gamma Q_f) + (s/\gamma)^2) + (1 + s\gamma/Q_f + (\gamma s)^2)]\,[1 + sG/Q_{eq} + s^2] = (1 + s/(\gamma Q_f) + (s/\gamma)^2)\,(1 + s\gamma/Q_f + (s\gamma)^2)\,(1 + s/Q_{eq} + s^2) \qquad [7]$$

We can multiply this out and obtain an equation of the form

$$a_0 + a_1S + a_2S^2 + a_3S^3 + a_4S^4 + a_5S^5 + a_6S^6 = b_0 + b_1S + b_2S^2 + b_3S^3 + b_4S^4 + b_5S^5 + b_6S^6 \qquad [8]$$

For the equality of Equation 8 to be satisfied, which will render Equation 1 satisfied with the transfer functions given by Equations 4, 5, and 6, the coefficients of like-order terms on the left- and right-hand side of Equation 8 must be equal. For example, a_0 must equal b_0. The task at hand then is to determine values for Q_f, Q_{eq}, and G, for a specified value of γ that satisfies the necessary constraints. After a little effort the following relationships emerged:

$$Q_{eq} = \frac{1}{(\gamma^2 - 1)}, \qquad [9]$$

$$Q_f = \frac{\gamma}{(\gamma^2 + 1)}, \qquad [10]$$

$$G = 1 + 1/\gamma^2 + 2/(\gamma^2(\gamma^2 - 1)) \qquad [11]$$

A plot of these relationships is shown in *Fig. 1.* Note the gain asymptotes to 1.0 as γ goes to infinity, and infinity as γ approaches 1.0. Q_{eq} also asymptotes to infinity as γ approaches 1.0, and 0.0 as γ approaches infinity. Q_f varies between 0.5 at $\gamma = 1.0$ and 0.0 as γ goes to infinity. I recommend that γ be restricted to between 1.5 and 2.0.

A schematic of an active circuit for the crossover is shown in *Fig. 2.* The component values for the filter and equalizer sections are given as follows:

For the high-pass (HP) section, for a given value of C_{HP},

$$R1 = \gamma/(4\pi f_x C_{HP} Q_f)$$

$$R2 = \gamma Q_f/(\pi f_x C_{HP})$$

For the low-pass (LP) section, for a given value of R_{LP},

$$C1 = Q_f/(\pi\gamma f_x R_{LP})$$

$$C2 = 1/(4\pi\gamma f_x R_{LP} Q_f)$$

and for the equalizer section, for a given value of R_e,

$$Le = Q_{eq} R_e/(2\pi f_x)$$

$$Ce = 1/(2\pi f_x Q_{eq} R_e)$$

$$Re1 = (G - 1)R_e$$

The inductor in the RLC resonate circuit can be replaced by an active inductor as shown in *Fig. 3*.

TESTING

I developed a spreadsheet-based CAD tool for designing the active crossover. A sample design, generated using the spreadsheet, is shown in *Fig. 4* for the case where $\gamma = 2.0$ and the crossover frequency, f_x, is 1kHz. The low Q behavior of the HP and LP sections is evident. The filter section Q_f is 0.4. The equalization for this case has a Q_{eq} of 0.333 with 3dB gain. The equalized HP and LP responses are also shown.

The actual spreadsheet also shows the phase response of the filters, as well as reference first-order and second-order Linkwitz-Riley filter functions, and computes the required values for the filter components shown in *Figs. 2* and *3*. You can download it from the Internet at: http://www.pvconsultants.com/audio/frdgroup.htm.

When using the spreadsheet, the values of C_{HP}, R_{LP}, and Re are specified and values for R1, R2, C1, C2, Ce, Le, and Re1 are computed. You must also specify Re2, and the spreadsheet then calculates the values of Re3 and Ce2 for the active inductor. I chose values of 0.22µF for C_{HP}, 2.21kΩ for R_{LP}, 4.42kΩ

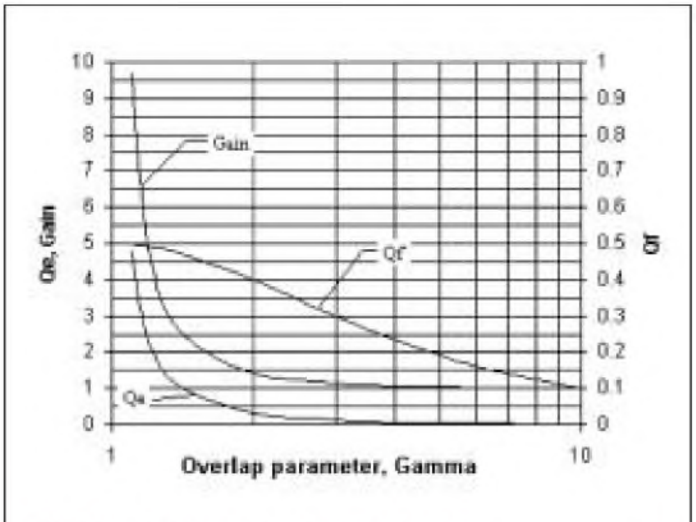

FIGURE 1: Variation of Q_{eq}, Q_f, and gain with γ.

B-2089-1

for Re, and 2.21kΩ for Re1. This yielded the following values for the remaining components:

R1 = 1.809kΩ
R2 = 1.157kΩ
C1 = 0.0222μF
C2 = 0.0347μF
Ce = 0.1080μF
Le = 234.5mH
Re3 = 2.21kΩ
Ce2 = 0.048μF

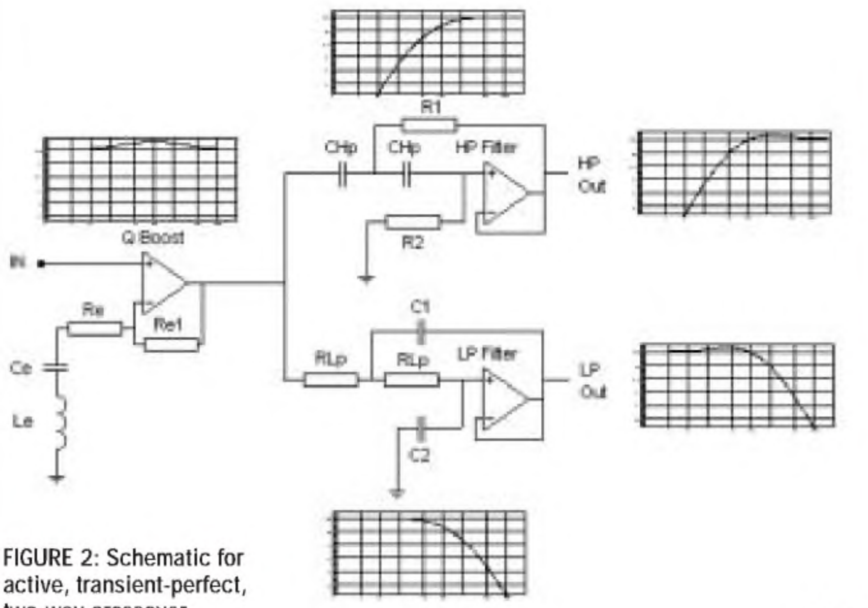

FIGURE 2: Schematic for active, transient-perfect, two-way crossover.

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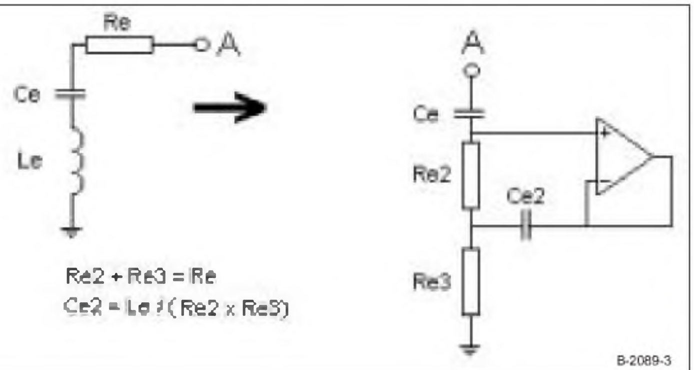
FIGURE 3: Schematic for eliminating the inductor, Le, from the active circuit.



B-2089-3

I then constructed the circuits shown in *Figs. 2* and *3* around the Burr-Brown OPA2134 dual audio operational amplifier using metal film resistors with standard values as close to the CAD-generated values as possible. Polypropylene capacitors were used, paralleled to obtain the necessary values. One 2134 was used for the active equalization and active inductor; the second 2134 was used for the high-pass and low-pass filters. The circuit was powered by a ±12 power supply salvaged from a circa 1980 RACAL computer modem.

I made measurements of the frequency response at the output of the equalizer circuit and at the output of the high- and low-pass filters using the Audio Tester[1] with a sine-wave sweep. The measured results are shown in *Fig. 5* and are in excellent agreement with the CAD design. I also measured the response of the system to a 250Hz square wave as generated using the Audio Tester.

The measured response of the high-pass, low-pass, and the summed output is compared to the input signal in *Fig. 6*. *Figure 7* shows the summed output and the input when the frequency is raised to 800Hz. Both figures show that the crossover retains waveform integrity within the limits of the generating/measuring system.

This transient-perfect nature of this type of crossover also lends itself directly to three- (or more) way crossovers

REFERENCES

Audio Tester, Shareware by Ulrich W. Muller, http://www.sumuller.de/audiotester/.

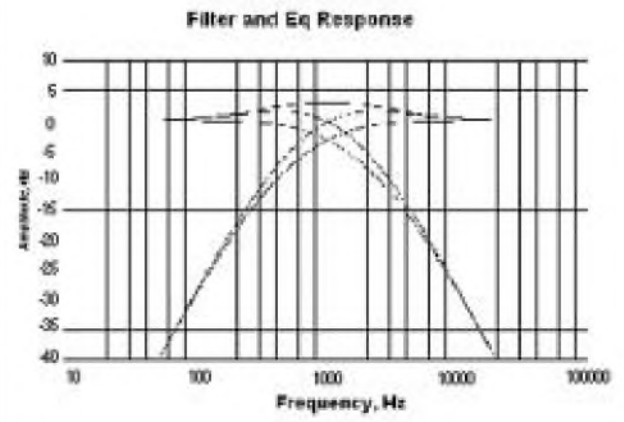


FIGURE 4: Sample filter design for an overlap parameter of 2.0 and crossover frequency of 1kHz.

B-2089-4

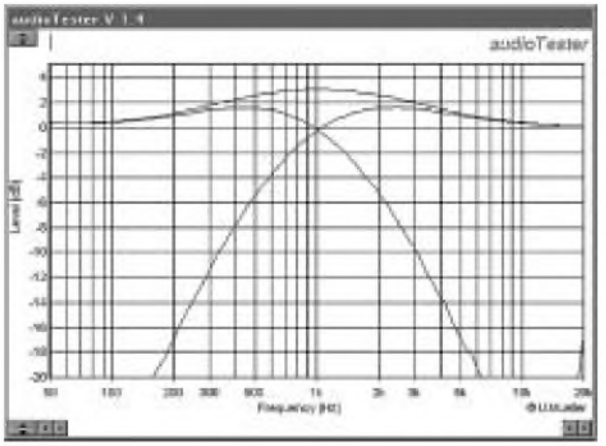


FIGURE 5: Measured response of the CAD designed sample filter. Upper traces are amplitude response. Lower traces show phase.

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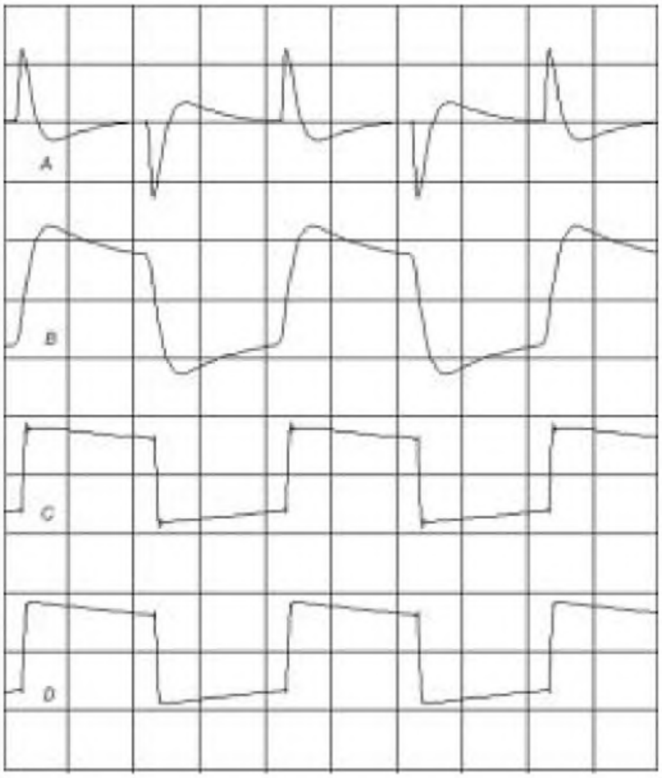

FIGURE 6: Measured 250Hz square wave response. A: high-pass output, B: low-pass output, C: summed output, D: input signal. Note: vertical scale varies for different sweeps.

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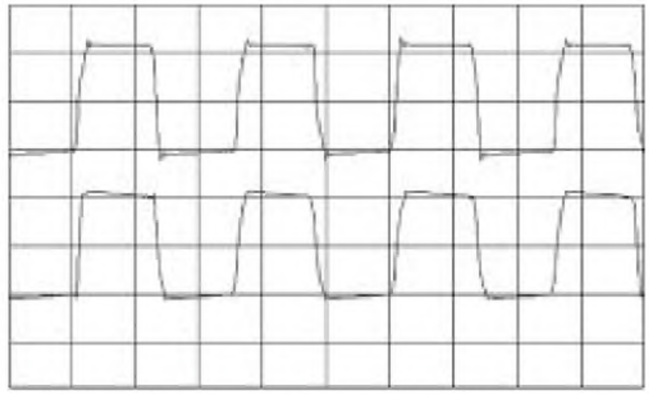

FIGURE 7: 800Hz square wave response. Upper trace: summed output, lower trace: input.

B-2089-7

with little additional effort. For example, the crossover can be cascaded without affecting the transient-perfect nature of the response. A second HP/LP/EQ block can be placed at the output of the HP or LP section of a previous stage to produce a three-way transient-perfect crossover that also has flat summed amplitude and phase response. For additional information, refer to the May '01 issue of *audioXpress*. ❖

6C33C-B Push-Pull Power Amplifier II

Some things are better the second time around . . . like this new, improved power amp from one of our tube enthusiasts in France.

By Daniel J.F. David

As you probably already know, hi-fi lovers are perfectionists. So it's no surprise that after publishing my article ("Triode Power Amplifier 6C33C-B," *GA* 2/99), remarks from certain friends encouraged me to continue studying and improving this amplifier, especially increasing its power, perfecting a lower-voltage phase inverter, and adding a new output transformer (*Photos 1–3*).

PREAMPLIFICATION

I kept the same preamplifier stage but used 12AT7 tubes (*Fig. 1*). With these the gain is 40, less than with the 12AX7, but with a larger input swing. The preamplifier and the phase inverter com-

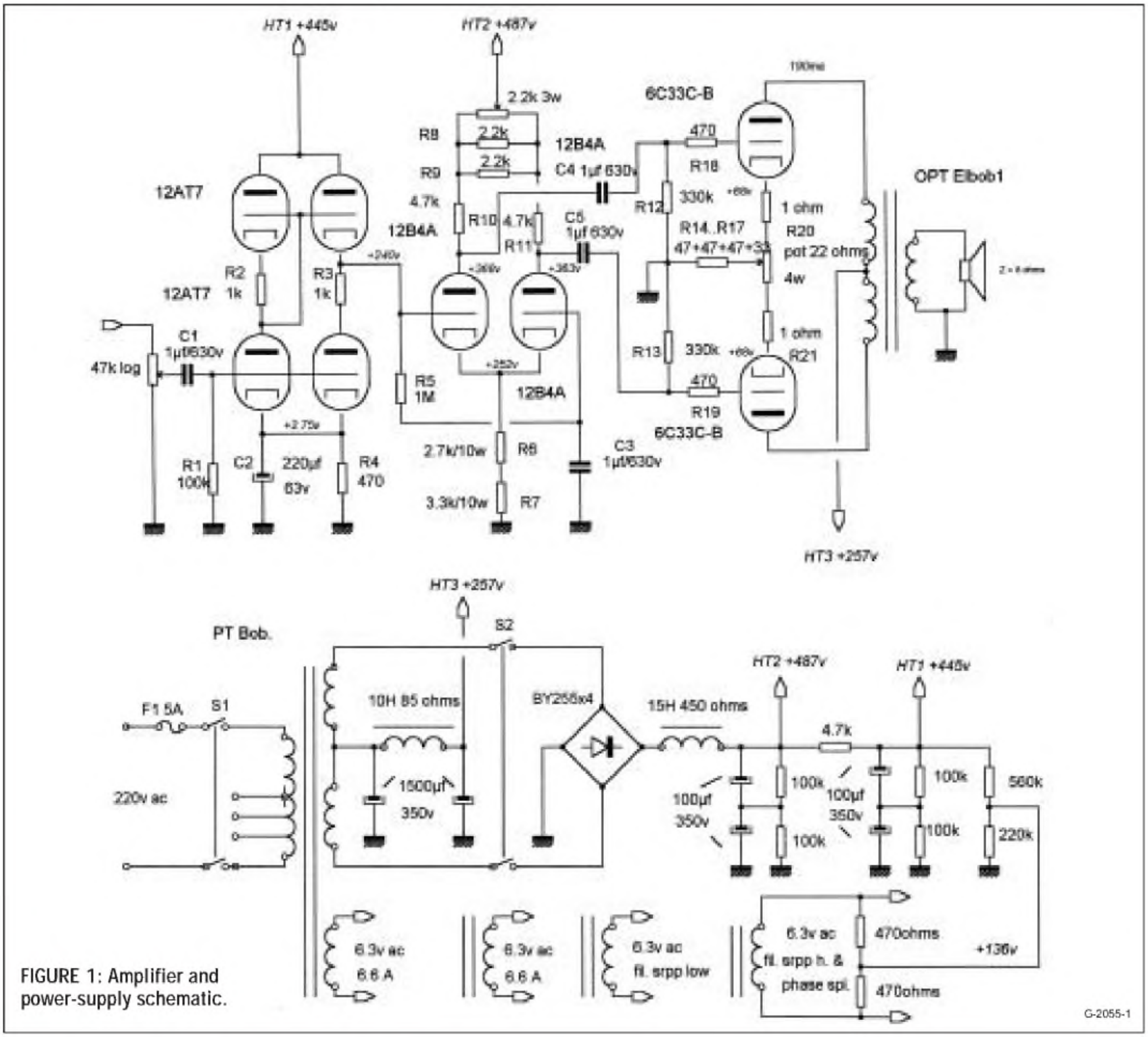

FIGURE 1: Amplifier and power-supply schematic.

bined have a 108 gain, which is more than sufficient. The preamplifier is composed of a shunted-regulated push-pull (SRPP) and a directly coupled cathode charge stage.

I discovered this schematic a few years ago as described by José Gomez in an issue of the *L'Audiophile* magazine. After having experimented with it, he allied a bass output impedance, a graduated distortion spectrum composed of only odd harmonics (regularly graduated from low to high), and a high gain and output swing.

The V1 and V2 tubes function as a normal SRPP. If the V1 receives a positive alternance, its grid becomes less negative, the V1 plate voltage decreases (the internal resistance of the V1 decreases, as well), and the R2 terminal voltage increases. A more and more negative voltage is applied to the V2 grid (the internal resistance of the V2 increases) and at the same time the same increasingly negative voltage is applied to the V4 grid. The voltage of the V4 cathode decreases as well.

The input signal is applied to the V3 grid. The grid voltage of this tube is less and less negative (the internal resistance of the V3 decreases), and the plate voltage of the V3 decreases. You now have a constant voltage on R3, which is of a constant ohm value, equal to a constant current in the R3.

Theoretically, a constant-current source has infinite impedance. Therefore, the V4 has a cathode load with infinite impedance (in alternate). R3 in AC has a resistance, which is superior to its ohm value. In theory, the higher the triode load, the more linear the stage.

SCHMITT CATHODE-COUPLED PHASE INVERTER

I needed to use a high voltage with the cathode-coupled phase inverter in order to obtain a direct liaison with the preceding stage. Despite its qualities, I had to rethink this stage, turning to the Schmitt phase inverter. I used two 12B4As for a gain of 2.7 without distortion and 170V peak-to-peak (pp) output with a voltage of 487V, conserving a direct liaison with the preceding stage. All the resistors in this stage are 11W, non-inductive.

The balance of the phase inverter is

PHOTO 1: The revised 6C33C-B power amp.

done with a 2.2k 4W potentiometer. You may use a voltmeter or a scope, adding the two signals and setting the signal to minimum.

POWER STAGE

The power stage is identical to my previous amplifier. The 25W cathode resistors are set up on an aluminum radiator. Since thermal dissipation is always a problem, a larger ventilated radiator, occupying the entire base of the amplifier, is being studied.

The stage adjustments can be made with an audio-frequency (BF) generator at 40Hz. A distortiometer is connected to a load of 8Ω at the output of the amplifier. Set the 22Ω potentiometer for minimum distortion.

Or as a second solution, you can connect two voltmeters to the 1Ω resistors. Set the two voltages to the same value.

OUTPUT TRANSFORMER

I studied a new mains transformer, which determines the final result of the amplifier. It was built by Bobinelec, a very competent Alsacien (France) winding company. I tried several transformers before finding a definitive and efficient model. The output transformer bandpass goes from 10Hz to 40kHz without any attenuation, and its distor-

TABLE 1
AMPLIFIER PARTS LIST
(DOUBLE ALL PARTS FOR STEREO)

AVAILABLE FROM CONRAD ELECTRONIC:

RESISTORS	
R1, R22. R23. R24, R25, R30, R31	100k, 2W, MO
R2, R3, R24	1k, 2W, MO
R5	1MΩ, 2W, MO
R6	2.7k, 10W, ceramic
R7	3.3k, 10W, ceramic
R8, R9	2.2k, 2W, MO
R10, R11	4.7k, 10W, ceramic
R12, R13	330k, 2W, MO
R14–R16	47Ω, 25W, wirewound, no inductive
R17	33Ω, 25W, wirewound, no inductive
R4, R18, R19, R26, R27	470Ω, 2W, MO
R20, R21	1Ω, 2W, MO
R28	560k, 2W, MO
R29	220k, 2W, MO
POTENTIOMETERS	
P1	47k, pot ALPS, log
P2	2.2k, 3W, pot, linear
P3	22Ω, 4W, linear
CAPACITORS	
C1, C3, C4, C5	1µF/630V MKS
C2	220µF/63V
C5–C10	100µF/350V electrolytic

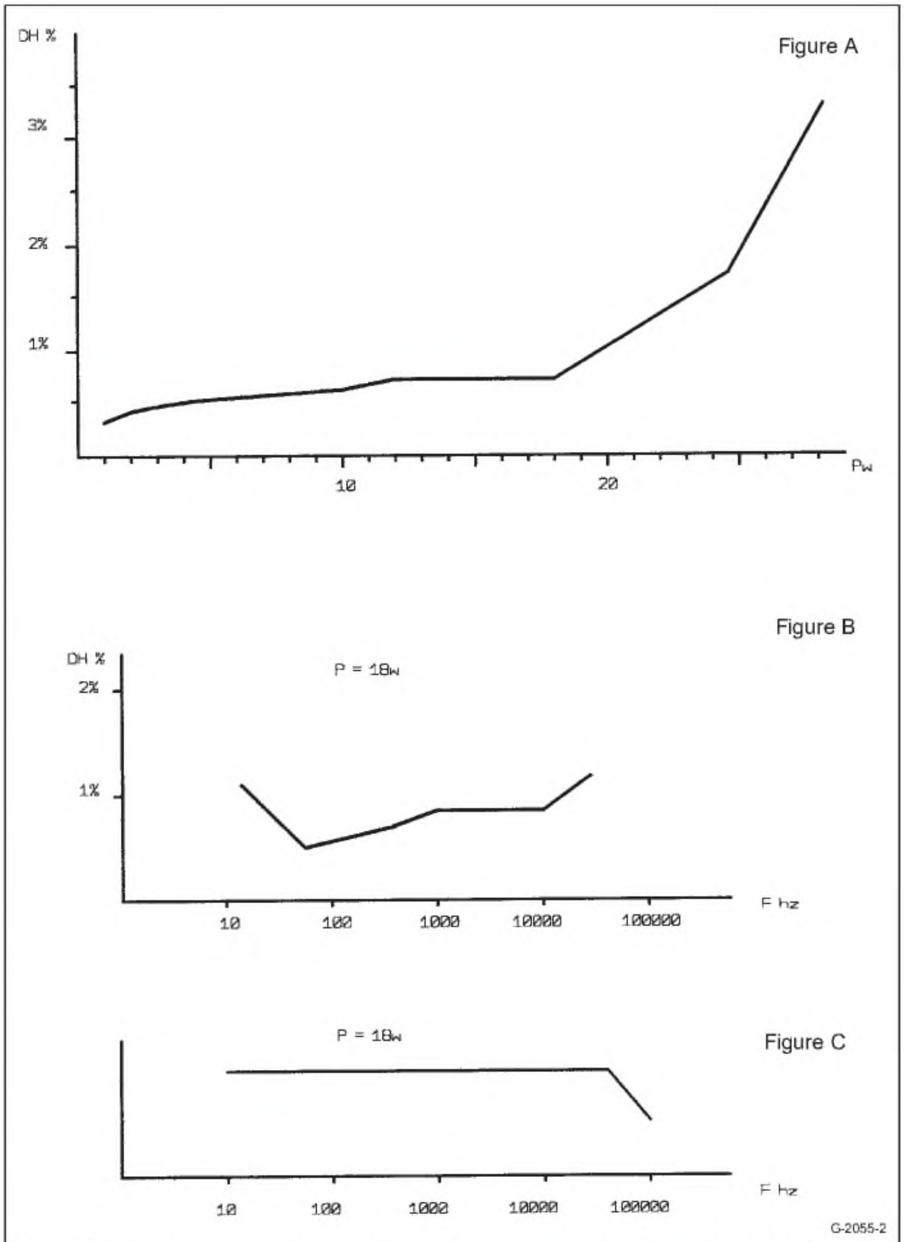

FIGURE 2: Distortion measurements (A and B) and output transformer bandpass (C).

FIGURE 3: Table of winding arrangements.

G-2055-3

tion rate goes from 10Hz to 40kHz, always below 1%, up to 20W (*Fig. 2*).

Finally, this transformer is easily reproduced at a competitive price. It is mounted on an EI magnetic circuit (105 × 126) of steel silicon sheets with oriented crystals. The theoretical primary inductance value of the transformer is 4.5H with a bandpass attenuation of 3dB to 10Hz with 1047 primary turns.

Due to winding constraints, 1092 primary turns are necessary to improve results of the lower frequencies. This was verified by electronic measuring.

The primary coil (*Fig. 3*) is sectioned into six parts and interlaced with six identical parallelized secondary coils. The leakage inductance is equally low and allows you to easily obtain high frequencies up to 40kHz. The transformation ratio automatically maintains the output power at 25W maximum in order to keep the total distortion rate below 1%.

POWER SUPPLY

The power-supply transformer was specially built by Bobinelec. It is composed of several primary coils, which allow us to accurately adjust the filament voltage on the 6C33C-B to 6.3V.

SOURCES

A.B.L
Z.A. Lou Quiquilhan
30260 GAILHAIN
France
33-4-66-80-14-34
FAX 33-4-66-80-14-34
output transformer

ANTIQUE ELECTRONIC SUPPLY
6221 S. Maple Ave.
Tempe, AZ 85283
480-820-5411
FAX 480-820-4643
info@tubesandmore.com
tubesandmore.com
aluminum frame

BOBINELEC
9 rue A.Kastler "Ila Vigie" BP 20
67451 Otswald Cedex France
FAX (as dialed from the US) 33-3-88-67-32-27
Also: ACEA 6 rue Fraçois Verdier
31830 Plaisance du Touch France
FAX 33-5-61-86-61-89
power transformer

CHELMER VALVE COMPANY, LTD
The Stables Baddow Park
Great Baddow
Chelmsford Essex CM2 7SY
44-1245-241300
FAX 22-1245-241309
Sales@chelmervalve.com
www.chelmervalve.com
6C33C-B and sockets

The voltages are given only as an indication and can vary depending on the tubes used. At the start-up both switches are off. Turn on S1 for five minutes to pre-heat the tubes, then turn on S2 for the high voltage. The temperature of the tubes is stabilized after 20 minutes.

CONSTRUCTION

The amplifier is built on a P-H129 aluminum frame supplied by Antique Electronic Supply. The 6C33C-B supports are slightly elevated to help make thermal dissipation easier. The cathode resistors of the power stage are mounted on an aluminum radiator (*Fig. 4*).

As usual, the wiring should be done with care so as to avoid hum, oscillation, and other unpleasant problems. Electrical efficiency should prevail over esthetic efficiency, especially with this powerful amplifier. The filaments of the 6C33C-B use 6.6A each, so wire them away from the audio-frequency generation (BF signal) wires! Don't forget to twist these wires to lower the magnetic field they can produce. As a usual precaution, use one ground per stage. These grounds should be wired together and connected to the frame at one point.

LISTENING RESULTS

With more power this unit represents one step further in quality. With better sound definition it achieves a softer, warmer, more melodious sound. The

PHOTO 2: Rear view.

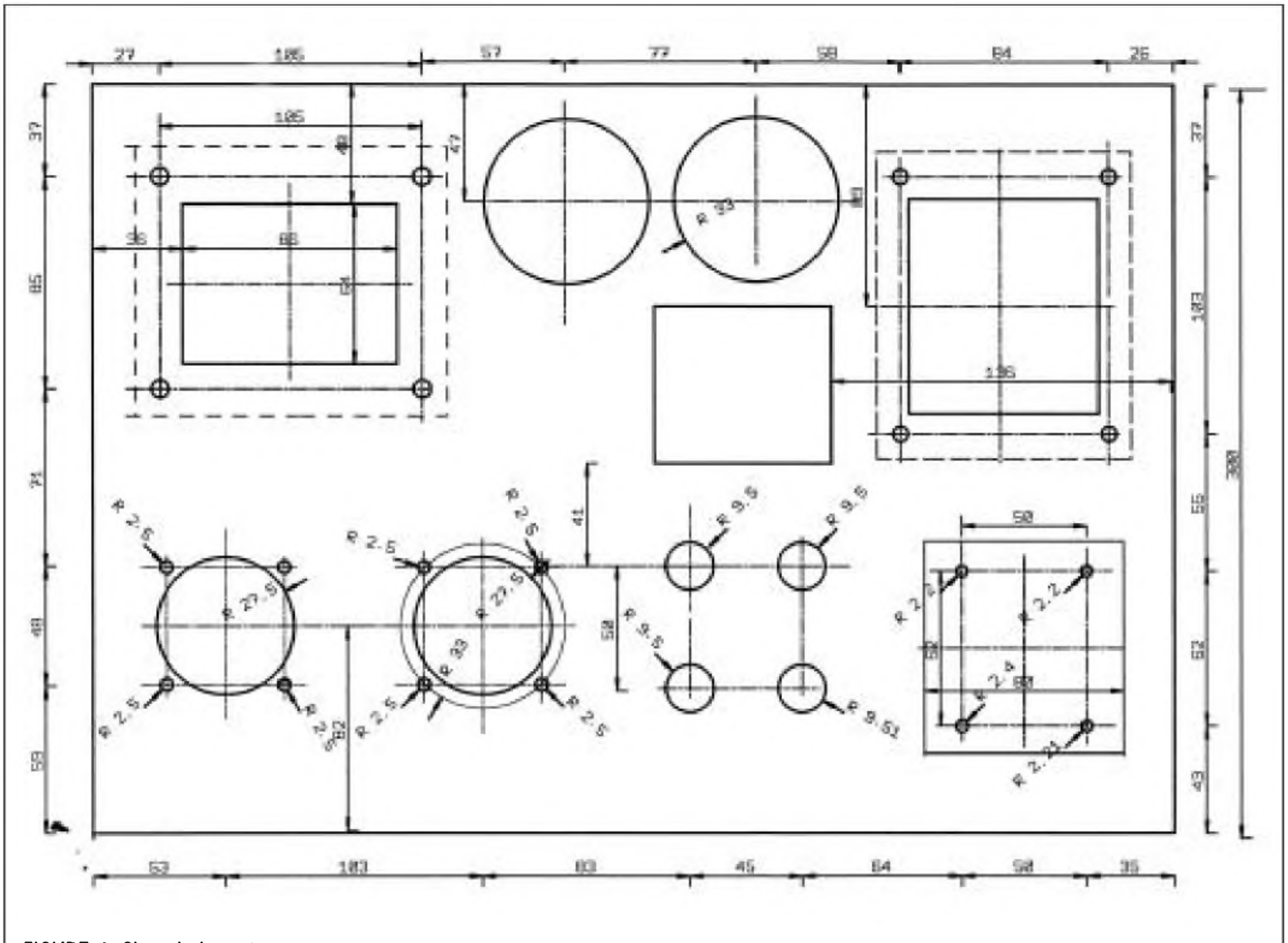

FIGURE 4: Chassis layout.

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transients are less powerful than with a split-load inverter; the bass tones are light and natural; the mediums are very definite; the high tones are more precise. Together this produces a very fine example of listening pleasure. I personally conducted my listening tests with CDs from the *Revue de Son* magazine of various styles of music. I found the results to be excellent. ❖

Translated from French by Gretchen Marie Hromyak.

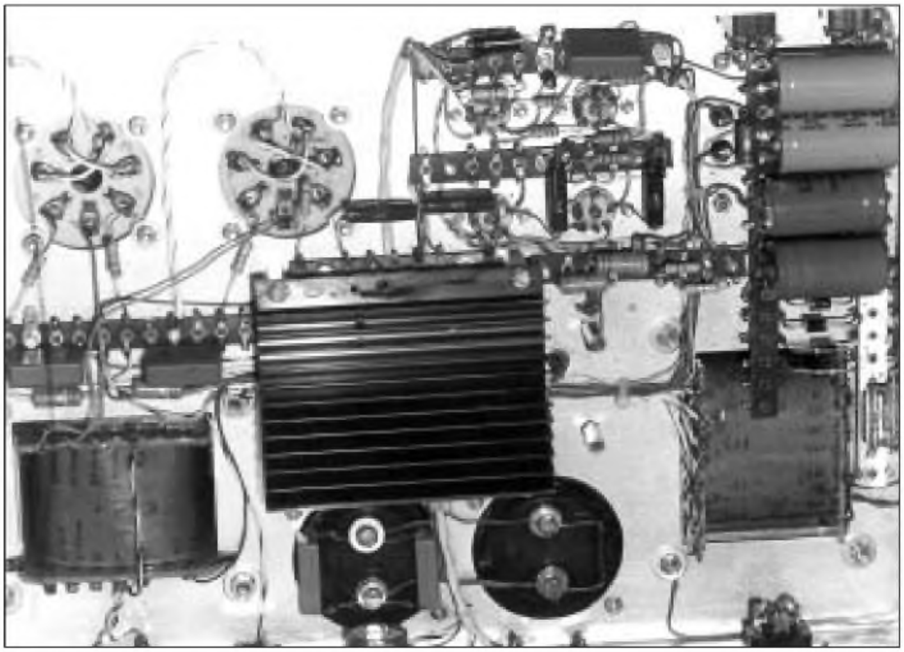

PHOTO 3: Bottom and internal view of amp.

Thagard's Power Supply Revisited

Norm Thagard's two-part series that appeared in our inaugural *aX* generated much reader response and appreciation of his sane, well-designed power supply. Here's a sampling. **By Tom Perazella**

Kudos on your article, "A Phono Pre-Preamplifier for the CD Era," in *audioXpress*, January and February 2001. I especially enjoyed the emphasis you placed on good, rather than expensive, power-supply design. The audio hobby is plagued with the notion that in order for something to be good, it must be prohibitively expensive. I sometimes think that there are people out there who try to cover up their lack of understanding by throwing money at a problem. That usually results in expensive failures rather than just failures.

I'd like to point out a few observations I've made over the years in designing my power supplies. When dealing with silicon rectifiers, or SCRs, Triacs, and other related devices, the surge current of the device is usually specified as the average current for a complete half cycle of conduction when the device is already carrying the maximum-rated current. In a capacitive power supply, as you mentioned, the current flows for only a short period of the cycle. In addition, the average current that each rectifier carries in a supply such as yours is far below its rating. Therefore, the peak current that the device can handle in this case should be in excess of the rated surge current.

CURRENT PROTECTION

Typical I^2T ratings for power dissipation also do not apply to rectifiers until the current is so high that the device transitions from a log response to a linear response. In a typical resistive device, if you double the current for half the time, you don't get the same dissipation, but twice the dissipation. That is because as the current doubles, the voltage drop across the device doubles, leading to four times the power loss but for half the time. Since the drop across the diode junction responds in a log fashion to the current, the voltage drop does not double, and the dissipation is not four times during the shorter time period.

At a company where I formerly worked, some of their products were commercial studio electronic flash equipment. That equipment used rectifiers to isolate the large capacitor storage banks from each other. This was done to protect the switches that selected the number of capacitors to connect to the discharge path from being destroyed if banks were connected to each other when not at the same voltage. When working with large banks of capacitors, it does not take much voltage differential to produce sufficient currents to weld the sturdiest switch contacts together as the switch is actuated.

To protect those contacts, rectifiers rated at 6A were used between the capacitors and the selector switches. The main discharge current of the flash—well into the high hundreds of amps for a few milliseconds—were easily handled by these rectifiers with no problems. This does not mean that rectifiers cannot be destroyed by high currents. Rather, unless you are already at the maximum operating current of the rectifier, the surge ratings are conservative. In all cases, this assumes that the junction temperature is not exceeded in normal conditions by excessively high ambient temperatures or inadequate heatsinking.

Your advice to move the transformer away from the rest of the circuitry is dead on. Using a toroidal transformer can help minimize radiated energy, but for whatever remains, nothing beats distance. As the frequency of the radiat-

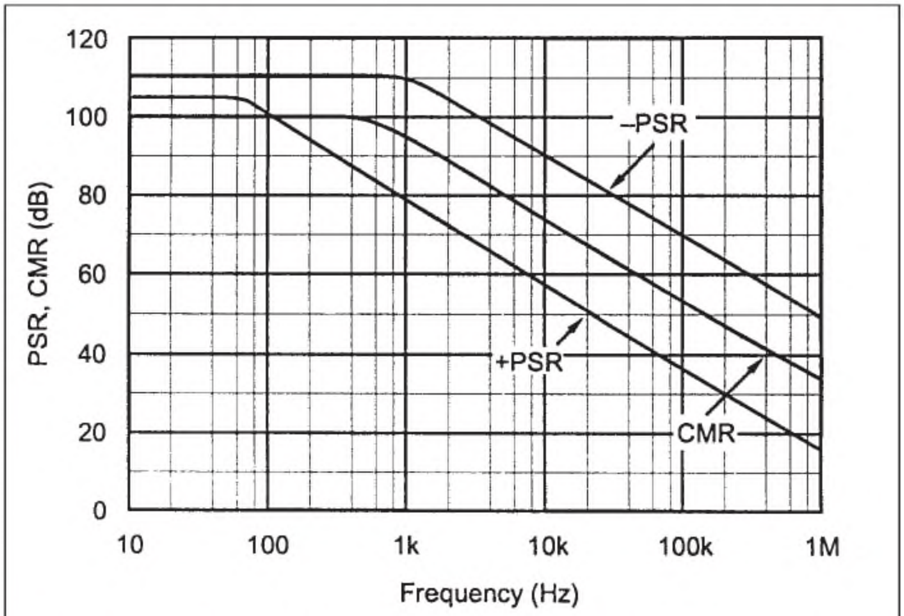

FIGURE 1: Power supply and common-mode rejection versus frequency. A-1962-1

ed magnetic field goes down, it becomes progressively more difficult to use materials to attenuate the field. Look at the VLF communications in use by the military to contact submarines under the ocean.

PSRR

Evaluating the effect of ripple in the power supply is just an extension of the general power-supply rejection ratio (PSRR) of the circuit. PSRR has an important impact on the design of the power supply. I don't know whether you have measured that with your circuit, but I would bet it is similar to what you would find with many of today's high-quality integrated circuit op amps.

One of the critical factors to address when evaluating PSRR is the relationship of ratio to frequency. As the frequency of the power-supply fluctuations—be they ripple or induced noise—goes up, the rejection of that fluctuation by the circuit goes down. If you look at *Fig. 1* showing the PSRR curve of the OPA134 op amp—a typical example of a very good, modern design for audio—you will notice that the rejection ratio is above 100dB at DC for both sides of the supply. The negative side tops out at 110dB and the positive side reaches "only" 105dB. However, the positive side starts losing its ability to reject power-supply fluctuations at 80Hz and the negative at 1kHz.

It is very clear that reducing lower-frequency components of the fluctuation is not as important as reducing the higher-frequency components. The circuit is less able to reject the higher-frequency components, and it is therefore more important to attenuate those before they reach the circuit.

For example, the circuit can reduce the effect of a supply fluctuation by 105dB at 80Hz, but only 60dB at 10kHz. For the same perturbation of the output signal, the power supply must have a 10kHz fluctuation that is 45dB below any at the 80Hz level. The loss of rejection capability appears to be around 6dB per octave.

ELIMINATING NOISE

Your power-supply circuit using a resistor between the two capacitor sections seems to be a very good choice to roll off any high-frequency components that may get through the transformer. Looking at even the first section of the filter, C101 or C102, you would think that those capacitors would eliminate any higher frequency noise that got through the transformer. In fact, that is probably true to a large degree.

However, depending on the capacitors used, the series inductance of the cap could block some of the HF noise from reaching ground, passing it on. To compound that problem, most IC regulators also have a decreasing ability to reject ripple as the frequency increases. Looking at *Fig. 2*—the ripple rejection curve of the 78XX series of regulators—

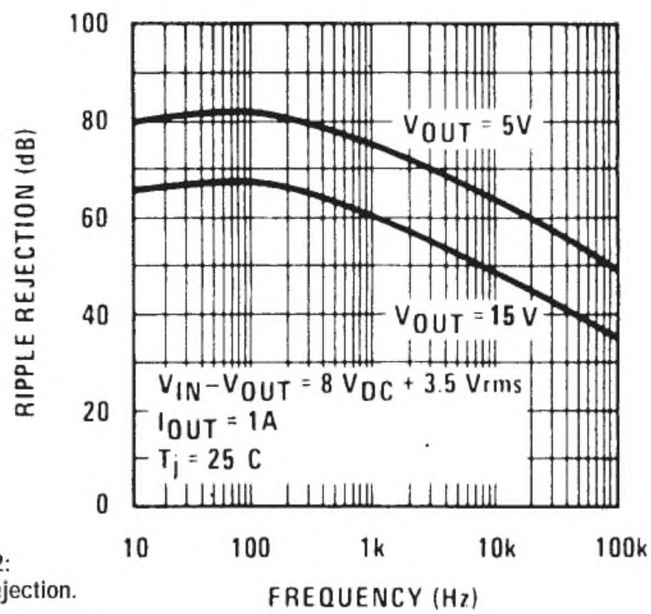

FIGURE 2: Ripple rejection.

you can see that the rejection peaks at about 68dB at around 100Hz, which is ideal for 60Hz ripple. Granted, capacitor-filtered supplies have higher frequency components to the ripple, but it's still a good start.

Your inclusion of the resistor between the two capacitors provides, in addition to ripple suppression at line frequencies, additional rolloff of higher frequency components that would not be effectively rejected by either the regulator or circuit. As far as the ripple is concerned in any capacitor input power supply, raising the secondary voltage and increasing the resistance into the first section will result in a more sinusoidal current flow, all things being equal. This resistance could be in the form of higher transformer secondary resistance or an external resistor. In fact, if you were to carry the concept to an extreme, an extremely high secondary voltage with a sufficiently high series resistor would effectively eliminate the average DC voltage that results on the capacitors from affecting the current waveshape. The charging current would closely approximate what you would see if the load were resistive.

There are obvious practical limitations to this concept, because if the load current were to decrease, the voltage division of the fixed resistor and effective circuit resistance would change, resulting in runaway voltages. In addition, efficiency decreases as the effective impedance of the feed increases in relationship to the circuit, although at the power consumption levels of supplies like this it is not a significant problem.

TWO PROBLEMS

There are two main problems that I have seen in supposedly high-end supply designs that fly in the face of logic. One is the concept that keeping the secondary voltage the same on a supply transformer feeding a low power consumption device, while vastly increasing its current capability, will greatly enhance the resultant sound quality. In fact, if the original transformer had sufficient current capability to keep the regulators from dropping out of regulation regardless of load and line fluctuations, there is no benefit to going to a higher secondary current capability at the same voltage level. Rather, you are better off having a higher secondary voltage at the required maximum current level expected. Then, you have a greater range of input voltages over which the supply can keep the regulators operating properly.

In addition, dissipation limits on the regulators can be a problem if the voltage and current capabilities are kept very stiff as the circuit current demands increase. When a higher secondary voltage transformer that has a higher secondary resistance is used, as the circuit demands increase, the input voltage to the regulator drops, limiting dissipation in the regulator, but shifting it to the transformer. This philosophy is obviously limited to relatively low power demands. The inefficiencies involved in this philosophy would be a problem in a multi-kilowatt amplifier application.

The second problem is the trend to try to speed up the response of the regulation circuit by using high-speed regulators with little or no output capacitance. On the surface this may look good. You might think that the lower capacitance plus fast regulators would reduce the time it took for the regulators to react to a change in output volt-

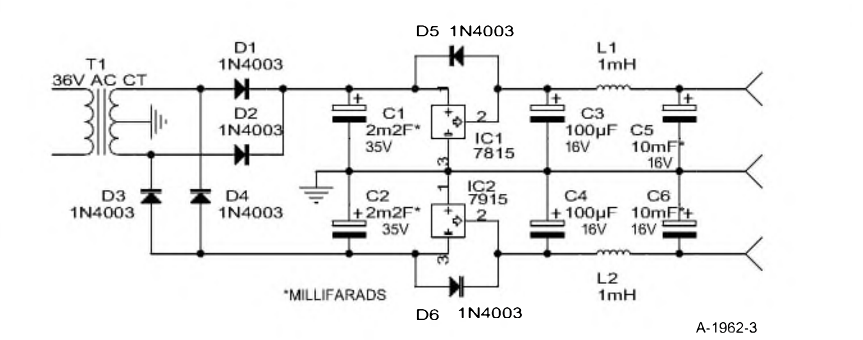

FIGURE 3: Schematic of a typical supply.

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age, resulting in smaller changes. However, speed does not necessarily relate to precision. In addition, even if the faster response resulted in smaller changes at the output of the regulation circuit, it does not necessarily lead to smaller changes in the output of the following circuitry.

Looking again at OPA134, assume the classical supply resulted in a fluctuation of 10mV at 120Hz. Due to the drop in PSRR with increasing frequency, if the higher speed regulators produced a voltage correction with a time constant relating to a frequency of 1kHz, the change in voltage would need to be less than 1mV to produce the same change in the output circuitry. In addition, at 1kHz, the resultant change would be far more objectionable. If the time constant equated to a frequency of 10kHz, to achieve the same level in output circuit fluctuation, the variation allowed by the regulator would need to be less than 0.1mV. With some devices, the PSRR can even go negative at very high frequencies; that is, the device can actually amplify the power-supply fluctuations.

SOLUTIONS

To solve some of these problems in the supplies I have built, I use an approach similar to yours, differing in only a few points. *Figure 3* is a schematic of a typical supply. First, I do not mount the regulators close to the output circuitry. Rather, I feed the rectified and filtered DC directly into the three terminal regulators. Small caps at the output of the regulators maintain stability.

Also note the protection diodes across both regulators, which do not take kindly to reverse voltages. In most cases, the load on the output of the regulators will assure that the output voltages always drop before the input voltages. However, if one of the input rectifiers or caps should fail, the regulator could also be destroyed. Ditto for an external voltage source inadvertently getting into the output circuit.

The other main difference is that I use two inductors and two additional capacitors before feeding the voltages to the rest of the circuitry. The combination of the inductors and capacitors provide a passive 12dB/octave rolloff of any high frequencies that manage to sneak through the rest of the circuit. In addition, these inductors have a DC resistance of only 0.55Ω, causing a very negligible regulated voltage drop.

The very high output capacitance combined with the typical load currents involved lead to a very long time constant, and small changes that are easily compensated for by the regulators, as at those lower frequencies; the chokes are essentially a short circuit. Any fluctuations that may occur are definitely in the flat, high PSRR range of the output circuits. The output changes are very minimal and at frequencies totally below audibility. For the inductors, I use J. W. Miller part number 5258, available from Digi-Key for a unit cost of $3.27.

Of course, to complete the process, I use local bypassing of the op amps directly at the power input leads. Typically I use 3.3μF tantalum caps in parallel with 100nF monolithic ceramic caps. It's also interesting to see how many audiophiles shun ceramic caps, when, in fact, they outperform all the so-called super foil caps in a bypass application. ❖

High g_m Smart Power Tubes, Part 2

Our three-part look at high-transconductance vacuum tubes with low power handling capabilities continues.

By Stefano Perugini

The results summarized in Figs. 4, 6, and 7 of Part 1 reveal that the more interesting Western tubes (with high values in g_m and σ) are also rare and expensive, as any DIYer can easily verify. Fortunately the opening of ex-Soviet markets allows us to discover noteworthy vacuum devices. At the moment, these alternatives are still cheap and easily available and, with their excellent electrical characteristics, represent a good opportunity for the advanced audio experimenter.

This second part looks at three of these Russian high g_m vacuum tubes able to cover most low power hi-fi needs.

THE 6C45-ΠE

This vacuum tube (*Photo 2*) with its predecessor, the 6C15-ΠE, first appeared, in all probability, in the late '60s/early '70s. This valve is a perfect replacement for tubes such as 3A167M/CV5112, WE437A, and EC8020. Since the introduction in my website[3], in 1997, code named by VT-Ω to avoid economic speculations (In Europe, for example, a 3A167M matched pair can approach $250), this tube has gained an excellent reputation in the advanced DIY circles, and for very good reasons, because the application fields are wide and stimulating:

- **RF:** Wideband linear amplifiers from DC to 80MHz
- **High End Audio:** Monotube single-ended low power amplifiers, driver stage for 211/845 SE amplifiers in two-stage configurations, headphone amplifiers, line preamplifiers transformer loaded, small OTLs, A2 class driving, impedance matching
- **Pro Audio:** Vacuum tube amplifiers in low-impedance OTL balanced lines, high-current vacuum tube drivers

Description

The 6C45ΠE is a 9-pin miniature triode intended primarily for use in the high-frequency voltage amplification stage of wideband amplifiers.

Weight: 20g

Pin functions (*Fig. 8*)

1–cathode
2–grid
3–cathode
4–filament
5–filament
6–cathode
7–anode
8–grid
9–cathode

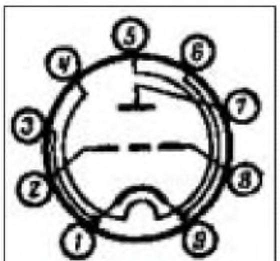


FIGURE 8: The 6C45-ΠE base pinout.

G-1832-8

PHOTO 2: The 6C45-ΠE triode.

CHARACTERISTICS

(Figs. 9 and 10)

Heater voltage	6.3V
Plate voltage	150V
Cathode resistor for self-bias	30Ω

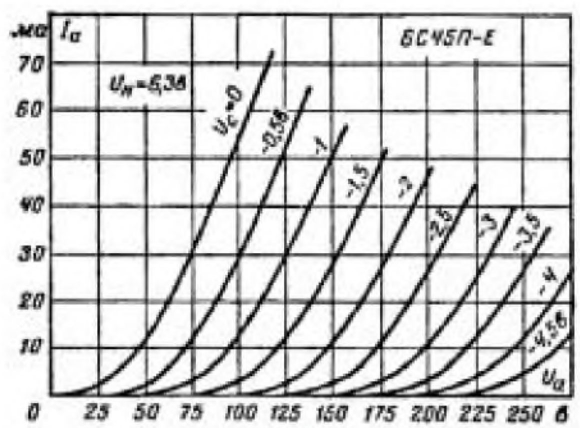
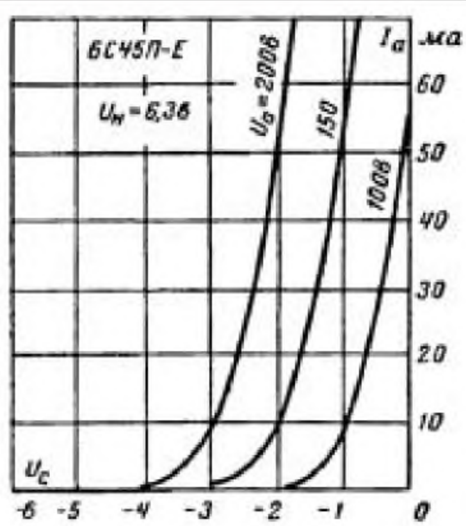


FIGURE 9: The 6C45-ΠE plate characteristics.

G-1832-9

FIGURE 10: The 6C45-ΠE transfer characteristics.

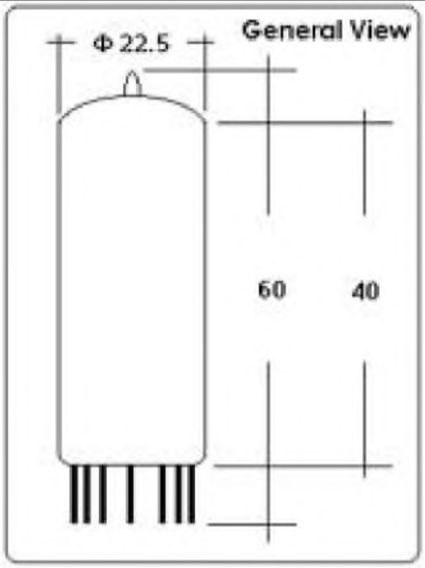


G-1832-10

General Characteristics

	Minimum	Nominal	Maximum
Heater current, mA	410	440	470
Plate current, mA	28	40	52

FIGURE 11: The 6C45-ΠE dimensions.



G-1832-11

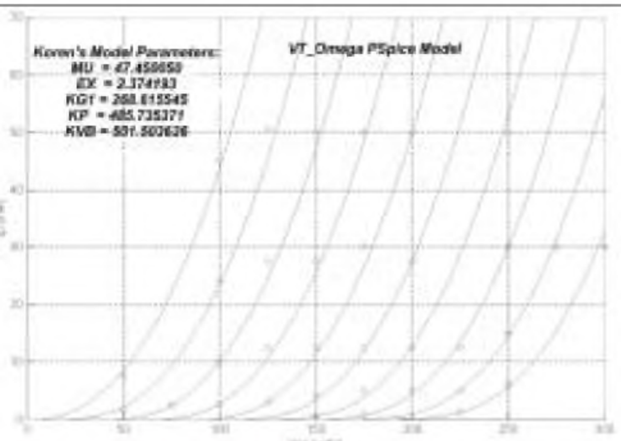


FIGURE 12: The plot of 6C45-ΠE PSpice model.

G-1832-12

Slew rate, mA/V	34	45	56
Amplification factor	36	52	68
Input impedance at F = 60MHz, kΩ	–	3.5	–
Vibro-noise, mV (RMS)*	–	–	100
Lifetime guaranteed, hours	3000	–	–
Direct interelectrode capacitances:			
Input, pF	9.5	11.5	13.5
Output, pF	1.8	1.9	2.2
Heater-cathode, pF	–	6.8	9.5
Heater-grid	–	–	0.13

Center Design Points

Maximum heater voltage	6.6V
Minimum heater voltage	6.0V
Maximum plate voltage	150V
Maximum cathode-heater voltage**	100V
Maximum cathode current	52mA
Maximum plate dissipation	7.8W
Maximum grid resistance	150kΩ
Bulb temperature (common environment)	210C
Bulb temperature (T air = 85C)	230C

*measured with plate resistor = 0.5kΩ

**potential of heater is negative relative to cathode

Mechanical Data

Continuous vibration 500–600Hz with acceleration	Up to 6g
Single beats with acceleration	Up to 500g
Multiple beats with acceleration	Up to 75g
Temperature range	–60C to +85C
Humidity (with T = 40C)	Up to 98%

PSpice Model *(Fig. 12)*

```
.subckt 6C45-PE 1 2 3 ; plate grid cathode
  + params: mu=47.4501 ex=2.374193 kg1=268.615545
  +kp=485.735371
```

```
  +kvb=501.503636
  +rgi=300
  + ccg=2.4p cgp=4p ccp=.7p
e1 7 0 value=
  +{v(1,3)/kp*log(1+exp(kp*(1/mu+v(2,3)/sqrt(kvb+
  +v(1,3)*v(1,3)))))}
re1 7 0 1g
g1 1 3 value= {(pwr(v(7),ex)+pwrs(v(7),ex))/kg1}
rcp 1 3 1g
c1 2 3 {ccg}
c2 1 2 {cgp}
c3 1 3 {ccp}
r1 2 5 {rgi}
d3 5 3 dx
.model dx d(is=1n rs=1
  +cjo=10pf tt=1n)
.ends
```

THE 6H30

This double triode (*Photo 3*), although designed for high-current pulse mode operation, can be seen as an improved version of the E182CC-SQ vacuum tube and therefore perfect as a line preamplifier and driver for power tubes. The official presentation appeared as an advertisement in *Glass Audio* (3/99, 5/99, and 6/99) from Balanced Audio Technologies. Target applications include:

- **High End Audio:** Line preamplifiers (SRPPs, mu-followers balanced stages), headphone amps, Class AB2 low power OTLs, impedance matching
- **Pro Audio:** Low impedance, low gain OTL balanced lines, high current cathode followers

PHOTO 3: The 6H30 double triode.

FIGURE 13: The 6H30 base pinout.

G-1832-13

Description

Double triode for work in pulse modes in various radio devices

Shape–miniature glass bulb
Mass–20g Size–H = 72.5, D = 22.5
Pin functions (*Fig. 13*)
1–plate 1
2–grid 1
3–cathode 1
4–heater
5–heater
6–plate 2
7–grid 2
8–cathode 2
9–shield

General Characteristics

(Figs. 14 and 15)

(with heater voltage = 6.3V; plate voltage = 80V; self-bias resistor = 56Ω)

Heater current	825(+75-100)mA
Plate current for each triode	40(+10–10)mA
Pulse plate current for each triode	2–3A
Plate current with Ug = –12V	<30μA
Reverse grid current	<1μA
Slew rate	18(+5–5)mA/V
Amplification factor for each triode	15(+3–3)
Vibro-noise voltage with Rp = 0.5kΩ	25–50mV
Direct interelectrode capacitances:	
Input(Cgc)	6.3(+0.9–0.9)pF
Output(Ccp)	2.4(+0.5–0.5)pF
Transfer(Cgp)	6.0–7.1pF
Between anodes	<0.2pF
Heater-cathode	8.8(+2.7–1.8)pF
Lifetime	>10000hrs
Term of workability-pulse plate current	>1.7A

Center Design Points

Heater voltage	6.0–6.6V
Maximum plate voltage	250V
Maximum plate voltage for closed tube	1050V
Maximum grid voltage in pulse mode t< = 100μs	–50V
Maximum heater-cathode voltage	400V
Maximum pulse current for each triode	6A
Maximum cathode current for each triode	100mA
Maximum plate dissipation for each triode	4W
Maximum grid dissipation for each triode	0.4W
Maximum grid resistance	300kΩ
Maximum bulb temperature	250°C

Mechanical Data

Continuous vibration 5–2000Hz with acceleration	20g
Multiple beats with acceleration	150g

PSpice Model *(Figs. 16 – 18)*

```
SUBCKT KTRIODE A G K
+ PARAMS: MU=37.58299 EX=1.132166 KG1=428.9745
  KP=402.1368 KVB=321.0256
+  LG=0.2U  VBIG=-0.2  EG=1.0389  KG=0.000769
  KRG=3.685584 KVG=0.090943
```

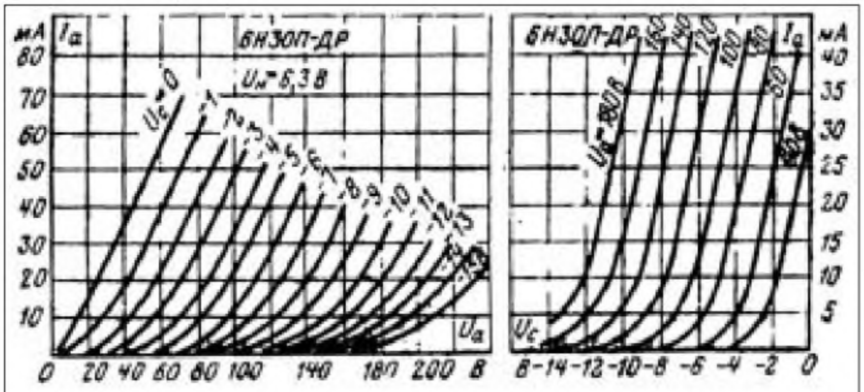


FIGURE 14: The 6H30 plate and mutual characteristics.

G-1832-14

```
+ CCG=2.4P CGP=3.9P CCP=0.7P CCH=5.6P
E1 7 0
  VALUE={V(A,K)/KP*LOG(1+EXP(KP*(1/MU+V(G,K)/SQRT
  KVB+V(A,K)*V(A,K)))))}
RE1 7 0 1G
G1 A K VALUE = {(PWR(V(7), EX) + PWRS(V(7), EX)) /
  KG1}
RCP A K 1G ;
```

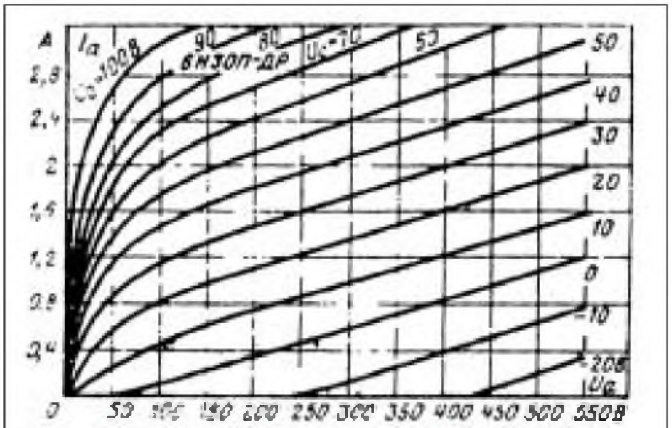
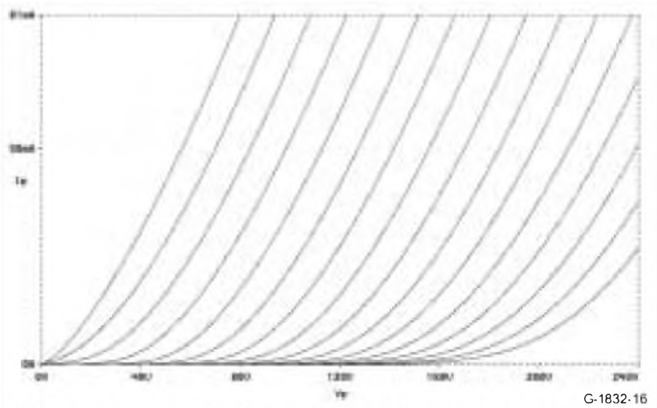


FIGURE 15: The 6H30 plate characteristics for positive grid driving.

G-1832-15



G-1832-16

FIGURE 16: The plot of 6H30 PSpice model (plate characteristics).

```
C1 G K {CCG} ;
C2 G A {CGP} ;
C3 A K {CCP} ;
C4 K 0 {CCH} ;
E10 10 0 VALUE = {IF(V(A) - V(K) > 0, V(A) - V(K),
  0)}
E11 11 0 VALUE = {IF(V(G) - V(K) > VBIG, V(G) -
  V(K) - VBIG, 0)}
E12 12 0 VALUE = {(KG * (V(11) ** EG) *
```

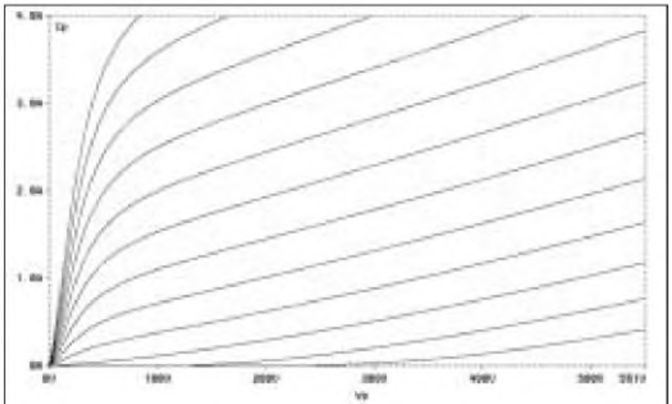
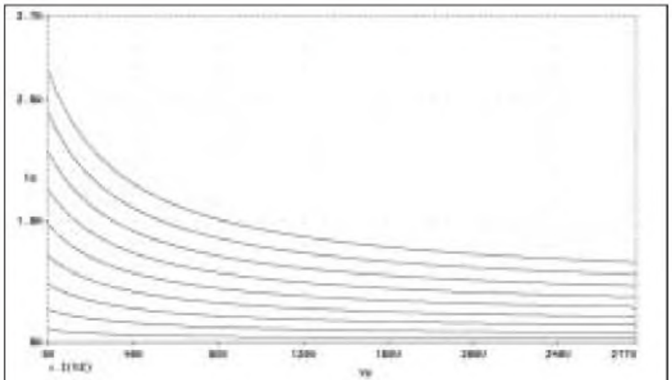


FIGURE 17: The plot of 6H30 PSpice model (plate characteristics for positive grid driving).

G-1832-17



G-1832-18

FIGURE 18: The plot of 6H30 PSpice model (grid currents).

```
+ (((KRG - 1) / (KVG * V(10) + 1)) + 1)) + LG}
G2 G K VALUE = {IF(V(12) > LG, V(12), LG)}
.ENDS
******************
.SUBCKT 6N30P-DR A G K
X1 A G K KTRIODE
+ PARAMS: MU=14.82339 EX=1.386938 KG1=255.6717
 KP=99.30537 KVB=1042.421
+ LG=0.2U VBIG=-0.1 EG=1.414474 KG=0.000769 KRG=5
 KVG=0.027177
+ CCG=6.3P CGP=6.5P CCP=2.4P CCH=8P
.ENDS
```

THE 6C46

This subminiature tube (*Fig. 19*) born as a low-power series regulator is a sort of 6C33 bonsai. The more interesting characteristic is the low voltage/high current mode of operation. This excellent feature permits you to easily build a truly high-performance low-voltage line preamplifier. Further, the interesting plate dissipation (Pd = 4.5W) allows you to experiment with low-voltage low-power amplifiers.

Description

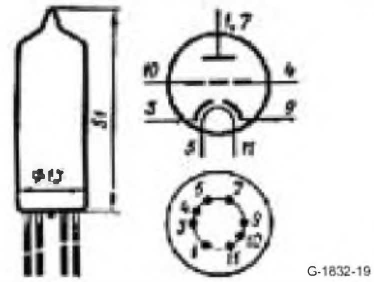
High reliability super-miniature glass triode with one anode and two separate grids and cathodes. Designed to be used in electronic voltage regulators. Lifetime >=500 hrs. *Figure 20* shows the tube's plate characteristics.

Mass	< 7g
Heater voltage	6.3 (+0.7–0.6)V
Heater current	500 (+50–50)mA
Plate voltage	42V
Grid voltage	–1V
Slew rate	20 (+10–5)mA/V
Amplification factor	7 (+2-2)
Plate current	60 (+15–15)mA
Reverse grid current	0.4µA
Maximum plate voltage	250V
Maximum plate voltage for closed tube	330V
Minimum grid voltage	–75V
Maximum cathode current	100mA
Maximum plate dissipation	4.5W
Maximum heater-cathode voltage	150V
Maximum grid resistance	0.25MΩ

PSpice Model *(Fig. 21)*

```
.subckt vt6c46pe 1 2 3 ; plate grid cathode
+ params: mu=8.955 ex=1.5966 kg1=282.77 kp=31.47
 kvb=796.640
+ rgi=2k ccg=6p cgp=7.5p ccp=1.76p
el 7 0 value=
+{v(1,3)/kp*log(1+exp(kp*(1/mu+v(2,3)/sqrt(kvb+v(1
 ,3)*v(1,3))))))}
rel 7 0 1g
g1 1 3 value= {(pwr(v(7),ex)+pwrs(v(7),ex))/kg1}
rcp 1 3 1g
c1 2 3 {ccg}
c2 1 2 {cgp}
c3 1 3 {ccp}
r1 2 5 {rgi}
d3 5 3 dx
.model dx d(is=1n rs=1 cjo=10pf tt=1n)
.ends
```


FIGURE 19: Sketch of 6C46 and base pinout.



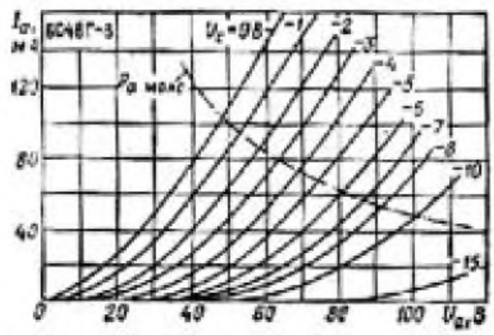


FIGURE 20: The 6C46 plate characteristics. G-1832-20

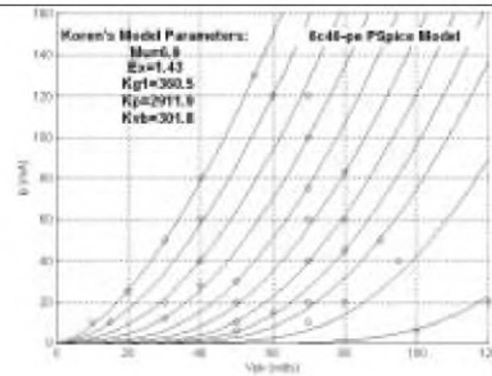


FIGURE 21: The plot of 6C46 PSpice model. G-1832-21

CONCLUSION

The triodes 6C45, 6C46, and 6H30 are capable of covering all your low-power audio needs. In the third part of this article I'll present a few practical examples. For pricing and availability information, you can refer to the Russian distributor STC Navigator[4]. ❖

ACKNOWLEDGMENT
I thank Pavel Skidan, who helped me in the search for technical documentation and in translating from Russian to English.

REFERENCES
3. http://www.paeng-design.com
4. http://www.tubes.ru

Product Review

Sony SACD Player Modification

Reviewed by Gary Galo

SACDmods.com Sony SCD-CE775 SACD Player Modification. SACDmods.com, 223 North 1st St., Cambridge, OH 43725, www.sacdmods.com. Prices (customer must supply the player for modification): $325 for two-channel mod with new RCA connectors; $355 for multi-channel mod with new RCA connectors.

SACDmods.com is a company specializing in modifications to Sony Super Audio CD players. The player I review here is the very popular SCD-CE775 (*Photo 1*), a recently-discontinued player that often sold on the street for less than $200, a far cry from the five grand price tag of Sony's original SACD player, and their current flagship, the SCD-1. SACDmods also offers modifications to the Sony DVP-NS500V and SCD-C222ES players.

The SCD-CE775 is a five-disc changer offering the full 5.1 channel surround playback capability for multi-channel SACDs. A pair of conventional left and right outputs are also included for two-channel stereo installations. The player comes with a full-featured remote. *Photo 2* shows the rear of the unit.

BURR-BROWN DAC

Sony—a company known for products filled with their own proprietary chips—

PHOTO 1: Front view of the SCD-CE775 Super Audio CD Player. The stock and modified units are identical in appearance.

PHOTO 2: Rear view of the SCD-CE775. The stock PC-mount RCA jacks for the two-channel outputs are replaced with Cardas GRFA connectors by SACDmods.

actually uses Burr-Brown (now a division of Texas Instruments) DSD1702 D/A converters. The DSD1702 is a multi-format, delta-sigma converter supporting both PCM and DSD formats. In the PCM mode, the chip will accept sampling rates up to 200kHz, with datastreams up to 24-bit, though it does not support the DVD Audio standard.

Typical of D/A converters used in lower-cost CD players, the DSD1702 is a combination DAC and digital filter, and actually contains two digital filters. The PCM filter is an 8×-oversampling type, and the DSD filter offers three –3dB passband choices—50kHz, 60kHz, or 70kHz. The SCD-CE775 includes three DSD1702 chips, for the six output channels. Like most combination DAC/filter chips, the DSD1702 has a voltage output, with I/V conversion done internally.

Each channel of analog circuitry consists of half of an NJM2114 dual op-amp manufactured by JRC, and configured as a three-pole analog filter. The left and right stereo outputs are fed from the same op-amp as the front-left and front-right outputs for surround installations. A multi-channel/two-channel mode button ensures that these outputs function correctly in your particular installation. The NJM2114 offers several performance improvements over the 5532, including a 15V/µs slew rate and a 15MHz unity-gain bandwidth.

Electrolytic coupling capacitors are used on both the input and the output of each of the six channels. It is strange that Sony didn't opt for a low-offset and low input bias current FET-input chip for the analog stages. This would have allowed them to eliminate the electrolytic output coupling capacitor. Output coupling caps seem to be a bad habit with Sony and other Far East manufacturers. Even Sony's $1700 SCD-C555ES and their top-of-the-line SCD-1 have capacitor-coupled outputs (these players do have separate D/A and digital filter chips, plus op-amps dedicated to I/V conversion).

The SCD-CE775 is not based on a DVD player. Consequently, it has a conventional power supply (*Photo 3*), rather than one of the noisy, sonically-menacing switching supplies found in all DVD players. This gives the player a head start, as far as modifications are concerned, because the modifier need not be concerned about trying to rid the player of the broadband, high-frequency noise inherent in switching supplies.

I know I will get flack for this statement, but I don't care: switching power supplies have no business in any piece of high-end audio equipment. There's only one reason to use them: they're small and cheap (OK, I guess that's two reasons). How many manufacturers of chips for digital audio have warned that low-noise supplies are important for best performance? Just read the data sheets, folks.

MODIFICATIONS

The modifications to the SCD-CE775 (*Photo 4*) consist of the following:

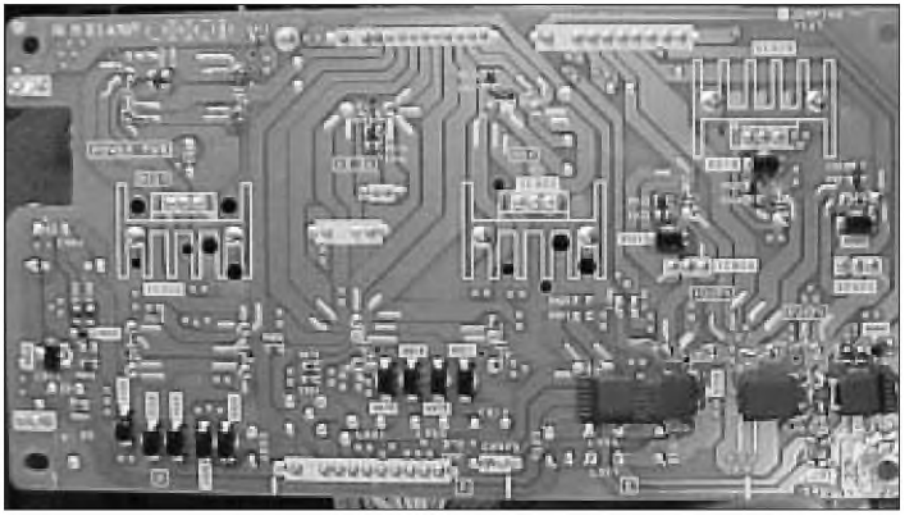


PHOTO 3: The modified power supply board. The Harris high-speed, soft-recovery rectifier diodes (FREDs) used in the power supply mod can be seen in the lower right, insulated with heat-shrink tubing (courtesy of SACDmods.com).

- Harris high speed, soft recovery diodes (FREDs). These are installed on the supply board that powers the audio and clock circuits. This is a tricky procedure, because four surface-mount chip diodes are replaced with TO-220 cases, and the fit is very tight.
- Black Gate Capacitors. These are 47µF non-polar electrolytics, NX Hi-Q series, which replace the stock capacitors coupling the inputs and outputs of the analog op-amps.
- LC Audio LClock XO2. This clock board (*Photos 5* and *6*) is made in Denmark, and replaces the presumably jittery clock supplied by Sony. Further details on this clock are available on the LC Audio website, http://www.lcaudio.dk/com/intlcl.htm. As you will see, the clock is quite expensive, and accounts for most of the modification expense.
- Cardas GRFA RCA jacks. These replace the cheap PC-mount RCA jacks supplied with the stock player. The jacks are connected to the PC board with Vampire "Continuous-Cast" copper wire with Teflon® insulation.

SACDmods proprietor Matthew Anker sums up his modification philosophy this way: "Unlike a lot of the other companies that modify SACD players, I believe that simplicity is the best way to go. I will not bypass the Black Gates with film capacitors or change resistors to metal foil. These methods get you a harsh, metallic sound. The high frequencies would be emphasized to the point that listening is no longer a pleasure; therefore, only a carefully selected blend of components is used in my mod. The carbon resistors in the player already are a very neutral sounding component, and match well with the Black Gate capacitors."

TEST TRACKS

SACDmods lent me a stock SCD-CE775,

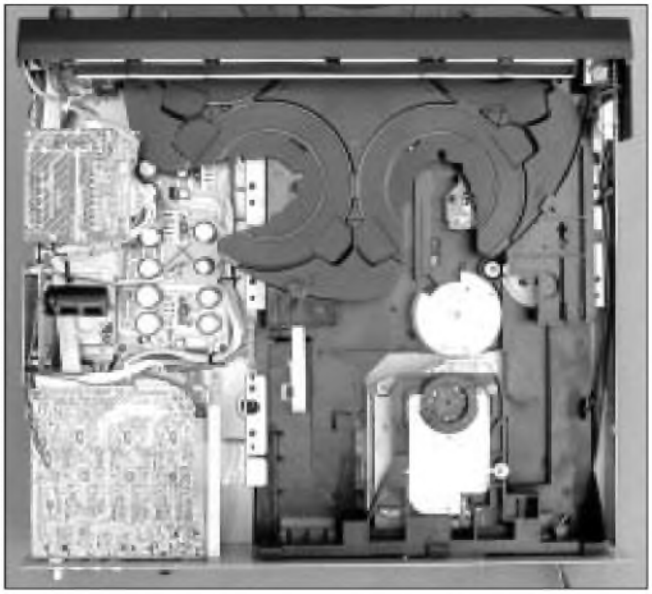


PHOTO 4: Inside the modified SCD-CE775. The analog board is in the lower left, and the new LClock XO2 board is just above the analog board. The power supply board is to the right of the clock board.

TELARC'S CLASSIC SOUNDSTREAM RECORDINGS RE-MASTERED

PHOTO A: On the left, a Direct Stream Digital remastering of an audiophile classic, Robert Shaw's 1978 Soundstream recording of Stravinsky's Firebird Suite. On the right, the remaking of a Telarc chestnut, Tchaikovsky's 1812 Overture with Erich Kunzel and the Cincinnati Pops Orchestra in an all-new DSD/SACD recording.

In 1978 Telarc, then a fledgling audiophile record label, made audio history with the issue of their first digitally-mastered, long-playing record, Frederick Fennell conducting the Cleveland Symphonic Winds in Gustav Holst's two *Suites for Military Band*, plus music by Bach and Handel. This album was quickly followed by Telarc's first digital orchestral recording, Robert Shaw and the Atlanta Symphony Orchestra performing Igor Stravinsky's 1919 *Firebird* suite, plus the Overture and "Polovetsian Dances" from Borodin's opera *Prince Igor* (*Photo A*). At the time these recordings were made, the classical recording industry was at an all-time low in terms of engineering, with most of the major labels capturing frustratingly unnatural sound due to over-use of multi-microphone techniques.

Telarc took several steps back, with Producer Robert Woods and Engineer Jack Renner opting for a purist approach. Borrowing the technique adopted by Robert Fine in his fabulous stereophonic recordings for Mercury in the late 1950s and early 1960s, Woods and Renner used only three spaced omnidirectional microphones (Schoepps Colette-Series SKM-52U). The results spoke for themselves—these recordings captured a symphony orchestra with remarkable realism, with natural imaging and depth, clarity and dynamics. Then, of course, there was that famous Telarc bass drum, taxing phono cartridges, power amps, and loudspeakers. (The drum head was facing the mikes. –Ed.)

Also noteworthy was the digital recording system used by Telarc, the Soundstream recorder designed by Dr. Thomas Stockham. Audiophiles were clearly divided on the merits of digital audio, and for years to come many would proclaim the virtues of analog sound. Indeed, many still do!

One thing was certain, however. Stockham's Soundstream recorder was well ahead of other digital recording systems in use at that time. To those of us who found Denon's digitally-mastered LPs of that period excruciatingly bad, the Soundstream recorder was a revelation, making a strong case for the potential of digital audio.

The Soundstream recorder was surprisingly advanced by late 1970s' standards. The sampling frequency was 50kHz, higher than what would be adopted a few years later for the Compact Disc. In addition, the system offered 16-bit performance at a time when anything over 14-bits was thought to be impractical.

The Soundstream system also used a Rockwell instrumentation recorder—modified, if my memory is correct—and arguably a better solution to the problem of digital audio storage than NTSC-encoding on videotape, which would become extremely popular in the early years of digital audio. Telarc's first digital recordings were made using a Studer Model 169 mixing console. By the time they recorded Stravinsky's *Le Sacre du Printimps* in 1980—with Lorin Maazel and the Cleveland Orchestra, another audiophile spectacular—they had switched to a Neotek console.

DSD CONVERSION

With the introduction of the Compact Disc, it was only natural that Telarc should transfer some of these pioneering recordings to the new consumer format. There were problems to be surmounted, however. Converting the 50kHz sampling rate of the Soundstream system to the 44.1kHz CD standard posed a nasty mathematical problem. You must remember that these transfers were done back in the 1980s, before the introduction of asynchronous sampling rate converters such as the Analog Devices AD1890-series and its successors. Surely, these CDs offered only an approximation of the original Soundstream masters.

Enter Sony's Direct Stream Digital recording, the mastering system for the Super Audio CD format. Within the past two years, Telarc has re-mastered some of their classic Soundstream recordings for issue in SACD format, using the best digital conversion technology. The conversion from the original Soundstream recordings to DSD has been done using a dCS 972 Sample Rate Converter with custom software. Fortunately, Telarc has had the presence of mind to offer all of their SACDs in dual-layer format, so they will also play on conventional CD players. For the mastering of the CD layer, the DSD transfer is converted to CD format using Sony's Super Bit Mapping Direct system.

I compared the CD layers of the Shaw *Firebird* and Maazel *Le Sacre* to Telarc's previous CD releases and was amazed at the improvement in sound. The most noticeable difference was a dramatic improvement in space. The old CD transfers sounded dry and "up-front," lacking in hall ambience and a sense of the original recording venues, and somewhat hard in the treble region.

The new transfers sound like a different microphone pickup altogether (though they obviously are not). There is a tremendous improvement in hall sound, with correspondingly greater depth perspective and accuracy of placement within the soundstage. Overall, the recording venues sound considerably larger on the new transfers. A layer of edge and glare on the old CDs is largely removed by the new conversion process.

The SACD layers are better still. As Telarc points out in the booklets to the new issues, the SACD transfers offer listeners the ability to hear—for the first time—the sound that the recording team heard on the original Soundstream masters. In fact, these SACD transfers may sound even better than that, due to considerable advances in the digital-to-analog conversion process.

What is most impressive is how well these Soundstream recordings have held up over the past 24 years. They don't have the detail and resolution, or the warmth and absence of high-frequency artifacts of today's digital recording systems—especially SACD. Nor do they capture the harmonic richness of a symphony orchestra, or the sound of the original acoustic space, as well as current digital technology. But they still offer a satisfying—indeed impressive—audio experience, which I cannot say about any other digital recordings made during that period. The Telarc DSD re-masterings are a testament to the remarkable achievement of Thomas Stockham and the Telarc recording team.

so I could compare it to the modified player. For this review, I purchased Telarc's SACD remasterings of two of their classic Soundstream digital recordings—Stravinsky's *Firebird* suite conducted by Robert Shaw (SACD-60039) and *Le Sacre du Printimps* conducted by Lorin Maazel (SACD-60563). I offer a few thoughts on these early Telarc digital recordings in the sidebar accompanying this review.

SACDmods also lent me Telarc's SACD of Tchaikovsky's *1812 Overture* (*Photo A*), an all-new Direct Stream Digital remake of one of Telarc's most famous audiophile recordings (SACD-60541). I also tried Sony Classical's SACD re-masterings of Leonard Bernstein conducting his Overture to *Candide*, the symphonic dances from *West Side Story*, the symphonic suite from *On The Waterfront*, and the *Fancy Free* ballet. All Telarc issues are dual-layer, so they will play on a conventional CD player, as well. Sony Classical's SACD discs are single-layer, and will only play on SACD players.

LISTENING CRITIQUE

The stock Sony CE775 did not make a very strong case for the SACD format, and exhibited many of the characteristics of an inexpensive CD player. The sound is rather dry, slightly edgy in the treble, and lacking in a precise, three-dimensional soundstage. If you compare the CD layers to the SACD layers on the Telarc recordings, the SACD layers are superior, but clearly well below the ultimate capabilities of the format. I actually preferred the CD layers of the Telarc recordings played on my reference CD playback system to the SACD layers as reproduced by the stock Sony player.

The modifications performed by SACDmods changed the balance entirely. On the Telarc *1812* the modified player has almost none of the high-frequency spit and trash of the stock player. There is more detail and less congestion in passages for full orchestra. The sonic presentation is more open and airy, especially striking in the reproduction of the chorus near the beginning of the work (the addition of the chorus is a re-touching by the conductor). The low strings are much warmer on the modified player, and the high strings are silkier and more life-like. There is much greater retrieval of hall ambience, particularly around the players near the back of the orchestra.

On the 1980 Telarc *Le Sacre du Printimps* the woodwinds have much greater harmonic richness on the modified player, and the entire sonic presentation is far more spacious and natural. The 1978 *Firebird* has more effortless full-orchestral transients, and the famous Telarc bass-drum sound is deeper and fuller. On all three recordings the message is the same: the stock Sony CE775 doesn't reveal enough of SACD's potential to justify the new format. The SACDmods remake most certainly does!

Sony Classical's *Bernstein Conducts Bernstein* SACD was not very impressive sonically. It is not categorically

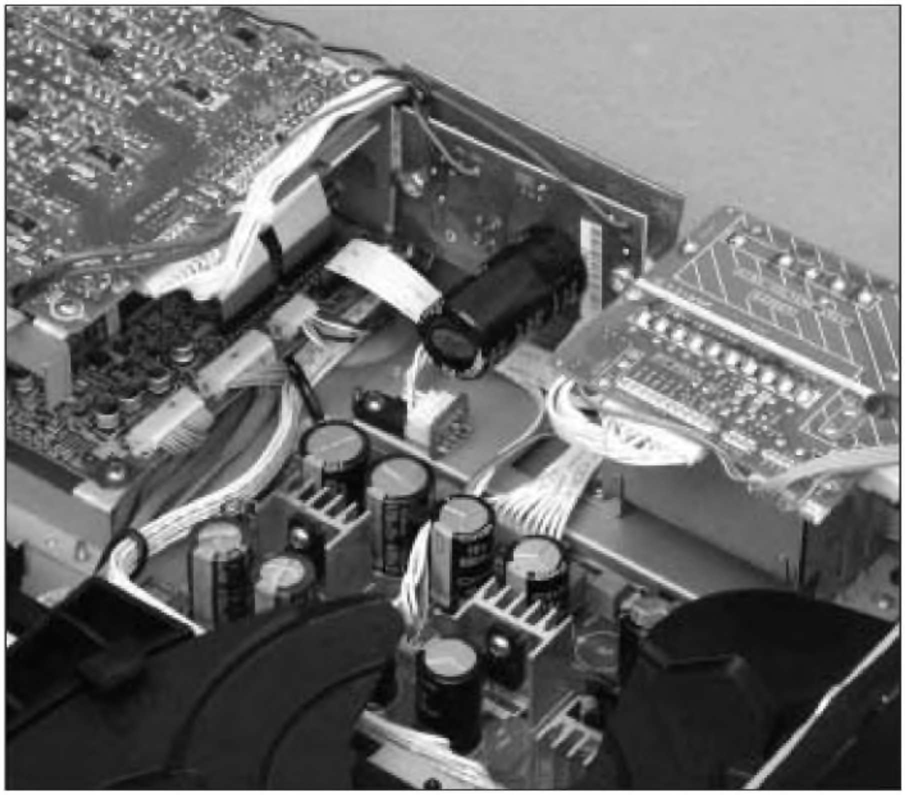


PHOTO 5: Close-up of the installed LC Audio LClock XO2 board. Most of the circuitry is on the bottom side of the board, hidden by the aluminum mounting plate.

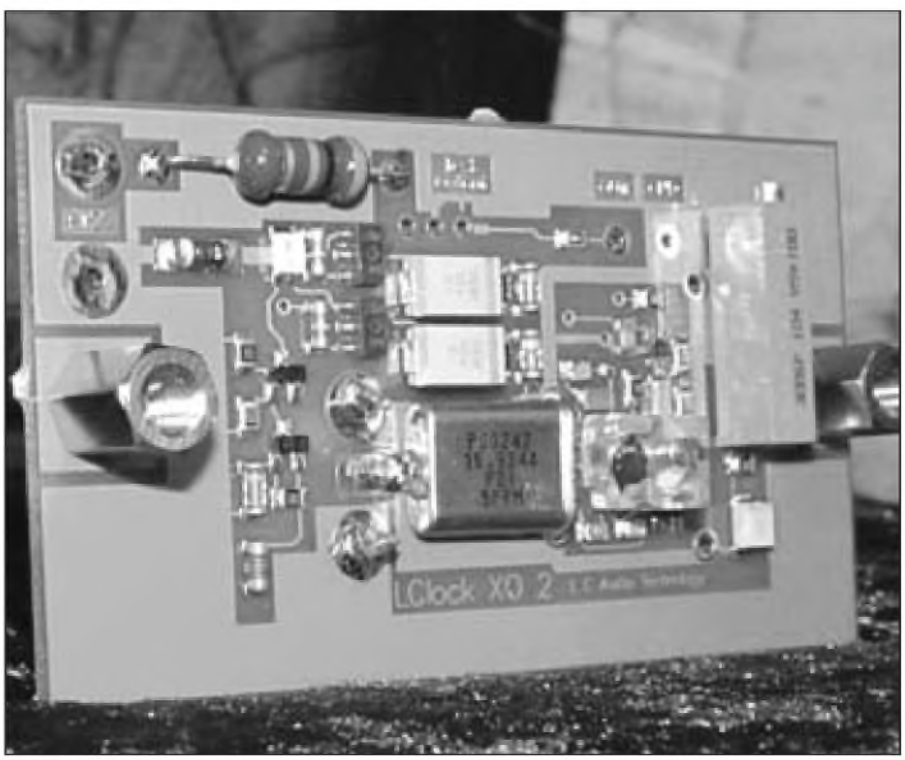


PHOTO 6: Component view of the LClock XO2 board (courtesy of SACDmods.com).

better than its 20-bit CD counterpart (SMK 63085, part of the "Bernstein Century" series, re-mastered with Sony Super Bit Mapping). In fact, the SACD sounds as though it was made from an analog production copy, rather than the original session tapes. The CD version was produced by John McClure, one of the original session producers, and someone who obviously cared a great deal about doing justice to the original tapes. I did not find this Sony SACD particularly useful as reference material.

BATTLE OF THE FORMATS

There is considerable debate regarding which format is superior: SACD or 192kHz/24-bit DVD audio. I don't think that matter can be put to rest until identical recordings can be compared in both formats. I hope some of the best recordings from the analog era—classics from the Decca/London, RCA Victor Living Stereo, and Mercury catalogs—will be re-mastered in both formats so that meaningful comparisons can be made.

Another contender is the Vox/Turnabout recording of Rachmaninoff's *Symphonic Dances* (already available in a superb 96kHz/24-bit DVD from Classic Records, and one of my most valuable reference recordings). I would also expect outboard digital processors to become available for these formats, once a standard digital interface becomes available. We should hope, even though the datastreams are incompatible, that both formats can be accommodated by one transport and outboard D/A converter, with a common digital interface, with automatic detection of the source format.

In the meantime, the SACDmods.com re-make of the Sony SCD-CE775 SACD player is a very impressive performer, one that makes a convincing case for the new format, at an affordable price. If you own this or one of the other low-priced Sony SACD players, you should strongly consider having SACDmods rework your player. ❖

Manufacturer's response:

Since Mr. Galo reviewed the early version modified SCD-CE775 I submitted, I have added several tweaks to my modification packages:

- *Class A biasing of the front channel op-amp. This smoothes out the high frequencies by simply adding 2mA of bias current.*
- *Installation of two 1000µF Black Gate filtering capacitors on the ±7V supply to the op-amps and DAC. This lowers the noise floor considerably.*

Recently added as an option:

Installation of a 1 lb, 25VA toroidal power transformer for the audio board's power supply. This makes the CE775's power supply nearly identical to that of the C222ES. The transformer gives the bass more "punch" and completely separates the digital supply from the analog rendering a cleaner sound. This is a $39.95 option that can be installed with a CE775 modification package.

Matthew Anker
www.SACDmods.com

LITTLE AMP

I wish to thank Mr. David Wolze for his kind words ("Express Mail," April '02 *aX*, p. 65) regarding my article "A Great First Amplifier Project" (Dec. '01). I welcome his suggestions for upgrading with a different output transformer. I tried other trannies besides the #1608 Hammonds, but found them to be the best value for the price and also to be up to the task of supplying the proper load required by this "little amp." (Actually, the best upgrade for this project would be to replace the coupling caps with the Hovland MusiCaps available from Antique Electronic Supply.)

As a matter of fact, I even installed a pair of Hammond #1620s and tried the circuit with some 5881 tubes I had in stock. They sounded pretty good but will sound better after I change the B+ voltage, heater power supply, and the bias circuit. The "snubber" circuit Mr. Wolze mentioned may work in some cases, but I found that it often duplicates the circuits in some crossovers. And, yes, I agree that a circuit with a single 6Y6 with a UBT-3 output would probably work fine—sounding OK and certainly keeping the parts count down—but it could drive the cost of the project up, which was what I was trying to prevent for the benefit of first-time builders who might not want to invest too much money in a project when they're not really sure of the outcome.

If someone really wants to build a single-ended amp with a low parts count, he could try using a 6L6, 6BQ5, 5881, or other similar type of tube in my single-ended project (Sept. '01, p. 20). You could do this with about 12 parts per channel and still have a sweet tube sound, by using a different power supply, adjusting the bias circuit, and the by-pass caps could be optional. As in all cases, the most expensive part in a single-ended amp is usually the output transformer.

The main purpose of the 12V6 push-pull amp was really to provide a good tube sound with a minimum of parts and an easy-to-construct project which is not very expensive for the first-time builder. Some other hobbyists who have built this amplifier say they really enjoy the sound it produces for such a small investment of their money and time. I truly hope others will enjoy it just as much!

Rick Spencer
Clovis, Calif.

CHOOSING TUBE CIRCUITS

I have enjoyed several of Rick Spencer's articles in the past two years. However, I would appreciate it if he could spend a little more space on the pros and cons of his circuit choices.

For his most recent article ("A Five-Channel Tube Home Theater Amp," June '02, p. 42), he used one separate cathode bias resistor for each 12V6 tube instead of a single one for both tubes. A shared cathode bias resistor is supposed to even out differences between tubes and provide better tube matching. He used a shared cathode bias resistor in a similar design in "A Great First Amplifier Project" (Dec. '01). Why the change? He also never mentioned whether tube matching is necessary for this type of circuit. Is it necessary?

I am new to tube amplifiers. Can he explain the sonic difference between a cathodyne phase splitter (Dynaco 70) versus the one used in his circuit? Both can achieve similar gain using two half tubes.

Bing Yang
bingyang99@hotmail.com

Rick Spencer responds:

I am glad you have enjoyed my articles in audioXpress. The pros and cons you mention in your letter can be narrowed down to the fact that all of the circuits used in my designs are very simple and, yet, are capable of delivering a good sound when used properly. My single-ended 6550 amp used a super simple, low parts count circuit, and when teamed with the proper speakers and signal sources will almost bring tears to your eyes with its pure sound. The "little amp" (Dec. '01) also uses a straightforward simple circuit, has a low parts count, and will deliver a very good sound at a minimal cost.

I used this same circuit in the five-channel home theater amp project with only a slight change to the cathode bias arrangement. The engineers at RCA used this type of cathode resistor-capacitor setup in their design of a 6V6 amp back in 1959. I was curious to see whether I could obtain a smoother sound, better stability, and maximum power from each tube, so I used this type of arrangement and found what seems to be an overall improvement in sound quality. However, if you want to reduce your total parts count and keep things even more simple, then use the same type of circuit found in the 12V6 amp in Dec. '01.

As far as tube matching goes, you may request matching when you order your tubes from Antique Electronic Supply. Matched tubes will usually perform better, and I try to match mine on my tube testers. I also try to match resistor and capacitor values in critical parts of the circuit.

Regarding the Dynaco phase inverter circuits, when you consider the famous engineers that developed them you can only assume that they produced the results they were trying to achieve. I've repaired and upgraded (POOGEd) a couple of Dynaco amps, and I have always enjoyed their laid-back sound.

Rather than use a lot of space explaining the different types of inverters here, I suggest you try to obtain a copy of Glass Audio 1/90 from audioXpress and read the article by Daniel Norman concerning the inner workings of phase inverters ("Understanding Tube Phase Inverters," p. 12). I have never seen it explained better.

The most interesting inverter circuit I have seen is the Fisher 500-C. It featured a control labeled "phase inverter adjustment"(!). If you're curious about it you can find the diagram in issue #6, 1997 of Vacuum Tube Valley. And, since the design was used in a FISHER—well, what more can one say!

The phase inverter circuit I used in my 12V6 amps is really simple but will have slightly more distortion and phase shift than the ones submitted by Joseph Still in his Glass Audio 5/00 article ("A 20W $260 Amplifier"). His style of inverter has reduced phase shift and only a small degree of imbalance remains between its two outputs, which makes it very desirable for obtaining a smoother overall sound. If you have the room on your chassis then use his style of voltage amplifier and inverter. You can't go wrong with any of his designs. The choice is yours.

MC STEP-UP TRANSFORMER

Peter Millett is to be commended on his excellent and very practical article on building a moving-coil (MC) input transformer unit (May '02, p. 58). There is no doubt that a properly designed passive unit is preferable to a specially designed active MC input stage, since it requires some unusual parameters of the input devices used to keep the noise level within acceptable limits. The main one is the rbe, which is the principal noise generator and must be as low as possible. Selected low-noise small-signal devices are not the prerequisites in this case, and, paradoxically, power transistors are more likely to be suitable! However, I'd like to expand on what he had to say with respect to the selection of the transformer (and I endorse his choice of the ones from Sowter, which are really microphone types).

My co-worker, the late Peter Baxandall, and I were involved with the design of a similar device for a British manufacturer of MC cartridges—the Mayware—sadly, no longer available, particularly because they carried a budget price and excellent value for money. The design brief was for a transformer interface that would accommodate a range of source Z cartridges, from 3Ω (typical) to as high as 50Ω. After ruling out the practicability of tapped transformers, we hit on the idea of a standard 1:40 designed specifically for a 3Ω cartridge. This lifted the nominal 125μV output to 5mV and so, ideal for most moving magnet (MM) input stages.

After examining a range of representative MC cartridges, we then confirmed that all had one outstanding useful characteristic in common—that the source Z was substantially resistive at all frequencies. This may well contribute to the excellent and consistent performance, since MM types are a complex mix of reactance and resistance and normally require careful loading, standardized as we know at 50kΩ. Even so, you must still consider shunt capacitance of the input cable and its effect.

None of these factors are a problem with MC cartridges. Loading was also immaterial, as long as the reflected load was at least 3× the source Z; a little arithmetic shows that the 50kΩ reflects to a trifle over 30Ω. So no problem there. Now, what about a higher source Z cartridge, say 50Ω? Two factors to consider here: the signal is going to be much higher at the secondary, far higher than we need. The technique we adopted has, until now, remained a secret.

Certainly, it baffled at least one reviewer of the product as to how it seemed that the transformer had an adjustable step-up ratio. Some of the obscure theories were highly amusing, since as is so often the case, he lacked the engineering experience and training to work out what was a straightforward solution to the problem. We simply shunted the transformer secondary internally with a lower resistor, that by virtue of its attenuating action in series with the reflected source Z, brought the signal level down to what the input stage required. It also had the desirable effect of swamping the reactive components of the transformer, now more dominant because of the higher source Z of the 50Ω cartridge.

In our case, an 18kΩ was the choice, and the reflected load now for a 3Ω cartridge becomes a trifle over 13Ω. I still have the protoype and a production model.

Reg Williamson
Kidsgrove Staffs, England

Peter Millett responds:

Thanks to Mr. Williamson for his compliments, and for an excellent way to simply make a "self-adjusting" MC step-up transformer.

If I understand correctly, the scheme used works basically by providing a relatively low impedance load for a higher impedance cartridge—in other words, the "3:1" rule for load:source impedance is intentionally violated. This works as an attenuator, because the source Z is nearly purely resistive—if it were not, the frequency response would be adversely affected. The result, then, is that output from a higher source impedance cartridge is attenuated more than that from a low source impedance cartridge.

It seems as though you could also provide multiple load resistors (selected with a switch), or even a potentiometer, across the secondary. This would allow you to adjust to your liking.

CAR SUB CROSSOVERS

I've enjoyed reading "The Infinite Box: Constructing a Subwoofer" (April and May '02). In part 2 of the article, Mr. Wright discusses crossovers for subwoofers. It's been my experience in car audio that subwoofer crossovers have some special requirements worth mentioning. These should apply to home audio as well.

1. *Crossover Slope:* Assuming the subwoofer may be located some distance from the satellites, it's important that the listener cannot localize the sub. As Mr. Wright mentions, a low crossover frequency (80Hz) will help mask the sub's location. I'd like to add that a steep crossover slope is equally important here. A shallow slope (first or second order) allows some midbass energy to be reproduced by the sub, giving away its location.

   The effect this produces in car systems is that the music is coming from in front of you, but the bass is coming from behind you. Depending on the crossover frequency and sub location, I would recommend a third- or fourth-order slope.

2. *Cutoff Frequency Control:* Just because you're using a fourth-order crossover slope for your sub doesn't mean the satellites need a complementary fourth-order crossover. In fact, it's common for satellites to be used in car audio without any crossover whatsoever.

   But to provide a seamless blend between the main speakers and sub, the sub crossover should have a continuously variable cutoff frequency, which (along with a means for inverting the sub's polarity) will provide the means to achieve a high-quality blend with virtually any satellite configuration. This isn't an obvious result, but computer simulations with sealed and ported satellites—both with and without crossovers—confirm this conclusion.

3. *Subsonic* Filter:* I'll lump this in with subwoofer crossover special requirements, too. A subsonic filter helps you stop torturing your woofer with infrasonic energy, especially with ported, TL, or IB designs.

The folks in car audio are already getting the message. Fourth-order continuously variable sub crossovers are readily available—even some entry-level car amps include this feature (e.g., US-Acoustics USX2050). After you've invested considerable time, money, and effort in your sub system, don't make your sub crossover the weakest link!

Mark Rumreich
Indianapolis, Ind.

**More and more this term is regarded as an oxymoron. Mark Rumreich uses the proper term in his second sentence of that paragraph.–Ed.*

R. O. Wright responds:

I wish to thank Mr. Rumreich for his comments. They are most timely and go into a much more detailed explanation and application of crossovers. G.R. Koonce and I had planned only a basic explanation of crossovers for the article.

DESIGNWARE

While cruising the web, I happened to stumble on a downloadable program that would be of great interest to *audioXpress* readers. It is a loudspeaker-modeling program named WinIsd. This is a very versatile program, which allows you to model sealed, vented, and bandpass designs. It contains a large driver database and has provisions for loading your own driver data. It automatically designs for the flattest response.

One extremely useful feature is the ability to change box tuning and/or box size simply by dragging your mouse. This allows you to see almost instantly how changes affect the speaker response.

A good example of the usefulness of this occurred when I recently modeled a utility speaker for which I wanted to use existing drivers on hand. The box with the flattest response was huge. In almost no time I was able to determine that I could knock off over a quarter of the volume and still have an acceptable response.

This program is available from http://www.linearteam.dk/, and is free. There is also now a "pro" version in its alpha state you can try that should be even more advanced. Hats off to the folks at Linear Team for providing such a useful and well-thought-out program at no charge. This is a true labor of love.

I have never used any other modeling program, so I cannot say how it compares to the "name brand" programs. It would make a good article if someone with experience were to make such a comparison.

Bob McHugh
Walker, Mich.

EDITORIAL RESPONSE

I just read your editorial for July '02 and wanted you to know that your child has grown into maturity quite nicely ("On Turning 32," p. 6). It made me realize that Madisound was born in 1972, and we have grown with you. It has been a good journey, and we appreciate all you have done for audio hobbyists over the years.

Larry Hitch
info@madisound.com

CORRECTION

In our October issue we published a "Classic Circuitry" item featuring the Grommes Stereo Preamp, Model 209. We quoted A.A. Hart, who had sent us the diagram in 1985, and incidentally stated that Grommes has turned exclusively to making public address equipment after 1972.

Ironically enough, just at that moment, a rejuvenated and reorganized Grommes Precision was in the final stages of jumping back into the high-quality audio business. They are very nearly ready to announce their first new tube product, a special limited reissue of their 260A 60W monoblock power amplifier. So we are pleased to correct our old 1985 news by announcing that Grommes Precision is "back." Even better news is that they are considering a later introduction of a kit version of the 260A. We will keep you posted on later developments. Meanwhile, you can look over their offerings at www.grommesprecision.com. ❖

STATEMENT OF OWNERSHIP, MANAGEMENT AND CIRCULATION

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Yard Sale

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Altec 1569A (60W monoblock, 4—EL34), 2 for $400; Teac 3340S 4-channel/4 track, $290; Sherwood S8000 IV tube receiver, $165. Packing/shipping extra. Rad Smith 734-769-2663.

2″ × 5″ horn piezo tweeters, 50W RMS, 3.5–25kHz response, 92dB SPL. Brand new, large quantity available, $5 each plus shipping. For details e-mail Chris at HAL84CH@cs.com.

Heathkit AP-1615 preamp, Heathkit AD-1304 Active Audio Processor (dynamic range expander, dynamic hiss filter), Heathkit Audio Control Center AD-1307 (8 AC Sockets, 2 power meters, separate hookups for: 2 amplifiers, 2 sets of speakers, electrostatic headphones, 2 sets of normal headphones). Manuals included. $100 plus shipping for all. cbest@santa-fe.net.

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