Crowhurst's Twin-Coupled Amp Revisited

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

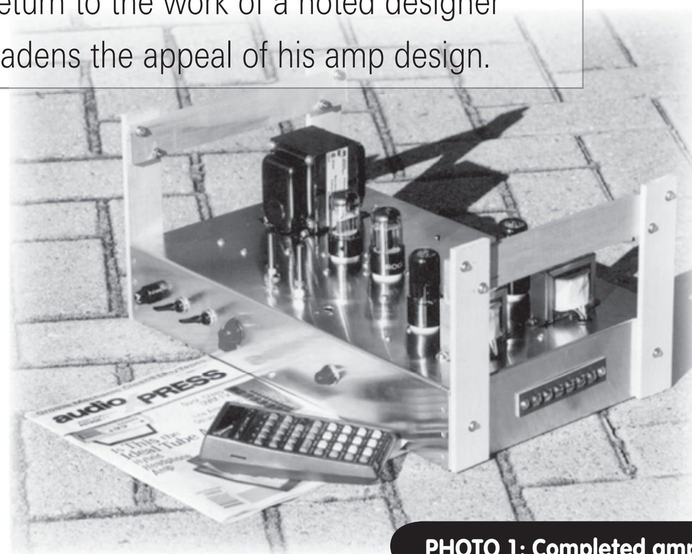


PHOTO 1: Completed amp.

By John Stewart

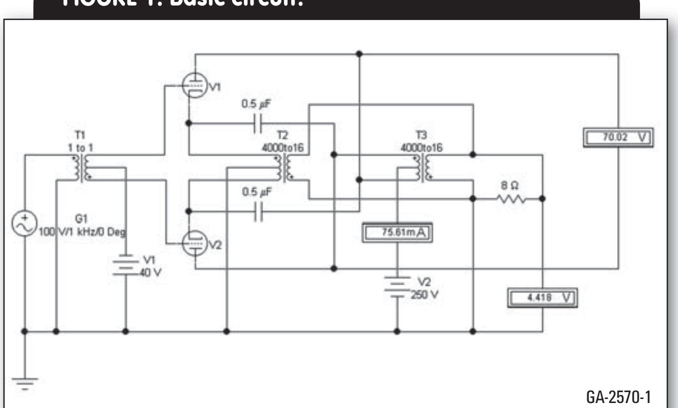
While completing my earlier version of Crowhurst's amplifier[1], I began to think there might be still more changes that would enhance its operation and at the same time further lower its cost. As well, not everyone wants or needs more than 30W as I had built into the previous amplifier. With those thoughts 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.

FIGURE 1: Basic circuit.



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.

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.

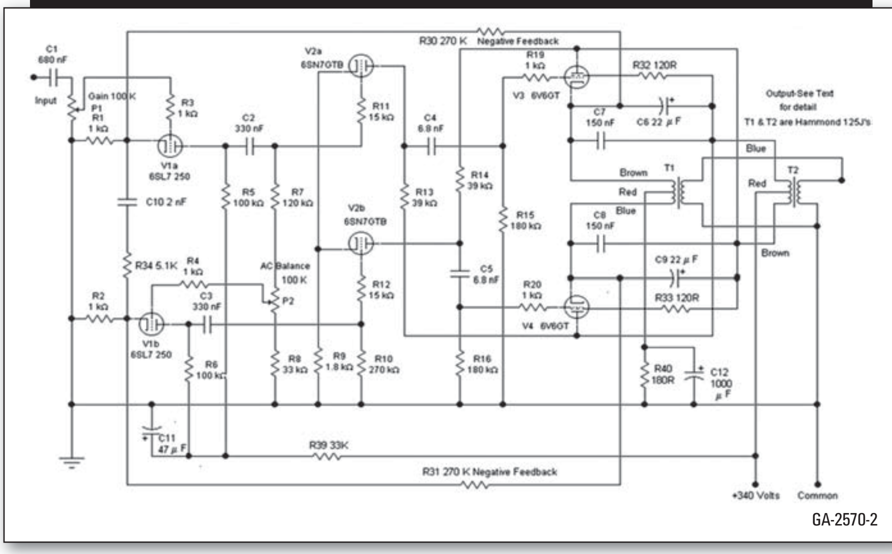
Furthermore, Crowhurst does not

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.

FIGURE 2: Cathode biased amplifier.

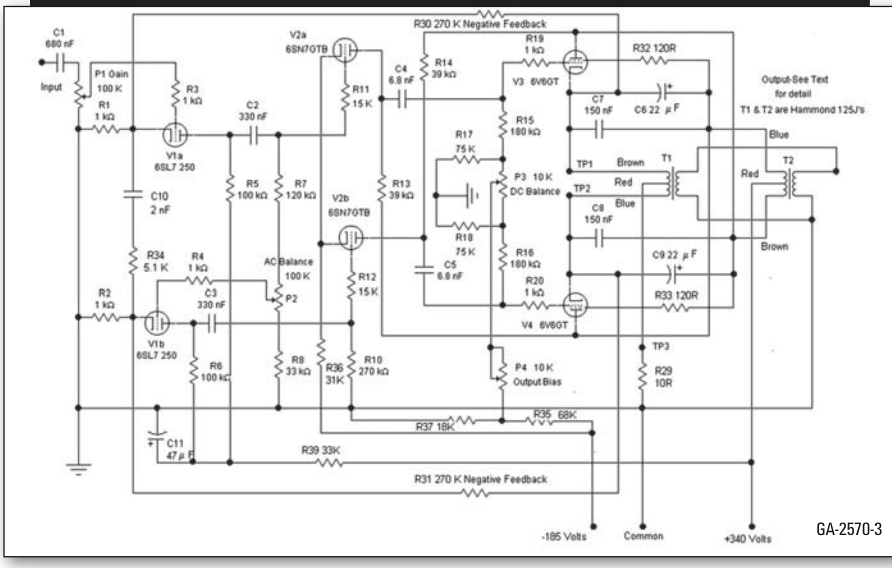


Also, you should notice that the output stage needs a large driving voltage. That problem needs to be solved in any amplifier in which part or the entire load is in the cathode circuit. Bootstrapping of the driver stage is used successfully in both the McIntosh and Circlotron. Crowhurst uses it in his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

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

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.

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.

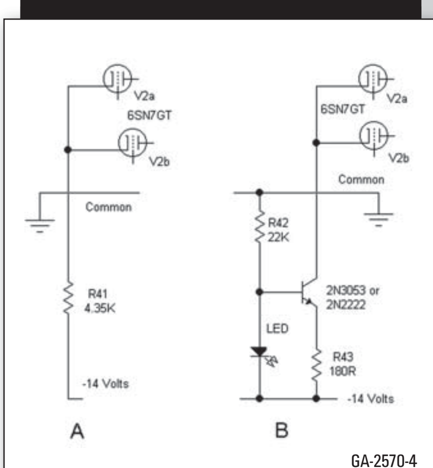
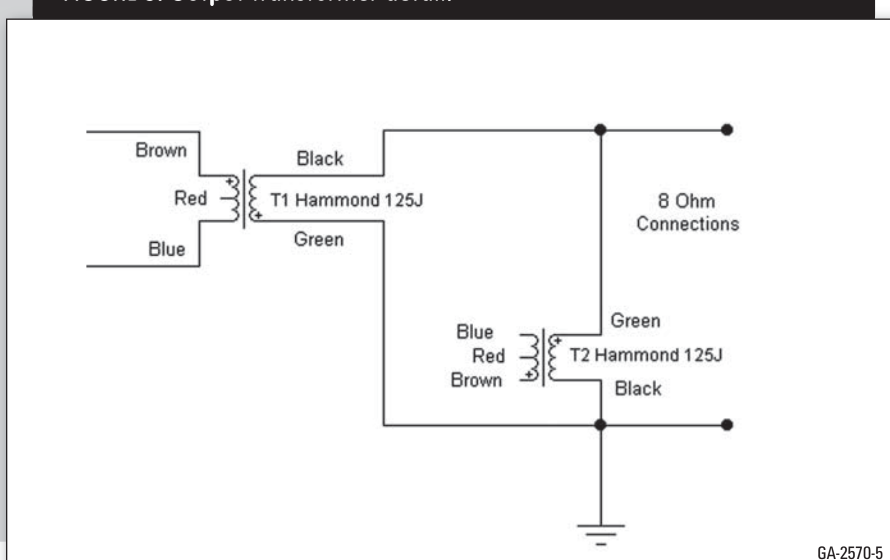


FIGURE 5: Output transformer detail.



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

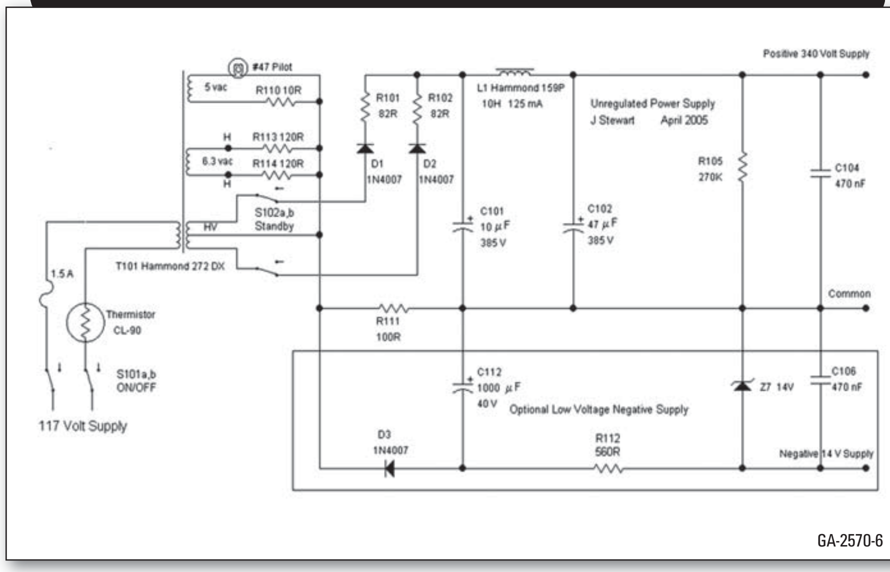
The power supply is relatively straightforward in its simplest form (**Fig. 6**). All of the power is easily provided by a Hammond 272DX transformer. It is rated for 125mA DC, while the amplifier connected for cathode bias operation runs about 100mA at no signal.

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.

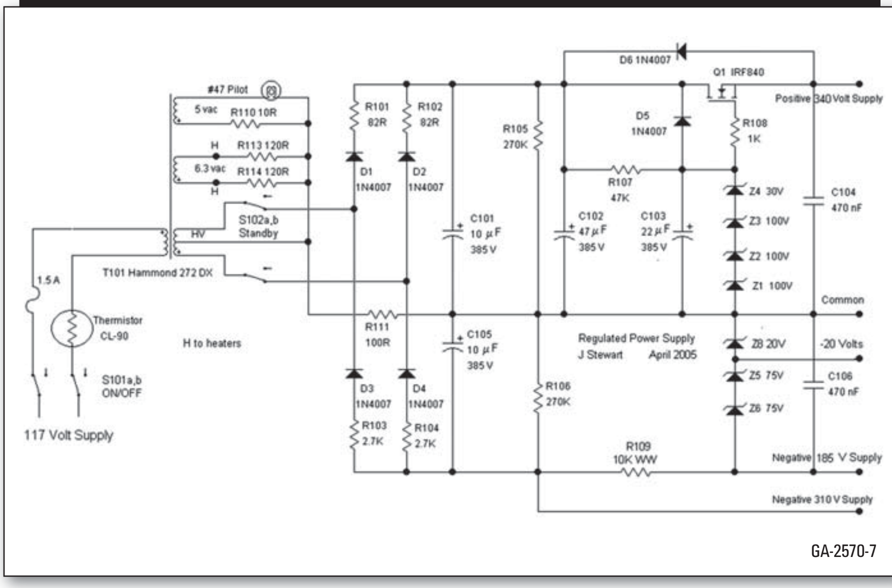


I've used a 47µF cap at the output of the filter so that even the lowest frequencies possible in this amplifier are well handled. After going through the Hammond 159P 10H choke, the HV ripple is less than ½V. The filter resonance is at 7.8Hz.

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 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 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.

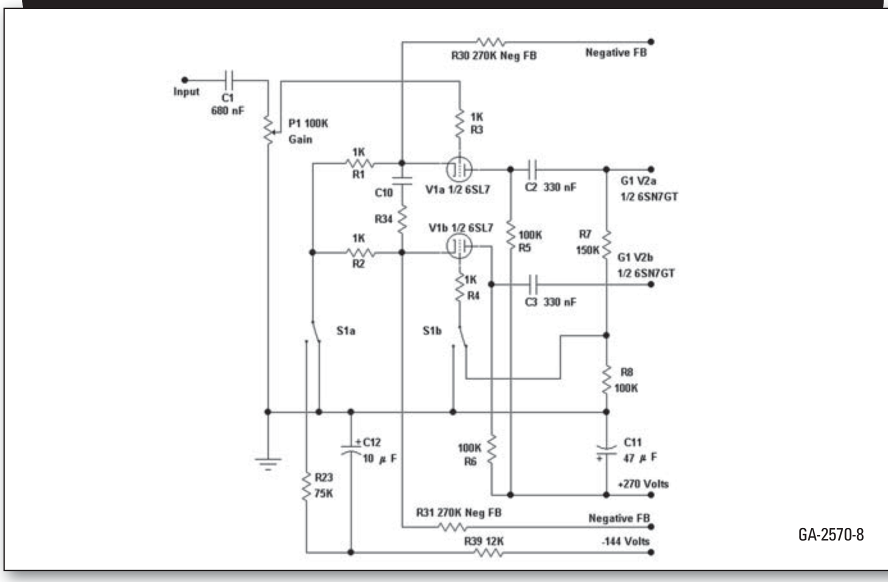
To optimize this mode of operation the CR time constant in the cathodes of the 6SL7GT needs to be changed. C10 is then 6.8nF while R34 becomes 2.2k. If you so choose, you can also easily connect this front end to a differential source. If you use the switches to change the operation mode, be sure the 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

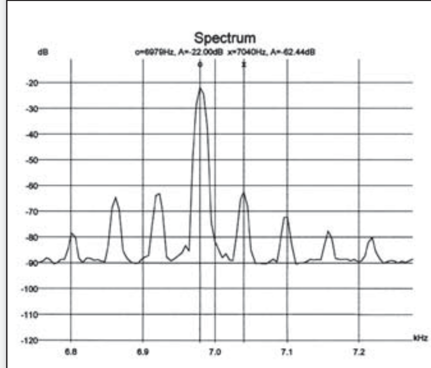
* Indicates no THD Data

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.

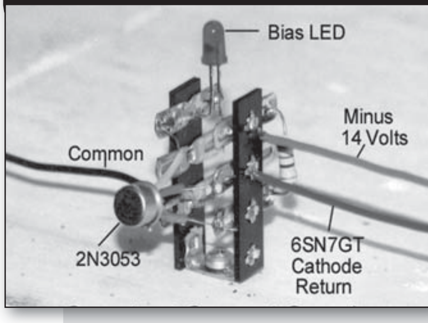


PHOTO 4: Negative filter and voltage regulator.

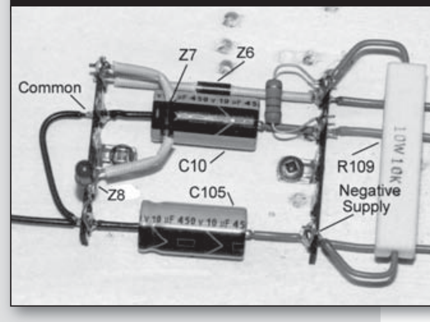


PHOTO 3: -14V supply.

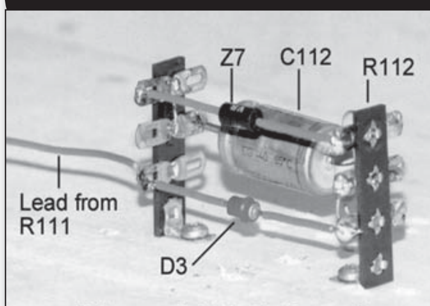


PHOTO 5: Positive filter on assembly board.

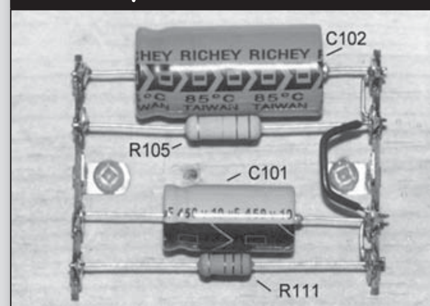


PHOTO 6: Positive voltage limit and delay.

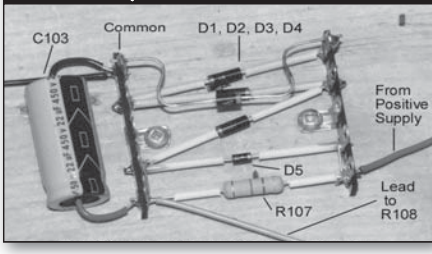
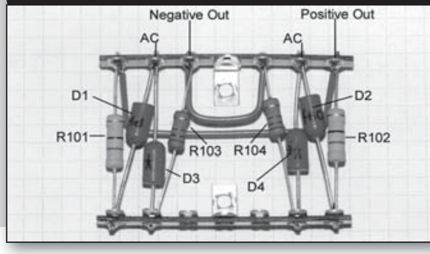


PHOTO 7: Rectifier module.



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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