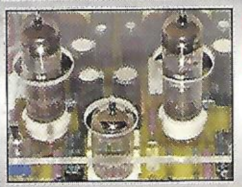
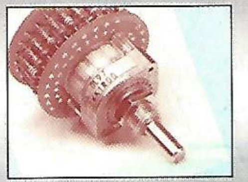


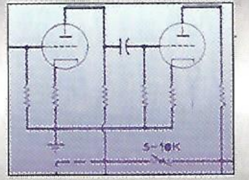
Frans Scheyving, p. 32

3 Stages 2 Tubes 4.3 Watts



Dan Norman, p. 38

Matched Precise Levels



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Better Power Designs

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GLASS AUDIO

70W 6L6GC Low-Distortion Amp

BY SCOTT K. REYNOLDS

This construction project began several years ago when I bought a pair of Magneplanars. It wasn't long after I set them up in my living room that I realized I would need a more powerful amplifier. I had been using a 20W-per-channel, Class A tube amplifier, and although it was delightful to listen to in combination with the Maggies at low or moderate volumes, it really didn't have enough punch to reproduce symphonic music or jazz at realistic levels.

I decided to try the Maggies with my Williamson 20/20 amplifier,[1] which (with a regulated power supply) will deliver 60+W into a 5Ω load. I found the 20/20 to be a more satisfying choice for reproducing higher-volume levels, and I suppose that a reasonable person would just have sat back to enjoy the music at that point. However, no one has ever accused me of being reasonable about stereo equipment, and I still thought there was more to gain sonically and aesthetically by building a more powerful tube amplifier.

Based on my experience with the 20/20, I set 70W per channel as a minimum design target. I also set a goal of matching the 20/20's distortion performance,

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which proved to be very difficult, since I wasn't willing to use a large amount of negative feedback in my design. Finally, I decided to use the 6L6GC as the output tube, because I liked the way it performed in my smaller class A amplifier, and also because it is used so widely in guitar amplifiers that there is a good selection of brands and types (the KT-66, 7581A, and 7027A are all essentially identical to the 6L6GC, and the Sovtek 5881 is similar).

I have found the 6L6GC to be rugged, reliable, and stable, but it does have some disadvantages. It is a very old design, with lower transconductance than more modern power tubes such as the EL-34, 6550, or 8417. Operated without feedback, it also tends to have higher measured distortion than those tubes. Nevertheless, the advantages outweighed the disadvantages for me, and I began work.

CIRCUIT DESCRIPTION

With my goal of achieving low distortion in mind, I decided to try a slightly unusual circuit topology involving nested feedback loops (*Fig. 1*). *Table 1* is the parts list. The push-pull-parallel output stage is conventional, but I used about 14.4dB of local negative feedback to linearize the 6L6GCs. This feedback is then surrounded by a conventional global feedback loop with about 15dB of negative feedback.

The result is a highly stable and low-distortion amplifier. It does not quite match the performance of the 20/20, but it does achieve THD of <0.15% at 70W and 1kHz, a respectable number in the world of vacuum tubes. Distortion decreases at lower power levels. The Plitron output transformer has a nominal power rating of 70W, but at midband (50Hz–10kHz) the amplifier delivers more than 90W at 1% THD. Other specifications are summarized in *Table 2*.

This description makes the design process seem quite simple and straightforward, but in reality I experimented with just about every aspect of the circuit before coming to the final result in *Fig. 1*. As I describe the circuit, I'll discuss some of the design tradeoffs that I made along the way, starting with the output stage.

The 6L6GCs are operated with a 500V plate supply, which is their maximum rating. Using the maximum plate-supply voltage obviously gives the most power output (assuming you don't exceed any other tube ratings), and I've found it also gives the lowest measured distortion. For push-pull-parallel Class AB_1 operation at 500V, a plate-to-plate load impedance in the range of 2.5k–3.2kΩ is appropriate, multiplying the suggested push-pull operating loads for the 6L6GC and 7027A in the RCA Tube Manual[2] by one-half.

The suggested screen voltage from the manual is in the 350–400V range, but here I have chosen to set it a bit on the low side, at 340V. Lower screen voltage yields lower screen dissipation, which in my experience results in a more stable operating point and longer tube life. Setting the screen voltage up to its 400V maximum will reduce distortion slightly and increase power output a few watts, so you can experiment here.

OUTPUT TRANSFORMER

The Plitron PAT4004 was not my first selection for output transformer, although it has proven to be an excellent choice. I began constructing the prototype using a pair of Thordarson transformers purchased many years ago, but I was very disappointed in their high-frequency power bandwidth, so I looked through *Glass Audio* to find a source for better ones.

I was initially concerned about the PAT4004's power-handling capability, since its 70W rating was at the low end of my design target. At midband, however, I discovered the Plitron transformer can handle over 100W. I have one caveat about the PAT4004's low-frequency power bandwidth that I will discuss later. I did not use the ultralinear taps on the PAT4004 in this design.

I've used a relatively large (47Ω) unbypassed cathode resistor for each 6L6GC. In most circumstances this would not be done, because it reduces the gain of the tube and increases both the plate resistance and the required signal voltage swing at the grid. In this case, however, you are controlling the gain and output resistance of the final stage with a local feedback loop, and so can get away with using the large cathode resistors. These have the benefit of equalizing the dynamic currents through the paralleled tubes, so that matched output tubes are not required. This was important for me because I had a quantity of unmatched, new old stock (NOS) RCA 6L6GCs that I wished to use in the amplifier.

I set the quiescent current in each

TABLE 1
AMPLIFIER PARTS LIST (EACH CHANNEL)

RESISTORS	
R1, R12–R15	100kΩ, ½W, 2%
R2	10kΩ, ½W, 2%
R3, R26, R27, R36, R37	100kΩ, 1W, 2%
R4	1.8kΩ, ½W, 2%
R5	27Ω, ½W, 2%
R6	27kΩ, 2W, 2%
R7	no longer used
R8	470kΩ, ½W, 2%
R9	1MΩ, ½W, 5%
R10, R11	50kΩ, 1W, 2%
R16, R17	910Ω, ½W, 2%
R18, R19	10kΩ, 2W, 2%
R20, R21	500Ω, ¾W, 15-turn trimmer
R22, R23, R28, R29	1kΩ, ½W, 5%
R24, R25, R30, R31	47Ω, 2W, 2%
R32, R35	100kΩ, 2W, 2%
R38, R39, R42, R43	10kΩ, ¾W, 15-turn trimmer
R40, R41, R44, R45	10kΩ, ½W, 5%
R46	330Ω, 3W, 2% (three 1kΩ, 1W resistors in parallel)
R47	5kΩ, ¾W, 15-turn trimmer
CAPACITORS	
C1	2µF, 400V, film
C2, C3	0.1µF, 600V, film (don't be tempted to substitute a higher capacitance. The relatively small value sets the low frequency dominant pole of the amp)
C4	1µF, 250V, film
C5	47pF, 1kV, "silver mica" style film or polystyrene
C6–C9	2.2µF, 600V, film
C10–C13	22pF, 1kV, "silver mica" style film or polystyrene
C14	8200pF, 100V, polystyrene
C15–C18	0.22µF or more, 600V, film (if both channels are built on the same chassis, C15–C18 need not be duplicated)
TUBES	
V1	5879 pentode (connected in the circuit as a triode with µ = 21. Could substitute 6J5, 6C4, half of a 6SN7GT/B, etc.)
V2, V4, V5	12AU7 or 5814 dual triode (I don't recommend Chinese 12AU7. Could substitute 6SN7GT/B.)
V3	6SN7GT/B (I prefer NOS.)
V6–V9	6L6GC, 7581A, 7027A, KT-66, Svetlana 6L6GC, or Sovtek 5881. (6L6, 6L6G, 6L6GA, 6L6GB, 5881 other than Sovtek, and all Chinese 6L6 types are unsuitable because they have lower ratings.)
T1	Plitron PAT4004, 2700Ω plate-to-plate primary, 5Ω secondary (ultralinear taps are not used. Don't substitute unless you're experienced; see text.)
Q1, Q2	MPF102, n-channel Si JFET

output tube to 50mA, equivalent to 2.35V across each of the 47Ω cathode resistors. The grid-bias voltage to obtain this value is approximately –28V, for a total grid-to-cathode bias of about –30V. The 50mA quiescent current gives a total plate-plus-screen dissipation of 25W, which is well within the 30W plate dissipation and 5W screen dissipation ratings of the 6L6GC. This conservative operation of the tube should result in long tube life. Higher quiescent currents will reduce distortion, as I discuss later in the article. The grid-bias voltage for each output tube is set independently with trimmer potentiometers R38–R39 and R42–R43.

DRIVER STAGE

The driver stage is unusual, so it deserves some attention. The active gain element for the top side of the push-pull circuit is one section of a 6SN7 (V3A) with a very large (10kΩ) cathode resistor (R18) to accept feedback from the 6L6GC's plates via two 100kΩ resistors in series (R32 and R33). The unbypassed 10kΩ cathode resistor would ordinarily reduce the gain too much, so I have provided a constant-current source load consisting of an FET (Q1) and the paralleled sections of a 12AU7 (V4).

A low-impedance output is taken from the common cathodes of the 12AU7 sections. This stage is a modified mu-follower, with the bias current set to 5mA by a small potentiometer (R20) in the source lead of Q1. If you break both the inner and outer feedback loops, you can measure a gain of approximately 105 (40.4dB) from the grids of the 6SN7 to the plates of the output tubes. With the inner feedback loop closed, this gain is 20 (26dB), so there is approximately 14.4dB of negative feedback.

This local feedback loop around the output and driver stages not only helps to reduce the distortion of the 6L6GCs, but also reduces their output impedance, making their performance less sensitive to variations in speaker-load impedance. Also, the output transformer produces less distortion and has wider bandwidth when driven from a lower impedance source. Menno van der Veen discusses this effect briefly in a letter published in *Audio Electronics* (1/98).[3]

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TABLE 2

AMPLIFIER SPECIFICATIONS

Power output:	70W per channel into 5Ω, nominal power rating. Actually delivers >90W continuous power into 5Ω at 1% THD, 50Hz–10kHz
Frequency response at 7W:	+0/–1dB from 9Hz–21kHz +/–3dB from 4Hz–46kHz
Power bandwidth at 70W:	+0/–3dB from 24Hz–46Hz
Total harmonic distortion:	0.15% at 70W and 1kHz. See *Figs. 3–4*
Intermodulation distortion:	Intermodulation distortion: 0.4% at 70W, 60Hz, and 6kHz, mixed 4:1. See *Fig. 5*
Closed-loop gain:	22dB
Sensitivity:	1.5V RMS for 70W into 5Ω
Open-loop gain:	37dB

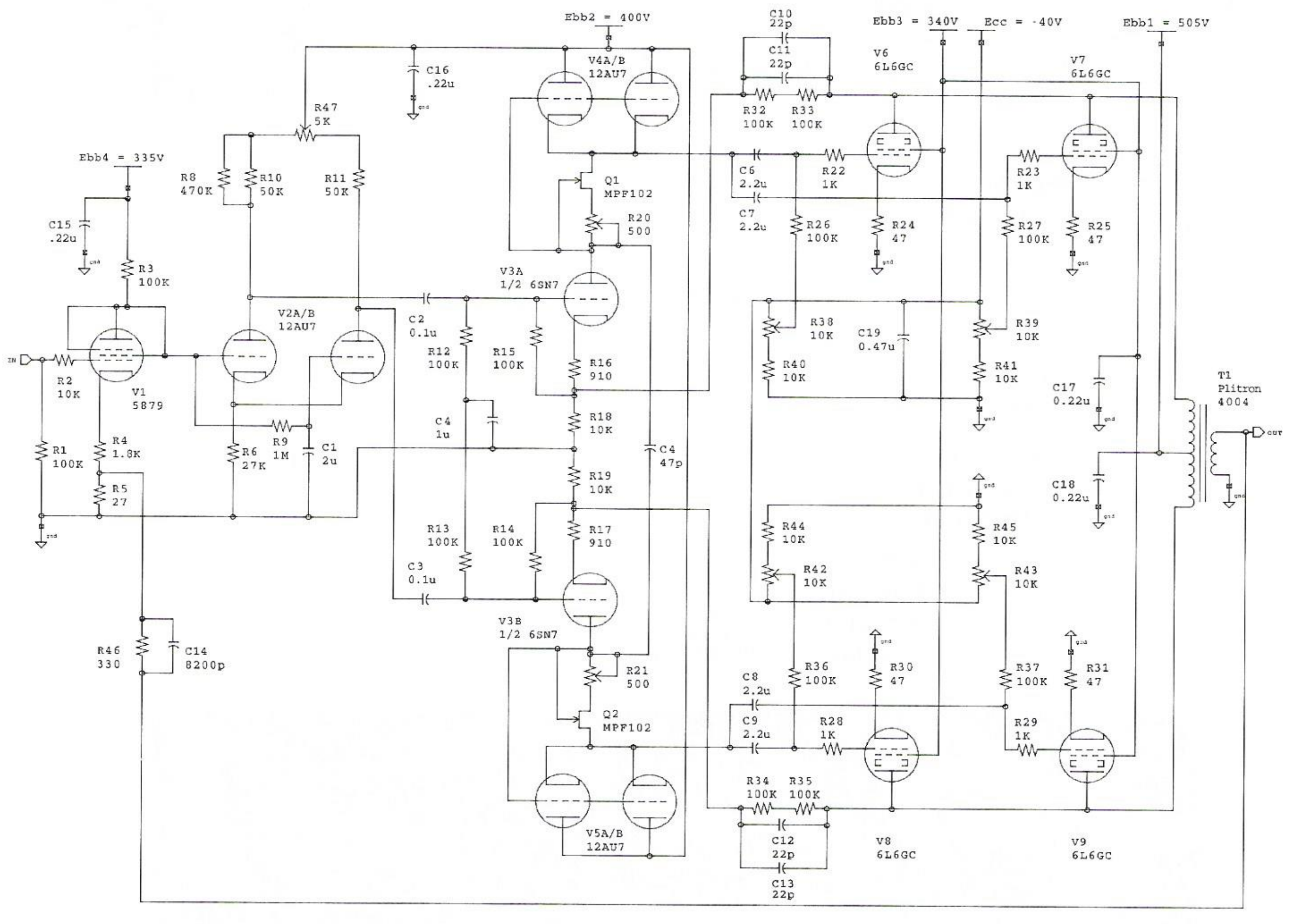

FIGURE 1: Amplifier schematic.

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INPUT STAGES

The input stages consist of a conventional voltage amplifier V1 and "long-tailed" phase splitter V2. I experimented with constant-current biasing of V2, as a replacement for the cathode resistor R6. This balanced the two out-of-phase outputs without the need for unequal plate-load resistors, but I found that I still needed an AC balance control (R47) in order to obtain minimum IM distortion for the amplifier as a whole. Therefore, I eliminated the constant-current source and reinstalled R6 to simplify the circuit.

A triode-connected 5879 pentode is used for the input-voltage amplifier. The 5879 is a low-noise, low-microphonics tube originally used primarily for microphone amplifiers. It produces very low distortion as a triode (it's pretty good as a pentode, too), and I've had good results with it. Note that both the screen and the suppressor are connected to the plate for triode operation. So connected, the 5879 has a mu of 21.

You may have noticed that all the tubes used in the voltage amplifier and driver stages are medium-mu triodes. In the parts list, I've suggested several possible tube substitutions, since a wide variety of medium-mu triodes exist. For instance, you could build the amplifier entirely with 6SN7 tubes if you wish. Note that I haven't built and tested all these combinations, so to be conservative, a beginner should stick to the schematic as drawn, but I can't think of a reason why the substitutions wouldn't work.

With the outer feedback loop open, the overall amplifier gain from input to speaker output is about 70, or 37dB. With the loop closed, the gain is about 22dB, so there is 15dB of negative feedback around the entire amplifier. In using negative feedback, you need to be concerned about stability. In this case, with nested feedback loops, stability considerations become a bit more complicated than usual, and the amplifier must be frequency-compensated by working from the inside feedback loop outward to the global loop.

CONSERVATIVE DESIGN

I've chosen to be very conservative in my design, favoring stability over extremely wide bandwidth. Experienced readers can experiment with frequency compensation to suit their own preferences, but inexperienced builders should follow the schematic exactly as shown to avoid problems. It is important to note that the values of frequency-compensation components depend very strongly on the choice of output transformer, so you should not substitute for the Plitron PAT4004 unless you are willing to compensate the amplifier yourself.

Capacitor C4 places a high-frequency dominant pole at roughly 13kHz within the inner feedback loop. Capacitors C10–C13 provide lead compensation for the inner feedback loop. This compensation decreases the closed loop gain of the driver and output stages at high frequencies, with a 3dB point at 15kHz. The location of this 3dB point is determined partly by the values of C10–C13 and R32–R35, and partly by the location of the dominant pole created by C4. This lead compensation of the inner feedback loop creates a 15kHz dominant pole for the outer feedback loop.

Finally, the outer feedback loop is lead-compensated by capacitor C14. Closed-loop gain decreases at high frequencies, with a 3dB point of 46kHz. The location of this 3dB point is determined partly by the parallel combination of C14 (8200pF) and R46 (330Ω), and partly by the location of the outer feedback loop dominant pole at 15kHz, as described earlier.

This frequency compensation gives the amplifier excellent high-frequency stability. You also need to be concerned

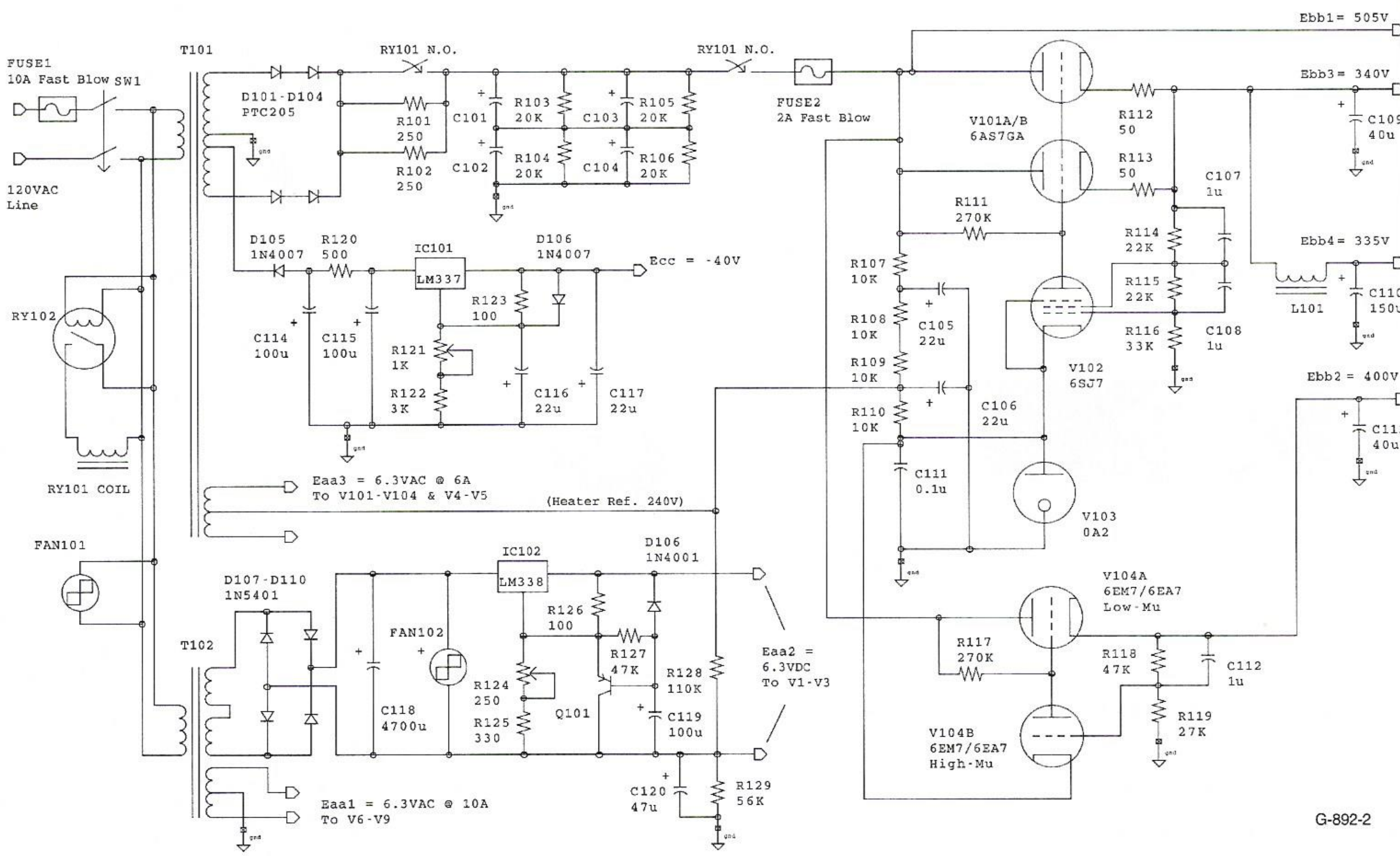

FIGURE 2: Power-supply schematic for two stereo channels.

TABLE 3
POWER-SUPPLY PARTS LIST (FOR TWO CHANNELS)

RESISTORS	
R101, R102	250Ω, 10W, wirewound
R103–R106	20kΩ, 10W, wirewound
R107–R110	5kΩ, 5W, wirewound
R111, R117	270kΩ, 1W, 5%
R112, R113	50Ω, 2W, 5%
R114, R115	22kΩ, 3W, 5%, metal oxide
R116	33kΩ, 3W, 5%, metal oxide
R118	47kΩ, 3W, 5%, metal oxide
R119	27kΩ, 3W, 5%, metal oxide
R120	500Ω, 1W, 5%
R121	1kΩ, ½W, trimmer
R122	3kΩ, ½W, 5%
R123, R126	100Ω, ½W, 5%
R124	250Ω, ½W, trimmer
R125	330Ω, ½W, 5%
R127	47kΩ, ½W, 5%
R128	110kΩ, 1W, 5%
R129	100Ω, ½W, 5%
CAPACITORS	
C101–C104	2500μF, 300WV DC, computer grade electrolytic
C105, C106	22μF 450WV DC, electrolytic
C107, C108, C112	1μF, 250V, film
C109, C113	40μF, 450WV DC, electrolytic
C110	150μF, 450WV DC, electrolytic
C111	0.1μF, 600V, film
C114, C115	100μF, 100WV DC, electrolytic
C116	22μF, 1000WV DC, electrolytic
C118	4700μF, 25WV DC, electrolytic
C119	100μF, 16WV DC, electrolytic
C120	47μF, 100WV DC, electrolytic
TUBES	
V101	6AS7GA or 6080, dual low-mu triode
V102	6SJ7GT, pentode
V103	0A2, 150V gas-filled regulator
V104	6EM7/6EA7, dual triode (the high-mu section is V104B, the low-mu section is V104A)
DIODES	
D101–D104	PTC205, 1000PIV, 2.5A rectifier diode
D105, D106	1N4007
D107–D110	1N5401
D111	1N4001
Q101	2N2907 or equivalent PNP Si bipolar
IC101	LM337T, TO-220 adjustable regulator
IC102	LM338K, TO-3 adjustable regulator (mount on large heatsink or chassis)
L101	15H @ 75mA filter choke (if no power take-off is desired, may substitute a 2.2kΩ, 1W resistor; see text)
T101	400-0-55-400V @ 800mA (420-0-60-420V no load) 6.3V @ 6A
T102	6.3V @ 10–12A, 5V @ 3A, 5V @ 3A
RY101	10A DPDT relay with 120V AC coil
RY102	Amperite 115NO30 thermal-delay relay (could substitute any electronic or mechanical delay relay adjusted for 30 seconds turn-on delay)
FAN101	120V AC fan, 4½″ diameter, mounted above T101
FAN102	12V DC fan to ventilate the underside of the power supply chassis; the sort used in PCs is suitable
SW1	10A DPST toggle

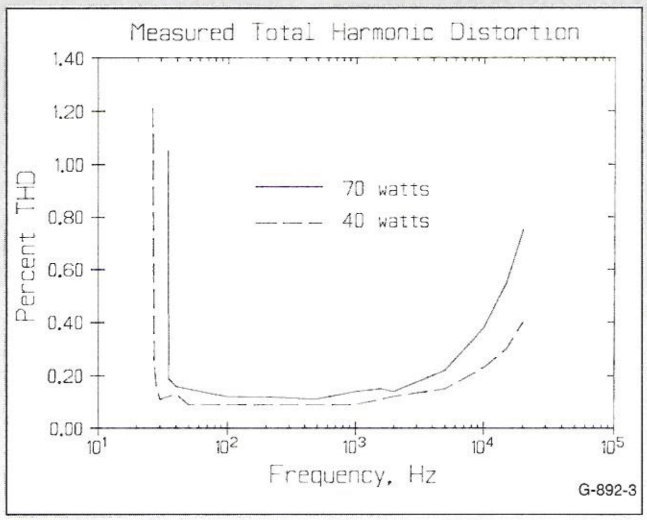

FIGURE 3: Measured total harmonic distortion versus frequency at 70W and 40W into 5Ω. Meter residual is 0.08%.

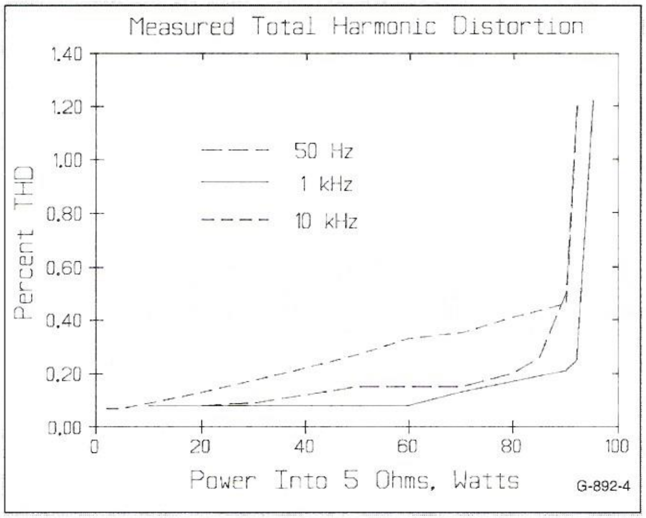

FIGURE 4: Measured total harmonic distortion versus power into 5Ω at 50Hz, 1kHz, and 10kHz. Meter residual is 0.08%.

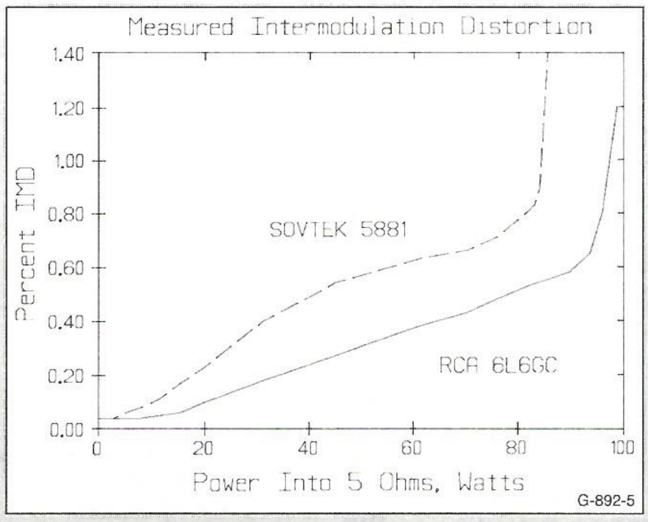

FIGURE 5: Measured intermodulation distortion versus power into 5Ω, 60Hz and 6kHz mixed 4:1. Meter residual is 0.06%. Data is shown for the RCA 6L6GC and Sovtek 5881. The Sovteks produced consistently higher distortion and lower power output than other tubes I tried.

about low-frequency stability, but that is much simpler. You establish a low-frequency dominant pole at about 11Hz by coupling capacitors C2 and C3. Any phase shift added by coupling capacitors C6–C9 or the output transformer T1 is at much lower frequencies. The resulting amplifier is stable with or without a secondary load on T1.

POWER SUPPLY

I chose to design a single power supply for both stereo channels, but there is no reason why you couldn't follow a dual monoblock approach, as long as you provide the same supply voltages and currents. Referring to *Fig. 2* (*Table 3* is the power-supply parts list), plate supply Ebb1 provides 505V under idle conditions, which is 200mA per channel. Current drain increases to 400mA per channel at maximum signal. Plate supply Ebb1 is unregulated, but computer-grade capacitors C101–C104 have a net capacity of 2500µF, which provides a very large energy reservoir.

With such large capacitors, the turn-on surge needs to be controlled, which is the function of relays RY101 and RY102. At turn-on, C101–C104 initially charge through power resistors R101 and R102, providing a controlled ramp-up of the capacitor voltage. After 30 seconds, thermal-delay relay RY102 energizes the coil of RY101, which in turn shorts out resistors R101 and R102 and supplies high voltage to the 6L6GC plates and the rest of the power supply.

The balance of the high-voltage supplies are regulated. Screen-supply Ebb3 is 340V at 45mA maximum per channel. Supply Ebb4 is 335V at 2mA per channel for the 5879 input tube. Supply Ebb2 is 400V at 15mA per channel for the phase splitter and driver.

The voltage regulators are a conventional design. Gaseous regulator tube V103 (0A2) provides a 150V reference for the error amplifiers V102 (6SJ7) and V104B (the high-mu section of a 6EM7/6EA7). V104A (the low-mu section of the 6EM7/6EA7) serves as the pass tube for supply Ebb2 and dissipates 3W. V101 (6AS7GA) serves as a pass tube for the screen supply.

Using a 6AS7 might seem a bit of overkill here, since V101 dissipates no more than 15W under maximum signal conditions, and much less than that normally. However, I wished to use the 335V supply Ebb4 as a possible power take-off for my tuner, preamp, or CD player at some point in the future, so I designed the regulator to have excess capacity. Similarly, decoupling choke L101 is grossly overrated at 75mA, since it passes only 4mA, but it will serve for the future power take-off. If no power take-off is desired, you can replace choke L101 with a simple 2.2kΩ, 1W decoupling resistor for much less cost.

Transformer T101 has a 55V tap that I used for the bias supply. A half-wave rectifier and an RC filter produces –72V at the input to the regulator IC101. The regulator output Ecc is adjusted for –40V DC. The amplifier requires several heater supplies. A separate transformer T102 has a 6.3V AC @ 10–12A winding, which I used for the 6L6GC heaters.

The heaters of tubes V4 and V5 in the amplifier and tubes V101, V102, and V104 in the power supply run off a 6.3V AC @ 6A winding on T101, which is biased 240V above ground to reduce the heater-cathode potentials. A 6.3V DC heater supply is provided for tubes V1–V3. Transformer T102 has two 5V AC @ 3A windings, which I wired in series for this supply. The circuit is a standard slow-turn-on regulator using the LM338K, a rugged three-terminal regulator in the TO-3 package, allowing for good heatsinking.

COOLING CONCERNS

I purchased transformer T101 from a surplus dealer several years ago, and I don't know of any current production product which is an exact duplicate. The T101 has a 400-0-55-400V secondary, rated at 800mA, so any transformer (or combination thereof) that provided this would be suitable. As an alternative, you could use two Hammond 278CX transformers (400-0-400V @ 465mA) in a dual-mono approach.

Transformer T101 proved not to be as conservatively rated as I had hoped, and in my prototype the temperature of T101 reached 60–65°C after several hours of operation, heating up the entire power-supply chassis. After prolonged high-power testing, it became even hotter. I considered this unacceptable, so I provided fan cooling for T101 and for the power-supply chassis. FAN101 is a 4″-diameter, 120V AC cooling fan mounted directly above T101. FAN102 is a small 12V DC fan (the sort used for cooling computers) that I used to ventilate the underside of the power-supply chassis. T101 and the supply chassis now barely rise above room temperature. If you wish to avoid fan cooling, I recommend a more conservatively rated power transformer.

In my setup, the power supply is mounted remotely from the amplifier, connected to it via a 12′ cable that runs from my listening room into the cellar, where the power supply sits on a shelf. This keeps the fan noise, transformer buzz, and power-supply heat dissipation out of the listening environment. The only caveat is that the heater supplies require heavy-gauge wiring to prevent excessive voltage drop; I used paralleled sections of 14-gauge stranded wire.

The overrated heater windings result in more than 6.3V AC at the transformer secondaries, so a drop of a few tenths of a volt in the cable is acceptable. With such a long cable, the DC power supplies need to be bypassed on the amplifier chassis; that is the function of C15–C19.

PERFORMANCE EVALUATIONS

Distortion measurements on the amplifier are summarized in *Figs. 3–5*. I measured harmonic distortion using the spectrum-analysis capability of a digital-storage-oscilloscope (DSO) card that I have for my lab computer. This DSO can resolve distortion down to about 0.08%, which is the noise floor in spectrum-analysis mode. Amplifier distortion is predominantly third harmonic, with a small amount of uncancelled second harmonic at low frequencies.

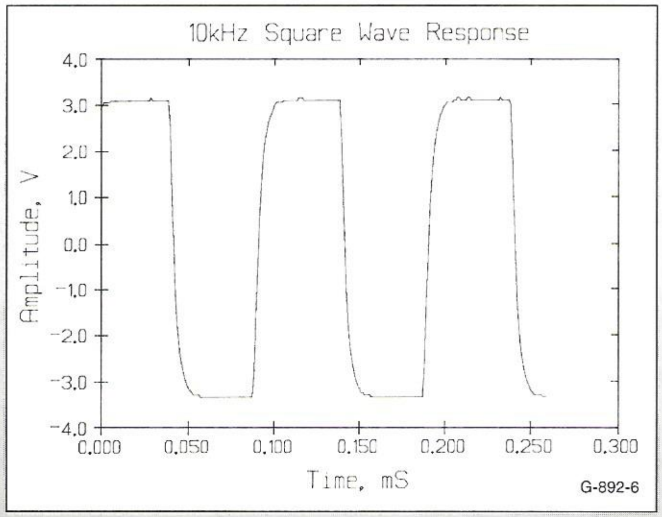

FIGURE 6: Square-wave response, 10kHz into 5Ω resistive load.

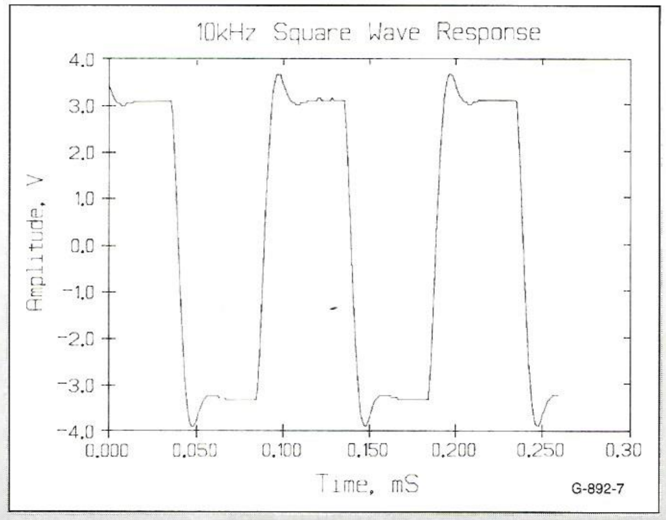

FIGURE 7: Square-wave response, 10kHz into 5Ω in parallel with 2.2µF, demonstrating stability into capacitive loads.

At 1kHz, distortion is at or below the noise floor of the DSO for power levels below about 50W. Distortion rises at high frequencies, as is common with most tube amplifiers. At 70W and 10kHz, harmonic distortion is 0.38%, and at 20kHz it reaches 0.75%. This rise in distortion at high frequencies is partly due to a reduction in negative feedback caused by falling loop gains, a byproduct of the frequency-compensation method used.

I measured intermodulation distortion (IMD) with a modified Heathkit IM-48, using 60Hz and 6kHz mixed 4:1. The meter residual in this case is about 0.06%. At 70W, the amplifier produces 0.4% IMD, a figure consistent with the THD measurements. In my experience, tube amplifiers usually have higher IMD than THD, sometimes much higher.

All distortion readings depend somewhat on the output-tube quiescent bias current, decreasing as the bias current increases. The measurements in *Figs. 3–5* were taken at 50mA per tube. Increasing the current to 60mA drops the THD at 70W and 20kHz from 0.75% to 0.62%. I experimented with these higher bias currents and decided I couldn't hear the difference, so I set the current back to 50mA to prolong tube life.

Distortion readings also depend somewhat on which individual tubes you use in the amplifier, suggesting that it is probably possible to select a set of tubes to give lower distortion. This is a small effect, however, and no brand of NOS tubes that I tried (RCA, GE, Raytheon) gave consistently lower distortion. The Sovtek 5881 gave consistently higher distortion and lower power output than the NOS tubes, as shown in *Fig. 5*, but in listening tests I couldn't hear the difference. I've also tried the new Svetlana 6L6GCs, which work well and sound as good as the others, but I haven't yet made distortion measurements with them.

SQUARE-WAVE PERFORMANCE

I also examined the amplifier's square-wave performance as a test of high-frequency stability. *Figure 6* is a plot of the amplifier's 10kHz square-wave response into a 5Ω resistive load, showing fast rise time with no ringing or overshoot. *Figure 7* shows the 10kHz square-wave response into 5Ω and 2.2µF. There is a small amount of overshoot (10%), followed by a half cycle of well-damped ringing, indicating good high-frequency stability.

Figure 8 is a plot of the amplifier being overdriven at 95W and 1kHz, showing symmetrical clipping. *Figure 9* is a plot of the amplifier being overdriven at 70W and 33Hz, showing output-transformer

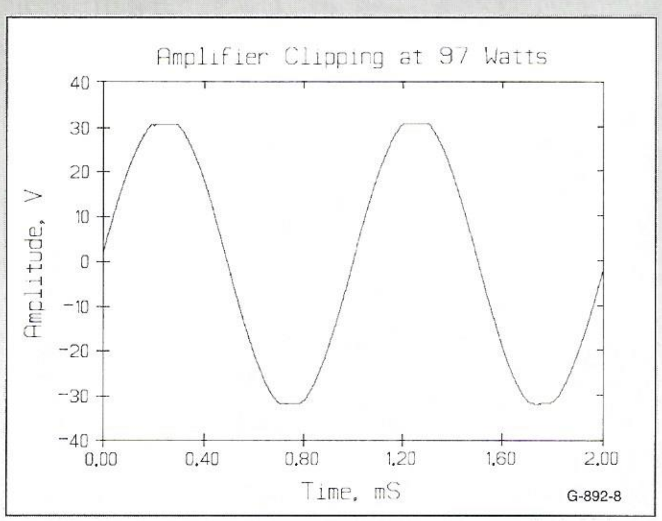

FIGURE 8: Amplifier clipping characteristic, 1kHz and 97W into 5Ω resistive load.

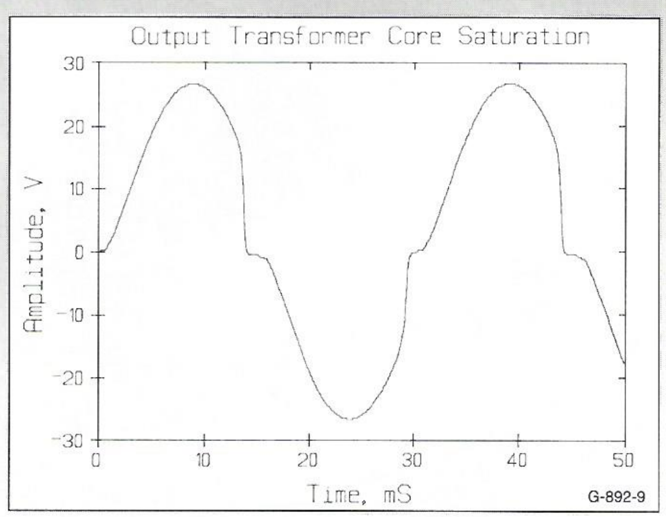

FIGURE 9: Output transformer core saturation at 70W and 33Hz into 5Ω resistive load. See text.

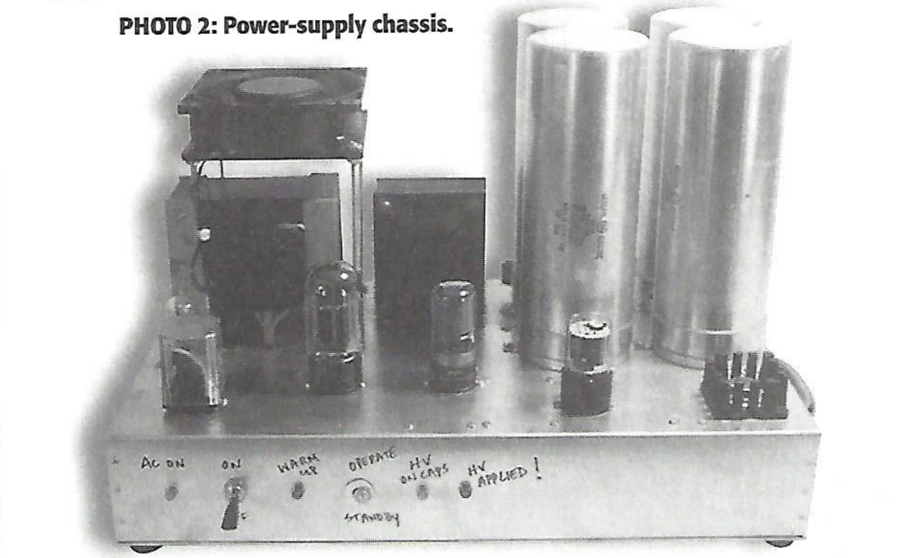

PHOTO 2: Power-supply chassis.

core saturation. The Plitron PAT4004 output transformers display the most abrupt onset of core saturation that I've ever seen. All transformers undergo core saturation at low frequencies, of course, but the PAT4004 goes from near-perfect performance to gross distortion within 1Hz! At 70W and 34Hz I measured THD of 0.18%, and at 33Hz I obtained the waveform shown in *Fig. 9*. At lower powers, the near-perfect performance extends to lower frequencies, as indicated in *Fig. 3*. This abrupt onset of core saturation may be a general characteristic of toroidal transformers.

I was at first concerned about what effect this core saturation would have on the amplifier's sound, since the gross distortion would be clearly audible. It turns out that the amplifier has excellent bass performance, as I discuss later in my listening comparisons. I suspect that most music has little power content below 30–40Hz, at least compared to the power in the upper bass or midrange. At any rate, I have been unable to detect core saturation on musical program material, either audibly or with my DSO.

To summarize the measurement results, I think the amplifier offers excellent performance. Although the distortion readings may seem high to those accustomed to the specifications of solid-state gear, the readings compare quite favorably to those of high-quality commercial tube amplifiers in the same power class, as reviewed recently in *Glass Audio*.[4,5] In making comparisons, note that some manufacturers (not those reviewed) advertise THD only at 1W and 1kHz, where it is low, rather than at higher powers or at the frequency extremes, where it is higher.

CONSTRUCTION AND SETUP

The Cover Photo shows the completed amplifier. The two stereo channels are built side by side on a 17″× 10″× 3″ aluminum chassis, and the power supply is built on a second identical chassis (*Photo 2*). I used a combination of point-to-point wiring and perforated board construction. The amplifier is sufficiently complicated that I recommend building it in sections and testing each section as you complete it. Perforated-board construction facilitates this by allowing you to test certain sections of the circuit outside the chassis on the bench. However, you could build the amplifier entirely with point-to-point wiring if you prefer.

The minimum test equipment required to build this amplifier is a digital multimeter (or VTVM) capable of accurate AC as well as DC measurements, and a sinusoidal signal generator. An oscilloscope is not absolutely required, but I highly recommended it because it simplifies checkout and troubleshooting.

I recommend building and testing the power supply first, beginning with the raw high-voltage supply. With a little planning, you can punch and drill the power-supply chassis before beginning any electronic assembly. You may need to modify the circuit slightly, depending on the transformer and filter capacitors you have available. The filter-capacitor bank should provide 1000–3000µF of capacitance at 600V DC, but you can obtain this with a variety of series or parallel connections.

Be aware that charged capacitance of this magnitude is dangerous and potentially lethal! With the bleeder resistors shown in the schematic, the capacitor bank can remain charged for 5–8 minutes after the power is switched off. Do not be tempted to discharge the capacitors by shorting across them, since the rapid release of energy will result in an enormous arc which may burn you or destroy components! You can test the power supply under load for brief periods using a 1250Ω, 200W resistor, which will draw 400mA (I made such a resistor by wiring ten 125Ω, 20W wirewounds in series). You should measure 505V ±20V with this load.

BUILDING THE REGULATORS

Once the raw HV supply is working, you can build the HV regulators and check them for the correct output voltages. The low voltage regulated supplies Ecc and Eaa2 are each built on their own perforated boards. You should mount the IC102 (LM338K) on a heatsink, as shown on the right-hand side of *Photo 2*, isolat-

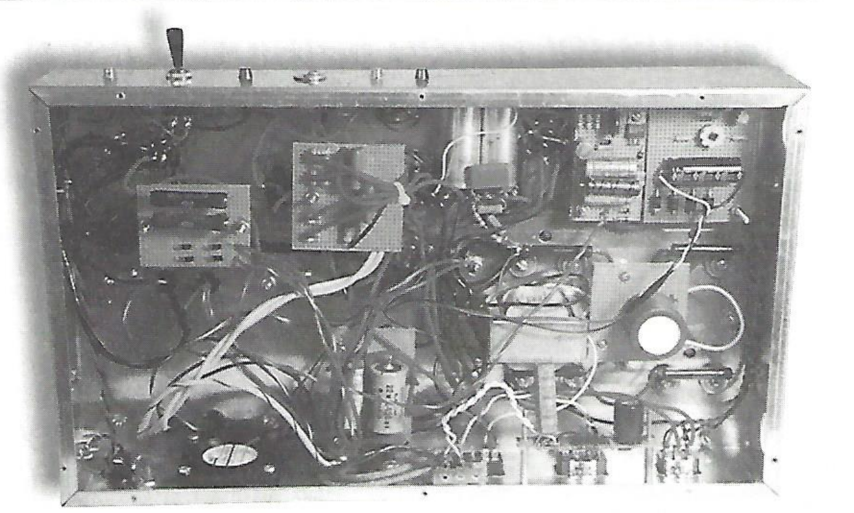
PHOTO 3: Underside of power-supply chassis.

ing it from the chassis with a thermally conductive washer or gasket.

Install the fans last, since they are a nuisance to work around if they are spinning. *Photo 3* shows the installation of FAN102 on an edge under the chassis, oriented to bring air into the chassis. Exhaust vents are shown on the right-hand side of *Photo 3*. As I mentioned, the power supply sits on a shelf in my cellar, where it can't be touched by accident. If it is in a room with children (or even adults), the exposed fans, hot tubes, and open vent holes represent a hazard, and you should install a screen cage.

Begin construction of the amplifier chassis by completing as much of the mechanical construction as possible. All grounds on the amplifier chassis are to a piece of 14-gauge solid wire connected to the chassis at a single point between the input jacks, as shown in *Photo 4*. The input jacks themselves are isolated from the chassis with an insulating washer, and the input grounds are connected to the grounded side of resistor R1 on the board containing tube V1.

COMPONENT ASSEMBLY

I assembled tubes V1 and V2 and their associated components (including C2–C4 and R12–R13) on a 3″ × 5″ piece of perforated board. Once they're assembled, you can power up the circuit (please be extremely careful when working with the HV applied!) and check for a plate voltage of about 120V on the 5879, a cathode voltage of about 125V on the 12AU7, and plate voltages around 280V on the 12AU7. If these are obtained, apply a small test signal (100mV RMS, 1kHz) to the input and check for a gain of 50–60 to both outputs of the phase inverter. You should adjust trimmer R47 at this point so the two outputs of the phase inverter are equal in magnitude.

You can also assemble tubes V3, V4, V5, and their associated components (including R32–R35 and C10–C13) on a 3″ by 5″ piece of board. For testing purposes, you can temporarily connect the unconnected ends of R33 and R35 to Ebb2, since you cannot yet connect them to the 6L6GC plates. When you power on the circuit, connect a voltmeter across R16 and adjust R20 for 4.55V.

Similarly, adjust R21 for 4.55V across R17. These two adjustments set the driver-stage idle current to 5mA. Under this condition, the plates of the 6SN7 should idle at about 225V, and the cathodes at about 65–70V. If you cannot achieve the correct 5mA bias current by adjusting R20 or R21, and there is no wiring error, try substituting a different MPF102 JFET. I've noticed that the variability in these JFETs can be very wide, and you may have an out-of-spec transistor.

When the idle currents are set, you can apply a 100mV test signal to the grid of one of the 6SN7 sections through a DC blocking capacitor. You should measure a gain of 10–15 to the cathodes of the corresponding 12AU7. Use a second DC blocking capacitor between the 12AU7 cathodes and your VTVM or oscilloscope. When you have verified the correct gains, disconnect R33 and R35 from Ebb2 and set the board aside.

WIRING

Finally, you can wire the output tubes and their associated circuitry, including the bias circuits, C6–C9, and the output transformer. Also wire the two circuit boards (per channel) into the larger amplifier circuit, connecting the inner feedback loops, but leaving the outer feedback loop temporarily open. Be careful to wire the amplifier exactly according to the schematic

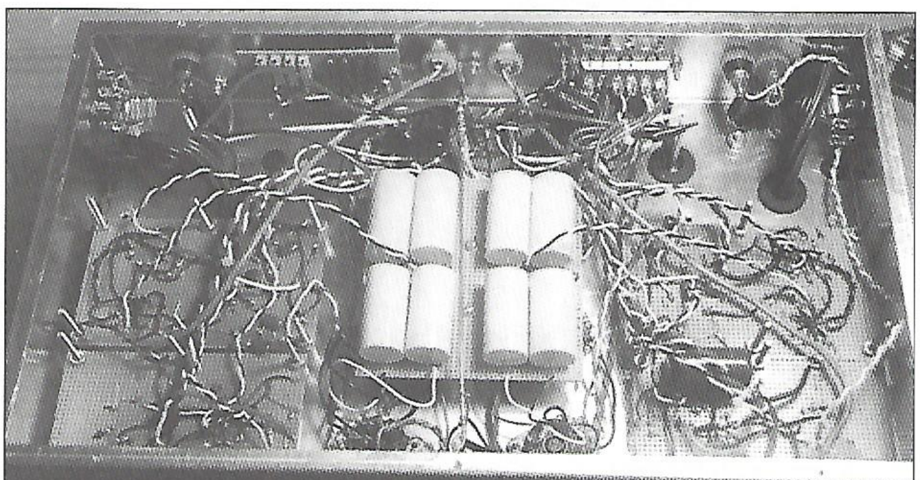

PHOTO 4: Underside of amplifier chassis.

in order to maintain correct phasing of the input and output. It is easy to invert the signal phase inadvertently when wiring the circuit boards or T1.

Apply power without inserting the output tubes and adjust R38–R39 and R42–R43 for –30V at the 6L6GC grids (pin 5 of the empty sockets). Then, with the power off, connect a 5Ω, 100W resistor to the speaker output, insert the tubes, and reapply power. If the amplifier oscillates with the outer feedback loop open, then the inner feedback loops must be out of phase or miswired. Adjust R38–R39 and R42–R43 for 50mA through the corresponding 6L6GC, equivalent to 2.35V across each of the 47Ω cathode resistors. With the 5Ω load resistor connected to the speaker output, apply a 50mV test signal and check for a gain of about 70. If you have an oscilloscope, check that the output is in phase with the input.

Complete the wiring by connecting the outer feedback loop. If the amplifier oscillates, then there is probably a phase reversal, which you can correct by reversing the primary leads of T1. Recheck the settings of R38–R39 and R42–R43 after an hour of operation. Also recheck the settings of R20–R21 and R47. If you have an oscilloscope, check the power output and frequency response. If you have distortion-measuring equipment, set R47 for minimum IM or THD at 70W.

LISTENING COMPARISONS AND CONCLUSIONS

I've compared this amplifier to my Williamson 20/20, to my 20W Class A tube amp, and also to a 14W single-ended amp using paralleled 2A3 triodes. My comparisons were not rigorous A/B tests in which I rapidly switched between the various amplifiers. Instead, I installed each amplifier into my system and listened to it over a period of days or weeks, trying to form an overall impression of their strengths and weaknesses.

I don't have golden ears, and I've always found it difficult to reliably distinguish similar amplifiers. In this case, the most obvious distinction between the amplifiers was on the basis of power. The new 70W amp gives the impression of being able to produce virtually any volume level I would wish with the Magneplanars, without apparent clipping or compression. I believe it is superior to the 20/20 in this respect.

One of the common weaknesses of tube amps is in the bass region, where the sound is not as clear as that of solid-state amps. This new 70W amp does not share that weakness; low notes on organ recordings are as clean as I've heard through the Magneplanars, equaling or even surpassing the performance of the 20/20.

I must admit that at lower power levels there was very little to distinguish between the four amplifiers. I did have the impression that the single-ended amp was less clear than the other three, but in all fairness, a 14W amp is severely overstressed by the relatively inefficient Maggies.

I'm pleased with my 70W 6L6GC amp. After almost three years, I've found nothing to fault in its sound or reliability. I hope some of you will be equally pleased. ❖

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