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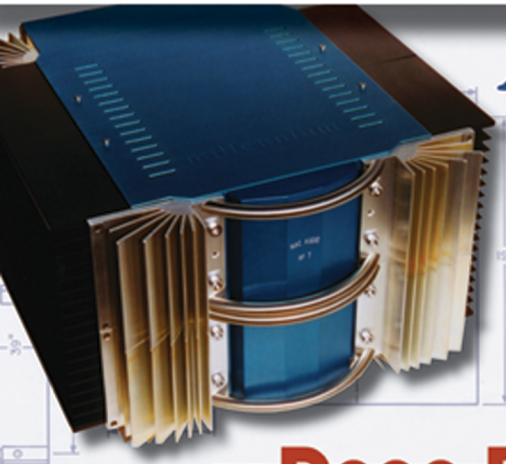

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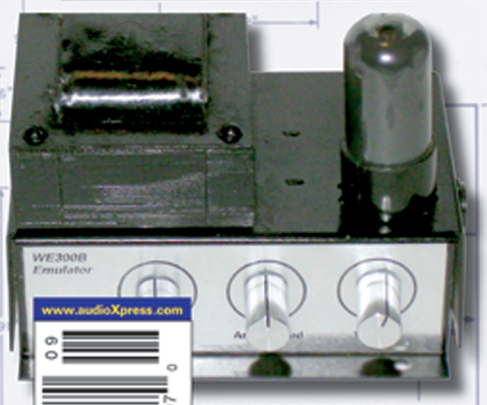

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Stereo Subs for the Swans M1

A pair of 1ft³ sealed boxes with 8″ Scan-Speak drivers extends the response of these excellent mini-monitors, cleanly and powerfully down to 30Hz (typical in-room –3dB frequency).

PHOTO 1: Swans M1.

By Dennis Colin

In *SB* 3/99 I assembled from a kit and reviewed a pair of Swans M1 speakers (**Photo 1**), a vented system with a 5″ mid/bass and ribbon tweeter. I thought then, and do now, that the sonic reproduction is at least as naturally lifelike as that from any other speaker I've heard. They are remarkable in their absence of audible coloration. Reproduction of live-miked voice and bells is uncannily like the live sound. Resolution, image focus, and spatial envelopment are excellent.

The subwoofers described here were designed as a stereo pair, each to be placed directly under the M1, also serving to elevate the tweeter to a typical seated listener's ear height (39″). However, these subs should integrate well (similarly placed) with just about any high-quality small speaker.

BACKGROUND

In my upstairs lab, a 2000ft³ room with 40° wall/ceiling slopes, I'd been using the Swans speakers (positioned as shown in **Photo 2, Fig. 1**) driven by an H.H. Scott 222D 20W per channel tube amp. The very nice sound lacked only two things: deep bass and high power. I took care of the latter by designing and building "Mad Katy," a 125W per channel stereo tube amp (**Photo 3**).

PHOTO 2: Lab/listening room.

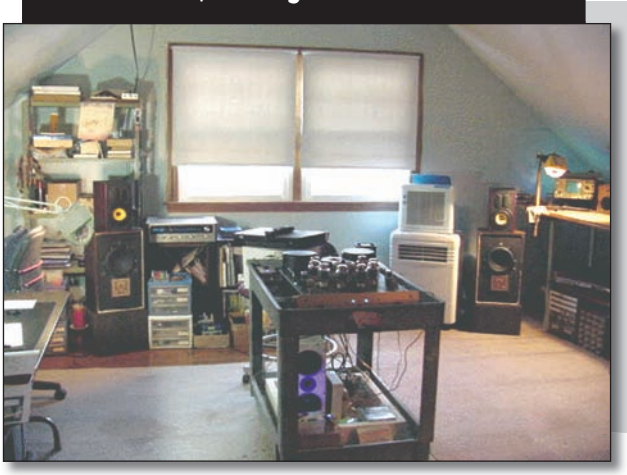


PHOTO 3: "Mad Katy" 125W per channel stereo amp.

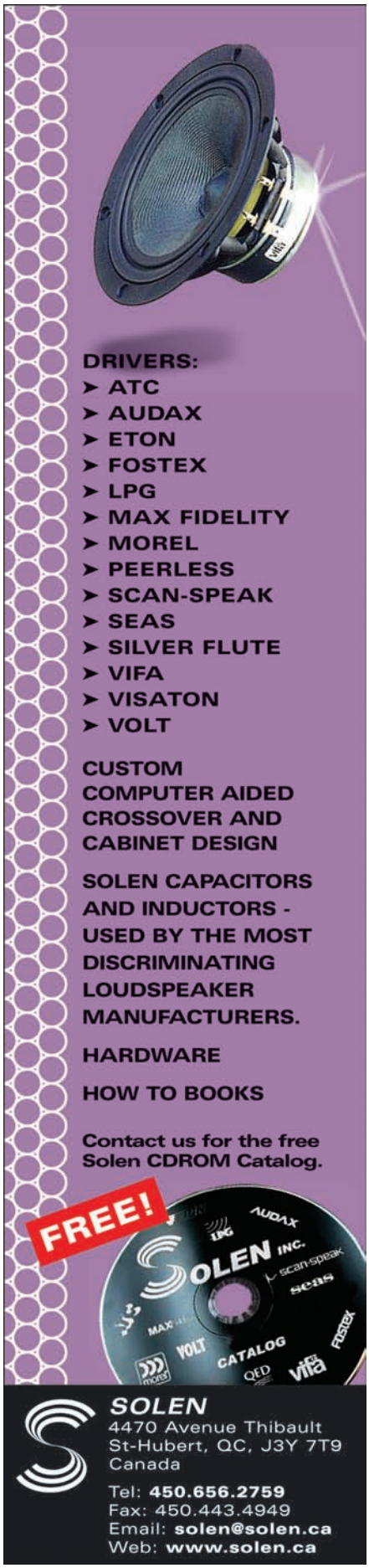


FIGURE 1: Speaker locations in 2000ft³ listening room.

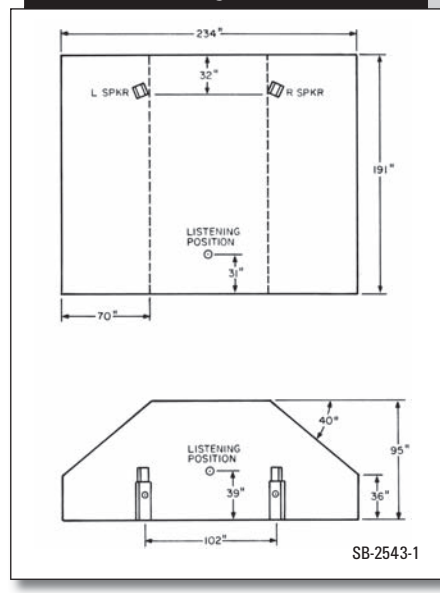
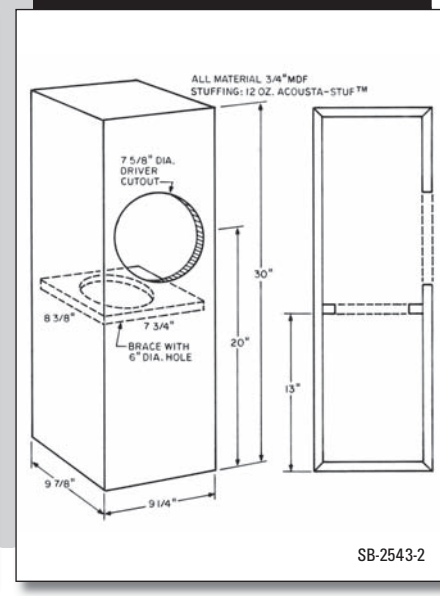


FIGURE 2: Enclosure detail.



THE SUBS

Each Swans M1 was sitting on an old "Advent/2" cabinet, which rested on a 9″ high box, the two elevating the M1 30″ off the floor; tweeter height = 39″ (**Photo 4**). The Advent cabinet, 21½″ × 13″ × 9″, has an internal volume of 1.0ft³.

I installed an 8″ Scan-Speak 21W/8555-00 in each cabinet (sealed) and measured f_{box} = 44.8Hz, Q = 0.72, f_{-3dB} = 44Hz. These are close to calculated values, taking into account the adiabatic-to-isothermal effective volume increase produced by the 12 oz. of Acousta-stuf™.

While the subs shown in **Photos 2** and **4** use the Advent cabinets, I recommend the enclosure specified in **Fig. 2** if you're building from scratch. Flush-mounting the woofer is not necessary, because the 210Hz crossover frequency has a wavelength of 64″.

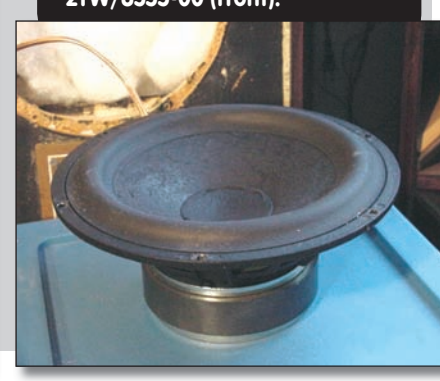
PHOTO 4: M1 and sub.

DRIVERS

These woofers (**Photos 5, 6**) are of very high quality, and unique (for an 8″ driver) in having an f_s of 20Hz. Q_{ts} = 0.31, V_{as} = 136 liters = 4.80ft³, DCR = 5.5Ω, and maximum linear excursion is ±6.5mm (about ½″ p-p). And they *are* linear: in free space, I drove the pair to ½″ p-p excursion at 20Hz, then held them next to my ears like giant headphones. The sensation of clean, ultra-deep bass was incredible!

The enclosure of **Fig. 1** has an internal volume of 1.055ft³, compared with 1.00ft³ in my Advent cabinets. With the same 12 oz. of Acousta-stuf™, the **Fig. 2** cabinet should produce a Q of 0.707, and f_{box} = f_{-3dB} = 43.8Hz.

PHOTO 5: Scan-Speak 21W/8555-00 (front).



DRIVER PLACEMENT

I made some room response measurements from my listening position, 31″ from the back wall, 39″ above the floor, 11′ from each speaker (**Fig. 1**). In the old Advent cabinets I use, the woofer is only 8″ center to cabinet top, so I could have the woofer (center) 8″ off the floor. This did not give the smoothest bass; it was better with some more elevation up to about 20″.

For curiosity, I tried the woofer in the room corner, 8″ off the floor (and about 8″ from the walls). This distance is ¼ wavelength at 423Hz, so this should be the lowest frequency dip from ½ wave reflections. Therefore, all room modes much below 423Hz should be fully stimulated. But this corner placement produced a large peak around 30-50Hz (I could enjoy that, but don't want it), followed by a very ragged response.

I returned the speaker to the location shown in **Fig. 1**. The woofer is 32″ from the front wall and 66″ from the (36″ high) side wall, 54″ from the 40° wall/ceiling slope.

In this position I obtained a smooth bass response with the woofer elevated 16″-24″ off the floor, hence the 20″ woofer height in **Fig. 1**. This elevation is also good for integration with the M1 mid/bass driver: center-to-center spacing is only 12¼″, which at the 210Hz crossover frequency is 0.19 wavelength, or 68°. The crossover is 2nd order; with the woofer electrical polarity inverted and the M1 positive, the two acoustic outputs are in-phase at the CO frequency, on a horizontal axis. The maximum summed acoustic reduction due to the 12¼″ driver spacing occurs on the vertical axis, but this is only –1.6dB, so the total integrated output is not significantly reduced.

FIGURE 3: Crossover schematic.

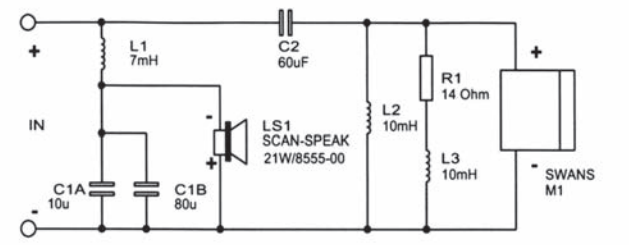


FIGURE 4: Crossover drive signals.

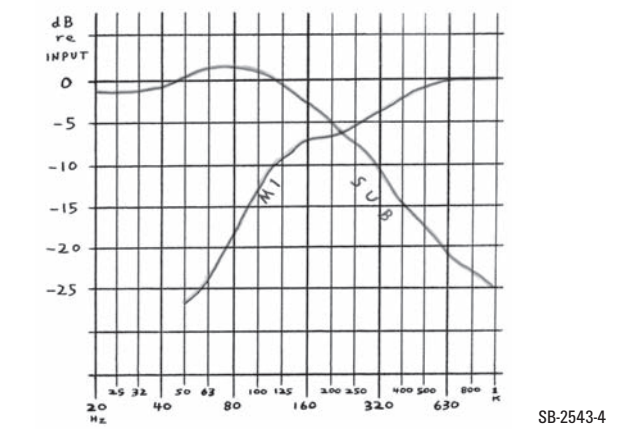
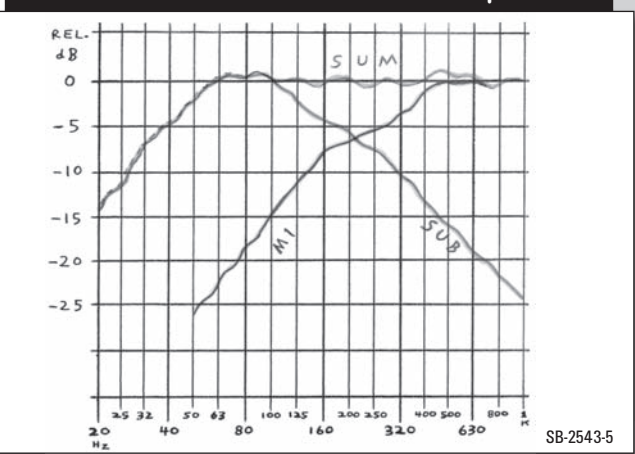


FIGURE 5: Crossover: nearfield acoustic outputs.



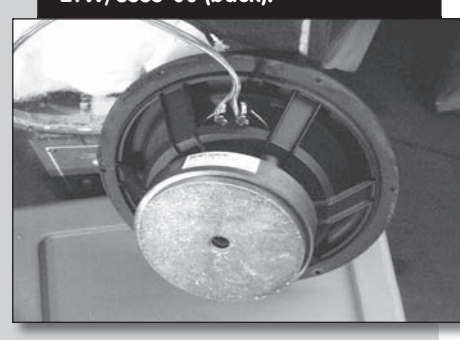
CROSSOVER

The crossover (CO), shown in **Fig. 3** and **Photo 7**, is 2nd order; CO frequency is 210Hz (acoustic), 220Hz (electrical). **Figure 4** shows the electrical responses; **Fig. 5** shows the acoustic responses (mike 1″ away from the cabinets, 5″ below the M1 mid/bass driver center).

The function of R1 and L3 is as follows: The M1 impedance has a peak at 85Hz (upper vented system peak); for a short span above 85Hz the impedance is capacitive. This reacts with L2, increasing its effective impedance. Without R1 and L3, this increased the M1 drive level below the CO frequency, particularly around 120Hz, where the rolloff was only –4dB. R1 and L3 lower the impedance below about 220Hz, thus functioning as a Zobel network. They also cancel some of the M1's capacitive reactance in the region above 85Hz. The rolloff at 120Hz (**Fig. 4**) is now –9dB, which is adequate to prevent overdriving the M1 mid/bass driver on powerful upper-bass fundamentals.

Note the Scan-Speak woofer polarity inversion, which is necessary to cancel that of the 2nd order network at the CO frequency.

PHOTO 6: Scan-Speak 21W/8555-00 (back).



CROSSOVER DESIGN

A 2nd order CO consists of a 2nd order lowpass filter (LPF) and a 2nd order highpass filter (HPF). The following analysis of the LPF also applies to the HPF:

PHOTO 7: Crossover.

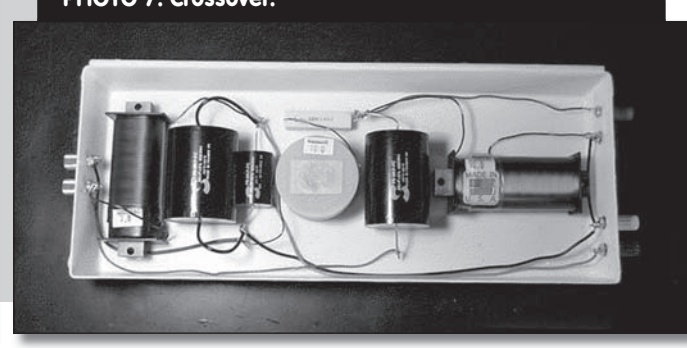


FIGURE 6: 1/3 octave smoothed in-room response, Swans M1 and sub.

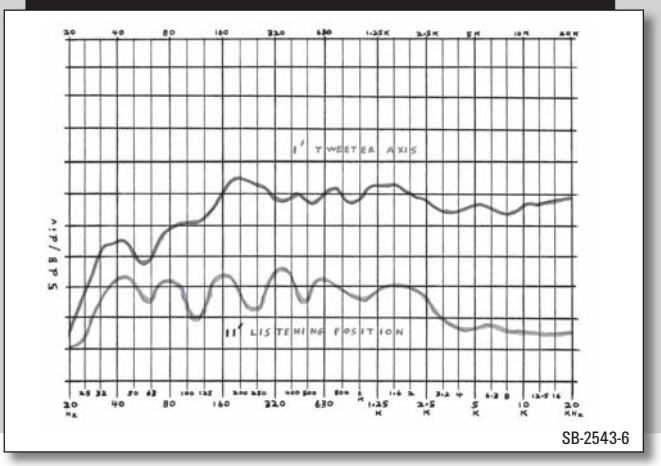
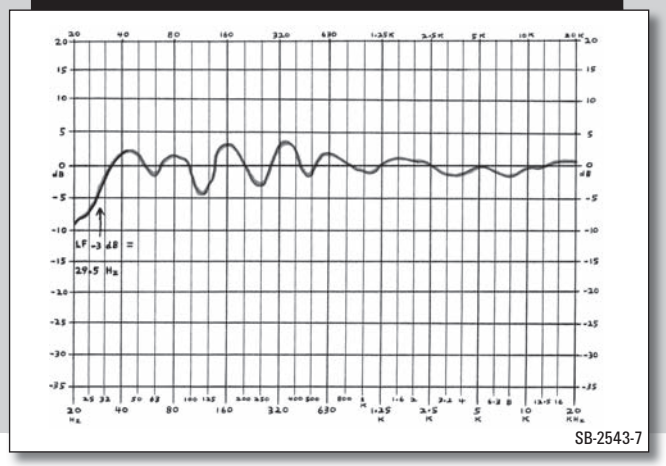


FIGURE 7: 1/3 octave smoothed in-room response, Swans M1 and sub, composite: < 2kHz = 11′ listening position; > 2kHz = 1′ tweeter axis.



The woofer load "sees" L1 in series (to the input), and C1 in parallel, forming a 2nd order LPF. The so-called cutoff frequency is $1/2\pi\sqrt{L,C}$; at this frequency (fo) the reactances of L1 and C1 are equal, and called the characteristic impedance Zo = $\sqrt{L1/C1}$. The ratio of the load resistance (if the load is resistive) to Zo is the filter's Q. The filter's voltage gain at fo is numerically equal to Q.

Because (ideally) each driver's output should be rolled off by 6dB (its input voltage reduced to half) at fo, so that with no phase shift the two drivers will combine to unity gain, the Q should be 0.5. So as a starting point (which will likely need adjustment), select inductor and cap values to have a reactance that's twice the driver's impedance at the CO frequency.

LISTENING POSITION MEASUREMENTS

The lower curve in **Fig. 6** is the response at the listening position (31″ from the back wall, 39″ high, 11′ from the speakers). The mike responds, of course, to both direct and reverberant sounds, and this room is fairly live. Thus the rolloff above 2kHz is from narrowing tweeter dispersion (mostly only vertical with the excellent M1 ribbon tweeter) and decreasing room reflectivity.

Figure 7 is a composite; below 2kHz is the listening position response; above 2kHz is the response 1′ from the tweeter, on axis. The rationale for this is as follows: At the lower frequencies, the ear has less discrimination both directionally and temporally within short delay intervals; thus spectral judgments tend to integrate direct and short-term reverberant (typical of listening rooms) sounds. But at the higher frequencies, spectral judgments tend to largely favor the first-arrival (direct) sounds.

So the composite response of **Fig. 7** is at least a crude approximation of the perceived in-room spectral balance. To my ears, and as a testament to Igor Levitsky, designer of the M1, the sound is very natural, spectrally balanced, and uncolored.

MAXIMUM BASS OUTPUT

From 20–40Hz, I drove the subwoofers to a level where I estimated 3% THD (noticeable, but not objectionable, low-order distortion), and noted the cone excursion and drive voltage. At 20 and 30Hz, these were 0.5″ p-p and 11.7V RMS; at 40Hz they were 0.6″ p-p and 14.0V RMS.

With both stereo subs driven in phase, the SPL at the listening position was approximately 106dB at 40Hz, 103dB at 30Hz, and 96dB at 20Hz. Scan-Speak data states ±6.5mm (0.51″ p-p) maximum linear excursion, which agrees with these tests. Maximum (not linear) excursion is stated as ±12mm (0.94″ p-p), which provides 5dB peak headroom (albeit with significant distortion) above the maximum linear SPL figures I estimated.

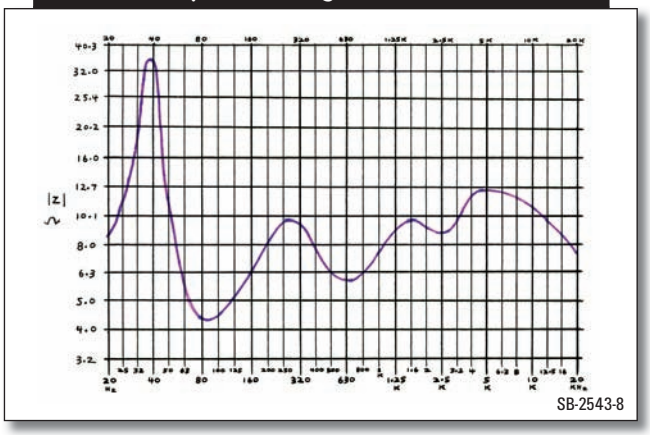
TABLE 1

Subwoofer Parts List (excluding cabinets)
Quantity: 1 each per channel
Source: Madisound

CO Schematic reference	Description	Part no.	cost
	8″ woofer, Scan-Speak	21W/8555-00	$162
C1A	Solen "fast cap," 80µF	CP80	17.75
C1B	Solen "fast cap," 10µF	CP10	3.80
C2	Solen "fast cap," 60µF	CP60	14.20
L1	Steel Laminate Inductor, 7.0mH	SL 7.0	14.05
L2	Steel Laminate Inductor, 10.0mH	SL 10.0	16.20
L3	Iron core bobbin inductor, 10.0mH	10.0 L8	10.00
R1	wire-wound resistor, 14Ω, 15W	15R14	0.40
	Terminal cup	D-cups	3.15

Total per channel	$241.55
Stereo pair	$483.10
Swans M1 kit, pair	$449.50
Swans M1 assembled, pair	$899.00

FIGURE 8: Impedance magnitude.



I think that clean 30Hz production at 103dB SPL in a 2000ft³ room from a pair of 8″ woofers "speaks" very well for these Scan-Speak drivers. Expensive at $162 each, but worth it.

SPEAKER PLACEMENT

The (sub) woofer center is 20″ off the floor; this distance is ¼ wavelength at 169Hz. At this frequency the reflection path (woofer to floor back to woofer) is ½ wavelength, whose phase cancellation produces the infamous "floor dip." But if you place the woofer twice this distance (40″) from the front wall, this reflection path will be a whole wavelength, returning in-phase, thus producing a reinforcement. This can largely cancel the floor dip.

In my setup (**Fig. 1**) the woofers are 32″ from the front wall, but this is because of space constraints in this crowded lab/listening room. Apparently the 32″ distance is good enough, because the bass response at the listening position (**Figs. 6, 7**) is relatively smooth (for a room response, that is). **Figure 8** shows the impedance magnitude.

BIGGER BROTHER

For more bass power, you could use the Scan-Speak 25W/8565-01, a 10″ woofer with the same motor structure as the 8″ unit. In a 2.0ft³ sealed box with 16 oz. of Acousta-stuf™, approximate parameters are f_{-3dB}, Q = 0.71. Maximum linear deep-bass output will be about 3.5dB greater than with the 8″ driver. Suggested sealed box dimensions are 12½″ square × 30″.

CONCLUSION

With the addition of these stereo integrated subs (making the M1 part of a floor-standing three-way system), excellent transparency and resolution are maintained with high power handling and deep bass extension. The system can handle the full output of the "Mad Katy" amp (155W per channel at clipping) without obvious distortion. Peak SPL in this 2000ft³ room is about 106dB; maximum average SPL with typical music waveforms is about 95dB; and with typical LF room gain, the LF –3dB rolloff point is about 30Hz. *aX*

Sources

Madisound Speaker Components
(608) 831-3433
info@madisound.com

Swans Speaker Systems, Inc.
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sales@swanspeaker.com

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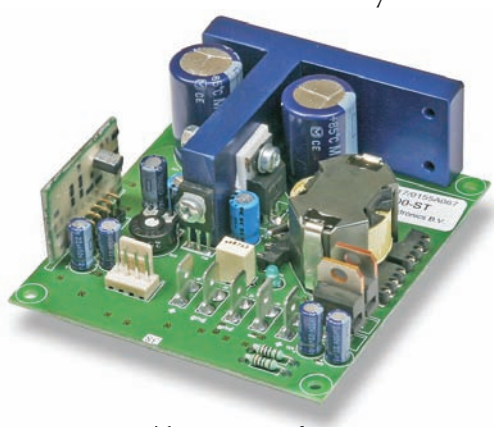

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Crowhurst's Twin-Coupled Amp Revisited

A return to the work of a noted designer broadens the appeal of his amp design.

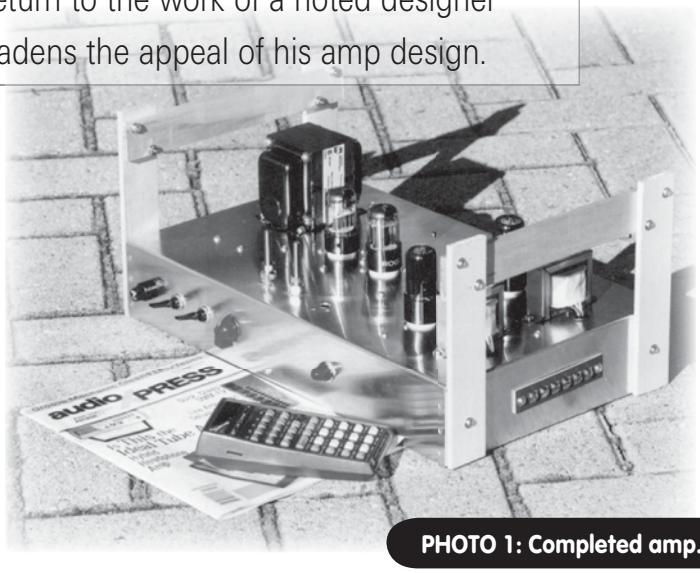


PHOTO 1: Completed amp.

By John Stewart

While completing my earlier version of Crowhurst's amplifier[1], I began to think there might be still more changes that would enhance its operation and at the same time further lower its cost. As well, not everyone wants or needs more than 30W as I had built into the previous amplifier. With those thoughts in mind, my goal became the design of a flexible circuit that could be easily adapted to the preferences of many. The result is shown in **Photo 1**.

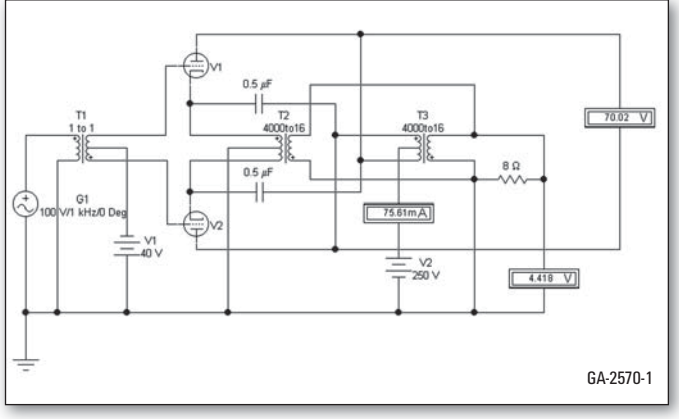
As background for those who may have missed the earlier article, Norman Crowhurst has provided for us a circuit that is a variant of the unity-coupled design introduced in the late 40s by F. H. McIntosh and G. J. Gow. Instead of an expensive and difficult-to-build bifilar-wound output transformer as used in the McIntosh amplifier, Crowhurst uses two separate identical but rather ordinary output transformers (OPTs). He calls the result his "Twin Coupled Amplifier"[2,3,4].

One of these OPTs is connected push-pull in the plate circuit, just as in an ordinary amplifier. The other is connected in the cathode circuit, again in a push-pull mode. The OPT secondaries are connected in series or parallel to match whatever load is to be connected. The result is that you have one-half of all the output stage gain applied as local voltage negative feedback (NFB).

In the design of this version of Crowhurst's amplifier I had two objectives in mind. I always like to present a circuit that performs well but at the same time can be built by others economically with easy-to-find parts. The other objective was to show step by step where minor circuit changes could result in improved performance. I've included plenty of test data so that you can decide how far along you would like to go. You might find some of the circuit and construction ideas in this article useful in other projects as well.

This version of Crowhurst's circuit does not provide the epitome of performance, but nevertheless for the cost and effort does perform extremely well. I believe it will easily outperform any other vacuum tube amplifier of similar cost and quite a few costing considerably more. If you should decide to build this amplifier, you should be aware that voltages in the circuit are dangerous, even to those of us with years of experience.

FIGURE 1: Basic circuit.



THE BASIC CIRCUIT

In the simplified schematic (**Fig. 1**), each of V1 and V2 is an imaginary triode with a mu of 5.4. Tubes used in the original circuit required a plate-to-plate load impedance of 8k. That means each of Crowhurst's output transformers should reflect 4k back to the tubes. The secondaries would each need to be 16Ω in order to match to an 8Ω loudspeaker.

In order to defeat the problems caused by leakage inductances between windings and stray capacities, capacitive coupling is applied to the ends of the transformer primary windings. This capacitive coupling also ensures that the load will be split evenly between the plate and cathode circuits should the secondary windings be series connected. Depending on circuit constants, when the secondary windings are paralleled, the resulting local NFB at the cathodes yields a damping factor (DF) of about 4, even before any loop NFB is applied.

Furthermore, Crowhurst does not

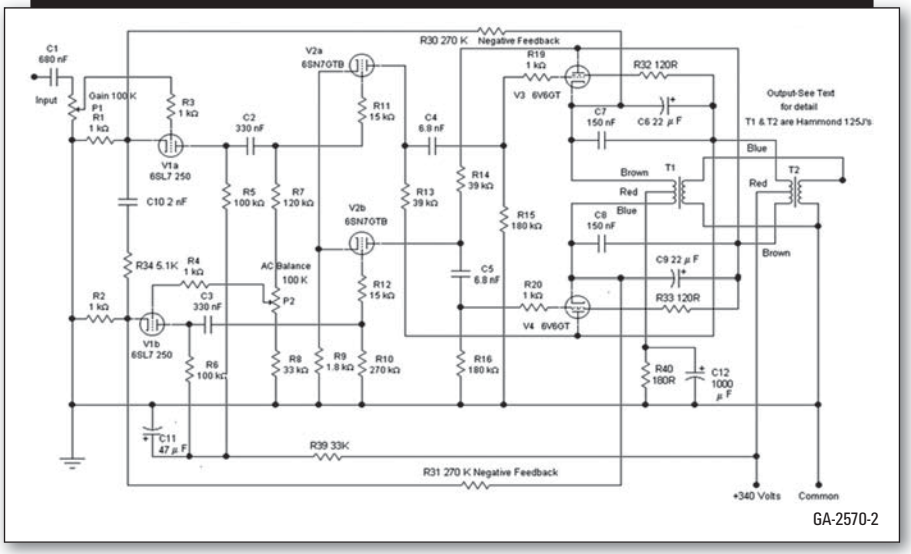
include the secondaries of his output transformers within the loop NFB circuit. That improves stability considerably. If you like, it is possible to operate this version of the amplifier with no loop NFB, and you would still have a very respectable damping factor.

If you apply a 100V signal to the grids of V1 and V2 as in the simplified schematic, the result at the output transformer primaries will be about 70V. With the impedance ratio of each transformer set at 4000 to 16, you would get about 4.4V into an 8Ω load. All this is based on ideal transformers.

Crowhurst's circuit[5] depends on good coupling of the two secondaries to each of their primaries to distribute the audio load equally between the plate and cathode circuits. At the higher frequencies he has overcome leakage reactance effects by using 500nF capacitors to tie the plate and cathode signals together. The original circuit used 6BQ5/EL84 pentodes so that each of their screens was wired directly to the opposite plate. That way they still operated as pentodes because the screens were held at a constant potential from their respective cathodes. It was also push-pull throughout, including the driver and predriver. As a result, Crowhurst was able to use push-pull NFB from the output cathodes to the amplifier input.

Also, you should notice that the output stage needs a large driving voltage. That problem needs to be solved in any amplifier in which part or the entire load is in the cathode circuit. Bootstrapping of the driver stage is used successfully in both the McIntosh and Circlotron. Crowhurst uses it in his Twin-Coupled Amp as well.

FIGURE 2: Cathode biased amplifier.



THE POWER AMPLIFIER

For the simplest version of the power amplifier, I've used cathode bias (**Fig. 2**), which ensures that as the tubes age

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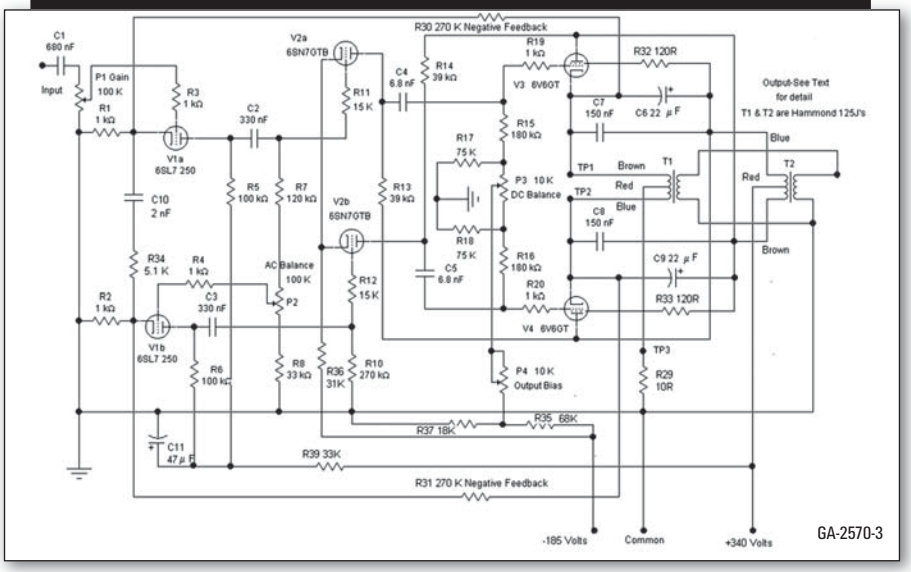
FIGURE 3: Fixed bias amplifier.



the circuit will be—to a large degree—self adjusting and trouble free. The detailed schematic shows resistors in all of the grid leads including the screen grids. Their primary purpose is to damp out any possibility that any stage of the amplifier might oscillate at radio frequencies. This does happen on occasion in some circuits and sometimes produces strange and difficult to track down performance problems, even in the simplest circuits. I've also included another amplifier schematic for a fixed bias version of the amplifier (**Fig. 3**).

All of the circuit designators I have used in the schematics are the same as I used in the previous article. I hope this helps those who have followed these projects. Most versions of this circuit are easily driven to full output by a ½V input.

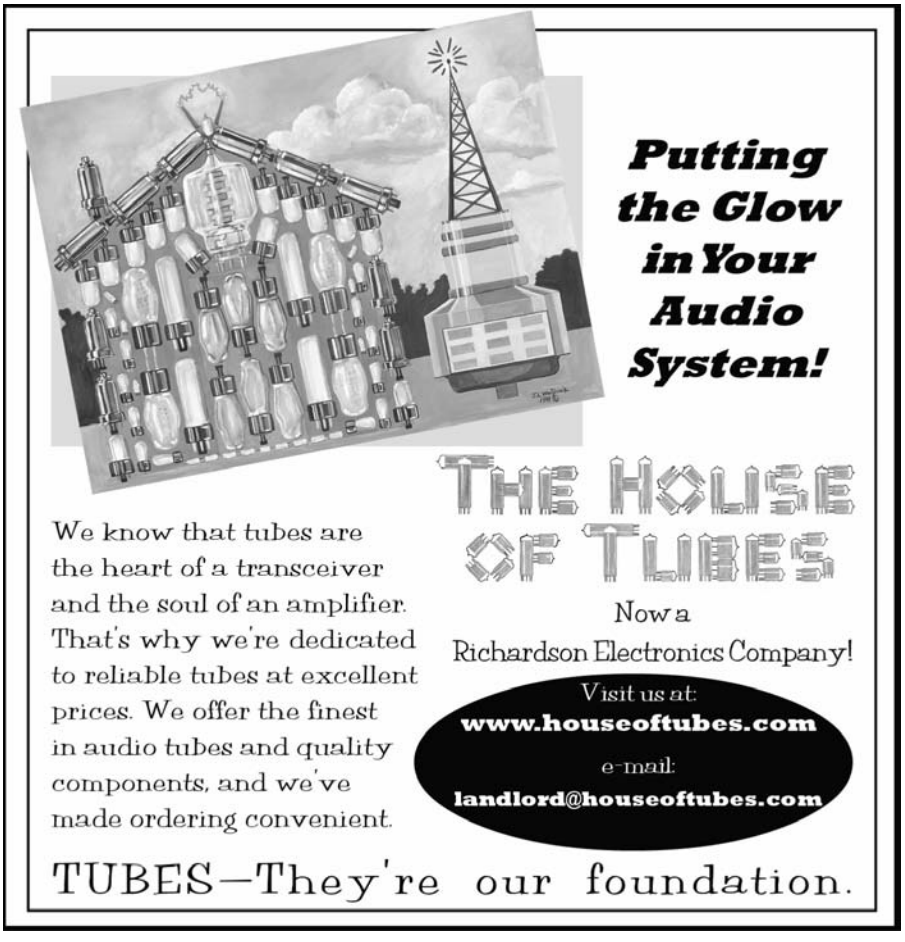
In order to keep the cost to a minimum, I've chosen a pair of 6V6GTs as the output tubes, but there are other possibilities. You will need to be somewhat careful with the selection of the output tube type because large heater to cathode voltage swings occur during normal operation. The 6V6GT allows up to ±200V difference between the heater and cathode.

In the cathode bias version of the amplifier, while delivering 10W to the load, the voltage at the 6V6GT cathodes varies from -100 to +140V. The offset is due to the approximately +20V bias at the cathodes. Some versions of the amplifier are capable of delivering close to 15W to the load, resulting in even larger voltages at the cathodes. The cathode biased version of the amplifier manages more than 12.5W before clipping. With the 180R cathode bias resistor R40, the zero signal current in the PP 6V6GT pair will be about 90mA.

For the fixed bias version of the amplifier, a good bias setting for the output tubes is 60mA. You can do that by adjusting the output bias pot P4 so that a voltage drop of 0.6V is developed across the 10R current sampling resistor R29. With no signal applied, you can set the DC balance of the output tubes by adjusting the pot P3 so that there is 0V between the test points TP1 and TP2.

You could also use a pair of 6AQ5As in this circuit, provided you keep the maximum plate dissipation in check. However, its close relative, the 6AQ5, would not be suitable, because its maximum heater to cathode voltage rating is only 90V. Another tube that would work is the 6K6GT. The 6F6 family won't, although you will see later in the section on results data I've included some test data.

In looking through the tube data sheets, I was surprised to find the 6BQ5/EL84s not suitable, even though Crowhurst had used them. The 6BQ5/EL84 H-K rating is only 100V. However, he probably correctly assumed that no one would normally listen to his amplifier at anywhere near full power on a continuous basis. Another condition in which the 6BQ5 might be possible to use is with an 8k load if such a combination of OPTs were available. That allows the power to be delivered at a lower voltage.



For this circuit I have settled on a total load impedance of 10k, which suits the 6V6GTs well. That would be 5k plate-to-plate and 5k cathode-to-cathode. For that you can use a pair of Hammond 125J output transformers (OPT). Fortunately, the 125J has a primary to secondary turns ratio that fits this application well.

In order to achieve a very low output impedance that translates to a high damping factor (DF), you need to be able to connect the secondaries in parallel. I was not able to do this when using the Hammond 125E Universal OPTs in the earlier article, so another problem is solved! In this mode the OPTs will run about double the recommended impedances, thus the output voltages are about 1.4× and the currents are about 0.7× normal for a given output power.

In order for the 6V6GTs to run as true pentodes, the screen grids are cross coupled to the opposite plate of each of the tubes. However, for the sake of completeness, I have included data in the tables for both ultralinear and triode operation. You will see by comparison that these alternatives do not offer much in the way of advantage as they do while in an ordinary push-pull circuit. As connected, most of their gain is already applied as NFB through the cathodes, even while connected as pentodes.

In the ultralinear connection the screen leads are simply taken to the DC HV supply, allowing you to run the plates at a higher voltage for more power if desired. That is what I did in the amplifier of the previous article. For triode operation I simply connected the screens directly to their plates through the damping resistors.

The circuit used to connect the 6V6GTs to the OPTs in this amplifier results in just 1250Ω from each cathode and plate to earth. These low impedances reduce the effects of capacitive shunting and thereby improve the high-frequency performance. Nevertheless, you will need to be careful with proper phasing of the transformer output interconnections.

THE DRIVER

I'm again using Crowhurst's original circuit with minor variations, although you will see as we progress there will be some changes or enhancements to improve the performance. Rather than using a 12AX7 and 12AU7 as voltage amplifiers, I used a 6SL7GT and 6SN7GT. That was partly for cosmetic purposes, but I also like working with the larger sockets in an experimental amplifier. They simply made changes easier as the design progressed and they looked better-fitted to my rather large experimental chassis. If you were to build this amplifier, any of the combination of tubes I've mentioned would work well.

I believed Crowhurst's application of a 1.8k tail resistor in the driver 12AU7 while his plate resistors were 39k was rather on the small side. A higher value tail resistor results in a better balanced drive to the output circuit. In this part of the circuit you need as much voltage at the plates as possible in order to drive the output tubes to full power.

Some way must be devised so that a larger tail resistor could be used, without using up more of the plate supply, which is done in circuits such as the Mullard 5-20. There are quite a few ways of developing a negative voltage with which you can supply a higher resistance tail resistor. Here are two of them and out of that three possible circuit solutions.

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In the first variant, I already had a 100R resistor inserted into the center tap (CT) lead of the high voltage winding of the power transformer, which I will discuss as part of the power supply. During normal operation there are 30V pulses at twice the line frequency between the common lead and the transformer CT. These pulses can be rectified and then put into a simple circuit to yield a regulated -14V supply.

By returning the 6SN7GT tail resistor to the -14V supply, it is possible to add another 2.55k for a total of 4.35k in the tail (**Fig. 4A**). I found the setting of the AC balance pot P2 to be much less critical after this minor circuit change. You will see later that there was some improvement in performance from the resulting distortion measurements.

Even better performance resulted when I replaced the tail resistor with a simple current source consisting of a 2N3053 NPN transistor, a common red LED, and two resistors (**Fig. 4B**). I used the 2N3053 because I had one in my parts box, but a 2N2222 would work just as well. If you were to use this optional circuit as your 6SN7GT tail circuit, you could eliminate AC balance pot P2 from the circuit. If you choose to do that, then R7 should be 150k while R8 is 100k. Their junction is then taken directly to R4, the one k resistor on the grid of V1b.

The third possible solution, which I will cover separately, results if you intend to connect your output circuit for fixed bias operation. If so, it allows a tail resistor of 31k and very good balance of the drive to the output tubes.

THE PREDRIVER

For the input tube I used the 6SL7GT connected in a circuit essentially the same as that used by Crowhurst. The AC balance adjustment I mentioned was included in Crowhurst's second article on the amplifier. He included the control, which is useful for setting an operating point for least distortion at the output. I found this control became almost irrelevant when the tail resistor of the 6SN7GT was increased, so much so that it could be replaced by fixed resistors, as I have mentioned.

You will see that the main loop NFB is again wired back to the cathodes of the 6SL7GT. In the results tables you will see considerable performance data both with the NFB connected and with no NFB.

THE OUTPUT TRANSFORMERS

While this circuit arrangement lets you use low-cost output transformers, there is still a need to make some circuit adjustments to accommodate them. One of these concerns the high-frequency rolloff in any transformer response. This rolloff is caused by leakage reactance and the self-winding capacity of the primary.

In order to get a reasonable high-frequency output, it is possible to modify the feedback loop by combining C10 and R34 in the circuit of the 6SL7 input cathodes. This CR pair reduces the feedback at the higher frequencies, just as a similar circuit does in Crowhurst's original version. If you were to use other than the Hammond 125J output transformers, then you would need to adjust the CR pair for that arrangement.

Another transformer problem is the imbalance of the interwinding capacity between the primary and secondary. This is not the same winding capacity I referred to. For most low-cost audio transformers, you will find the secondary winding is wound around one end of the primary.

Whatever circuit you might connect this to acts as though an additional capacity had been connected to one end of the circuit. In the Hammond 125J that amounts to about 200pF. If you were to test for this parasitic capacity, you would find it at the end of the winding connected to the brown lead. For a single-ended circuit you simply connect the primary so that this extra capacity appeared at the power supply lead of the transformer, away from the plate connection.

However, in a PP circuit there will be considerable imbalance, because the extra capacity will show up on one of the signal leads. As a result you need to connect the output transformers in this circuit so that both their primaries and secondaries are connected in opposite phase to each other. You can see on the amplifier schematic that the primary leads of the transformers are connected in such a way that the parasitic capacities will balance. If you are unfortunate enough to make the other possible connection, you will find response to a square wave to be unsymmetrical. Output is also considerably reduced before distortion shows up.

Because the primaries are connected in opposite phase, you will need to connect the secondaries so that the resulting

FIGURE 4: Alternative cathode returns for the 6SN7GT driver.

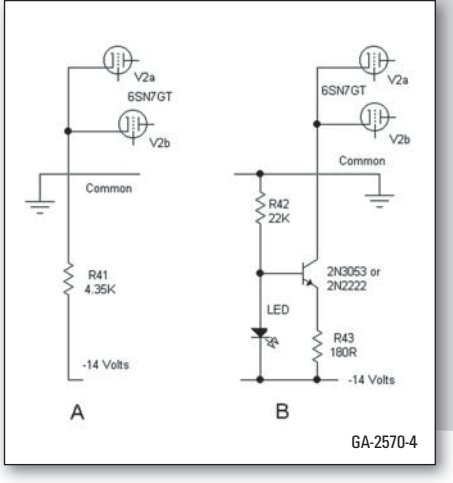
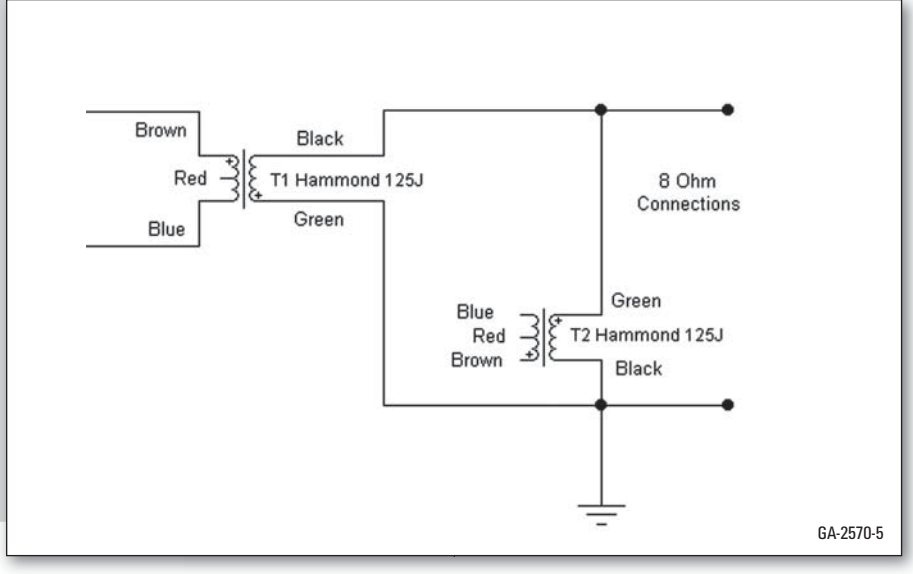


FIGURE 5: Output transformer detail.



signal adds in order to drive your loudspeakers properly. Refer to the output transformer detail (**Fig. 5**). If you intend to drive a 4Ω loudspeaker, then substitute the yellow leads where I have indicated the green lead connection. Either way, you may leave the OPT secondary circuit ungrounded if you like.

THE POWER SUPPLY

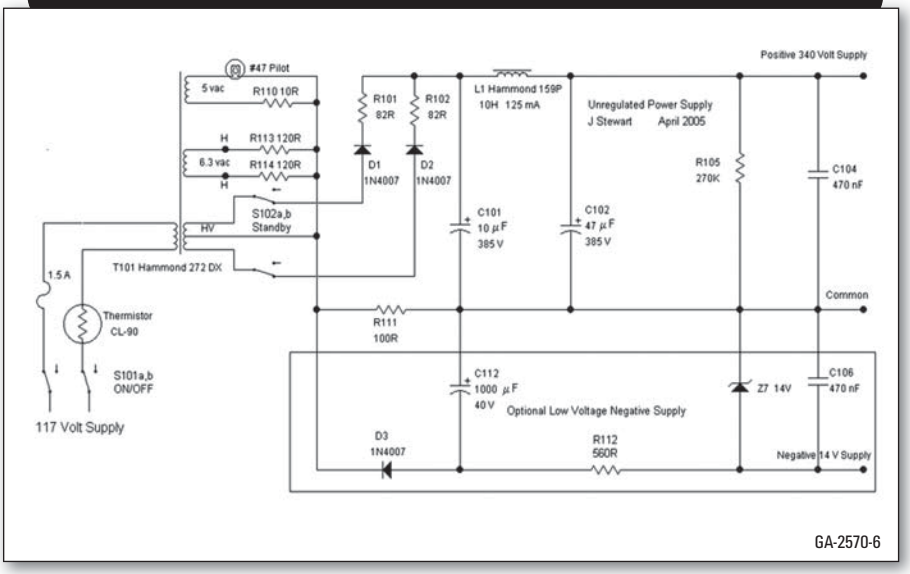
The power supply is relatively straightforward in its simplest form (**Fig. 6**). All of the power is easily provided by a Hammond 272DX transformer. It is rated for 125mA DC, while the amplifier connected for cathode bias operation runs about 100mA at no signal.

Rather than using a vacuum tube rectifier, I prefer silicon diodes such as the 1N4007, so in this amplifier the 5V winding is free to be used for other purposes. In this case it simply powers the pilot light. If I were to use a vacuum tube rectifier, my preference would be the 5V4GA, as it was in many projects in earlier days. Since the 272DX was designed for use with vacuum tube rectifiers, I've included an 82R resistor in series with each of the silicon diodes. This serves two important purposes.

First, I want to limit the charging current to the first capacitor, just as a vacuum diode would. Second, if one of the rectifier diodes were to fail shorted, the resulting current would be limited at least until the fuse blows. You need one resistor in series with each diode to do that properly.

The first electrolytic capacitor after the rectifiers is 10µF. To many familiar with current vacuum tube thinking, this may seem much too small. The ripple at this point is about 50V peak-to-peak. However, the simplest form of the amplifier is cathode biased so that the current change from no signal to maximum is not particularly great.

FIGURE 6: Unregulated power supply.



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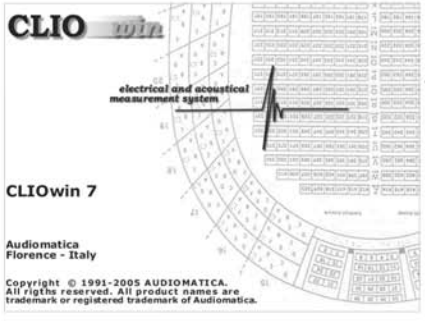

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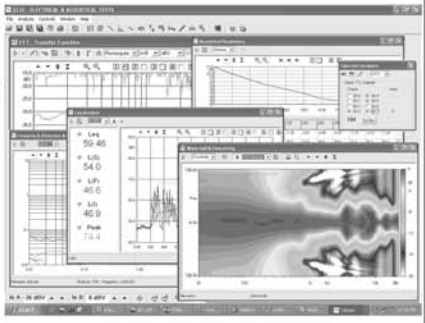

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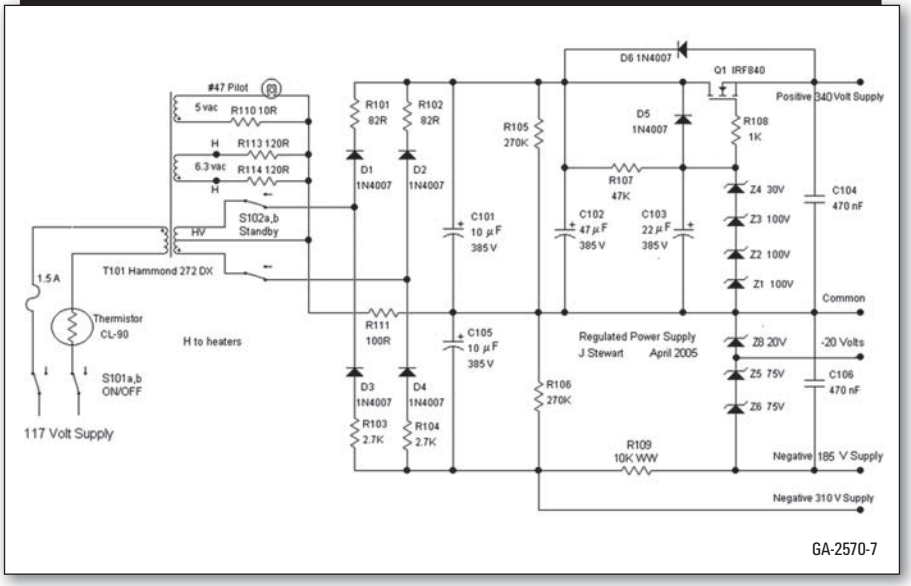
I've used a 47µF cap at the output of the filter so that even the lowest frequencies possible in this amplifier are well handled. After going through the Hammond 159P 10H choke, the HV ripple is less than ½V. The filter resonance is at 7.8Hz.

The combination of a rather small capacitor input to the filter and the 82R resistors in series with the rectifier diodes comes close to providing the voltage I needed to run the output circuit without exceeding the 6V6GT dissipation ratings. But I needed a little more drop, so I added the 100R resistor in series with the 272DX HV CT. That is where I was able to derive a -14V supply I mentioned previously in the text. The result was about 313V between the cathodes and plates of the 6V6GTs after voltage drops in the output transformers.

The 6.3V winding is rated at 4A. In this circuit it need deliver only 1.8A, so I have been very conservative and reliability will be very good. As further protection, I included a 2A fuse and a CL-90 thermistor, which is a very useful device ensuring a soft start at turn-on for the tube heaters. After reaching operating temperature, its voltage drop is only a volt or two.

FIGURE 7: Regulated power supply.



POWER SUPPLY ENHANCEMENTS

My final version of the amplifier output circuit was to run in the fixed bias mode, so I needed better power supply regulation. I simply used a cut-down version of the regulator I had used in the circuit of my previous article (**Fig. 7**). This circuit is somewhat simpler because the voltages involved are lower. Only one of the IRF840 FETs is required.

The regulator includes soft-start so that when the standby switch is closed the 6V6GT cathode currents don't rise to dangerous levels as they would without that delay. The FET is connected simply as a source follower whose gate is driven

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by the controlling voltage on an RC time constant. The upper limit voltage is set by a series of zener diodes. The physically small 1k resistor in the gate lead guards against the FET oscillating on its own.

For experimental use, having a standby switch included in the circuit helps me a lot in making changes and adjustments while I work on the circuit. You don't ordinarily need this, but you might want to have one for your own reasons. I use an ordinary two-pole switch connected to the power transformer secondary to unhook the rectifier diodes. Although the DC is defeated, you should always keep in mind that the transformer HV is still in the chassis and can give you a nasty bite.

You may have looked at the photo of the rectifier sub-assembly and noticed four rectifier diodes. The extra diodes are used to wire the output stage for fixed bias operation. The negative supply provides the fixed bias potentials.

I've used current limiting resistors, each of 2.7k in series with the diodes into their 10µF filter cap. At that capacitor you will find a little more than –300V. After passing through a 10k WW resistor into the zener diode circuit, I measured –185V, even though the diodes totaled only 170V.

A look at the diode spec sheet for the 75V 1N5368B indicated that heating in the diodes causes the zener voltage to drift up. These zeners are rated at 5W, but that would be in ideal conditions. As used in this circuit, a little more than 1W is dissipated in each of the 75V zeners.

I had originally used spaghetti on the zener leads. It turns out the zeners depend to a large degree on cooling through the lead structure. Even after removal of the spaghetti, there was still quite a voltage rise. However, I later didn't find this to cause any problems.

If you find that the power line in your area tends to run on the low side, you may need to use a different rectifier limiting resistor set than I have shown in the schematics and parts list. You could reduce R101 and R102 to 47R in order to get enough voltage to properly run the positive side of the power supply. I don't think you will need to change the values of R103 and R104. Be sure to leave some resistance in the circuit so that if a rectifier diode should fail there will be a margin of safety for that part of the circuit.

RESULTS DATA

I have included several tables of test results in order to better illustrate why I made the various modifications. For the total harmonic distortion (THD) and intermodulation distortion (IMD) tests, I first set the AC balance pot for a 2nd harmonic minimum while the output was 1W. I left the AC balance pot in that position for the remaining tests within each test group. The NFB resistors R30 and R31 remained the same for all measurements.

Measurement of the DF was done using the rated load versus the no load voltages at the output. From that I used the internal resistance at the amplifier load terminals to compute the DF. The distortion data, given in percent, was taken at Rload of 8Ω and 1000Hz, except as noted.

The test signals were provided by an HP 200CD wide-range oscillator and a GW GAG-810 sine- and square-wave audio generator. I took the THD measurements using an HP 334A analyzer and

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A470S reproduction of the Dynaco A470 output transformer as used on the Dynaco (tm) ST70 and MkIV amplifiers

A431S a reproduction of the Dynaco A431 output transformer as used on the MkII and MkIII amplifiers and Sunn guitar/bass amplifiers.

Other stuff we're working on: A420 (6600 ohm 30W) and A442 (4300 ohm 60W with tertiary screen winding)..

measured the figures for the individual harmonics up to the 5th with a Pico Technology ADC-216 virtual instrument running into a PC. With a 16-bit resolution, it has a dynamic range of 96dB. The resulting data sets cross-check well.

I made IMD tests by the SMPTE (Society of Motion Picture & Television Engineers) method using a 4:1 ratio of 60Hz and 7kHz signals. Other IMD tests were done by the CCITT method, using equal signals of 1 and 1.6kHz.

Take great care in order to get accurate results from the tests. I spent quite a bit of time chasing down ground loops in the measurement setup. Using a PC as part of the setup adds another problem because the display raster can cause interference if you are not careful. This shows up as a false indication while using a digital spectrum analyzer such as the Pico ADC-216 and is known as "aliasing," an artifact of the digitizing process. You must be familiar with advanced measurement techniques in order to properly evaluate the resulting data.

Table 1 is a comparison of the various possible 6SN7GT tail circuits. Looking at the results, you would not think there is any significant advantage of one circuit over any other. However, as more resistance was added into the tail circuit, the setting of the AC balance control became less critical.

If you don't have any distortion measurement equipment available, this is an important feature to keep in mind. The best option in that case is the simple constant-current (CC) circuit using the 2N3053 NPN transistor connected to a negative supply.

Table 2 compares pentode operation of the 6V6GTs with the same tubes and circuit connected for triode and ultralinear (UL) operation. As you would expect, pentode operation provides the most power and triode operation the least, just as in an ordinary amplifier. The DF for triode and ultralinear operation is only marginally different than while pentode connected, so there is no advantage with respect to that.

The only other advantage while not pentode connected is possible in the UL mode. Because the screens would then be supplied separately from the plates, it becomes possible to run the plates at higher voltage and hence more power could be available. That is what I did in the previous project.

Table 3 shows the advantage of using the 2N3053 CC source when compared to the original 1.8k in the tail of the 6SN7GT cathodes. Distortion is reduced at all levels and particularly at 12.5W, where the THD is reduced from 0.6% to 0.06%. That is a significant drop and is much less critical to adjustment of the AC balance pot.

In **Table 4** I have made minor changes so that the amplifier is set for fixed bias operation from its regulated power supply. It now clips at about 14.5W. You can see by the measurement results that even as close as 14W is to clipping, the distortion is still quite low. You can see in tests H and I the minor difference in performance when the 6SN7GT cathode tail is taken from the 31k resistor as compared to the 2N3053 CC source. But here again the CC version is much less critical to adjustment of the AC balance pot.

In tests J and K the front-end 6SL7GT circuit is set up for differential input and zero NFB (**Fig. 8**). I used a DPDT switch S1a,b to alternate between the regular circuit and differential input. Another DPDT switch (not shown) is used to dis-

TABLE 1 Twin Coupled Amp Performance Summary

Output 6V6GTs ultralinear connected & cathode biased

CONDITION	WATTS	2H	3H	4H	5H	THD	Other
A) 6SN7GT driver cathodes through 1.8k to common	1	0.037	0.029	~	~	0.055	cathode current = 91mA
	3	0.024	0.066	~	~	0.078	
	10	0.039	0.634	0.056	0.364	0.640	
B) 6SN7GT driver cathodes through 4.3k to -14V	1	0.045	0.025	~	~	0.060	
	3	0.052	0.046	~	~	0.075	
	10	0.014	0.650	0.057	0.436	0.870	
C) 6SN7GT driver cathodes through 2N3055 CC source to -14V	1	0.044	0.019	~	~	0.054	DF = 9.43
	3	0.046	0.028	~	~	0.056	NFB = 12.6dB
	10	0.094	0.432	0.048	0.311	0.450	
D) 6SN7GT driver cathodes through 31k to -160V	1	0.046	0.024	~	~	0.058	
	3	0.042	0.029	0.006	~	0.060	
	10	0.067	0.433	0.060	0.310	0.440	

TABLE 2 Twin Coupled Amp Performance Summary

6V6GTs cathode biased; 6SN7GT driver cathodes through 2N3055 CC source to -14V

CONDITION	WATTS	2H	3H	4H	5H	THD	Other
E) Output Triode Connected	1	0.048	0.021	0.004	~	0.058	DF = 9.89
	3	0.041	0.028	0.003	~	0.057	NFB = 10.8 dB
	6	0.061	0.811	0.108	0.587	0.860	cathode current = 91mA
C) Output UL Connected	1	0.044	0.019	~	~	0.054	DF = 9.43
	3	0.046	0.028	~	~	0.056	NFB = 12.6 dB
	10	0.094	0.432	0.048	0.311	0.450	cathode current = 91mA
F) Output Pentode Connected	1	0.047	0.017	~	~	0.055	DF = 9.92
	3	0.041	0.023	~	~	0.051	NFB = 18.4 dB
	10	0.035	0.045	0.003	0.006	0.060	
	12.5	0.033	0.044	0.003	0.007	0.060	cathode current = 91mA

TABLE 3 Twin Coupled Amp Performance Summary

Output 6V6GTs Pentode connected & Cathode Biased

CONDITION	WATTS	2H	3H	4H	5H	THD	Other
F) 6SN7GT driver cathodes through 2N3053 CC source to -14V	1	0.047	0.017	~	~	0.055	DF = 9.92
	3	0.041	0.023	~	~	0.051	NFB = 18.4dB
	10	0.035	0.045	0.003	0.006	0.060	
	12.5	0.033	0.044	0.003	0.007	0.060	cathode current = 91mA
G) 6SN7GT driver cathodes through 1.8k to common	1	0.044	0.029	0.006	0.004	0.062	DF = 8.91
	3	0.101	0.082	0.016	0.006	0.125	NFB = 14.4 dB
	10	0.271	0.212	0.004	0.032	0.300	
	12.5	0.262	0.490	0.074	0.175	0.600	cathode current = 91mA

connect the NFB for zero feedback. The differential circuit provides a much better balance for the drive signal to the output 6V6GTs than is possible with the original front-end circuit.

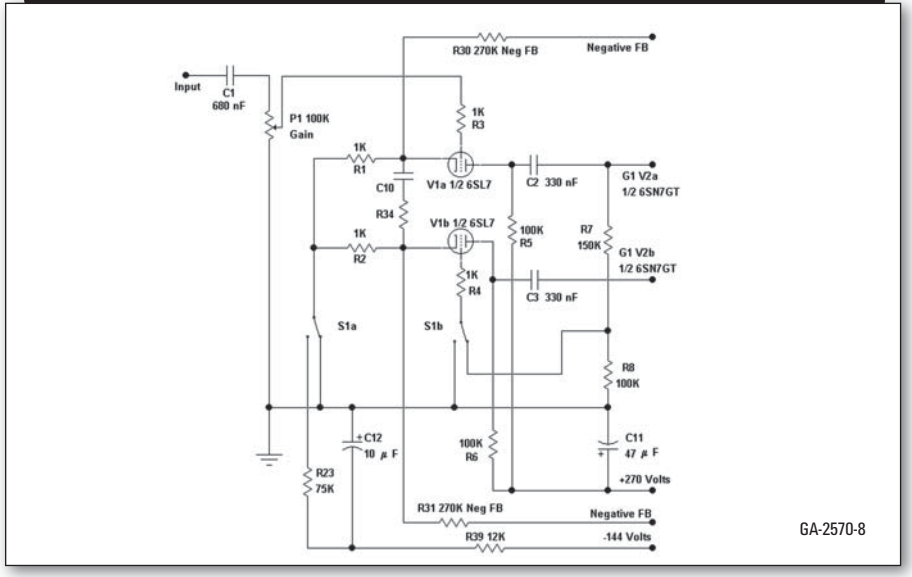
To optimize this mode of operation the CR time constant in the cathodes of the 6SL7GT needs to be changed. C10 is then 6.8nF while R34 becomes 2.2k. If you so choose, you can also easily connect this front end to a differential source. If you use the switches to change the operation mode, be sure the amplifier is shut off before switching. Otherwise, damage to your loudspeakers or the amplifier could result. Also, there is another comparison of the CC tail with the 31k resistor tail on the 6SN7GT driver tube.

The tests L, M, N, and O are provided so that you can see how well twin triodes of the 6BK7/6BQ7/6BZ7 family compare to what you would normally use. These tubes are often overlooked for this kind of operation, yet they are surprising performers as you can see by the results. I did a similar comparison in the earlier project. It was easy to try again because the socket adapters I had made up and the tubes used are still here on my workbench.

If you decide to go with tubes of this type, the substitution is rather easy. For the 6SL7GT, you can wire a 6BQ7 straight into the circuit with no changes at all. For the 6SN7GT substitution you will need to add one each 2.2k resistor in series with each of the 6BQ7 cathode leads. Their junction then simply goes to whatever cathode tail circuit you choose. These resistors serve to reduce the gain of the 6BQ7 so that it is more like that of the 6SN7GT. At the same time they provide a degree of linearization of the 6BQ7's characteristics.

Also notice in this group of tests the negative supply for the 2N3053 has increased from 14V to 20V. The new volt-

FIGURE 8: Alternative front end.



age is derived differently from what was used in the cathode biased version of the amplifier. Refer to the regulated power supply schematic.

Tests P and Q return you to the 6SL7GT/6SN7GT front end. Test P is regular operation with the CC source as the tail for the 6SN7GT cathodes. Test Q shows the performance with the front end wired for differential input and zero NFB.

In test R I have set the test signal to 60Hz. Even though the OPTs are on the small side they perform very well. In test S, a pair of cathode biased 6F6s have been used in the output circuit. If you use 6F6s the cathode bias resistor is 330Ω.

For the SMPTE IMD test I set the two signal sources so that the amplifier output was 8W as indicated on the Metrahit 29S Wattmeter. That is equivalent to an 11.76W sine wave. The resulting IMD sidebands in the amplifier output measured about 40dB below the 7kHz signal.

Refer to the spectrum for the differential input mode with zero loop feedback showing 7kHz and the upper and lower 60Hz sidebands. In the CCITT IMD test I found the highest IMD product to be always more than 60dB down from either of the test signals, even at 14W equivalent. The 600Hz difference signal was always the largest of these.

TABLE 4 Twin Coupled Amp Performance Summary

Fixed Bias / Regulated PS / 6V6GTs Pentode connected

CONDITION	WATTS	2H	3H	4H	5H	THD	Other
H) 6SN7GT driver cathodes through 29k to -160V	1	0.043	0.024	~	~	0.056	DF = 8.95
	3	0.044	0.035	~	~	0.061	NFB = 15.4 dB
	10	0.044	0.020	0.010	0.007	0.042	
	14	0.029	0.015	0.018	0.028	0.070	cathode current = 60mA
I) 6SN7GT driver cathodes through 2N3055 CC source to -14V	1	0.044	0.022	0.004	~	0.056	DF = 9.59
	3	0.038	0.033	~	~	0.056	NFB = 18 dB
	10	0.029	0.012	~	0.007	0.035	
	14	0.033	0.043	0.006	0.004	0.055	
J) 6SN7GT driver cathodes through 29k to -160V Differential input circuit	1	0.143	0.083	0.005	~	0.145	DF = 3.75
	3	0.257	0.182	0.007	0.013	0.270	NFB= Zero
	10	0.486	0.054	0.010	0.051	0.470	
	14	0.571	0.213	0.046	0.030	0.580	
K) 6SN7GT driver cathodes through 2N3055 CC source to -14V Differential input circuit	1	0.115	0.088	0.010	~	0.124	DF = 3.62
	3	——————No Data——————				0.254	NFB = Zero
	10	0.450	0.074	0.007	0.046	0.440	
	14	——————No Data——————				0.560	
L) 6BQ7s substituted for the driver & predriver. Driver cathodes through 2N3055 CC source to -14V	1	0.040	0.036	~	~	0.066	DF = 6.76
	3	0.043	0.068	0.007	~	0.086	NFB = 14dB
	10	0.058	0.054	0.005	0.011	0.075	
	14	——————No Data——————				0.160	
M) 6BQ7 driver & predriver Balance Pot reset to Series I Driver cathodes through 2N3055 CC source to -14V	1	0.042	0.030	~	~	0.062	2x 6V6GT
	3	0.045	0.051	0.003	~	0.076	cathode current = 46mA
	10	0.054	0.033	0.008	0.010	0.170	
	14	0.059	0.109	0.005	0.013	0.286	
N) 6BQ7 driver & predriver Driver cathodes through 2N3055 CC source to -20V	1	0.040	0.016	~	0.006	0.050	2x 6V6GT
	3	0.036	0.007	~	0.013	0.050	cathode current = 46mA
	10	0.039	0.156	0.010	0.019	0.175	
	14	0.048	0.242	0.014	0.051	0.280	
O) 6BQ7 driver & predriver Driver cathodes through 2N3055 CC source to -20V	1	0.051	0.039	~	~	0.072	DF = 7.66
	3	0.056	0.071	~	~	0.094	NFB = 15dB
	10	0.089	0.029	0.007	0.009	0.100	2x 6V6GT
	14	0.098	~	0.006	0.010	0.108	cathode current = 60mA
P) Return to 6SL7/6SN7 6SN7GT driver cathodes through 2N3055 CC source to -20V	1	0.045	0.024	~	~	0.054	DF = 9.74
	3	0.043	0.033	~	~	0.056	NFB = 17.9dB
	10	0.047	0.015	~	~	0.045	
	14	0.032	0.029	0.034	0.030	0.071	cathode current = 60mA
Q) Differential input circuit 6SL7GT cathodes through 75k to -155V	1	0.117	0.113	~	~	0.15	DF = 3.72
	3	0.269	0.250	0.011	0.014	0.320	NFB = Zero
	10	0.537	0.218	0.019	0.060	0.560	
	14	0.690	0.071	0.025	0.020	0.700	cathode current = 60mA
R) Same as P above But at 60Hz	1	0.081	0.051	~	0.020	0.100	DF = 9.20
	3	0.167	0.042	~	0.167	0.160	NFB = 18dB
	10	0.329	0.167	0.046	0.037	0.320	
	14	0.352	0.189	0.034	0.153	0.420	
S) PP 6F6s as output 6SN7GT driver cathodes through 3.9k to -20V	5	0.225	0.140	0.013	0.009	*	DF = 7.64
	9	0.582	0.605	0.200	0.259	*	NFB = 14.6dB

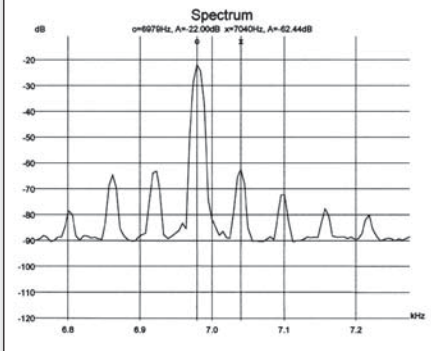
* Indicates no THD Data

CONSTRUCTION

This amplifier was built as an experimental unit and not as a showpiece. As a result, I did not spend much time on cosmetics. My primary objectives were the design of a reliable and affordable vacuum tube amplifier of outstanding performance and flexibility, easily built by someone of moderate experience. As usual, I will leave it to you, the builder, to decide what the appearance should be for your own listening environment.

I built my version of this amplifier on a rather large Hammond 1444-32 chassis measuring 17 x 10 x 3″. It has a matching Hammond 1434-30 bottom plate. I like to have room to move around while

FIGURE 9: Input 60 and 7000Hz mixed 4:1 output is 8W on the watt meter, equivalent to 11.76 sine wave.



GA-2570-9

making changes, adjustments, and measurements.

The power and two output transformers are all mounted on the top surface while the choke is inside the chassis. You may not need that much space and will find you can get by with something smaller, in particular if you should choose to use a miniature tube set. For example, if you go with an output section using the 6AQ5As and the 6BQ7s in the driver circuitry, you will need somewhat less space power. You could get along with a power supply consisting of something like a Hammond 270DX and one of their 157M chokes.

You can see in **Photo 1** that I've again installed the wood pieces on the chassis ends as I have for earlier projects. During the development stage this structure supports the amplifier while it is inverted on the test bench. The wood pieces also work well as handles so that I can easily move the amp from the workbench to the shelf.

Several of the circuit parts were built as sub-assemblies, rather than trying to put them together while inside the chassis. You simply attach two of the terminal

PHOTO 2: Constant current supply.

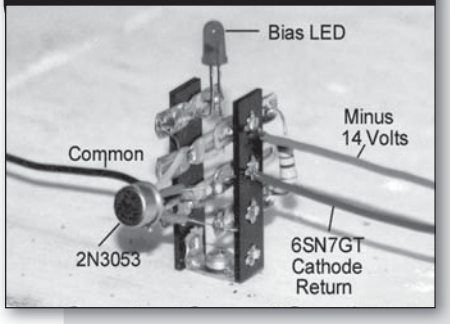


PHOTO 4: Negative filter and voltage regulator.

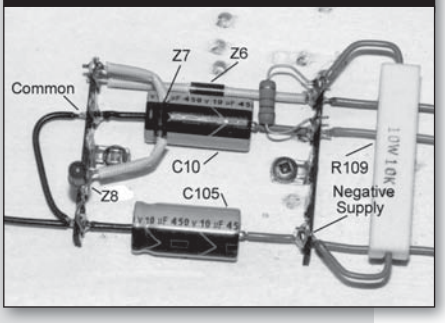


PHOTO 3: -14V supply.

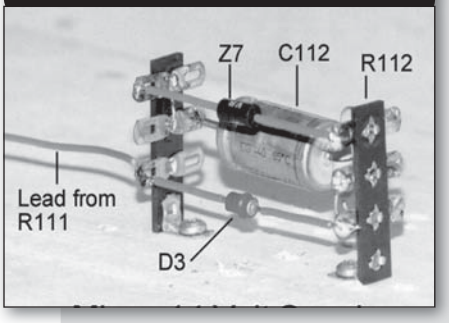
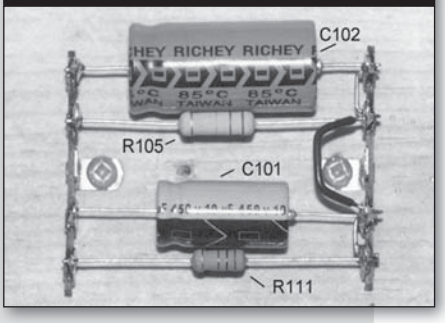


PHOTO 5: Positive filter on assembly board.



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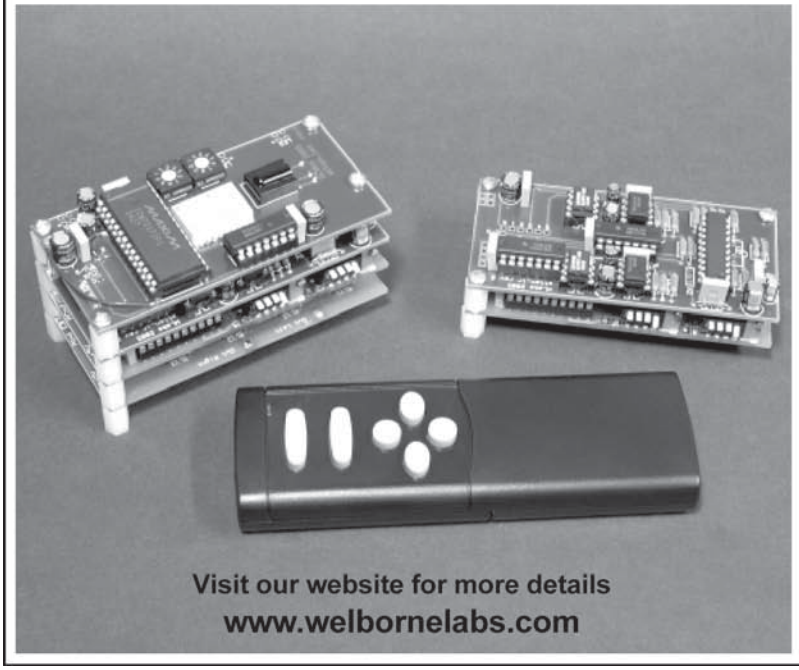

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PHOTO 6: Positive voltage limit and delay.

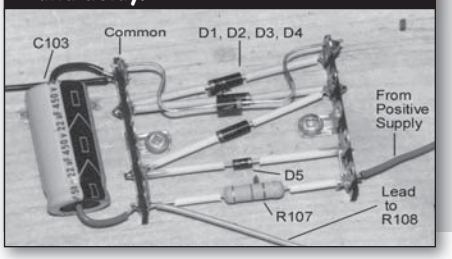
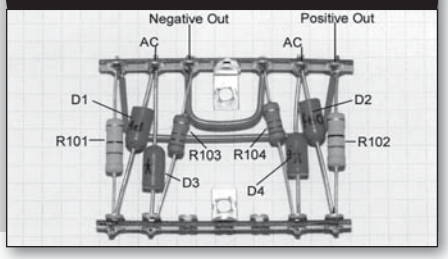


PHOTO 7: Rectifier module.



strips to a clean scrap of wood. Then solder all of the parts in place on the terminal strips. You can then mount the whole assembly in the chassis and wire into the circuit. I found this technique to be very useful while building a "one off" circuit. The various component parts are identified in **Photos 2-7**.

The heatsink for the IRF840 FET is nothing more than a scrap of ⅛″ thick aluminum extrusion. It measures in total about 2.5″ x 4″ and is more than enough to dissipate heat generated in the FET. Don't be tempted to feel the heatsink for temperature as you might a solid-state amplifier. There are almost 400V on the sink during normal operation! The heatsink is supported and insulated from the chassis by three

EQUIPMENT USED IN THE DEVELOPMENT OF THIS AMPLIFIER

HP 334A	Total Harmonic Distortion (THD) Analyzer
HP 200CD	Wide Range Audio Generator
GW GAG-810	Sine and Square Wave Audio Generator
Rohde & Schwarz	Dual Channel, 100MHz Oscilloscope
Metrahit 29S	Precision RMS DMM and Power Meter
Pico Technology ADC-216	Dual Channel Spectrum Analyzer/Oscilloscope
Pico Technology M1053	Differential Probe
Powerstat	8A Variable Transformer

PARTS LIST

1—Amplifier

C1	680nF, 630V	
C2, C3	330nF, 630V	
C4, C5	6.8nF, 630V	
C6, C9	22µF, 450V	
C7, C8	150nF, 630V	
C10	47nF, 630V	
C11	1000µF, 40V	
C12 - See Text	10µF, 450V	Diff Amp Input Stage Only
LED	A common red LED	
P1	100k	Gain
P2	100k	AC Balance (optional)
P3	10k	DC Balance
P4	10k	Output Bias
Knob for Gain Control		
3 Pot. Locks for the Bias, AC and DC Balance Adjustments		
Q1	2N3053 or 2N2222	
R1, R2, R3, R4, R19, R20	1K, ½W	
R5, R6	100k	
R7	120k	
R8	33k	
R9	1.8k	
R10	270k	
R11, R12	15k, ½W	
R13, R14	39k	
R15, R16	180k	
R17, R18	75k	
R21 to R28	Not Assigned	
R29	10R, ½W	
R30, R31	270k (Negative Feedback)	
R32, R33	120R, ½W	
R34	5.1k	Front End FB Adjustment
R35	68k	
R36	31k	
R37	18k	
R38	Not Assigned	
R39	12k	
R40	180R	Cathode Bias
R41	4.3k	Optional, see Figure
R42	22k	"
R43	180R	"

All resistors are 2W except as noted

T1, T2	Hammond 125J Output Transformer
V1	6SL7GT
V2	6SN7GTB
V3, V4	6V6GT

4 Octal Sockets

RCA Phono Connector (Input) AES p/n S-H310
One 6-Screw Terminal Strip (Output) AES p/n S-H317P

2—Power Supply

C101, C105	10µF, 385V
C102	47µF, 385V
C103	22µF, 385V
C104	470nF, 630V
C106	220nF, 630V
C107 to C110	Not Assigned
C112	1000µF, 40V
CL-90	NTC Thermistor
D1 to D6	1N4007 or equivalent
Q1	IRF840 FET
R101, R102	82R
R103, R104	2.7k
R105, R106	270k
R107	47k
R108	1k
R109	10k 10W WW
R110	10R
R111	100R
R112	560R
R113, R114	120R

All resistors are 2W except as noted

S101, S102	DPDT Switch
T101	Hammond 274BX Power Transformer
Z1, Z2, Z3	1N5378B, 100V Zeners
Z4	30V Zener
Z5, Z6	1N5368B, 75V Zeners
Z7	14V Zener
Z8	20V Zener
Chassis	Hammond 1444-32, 10 x 17 x 3 Chassis
Bottom Plate	Hammond 1434-30, 10 x 17 Bottom Plate
Small Aluminum Extrusion	Heatsink/Mounting for Q1

nylon screws.

I especially thank the people at Hammond Manufacturing for their generous assistance towards the successful completion of this project. *aX*

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barrel, cone, truncated cone, cube, cylinder, domed cylinder, truncated cylinder, ellipsoid, square prism, opt square prism, reg polygon prism, slanted front prism, truncated edge prism, 4-sided pyramid, 3-sided pyramid, truncated pyramid (trapezoid), sphere, wedge, 2-chamber BP cylinder, 3-chamber BP cylinder, 2-chamber BP prism, 3-chamber BP prism

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Odds & Evens Revisited

Including how to improve the performance of two-stage amps.

It has been a few years since I wrote "Odds and Evens" (*July 2001, p. 56*), but since then my work has continued to focus on the transfer curves of tubes and tube amplifiers.

By Graham Dicker

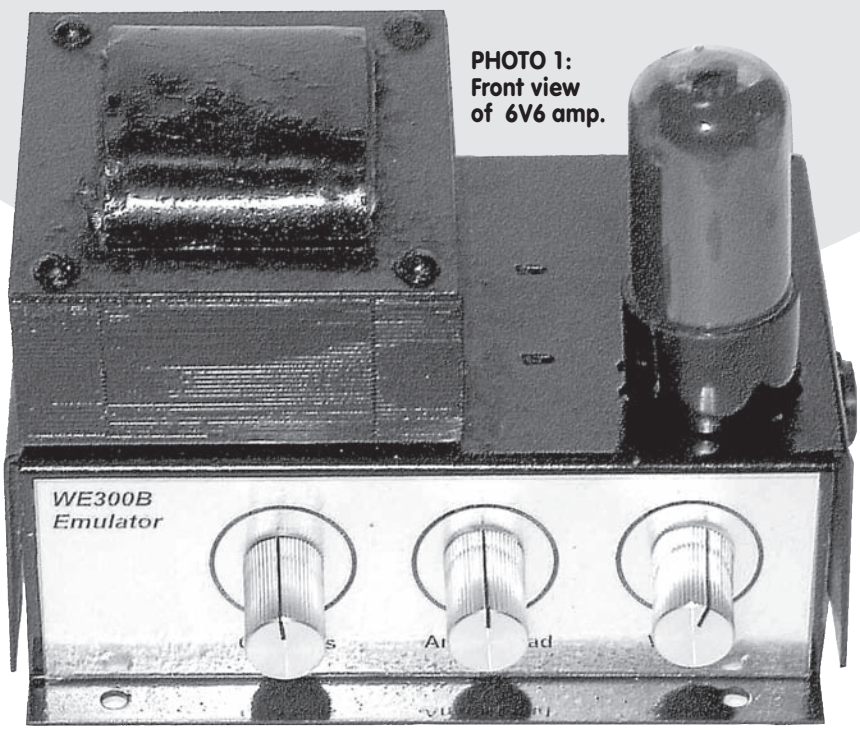


PHOTO 1: Front view of 6V6 amp.

THEORY ONE:

Two transfer curves equal and opposite in phase will cancel any distortion (from original article). One of my previous ideas is it is important to try to optimize each stage's performance as much as possible, using nonlinear devices—i.e., LEDs and diodes—in the local feedback path of each tube stage to make them as linear as possible.

My current thinking has changed somewhat. The basic idea of an even number of stages with the same transfer function to reduce distortion still holds, but I believe that trying to linearize each stage does not matter; you simply select stages which have similar transfer functions over the signal excursions actually used. This then broadens the scope to include dissimilar technologies, which are now co-compatible.

THEORY TWO :

The net algebraic sum of the two transfer functions over the used section of each curve should provide a linear end result. For example, a tube output stage could be complemented with a bipolar transistor and still achieve a linear result without feedback. Indeed, the end result need not be linear.

For example, should you choose to replicate the transfer function of a favorite tube (say a WE300b), you could now do so by using these two simple rules. Many amplifiers over time have accidentally ended up with a nice-sounding result, as trial and error experimenters and designers alike have found that certain tubes and operating parameters combine to produce the desired result. This article attempts to explain why.

LINEARITY NOT NECESSARILY A GOOD THING

Strange for me to say it, but as I listen to people telling me about amplifier sound (as I have done for the last 40 years or so), I have concluded that as an engineer I have always worked on the premise that an amplifier should do as its name implies (amplify a signal) and nothing else. This has always been the standard for 60 years or so. Today many of the top-sounding amplifiers certainly add coloration to the sound, but it is this character that is pleasing to the ear and sought after by many audiophiles and musicians.

This takes me back to the ol' black art days, when an amplifier may test terrible but sound great (and vice versa). Without doubt my conclusion (while still keeping an open mind and ear) is that the sound from a good amplifier using WE or other 300B or 2A3 tubes has more sonic merit than the sound produced from other tubes. I found, on initial investigation, a clean character was mainly due to the very linear nature of the tubes used and the previously stated rules.

In others the amplifier was not linear but had substantial second harmonic distortion. However, it was this sound that the owners liked. These amplifiers usually employed a less than linear triode or pentode which was triode-con-

nected as the voltage amplifier.

Tubes such as the 300b and 2A3s do offer better linearity than many others available, especially when you look at the plate transfer curves. This inherent linearity is what normally allows for low distortion without feedback in an amplifier. As I've already noted, you can employ other methods without feedback to achieve the same end result.

To many, the sound characteristic is worth the price paid. Being a little frugal myself, I thought it would be good to be able to achieve the sound qualities these "golden" tubes have to offer but without the cost. In my research I found that tubes such as the Eimac 300 TH and 100 TL—among other transmitting tubes—also offer great linearity for use in audio, so you should not discount them just because the anode lead comes out the wrong end of the tube.

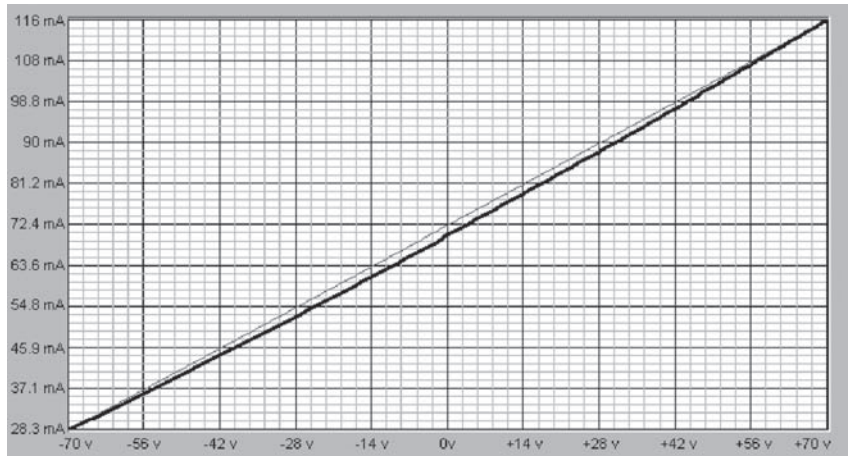
My current work has tried to emulate the transfer curves of classic tubes by using common off-the-shelf tubes. Because pentodes and tetrodes are readily available at low cost, these were the focus of my work (how to make a sow's ear sound like a silk purse or something like that).

EMULATING THE WE300B

Having recently purchased some NOS WE300Bs and some RCA 2A3s for customer projects and experimentation, I quickly managed to empty my bank account. These tubes, while nice triodes, will become harder to find over the next ten years or so. My advice is to buy now and put them away as an investment (you can add the Telefunken VF14 to this list as well). With this in mind, I spent some time trying out common tubes such as the 6V6, 6L6, 807, and EL34 in triode-connected mode, with generally disappointing results in comparison to the 300Bs and 2A3s. There must be a better way.

My experimentation in using composites in audio technology over the years has led me closer to a solution than I had imagined. My new theory, based on my odds and evens rethink, started me looking again at the transfer curves of the "golden" tubes with the thought of how to get a common tube to behave in the same manner, and emulate the transfer function. Initial experiments with the generic 6V6 were quite promising, but ended up with an extremely complex network of diodes, zeners, VDRs, and resistors to make the same shape transfer curve of a WE300B. The net result apart from circuit complexity was that the emulated device had less voltage gain than the 300B it was to replace.

FIGURE 1: Transfer function of the 300B exported from Glassware's SE AmpCAD.



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The next complication was to try to answer the question, Which part of the curve do you wish to emulate? A little explanation is due here. Obviously a 6V6 has less anode dissipation than a 300B so you could not use it as a one-for-one replacement. However, you could use other tubes such as a KT88 or 12E1 in a similar environment.

With this restriction, do you now look at how a 300B would perform in the standard 6V6 circuit? Or do you

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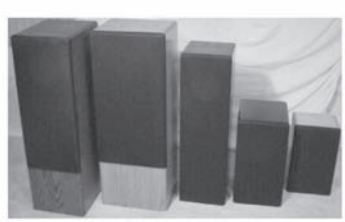
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look at the small part of the overall transfer function to emulate? Or do you look at the overall performance of a 300B run to its limits, and scale that performance down for a 6V6? I do not have the answers to these questions, but my experiments did lead me in the end to the latter solution as having the nicest sound qualities.

Where do you go when you have some serious number crunching? One of the main problems that I had in solving the 300B dilemma was trying to plot the difference between the transfer functions of many of the tubes available. Let me explain.

I needed a 300B transfer function and what I had was a 6V6. The difference between them is what I needed to apply to the 6V6 to get the result. Messing around with lots of graph paper in the early days solved the problem, but it also was time-consuming to plug in different tube curves.

The answer was to use a simple spreadsheet to just cut and paste different values to compare against and calculate the results. Better still, Excel allowed me to output to a graph so I could easily see what was going on. As a worked example, in **Fig.1** you can see the transfer function of a WE300b; **Fig.2** shows a 6V6; and when you crunch the numbers with Excel you get the result in **Fig.3**.

When I started to look at the transfer functions of the various tubes available, I found (when plotting the tube curves) that some tubes are more linear than others, with the pentodes and tetrodes shown triode-connected. In decreasing order of linearity, the results were tri-

FIGURE 2: Transfer functions of the 6V6 exported from Glassware's SE AmpCAD.

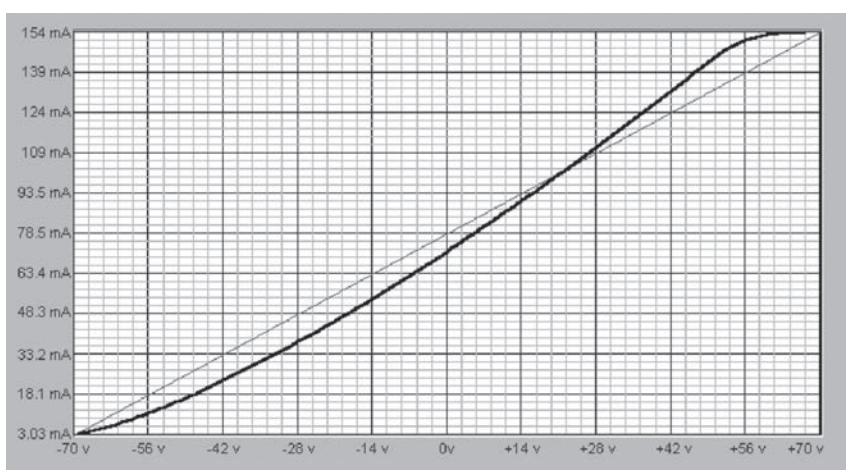

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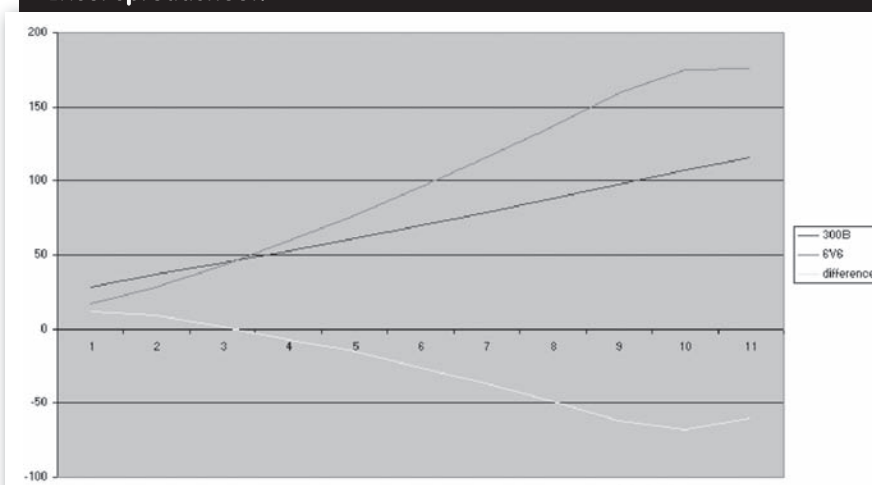
FIGURE 3: Difference between 300B and 6V6 transfer function from Excel spreadsheet.



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odes WE300B, CE300B, 2A3, and 811A, then the pentodes and tetrodes 6AQ5, 6V6, 12E1, 807, 6L6, KT66, KT88, EL34, and 6BQ5.

From the list, I found that of the pentodes and tetrodes available, both the 6AQ5 and the 6V6 were the best contenders in the linearity stakes to use for transfer function modification. The WE300B wins the linearity stakes in native mode with the 2A3 a very close second. Interestingly, the EL34 and 6BQ5 came in last.

MIXED MODE OF OPERATION

While operating in triode-connected mode did offer a substantial improvement in linearity over native pentode and tetrode mode, low efficiency was the drawback. The transfer function of a bipolar transistor is similar to a pentode tube, and as it transpires, by control of the voltage gain and the operating parameters, I was able to control the peak-to-peak grid voltage swing needed to drive the tube. Placing a resistor in series with the cathode bypass capacitor allowed the voltage gain to be varied by applying local negative feedback. By increasing the DC supply voltage and increasing the plate load, you change the transfer function and increase the grid drive requirements.

By adjusting the drive requirement, picking an appropriate DC supply voltage for a bipolar transistor, and setting the value of collector current (and hence stage voltage gain), I was able to adjust the overall linearity of a composite stage consisting of a bipolar transistor and a pentode or tetrode tube. This resulted in the pentode output stage operating at peak efficiency and still providing overall linear performance (by having an inverse nonlinear grid drive) with minimal feedback—all of this to emulate a triode valve. How close a match you produce will determine how close you get to the transfer function of a triode or to linear performance.

One of my experiments was to build a 2A3 amplifier using two 2A3s—one in the output stage, and one as the voltage amplifier—under the same operating conditions, and with an attenuator between the voltage amplifier plate, and the output stage grid, so that both tubes worked over the same operating range and conditions. The net result was exceptional but with even lower overall efficiency. The bigger problem was a lack of gain as these tubes have a relatively low mu.

To build a useful amplifier required four stages for the design. I ended up with two halves of a 12AU7 in front of the 2A3s as a gain stage. The overall performance was not as linear as using a single output stage and a matched characteristic voltage amplifier.

Tube Difference Spreadsheet.

	TUBE 1 300b	TUBE 2 6V6	
Grid V	Anode I	Anode I	Difference
-70	28.3	16.7	11.6
-56	37	27.9	9.1
-42	44.4	42.9	1.5
-28	52.3	59.4	-7.1
-14	61.7	77	-15.3
0	70.2	96.3	-26.1
14	79	116	-37
28	88.3	137	-48.7
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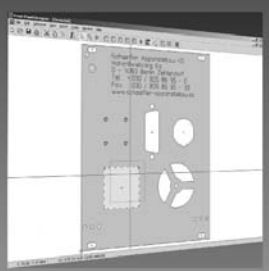

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WE300B LOOK-ALIKE AMPLIFIER

An example of my WE300B amplifier is shown in **Fig. 4**. To make it a little more nonlinear to simulate a 2A3 circuit, I just added a 100R resistor and a 100mF capacitor in series across R1 (**Fig. 5**). A variation on the theme is shown in **Fig. 6** where, instead of using a separate negative supply voltage, I used a NPN transistor and a cathode regulated supply to achieve the same result. (**Figure 7** shows an all-tube WE300B circuit).

The difference in the grid and cathode voltages results in the appropriate bias voltage. The downside to this variation is that the cathode voltage is twice that needed for a conventional self-biased output stage. This results in more wasted DC power and an overall lower complete circuit efficiency. The approximate cathode voltage to use is the HT voltage divided by half the mu of the tube.

FIGURE 4: WE300B emulator schematic.

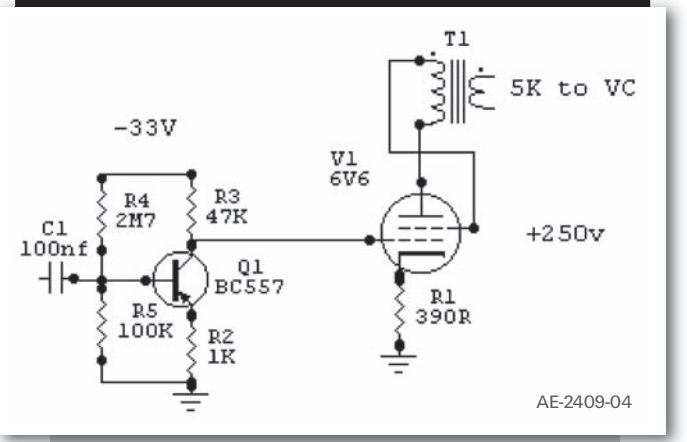


FIGURE 5: 2A3 emulator schematic.

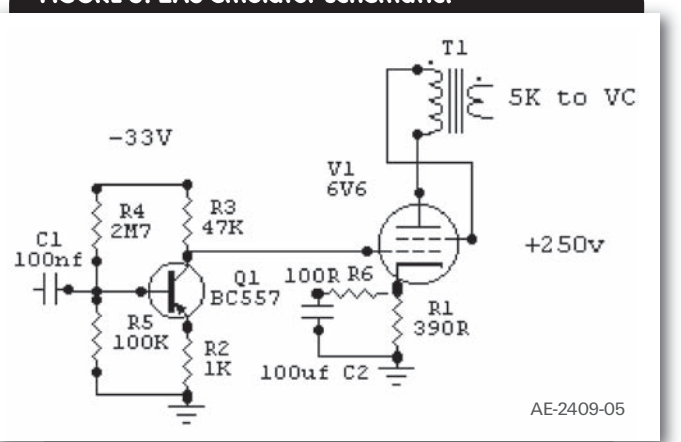
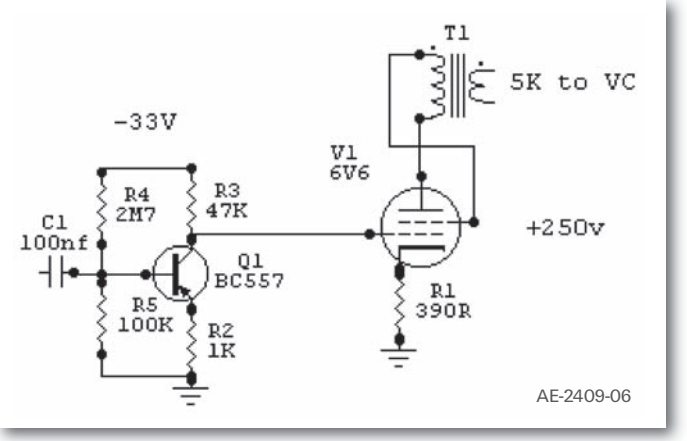


FIGURE 6: Variant of emulator schematic.



OPTIMIZING A TUBE DRIVER STAGE

Figure 8 shows the transfer curves of a 12AX7 tube. I noticed in practice that the theoretical curves apply quite well. Under low anode current conditions and high grid bias the tube transfer curve has quite a bend, whereas at higher anode currents and less grid bias the transfer curve is almost a straight line.

To optimize the voltage amplifier stage which drives the output tube, I needed to choose a part of the curve complementary to the curve of the output tube. You can work backward if you know the plate supply voltage, the load impedance, and quiescent current of the output tube. By looking at the tube characteristics and dividing the plate peak-to-peak voltage by the stage voltage gain, I ended up with the peak-to-peak grid driving voltage. With this symmetrically placed on the transfer

curve of the output tube and the correct grid bias offsets, I was able to draw the real transfer curve and transfer the information to my spreadsheet.

With a little juggling of the selected anode current and bias voltage on the grid of the 12AX7, I was able to override the peak-to-peak output voltage required by selecting a portion of the transfer curve similar in shape to that of the 6V6. In practice this reduced the THD of the amplifier, without feedback, from 11% to 4%—a large improvement. Not content with this, I eventually came up with a simple solution of having both a variable anode load and cathode bias resistor. These can simply be adjusted for the minimum distortion overall or to synthesize the sound of whatever tube combinations you wish.

With some juggling I was able to reduce the THD to a level of a little over 1% at full output and less than 1% at 1W. The downside to this method of operation is that you may run out of usable peak-to-peak output from the driver stage before you obtain the most linear overall results. The reason is that instead of optimizing the stage for maximum output, the Q point will now move away from the nominal half HT supply voltage on the anode, to a point somewhere above or below this figure to achieve the amount of nonlinearity required.

I have tried a few other audio tubes with great success, such as the 12AU7 and the 12AT7, which you may use for curve matching. For hard-to-match nonlinear tubes such as the EL34 and the 6BQ5, I have tried the 6ES8, which is pin-compatible with the 12Axx series tubes if they are wired for 6.3V filaments. The 6ES8 is an interesting tube to play with, because it is a semi-remote cutoff dual high mu triode normally used in TV tuners. The noise figures

FIGURE 7: All-tube emulator schematic.

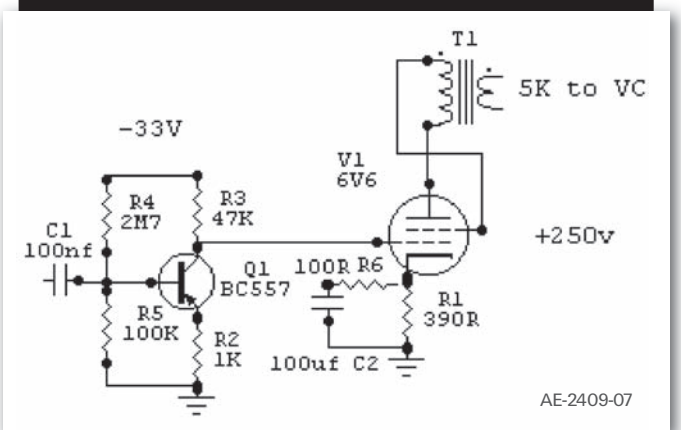
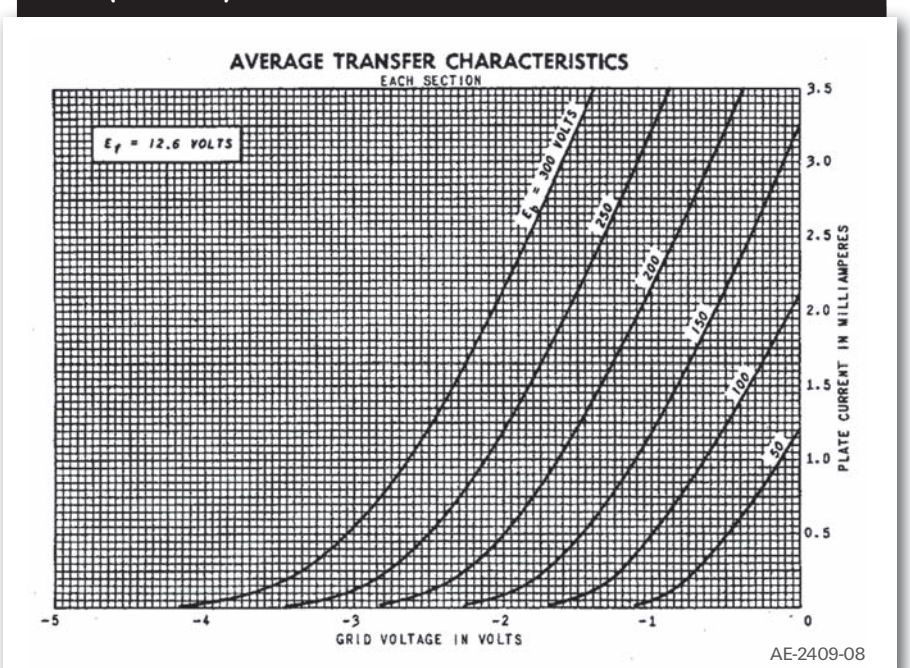


FIGURE 8: 12AX7 transfer function from the *STC Valve Data Handbook* (Volume 2).



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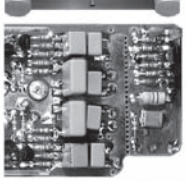

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and gain are quite high, but the tube has variable gain (slope) as well as a nonlinear (bendy) section of its transfer curve. This gives quite a bit of flexibility in this application.

As a side note, I have found that these tubes also make for an interesting-sounding overdriven cascade stage in a guitar amplifier. Amplifiers such as the Fender twin reverb produce a completely different character with the 6ES8 replacing the front end 12AX7. They sound particularly nice with a Stratocaster. I have also made the same transplant in a Marshall JCM800 with equal results.

LITTLE AMPLIFIER FROM DOWN UNDER

My all-tube prototype amplifier has now been passed around through the hands of many listeners here in my hometown of Adelaide, Australia, over the last two years. So far all comments have been complimentary, and several clones of this amplifier have been produced. Of the few people who have high-grade tube amplifiers to compare it with, only one has preferred the sound of his existing amplifier.

The 6V6 is a good low-cost tube. If you have a good junk box, you could build this amplifier in a day for little cost. Other tubes may also be used for higher output power. **Photos 1** and **2** show the prototype "little amplifier from down under."

Figure 9 shows the uncluttered front panel layout. One of my easy methods for making front panels is to simply use a laser printer to print onto a metalized art card and then laminate the panel. You can then cut it to size and spray the rear surface with craft adhesive. You can easily punch out holes for potentiometers and switches with a wad punch.

The result is a professional-looking front panel at nearly zero cost. The art card is available in a range of colors and finishes. My two favorites are aluminum and brass. The card does not cause any damage to the laser printer because the surface is non-conductive.

You can photocopy **Fig. 9** onto an art card instead of printing using a laser printer. Assembly is straightforward. There is nothing critical in the circuit, but be sure to heed the following comments about output transformers.

FIGURE 9: WE300B emulator front panel design.

AE-2409-09

OUTPUT TRANSFORMERS

As in any high-quality amplifier the final results are only as good as the quality of components used, especially transformers. My current experiments allowed me to try normal series-connected SE output transformers and also shunt-fed circuits. I found the latter to be problematic in achieving good phase distortion and low leakage induc-

FIGURE 10: 6AQ5/6A3 adapter.

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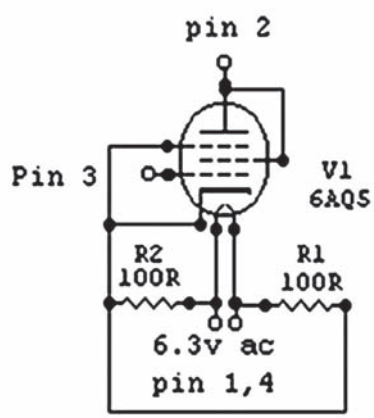

FIGURE 11: 6AQ5 pinouts.

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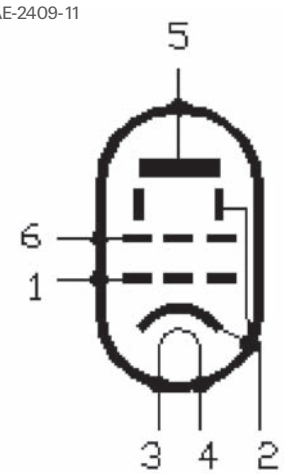

tance. The problem of using a series-fed SE output transformer is that the transformer will always be optimized at a single current by the choice of air gap at design and manufacture time.

During my experimentation I wound several output transformers—both E&I scrapless lamination types and also some "C" core designs. It was while experimenting with the latter that I replaced the usual soldered "C" core straps with some high-quality stainless-steel automotive radiator clamps, which allowed for easy tensioning of the cores and make for a variable gap output transformer. By tightening the bands you reduce the gap width, which decreases the standing current rating and also increases the primary inductance. Loosening it has the opposite effect. This allows for easy optimization with a change in operating conditions of the output stage.

MAKING A 6AQ5/6A3 ADAPTER

The 6A3 is a well-known 6.3V filament version of the classic 2A3 tube and a nice-sounding unit to use for developing and testing different circuits. When experimenting, it is always handy to have a good used set of tubes as testers until the circuit bugs are all worked out. It is better to destroy the tester rather than a brand new expensive NOS tube. 300b tubes and the 2A3/6A3 are in this category, but even testers can be expensive.

My simple solution to this problem was worked out about 40 years ago, but well-forgotten today. I often make small adapters to use available B9A and B7A base tubes, in place of less common 4-pin ones. This was often done in the '60s, making octal plugs to B9A bases for rectifiers. Making an adapter is a little more difficult for a directly heated cathode triode.

Figure 10 shows the schematic diagram of my 6A3 to 6AQ5 adapter. The 6AQ5 is wired triode-connected, which gives it approximately the same plate dissipation as a 6A3 (15W). The 6AQ5 has a plate dissipation of 12W and a screen dissipation of 2W, together 14W. The 6AQ5 pinouts are in **Fig. 11**.

To emulate a directly heated cathode is a little difficult. To solve this problem, wire two 100R ½W resistors from the 6AQ5 cathode to each of the filaments. The overall mu is a little higher than a 6A3, but the adapter is most workable, and the 6AQ5s are really easy to obtain at low cost. These are nice B7A base audio pentodes, electrically similar to a 6V6, very common, and available at a few dollars each. These are one of my useful sleeper tubes, such as the 6ES8, that everyone ignores and forgets about.

CONCLUSIONS

After much testing of classic and high-end tube amplifiers, I have come to the conclusion that those of two-stage design without any negative feedback tend to end up linear (to a degree) only by luck more so than by design. By applying the methods in this article, you may improve the performance of most two-stage amplifiers. By changing the operating conditions of the driver and/or the output stage, you can achieve better overall linearity. The simple modification to an existing amplifier is to replace the voltage amplifier fixed resistors with variable anode and cathode resistors. *aX*

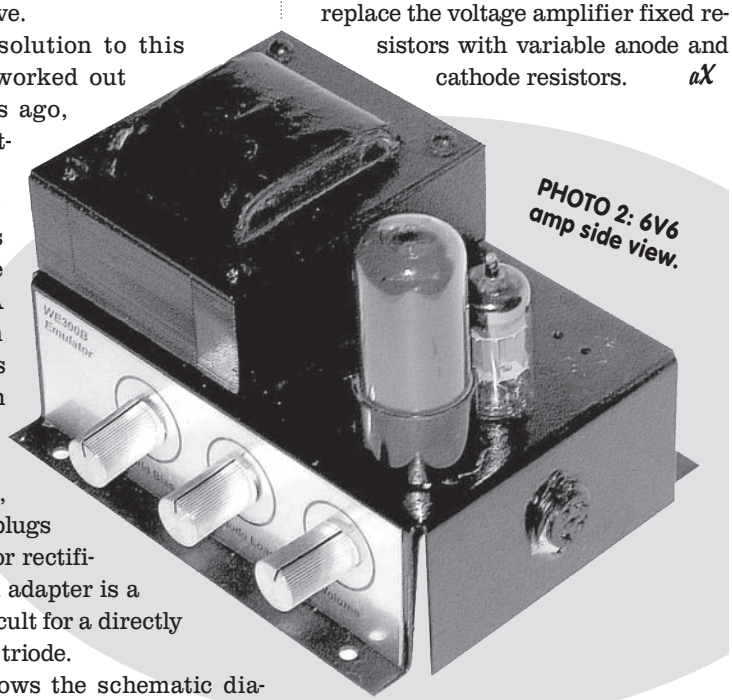

PHOTO 2: 6V6 amp side view.

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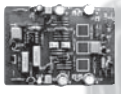

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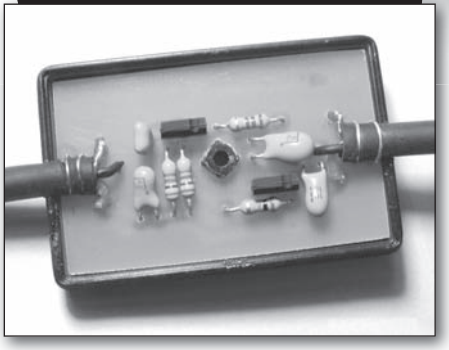
This mike amp design expands computer and speech recognition capabilities.

By Paul Loewenstein

This microphone amplifier is designed to interface the Shure SM10A dynamic microphone to a SoundBlaster card or other computer microphone input with a 5V through 2.2kΩ bias to the microphone connector ring. It enables speech recognition veterans of DOS DragonDictate to continue using the DragonDictate microphone with Dragon NaturallySpeaking.

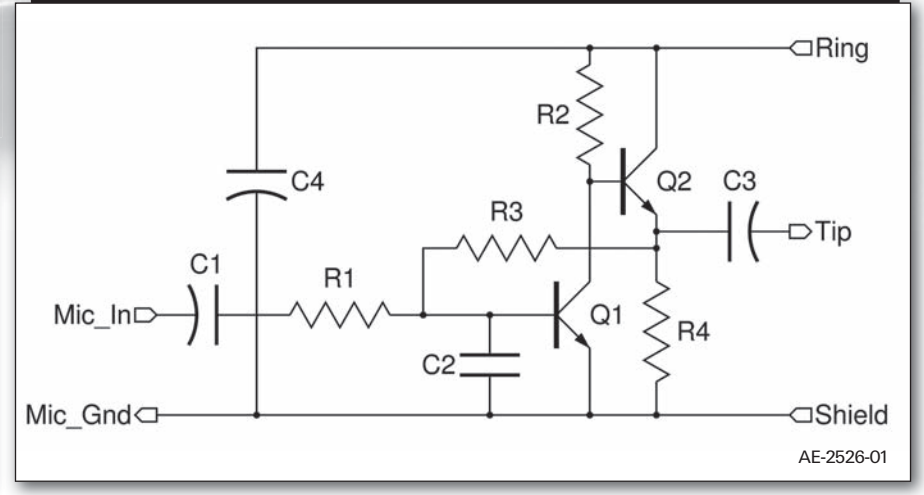
PHOTO 1: Amplifier in Radio Shack box.



PRINCIPLE OF OPERATION

Referring to the schematic in **Fig. 1**, the open circuit voltage gain is provided by the grounded emitter stage formed by Q1 and R2. This is followed by a unity-voltage-gain emitter-follower stage formed by Q2 and R4. R3 provides negative feedback to better define the gain,

FIGURE 1: Amplifier schematic.



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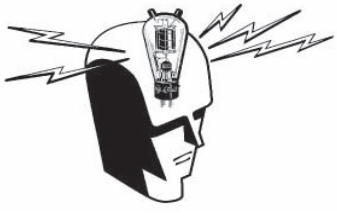


improve linearity, provide DC bias, and reduce the output impedance.

C2 rolls off the open-loop gain above audio frequencies to prevent oscillation. It also attenuates any radio frequencies picked up (see "Effects of Nonlinearities" section).

C1 and C3 provide DC isolation at the input and output, respectively. C4 reduces power rail excursions.

DC BIAS

Q1 base is one diode drop (about 0.6V) above ground, drawing current from Q2 emitter through R3, across which there is a negligible DC voltage. Q2 base is therefore about 1.2V above ground. Taking into account the 2.2kΩ in the soundcard between the ring and the 5V supply, the ring settles at about 3.5V.

OPEN-LOOP GAIN

The DC voltage across R2 is 3.5 – 1.2 = 1.3, so at room temperature the voltage gain of the grounded emitter stage is about 40 × 1.3 = 52.

CLOSED-LOOP GAIN

The model for calculating the closed-loop gain is shown in **Fig. 2**. It comprises:

- An ideal amplifier of gain –A representing the gain of the grounded-emitter stage with A = 52.
- Resistance *x* representing R1 in series with the microphone impedance; the impedance of an SM10A is 223Ω, so x = 323Ω.
- Resistance *z* represents the input impedance of the grounded-emitter stage. This is ß × the emitter impedance, which is $\frac{R2}{A}$. The lower bound on ß is 270, giving a lower bound on z of 29kΩ.
- Resistance *y* represents R3 (10kΩ).

Analysis of this circuit gives the closed-loop gain as:

$$\frac{-Ayz}{xy + yz + (1 + A)zx}$$

Substituting in for *A, x, y,* and *z* gives a (negative) closed-loop gain of 19.

NONLINEARITIES

The emitter impedance of a bipolar transistor is inversely proportional to the emitter current. This results in two sources of nonlinear gain:

1. The gain of the grounded-emitter stage.
2. The output impedance of the emitter follower, which results in nonlinear gain depending on how the output is loaded.

THE GROUNDED-EMITTER STAGE

The gain of a common emitter stage is proportional to the emitter current and therefore proportional to the voltage drop across R2. The gain when the output signal is high is less than when the

FIGURE 2: Closed-loop gain model.

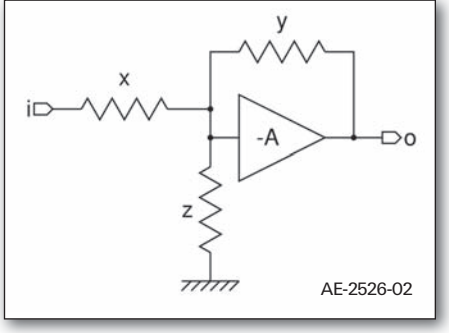
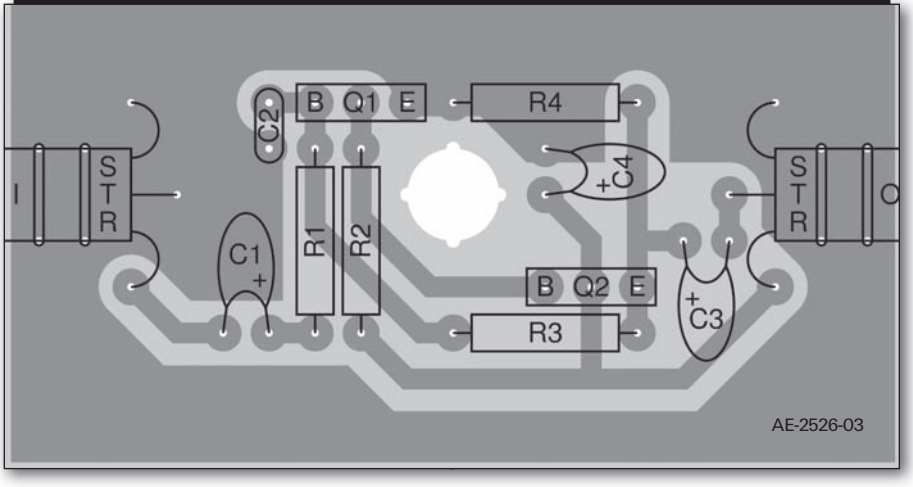


FIGURE 3: Printed circuit layout.



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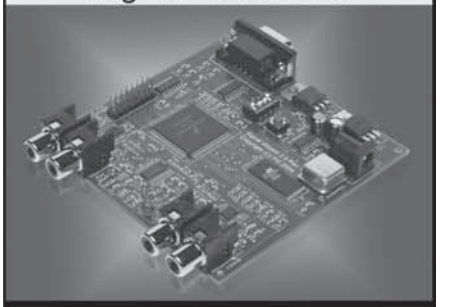

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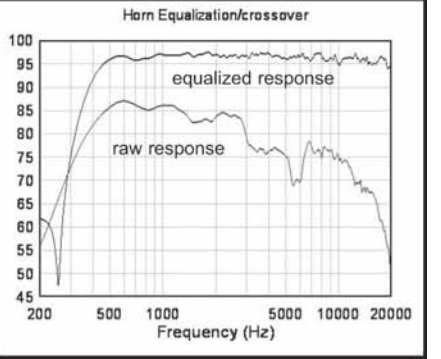

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output signal is low. Assuming a 1mV signal from the microphone, the signal is about 20mV at Q1 collector, which means that the open-loop gain can deviate by almost 2%. However, the negative feedback limits the deviation of the closed-loop gain to about 0.7%.

THE EMITTER-FOLLOWER STAGE

The output impedance of the emitter-follower stage is inversely proportional to Q2 emitter current. With a 20mV signal, this can vary by about 6% around about 200Ω. However, the soundcard microphone input impedance is about 2.2kΩ, so the effect on open-loop gain is only about 0.3%. The effect on closed-loop gain is only about 0.1%. In addition, the nonlinearity is in the opposite sense to that of the grounded emitter stage, so it subtracts from the effect due to the grounded emitter stage.

Capacitively loading the output can introduce more nonlinear distortion. However, a significant effect would require several nanofarads of capacitance, much more than any likely parasitic.

EFFECT OF NONLINEARITIES

For variation of gain as a function of signal level introduces two components to the signal:

1. Frequencies double that of the original (0.6% deviation of gain introduces 0.3% second harmonic distortion). Given the high harmonic content of human speech, this is unlikely to adversely affect speech recognition.

2. A DC offset proportional to the signal level. At first sight this may seem inconsequential, but it turns the amplifier into a radio receiver un-

Parts List

C1	Tantalum capacitor	10µF 16V
C2	Ceramic capacitor	47nF 50V
C3	Tantalum capacitor	10µF 16V
C4	Tantalum capacitor	10µF 16V
R1	Carbon film resistor	100Ω 0.25W
R2	Carbon film resistor	5.6kΩ 0.25W
R3	Carbon film resistor	10kΩ 0.25W
R4	Carbon film resistor	2.2kΩ 0.25W
Q1	Silicon NPN transistor	NTE16
Q2	Silicon NPN transistor	NTE16

less steps are taken to attenuate radio frequencies[1].

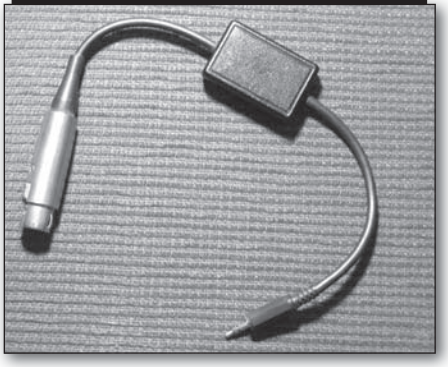
MECHANICAL DESIGN

The printed circuit design is shown in **Fig. 3**.

To house the amplifier, you can use RadioShack's Mini Project Enclosure, part number 270–288, which has a convenient stud in the middle, providing a press-fit mount for a 0.1″ pitch perforated board with one hole drilled out to 5mm.

Holes for the cables are drilled in each end with the box clamped tightly closed. The board is then inserted, cables attached, into the open box (**Photo 1**). The box is then closed and secured with a single self-tapping screw. The complete assembly with cables and connectors is shown in **Photo 2**. aX

PHOTO 2: Completed amplifier with cables.



Paul Loewenstein *received a BA in 1971 from the University of Cambridge, an MA in 1974, and a Ph.D in 1994. His early career at Schlumberger in the UK and France involved analog systems from high-frequency IC amplifiers, vibrations of thin shells, and transients on high-voltage power lines. His professional activity has been purely digital since his move to the US in 1981, working for Schlumberger, National Semiconductor, Mitsubishi, and Sun Microsystems.*

REFERENCE

1. The first version of this amplifier rolled off the loop gain using a capacitor to the collector of Q1. When conditions were right it happily picked up KGO at 810kHz AM, causing great confusion to the speech recognition.

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I wish to record and play monophonic LPs and 45s (vertical cut). I am aware that you can physically install a switch on the turntable plinth or wherever possible. But, I am not sure how to electrically wire the switch to play vertical cut discs, so would appreciate the schematic of the mono/stereo switch.

P.S. My phono/preamp does not have a mono/stereo switch, and I understand the need for employing the switch at the turntable. Page 77 in the book *The LP Is Back* shows a print of a lateral/vertical switch. Is this one correct?

Lisa Marie Jastrzemski,
Eastpointe, Mich.

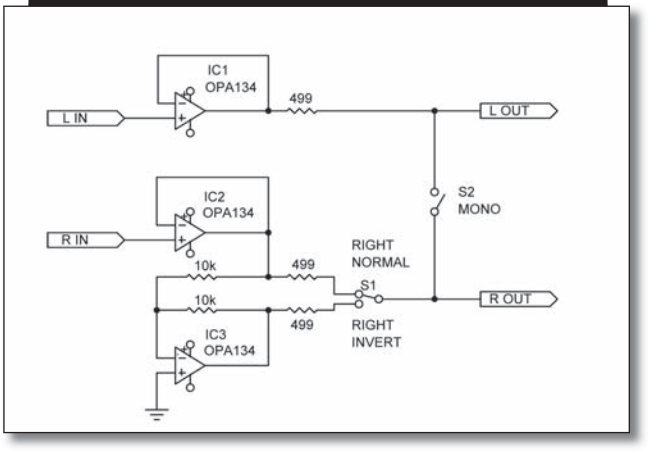
Gary Galo responds:

Your question is a little confusing, because monophonic LPs and 45-rpm records are always laterally-cut. I assume that you want to be able to combine the two channels to mono, and be able to switch between laterally-cut and vertically-cut records (vertically-cut recordings include all cylinders, Edison Diamond Discs, and most Pathé discs). The illustration on page 77 of *The LP Is Back* does show correct switching for laterally- and vertically-cut records at the phono cartridge input ("A Preamp for Vintage 78s," a modification of the McIntosh C-8 preamp originally published in *Audio Amateur*, 1/1985).

It is important that you mount this switch right next to the phono input jacks on the back of your preamp, minimizing lead lengths which can cause hum pickup at phono cartridge signal levels. You will also need a tonearm cable with a floating shield, as described elsewhere in that article.

If you are using a stereo preamp (or a pair of modified C-8s), I recommend doing the lateral/vertical switching at line level. **Figure 1** shows a circuit that will accomplish this. IC1 and IC2 are configured as

FIGURE 1: Line-level lateral/vertical/mono switching for 78-RPM recordings and cylinders.



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unity-gain voltage followers for the left and right channels. IC3 is a unity-gain inverter for the right channel only. S2 is the mono switch, and S1 is the lateral/vertical switch.

In the "Right Normal" position, the two channels are summed to mono "in-phase" for laterally-cut recordings. The vertical noise component is cancelled. In the "Right Invert" position, the two channels are summed to mono "out-of-phase" for vertically-cut recordings. In this case, the lateral noise component is cancelled.

Use high-quality, low-offset FET-input op amps for the ICs 1-3, such as the Burr-Brown/TI OPA134 or OPA604, or Analog Devices' AD825. Supply rails can be ±15V, and be sure to include local supply bypassing close to the op amps. You can mount this circuit either inside an existing preamp or in an external enclosure. In the latter case, you could connect it through a tape-monitor loop in your preamp. You could probably use the preamp's own mono switch, assuming that it is located after the tape monitor loop (this is usually the case).

Bear in mind that a normal RIAA stereo phono preamp will not provide correct equalization for older recordings.

TEST EQUIPMENT

After reading the *Audio Electronics* 5/00 article about the Pico scope ADC-216, I decided to purchase the high-resolution 12-bit ADC 212/3 because at +$400 it was about half the price of the reviewed $800 16-bit version. I now have a situation in which I am setting up a test bench at a new home in the Portland, Oreg. area, so I would like to duplicate or improve upon the existing equipment at my test bench in northern California. I primarily use the ADC 212/3 as a spectrum analyzer to check for distortion, hum, and noise for the SE and PP tube amplifiers at 1W, 2W, and so on.

The only undesirable attributes of the Pico scope are the inability to set the 1kHz test signal at 0dB, fast response times versus resolutiontrade-off, and the 20V input limit. I saw the new 3000 series from Pico, but this series emphasizes the DSO (only 8 bit) rather than the spectrum analyzer, which is my primary use of this product.

Do you know of any new products by other manufacturers that now perform the spectrum analyzer functions of the Pico 212/3 better or for less money?

Nsmurro@aol.com

Charles Hansen responds:

I'm not sure why you could not set a 1kHz test signal at 0dB. The 0dB level is referenced to the selected input range. If the input signal exceeds the 20V input limit of the ADC212/3, you can use an input attenuator network to divide down the input voltage. Response time versus resolution is the trade-off you must always make with FFT measurements, no matter whose equipment you use. I prefer a longer response time in order to achieve the best accuracy and resolution. Pico has an excellent application knowledge base at their website, and I have found their online support to be very helpful as well.

There are a number of PC-based test equipment suppliers. Some use interface units such as the Pico scope, some use PC sound cards. A quick web search turned up ATSpec Pro, Audio Precision, BitScope, National Instruments, Prism Audio, MultiScope, ProTek, RMAA Rightmark, PHS Spectra-Plus, Sound Technology SpectraSeries and Velleman. I leave it to you to determine your own cost-performance trade-off. *aX*

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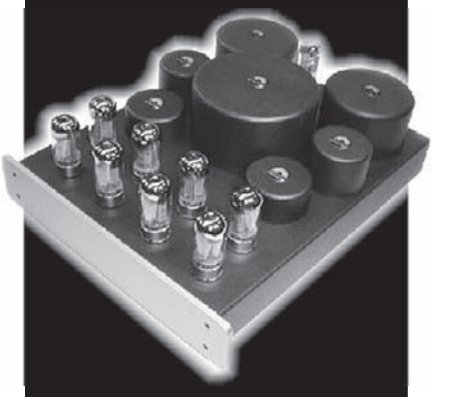

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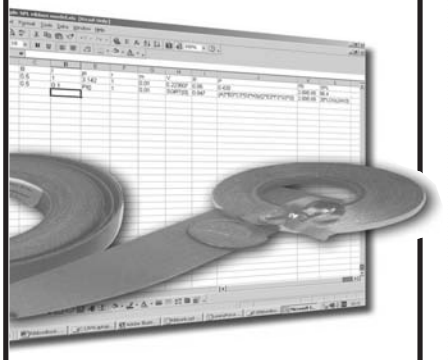

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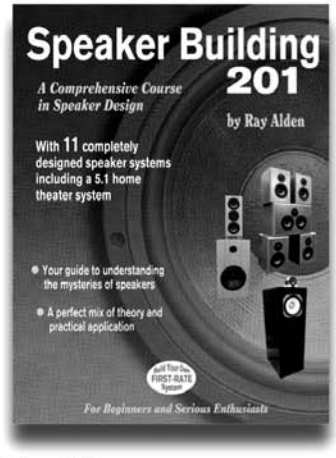

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CORRECTION

I just noticed that in my article, "Isolator Tube Interface to Computers," appearing in the July issue of *audioXpress*, the first four sentences were left out. These sentences contained the name of the software and the company that provides the software, plus the website that it can be downloaded from, so it might be confusing to the reader not to have this information. The website address is: **http://www.nch.com.au/tonegen/index.html**, which will take you directly to the downloading site. The main company web address is: **http://www.nch.com.au/software/index.html.**

Frank Simonsen
fsimonsen@telus.net

TRANNY TALK

The timing of Larry Lisle's article on the 71A project ("A Minimalist Amplifier Design," March '05 *aX*, p. 16) couldn't have been better! I have all the components to build except the output transformers.

I have a Stancor A3311 (P.P 10,000Ω plates to 500/15/8/4Ω primary DC 70mA max audio power 25W). I know the 8054 is a high-fidelity model. How limited will my results be with these A3311s, whose application may be only for a more limited bandwidth without the extension at both ends of the frequency response that you must be enjoying? I have 110dB efficient horns. Specifically,

1. What modern equivalent for the 8054 would you recommend?
2. What would you recommend as a SE replacement and how would the audio circuit you show change?
3. Do you know where I might find a pair of 8054s?
4. Do you have a recommendation in NOS (50s vintage) to the 8054?

Thank you for your continuing series of articles over many years of service to the audio industry.

Paul
phat@mountaincable.net

Larry Lisle responds:

Thank you for your letter and kind remarks. Your A3311 was a very successful transformer for Stancor. It was listed in their catalogs from at least 1939 to 1959 for use in radios with push-pull 6F6s or similar output tubes. Therefore, you can count on it to reproduce the audio spectrum of a good-quality AM radio. It's certainly worth a try.

As you pointed out, though, it isn't a high-fidelity transformer. For comparison, the following prices are from a 1957 Allied Radio parts catalog: Stancor A3311—$5.94, Stancor A8054—$13.36, UTC LS-52—$21.56, Acrosound TO300—$24.75, Altec-Lansing Peerless S238—$25.86. These push-pull transformers are approximately equal in power handling ability, and I think the price is a rough guide to their quality.

Whether you can actually perceive the differences among them in your listening environment with real music is something that can only be determined by experiment. I don't believe the small amount of unbalanced direct current in the primary drawn by a single-ended 71A would be enough to seriously degrade the performance of these large transformers, but again, there's no way to be sure without actually trying it.

Where to find these and other transformers from the golden age is a good question. They sometimes turn up at ham or antique radio flea markets, but I've had better luck using the Internet. In any case, be prepared to pay a much higher price than in 1957!

There are many modern transformers, both single-ended and push-pull, that are very good, and I hesitate to flat-out recommend a particular brand. You might check previous issues of *audioXpress* and *Glass Audio* for types that have been used in vari-

ous projects and then send for a catalog or check specifications on the Internet.

Thank you again for your interest.

REPLACING BATTERIES

I read Charles Hansen's article on the Fluke 8050A DVM (April '05, p. 36) with great interest, especially about the 4.5 digit wide-bandwidth RMS accuracy, the option 01 battery feature, direct dB measurement, and the relative setting which allows you to store a reference measurement to compare against future readings. I acquired one from eBay for $56 and I am very happy with it. The meter's LCD is in good shape (replacing it costs $36 + s/h from Fluke!). My meter displays the BT warning when on batteries, although one of the four cells is borderline.

I started researching battery replacement per your article, which brings me to the point of this letter. I found out from the Internet that you should not replace NiCad with NiMH cells unless the charger is specifically designed to do so. This is per a very good battery company (Zbattery) website **http://www.zbattery.com**, which has an excellent FAQ section on this. Their model #NCD3000T is a 3000mAh NiCad C-cell with solder tabs at $3 each, and is my candidate to use for this project. This obviates the need to install battery holders; you can re-use the existing plastic holders that housed the original cells.

This brings up the question of what it would take to modify the charger circuit, because most of us would prefer to use NiMH cells for their ability to store more power per weight and their lack of memory effect problems. Also, note that the manual containing a schematic and parts list plus an errata update is available as a free download from the Fluke manual support page as a PDF file.

Philip Cooper
picooper@optonline.net

Charles Hansen responds:

I'm glad that Mr. Cooper is happy with his Fluke 8050A DMM. His concern over the charging requirements for NiMH batteries is well founded in the case of fast charge times, such as the chargers used for power tool batteries where rapid cycling is needed to maintain productivity on the job. I have two identical Makita cordless drills. The older one came with NiCad batteries, and

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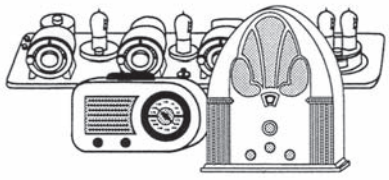

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I won't charge NiMH batteries with it because the charge current is too high and the charger does not have the necessary "smarts" to prevent damage. The newer charger will handle both battery types, although a NiCad will take a bit longer to charge.

With regard to low (trickle) charging rates, however, you will not damage a NiMH battery using a charge circuit originally designed for NiCad batteries. As long as the charge current is 10% of the cell AH rating, either type of battery can remain on charge indefinitely without damage. The 8050A manual (Table 1-4) states that the full charge time for a new uncharged battery is 14 hours. We have been using NiMH batteries in four Fluke 8050A and 8010 DMMs at work for two years without any problems. I don't think it is necessary to modify the Fluke charging circuitry.

If you do not regularly use the DMM, NiCad batteries might be a better choice because of their lower self-discharge rate. Otherwise, the higher energy density characteristics of NiMH will deliver more continuous operating hours, assuming the same physical battery size.

Mr. Cooper's information on the solder-tabbed Zbattery is useful for those not wishing to modify their Fluke with new holders.

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E-LATED

Pete Millett really hit a home run with his E-Linear single-ended pentode amp (April '05, p. 6). It's amazing the kind of performance you can get with a few parts and some thoughtful engineering. I'm building this one, for sure!

I do have a few suggestions, as well as a concern that I'd like to share:

1. You can substitute the 7-pin miniature 0A2 for the octal 0D3 regulator. This little guy has better regulation but a lower current rating than its big brother, so you'll need to increase the value of R1 to about 9100Ω. Whatever you do, don't change the value of C2—you'll ruin the tube!

You can get the basing info from a neat little program called "Tube Data Sheet Locater (TDSL-PE)," downloadable for free from **http://www.duncanamps.com/tdslpe/download.html.**

2. I agree with Pete that ordinary capacitor clamps are pretty ugly, but you don't need to machine fancy ones. Simply turn the clamps upside-down and mount them beneath the chassis. Just be sure to lay out the clamps from the underside, or the holes may not end up where you want them.
3. The ASC 40µF polypropylene motor-start capacitors Pete recommends are economical and really sound great, but higher values do not sound as good. This is definitely a case of bigger not being better.
4. I'm concerned about the heater-to-cathode voltage applied to the rectifiers. Design-maximum rating for these tubes is no more than 100V continuous, or 300V peak.

Although these tubes are on their own separate heater winding, the winding is grounded. This places the entire 425V supply across the heater-cathode insulation.

It is far better to just let the heater winding float. Heater-cathode leakage will then develop a positive bias on the winding, which is actually desirable.

CAUTION—it is potentially hazardous to power any other tube from the rectifier heater winding!

5. It used to be standard practice with pentodes to put a small capacitor across the plate winding of the out-

put transformer. This maintained transformer loading at the higher frequencies, where the speaker had given up.

I believe that such a capacitor may reduce the slight ringing noticeable in the square waves, and it might allow compensation capacitor C9 to be reduced or eliminated.

Just a small capacitor would be needed—I guesstimate about 1nF (1000pF)—but the voltage rating should be around three times plate voltage. Since this capacitor handles only ultrasonic frequencies, silver mica or ceramic will work fine, and are readily available at these voltages.

Bob McIntyre
musitronix@yahoo.com

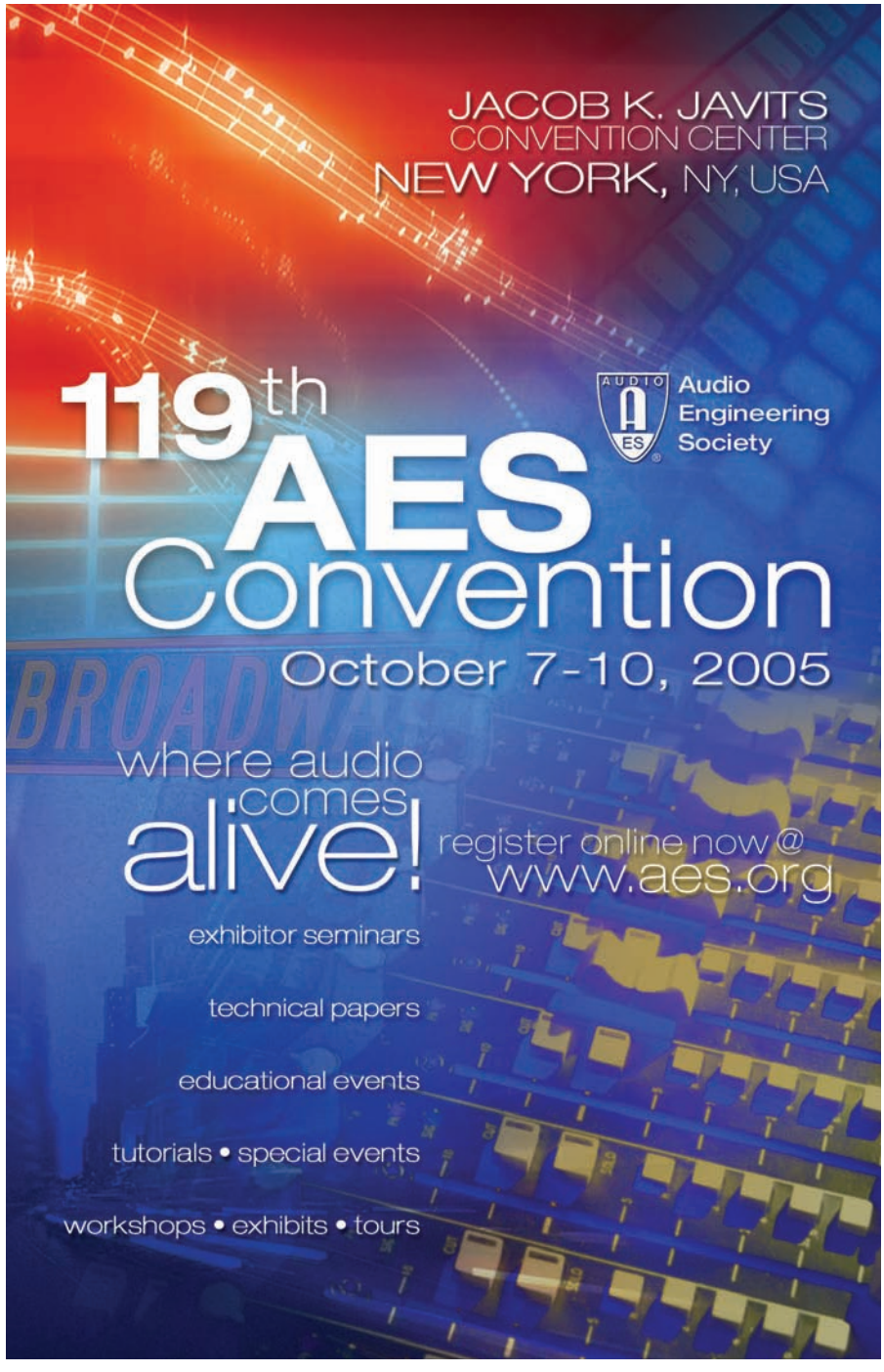
Pete Millett responds:

Thanks for the comments.

1. The 0A2 can be used in place of the 0D3 if desired. I like the 0D3/VR150 largely for its looks!
2. You can use regular capacitor clamps inside the chassis to hold the large can capacitors, as Bob suggests. It does require quite a bit of space under the chassis, though. With the clamps I have, the terminals of the caps would wind up 2″ below the chassis, so you'd need about a 2-½″ deep chassis to provide some clearance. One benefit of mounting them the way I did is that you can use a pretty shallow chassis.
3. I haven't played with larger polypropylene-in-oil motor run caps. The parasitics (inductance and so on) do increase with size, so it's entirely possible the larger ones wouldn't sound as good.
4. I think Bob picked the wrong rating off the data sheet for the rectifier tubes. The heater-cathode rating of the 6CJ3 and 6CL3 is only 100V if the heater is *positive* with respect to the cathode. The rating for having the heater *negative* with respect to the cathode is 900V continuous, 5500V peak—and with the heater grounded and cathode at B+, the heater is negative with respect to the cathode.

In television damper diode service, which is what these tubes were designed for, the heater is normally run off the same (grounded) filament supply as other tubes in the set, and the cathode is at high voltage (typically the same DC voltage as the plate of the horizontal output tube, 200–800V), so it is perfectly safe to use them this way. If desired, assuming you have a separate filament winding, you could connect it to the cathode; but I'd be as worried about the isolation of the filament transformer as the heater-cathode insulation of a damper diode tube. Floating the winding will work OK as well—I've done it—but in general I don't like to have floating heater windings.

5. I didn't try a capacitor across the output transformer. It's possible it might allow the capacitor on the driver plate to be eliminated, but I suspect it is still desirable to have the capacitor on the driver. The driver pentode has such high Gm that any parasitic inductance in the plate circuit causes an increase in gain, possibly at hundreds of megahertz, which is undesirable in and of itself. The ringing in the square wave output is pretty inconsequential—the amplitude is very small, and the square wave response looks much better than most amplifiers that I have looked at. You can increase the value of the capacitor on the driver plate if desired to eliminate it, but I chose a capacitor value that provided close to critical damping. Sweeping a sine wave up to



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over 100kHz didn't show any significant peaks at high frequency, another indication that it is adequately damped.

PERSONAL USE ONLY

I just scanned the latest edition (April '05) of *audioXpress* that I received in the mail today. At the end of Erno Borbely's article (p. 25), there's a box containing a statement regarding intellectual property. I'm sorry to hear that this sort of thing is an ongoing problem that has a direct effect on Mr. Borbely.

Thank you for so generously sharing your knowledge over the years with the DIY audio community. I have learned quite a bit recently by reading (and rereading) your series on designing with JFETs from a few years ago. Given the problem with people using your designs for commercial purposes, it's truly wonderful that you remain willing to provide this information to us amateurs. Obviously something motivates you beyond the simple desire to make money and a living, and I hope that you get a few of these from time to time that encourage you to continue to contribute.

It's great, and we like it!

Vince Harris
Philadelphia, Pa.

CABLE CONNECTIONS

I am perplexed by Richard Honeycutt's response to John Volanski's book review critique (April '05, Xpress Mail, p. 48). Without consulting many references on signal cable construction, but relying on memory, I don't understand Mr. Honeycutt's comment on "one way" cables. Of course, he recognizes the reason for marking them, but, I think, confuses the two types of signal cables and the way the shield is used.

In balanced cable, there are two conductors and a shield. Neither conductor is (or should be) connected to ground at either end. The shield is the only wire connected to ground, and can be connected at both ends without concern for ground loops.

In unbalanced lines, there are two conductors and a shield, but one of the conductors is connected to ground at both ends. If the shield is connected to ground at both ends, a ground loop will exist. This may or may not cause noise, hum, and/or EMI interference depending on the design and circumstances, so it has become common to leave the shield unconnected at

one end. I have made up all of my signal cables from Mogami cable in this configuration and have zero noise, hum, and so on. Contrary to Mr. Honeycutt's remark, I connect the grounded end of the shield to the output jack of the source component, not the destination end.

This configuration has been recommended by many designers over the years, and it has worked for me for more than ten years.

Frank S. Thomas III
Ballwin, Mo.

Richard Honeycutt responds:

Thank you for your comment. Your letter leaves me with the impression that you believe that there is a difference between the "low" side of the signal and "ground" in unbalanced equipment. This is very rarely the case. Therefore, whether the shield is connected at both ends makes no difference, because the "low" side of the signal pair already connects the grounds in the "source" and "destination" equipment.

Your letter also assumes that unbalanced cables use two-conductor shielded cabling. Normally this is not the case either: single-conductor shielded cable is most common. Your solution certainly would help if you were connecting unbalanced equipment with isolated grounds.

Thank you for identifying my error in recommending grounding at the destination end: you are correct in that for balanced circuits and isolated unbalanced circuits, the shield should be grounded at the source end. Further information on grounding can be found in Davis's *Sound System Engineering,* Ballou's *Handbook for Sound Engineers* (available from Old Colony Sound Lab), and in several books by Ralph Morrison.

PCB PICTURES

I read with great interest Erno Borbely's two JFET Preamp articles (June and July '05 *aX*). Unless I missed it, though, neither installment contains pictures of the PCB traces. This makes them little more than a two-installment advertisement in my opinion. Yes, the schematics are there, complete with difficult-to-find-except-from-Borbely devices, but I'd be far more likely to patronize your advertisers for this one if you could make my life a little easier, like providing PC artwork. Most of your other authors seem capable of this task.

Thanks for letting me vent.

Steve
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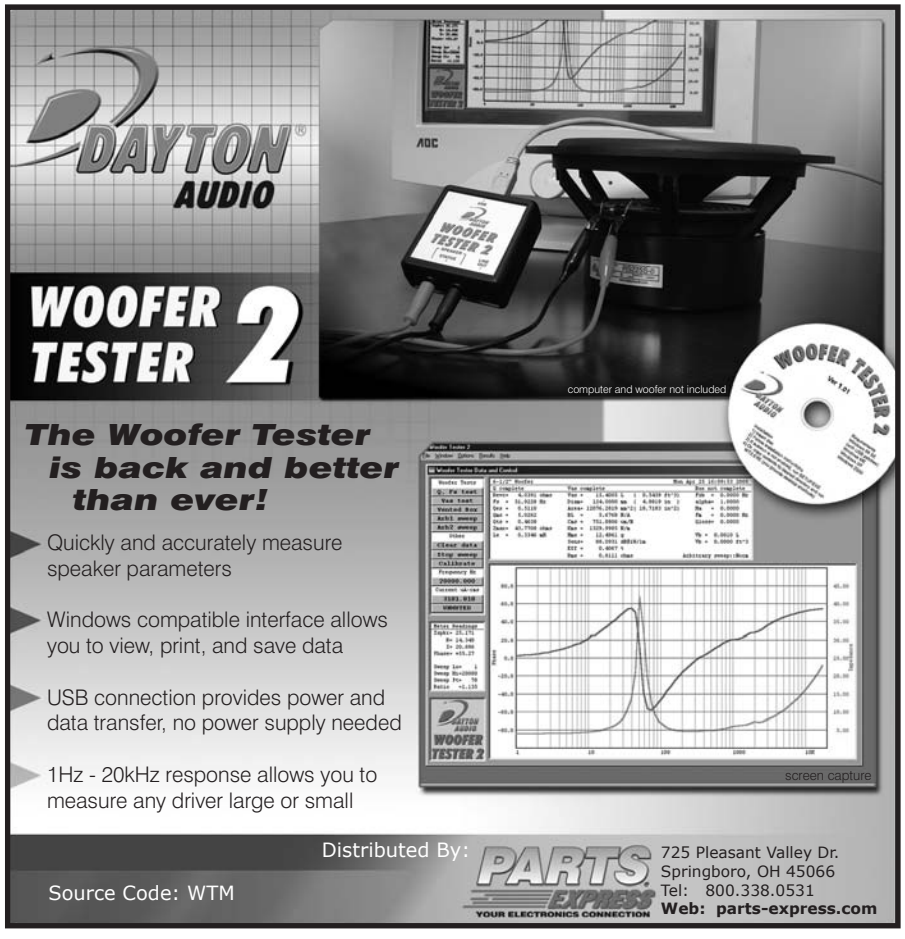


Erno Borbely responds:

There is no question of being "capable of this task," but reproducing PC artwork 1:1 in the magazine is not easy. I send my manuscript in Word and/or PDF format, both of which change the size of my originals. Limited space in the mag might also dictate size reduction or even omission.

The original size of the boards is: 105 × 145 and 210 × 145. You can find .pdf versions of the line amp and phono preamp on the *aX* website **www.audioXpress.com**. For hard copies, please send a self-addressed stamped envelope to audioXpress, PO Box 876, Peterborough, NH 03458. Good luck with your JFET projects.

CAP TYPES

As Paul Stamler points out in "Capacitor Bypass: Proceed Carefully" (*aX* 5/05), bypassing can be counter-productive if not done with a perception into the interaction between capacitor types. Analog Devices has some very good application notes on grounding and bypassing on their website **(www.analog.com)**, as do many of the capacitor manufacturers.

I'd like to add to his comments under "Raw Results" on page 60. Comments 1 and 2 concern the expected relationship between ESR and ESL in aluminum filter caps. Low ESR requires a large plate area that minimizes the electrode resistance, and closely spaced electrodes.

Interestingly, the tall, smaller diameter caps tend to have lower ESR than short, larger diameter caps of the same capacitance and voltage ratings. This may be because the taller caps have foil lengths and widths that are closer to square, which maximizes the foil area. Etching the foil further increases the available surface area. Capacitors with thinner oxide layers and separators produce higher capacitance and lower ESR, but at the cost of reduced voltage ratings.

You might think the close plate spacing that produces low ESR would also result in lower ESL, because the loop area available to produce a magnetic field is reduced. This would be true if the aluminum capacitor foils were laid flat, but they are instead wound into a cylinder (this is also true of wound film-foil and metallized-film caps, comment 6). Think of them as being similar to

air-core inductors, so their inductance is essentially independent of frequency. The terminal leads also add a small amount of additional inductance.

Tan theta (comment 3) is the ratio of capacitive reactance to ESR, and is a frequency-dependent term. For low loss capacitors, tan theta is essentially equal to dissipation factor (DF).

I would think the higher number of turns in short, larger diameter capacitors of the same voltage and capacitance as tall, small diameter caps should increase the ESL, not decrease it (also comment 3).

The multi-layer parallel-plate ceramic disc (comment 7) is indeed the lowest ESL capacitor type, if you can put up with the variations in capacitance with temperature, age, applied voltage (AC and DC), and frequency. The high-K X5U types might vary as much as –55% with all these combined effects.

While solid tantalum capacitors are generally superior to aluminums in terms of ESL, they are not as good as the disc ceramics. They also have inrush current limitations and thus cannot be used as the main reservoir filter capacitor. A combination of aluminum, tantalum, and ceramic types may give the best results.

Charles Hansen
Ocean, N.J.

TUBE SAFETY

Kudos to Larry Lisle for his excellent article "Reducing Filamentary Triode Amp Distortion" (*aX* May '05, p. 35). This is truly groundbreaking work. But I am curious about the stated precautions on too-low filament voltages: What harm could this cause?

We know for incandescent light bulb filaments that life is inversely proportional to a high exponent of the operating voltage, such that a small reduction in voltage causes a large life extension (at the loss of much light emission, alas). It would seem that in the triode, lower filament voltages should result in less material loss from the filament with any given plate voltage. Is there some other effect going on here?

Also, with regard to reducing the space charge, is there any reason this technique could not be employed with separate-cathode tubes? Or does the high electron emissivity of the cathode coating make this concept too voltage-sensitive to be practical?

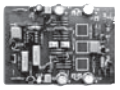
Many thanks for a fine and thought-provoking article!

Bob Masta
audio@daqarta.com

Larry Lisle responds:

Thank you for your comments. As to your questions—in writing for publication I think it's best to err on the side of caution. While I haven't experienced any apparent shortening in tube life by reducing filament voltage to values above the critical point where the space charge disappears, I don't have enough evidence to make a categorical statement that there couldn't be any danger with any filamentary tube under any conditions. I'd hate to be responsible for a reader blowing away some expensive tubes without being aware that there may be risks.

To your second question, I have noticed distortion reduction with separate cathode tubes, but to eliminate a variable I did most of my research on filamentary types. If you choose to experiment, be sure to allow enough time for the emission from the cathode to stabilize after changing the heater voltage. *aX*



All FET Class-A 75W Amplifier

Starting life as a Borbely kit, this project goes a step further and utilizes an external power supply, normally associated with line-amps, headphone amps, and phono preamps.

By Rick MacDonald

Back

While it is impractical (not to mention expensive) to power a Class A amp using batteries, it is, however, practical to build a separate power supply using umbilicals. Some of the benefits are obvious, but there are other reasons that some DIY builders may not have considered.

The idea for this project (**Photo 1**) began when I was listening to a headphone amp connected to a separate battery power supply. The sound of this amp was so quiet, with extended bass and very detailed mid and high frequencies, that I thought, "There must be a way to realize some of these benefits with an amplifier driving speakers, and not just any, but a full Class A amplifier." If I built this amp, it would need to be constructed from the beginning this way, and not just as an afterthought to solve problems with hum or other noise caused by transformers and their stray magnetic fields. It would need heavy well-made umbilical cables with connectors that are up to the task. With these things in mind, I decided to use a tested design for the amp and power supply, which brought me to the Borbely 75W Millennium.

DESIGN OVERVIEW

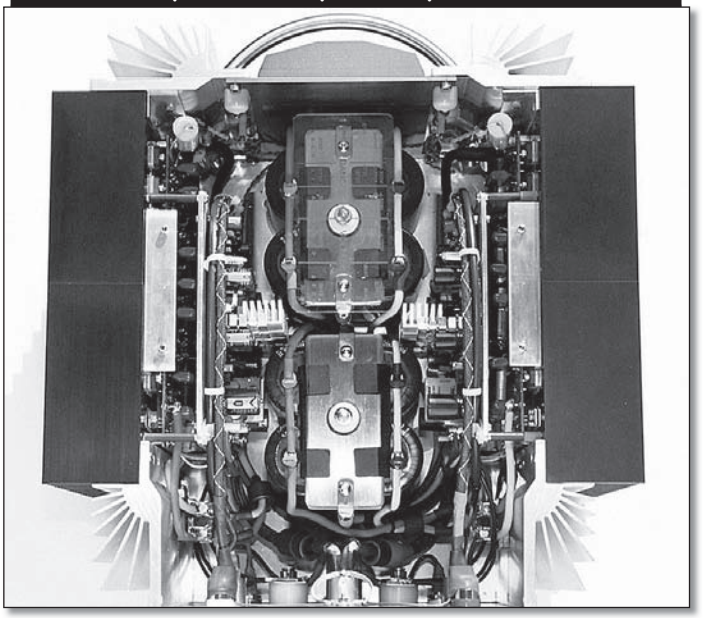
The basis for this amp comes from Erno Borbely's All FET Class A Millennium Amp (borbelyaudio.com). Intended to be built as two monoblocks, it can also be constructed as either stereo or dual mono, which is how I chose to build it (**Fig. 1**). With an array of ways to purchase it (you can buy the full kit, just the PCB, or almost anything in-between), I went with the PCBs and the semiconductors only; used my own sources for resistors, caps, and transformers; and also built my own enclosures. The amp consists of one driver board, one output board, and one power-supply board per channel. You can save yourself a lot of hunting for parts by ordering the entire kit, but if you are determined, I recommend getting the semis, as some may be tough to source, particularly the matched FETs.

The driver is an all-FET high-speed circuit with onboard regulators. Maximum supply voltage is ±55V (unregulated) and is adjustable from 30–50V. The unit can deliver 100W, but only 75W in pure Class A. Fully complementary, the input stage is differential cascode, using dual monolithic JFETs as input devices and MOSFETs as cascodes.

DC offset is tracked/controlled by the AD820, a JFET op amp powered by a shunt regulated ±10V supply. A relay out is provided to avoid the small turn-on thump (I used the direct out also provided). To handle any overly capacitive speaker loads, I used an L1R1 parallel combination on the output to control oscil-

PHOTO1: Front view of the completed amp.

PHOTO 2: Top view of completed amp.



lations, but it is easily bypassed for lower damping.

The feedback network (called a T-network) was developed for the all-FET line of power amps. It accommodates both balanced and unbalanced operation, without changing the feedback network. A switch at the input equalizes gain in both modes, and ensures equal impedance for the two inputs.

The output stage uses eight pairs of TO-3P power MOSFETs (2SK1058/2SJ162), which come tested and matched, and this batch had the same lot code, which helps ensure good matching even with bias currents other than those used for the matching process. For this reason, no source resistors are required. The sole purpose of this circuit is to move huge amounts of current, and no voltage gain is introduced. These MOSFETs become very hot and must have significant heatsinking to run Class A.

As you can see in **Photo 1**, the enclosure provides just that . . . in abundance. The big horseshoe-shaped heatsink (**Photo 2**) makes a good strong chassis. Vertical corner sinks supplement the horizontal fins of the main heatsink (black). For reasons of assembly and other design issues, it was the only way they would all fit together. Considering the enclosure was built using available parts, surplus heatsinks, and scrap pile aluminum, it came out reasonably well.

The power supply is simple, consisting of independent supplies for driver and output stages and including delay timers for power-on and speaker relays. The main power caps are separate from the kits, and 100,000µF per rail is recommended. I used NTC thermistors on the power-supply board to reduce inrush current, and you can order transformers along with the kits. I used separate surplus transformers (for driver and output boards) on this project (625VA—minimum required).

WHY EXTERNAL POWER SUPPLY?

The most obvious benefits—reductions in noise, hum, and EMI—have been proven in line and headphone amps. AC mains can bring in all kinds of garbage picked up along the way. Microwaves, RF, and switching power supplies all contribute to the hash that can cause

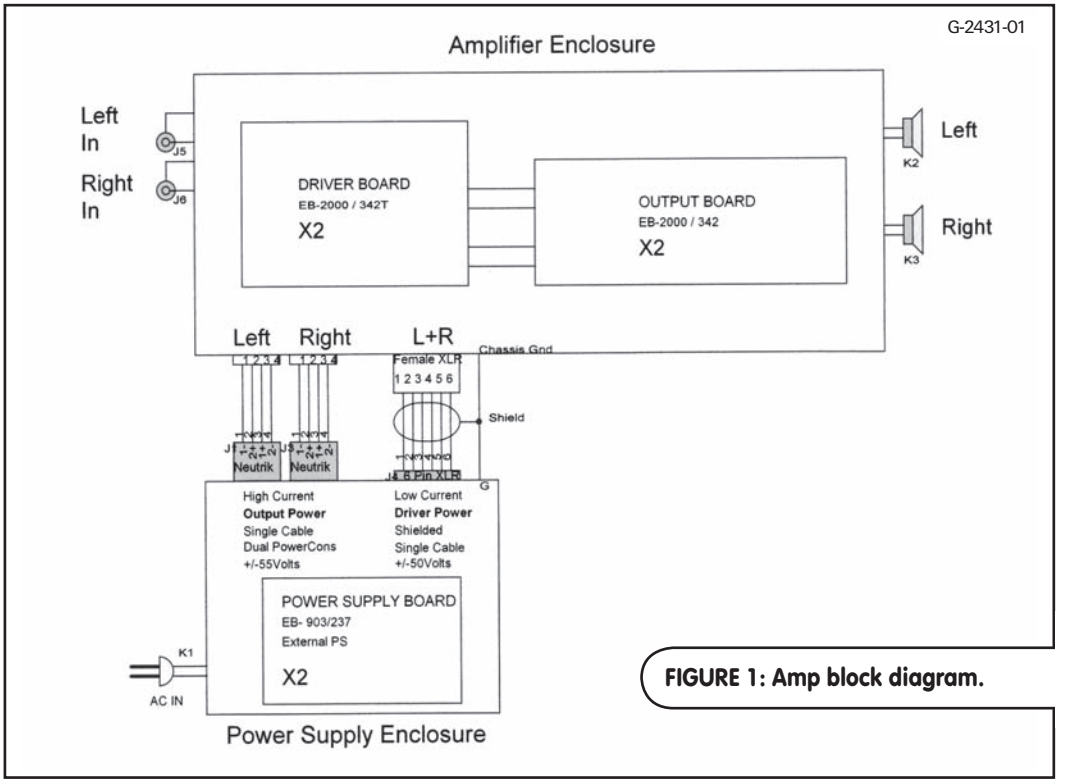


FIGURE 1: Amp block diagram.

problems in audio gear. Removing the power supply from the amp and away from the high voltage gain of the driver stage can have beneficial results similar to using a battery supply.

The lowered noise and audible improvement in the final amplifier's sound outweigh very minor losses in channel separation. And because most of the reservoir caps are in the amp chassis, high current demands are met, and minuscule losses due to umbilicals are overcome. This works because approximately one-third of the filter caps (90,000µF) are in the power supply, while the rest (200,000µF) are in the amp (**Photo 2**). Combined with the 2m umbilicals' small inductance and resistance, a pseudo pi network is produced, helping to attenuate any residual ripple or noise, essentially making the cables part of the power-supply design.

Less obvious are the benefits to DIY builders. This power supply will be used on several other different amps, some MOSFET, some bipolar, all with different topologies. Up to half the cost of building any amp is its power supply. This will save time and expense when experimenting with other designs. But even more important, eliminating the power supply as a variable will give you insight into the amp's subtleties, by knowing that changes in amp design alone have caused improvements (or not) in sound and performance.

Because there are no transformers in the amp's chassis, it is much lighter and easier to move, and transformer-induced hum is eliminated. Placement of the power-supply unit is flexible so it can be put on the floor out of the way or hidden. While working with these power-supply voltages imposes some limitations, you can use methods such as regulation and others to adjust them to required levels.

OUTPUT CONSTRUCTION

Start with the output boards, which I constructed first and mounted to the main heatsinks (**Photo 3**). This gave me the dimensions needed to mount the drivers on top and also gave me an idea as to the size of the enclosure. Because of the integral nature of the output modules, drilling and tapping were done first. I soldered in all passive components and left the MOSFETs unsoldered until the last step. Using the stuffing guide as a template (**Photo 4**), I marked and tapped all holes on the heatsinks. This included measuring and tapping four holes for the driver board mounting plate.

You must take care in the final step when soldering in the MOSFETs. You must bend and test-fit leads onto the mounted boards before soldering. Tighten the MOSFETs down first (with silpads!), then solder them in last to avoid putting pressure on the devices or their leads. Because all parts (semis

PHOTO 3: Output boards mounted to heatsinks.

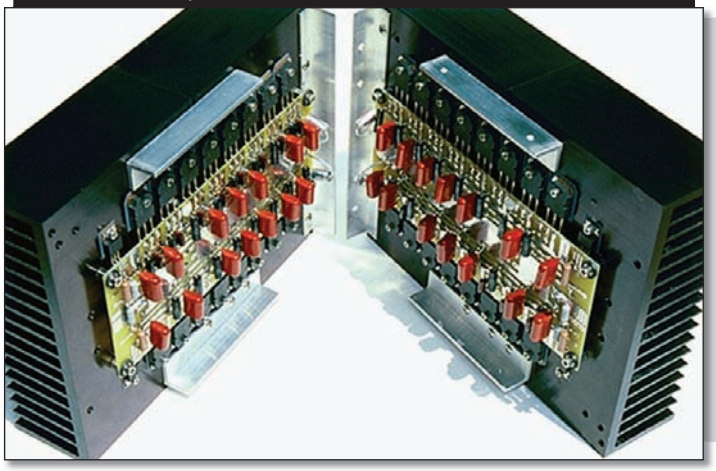


PHOTO 4: Stuffing guide used to mark MOSFET holes.

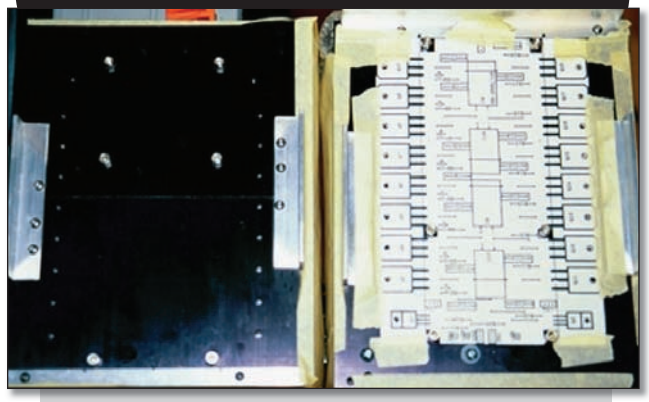


PHOTO 5: Completed driver.

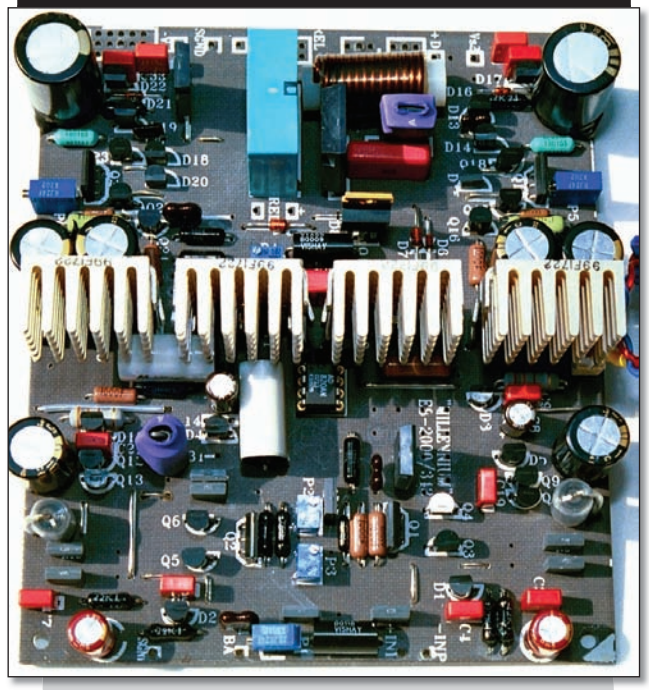
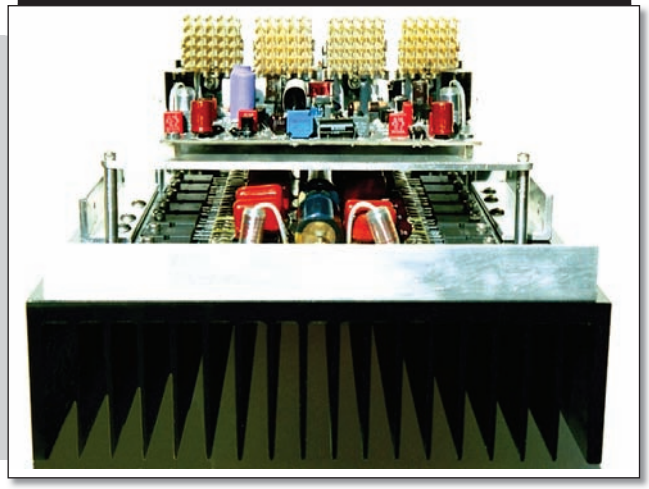


PHOTO 6: Driver and output mounted to heatsink.



in this case) are clearly labeled and placed in their own plastic bags, sorting and keeping track of what you are doing and what needs to be done is made that much easier.

DRIVER CONSTRUCTION

This PCB requires some degree of skill in soldering, and might be beyond the weekend builder. But if you have military or other soldering experience, you should have no trouble. You can purchase these PCBs as either standard fiberglass (FR-4) or Teflon (more expensive). Because of Teflon's excellent heat resistance and dielectric properties, I used it here for the driver board (**Photo 5**). The stable dielectric properties benefit this circuit because of the high gain that takes place and therefore have a more significant effect on the sound. Along with the PCB and semiconductors, I also received the relay, the inductor coil (L1), and the TO-220 heatsinks.

The rest of the components are made up of Vishay, Holco, Caddock, and Dale RN60 resistors. Cerafine, Wima, and Blackgate caps make up the bulk of the caps. The heatsinks provided are adequate, but I modified them by mounting crosscut sinks designed for a video chip (**Photo 5**). I attached these using 440 screws and thermal epoxy (Arctic silver).

There are two reasons for this: First, the boards will be mounted on their sides, reducing the cooling efficiency of the standard sinks (new ones are crosscut and work well at almost any angle). Second, I will be trying higher bias currents, with various other changes, and want to avoid heat stressing the MOSFETs (**Photo 6**).

The driver is a fairly complex PCB—some traces are thin as well as close and must be examined for solder bridges, splashes, and small metal objects, which could short these traces together. I recommend using a small handheld blue LED flashlight, which is popular and widely sold. Just make sure it uses the blue (I believe near ultraviolet wavelength) LED. After virtually inspecting with white light, I use the blue light to illuminate small particles and other debris missed by the first inspection. Blue light also illuminates cold joints and flux residue missed by other visual methods.

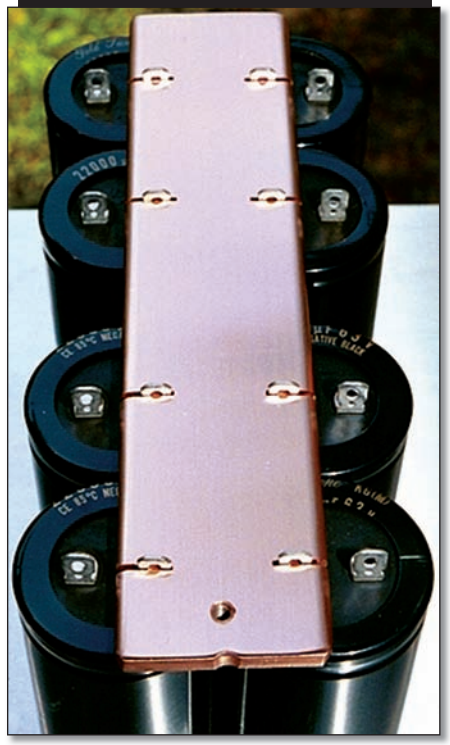
MOUNTING THE DRIVER

Once this was completed, I mounted the driver to the main heatsink (**Photo 6**), above the output board on a 5″ × 6″ aluminum plate. I installed insulating plastic (stuff used in switching power supplies) directly under the driver PCB to avoid any accidental shorting. On the bottom of the driver's aluminum plate, I installed a ⅛″ thick silicon steel plate roughly the size of the driver board to increase shielding from the output board directly below it. High currents run through the output board and induce electric fields that can cause problems on the high gain driver circuitry when it's that close. You can't see the plate from the photos, but it's there.

RESERVOIR CAPS

Because even Class A circuits draw more current on peaks, most of the power caps are placed in the amp itself. Not wishing to use the three leg brackets typically used to mount caps, I devised a way to mount them that also serves as the ground bus (**Photo 7**). I picked up these solid copper brackets (**Photo 8**) from a metal recycler and cut them to hold two banks of four caps.

PHOTO 7: Copper Gnd bus.



I had 12 Gold Tune 22,000µF caps that use a solder tab, though screw terminals would be the best way to do this. I needed to use a 100W iron and let the copper bar cool down after soldering each terminal to avoid damaging the caps. The trick was getting it hot fast, flowing the joint, and helping it cool down fast with a damp sponge, before the caps became too hot.

Once this was completed, I used large chrome Allen bolts fed down through the middle of the four caps and screwed into a large standoff or hardware store coupler (**Photo 9**). You must take care to insulate these with heatshrink and Teflon or nylon washers so that contact is not made with the GND bus bar (**Photo 8**). I used high-temp silicone on top of the caps just underneath the bus bar (sparingly), running down the four corners of the caps to strengthen everything, but not so much as to make it hard to replace one of the caps if needed. I placed some very thin high-density foam (1/16″) on the bottom of the four-cap bundle to keep pressure even on all four.

Finally, there is no need to overtighten the Allen bolt. I used Loctite to keep the hot/cold cycling from working it loose. This method has worked extremely well, takes up very little space compared to other brackets, and is rock solid. Almost ¼″ thick, the copper I used was overkill, but it conducts very well, so you could use thinner copper or even aluminum.

I used insulating plastic to keep accidental shorts away from the ±50V terminals and the copper GND bus (**Photo 10**). I installed four 2µF metalized polypropylene caps across each 50V supply on the output board (**Photo 10**, top left, right). These are shunted by four Seimens 10nF polystyrene caps to improve the power-supply response to high-frequency demands.

REAR PANEL CONNECTIONS

I used a GND star, which simply involves two 1″ × 1½″ copper tabs mounted on the back panel on Teflon spacers to isolate them (**Photo 11**). All GNDs (star Gnd) are tied to these tabs—one tab per channel. Power for driver and output sections enters the panel in the center (**Photo 12**). These consist of Neutrik four-pin Powercons (2), one per channel for output power (**Fig. 1**). They use heavy silver-plated contacts rated at 20A per contact, and are keyed as well as locking. Each cable carries ±50V, PSGND, and AMP-GND.

This scheme works well, and has no GND loop problems or hum (in fact, it's absolutely quiet). I slipped large ferrite cores over each channel's bundle of power cables where they enter the amplifier. These help attenuate high-frequency EMI and other RF noise.

Driver power is actually two separate cables (L+R) in one sleeve, connected to one Neutrik 6-pin XLR. Each three-wire cable within (±55V, Gnd) has its own braided copper shield, which is grounded to the chassis only—on both ends. They are also connected every foot with a piece of the braid wrapped around both shields and soldered. This dual shield connects the two chassis (GND), keeps left and right channel power separated, and shields them from external noise. Remember that this is power for the high-gain driver boards, so it is important to keep the

PHOTO 9: Ready to mount right channel cap bank.

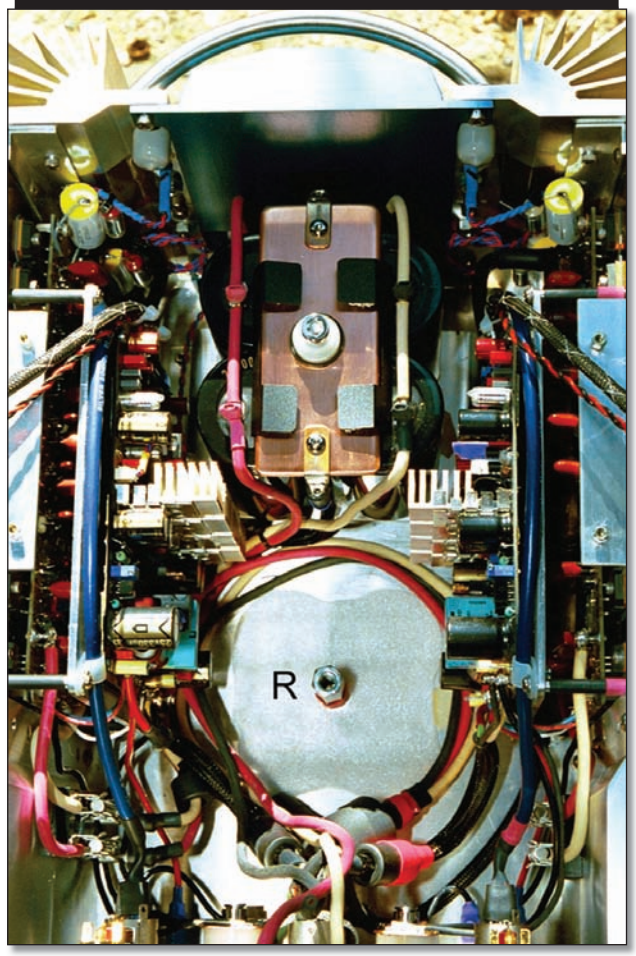


PHOTO 10: Left channel caps mounted with Gnd bracket.

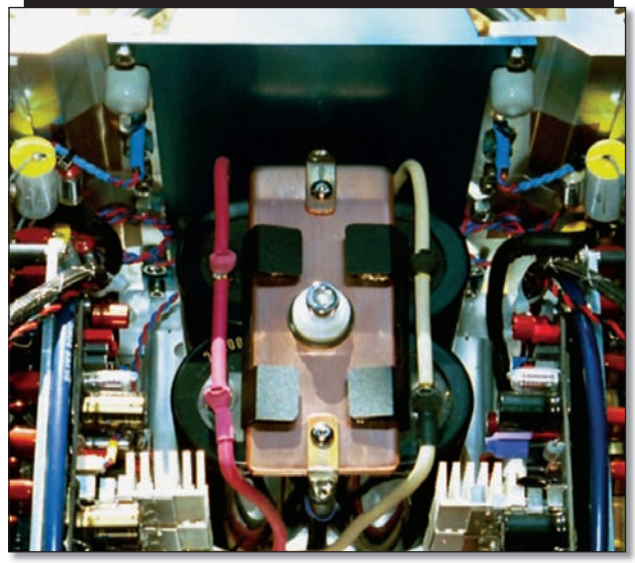


PHOTO 8: Gnd bus brackets and mounting hardware.

power coming in noise free.

Final connections include left and right RCA inputs, XLR balanced with a locking switch for balanced or unbalanced selection. The WBT heavy-duty speaker connectors are on the bottom. Splitting up the left and right inputs (usually side by side) works well here because there are essentially no stray magnetic fields from a transformer. In some cases transformers cause hum when inputs are divided in this way, but here it serves to improve channel separation.

Unbalanced connection from RCA to the driver input is made with solid silver Teflon insulated wire, which is twisted (pair) and pulled through copper braid. The braid is chassis-grounded on one end, floating on the other. Balanced operation uses Silver Sonic BL-1 Series II silver-plated cable.

The amp is positively quiet, with no hint of hum or hiss. Even with the volume way up, I can barely hear any hiss at all when it is driven by my McIntosh line-amp. This speaks volumes about the excellent Borbely design, and I believe the external power supply certainly has helped in this regard, but I must compare this amp with the monoblock version before I can make an unbiased judgment.

POWER SUPPLY UNIT

Power Supply Construction—Building

PHOTO 11: Front/top view of completed amp.

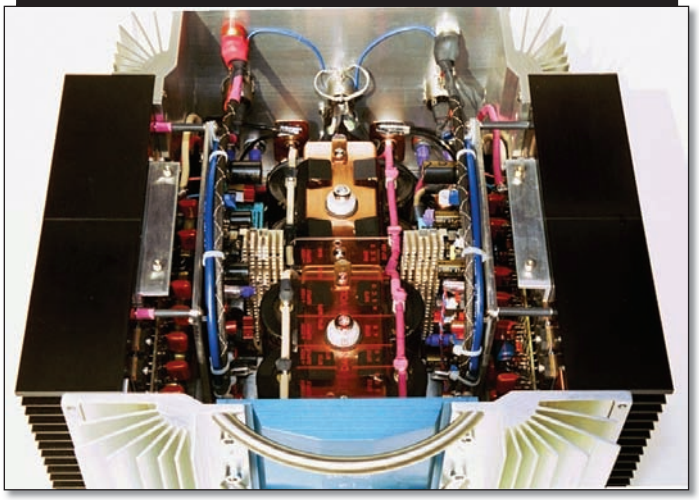
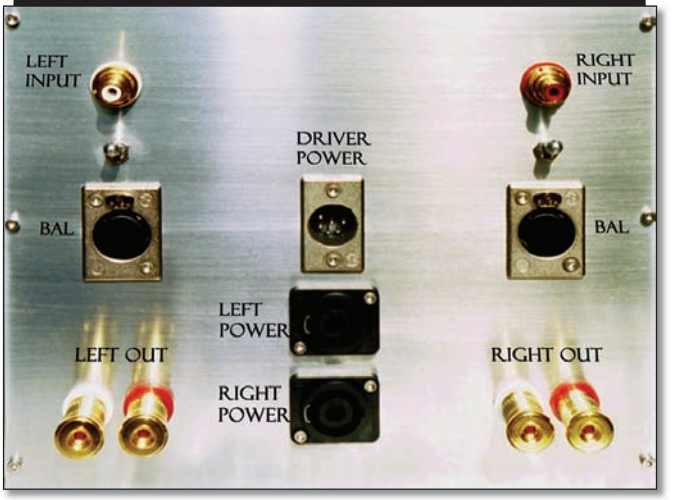


PHOTO 12: Amplifier rear panel.



the external power supply (**Photo 13**) for this amp required an extra heavy chassis to support the heavy transformers and other components. I turned to an old Hewlett-Packard tube pulse generator that had been parted out years earlier because of an open high voltage transformer and burned PCB. I chopped down the width, added some paint (auto), hardware (**Photo 14**), and aluminum to cover it.

Mounting the transformers involved a bit of luck, as they just fit onto the sides, but I needed to pot them with hi-temp epoxy and thread some ¾″ nylon rod for mounting bolts. Plitron does this all the time, and if you use the correct potting epoxy (McMaster-Carr has it), it will work (**Photo 19**). You must make sure that the mounting bolts don't touch when mounting the smaller driver transformer (**Photo 15**), which could cause a single loop, overheating, and the transformer to fail. Grain oriented silicon steel (GOSS) shields mounted around the periphery of the output transformers provide attenuation of any stray fields. 10Ga copper wire is used for all output power connections, and 16Ga for driver power.

Photo 15 shows details of each PCB mounted at each end of the enclosure, separating left and right power as much as possible. I made brackets to fit behind the PCB supporting the top and bottom of the boards.

Power caps are supported with aluminum brackets on the bottom and copper brackets on top, which also serves as the GND bus. I used separate transformers, so I needed a shielded 12V transformer to power timers and relays (bottom left of **Photo 14**). This means all 625VA goes to the output board and there is no crosstalk between windings as with transformers containing windings for both driver and output power.

For power rectifiers I used IXYS DSEP 30-12 soft recovery (called Hyperfreds) diodes for the output section. These work well and exhibit minimal noise. The power-on switch is a NOS Cutler-Hammer heavy-duty 20A DPST.

REAR PANEL CONNECTIONS AND CABLES

The power-supply rear panel (**Photo 16**) brings out all the power for the output and driver boards. As with the amp, I used Neutrik Powercons for the high current output power. **Photo 16** shows how these are mounted sideways, allowing you to grasp the locking mechanism to remove them. Two four-pin cons are used, but Neutrik makes an eight-pin version with the same ratings that you could also use.

I like these connectors so much, in fact, that I used the three-pin AC Powercon for the AC mains coming into the power supply. Originally using an expensive IEC AC connector (**Photo 17**), it kept working loose and pulling out with heavy 10Ga cable, so I switched to the Neutrik (**Photo 17**, white on blue). With 20A silver-plated contacts and heat-resistant plastic, you can screw on or solder wires. These are tough, well-made connectors. Mouser carries all of the Neutrik connectors (see resources). I used double-insulated 10Ga, 600V copper wire for the output umbilical (**Photo 18**).

Driver power uses the Neutrik six-pin XLR (female on power side, **Fig. 1**), with silver-plated 8A contacts. The

PHOTO 13: Front view of completed power supply.

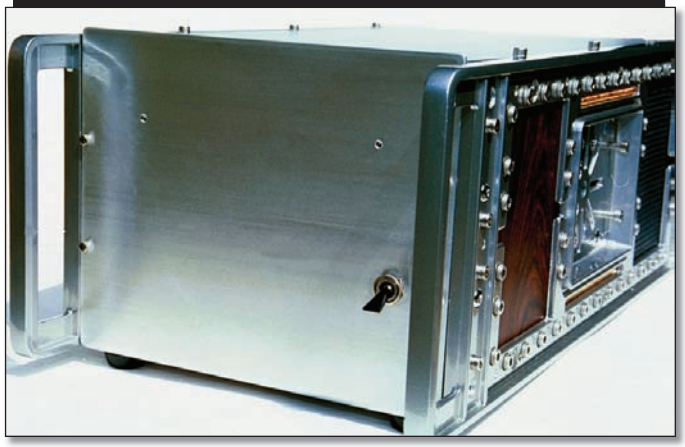


PHOTO 14: Recycled HP chassis after paint and hardware.

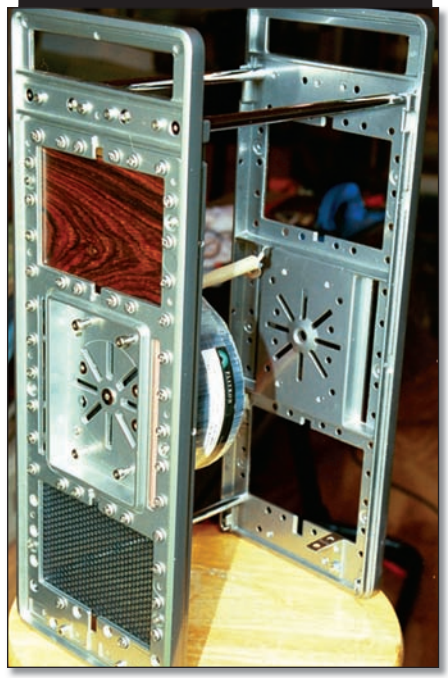


PHOTO 15: Top view of power supply.

twin copper braid used to shield each channel's power wires (±55V, GND) also ties the two chassis together. This keeps any external noise sources from contaminating driver power, and the twin shields keep interchannel crosstalk down.

I was concerned that ground loops might be produced using this method—it was so simple, but it performed better than I ever expected. I used several GND straps to ensure good connections between the sides, front, and back panels of the power-supply enclosure. The last strap goes to the safety GND on the AC Powercon coming in. After painting the power-supply chassis, I wanted to make sure all these panels were connected electrically. This was not a concern with the amp enclosure, because I used no paint.

FINAL TOUCHES

To complete the power supply, I attached two pieces of cocabola wood with brass brackets in the front and inserted perforated aluminum in the back to allow airflow (**Photo 14**). All of the (handy) mounting holes not used were filled with 10-24 Allen screws and Loctite to avoid vibrating out. Using the old HP test equipment for this beats dumping it in a landfill—even if it was beyond repair, it had a ton of useful parts. The big heavy handles make moving the power supply so much easier, and the innumerable mounting holes make it perfect for building just about anything, but please don't tear one of these apart if it can be repaired. Old tube test equipment like this is classic, and should be restored if possible.

Because I made the amplifier enclosure using surplus heatsinks and custom-made parts, you can't go out and buy them. But with some imagination, anyone can build something that looks unique or one-of-a-kind for next-to-nothing. It just takes more time rather than money. You can have some of these parts made.

The top (lid), and the front panel (blue) were machined to my drawings by Maxxus in Richardson, Tex., and blue anodized for a reasonable amount. The handles are heavy stainless steel, and made for cabinetry. The entire project was done for less money than a mid-priced, mass-produced amplifier.

PHOTO 16: Rear view of power supply.

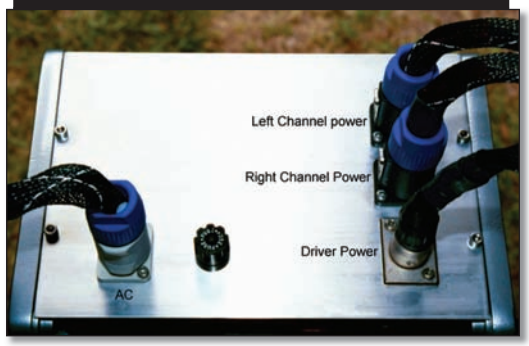


PHOTO 17: Power cables.

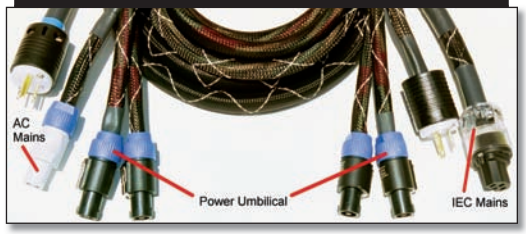
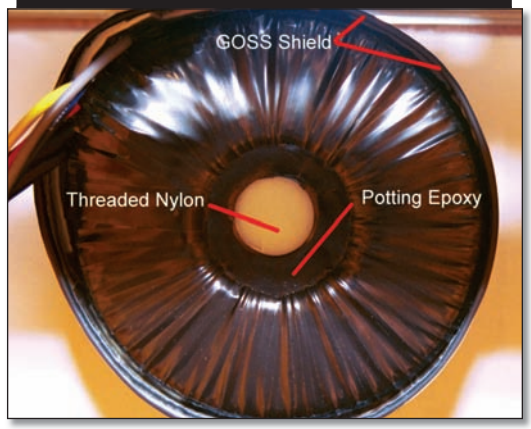


PHOTO 18: Amplifier power umbilical.

PHOTO 19: Plitron transformer after potting.



LISTENING

The sound of this amp is very good. Imaging is wide and very stable, with detail in the mids and highs, bringing out subtleties missed by lesser amps. Borbely's designs are sophisticated and rival just about anything out there.

Mr. Borbely provided helpful advice and I know of no one else who sells kits in such a variety of ways. Larry at audiokits.com also carries most of the Borbely kits, so the support is there if you have trouble. Thanks to Erno and Ed for their patience and help getting this project done. *aX*

***Rick MacDonald**, after 15 years as an engineering tech designing and prototyping broadband RF antennae and GPS satellite subsystems, moved into software engineering with E-Systems/Raytheon. Rick's most recent work has been with digital audio, composing music with MIDI and editing audio for games, software, kiosks, and websites for Gateway, Tandy, and Voyetra, among others. Rick has also written articles for **Recording** magazine, and has published several books on interactive audio and computer music. He continues to pursue the perfect power supply for all his audio projects.*

RESOURCES

Borbelyaudio.com—Amp PCBs, kits, parts, cases, Cerafine caps, tantalum resistors, silver mica caps, FETs, transfomers.

Audiokits.com—Amp PCBs, Borbely kits, audio parts, enclosures for Hpamp, FETs, MOSFETs.

Percyaudio.com—Components, braided shield, Caddock, Kiwame resistors, MUSE caps, Goldtune caps, silver wire, audio parts.

Mouser.com—Components, Neutrik Powercone, XLR connectors, RCA jacks, more.

bgmicro.com—Surplus heatsinks, batteries, surplus parts, much more.

DIYHIFISupply.com—DACT attenuators, audio components, Kiwame resistors, kits.

McMaster.com—High temp epoxy, adhesives, nylon rods, copper, aluminum.

Digikey.com—IXYS Hyperfreds, components.

CELEBRATING 75 YEARS

California's only audio-video and music show for 2005, the **Southern California Audio, Video and Music EXPO**, to be held Nov. 19 and 20 at the Embassy Suites in Pasadena/Arcadia, will be commemorating the 75th anniversary of high quality sound in movies. Former Technical Editor of *Vacuum Tube Valley*, John Atwood, will present an 8-hour comprehensive seminar, while manufacturers will display and demonstrate the latest in home theater systems, loudspeakers, cables, and more. For more information, visit **www.vacuumtube.com.**

THE TSMD2

With 97dB sensitivity and 100W maximum power handling, the **Tom Simmons Memorial Design number 2**, from TDL Technology is now available. Each speaker is a vertical line-array with ten 3″ drivers. To learn more, contact Ron Tipton at 505-382-3173, FAX 505-382-8810.

QUARTERLY CATALOG

Mouser Electronics, Inc., has released its newest catalog, with 1,532 pages of the latest products and technologies from its more than 300 electronic component suppliers. As the only distributor in the industry to publish a new catalog every 90 days, the most recent one features 56 pages of new product information from Mouser's broad-based product line. To learn more about Mouser Electronics, Inc., visit **mouser.com.**

CLIOWIN 7

Audiomatica has released its latest CLIO version, **CLIOWin 7**, a professional suite to execute electrical and acoustical measurements under Windows. Possible measurements include: sinusoidal analysis, linearity and distortion analysis, waterfall plots, frequency counter, Thiele and Small parameters, and so on. For more information, go to **www.audiomatica.com** or **www.cliowin.com.**

NEW TECHNOLOGY FROM IMPACT ACOUSTICS

All **Impact Acoustic Velocity interconnects**, from Impact Acoustics, feature 24k gold-plated contacts, high purity copper conductors, and tinned copper braid and 100% foil shield. To find out more, visit **www.impactacoustics.com.**

IT'S CATALOG TIME AGAIN

MADISOUND has issued a big revised catalog marking over 25 years of selling loudspeaker components from manufacturers from around the world. Though there are no page numbers, the book is about ⅜″ thick and crammed with a dazzling array of loudspeaker drivers and speaker related products. Madisound is at 8608 University Green, PO Box 44283, Madison WI 53744-4283, USA or e-mail at info@madisound.com. Their web address is **http://www.madisound.com.**

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Note: we are always pleased to receive catalogs since they provide the starting point to any amateur's project. Send yours to Editorial, Audio Amateur Inc., P.O. Box 876, Peterborough NH 03458, USA.**—ETD.**

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