The C70 KT-88 Vacuum Tube Stereo Amplifier

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1. Introduction

Those who know me may be surprised to see an article on a vacuum tube amplifier written by me. Several years ago I needed a vacuum tube amplifier to compare with solid state designs for use in listening and measurement workshops that a few friends and I were presenting at the Rocky Mountain Audio Fest. I had built numerous tube amplifiers in the late '60's, so I decided to build one for this purpose using some updated technology and design approaches. I re-designed a 35 wpc amplifier that

Figure 1: The C70 Power Amplifier.

I had built in 1968, keeping little more than the chassis, output transformers, power transformer and power supply choke. This amplifier is the result of that effort.

Vacuum tube amplifiers are alive and well in the high-end audiophile community. Indeed, several people asked me if I had any material on vacuum tube amplifiers in my new book, "Designing Audio Power Amplifiers" [1]. Unfortunately, there was just no room at the time, and some will be distressed to know that I devoted 4 chapters to class D amplifiers instead.

In many ways the amplifier described here is a classic design, using KT88s in a class AB pentode arrangement. Why not ultra-linear? The transformers from the old amplifier I used did not have ultra-linear taps on them – simple as that. The amplifier is pictured in **Figure 1**.

This amplifier differs from many classic designs in some ways as well. The topology comprises two differential amplifier stages in tandem and avoids the use of an explicit phase splitter stage – the differential pairs implicitly provide the phase splitter function. Solid state current sources are used to supply the tail currents to these long-tailed pairs (LTPs). Providing the tail current source for the input stage from the -35V supply allows the input stage grids to be biased at 0V while having a high-impedance current source for the tail.

MOSFET power transistors are used to implement two pass regulators for supply of the screen, driver and input stage rail voltages. The output voltages for these regulators slowly track the main power

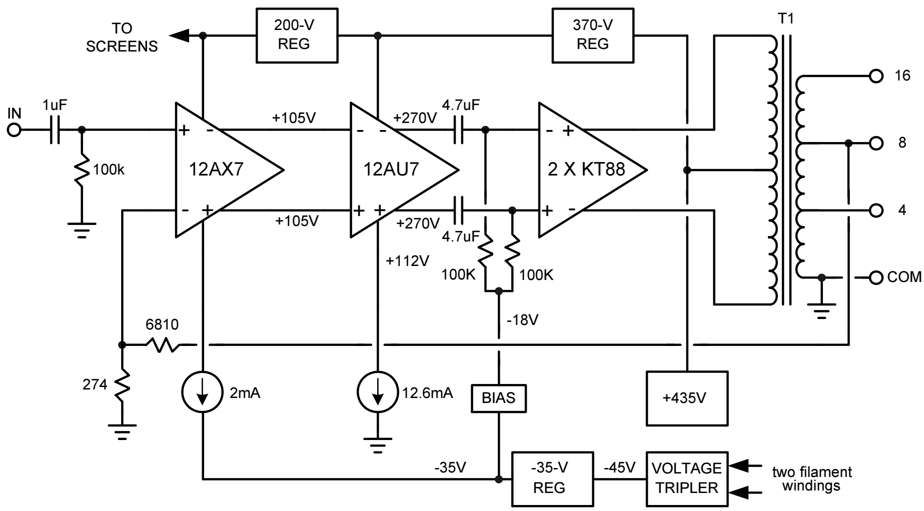

Figure 2: Simplified block diagram of the amplifier showing its differential signal path.

supply voltage, so technically they are more akin to capacitance multipliers. Regulated screen supplies in a Pentode design are especially advantageous, but rarely found in practice. Fixed bias is used in the output stage.

The design is "over-tubed", employing the large KT88s to produce only 35 watts using a 7200-ohm output transformer with a 435V power supply. In essence, the KT88s are loafing. This, in combination with the 7200-ohm output transformer's larger turns ratio (30:1), provides a higher output current capability than one would normally obtain. Tighter bass and better bass extension are among the advantages of this approach. The higher plate dissipation of the KT88 also allows a somewhat higher class AB1 output stage bias setting that reduces distortion.

Figure 2 is a block diagram showing the arrangement of the amplifier. The signal path is fully differential from input to output, and does not employ an explicit phase splitter. Instead, the 12AX7 (V1) input differential pair provides implicit phase inversion by having the input applied to one side and the negative feedback applied to the other side of its input. The use of a high impedance transistor current source (Q1) for the tail of the pair provides exceptional common mode rejection and phase splitting action, even out to high frequencies. The current source is connected to a –35V supply. There is no capacitor in the feedback shunt return path of the amplifier, and instead a DC balance adjustment is incorporated in the input stage cathode circuit. It is adjusted to achieve equal driver plate voltages in the presence of minor vacuum tube offset voltages.

The driver employs a 12AU7 differential amplifier (V2), which is direct coupled from the input differential pair. It employs a transistor tail current source (Q2) referenced to ground. This provides good common mode rejection and further enhances the balance and symmetry of the phase-split signals provided by the input stage.

There is only one coupling capacitor (per differential side) within the amplifier's feedback loop. It is in the forward path, coupling the differential driver outputs to the KT-88 grids. The presence of only a single capacitive coupling within the feedback loop greatly enhances low frequency stability, tightness and extension. The coupling capacitors are a relatively large 4.7 uF. Both AC balance and DC balance adjustments are provided for the KT-88 output stage.

The power supply provides 435 volts from an over-sized power transformer (actually a Sixties-vintage television power transformer), silicon rectifiers and 600 uF of capacitance with a choke in a pi configuration. The power supply is heavily bypassed and snubbed with smaller-value capacitors. A MOSFET voltage regulator implemented with Q3 provides 370V to the driver stage, while a second MOSFET voltage regulator (Q4) provides 200V to the input stage and the screens of the output tubes. A pair of filament windings connected in series provides -45V via a voltage tripler. This voltage is regulated to -35V for use by the input stage current source and the fixed bias circuit for the output stage.

The important features of the C70 are summarized as follows:

- 35 Watts per channel
- KT-88 output tubes, Class-AB pentode, fixed bias
- 435V power supply with 600 uF capacitance and choke
- Regulated screen, input stage, driver and fixed-bias supplies
- Fully balanced differential design with no explicit phase splitter
- Transistor current sources for the long-tailed pairs
- No electrolytic capacitors in the signal path
- Only one coupling capacitor stage within the feedback loop
- Extensive input, output and power supply RFI immunity enhancements
- 4 Hz to 45 kHz, +0/- 1 dB
- Medium feedback design (24dB) with damping factor of 23

The key components, the output transformers, are from the 1960's and are no longer available. However, the design can be made to work some modern output transformers with appropriate changes to the feedback compensation. In fact, a modern transformer that is compatible has been identified and will be discussed later. It is available from Dynakit® Vacuum Tube Audio Products.

2. Input and driver circuits

Figure 3 shows the input and driver circuits, consisting of two differential amplifier stages implemented with 12AX7 and 12AU7 devices.

The input pair is biased with a tail current of 2 mA provided by Q1 and the driver pair is biased with a tail current of 12.6 mA provided by Q2. Resistors R10 and R22 absorb some of the voltage drop and power dissipation to enable the use of TO-92 and TO-126 transistors for the current sources. The voltage reference for the input current source is an 8.2V Zener diode, while that for the driver current source is the +10V regulated power supply. Notice that the input stage tail is powered from a -35V power supply. The cathodes of the driver stage float at about 110V so as to permit DC coupling from the input pair to the driver pair.

Each half of the V1 input pair operates at a plate current of 1 mA with a transconductance of about 1700 uS. Plate resistance of V1 is about 60k on each side. Each half of the V2 driver pair operates at a plate current of 6.3 mA and a transconductance of about 2600 uS. Plate resistance of each half of V2 is about 7.5k. The low-frequency differential gain of the input stage is about 45, while that of the driver stage is about 12.

R4 and C2 form an input low pass filter at about 160 kHz. Negative feedback is established by R15 in series from the 8-ohm tap of the output transformer feeding into the shunt resistance R14. Many classic amplifiers take the feedback from the 16-ohm tap, but the distortions of output transformers

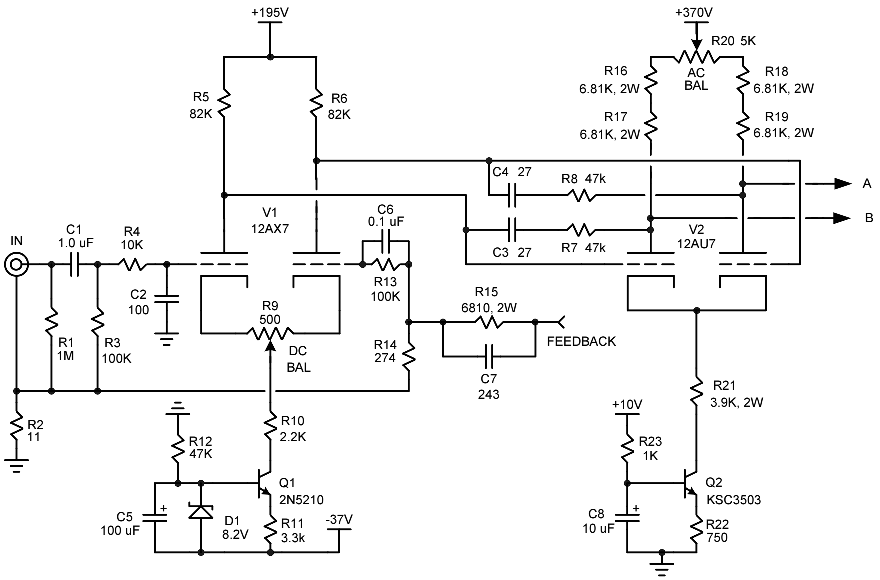

Figure 3: The DC-coupled differential input and driver stages of the amplifier

are such that lower distortion and more uniform frequency response result when the feedback is taken from the tap actually used to drive the load. R2 is employed to break ground loops at the input.

Potentiometer R9 provides adjustment of the DC balance of the input stage. It is adjusted to make the plates of V2A and V2B be at equal potentials. Potentiometer R20 is the AC balance adjustment for the output stage. In the absence of a more detailed adjustment procedure using a distortion analyzer, it is adjusted so that the AC signals at the plates of V2A and V2B are of equal amplitude.

The frequency compensation used with this amplifier is called differential Miller compensation. It is implemented by series R-C networks connected from grid to plate on V2. The advantage to this compensation as compared to conventional shunt compensation is that it encloses the driver in a local feedback loop that reduces driver distortion and driver output impedance. This is not unlike the way that Miller compensation is used in solid-state amplifiers. As compared with conventional shunt compensation at the plates of V1, the value of the Miller capacitors will be smaller by a factor of roughly G2+1 (where G2 is the voltage gain of the driver stage). Similarly, the value of the series resistors will be larger by about that same factor. C3/C4 and R7/R8 provide a pole-zero compensation frequency characteristic, while C7 provides some lead compensation in the feedback path.

The details of the negative feedback compensation in any vacuum tube amplifier depend on the characteristics of the output transformer, since it has a strong influence on the open-loop gain and phase of the amplifier. The transformers in this amplifier came from an amplifier that the author built in the 1960's and are of unknown origin and are certainly no longer available. For this reason, the feedback compensation used in this amplifier might not work with some modern output transformers. Altering negative feedback compensation to fit a different transformer can be a tricky process requiring substantial skill and good test equipment. Some general guidelines for frequency compensation of a VT amplifier are discussed below.

Overall negative feedback in the audio band is about 24 dB. Closed-loop gain of the amplifier to the 8-ohm tap is about 25. An input level of 670 mV rms produces 35 watts into an 8-ohm load.

3. Fixed bias vs. cathode bias

Many classic VT amplifiers use cathode bias in their class AB output stages. Cathode bias connects the pair of output tube cathodes to ground through a resistor that is bypassed. Sometimes separate bypassed cathode bias resistors are used. The control grids are referenced to ground and the voltage drop across the cathode bias resistor provides the necessary negative grid-cathode bias. This works fine in the quiescent state but causes the negative grid bias to increase as output power increases. This results in program-dependent bias. Moreover, the increase in bias with signal is given a time constant by the cathode bypass capacitor, usually an electrolytic. After a high-power interval, the bias may remain high briefly, under-biasing the amplifier for the subsequent softer interval, possibly resulting in class B operation with its attendant crossover distortion. Finally, cathode bias wastes HT voltage, as the useable amount of HT voltage is the actual HT voltage less the amount of the cathode bias.

Fixed bias eliminates the shortcomings of cathode bias, but incurs the increased cost of a separate negative voltage supply. Fixed bias also provides the opportunity for a DC bias balance pot to compensate for minor tube differences (even present in matched pairs).

4. Over-tubed amplifier design

Over-tubed amplifier design encompasses two things. First the primary impedance of the output transformer is greater than one normally uses for the given amount of power output. In other words, the turns ratio is larger. Secondly, larger output tubes, capable of conducting more current than normal for the given rated output power are used. The combined effect of these two practices makes for an amplifier that is capable of much higher current delivery than normal. In a sense, this is analogous to high-current solid state amplifiers. This is not unlike an automobile with a higher gear ratio in the drive train, resulting in higher torque capability. At the same time, the over-tubed amplifier requires a higher-voltage power supply to achieve a given output power. This is akin to the engine in the car needing to be capable of higher RPMs. The over-tubed amplifier is also capable of delivering enough current to overcome some reduced transformer inductance as mild saturation occurs, allowing for more bass to be delivered through a given-sized-core transformer.

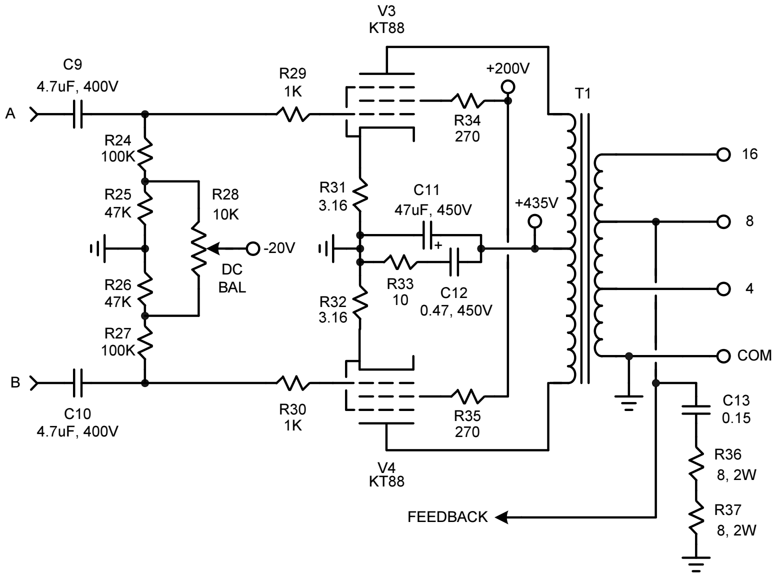

Figure 4: The KT-88 output stage operating in pentode mode.

5. Power output stage

The output stage is shown in **Figure 4.** It comprises a pair of KT-88 output tubes operating in Class AB1 pentode mode with fixed bias of about -18V. Grid drive is about 40V p-p at full output. R31 and R32 are for measuring cathode current. The output transformer has a 7200-ohm primary, higher than that normally employed with KT-88 output tubes. The higher turns ratio (30:1 to the 8-ohm tap) provides a lower output impedance and greater current drive capability at the expense of reduced power capability into an 8-ohm load for a given supply voltage.

The amplifier output circuit includes a series R-C Zobel network. This is common for solid state amplifiers, but is not often seen in vacuum tube amplifiers. This improves high-frequency stability and immunity to RFI. Potentiometer R28 controls DC balance. The fixed bias voltage is controlled by R67 in **Figure 7**. Bias is set to about 63 mA per output tube. This corresponds to about 200 mV across each of the 3.16-ohm cathode resistors. Plate dissipation for each KT88 is thus about 27W.

6. Quiescent power dissipation

One reason why tube amplifiers may sound better is their relative absence of crossover distortion. Although any class AB design will in principle generate crossover distortion, the higher relative quiescent bias current combined with the softer cutoff characteristic of vacuum tubes will mitigate it. Consider the ratio of quiescent output stage power dissipation to rated power for this vacuum tube amplifier to that for typical BJT and MOSFET power amplifiers. At 63mA per tube and a 435-V HT

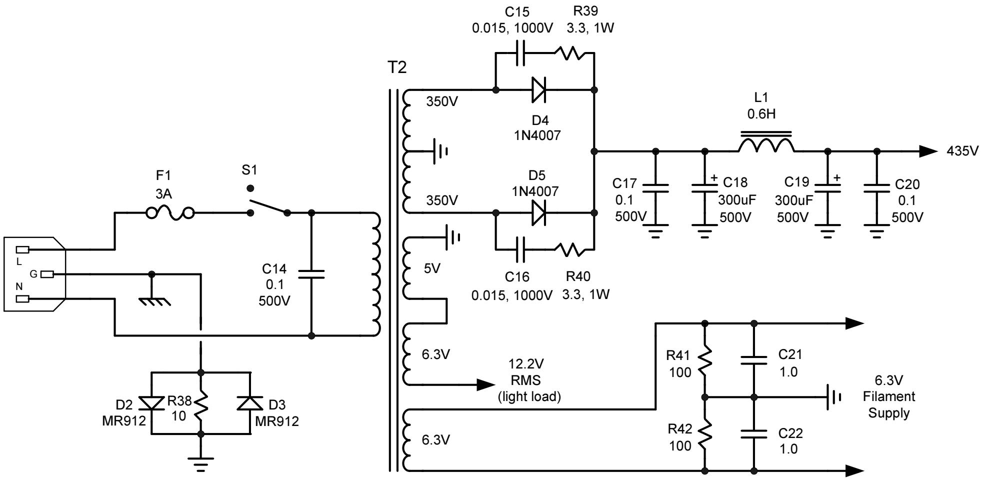

Figure 5: The main power supply delivers 435V and includes substantial reservoir capacitance and filtering.

supply, this amplifier dissipates 54W at idle and is rated at 35W, for a quiescent-to-rated power ratio of 154%. In contrast, a 35W BJT amplifier biased at 115mA (RE=0.22) with 38V rails dissipates 9W for a ratio of 26%. A 35W MOSFET power amplifier biased at 150mA with 42V rails dissipates 13W for a ratio of 37%. The typical vacuum tube amplifier is operating closer to class A than the typical solid state amplifier. In this amplifier, this can be verified by looking at the voltage across the cathode resistors as a function of power level with an oscilloscope. The cathode currents will remain sinusoidal to a larger power level.

7. Main power supply

The main high-voltage power supply is shown in **Figure 5**. This was built with an old television power transformer that included several secondary filament windings.

Diodes D2 and D3 and R38 connect the mains ground to the chassis ground in such a way that ground loops are broken, but where any potential difference is limited to one diode drop.

The high voltage supply is a conventional full-wave design employing a center-tapped secondary. The power transformer delivers 350-0-350 VAC at the idle load current of 330 mA. C15, C16, R39 and R40 provide snubbing for the silicon rectifiers. The main supply includes a 0.6H choke that is surrounded by 300 uF on each side. The loaded output voltage of the supply is about 435V with ripple of about 50mV p-p. The choke can be replaced by a 15-ohm resistor with ripple increasing to 1.2V p-p with a negligible effect on amplifier hum due to the use of regulated HT supplies for everything except the output stage plate supply.

Two filament windings are connected in series to provide about 12.2 V rms for use by low-voltage power supplies described later. These two filament windings are lightly loaded, so they produce a bit more than the rated 6.3V + 5V. C21 and C22 provide additional RFI filtering for the filament supply.

8. Screen and low-voltage supply

The plate voltages for the input stage and driver are quasi-regulated, as is the screen voltage for the output stage. The voltages will slowly track changes in the mains voltage, and the arrangement is somewhat of a cross between a regulator and a capacitance multiplier. These supplies are very quiet and free of ripple and EMI. The HT regulator circuit employs IRFP240 power MOSFET devices as pass transistors and is shown in **Figure 6**. The power MOSFETs dissipate some power and are mounted to the aluminum chassis of the power amplifier for heat sinking. The V_{DSS} rating of 200V for these devices is a bit marginal and MOSFETs with a higher V_{DSS} would be desirable.

The two necessary voltages, +370V and +200V, are established by resistive voltage dividers feeding the gates of the MOSFET pass transistors. These voltages are filtered, but largely un-regulated, in that they are intended to track changes in the available main power supply voltage. If I had it to do again, I would fixed-regulate these supplies (at greater cost) so that output stage bias would be less affected by line voltage changes causing screen grid voltage changes.

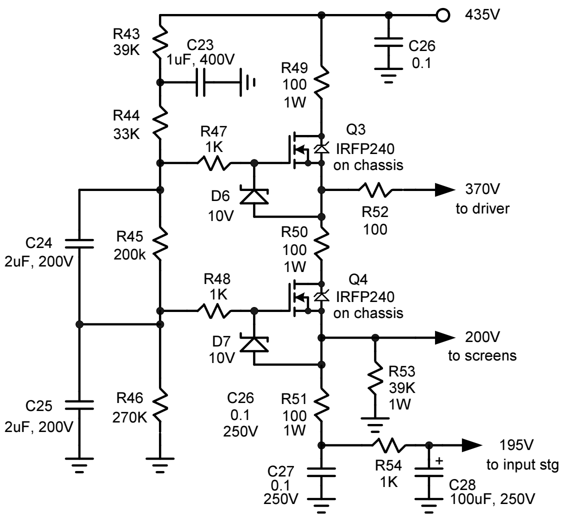

Figure 6: MOSFET pass regulators provide clean input stage, driver stage and screen voltages at low impedances.

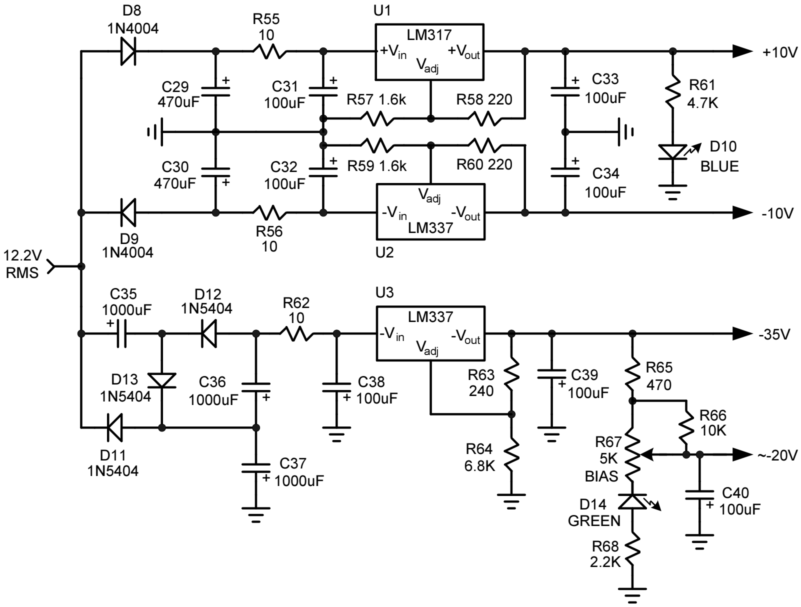

Figure 7: Regulated low-voltage supply provides -35V for input stage current source and output stage fixed bias.

Gate stopper resistors R47 and R48 provide stability against parasitic oscillations, while Zeners D6 and D7 protect the gates from voltage swings that could harm them. Resistors R49 and R50 also help assure high-frequency stability.

Resistor R53 guarantees a minimum amount of current flow in the pass regulators in the absence of load while the vacuum tubes are warming up. R54 and C28 provide some additional filtering for the input stage plate supply. It is important to minimize the influence of mains voltage transients on the input and driver supplies because in most vacuum tube amplifiers such changes can be propagated to the grids of the output tubes, disturbing their bias.

9. Fixed bias and +/-10-volt supplies

The fixed bias supply for the output stage grids is shown in **Figure 7**. This circuit applies 12.2V rms from the main power transformer to a 3-diode voltage tripler to provide about –47 volts DC. This is then regulated by an LM337 to a constant –35 volts. This voltage is then applied to an adjustable voltage divider to provide the approximate –18 volts used for the fixed bias. R67 is the bias adjusting potentiometer. The –35-volt supply is also used for the input stage tail current source. One can optionally use separate pot arrangements to be sure that both channels are operating at the same bias current.

Plus and minus 10-volt power supplies are also provided in the amplifier. This circuitry is also shown in Figure 7. It is a simple half-wave-rectified power supply whose output is regulated by LM317 and LM337 IC regulators. The +10-volt supply is used as a reference for the driver tail current source and for the Blue power-on LED. The –10-volt supply is not used. These supplies were included for possible use by auxiliary circuits.

10. The output transformer

The output transformer is the heart and soul of a vacuum tube amplifier. A very simplified model of the output transformer I used is shown in **Figure 8**. It is a good quality unit of modest size from the 1960’s.

The primary inductance can introduce a low-frequency roll-off. Imperfect magnetic coupling between the primary and secondary results in what is called leakage inductance [2], which is effectively in series with the primary. It can be determined by measuring the inductance of the primary when the secondary is shorted. Leakage inductance can introduce high-frequency roll-off and excess phase lag. It can be reduced by interleaving the transformer windings to improve magnetic coupling. Interleaving can be done among segments of a given winding and/or among different windings. Without interleaving, the two halves of a center-tapped primary will not necessarily have the same transmission characteristics to the secondary, for example.

The effective primary capacitance results from the proximity of different portions of the windings. It includes the reflected secondary winding capacitances scaled by the inverse square of the turns ratio. Primary capacitance causes high-frequency roll-off. It is generally increased by interleaving of the windings. In reality, the leakage inductances and primary capacitances are distributed along with the winding resistances, and can actually result in some transmission-line-like behavior.

Using a different output transformer can affect amplifier stability in a number of ways. First of all, a transformer with a different turns ratio can alter the open-loop gain. This can have a big effect since transformer primary impedances typically range from 8000 ohms down to 2000 ohms, and the turns ratio goes as the square root of the rated impedance. Theoretically this could result in a 6 dB range

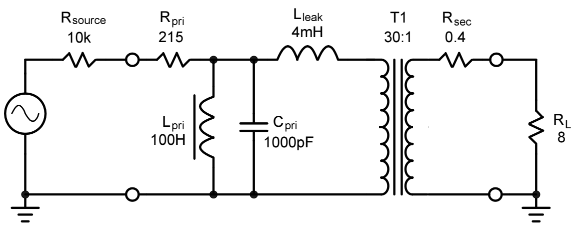

Figure 8: A very simplified model of the output transformer.

of open-loop gain. However, in a given amplifier the reasonable range of useable output transformer impedances might be from 4000 ohms to 8000 ohms (as in this design), resulting in a turns ratio range of less than 1.4:1, corresponding to a 3 dB range in open loop gain at most.

The effective primary capacitance can have a significant influence on the high-frequency portion of the open-loop gain characteristic. This capacitance forms a pole with the net impedance at the primary, consisting of the output tube plate impedance and the impedance of the load reflected by the square of the turns ratio. In this amplifier, with a 30:1 turns ratio, an 8-ohm load, and a 10k plate to plate output tube resistance, the effective primary impedance is about 4200 ohms. With a 1000pF primary capacitance, the pole lies at 4.2k and 1000pF, or 38kHz. If the transformer is replaced with one whose primary capacitance is only 500pF, this pole will rise to 76kHz, increasing the HF open-loop gain and possibly causing instability. In principle, this could be compensated by adding a 600pF capacitor across the primary, but the capacitor would have to have a voltage rating of 1000V or more. The effective primary capacitance is strongly influenced by the amount of interleaving in the transformer windings.

11. Frequency compensation

Compensating a vacuum tube amplifier can be a difficult and frustrating process, due to the presence of the output transformer in the loop. They are distributed LCR devices with a complex frequency response and numerous opportunities for resonances. Here I share a few pragmatic guidelines.

The first step in compensation should be to characterize the output transformer. Here we assume that the feedback will be taken from the 8-ohm tap and that the transformer will be loaded at the 8-ohm tap with an 8-ohm load. Feed one-half of the primary with a sinusoid through a 10k resistor and load the 8-ohm tap with an 8-ohm resistor. Measure the output at the 8-ohm tap. The 10k resistor approximates the plate resistance of a pentode, and is not critical. The load impedance seen across one-half of the primary will be 7200 ohms/4 = 1800 ohms for the transformer used in the C70.

We do not want to see any frequency response peaks that rise much above the 0-dB frequency response reference at 1 kHz. We also do not want to see any frequency response dips. In general, we like to see a well-behaved approximation to a 6 dB/octave roll-off caused by the effective primary capacitance of the transformer working against the net impedance seen at the primary. A second pole will often define the higher-frequency roll-off. Take note of where the higher-frequency roll-off of the transformer begins. This may be in the neighborhood of 200 kHz. Choose your amplifier gain crossover frequency to lie about an octave below this frequency. The gain crossover frequency for the C70 amplifier lies below 100 kHz.

As a check, repeat this test using the other half of the primary. You might be surprised at how big the difference is for some output transformers. The difference will tend to be smaller for transformers with better interleaving of the primaries. You will also often see differences in the DC winding resistance from either end of the primary to the center tap. You may also want to look for differences in the measured frequency response at the 16-ohm and 4-ohm taps while the transformer is still loaded at the 8-ohm tap.

Sometimes the frequency response of the output transformer can be tamed and smoothed a bit by incorporating a heavier-than-normal Zobel network at the output. Try loading the 8-ohm tap with a Zobel network sufficiently heavy to eliminate any peaks in the transformer frequency response. A good starting point is a network with about 16 ohms resistance and a capacitor that is large enough to attenuate any peaks. Do not use a capacitor that is larger than necessary, nor a resistor that is smaller than necessary, as power dissipation in the resistor must be considered and overly heavy loading at high frequencies will increase HF distortion.

Measure the mid-band loop gain of the amplifier at 1 kHz. This is done by disconnecting the feedback network from the output transformer and feeding it from an audio oscillator while measuring the signal amplitude at the output tap from which the feedback is normally taken. Choose the compensation Miller capacitance so that the loop gain will fall to unity at the chosen gain crossover frequency as if the transformer response were flat. Verify the chosen roll-off by measuring loop gain at 20kHz. In this case it should be 14 dB for a 100kHz gain crossover frequency.

The resistance in series with the compensation capacitance provides a zero that partially compensates for the initial phase lag and gain roll-off of the output transformer. Choose its starting value to provide a zero in the vicinity of the -3-dB response of the transformer as measured above.

Close the loop and check the frequency response of the amplifier. Hopefully it will not oscillate. It may show a significant gain peak in the frequency range near the chosen gain crossover frequency, perhaps as much as 6 dB. Adjust the resistor to find the value that results in the smallest peak, hopefully 3 dB or less. If you cannot get the peak down to 3 dB, you should consider choosing a lower gain crossover frequency and start over.

Now add the lead capacitor in the feedback loop. Start with a small value that forms a corner with the feedback resistor in the vicinity of the second pole of the transformer frequency response. This should decrease the peak in the frequency response. Increase the value of the lead capacitor until there is no peak, and perhaps as much as 3 dB of attenuation at the chosen gain crossover frequency. Don't use a capacitor that is larger than necessary. The response should show a reasonably smooth roll-off as frequency is increased with no local peaks, up to frequencies more than an octave above the gain crossover frequency.

Check the gain margin by shunting the feedback resistor with one of equal value. Also shunt the lead capacitor by one of equal value. If the amplifier does not oscillate, you have at least 6 dB of gain margin. It is also useful to check the response to a 10-kHz square wave under these conditions to see how close to oscillation the ringing is.

Check the square wave response with the input filter capacitor C2 disconnected. There should be minimal ringing and perhaps only a small amount of overshoot, maybe 10%. The initial overshoot peak and one much smaller one after that is acceptable. There should be no sign of a small-amplitude higher-frequency ringing superimposed on this waveform. The compensation elements above can be adjusted to improve the square wave response, but always go back and check the closed-loop frequency response and gain margin to make sure that they have not be unduly compromised.

This process sounds straightforward, but it can take a lot of time, experimentation and iteration. Moreover, it is only one pragmatic way to arrive at satisfactory frequency compensation for the amplifier.

12. Low-frequency negative feedback

Most vacuum tube amplifiers suffer from potential low-frequency instability due to the use of interstage coupling capacitors and the phase shift that they add to the overall feedback loop. This can lead to motor-boating. Just as with frequency compensation in amplifiers at high frequencies, the amount of negative feedback at low frequencies is limited by stability issues in these vacuum tube amplifiers. This inevitably leads to decreasing amounts of negative feedback at low frequencies, exacerbating low-frequency distortion. A vacuum tube amplifier whose LF response falls 3 dB at 10 Hz probably has only 6 dB of negative feedback at 20 Hz. The very low LF cutoff of the C70 is made possible by the use of only one coupling capacitor in the loop. This means that the amount of negative feedback available at 20 Hz is still nearly its full amount, leading to much lower distortion at low frequencies. As a side benefit, the amplifier frequency response extends to unusually low frequencies.

13. Performance

Figure 9 shows the frequency response of the amplifier. Frequency response is 3 Hz to 45 kHz +0/-1 dB. The –3dB frequencies are 1.8Hz and 75kHz. The low-frequency response is exceptional, being down only 1 dB at 3Hz. This extended low-frequency response is made possible by the design topology that requires only one coupling capacitor in the feedback loop. The high-frequency response is limited by the output transformer's primary capacitance and leakage inductance along with the necessary feedback frequency compensation when using this output transformer.

The A-weighted noise is about 112 dB below full output and -97 dB with respect to 2.83V output (1W, 8 ohms). Hum is 100uV at the output, which, referred to 2.83V, is 89 dB down. It is predominantly 60Hz.

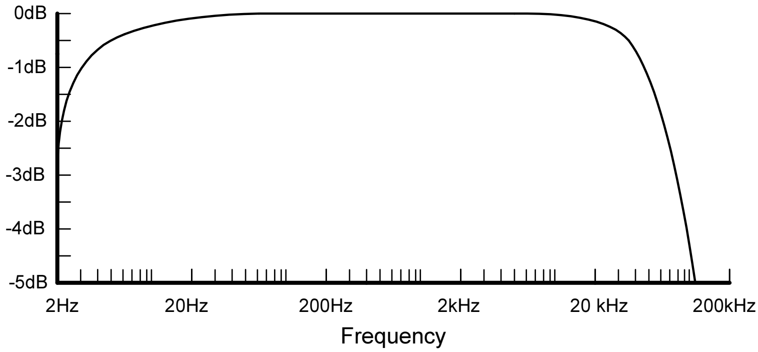

Figure 9: Frequency response of the amplifier extends to unusually low frequencies.

Figure 10 shows THD+N vs. power into an 8-ohm load at 50 Hz, 1 kHz and 20 kHz. Output power at 1% THD+N at 1 kHz into an 8-ohm load is about 35 watts. **Figure 11** shows THD+N vs. frequency at 1 watt, 8 watts and 28 watts into an 8-ohm load. Distortion is below 0.05% from 20Hz to 20kHz at 1 watt output. At 8 watts, THD+N is below 0.2% from 25Hz to 20kHz. As a reference, THD+N at 1 watt for the legendary McIntosh 275 is shown with the dashed curve [3]. The '275 is the gold-standard of classic vacuum tube audio power amplifiers. Putting aside the fact that the '275 is rated at twice the power, this amplifier compares very favorably to the '275 in distortion performance.

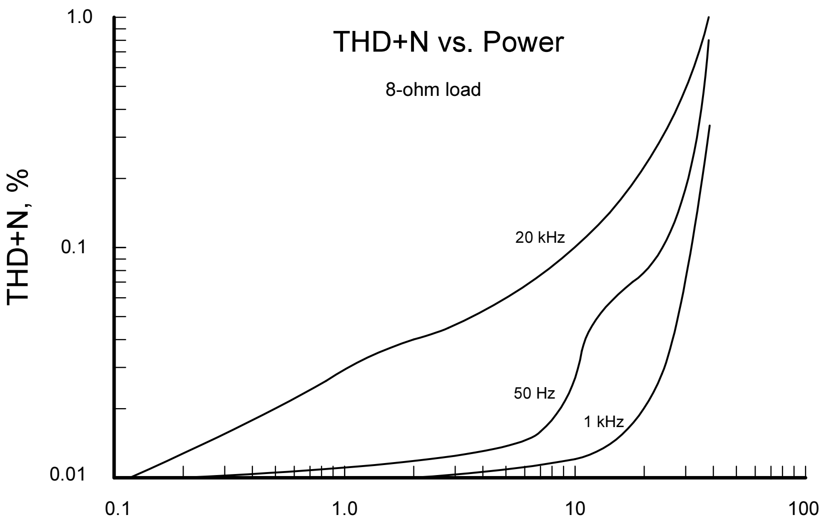

Figure 10: THD+N vs. power at 50Hz, 1kHz and 20kHz driving 8 ohms.

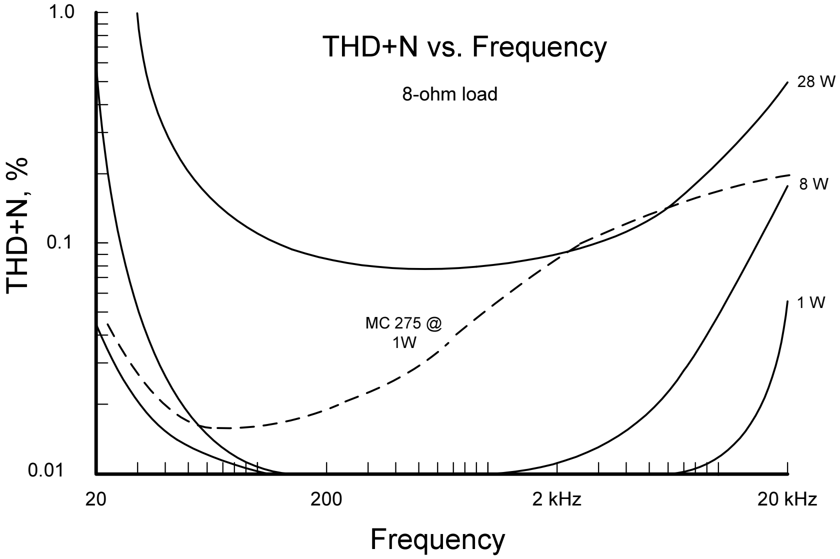

Figure 11: THD+N vs. Frequency at 1W, 8W and 28W driving 8 ohms. Note comparison with McIntosh MC275 at 1W [3].

Notice the substantial increase in THD+N at 8w and 28w as frequency decreases below 50Hz. This is evidence of output transformer core saturation. Larger output transformers would have improved the THD+N substantially at 20 Hz.

14. Listening tests

As one who has been designing solid-state amplifiers since the 1970's I must say that this VT amplifier sounded very good. Indeed, it sounded much better than it had a right to sound based on its measured power and distortion performance. This is often said of vacuum tube amplifiers. The C70 was demonstrated at a "bottle-head" meeting and met with praise. It was auditioned at the 2006 RMAF and HE2007 audio shows where it was used in listening and measurement workshops. There it was compared against a 200+ wpc solid-state amplifier and was also very well-received. To my ears, the C70 is solid across the full audio band and not the least bit sloppy or mushy like some VT amplifiers. The bass is surprisingly authoritative, but not at the expense of the midrange and highs.

15. Adjusting the output stage bias

In general, too little output stage bias will result in crossover distortion, and increased bias will reduce distortion. However, too much bias may cause excessive plate dissipation and actually reduce maximum output power. Once DC and AC balance have been optimized at a reasonable bias current setting, one can adjust the bias and observe distortion and maximum output power versus bias current. Distortion at low frequencies should also be checked as a function of bias. DC balance and AC balance should be trimmed after making any changes to the bias current. This amplifier has been adjusted for an output stage bias of 63mA per tube. Always let the amplifier warm up for at least one-half hour before making any adjustments.

Rated plate dissipation for the KT88 is 42W. With the existing bias of 63mA and a plate supply of 435V, plate dissipation at idle is 27W. This is fairly conservative and yet allows a healthy idle bias for low distortion. This is another advantage of the over-tubed design.

16. Adjusting DC balance

If the two output tubes do not operate at the same bias current, the difference in current can act to increase distortion and output transformer core saturation effects. It is always important to employ matched pairs of output tubes, but modest differences can occur between even matched pairs. This means that the two tubes in a pair will not operate at the same current even if their grid voltages are the same. DC balance should be adjusted so that the voltage drop across each of the cathode resistors is the same. VT amplifiers that do not have a DC balance adjustment will often be operating sub-optimally.

17. Adjusting the AC balance

The AC balance adjustment takes care of driver signal balance and differences in output tube transconductance. Once again, a "matched pair" of output tubes cannot be assumed to be perfectly matched. Moreover, even when operated at the same bias current, the two tubes in a pair may not have exactly the same transconductance.

AC balance strongly influences second harmonic distortion. Although it is sometimes said that vacuum tube amplifiers produce more second harmonic distortion, the symmetrical design of push-pull vacuum tube amplifiers suggests that second harmonic distortion should actually be fairly low. Tube amplifiers that do not benefit from DC and AC balance adjustments are likely to suffer from increased amounts of second harmonic distortion. For those who like second harmonic distortion, the AC balance pot provides an opportunity to adjust it to taste.

AC balance is best adjusted using a distortion analyzer. The amplifier should be operated at 1/3 power at 1 kHz and the AC balance should be adjusted for minimal distortion. I have seen proper adjustment of AC balance reduce distortion by a factor of two in otherwise balanced amplifiers using matched output pairs. The optimum position will sometimes be a bit different for different frequencies, and the choice of frequency at which to set the optimum is a compromise.

18. An alternative output transformer

As mentioned earlier, this amplifier was built with output transformers from the 1960's that are unfortunately no longer available. However, I have tested the amplifier with a transformer that is in current manufacture and is readily obtainable. It is the A-470 from Dynakit Corporation. They sell it as a drop-in equivalent of the 35-watt transformer used in the original Dynaco ST-70. Dynaco has long had a great reputation for its output transformers, traceable back to the Acrosound output transformers developed by David Hafler. These modern versions are wound to the same specifications and measure very well. They are specified as having a 4300-ohm primary, which is quite a bit different from the transformers I used. However, measured performance of the amplifier is very close to the original, and bass response is a bit better at higher power levels because the A-470 transformers have a bit more iron. Listening tests that I conducted after the amplifier was fitted with the A-470 output transformers confirmed the high sound quality of these transformers. Optimal compensation for these transformers is slightly different, with C3/C4 at 22pF and R7/R8 at 100k. The A-470 transformers are available from:

Dynakit Inc.
55 Walman Ave
Clifton, NJ 07011
www.dynakitparts.com
dynakitparts@aol.com
(973) 340-1695

The Dynakit PA-060 power transformer, also designed for the Dynaco ST-70, is probably a compatible choice for the C70, but I have not evaluated it.

19. Conclusion

The topology and design techniques discussed here should be useful to those who wish to design and build modern vacuum tube amplifiers. Moreover, this amplifier demonstrates the level of performance that can be achieved in a classic push-pull power amplifier. For more audio information visit my site at www.cordellaudio.com.

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