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[Second Edition.]

PATENT SPECIFICATION

Application Date: Sept. 7, 1940. No. 13972/40 **564,250**

Complete Specification Left: July 17, 1941.

Complete Specification Accepted: Sept. 20, 1944.

PROVISIONAL SPECIFICATION

Improvements in or relating to Thermionic Valve Amplifier Circuit Arrangements.

I, ERIC LAWRENCE CASLING WHITE, a British subject, of 7, Vine Lane, Hillingdon, Middlesex, do hereby declare the nature of this invention to be as follows:—

This invention relates to thermionic valve circuit arrangements known as cathode follower circuits. Circuits employing a cathode follower type of valve are now well known as providing a useful output circuit having a low value of output resistance approximately equal to $\frac{1}{g}$, where g is the mutual conductance of the valve. The valve is thus capable of driving an output load of low impedance with only a small loss. The output amplitude is restricted, however, being dependent fundamentally on the current swing available from the valve before over-loading occurs.

The required current swing is known from the output voltage and the nature of the load impedance. A simple cathode follower circuit must then be designed to be capable of passing without over-heating, a mean current at least equal to one-half the required swing and to be capable of passing a peak current at least equal to that current swing without passing grid current, although the flow of grid current may in some cases be permitted.

It is assumed that for maximum economy the arrangements for feeding the mean D.C. to the cathode have a high A.C. impedance compared with the load, for example, such arrangements may consist of a high inductance choke coil or a high resistance feed from a source of considerable negative potential with respect to the mean grid potential.

The object of the present invention is to provide a circuit arrangement which will economise in the mean D.C. feed current required for a cathode follower circuit, or alternatively, for a given size of valve will permit a considerable increase of output. A further object is to reduce the output resistance to less than $\frac{1}{g}$ as defined above.

According to the present invention in a circuit arrangement including a cathode follower valve, in addition to the connection of the cathode to the load, further connections are provided between said cathode and a second valve which is so coupled to the cathode follower valve that the outputs from the two valves are in the correct phase for connecting to the same point of the load. In particular forms of circuit embodying the invention, the cathode of the cathode follower valve is connected not only to the load but also to the anode of the second valve, the grid of which is coupled to the anode of the cathode follower valve, a suitable anode load resistance being provided.

In operation of such an arrangement both valves, in the absence of input signals, may be arranged to pass a quite small current, sufficient merely to bring them to a reasonably linear part of their characteristic curves. On an increase in the input voltage the cathode follower valve is caused to pass more current and raises the output potential. Its anode potential drops, thus reducing the current of the second valve and adding a further contribution to the output. Owing to the small initial current this contribution is limited and for large positive excursions of the input voltage the load current is supplied almost entirely by the cathode follower valve.

On a decrease in the input voltage the reverse action occurs. It is arranged that reduction of the anode current of the cathode follower valve to zero raises its anode potential sufficiently to cause the second valve to pass maximum current, which, like that of the cathode follower valve may be many times the steady feed for zero input signal.

The arrangement outlined is analogous to a class B push-pull amplifier, but in

which the output from the two valves are both in the correct phase for connecting to the same point of the load, the other end of which is effectively earthed; the arrangement not requiring a reversing transformer.

The invention may find various applications but it is particularly useful in television systems. Two applications to television arrangements will be described by way of example with reference to Figures 1 and 2 of the accompanying drawings.

Figure 1 shows the invention applied to an output stage feeding a low impedance concentric cable represented by the resistance 1, fairly heavy currents at a low voltage being required in a case of this kind. It is convenient in the case illustrated, to have a direct connection from the cathode of the cathode follower valve 2 to the load 1 and to arrange that some intermediate point of the signal range, for example the black level, should correspond to zero current in the load. The picture signals may then be positive and will be applied mainly by current from the cathode follower valve 2 and the synchronising signals may be negative and supplied mainly by current from the second valve 3. The cathode follower valve is never entirely switched off so that the output impedance is always low. $\left(\frac{1}{g}\right)$. It will be seen that in applying the invention, the cathode of the valve 2 is connected not only to the load 1 but also to the anode of the valve 3, the control grid of which is coupled to the anode of the valve 2 through a condenser 4.

The arrangement of Figure 2 shows an output stage feeding capacity load 5, such as the modulating electrode of a cathode ray tube, where a large voltage swing may be required. In such an application, the arrangement is particularly economical as regards current, as high currents are only drawn when current sudden changes of input signal voltage occur, necessitating sudden bursts of current to charge or discharge the capacity 5. As in the arrangement of Figure 1, it will be seen in Figure 2 that the cathode of the cathode follower valve 2 is connected directly to the load and to the anode of the valve 3, the control grid of which is coupled through condenser 4 to the anode of the valve 2.

Because in the arrangement of Figure 2, the output swing is large, pentode type valves are employed, but in Figure 1, in which the voltage output swing is low, triode valves are suitable.

Apart from the application of the invention to television purposes, it will be understood that other applications may be found and, for example, the invention is of particular use in any systems involving signals consisting largely of sharp pulses.

Dated this 5th day of September, 1940.
F. W. CACKETT,
Chartered Patent Agent.

COMPLETE SPECIFICATION

Improvements in or relating to Thermionic Valve Amplifier Circuit Arrangements.

I, Eric Lawrence Casling White, a British subject, of 7, Vine Lane, Hillingdon, Middlesex, do hereby declare the nature of this invention and in what manner the same is to be performed, to be particularly described and ascertained in and by the following statement :—

This invention relates to cathode follower circuit arrangements, that is to say, circuit arrangements including a valve having a load connected in its cathode circuit in such a manner as to provide negative feedback into its grid circuit so that the apparent output impedance of the cathode circuit of said valve as seen by said load is low and of the order of the inverse of the mutual conductance of said valve.

One of the objects of the present invention is to provide a circuit arrangement employing a pair of valves, one of which is arranged in a cathode follower circuit, for affording what is effectively a push-pull output without the necessity of employing a push-pull output transformer such as is usually required with push-pull circuits.

According to the invention, a circuit arrangement is provided comprising a thermionic valve and a load arranged in a cathode follower circuit, means for applying signals to the control electrode of said valve, a further valve having its output circuit arranged to feed current to said load in opposite sense to the current fed thereto by said first mentioned valve, said further valve also being arranged to be controlled by said signals, the arrangement being such that if the current flowing through said load under the control of said first mentioned valve is caused to

increase on the application of said signals the current flowing through said load under the control of said further valve is caused to decrease and *vice versa*.

The output power which can be delivered to the output by a cathode follower circuit arrangement is limited by the available cathode current swing of the valve and it is usual to choose a valve capable of passing, without overheating, a mean current at least equal to one-half of the required output current swing and a peak current at least equal to the required current swing. One of the disadvantages of employing a valve in this manner is that the mean power output for a given size of valve is limited to a fraction of the maximum power output of the valve. A further disadvantage is that the mean current drawn by said valve is substanial and the arrangement is therefore uneconomical. By employing the arrangement according to the invention, the available power output of the valve can be more fully utilised. The mean current can also be reduced by so biassing said valves that in the absence of signals the anode current of each valve is small. The biasses are, however, preferably arranged to be sufficient to ensure that the valves will operate on reasonably straight portions of their characteristic curves. By so biassing the valves it can be arranged that for large positive excursions of input signal voltages the load current is almost entirely supplied via one of said valves and for large negative excursions the load current is almost entirely supplied via the other valve. By employing two valves biassed in the above manner, the mean current can be substantially reduced compared with the mean current which would be required for a single valve operating in a cathode follower circuit providing the same power output.

Preferably, the control electrode of said further valve is coupled to an impedance associated with the valve in the cathode follower circuit in such a manner that if signals are applied to said valve signals are also applied to said further valve but in opposite phase to the signals applied to said valve. The impedance is preferably provided in the anode circuit of said valve. An impedance may be connected in the cathode circuit of said further valve so as to provide negative feedback to linearise said further valve.

In order that the invention may be clearly understood and readily carried into effect it will now be more fully described with reference to the drawings accompanying the Provisional Specification, in which :—

Figure 1 illustrates a circuit arrangement according to one embodiment of the invention, and

Figure 2 illustrates a further embodiment of the invention.

Figure 1 of the drawings illustrates the invention as applied to an output stage of an amplifier feeding a load in the form of a low impedance concentric cable represented by the resistance 1. The triode valve 2 has the resistance 1 connected in its cathode circuit so as to provide negative feedback in its grid circuit, and the arrangement functions as a cathode follower circuit in which the impedance seen by the load is low and is substantially equal to the inverse mutual conductance of the valve 2. The cathode of the valve 2 is connected to the anode of a further valve 3, the control electrode of which is coupled to the resistance shown in the anode circuit of the valve 2 via the coupling capacity 4 and is also connected to the cathode of the valve 3 via the leak resistance shown. The cathode of valve 3 is connected via the source of anode current represented conventionally by a battery to the negative terminal of the source of anode current for the valve 2 to which the resistance 1 is also connected.

Thus, when signals are applied to the input terminals shown in the figure, the anode current of valve 2 is varied and the voltage changes across the resistance in the anode circuit of valve 2 are communicated via the coupling capacity 4 to the control electrode of valve 3. The signals so applied to valve 3 are, however, in opposite phase to those applied to valve 2. It will be appreciated from the circuit illustrated that the current fed by the valve 2 through the load 1 is in opposite sense to the current fed through the load by the valve 3. When the control electrode of the valve 2 is made positive on the application of the signals, the anode current of the valve 2 increases and the anode potential of the valve 2 decreases and this decrease of anode potential is applied via the resistance-capacity coupling to the control electrode of the valve 3 so that the anode current of the valve 3 is reduced. On the application of a signal causing the control electrode of the valve 2 to become less positive, the anode current of the valve 2 diminishes the increase in anode potential which result is communicated to the control electrode of the valve 3 so that the anode current of this valve increases.

The bias potentials applied to the control electrodes of the valves 2 and 3 are so chosen that in the absence of input signal voltage the anode current of each valve is small but sufficient to ensure that the valves will operate on reasonably straight portions of their characteristic curves. It

can therefore be arranged that for large positive excursions of the input signal voltage the current flowing through the load 1 is almost entirely supplied via the valve 2 and for large negative excursions of input signal voltage the load current is also entirely supplied via the valve 3. The biasses should be so chosen in relation to the signal voltages that if the anode current of the valve 2 were reduced to zero the anode potential of the valve 2 would be raised sufficiently to cause the valve 3 to pass maximum current. In operation, however, the bias potential applied to the valve 2 should be so chosen that the anode current of this valve is never entirely reduced to zero so that the output impedance always remains low. The output impedance of the two valves combined as seen by the load 1 can, if desired, be arranged to be less than the inverse of the mutual conductance of the valve 2.

If the signal voltages applied to the valve 2 comprise a television signal of the usual waveform consisting of picture signals and synchronising signals in different amplitude ranges, it can be arranged that the picture signal output is supplied mainly by current from the cathode follower valve 2 and the synchronising signal output is supplied mainly by current from the valve 3.

The circuit shown in Figure 1 thus provides what is effectively a push-pull output without the necessity of employing a push-pull output transformer and enables a substantial economy to be effected compared with a cathode follower circuit having a single valve producing the same power output.

The arrangement of Figure 2 shows an output stage feeding a capacity load 5, such as the modulating electrode of a cathode ray tube, which is of low impedance at high signal frequencies. As in the arrangement of Figure 1, it will be seen in Figure 2 that the cathode of the valve 2 in the cathode follower circuit is connected directly to the load and to the anode of the valve 3, the control grid of which is coupled through the condenser 4 and leak resistance to the anode resistance of the valve 2. A resistance is shown connected in the cathode lead of the valve 3, the value of the resistance being selected to provide negative feedback for the valve 3 such that it will operate substantially linearly. Since the output voltage swing is likely to be large, pentode type valves are employed instead of triodes which are shown in the arrangement of Figure 1, in which the voltage output swing is low. The operation of the arrangement is in substance similar to that already described with reference to Figure 1.

This application of the invention permits great economy in anode current, as high currents are only drawn when sudden changes of input signal voltages occur necessitating sudden bursts of current to charge or discharge the capacity 5.

Although the invention has been described with reference to output stages delivering television signals to low impedance loads, it will be understood that other kinds of signals may be used and the invention is particularly useful when the signals consist largely of sharp pulses.

Having now particularly described and ascertained the nature of my said invention and in what manner the same is to be performed, I declare that what I claim is :—

1. A circuit arrangement comprising a thermionic valve and a load arranged in a cathode follower circuit, means for applying signals to the control electrode of said valve, a further valve having its output circuit arranged to feed current to said load but in opposite sense to the current fed thereto by said first mentioned valve, said further valve also being arranged to be controlled by said signals, the arrangement being such that if the current flowing through said load under the control of said first mentioned valve is caused to increase on the application of said signals the current flowing through said load under the control of said further valve is caused to decrease and *vice versa*.

2. A thermionic valve circuit according to Claim 1, wherein said valves are so biassed that in the absence of signals the anode current of each valve is small.

3. A circuit arrangement according to Claim 1 or 2, wherein the control electrode of said further valve is coupled to an impedance associated with said first mentioned valve, preferably an impedance in the anode circuit of said valve, in such a manner that if signals are applied to said first mentioned valve signals are applied in opposite phase to said further valve.

4. A circuit arrangement according to Claim 1, 2 or 3, wherein an impedance is connected in the cathode circuit of said further valve so as to provide negative feedback to linearise said further valve.

5. A circuit arrangement substantially as described with reference to Figure 1 or 2 of the drawings accompanying the Provisional Specification.

Dated this 16th day of July, 1941.

F. W. CACKETT,
Chartered Patent Agent.

2nd Edition

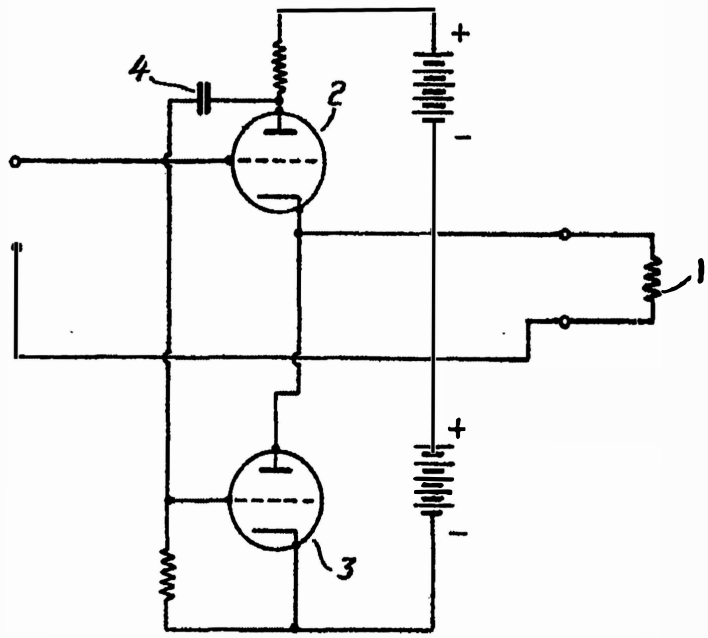


Fig. 1.

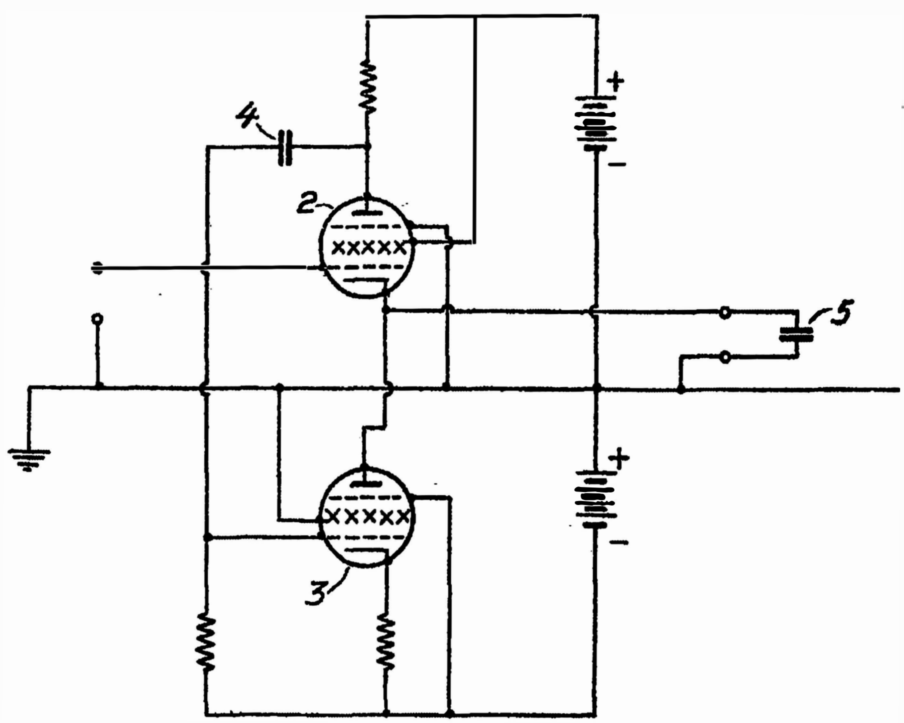
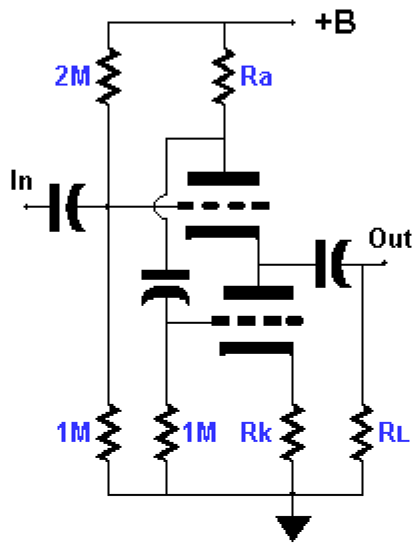


Fig. 2.

[This Drawing is a reproduction of the Original on a reduced scale.]



White Cathode Follower

The White Cathode Follower

Mr. White's improvement on the Cathode Follower was to create a buffer that boasted a much lower output impedance and the ability to sink as well as source current, i.e. push-pull operation. The lower output impedance results from the use of a feedback loop from the plate resistor to the bottom tube and the use of two tubes allows the buffer actively to draw current in both directions.

Because of the increased complexity of the circuit, the math is much more complicated than that of a simple Cathode Follower:

As an example, given a setup that consists of a 6DJ8 with a bypassed cathode resistor of 200 ohms and a 10k plate resistor, the results are

Gain = 0.97
Zo = 3.44 ohms
PSRR = -65 dB.

$$\text{Gain} = \frac{mu^2 + murp/Ra}{(mu^2 + mu + 1) + (mu+2)rp/Ra}$$

$$Zo = 1 / \left[(1 + mu)/(rp + Ra) + \frac{1 + mu(mu + 1)/(1 + rp/Ra)}{((mu + 1)Rk + rp)} \right]$$

$$\text{PSRR} = \frac{2rp + (2mu + 2)Rk}{[2rp+(2mu+2)Rk+Ra+murp+(mu^2+mu)Rk]\ [(Ra+rp)/(mu+1)+rp+(mu+1)Rk)/(muRa)]}$$

If this seems too good to be true, that's because it *is* too good to be true. Yes, the gain is almost unity and the Zo is amazingly low, yet the circuit cannot deliver very much current into a low impedance load, such as a Grado headphone, 32 ohms. Imagine a car with 340 horsepower, yet which could only do 10 miles per hour. Surprisingly, if we try to output more than a few millivolts into the 32 ohm load, we will overdrive the circuit, as we will break out of Class A operation.

Here is what happens in detail. Any variation in the current flowing though the top triode will produce a variation in the voltage developed across the plate resistor. In turn, this voltage will be transmitted to the bottom triode's grid, which can only see a few positive volts before it is driven into positive grid voltage or, if the voltage swings negatively, it is completely turned off. The greater the value of the plate resistor, the easier it is to overdrive the bottom triode, as a smaller amount of current is needed to develop a large voltage change across the plate resistor.

On the other hand, if we make the plate resistor smaller in value, we gain dynamic headroom, but lose the stellar specifications. In fact, if we set the plate resistor to zero ohms, we end up with a classic Cathode Follower with an active load, i.e. the bottom triode. Of course, if the load we wish to drive is not a punishingly low 32 ohms, the headroom issue is much less of an issue. But if the load is a high impedance one, such as a 100k potentiometer, then we must ask: Why do we need to use a super low output impedance buffer?

Optimal White Cathode Follower

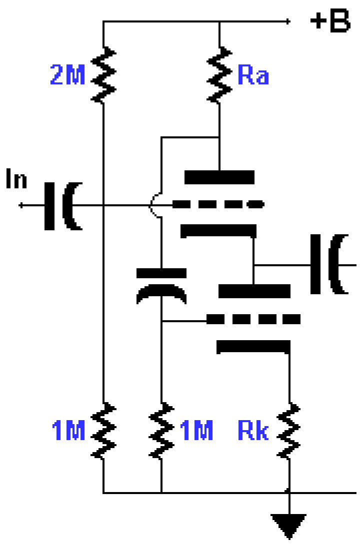
We found that too large a value plate resistor limits the potential output current from this buffer and that too low a value reduces the buffer's specifications. So the question is what would be the optimal value for a given load and a given desired out voltage swing?

This was the question I had asked myself, when I was disappointed by the results of an experiment wherein I had built a White Cathode Follower with the aforesaid tube and resistor values for driving a much more reasonable load: the Sennheiser headphones, which have an impedance of 300 ohms. After only a few millivolts, clipping occurred. My expectation was that the circuit should be able to deliver the idle current of at least 10 mA into this load, if not almost 20 mA, which would conform to classic Class A, push-pull amplifier standards. I then replaced the plate resistor with a 10k potentiometer with its center tab connected to one of its outside tabs, which allowed for easy adjustment of the plate resistor value.

After adjusting the potentiometer, I found the optimal value according to the trace on the oscilloscope to be 100 ohms. The lowness of the value surprised me. I then wondered what the optimal value would be for the 32 ohm load represented by the Grado headphones. Even more surprising was that the same 100 ohm plate resistor value yielded the best performance into the 32 ohms, in spite of this load being 10 times lower in value than the previous load. Moving to the other extreme, I replaced the 32 ohm resistor with a 3k resistor and retested. The 100 ohm plate resistor value once again made for the biggest and most symmetrical voltage swings. After some mathematical introspection, everything made perfect sense to me.

For any push-pull tube amplifier to work well, there most be an almost identical signal presented to each tube. (The signals must differ in phase.) In this circuit, if the top triode sees an increase in its grid-to-cathode voltage, then the bottom triode must see an equal decrease in its grid-to-cathode voltage. How do we ensure equal drive voltages for top and bottom triodes?

Let us start our analysis with the severest load possible, not Grado headphone, but 0 ohms, in other words, a dead short to ground via a large valued capacitor.



White Cathode Follower with a shorted output

The top triode now functions as a Grounded Cathode amplifier and does see the bottom triode at all. The amount of current flowing from ground into the capacitor then into the cathode of the top triode is given by the formula:

Ip = VgGm´,

where

Gm´ = (mu + 1) / (Ra + rp).

Now as the bottom triode current flow is governed by the top triode's current flow into the the plate resistor, the amount of current flowing from the bottom triode's plate into the capacitor is given by the formula:

Ip = VgGm

where Gm is the transconductance and

Gm = mu / rp.

By rearranging the formulas for current we get Vg = Ip / Gm´ for the top triode and Vg = Ip / Gm for the bottom triode. Obviously, the only way that the two grid voltages can match is if Gm´ = Gm. Expanding this formula out yeilds:

(mu + 1) / (Ra + rp) = mu / rp,

which when we solve for Ra becomes

(Ra + rp) / rp = (mu + 1) / mu

Ra = rp(mu + 1) / mu - rp

Ra = (rpmu) / mu + rp / mu -rp

Ra = rp / mu

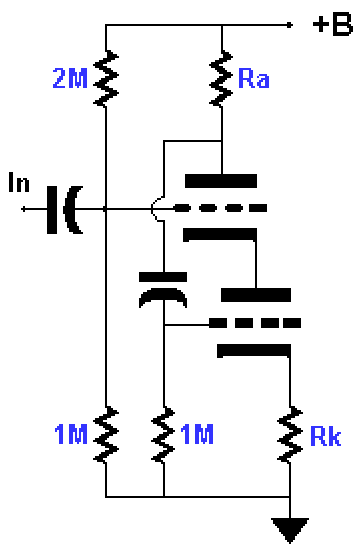
and as rp / mu = 1 / Gm

Ra = 1 / Gm.

Thus, the only way the Vg of the top triode can equal Vg of the bottom triode is if the plate resistor equals the inverse of the transconductance of the triodes being used. (The test to put any tube circuit equation through is to try the equation with a 6AS7 and then with a 12AX7 to check the equation for absurdities.)

What happens if we chose to start with infinite ohms as a load instead of 0 ohms. The answer is the same, the optimal value for the plate resistor is the reciprocal of the Gm of the triodes used, or what is the same quantity, rp/mu.

White Cathode Follower with an infinite impedance output load



Now let us step back and look at what is happening with this circuit in broad terms. Without an external load the rp of the bottom triode will define the sole load impedance for the top triode, remember we had defined an infinitely high impedance load. Since the gain of this circuit is less than unity, the cathode voltage will slightly lag the grid's and this gap is the change in the grid-to-cathode voltage that will prompt a change in current flow through both the top triode and plate resistor, which in turn will give rise to a change in voltage across that resistor, which will then be relayed to the bottom triode's grid. We need to ensure that that bottom tube receives an identical grid-to-cathode voltage signal as the top tube.

The math can become quite thick here, but if we think abstractly enough, it will not be too difficult to follow. We know that if the top triode sees a +1 volt pulse at its grid, its cathode will follow to some degree less than +1 volt. Whatever this outcome may be, we will refer to it as "Vg." Now Vg/rp equals the increase current (Ip) flow through the entire circuit, as all components are in current series with each other. Ip times the plate resistor (Ra) equals the voltage pulse that the bottom triode sees, which times the Gm of the bottom triode will equal Ip, if the right value of Ra has been chosen. Thus,

VgRa/rpmu/rp = Vg/rp,

which when we solve for Ra equals:

muVgRa/rp² = Vg/rp

muRa/rp = 1

muRa = rp

Ra = rp/mu.

Okay, what if we choose a load impedance somewhere between zero and infinity, say, 10k. Same result, Ra = rp/Gm. In this case, the load impedance is in parallel with the rp of the bottom triode. So Vg/(rp||R_L) equals the increase current (Ip) flow through the top triode and IpRa equals the pulse voltage to the bottom triode. In this case, like the one with a shorted output, we have true Class A output current swing capability, so as the bottom tube approaches cutoff, the top tube's current conduction will near twice its idle value. And, of course, vice versa for negative input voltage swings. Thus,

VgRa/(rp||R_L)mu/rp = Vg/(rp||R_L),

which when we solve for Ra equals:

VgRa/(rp||R_L)mu/rp = Vg/(rp||R_L)

VgmuRa/rp = Vg

muRa/rp = 1

muRa = rp

Ra = rp/mu.

Optimization and Zo

We can use the stock, long, complex equation for output impedance for the White Cathode Follower or we can realize that we

stipulated that Gm´ = Gm as a condition of satisfaction in the quest for the optimally valued Ra, and we found that Gm´ = (mu + 1)/(Ra + rp). In effect, what we have actually done by specifying the correct value for Ra is to balance the push-pull aspect of the circuit, which includes each triode offering the same output impedance to the load. Consequently,

Zo = 1 / 2Gm,

or

Zo = rp / 2mu.

Conclusion

We find once again that we cannot get something for nothing: spectacularly low output impedance came at the price of a disappointingly low input overload voltage and a miniscule output current ability. But what we did get, when we gave the White Cathode Follower the optimal plate resistor value to work with, was a buffer circuit twice as good as a textbook Cathode Follower: half the output impedance and a symmetrical output current swing with twice the output current swing than a single triode Cathode Follower.

Optimal White cathode followers revisited

I coined the phrase "optimal White cathode follower" to describe a method to find the optimal value for a White cathode follower's plate resistor, Ra. The optimal value is the one that yields the largest, most symmetrical voltage and current swing from both top and bottom triode, in other words, an optimally adjusted push-pull follower. In my article on follower circuits, I determined that the optimal plate resistor equaled the inverse of the triode's transconductance, or rp/mu. Thus, for example, a triode with a transconductance of 10mA per volt would require a plate resistor equal to 1/0.01, or 100 ohms.

About a year later Alex Cavalli e-mailed me, pointing out that the Ra = 1/gm was close, but not exactly correct, as far as SPICE simulations and empirical testing revealed. So, I reevaluated my mathematical reasoning and I found that he was right, that I had missed a term. I wrote back to him explaining that I found the error and I had fixed the optimal-Ra formula. His reply was that he too had come up with the true optimal-Ra formula and that he planed on revealing it in an article he planned on writing. I left the topic with him and I never posted the new formula. We never shared formulas, so I don't know if we agree and I don't know where his article appeared or if it did appear at all. Thus, this might be old news, but the formula for finding the optimal plate resistor value is:

Ra = (rp + 2RL)/mu,

where RL equals the load impedance. From a quick inspection, we see that the lower the load impedance, the closer the formula comes to:

Ra = rp/mu

On the other hand, when the load impedance is as much as 300 ohms and the rp

as low as 280 ohms, as it is with the Sennheiser HD-650 headphones and a 6AS7, then the adjustment is fairly large. The 6AS7's mu of 2 and rp of 280 ohms yields 140 ohms in the simple inverse of gm calculation, but they yield 370 ohms in the new optimal-Ra formula.

So how important is this fine-tuning? If you are using a 6AS7, very important; if you are using a 12AX7, very little, as the 12AX7's 80,000-ohm rp swamps out the 300-ohm load.

Nonetheless, as much as I am a fan of increased exactitude, I am also opposed to the practice of false exactitude. For example, some poor tube fancier, armed with the new optimal-Ra formula and a tube manual, looks 6DJ8 and he finds that the 6DJ8 holds an rp of 2,640 and a mu of 33.

RATINGS (Design Center Values—Each Section)

Plate Supply Voltage (Ib = 0 Ma)	550 Volts	Max.
Plate Voltage[2]	130 Volts	Max.
Plate Dissipation	1.8 Watts	Max.
Cathode Current	25 Ma	Max.
Negative Grid Voltage	50 Volts	Max.
Grid Circuit Resistance	1.0 Megohm	Max.

CHARACTERISTICS

Class A1 Amplifier (Each Section)

Plate Voltage	90 Volts
Grid Voltage	−1.3 Volts
Plate Current	15 Ma
Transconductance	12,500 μmhos
Amplification Factor	33
Equivalent Noise Resistance	300 Ohms

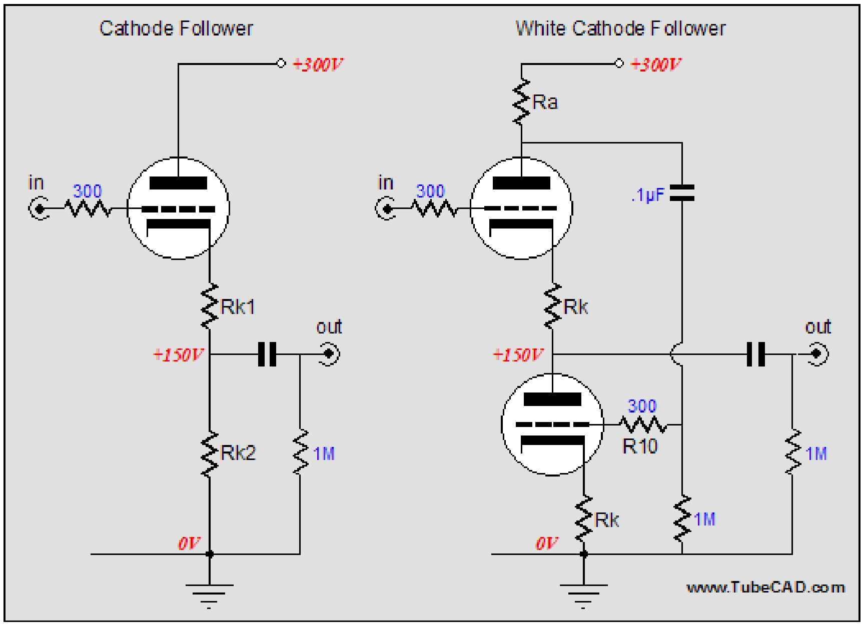
Hoping to drive 300-ohm headphones, he then calculates that Ra should equal 98.181818181818181818181818181818, which he then tries to find. Aside from the ridiculous precision, this resistor value is fundamentally ludicrous, as the specified tube characteristics are only correct at one cathode-to-plate voltage and cathode current, otherwise they are way off. What? How can that be? The tube manual's list of characteristics is about as accurate as your weight and height was in high school is today. Plate resistance and transconductance vary. Amplification factor comes the closest to being a true constant, but only because the other two vary in a related fashion. (In fact, mu is the least concrete characteristic of the three.)

So what good is the optimal-Ra formula—or any other tube circuit formula? Well, if nothing else, the formulas get you close to where you should be. Even 98.1818... is much better than the old assumption that the higher the plate resistor value, the better, so a constant-current source would have to prove best. And if someone just happened to build a 6DJ8-based White cathode follower with a 90-volt cathode-to-plate differential and 15mA idle current, as the tube manual specified, the 98.1818... value would be right on the mark. On the other hand, if the 6DJ8 sees a 125-volt cathode-to-plate differential and a 10mA idle current, where the mu is 29.7 and the rp is 3,030 ohms, and where the load is 300 ohms, the better value would be 122 ohms, with 120 ohms being close enough for all but the most anal-retentive of solder slingers. (If 32-ohm headphones are the intended load, with the same current-voltage values, then Ra should equal 104 ohms, which isn't that far off from 122 ohms.)

Improved followers (line stage and headphone amplifier emphasis)

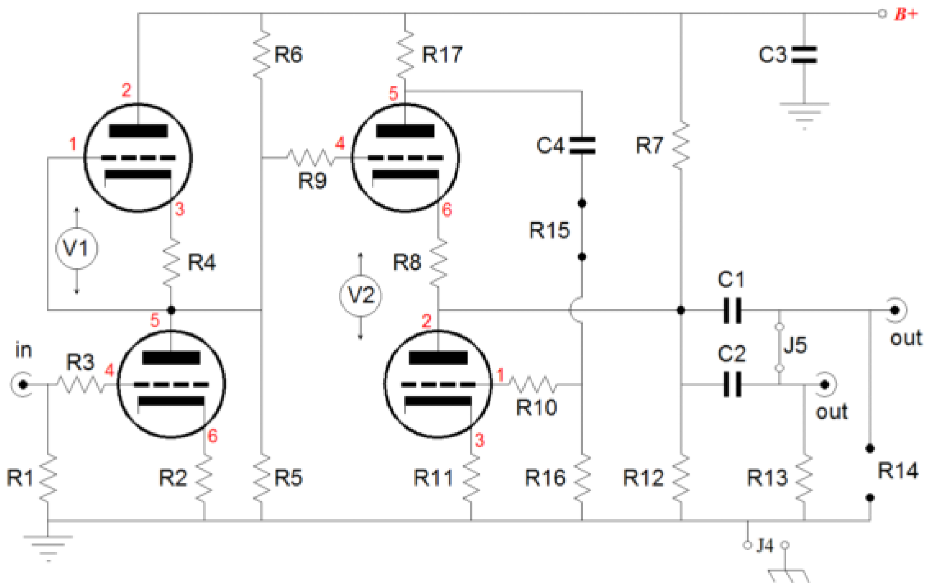
I don't like to attach my cathode followers naked, preferring to use a small-

valued cathode resistor as a buffer instead. This extra resistor greatly troubles many readers; isn't less more? No, not always; sometimes more is, indeed, more. The added cathode resistor further linearizes the cathode follower and buffers it from excessive capacitance. In fact, the Aikido amplifier uses this technique as part of its ear-pleasing arsenal. In the schematic below, we see both a cathode follower and a White cathode follower with extra cathode resistors.



Actually, in keeping with the first law of engineering ("Nothing of value comes without a cost"), this extra resistor's added benefits must be paid for with a higher output impedance and slightly less voltage and current swing. Whereas a standard 6SN7-based cathode follower would offer an output impedance of 400 ohms, the extra-resistor modified cathode follower can easily present twice that impedance. When driving interconnects, this increase is negligible; but it becomes significant when driving low-impedance headphones. Yet this is what I am about to recommend, i.e. using un-bypassed cathode resistors in all cathode followers, including the White cathode follower output stage. Why? The un-bypassed cathode resistors improve the performance of a line-stage amplifier and headphone amplifier to such a high a degree that the accompanying higher output impedance must be both paid for and accepted.

True, Grado headphone listeners are not going to like this modification, but those who listen to Sennheiser HD-580, HD-600, HD-650 and the higher-impedance AKG headphones are going to be well pleased. I am. I own a pair of Sennheiser HD-650s and they are supremely revealing of associated equipment, whereas the Grado SR-225s that I also own are much more forgiving, which is a virtue with 99% of audio equipment. (No, I haven't heard the new Grado GS100 reference headphones but I would love to give them a test drive.) And those who just want a supremely fine line-stage amplifier will find this new option a welcome increase in flexibility.



In the above schematic, we see the complete Aikido line-stage/headphone amplifier, with its White cathode follower with extra un-bypassed cathode resistors. The schematic refers to the new octal mono PCBs, but it applies equally to the 9-pin versions, save for the pin numbers (text in red). The critical resistor is Ra, the plate resistor of the White-cathode-follower output stage. The revised formula, given above, was based on the assumption that the bottommost cathode resistor would be bypassed or that fixed bias would be used for the bottom triode and that resistor R8 between triodes wouldn't be used. So how does the formula perform, given these changes to the topology? It must be modified, but the modification is simple enough:

Ra = (rp + 2RL)/mu + Rk

Thus, for example, using a 6DJ8 working under a 125-volt cathode-to-plate differential, a 10mA idle current, 280-ohm cathode resistors, and into a 300-ohm load, the optimal plate resistor value is 402 ohms (122 ohms plus 280 ohms, in other words).

So is this modification really worth doing? I think so, for two major reasons. The first is that I hate placing anything in series with a cathode, other than an un-bypassed resistor. Many forget that the cathode has a slightly greater gain than the grid, which is a big oversight, as any weirdness or failing at the cathode will be amplified in phase at the plate (remember the ground-grid amplifier). A resistor placed in series with the cathode, on the other hand, linearizes the triode's transfer function. The second reason is that the distortion spectrum changes to the ear's benefit. Often with the added resistors the second harmonic does not drop—in fact, it might rise a tad—but the higher harmonics, say the 4th, 5th, 6th, 7th, fall a great deal. This a great deal, swapping away some second harmonic for getting rid of higher harmonics.

The downside is the increased output impedance, which may prove a deal breaker for some. My recommendation is to experiment a bit and try the added cathode resistors; I am sure that you will like what you hear.

//JRB

*In point of fact (as the Brits like to say), after I had had exhausted the obvious sources for tube gold, such as the *Radiotron Designers Handbook*, *Vacuum Tube Amplifiers*, and old issues of *Audio Engineering*, I had to dig into less obvious

sources. I am glad I did, as veins thick with nuggets were there for the taking. For example, look into articles and books that deal with tube-based DC amplifiers and instrumentation for some easy gold mining.

The White Cathode Follower as An OTL Power Stage for Headphones

Alex Cavalli

There are still many tube aficionados who design and build tube headphone amplifiers. Many of these designs are output transformer-less (OTL). Choosing an OTL topology, however, severely limits the amount of power that can be delivered to the headphones for many of the common and cost-effective tube types that are used in these designs. In order to achieve the maximum power from tube OTLs, a number of these designs use the White Cathode Follower (WCF) as the output stage. There are several reasons why the WCF is suitable, if not the most suitable, tube topology for this purpose:

1. The WCF has low Zo
2. The WCF can operate in push-pull mode delivering twice the idle current into the load compared to a standard Cathode Follower for the same quiescent current
3. The WCF's push-pull operation will cancel some amount of distortion
4. The input to the WCF is at high enough DC voltage to be coupled directly to the gain stage's plate output, thereby avoiding a coupling capacitor and associated grid bias resistor(s)
5. The WCF does not need a separate phase splitter to achieve push-pull operation

Items 4 and 5, while not affecting power delivery, are helpful in reducing complexity and component count.

Modern headphones impose severe conditions for tube OTLs because their impedances are so low. For common, good quality headphones these impedances typically range from 32R to 300R. These impedances impose nearly vertical load lines on the output stage. There is no practical way to eliminate this difficulty, but there is a set of optimizations that offer good performance and make the WCF an excellent headphone driver.

The optimizations involve:

1. Setting high enough quiescent current in the output stage (WCF is class A)
2. Setting the correct plate load resistor value to achieve push-pull operation
3. Application of NFB to reduce distortion at smaller loads

This article will address #2.

Achieving Balanced Push-Pull Operation

The goal in selecting the operating point and component values for the WCF OTL stage is to achieve push-pull balance in the amplifying and regulating triodes. Balanced push-pull operation will maximize the power and minimize the distortion.

In an article published in the TubeCad Journal Webzine (www.tubecad.com) October 2000, John Broskie proposed a plate load resistor optimization to achieve balanced push-pull operation for all values of the load. This optimization, as will be shown below, only achieves balanced push-pull for zero load. The question then is: is it possible to achieve balanced push-pull operation for non-zero load under conditions typical to headphone amplifiers and what component values should be selected to do this?

Small Signal Analysis

Figure 1 shows the WCF topology with the important voltages and currents noted. The subscripts $_a$ and $_r$ refer to the amplifying and regulating triodes respectively.

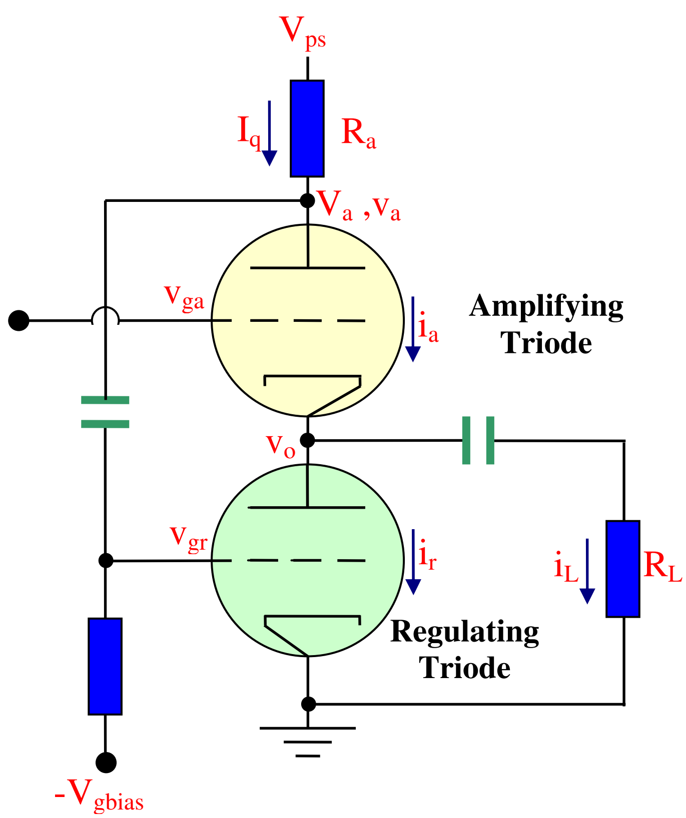


Figure 1 – The White Cathode Follower

The small signal parameters are:

i_a = current in the amplifying triode
i_r = current in the regulating triode
i_L = current in the load
r_p = triodes' plate resistance

R_a = plate load resistor
R_L = load resistor
v_a = plate voltage of the amplifying triode
v_o = output voltage
v_{gr} = grid voltage of the regulating triode
v_{ga} = grid voltage of the amplifying triode

The DC parameters are:

V_{ps} = the power supply voltage
V_a = amplifying triode plate voltage
V_{gbias} = regulating triode grid bias
I_q = the quiescent current

The following standard triode relationships will be used:

$$g_m = \frac{\mu}{r_p}$$

$$i_p = g_m v_g$$

where:

g_m = transconductance at the specific operating condition
μ = amplification factor

and i_p is the variation in plate current due to variation in grid voltage v_g.

A critical assumption for most of this analysis is that the regulating triode does not go positive grid. This will be addressed, approximately, below.

For simplicity we will assume that the triodes are identical. It is straightforward to extend the analysis for different amplifying and regulating triodes, which will also account for the same type triodes with different characteristics.

By inspection of the diagram and using the triode relationships we can establish five simple relationships among the various voltages and currents:

$$i_r = i_a - i_L \quad (1)$$

$$i_r = g_m v_{gr} + \frac{v_o}{r_p} \quad (2)$$

$$v_o = i_L R_L \quad (3)$$

$$v_a = v_{gr} = -i_a R_a \tag{4}$$

$$i_a = g_m\left(v_{ga} - v_o\right) + \frac{\left(v_a - v_o\right)}{r_p} \tag{5}$$

Substituting Eq. 3 & Eq. 4 into Eq. 2 gives:

$$i_r = -g_m i_a R_a + \frac{i_L R_L}{r_p}$$

Substituting Eq. 1 for the left side we have:

$$i_a - i_L = -g_m i_a R_a + \frac{i_L R_L}{r_p}$$

Solving for i_L gives:

$$i_L = i_a\left[\frac{r_p\left(1 + g_m R_a\right)}{\left(R_L + r_p\right)}\right] = i_a \frac{\left(r_p + \mu R_a\right)}{\left(R_L + r_p\right)} \tag{6}$$

Using Eqs. 1 and 6 we can obtain the relationship between i_r and i_a:

$$i_r = i_a \frac{\left(R_L - \mu R_a\right)}{\left(R_L + r_p\right)} \tag{7}$$

Equations 6 and 7 define the relationships among the currents in the circuit. It is instructive to observe the behavior of Eq. 7 as R_L is swept from 0 to ∞ for a given R_a.

$$i_r = -i_a \frac{\mu R_a}{r_p} \quad for \quad R_L = 0 \tag{8}$$

$$i_r = i_a \qquad for \quad R_L = \infty$$

When i_r is negative with respect to i_a the WCF is in push-pull mode. When it is positive push-pull operation is lost. These equations show that the WCF is in push-pull when $R_L = 0$, but it is not in push-pull when $R_L = \infty$. The phase transition occurs when Eq. 7 evaluates to zero or when:

$$R_L = \mu R_a \tag{9}$$

This establishes the first condition for maximizing the power from the WCF. The RHS of Eq. 7 must be negative, leading to:

$$R_a \geq \frac{R_L}{\mu} \tag{10}$$

Returning to Eq. 8 we note that for perfect push-pull behavior we must have $i_r = -i_a$ so that the triode currents are $180°$ out of phase. In general, Eq. 8 does not evaluate to $i_r = -i_a$ showing that the WCF is not in perfect push-pull at $R_L = 0$. There is the special case where $\frac{\mu R_a}{r_p} = 1$ that will be addressed below.

Using the fact that for perfect push-pull we must have $i_r = -i_a$, we solve Eq. 7 for the value of R_a that delivers perfect push-pull of any given R_L. Substituting $i_r = -i_a$ into Eq. 7:

$$-i_a = i_a \frac{(R_L - \mu R_a)}{(R_L + r_p)}$$

Solving for R_a:

$$R_a = \frac{1}{g_m}\left[1 + \frac{2R_L}{r_p}\right] \tag{11}$$

This establishes the second condition for maximizing the power from the WCF. For any given R_L setting the value of R_a according to Eq. 11 will place the WCF in balanced push-pull operation and maximize the power delivery into the load.

It is worthwhile to verify if this condition satisfies Eq. 10. Rearranging Eq. 11 gives:

$$R_a = \frac{2R_L}{\mu} + \frac{r_p}{\mu}$$

Therefore, Eq. 11 always satisfies Eq. 10.

Broskie's Special Case

In Broskie's article he determined that setting $R_a = \frac{1}{g_m}$ would balance the WCF for all possible loads. Substituting this value into Eq. 7 and setting $i_r = -i_a$ we have:

$$-1 = \frac{\left(R_L - \frac{\mu}{g_m}\right)}{\left(R_L + r_p\right)} = \frac{\left(R_L - r_p\right)}{\left(R_L + r_p\right)}$$

The only value for which this can be true is $R_L = 0$. Therefore, Broskie's special case achieves balanced push-pull operation only for $R_L = 0$ and cannot provide this balance at any other value of R_L.

Back to the Current Balance

It has been established that, for maximum power delivery into a specific load, it is possible to compute the correct R_a. Unfortunately, the headphone amplifier must drive 32R to 300R loads. Given that balance is only obtainable for one load, it becomes important to know what level of imbalance there will be at other loads.

As an example, select a load R_B where balance is to be achieved:

$$32 \le R_B \le 300$$

Substituting R_B into Eq. 11 gives the correct plate resistor:

$$R_a = \frac{1}{g_m}\left(1 + \frac{2R_B}{r_p}\right) \tag{12}$$

To calculate the imbalance at the extremes of R_L it is necessary to know the ratio $\frac{i_r}{i_a}$. Eq. 6 can be rearranged to give this result:

$$\frac{i_r}{i_a} = \frac{\left(R_L - \mu R_a\right)}{\left(R_L + r_p\right)} \tag{13}$$

Putting Eq. 12 into Eq. 13 gives:

$$\frac{i_r}{i_a} = \frac{\left(R_L - 2R_B - r_p\right)}{\left(R_L + r_p\right)} \tag{14}$$

As a test, setting $R_L = R_B$ gives $\frac{i_r}{i_a} = -1$, the correct result because, by definition, when $R_L = R_B$ the currents are balanced.

A 6922 will be used for the example calculations. Datasheet values for 6922 are typically:

$r_p = 3000$, $g_m = .011$, and $\mu = 33$

Choosing R_B arbitrarily in its range:

$R_B = 150\Omega$

The plate resistance calculated from Eq. 12 is:

$$R_a = \frac{1}{g_m}\left(1 + \frac{2R_B}{r_p}\right) = \frac{1}{.011}\left(1 + \frac{300}{3000}\right) = 100$$

So that when $R_B = 150$ and $R_a = 100$ the WCF will be in perfect balance.

If $R_L = 32$ then:

$$\frac{i_r}{i_a} = \frac{(R_L - 2R_B - r_p)}{(R_L + r_p)} = \frac{32 - 300 - 3000}{3032} = -1.0778$$

If $R_L = 300$ then:

$$\frac{i_r}{i_a} = \frac{(R_L - 2R_B - r_p)}{(R_L + r_p)} = \frac{300 - 300 - 3000}{3300} = -0.909$$

If the amp is balanced into 150R then this is how much it gets unbalanced at the extremes of load impedance.

Calculating the Gain of the WCF

To verify this analysis, we can derive the current into the load as a function of the input signal, $v_{ga.}$ To start we must find the current in the amplifying triode as a function of $v_{ga.}$ This is given by Eq. 5:

$$i_a = g_m\left(v_{ga} - v_o\right) + \frac{\left(v_a - v_o\right)}{r_p} \tag{5}$$

Using Eq. 3 & 4 and solving for i_a yields:

$$i_a = \frac{\mu\, v_{ga} - (\mu + 1) i_L\, R_L}{R_a + r_p} \tag{15}$$

Substituting this value for i_a into Eq. 6 and solving gives:

$$i_L = \frac{\mu\,\left(\mu R_a + r_p\right) v_{ga}}{\left(R_a + r_p\right)\left(R_L + r_p\right) + (\mu + 1) R_L\left(\mu R_a + r_p\right)} \tag{16}$$

This agrees with the equation for A_v in VA3 pp 112 assuming that the triodes are identical. This result, however, is valid for any load, whereas the result on pp 112 is only valid for infinite load. Using Eq. 16 to calculate the gain and putting $R_L = \infty$ gives the result from Amos and Birkinshaw shown in VA3. To show this we use $v_o = i_L\, R_L$ and $A_v = \frac{v_o}{v_{ga}}$. Eq 16 becomes:

$$A_v = \frac{\mu\,\left(\mu R_a + r_p\right) R_L}{\left(R_a + r_p\right)\left(R_L + r_p\right) + (\mu + 1) R_L\left(\mu R_a + r_p\right)} \tag{17}$$

Letting $R_L \to \infty$ we have for A_v:

$$A_v = \frac{\mu\,\left(\mu R_a + r_p\right)}{\left(R_a + r_p\right) + (\mu + 1)\left(\mu R_a + r_p\right)}$$

Or

$$A_v = \frac{\mu\,\left(\mu R_a + r_p\right)}{r_p(\mu + 1) + r_p + R_a\left[\mu(\mu + 1) + 1\right]} \tag{18}$$

This is the VA3 result if the triodes are identical.

Large Signal Considerations

When large signal excursions are present, there are several limiting features of the WCF that require modification to the linear analysis for optimum performance. This section will try to account for these.

Referring the Figure 1, there are two critical limiting voltages: (V_{ps} - V_a) and $V_{gbias.}$ These define two regimes of operation separated by a fuzzy transition regime. The regimes are $\left|V_{\text{gbias}}\right| \leq \left|\left(V_{ps} - V_a\right)\right|$ and $\left|V_{\text{gbias}}\right| \geq \left|\left(V_{ps} - V_a\right)\right|$.

When $|(V_{ps} - V_a)|$ is larger than $|V_{gbias}|$, generally when R_a is large, then V_{gbias} limits the maximum output. The larger (V_{ps} - V_a) is compared to V_{gbias} the sooner the regulating triode will go to positive grid.

But, when the opposite is true, (V_{ps} - V_a) limits the output because it defines the maximum signal that can reach the regulating triode. Furthermore, this limit is not reached abruptly, but nearly asymptotically as the amplifying triode enters cutoff. This limit also introduces an asymmetry in the current flow because the amplifying triode can pull V_a down more than it can pull it up. The limiting positive swing is (V_{ps} - V_a) which limits the maximum signal that can be applied to the regulating grid to $v_a = v_{gr} = \left(V_{ps} - V_a\right)$.This condition, as a limit, is a soft limit dependent on the specific non-linearity of the amplifying triode.

The transition region is fuzzy because of the asymptotic approach to cutoff and because the triode enters positive grid as a forward-biased diode.

The Regime $\left|V_{\text{gbias}}\right| \leq \left|\left(V_{ps} - V_a\right)\right|$

It is possible to compute when the WCF will cause the lower triode to enter positive grid operation. Setting aside the non-linearity exhibited beyond the small signal regime, maintaining Class A leads to the requirement:

$$\left|v_{gr}\right| \leq \left|V_{gbias}\right| \tag{19}$$

The first step is to find the maximum v_o as a function of v_{gr} and then apply Eq. 19. Starting with Eq. 2:

$$i_r = g_m v_{gr} + \frac{v_o}{r_p} \tag{2}$$

and substituting Eq. 1 leads to:

$$i_a - i_L = g_m v_{gr} + \frac{v_o}{r_p}$$

Using the relationships $i_a = -\frac{V_a}{R_a} = -\frac{V_{gr}}{R_a}$ from Eq. 4 and $i_L = \frac{v_o}{R_L}$ gives:

$$-\frac{V_{gr}}{R_a} - \frac{v_o}{R_L} = g_m v_{gr} + \frac{v_o}{r_p}$$

Solving for v_o:

$$v_o = -v_{gr}\left[\frac{R_L(r_p + \mu R_a)}{R_a(r_p + R_L)}\right] \tag{19}$$

Which, using Eq. 18, gives the maximum v_o as:

$$v_{o_{max}} = |V_{gbias}|\left[\frac{R_L(r_p + \mu R_a)}{R_a(r_p + R_L)}\right] \tag{20}$$

Now, using Eq. 17 written for the voltage relationship:

$$v_o = v_{ga}\,\frac{\mu\ (\mu R_a + r_p)\ R_L}{(R_a + r_p)(R_L + r_p) + (\mu + 1) R_L (\mu R_a + r_p)}$$

Substituting Eq. 20 for v_o and solving for v_{ga} we obtain:

$$v_{ga} = -v_{vgr}\left[\frac{(R_a + r_p)(R_L + r_p) + (\mu + 1) R_L (\mu R_a + r_p)}{\mu R_a (R_L + r_p)}\right]$$

Therefore the maximum input signal that can be applied before the regulating triode leaves class A mode is theoretically:

$$v_{ga_{max}} = |V_{gbias}|\left[\frac{(R_a + r_p)(R_L + r_p) + (\mu + 1) R_L (\mu R_a + r_p)}{\mu R_a (R_L + r_p)}\right] \tag{21}$$

Application of this input signal gives the maximum output voltage in Eq. 20.

The Regime $|V_{\text{gbias}}| \geq |(V_{ps} - V_a)|$

For many of the triodes used in WCF headphone stages the optimization given in Eq. 11 leads to plate resistors that are in the range of 100Ω–200Ω. This optimization, coupled with idle currents in the 10mA to 20mA range, often places the WCF into this regime.

Even though the resistors are chosen for small-signal balanced push-pull operation, the WCF cannot reach maximum current into the load and becomes unbalanced at large signals because (V_{ps} - V_a) is not large enough and because of the asymmetries in the plate curves.

It is nearly impossible to calculate the behavior of the WCF in this regime since the triodes are entering highly non-linear parts of their curves, but it is possible to make a scaling argument that will give a good idea of how to modify the small-signal optimum value of the plate resistor R_a.

Let us make two scaling assumptions: that the both amplifying and regulating triodes conduct more than they cutoff. If we label the peak conduction current as i^p then we can define:

$$i_a^+ = i_a^p \text{ and } i_a^- = k_a i_a^p \tag{22}$$

where i_a^+ and i_a^- are the peak positive and negative instantaneous currents in the amplifying triode and $k_a \leq 1$ is the proportionate reduction of current in the cutoff regime. Similarly for the regulating triode:

$$i_r^+ = i_r^p \text{ and } i_r^- = k_r i_r^p \tag{23}$$

Where i_r^+, i_r^- and k_r are the equivalent quantities for the regulating triode. Note that i_a^p and i_r^p are not assumed to be equal.

For peak positive current into the load we have current relationship from Eq. 1:

$$i_a^+ = i_L^+ + i_r^-$$

and similarly for the peak negative current:

$$i_a^- = i_L^- + i_r^+$$

Using Eqs. 22 and 23 and rearranging to solve for i_L:

$$i_L^+ = i_a^p - k_r i_r^p \tag{24}$$

and

$$i_L^- = k_a i_a^p - i_r^p \tag{25}$$

For current balance into the load we must have $i_L^+ = i_L^-$:

$$i_a^p - k_r i_r^p = k_a i_a^p - i_r^p$$

and solving for i_r^p:

$$i_r^p = i_a^p \frac{(1+k_a)}{(1+k_r)} \tag{26}$$

Now we must solve for the value of R_a that makes this true. Returning to Eq. 2 and arbitrarily selecting the positive peak of the regulating triode, we begin with:

$$i_r^p = g_m v_{gr}^p + \frac{v_o^p}{r_p}$$

and substituting the various quantities:

$$i_r^p = k_a g_m i_a^p R_a + \frac{i_L^- R_L}{r_p} \tag{27}$$

Equation 26 gives us i_r^p but we still need i_L^-. We can get this from Eq. 25:

$$i_L^- = k_a i_a^p - i_r^p = k_a i_a^p - i_a^p \frac{(1+k_a)}{(1+k_r)}$$

Substituting this result together with Eq. 26 into Eq. 27 yields:

$$i_a^p \frac{(1+k_a)}{(1+k_r)} = k_a g_m i_a^p R_a + i_a^p \frac{R_L}{r_p}\left[k_a - \frac{(1+k_a)}{(1+k_r)}\right]$$

Solving this for R_a:

$$R_a = \frac{1}{g_m}\left[\frac{1}{k_a}\frac{(1+k_a)}{(1+k_r)}\frac{(r_p + R_L)}{r_p} + \frac{R_L}{r_p}\right] \tag{28}$$

As a test, putting $k_a = k_r = 1$ (perfect symmetry in the positive and negative peak values) gives:

$$R_a = \frac{1}{g_m}\left[\frac{1}{1}\frac{(1+1)}{(1+1)}\frac{(r_p + R_L)}{r_p} + \frac{R_L}{r_p}\right] = \frac{1}{g_m}\left[1 + \frac{2R_L}{r_p}\right]$$

which agrees with Eq. 11. As an example, the optimum small signal R_a for the 6922 WCF for 150Ω load was 100Ω. For the sake of this example, using some observed simulation behavior, we set $k_a = .7$ and $k_r = .75$. Substituting these values into Eq. 28 give the value for R_a.

$$R_a = \frac{1}{.011}\left[\frac{1}{.7}\frac{(1+.7)}{(1+.75)}\frac{(3000+150)}{3000} + \frac{150}{3000}\right] = 137\Omega$$

This value of Ra will better balance the large signal push-pull behavior of the WCF.

The calculations in the current balance section above together with Eq. 28 make it clear that there is no single value of R_a that will balance the WCF for a range of loads and for both small and large signal regimes. Any choice of R_a is, therefore, a compromise to get the best possible behavior for these two separate constraints.

The Morgan Jones Mini Tube Headphone Amplifier

by Chu Moy

Back in 1999, I received an email with an attached schematic of a tube amplifier. The sender told me that it was supposed to be the schematic for a "clone" of the famous **EarMax** miniature headphone amplifier. Since he had no more information about the design and had not built it, I filed it away to be referred to at a later date. Months later, I saw the schematic again in a book called **Valve Amplifiers (2nd ed.)** by Morgan Jones. In the book, Jones described it as a reverse-engineered version of the EarMax. That is, the schematic was not of the true EarMax, but was derived from the published specifications of the EarMax (e.g., 3 tubes, the power supply voltage, the input and output impedances). Jones had created the circuit as an academic exercise, but had not actually built it. I once again put the schematic away, hoping later to find a DIYer who could give construction details.

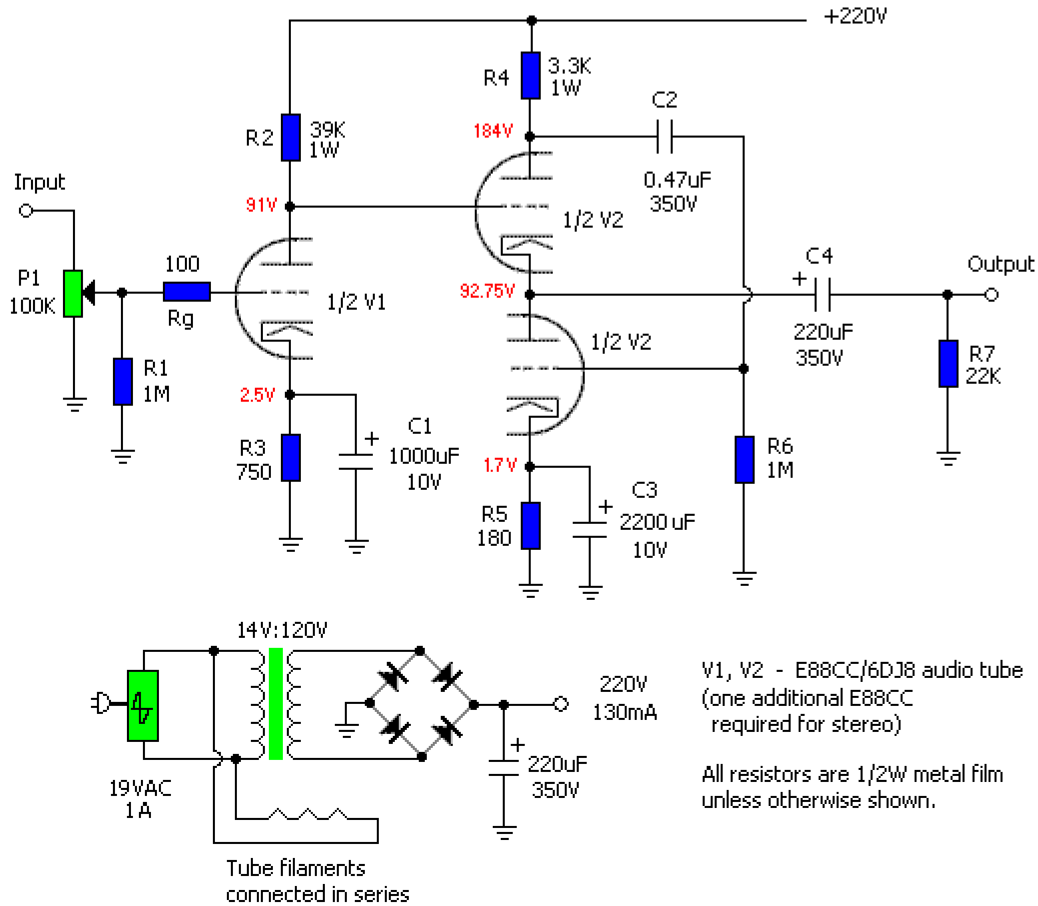
While doing research on the internet, I came upon a reference to the Jones design in an archive for the Sound Practices mailing list. Lance Dow, who knew Morgan Jones personally, had posted the schematic in that newsgroup way back in 1996. Dow had not built the circuit either. Given all the interest in the audiophile community about the EarMax, I thought that surely someone, somewhere, must have tried to build it. I scoured the Sound Practices archives, downloading year after year of digests, and finally found a posting by Johannes Chiu, who described enthusiastically his DIY work on

this design. I contacted Chiu for more information, but apparently he no longer remembers many specifics about it. This previous version of this article was a summary of the information collected from the Sound Practices archives and from the HeadWize forums about the original Morgan Jones circuit.

Since the publication of that article over a year ago, there have been several inquiries regarding the possible topology of the **EarMax Pro**. The Pro has basically the same output tube complement as the EarMax, but can provide higher current output into low impedance headphones. The specifications suggested that the topology of the EarMax Pro was similar to the EarMax (both having White cathode follower output stages), but the mystery remained as to how two amplifiers with similar output stages (no output transformers) could have substantially different output characteristics. Then in 2002, Alex Cavalli submitted revised Morgan Jones circuits with new parts values, based on the White cathode follower optimization techniques developed by John Broskie. The optimized Morgan Jones amplifiers (with and without feedback) can output more than 3 times the current of the original into a 32-ohm load. The amp shown at the top of this article is an optimized Morgan Jones amplifier (without feedback) built by Bryan Ngiam.

The Amplifier Designs

1. The Original Morgan Jones Amplifier



Original Morgan Jones Headphone Amplifier (one channel)

Figure 1

Figure 1 is the schematic for the original Morgan Jones amplifier. It has a grounded cathode input stage with an idling of about 3mA. The output stage is a push-pull White follower, which provides low output impedance without the need for global feedback. It idles at about 10mA and can swing ±20mA in push-pull. The output impedance is about 10 ohms (the calculated value was 6 ohms). The overall gain of the amp is about 22 (the calculated value was 28). Jones used an ECC88 input tube. The original EarMax has an ECC81 input tube, but he felt that the low anode current required for this stage would lead to noise and gain problems with the ECC81. The EarMax output tubes are ECC86. Jones used the similar ECC88, later used in the EarMax Pro's output as well. He rated the amplifier to drive headphones from 200 to 2000 ohms.

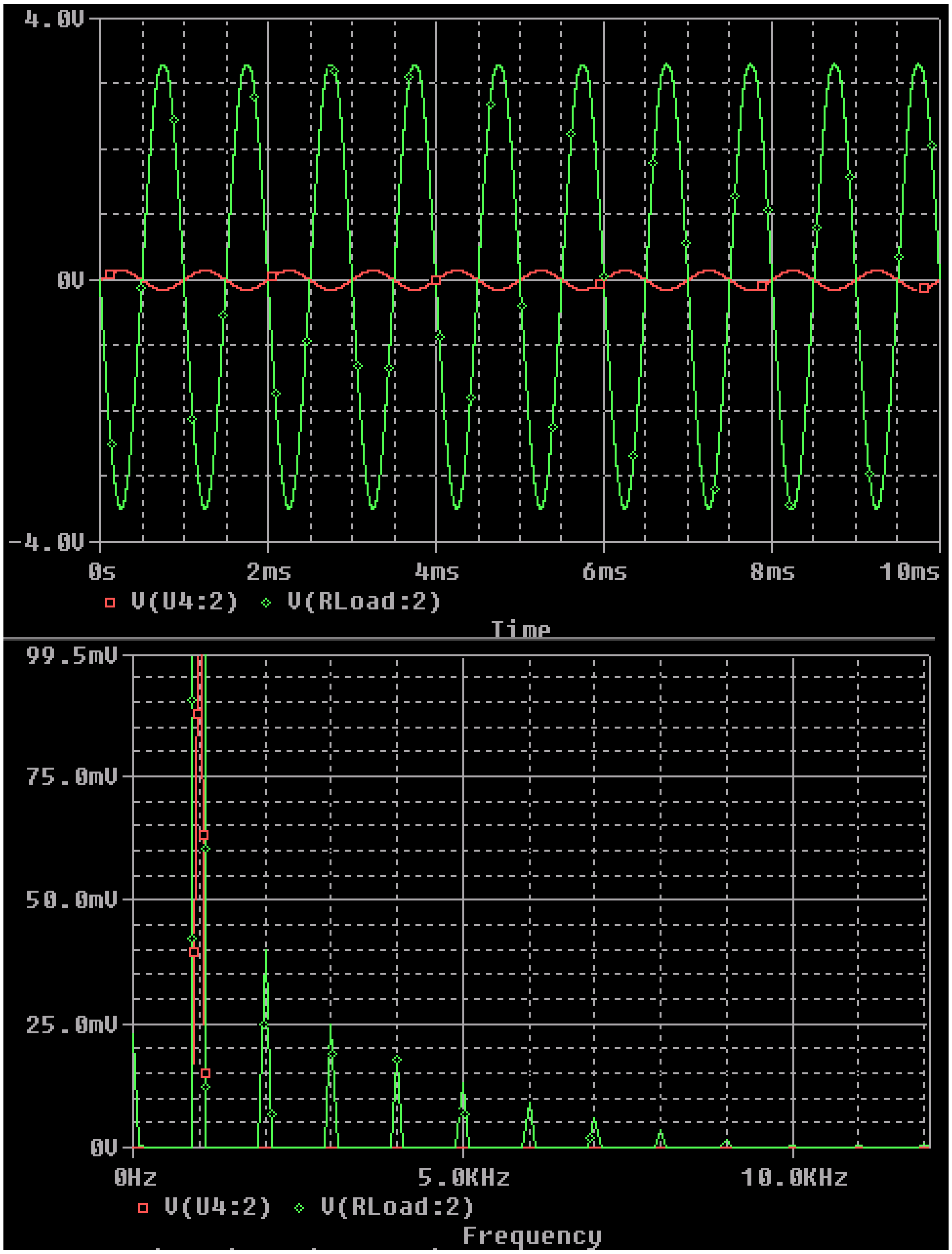


Figure 2

In all of following voltage and Fourier analysis graphs, the red curve is the input; the green curve is the output. The top graph in figure 2 shows the output voltage waveform into a 300-ohm load (the input is a 0.15V, 1 KHz sine wave). The bottom graph is a Fourier analysis of the output waveform to determine the harmonic distortion, which turns out to be about 2%. The output current (not shown) is 9mA, so the amplifier is driving the load with 36mW (the maximum power into 300 ohms is about 120mW).

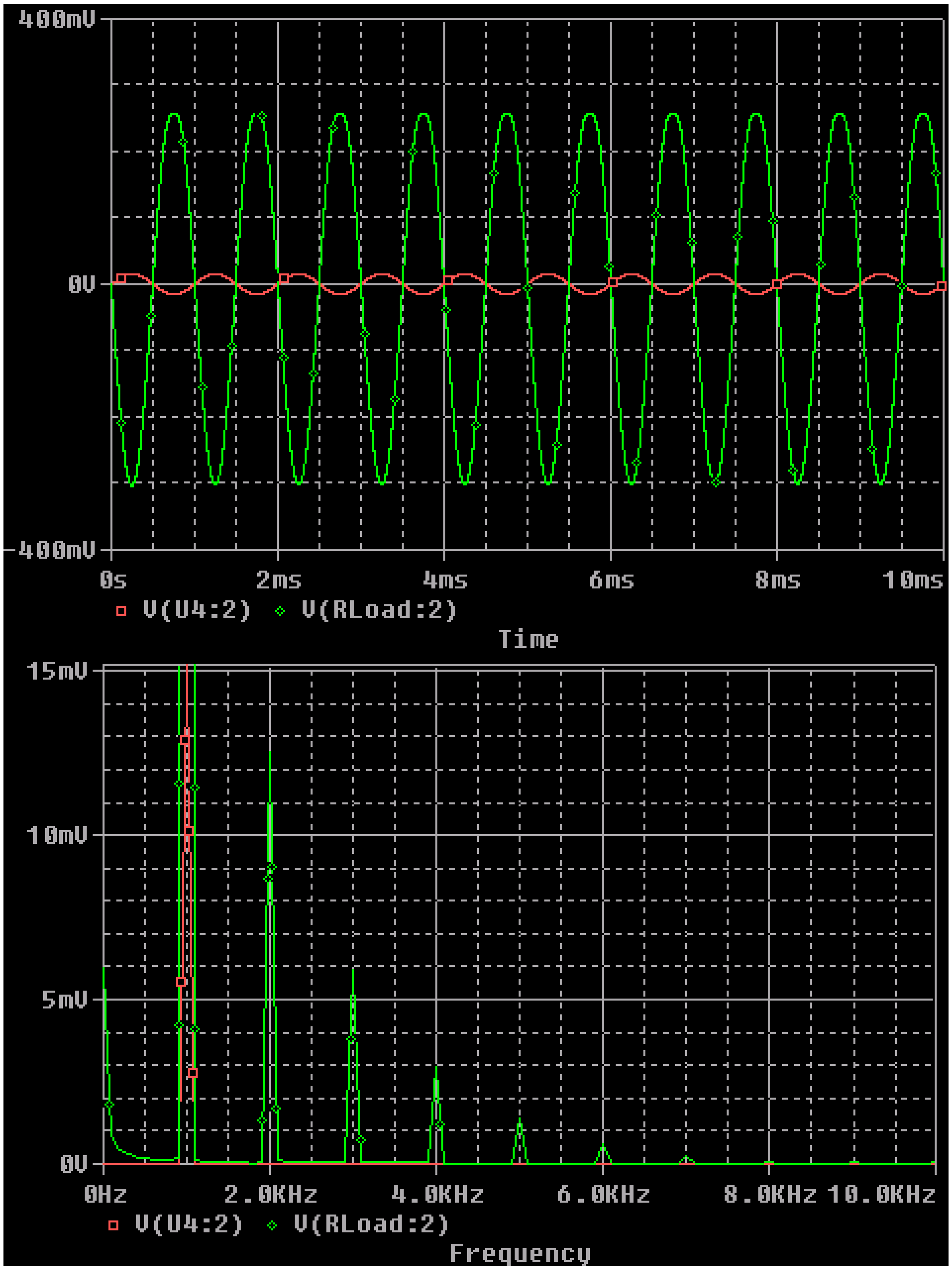


Figure 3

Figure 3 shows the same set of graphs for a 32-ohm load, with a 0.016V, 1 KHz sine wave. Like the example in figure 2, the output is being driven to a total harmonic distortion of 2%, but here, there is a distinct imbalance between the top and bottom halves of the waveforms, because the lower impedance load draws more current and is unbalancing the push-pull configuration. The amplifier is supplying about 2.6mW into 32 ohms (the maximum power into 32 ohms is about 13mW). Thus, like the EarMax, the original Morgan Jones amplifier is not truly suited to power low impedance headphones such as the Grados, despite the low output impedance.

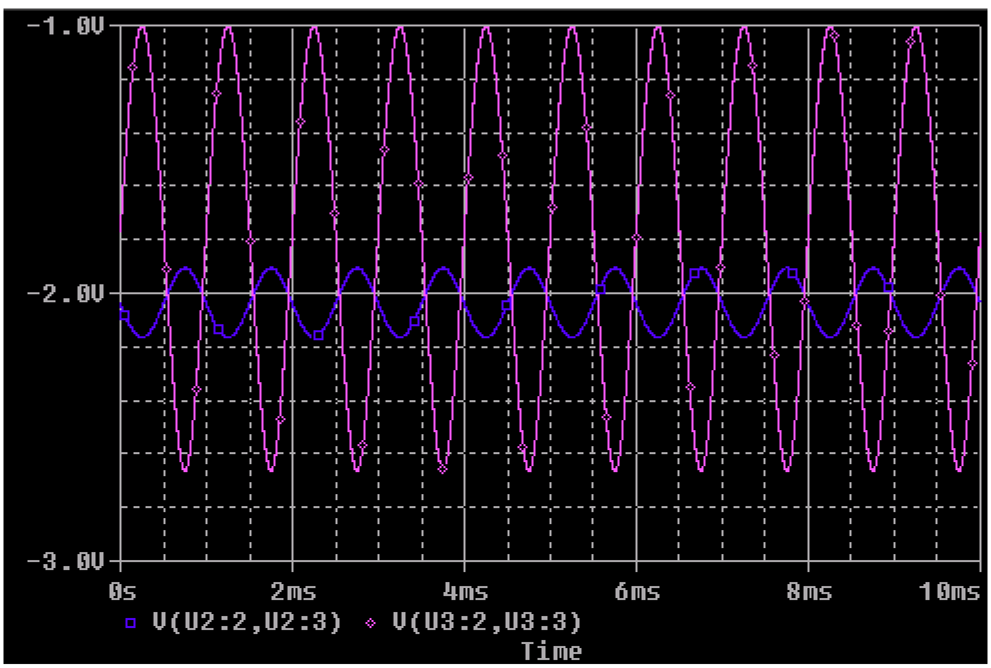
2. Analysis of the Performance of the Original Morgan Jones Amplifier

Theoretically, this type of output stage should be able to drive low impedance loads well, because it has a very low output impedance. In his article The White Cathode Follower, TubeCad editor John Broskie investigated the poor performance of the White cathode follower when driving low impedance loads. He discovered that the voltage drop across the anode load resistor (R4) of the top tube (V2a) varies with the current flowing through the tube. If the voltage across R4 is high enough, it will overdrive the bottom tube (V2b).

Alex Cavalli viewed the problem another way: the imbalance was caused by the gain of the V2a being greater than 1:

> Assume that the output of the amplifier is shorted (an AC short at the junction of the upper and lower triodes in the output stage) and ignore the fact that the tubes are in series. Under this condition, both the upper and lower triodes are operating as simple grounded cathode amplifiers, where the output of the upper section is fed directly into the lower section.
>
> Now the gain of the first stage is about 25. The gain of the upper triode in grounded cathode mode is about 15. If a 0.01V sine wave is applied to the input, the first stage will produce about 25 x 0.01 =.25V at its plate. Thus the upper triode sees 0.25V on its grid. The upper stage in turn produces 15 x 0.25 = 3.75V, which is coupled to the grid of the lower triode.
>
> There are two things to note here:
>
> 1. the grid drive to the push-pull sections is unequal, 0.25V (upper) vs. 3.75V(lower) and
> 2. the bottom triode, which has a bias of about 1.75V, is being driven hard into cutoff and positive grid.
>
> In this design the upper and lower output sections are not working together equally and so they are not producing the maximum possible current swing. Furthermore, the enormous gain feeding the lower section makes the amp extremely sensitive and sends it out of class A mode quickly.

The maximum voltage that can appear at the grid of the bottom tube is determined by the DC biasing voltage across the cathode resistor. In figure 1, for example, the bias or idling voltage across R5 is 1.7V, so the maximum peak-to-peak voltage into the grid of V2b is 1.7V. A higher grid voltage will either turn the tube off or drive the grid positive and will push the triode out of Class A operation.



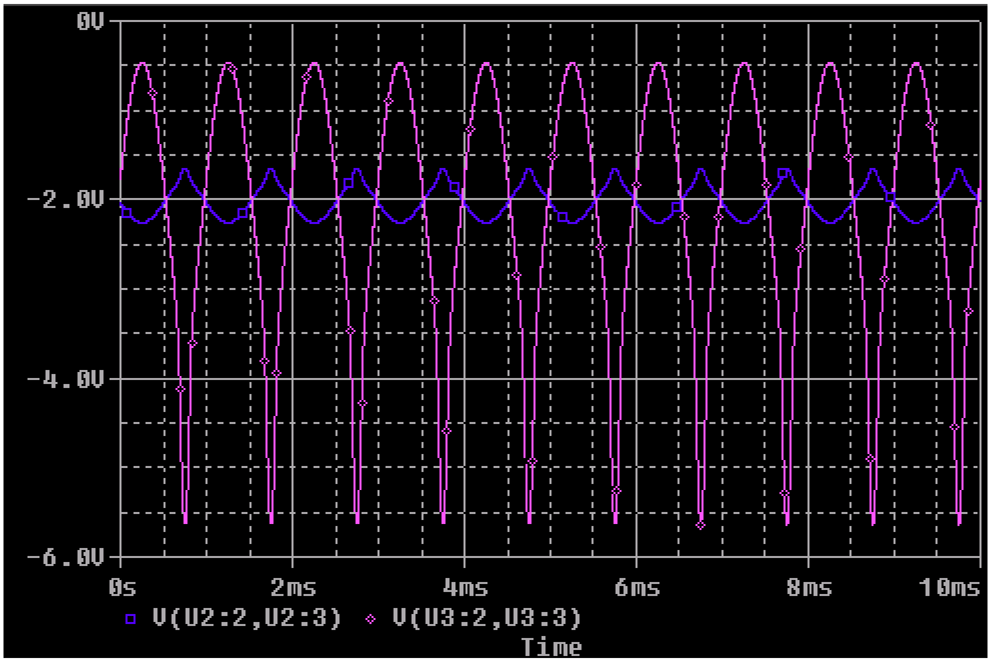


Figure 5

Each graph in figure 5 shows the differential voltages between the grid and cathode (V_{gk}) of the output tubes V2a (blue) and V2b (magenta) in the original Morgan Jones amplifier driving a 300-ohm load. In the top graph, V_{gk} for V2b is 1.66Vp-p, which is just less than the 1.7V bias voltage across R5. The V_{gk} for the top and bottom tubes appear symmetrical, but have unequal amplitudes. At this V_{gk}, the amplifier is outputting 2V into a 300-ohm load, which is approximately 13mW. This is the maximum output power of the original MJ amplifier into 300 ohms before push-pull stage leaves class A mode. The Fourier analysis indicates that the harmonic distortion at 13mW is only 0.3% (output voltage and Fourier graphs omitted).

The bottom graph shows the V_{gk} waveforms when the White follower stage is severely unbalanced. The V_{gk} for V2b exceeds 1.7Vp-p and the shapes of both waveforms are grossly distorted. Here, the original MJ amplifier is driving a 300-ohm load with an output voltage of 3.3V (about 36mW) and the harmonic distortion has risen to 2% (output voltage and Fourier graphs omitted).

After employing a similar White follower balance analysis for a 32-ohm load, the maximum output power of the original MJ amplifier into that load is actually less what was determined from figure 3: a mere 1.6mW at 1.3% distortion (0.228V output). Even for high efficiency, 32-ohm headphones like the Grados, 1.6mW is not enough to achieve clean volume levels - especially not if the music has wide dynamic range.

Broskie concluded that in order for the White follower to perform optimally (and maintain the balance in the push-pull pair), the anode load resistor R4 should be chosen so that the bottom tube receives an identical grid-to-cathode voltage signal as the top tube. In other words, V_{gk} for V2a should equal V_{gk} for V2b (the bias voltage across the cathode resistor still determines the limit of V_{gk}). His solution was to calculate a lower value for the anode load resistor (which he called R_a) based on the equation:

$$\mathbf{R_a = r_p / mu = 1 / G_m}$$

where G_m is the transconductance of the tube.

Alex Cavalli provided this explanation:

> The way to balance the grid drives where the output is shorted (a load of zero ohms) is to ensure that the upper output section has a gain of 1. This will cause the lower triode to see exactly the same grid signal as the upper triode. According to Broskie the effective G_m of the upper stage is:

$$G_m = (mu + 1) / (r_p + R_a)$$

where R_a is the anode load resistor and r_p is the plate resistance of the triode. The gain of upper stage is given by:

$$Gain = ((mu + 1) \times R_a) / (r_p + R_a)$$

To have a gain of 1:

$$((mu + 1) \times R_a) = (r_p + R_a)$$

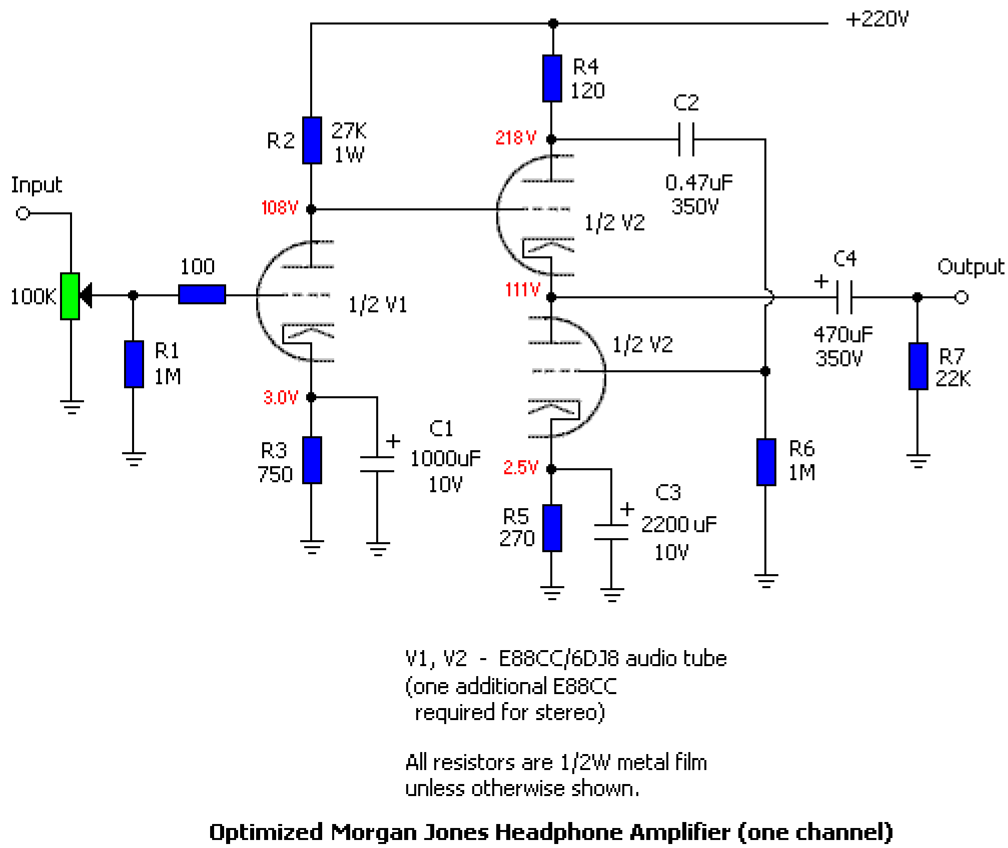
Then solving for R_a:

$$R_a = r_p / (mu) = 1 / G_m$$

This result is the same as Broskie's, except that he proves this result for all load impedances.

The transconductance for a 6DJ8 is 11mA/V, so R_a ~ 90 ohms.

3. The Optimized Morgan Jones Amplifier



Optimized Morgan Jones Headphone Amplifier (one channel)

Figure 4

Alex Cavalli's revised Morgan Jones circuit is shown in figure 4. It is identical the original, except that the power supply's current rating has been doubled and three resistor values in the amplifier have been changed. R2, R4, and R5 determine the balance for the White push-pull output stage. R2 determines the quiescent plate voltage on V1 which sets the grid bias on the V2a in combination with R4 and R5. These seemingly minor changes in the resistor values have a huge impact on the performance of the amp as discussed below.

He used PSpice simulations to determine the best values for R2, R4 and R5. Although Broskie determined that the optimal anode load resistor value R4 was 1/Gm (or 90 ohms for a 6DJ8), the

simulations indicated that the amplifier had better output characteristics with a higher value - 150 ohms. The output stage still idles at around 10mA. These modifications resulted in better performance into both 300-ohm and 32-ohm loads.

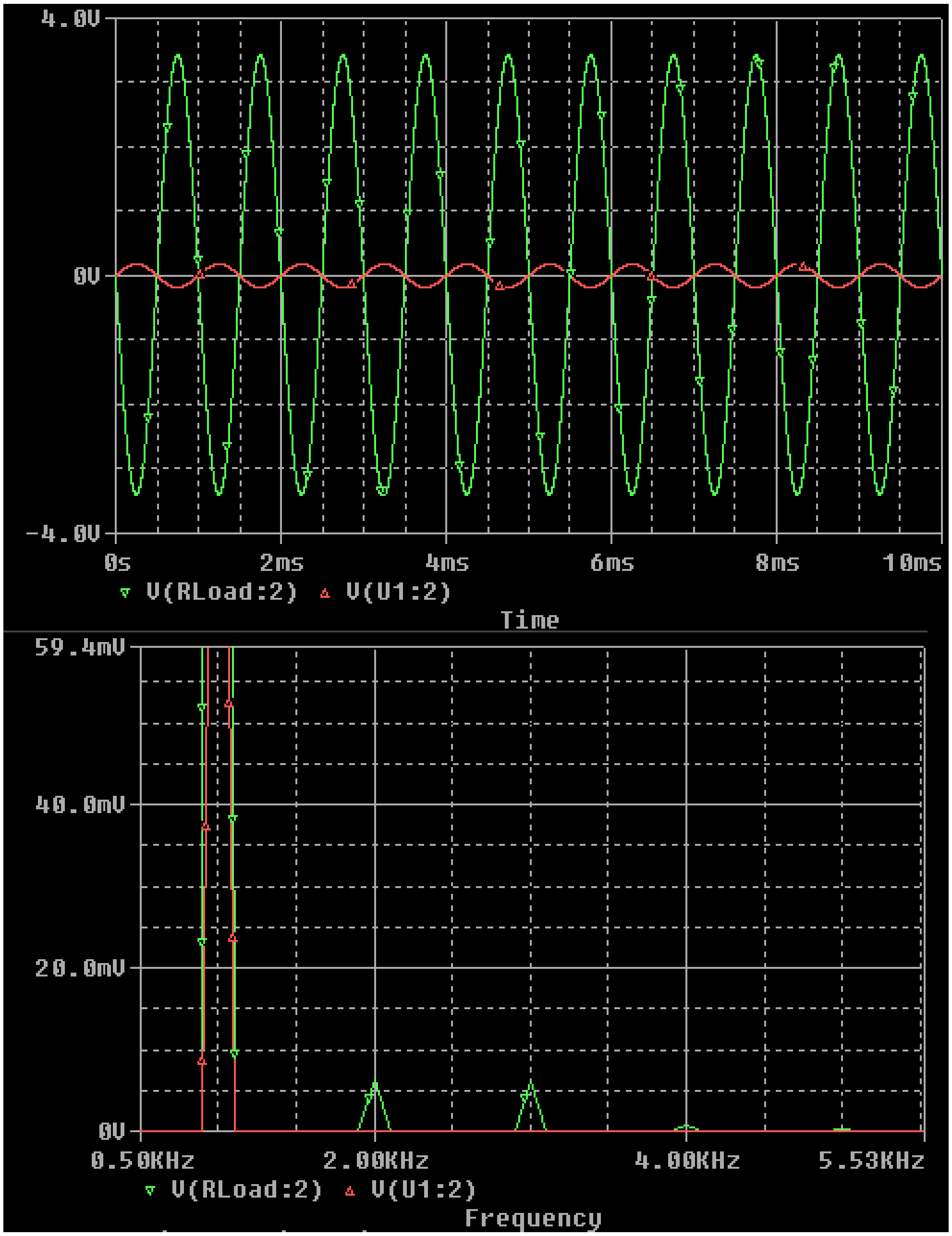


Figure 6

The output voltage of the Cavalli-optimized amplifier in figure 6 (top graph) was chosen by monitoring the V_{gk} for V2b until it reached about 2.5Vp-p, the same value as the DC bias voltage across R5. At that point, the amplifier's output voltage into 300 ohms is 5V or 83mW, a six-fold improvement over the 13mW maximum for the original Morgan Jones. The harmonic distortion at 83mW is about 1%.

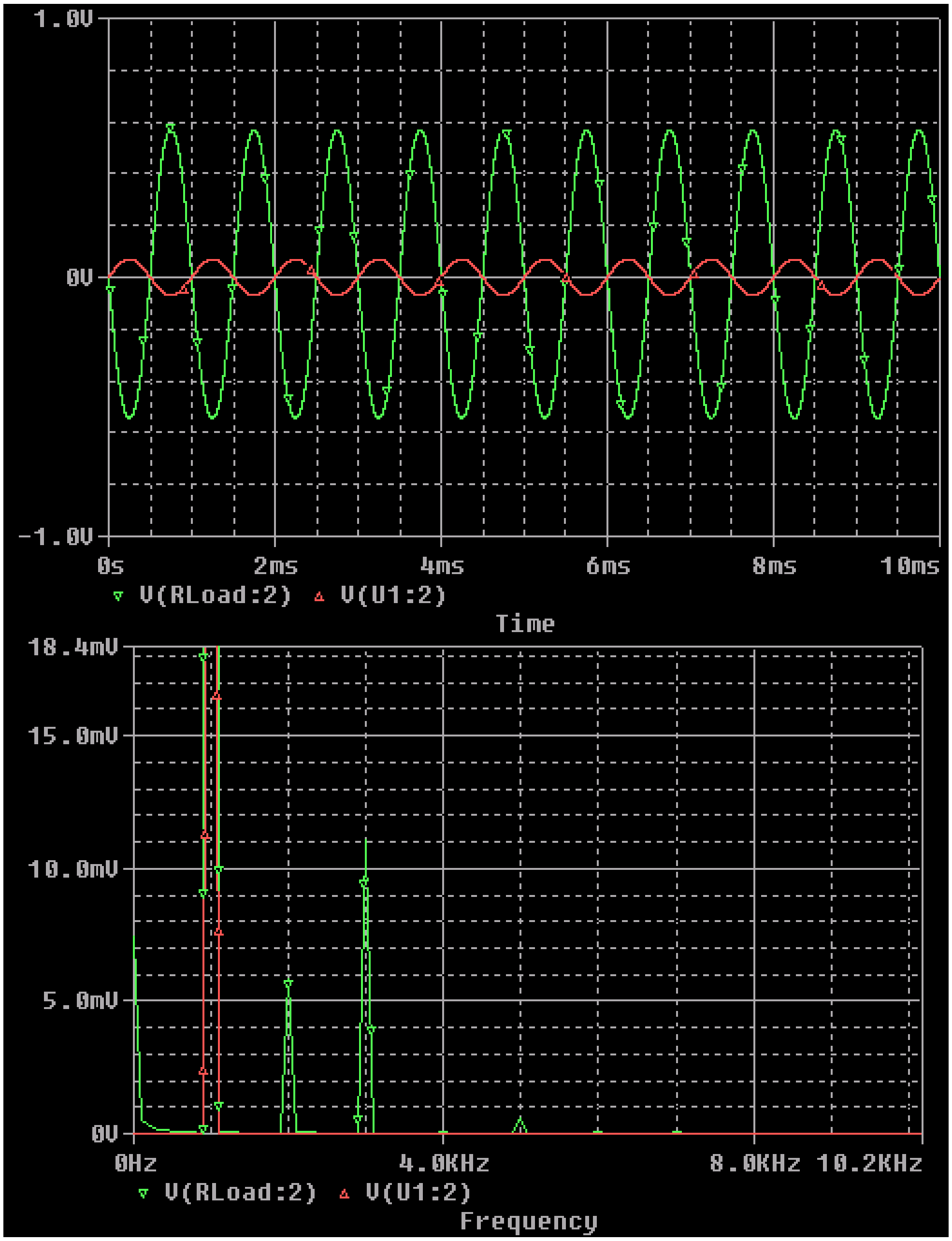


Figure 7

The performance of the optimized amplifier into a 32-ohm load is as remarkable. Again, the output voltage in figure 7 was chosen by monitoring the V_{gk} for V2b until it reached about 2.5Vp-p. Based on the results of the White follower balance analysis, the maximum output power of the amplifier into 32 ohms is 10mW, a six-fold improvement over the 1.6mW maximum of the original MJ amplifier, although the harmonic distortion at 10mW is also higher: 2.1%.

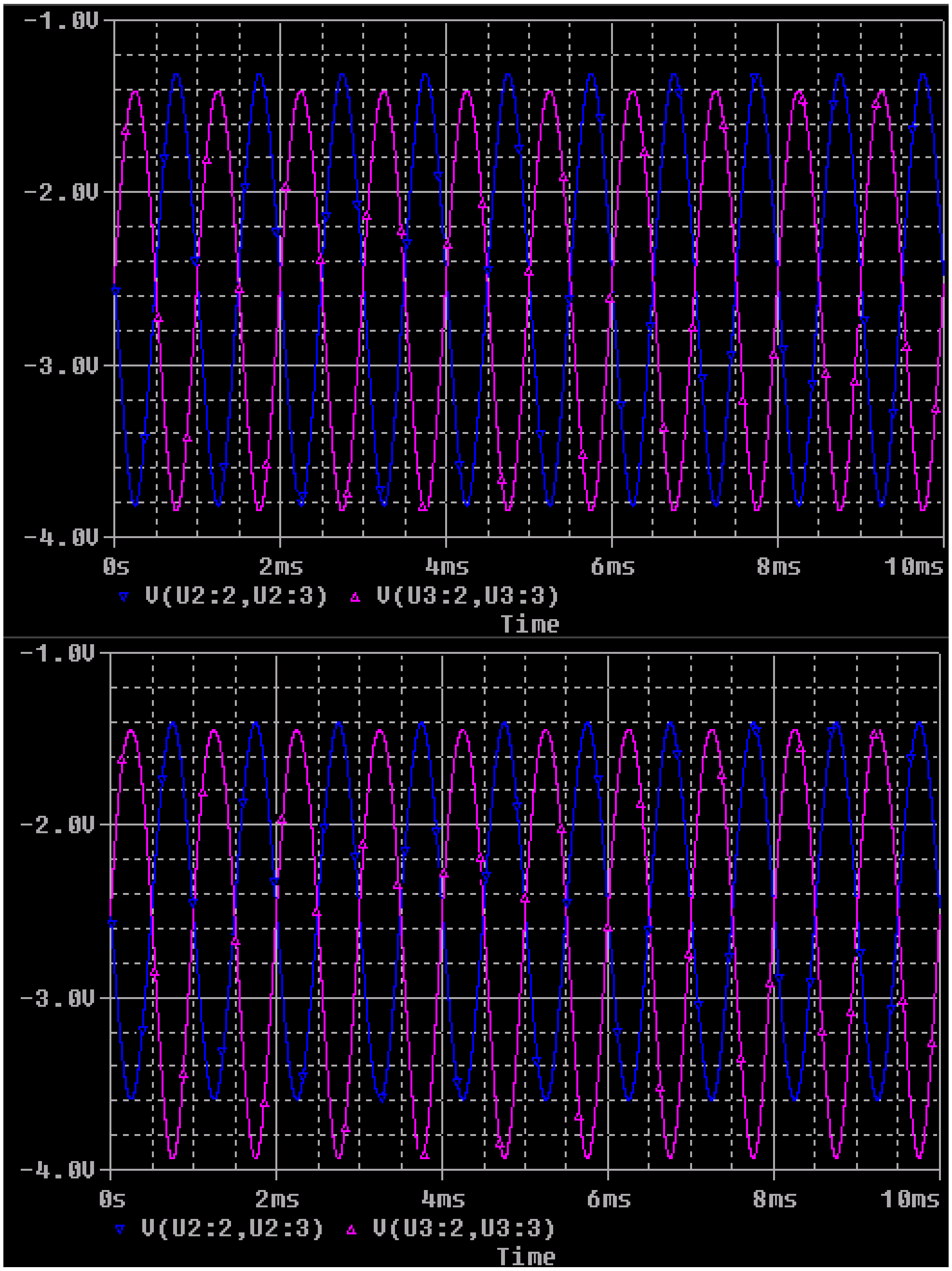


Figure 8

The graphs in figure 8 compare the V_{gk} for the output tubes V2a (blue) and V2b (magenta) for the setups in figures 6 and 7 respectively. When the top graph here is contrasted with the top graph in figure 5, the V_{gk} waveforms in the optimized amplifier when driving a 300-ohm load have achieved virtually perfect balance. The bottom graph shows that when the optimized amp is connected to a 32-ohm load, the V_{gk} waveforms are not quite as balanced (the amplitude of V_{gk} for V2a is slightly larger than that for V2b), but are vastly more balanced than the curves in figure 5.

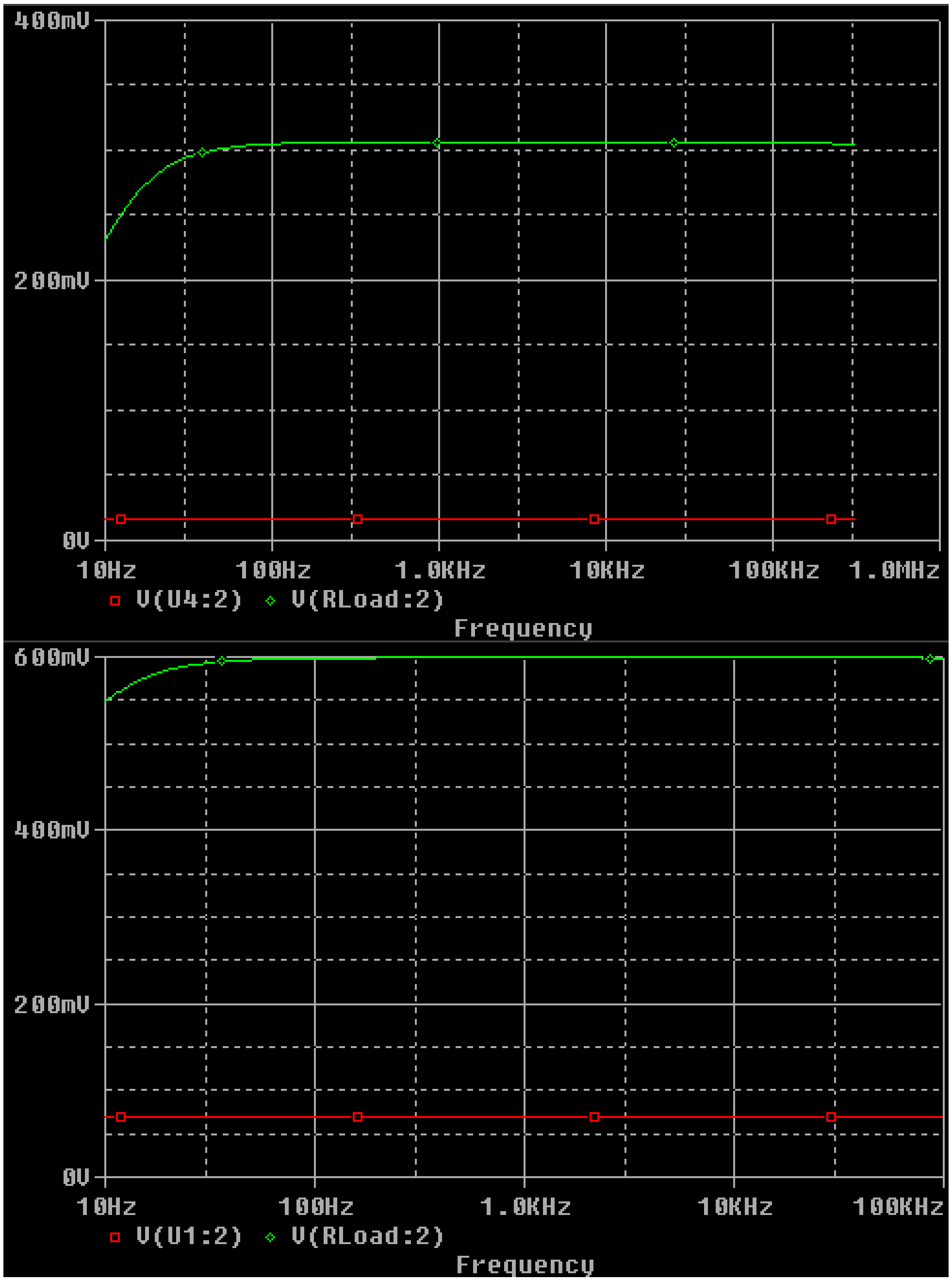
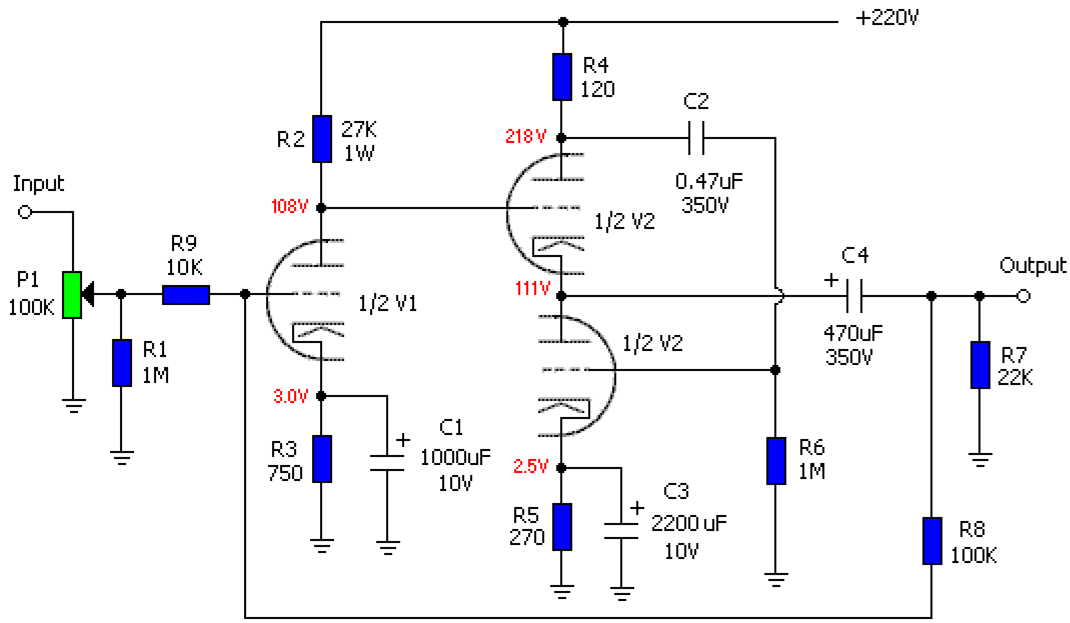


Figure 9

The frequency responses of the original (top graph) and optimized amplifiers (bottom graph) driving a 32-ohm load are shown in figure 9. The low frequency response of the optimized version has a more extended low-end than the original. The overall gain of the optimized amplifier is about 8, whereas the original has a gain of 19 (the graph scales do not make the differences in gain obvious, however). With the 300-ohm load, the difference in gains is far less: 23 and 19 for the original and optimized amps respectively (graphs not shown). The drop in gain with the 32-ohm load is due to the higher output impedance of the optimized amplifier, one of the tradeoffs of optimization. The original has an output impedance of about 10 ohms, but in the optimized version, the output impedance is 53 ohms - high enough to cause strong loading effects on a 32-ohm load. More than half of the amp's output voltage is absorbed by the output impedance in this case.

4. The Optimized Morgan Jones Amplifier with Feedback



V1, V2 - E88CC/6DJ8 audio tube
(one additional E88CC
required for stereo)

All resistors are 1/2W metal film
unless otherwise shown.

Optimized Morgan Jones Amplifier with Feedback (one channel)

Figure 10

While the optimized amplifier can dump 10mW into 32 ohms, adding feedback improves the performance even more. The schematic in figure 10 is identical to the one in figure 4, except for the resistors R8 and R9 that form a feedback loop and the removal of the 100-ohm grid stop resistor on V1. A separate grid-stop resistor is not needed, because of the presence of R9. Cavalli selected the feedback resistor values by examining trade-offs. Too much feedback and the gain was reduced too much. Too little feedback and there was no benefit. He tried to select values that reduced the output impedance down substantially below 32 ohms and reduced distortion, while still leaving enough gain to be sensitive to conventional CD inputs. Cavalli recommends experimenting with these resistor values to get the best tradeoff for specific headphones.

The output graphs for this amplifier driving a 300-ohm load are not shown here, but they indicate that it delivers the same power (83mW) as before, but at a lower distortion: 0.5%. Feedback also lowers the output impedance to about 20 ohms and the gain to about 6 for a 300-ohm.

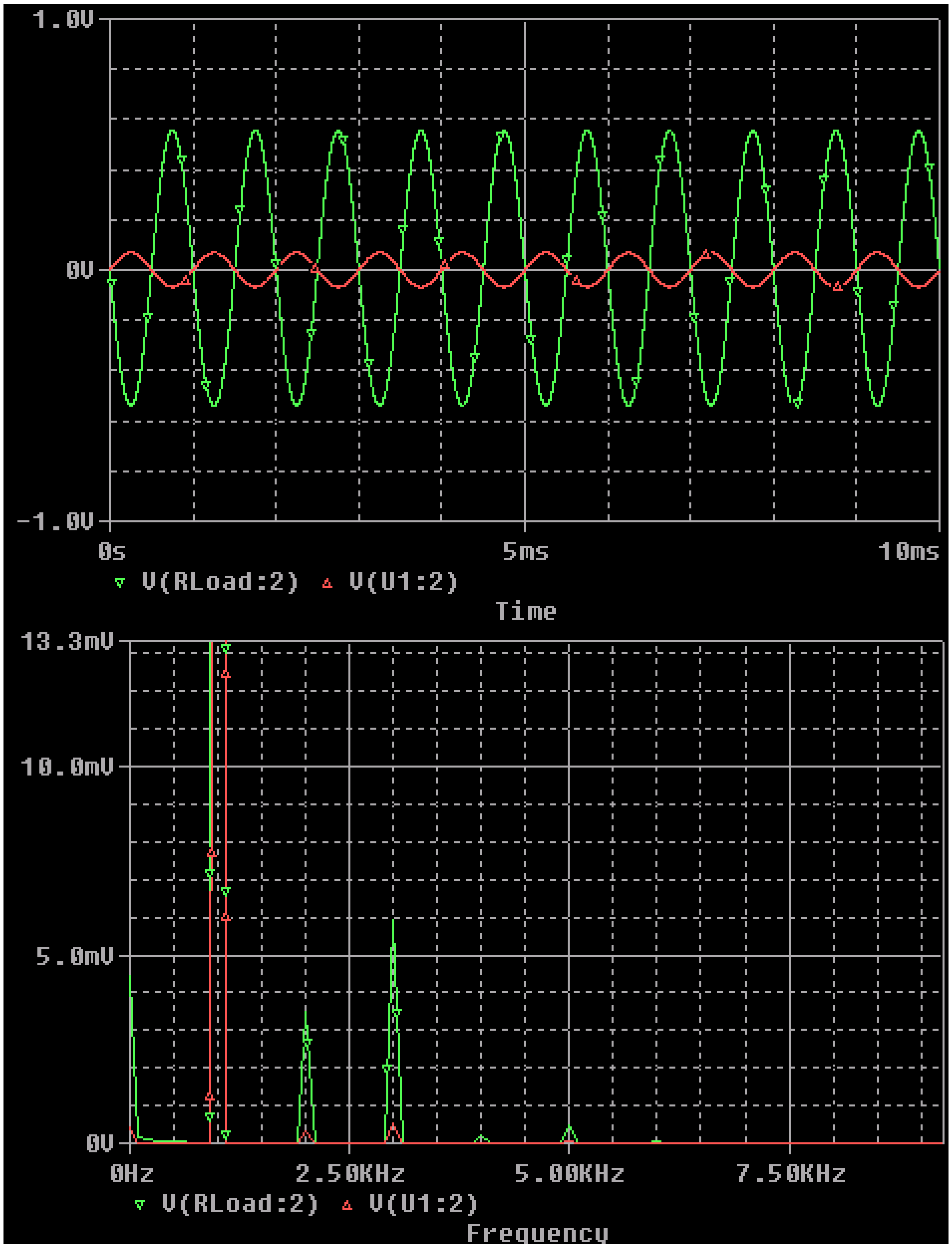
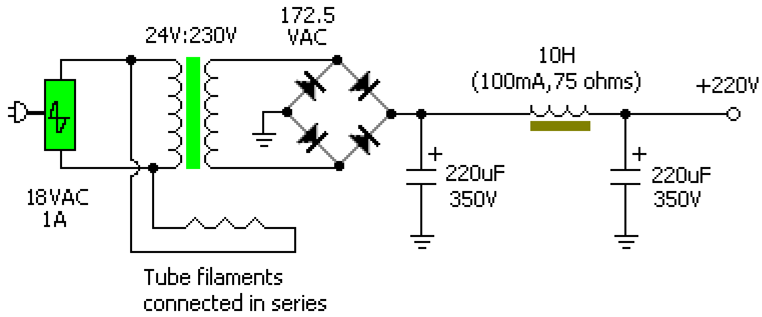


Figure 11

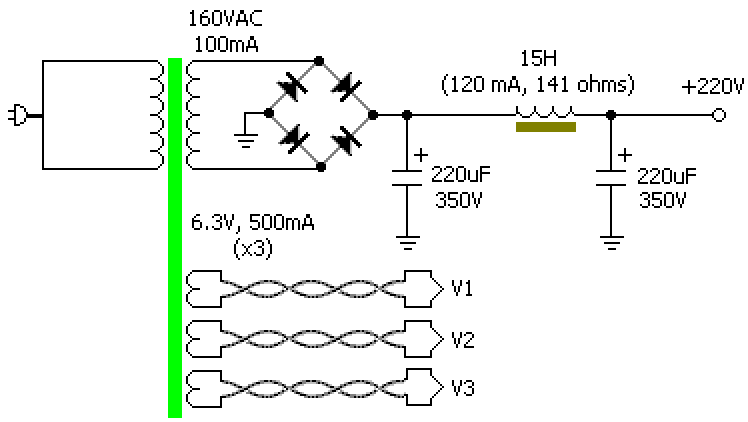
The output characteristics of the optimized MJ amplifier with feedback into 32 ohms are shown in figure 11. The amp's voltage gain is about 4, because the output impedance, while less, is still significant compared to 32 ohms. Again, the maximum output power under a White follower balance analysis is the same as for the non-feedback version in figure 4, but the distortion is lower: 1.4%. Thus, the primary effect of feedback in this circuit is to provide cleaner output power for low and high impedance headphones.

5. Revised Power Supplies

The power supply used by Johannes Chiu (figure 1) was "bare bones" and provided modest filtering. A 19VAC wallwart directly powered the tube filaments connected in series. A step-up transformer converted the 19VAC to about 156VAC, which was then rectified and filtered with a small 220uF capacitor to output 220VDC.



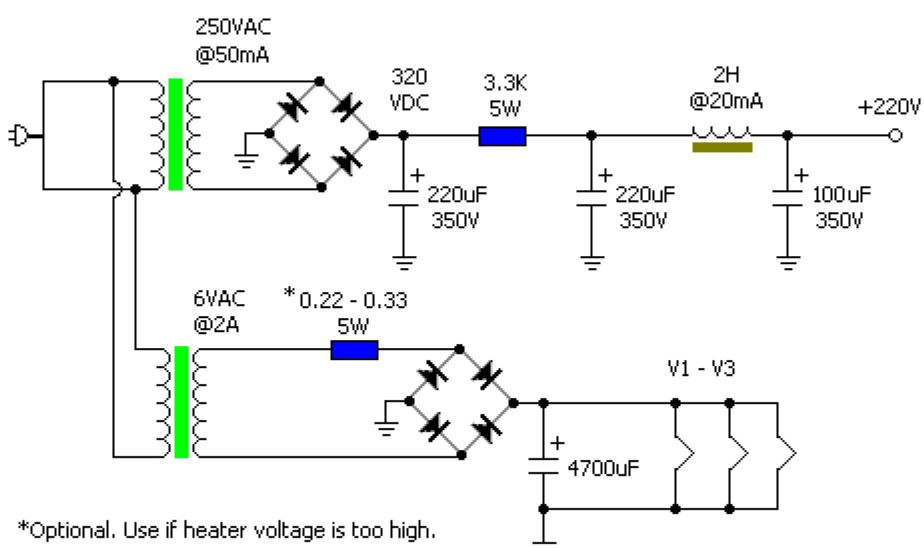
Bryan Ngiam's MJ Power Supply (1)



Bryan Ngiam's MJ Power Supply (2)

Figure 12

Bryan Ngiam and Rudy van Stratum have modified the Chiu design to reduce noise and hum. Ngiam built two power supplies (figure 12). The first supply is the closest to the original. It uses an 18VAC wallwart to power the tube filaments connected in series, a different step-up transformer and a L-C pi output filter for improved noise filtering.

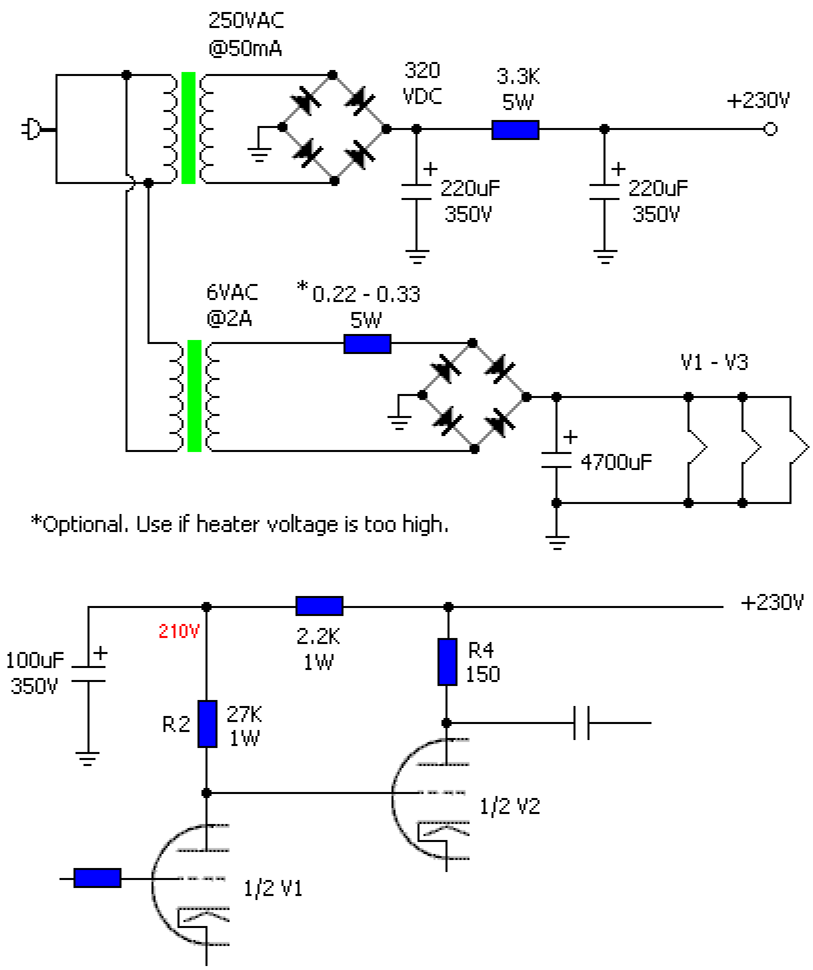


Rudy van Stratum's MJ Power Supply (1)

Figure 13

Ngiam's second supply has a single custom power transformer with three 6.3VAC, 500mA secondaries. Like the first supply, the high-voltage primary employs a pi filter, but each 6.3VAC secondary powers

the heater of one tube. Ngiam recommends twisting the filament supply wires "to reduce AC currents into the audio circuit." With either supply, Bryan recommends removing the 1 Megohm and 100 ohm resistors at the input stage.



Rudy van Stratum's MJ Power Supply (2)

Figure 14

Rudy van Stratum's designs are shown in figures 13 and 14, and incorporate power transformers that he already had in his possession: a 250VAC/50mA unit and a 6.3VAC/2A unit. In both circuits, the tube heaters are connected in parallel across the DC filament supply. Like the Ngiam supplies, the high voltage output of Stratum's first supply uses an L-C pi filter.

For DIYers who cannot find an appropriate inductor, Stratum's second circuit has several stages of RC filtering with two high voltage outputs: 230VDC and 210VDC. Originally, he powered the entire amplifier off the 230VDC tap. Later, he decided to increase the filtering to the input stage supply by adding a 2.2K resistor and a 100uF electrolytic capacitor, because most of the hum was coming from the input stage. The extra RC filtering also reduced the voltage to about 210VDC. The necessity of two high voltage taps can be avoided if the power transformer is replaced with one that outputs enough voltage to give 220VDC. It might also be worth trying to power the entire amplifier with 210VDC.

Construction

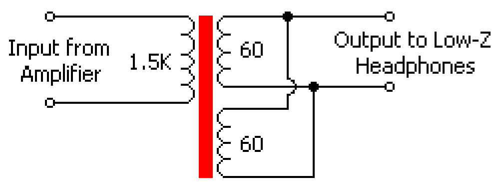
Johannes Chiu constructed the **original** Morgan Jones amplifier. He made one change: a 100-ohm grid stopper resistor to the input tube for each channel to improve stability. The EarMax uses a 19VAC, 350mA wallwart. Chiu created the simple power supply in figure 1 based on a 19VAC wallwart. The 19VAC directly powers the tube filaments connected in series, and could be rectified to

DC for lowest noise. To step the 19VAC up to 220V, Chiu used a 10V filament transformer in reverse. However, Frank Nikolajsen pointed out that a 10V filament transformer would actually give a rectified voltage of 310V. Therefore, I have scaled the transformer secondary to 14V, but recommend experimentation. I have increased the wallwart's current spec to 1A (Chiu's wallwart had a capacity of 840mA). DIYers in a country with an AC standard different from the US standard should select the transformer accordingly. I have drawn the power supply with a small value filter capacitor (there were no instructions from Chiu about this). DIYers will probably want to increase the value.

Several DIYers have found that Chiu's supply can introduce excessive hum in the amp. Bryan Ngiam and Rudy van Stratum designed powers supplies with superior filtering (figures 12-14). Bryan recommends that all input cables should be properly shielded and that a star ground should be employed whenever possible. If using a ground "strip" inside the chassis, Stratum suggests experimenting with the positioning of audio grounds on the strip to get the lowest hum and noise.

Alex Cavalli recommends a minimum power supply current rating of 220V @ 50mA for the optimized and feedback amplifiers, figuring 20mA peak per output section and 8mA for the input stages for about 48mA total. Otherwise there will be major power supply sag. He also recommends that the tubes be actual 6922/6JD8. The 6N1P is sometimes subsituted by vendors and is NOT a true substitute (see Bruce Bender's 6N1P OTL headphone amplifier for a modification of the Morgan Jones design using that tube).

Chiu used a trapezoidal-shaped chassis measuring 2" (top) x 4" (bottom) x 6.5" (length) x 2" (height), which gives a volume 1.5 times larger than the chassis for the EarMax (3.75" x 3.5" x 4.0"). The additional space is required by the filament transformer. The EarMax probably has a smaller custom transformer. For the volume control, Chiu selected a 100K dual audio pot from Radio Shack (RS 271-1732).



Hi-Lo Output Impedance Converter for Original Morgan Jones Amplifier (one channel)

The original Morgan Jones amplifier does not have enough current drive for low impedance headphones like the Grados, and Chiu did not have the schematics for the optimized versions. Instead, he experimented with a small impedance matching transformer (1.5K/60-60 from Antique Electronic Supply) for higher power transfer to his Grados. He attached the 1.5K primary to the output of the amplifier and connected the dual secondaries in parallel, which then became the output for the headphones. On the matter of selecting the impedance transformer, Chiu writes:

> People think that the output transformer is crucial in the sound, and hence must be high quality and expensive. HOWEVER, the power involved here to drive the headphones is in the milliwatts. Given that the power is so low, it greatly relaxes the requirements for the transformer. The transformer I got was about the size of a nail, and looks as cheap as many transformers found on computer modems and the like. I would also argue that the specified frequency response is perhaps 100Hz-15kHz, which most people would frown at. BUT, these specs are at full power, which could be up to 1/2 watt. At 10mW, who knows what the response is. All I know is that I tried it, and one would be hard pressed to distinguish and sonic feature that was added because of the transformer. I would encourage people to try out a few transformers. Make sure the impedances are more less correct, such that you have enough current drive, while at the same time not lose too much signal level due to the step down.

The Result

Chiu compared his original Morgan Jones amplifier to a LT1010 buffer headphone driver and a tube headphone amplifier (a "monster" 6AS7G cathode follower driven by a 6DJ8 diff amp) he had built

earlier. The headphones were a pair of Sennheiser HD420. He found that possibly "because of the higher gain of this circuit, it feels to have more punch and wallop...like moving from a cathode follower pre-amp to a mu-follower, or like adding a huge cap to your pre-amp power supply."

With a borrowed pair of Grado SR60s and the impedance converter, Chiu noted "I think the transformer will give you 90% of what is there. Maybe the bass is a tad weaker, the highs are different, but still better than straight out from those personal walkman or cd players." He tried the Grados without the impedance converter "and the transformer is way better." Chiu's final verdict on this project: "there is one thing I am certain nobody would deny if they listened to it, and that is: the amp is really fun to listen to."

DIYers building the Morgan Jones amplifier today should try the **optimized** versions, which garner favor through lower distortion, higher output power and more stable output stages that benefit both high and low impedance headphones. Although low impedance headphones were starved for current with the original Morgan Jones design, the optimized amps can drive these types of headphones to reasonable volumes without the need for an impedance-matching transformer. Low impedance headphones not only get more power from the optimized amps, but also get a flatter, more extended low frequency response. The mystery of the EarMax Pro, at last, is solved.

Appendix: Simulating the Amplifier in OrCAD PSpice

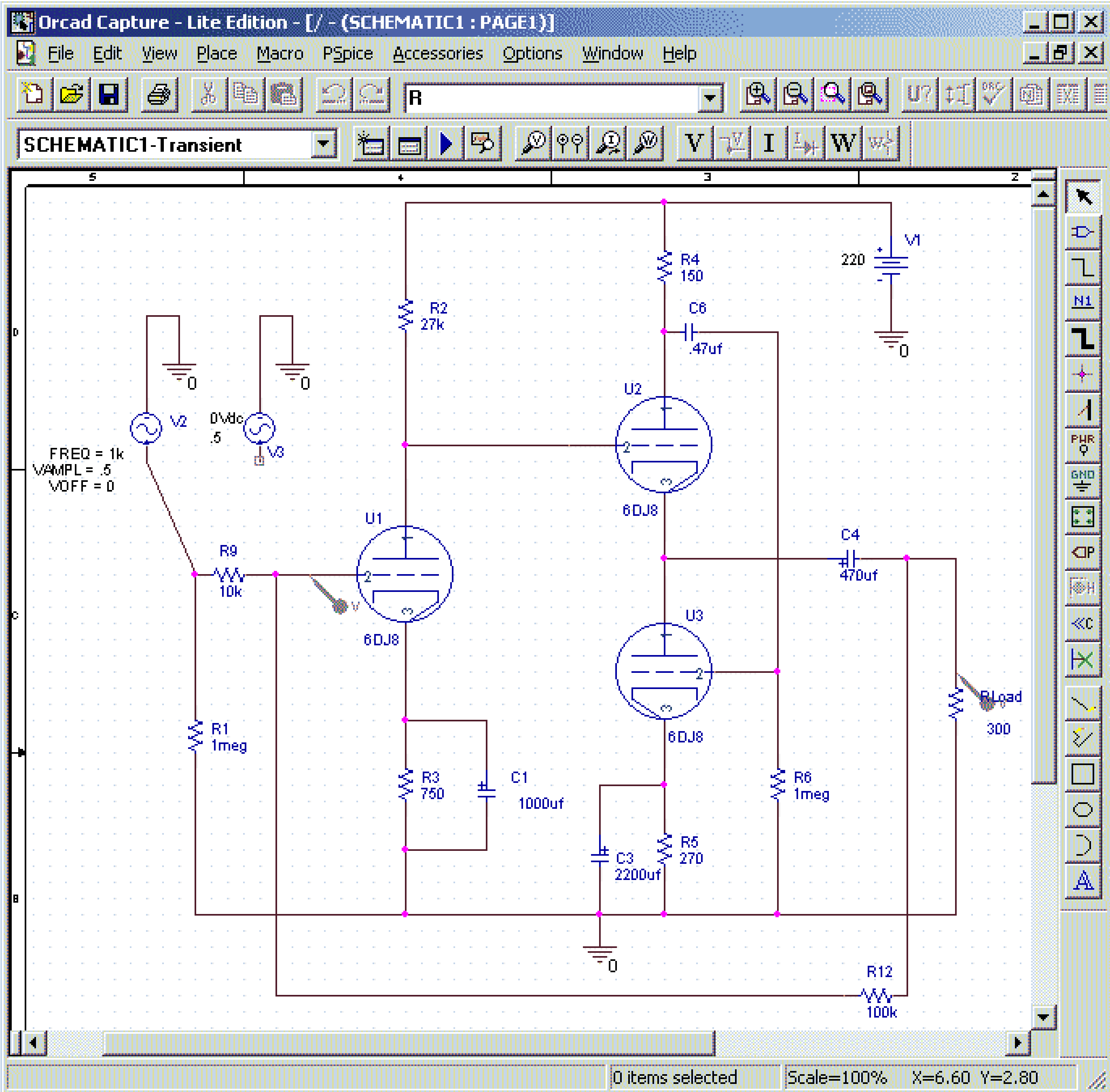
This section discusses how to use OrCAD Lite circuit simulation software to simulate Alex Cavalli's optimized Morgan Jones amplifier. OrCAD Lite is **free** and the CD can be ordered from Cadence Systems. At the time of this writing, OrCAD Lite 9.2 is the latest version. OrCAD Lite 9.1 can be downloaded from the Cadence website (a very large download at over 20M) and should work as well. There are 4 programs in OrCAD suite: Capture, Capture CIS, PSpice and Layout. The minimum installation to run the amplifier simulations is Capture (the schematic drawing program) and PSpice (the circuit simulation program).

Download Simulation Files for Alex Cavalli's Optimized Morgan Jones Amplifier

Download OrCAD Triode Simulation Libraries

After downloading mj_sim.zip and orcad_triodes.zip, create a project directory and unzip the contents of the mj_sim.zip archive into that directory. Then extract the contents of the orcad_triodes.zip archive into the <install path>\OrcadLite\Capture\Library\PSpice directory. The files triode.olb and triode.lib are libraries containing simulation models for several popular types of triode vacuum tubes, including the ones used in this amplifier. They are based on tube SPICE models found at Norman Koren's Vacuum Tube Audio Page and Duncan's Amp Pages. **Note:** heater connections are not required for any of the triode models.

The two basic types of simulation included are frequency response (AC sweep) and time domain. The time domain analysis shows the shape of the output waveform and can be used to determine the amplifier's harmonic distortion. They both run from the same schematic, but the input sources are different. For the frequency response simulation, the audio input is a VAC (AC voltage source). The time domain simulation requires a VSIN (sine wave generator) input. Before running a simulation, make sure that the correct AC source is connected to the amp's input on the schematic.



The following instructions for using the simulation files are not a complete tutorial for OrCAD. The OrCAD HELP files and online manuals include tutorials for those who want to learn more about OrCAD.

Frequency Response (AC Sweep) Analysis

1. Run OrCAD Capture and open the project file "Morgan Jones.opj".
2. In the Project Manager window, expand the "PSPICE Resources|Simulation Profiles" folder. Right click on "Schematic1-freq_resp" and select "Make Active."
3. In the Project Manager window, expand the "Design Resources|.\morgan jone.dsn|SCHEMATIC1" folder and double click on "PAGE1".
4. On the schematic, make sure that the input of the amp is connected to the V3 AC voltage source. If it is connected to V2, drag the connection to V3. By default, V3 is set to 0.5V. (Note: the tubes in the OrCAD schematic are labelled U1, U2 and U3. In the article schematics, they are referred to as V1, V2a and V2b.)

5. To add the triode library to the Capture: click the Place Part toolbar button (). The Place Part dialog appears. Click the Add Library button. Navigate to the triode.olb file and click Open. Make sure that the analog.olb and source.olb libraries are also listed in the dialog. Click the Cancel button to close the Place Part dialog.
6. From the menu, select PSpice|Edit Simulation Profile. The Simulation Settings dialog appears. The settings should be as follows:

   Analysis Type: AC Sweep/Noise
   AC Sweep Type: Logarithmic (Decade), Start Freq = 10, End Freq = 100K, Points/Decade = 100

7. To add the triode library to PSpice: Click the "Libraries" tab. Click the Browse button and navigate to the the triode.lib file. Click the Add To Design button. If the nom.lib file is not already listed in the dialog list, add it now. Then close the Simulation Settings dialog.
8. To display the input and output frequency responses on a single graph, voltage probes must be placed on the input and output points of the schematic. Click the Voltage/Level Marker () on the toolbar and place a marker at the junction of R9 and the grid of U1. Place another marker just above R_{Load} at the amp's output.
9. To run the frequency response simulation, click the Run PSpice button on the toolbar (). When the simulation finishes, the PSpice graphing window appears. The input and output curves should be in different colors with a key at the bottom of the graph.
10. The PSpice simulation has computed the bias voltages and currents in the circuit. To see the bias voltages displayed on the schematic, press the Enable Bias Voltage Display toolbar button (V). To see the bias currents displayed on the schematic, press the Enable Bias Current Display toolbar button (I).

Time Domain (Transient) Analysis

1. On the Capture schematic, make sure that the input of the amp is connected to the V2 sinewave source (the default values are: VAMPL=0.5, Freq. = 1K, VOFF = 0). If it is connected to V3, drag the connection to V2.
2. In the Project Manager window, expand the "PSPICE Resources|Simulation Profiles" folder. Right click on "Schematic1-transient" and select "Make Active"
3. From the menu, select PSpice|Edit Simulation Profile. The Simulation Settings dialog appears. The settings should be as follows:

   Analysis Type: Time Domain(Transient)
   Transient Options: Run to time = 10ms, Start saving data after = 0ms, Max. step size = 0.001ms

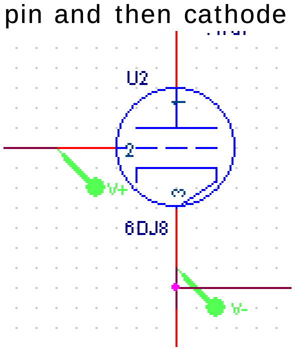
4. To display the input and output waveforms on a single graph, voltage probes must be placed on the input and output points of the schematic. Click the Voltage/Level Marker () on the toolbar and place a marker at the junction of R9 and the grid of U1. Place another marker above R_{Load} at the amp's output.
5. To run the time domain simulation, click the Run PSpice button on the toolbar (). When the simulation finishes, the PSpice graphing window appears. The input and output curves should be in different colors with a key at the bottom of the graph.
6. To determine the harmonic distortion at 1KHz (the sine wave frequency), harmonics in the output waveform must be separated out through a Fourier Transform. In the PSpice window, press the FFT toolbar button (FFT). The PSpice graph changes to show the harmonics for the input and output waveforms. The input and output curves should be in different colors with a key at the bottom of the graph.
7. The fundamental frequency at 1KHz will have the largest spike. The other harmonics are too

small to be seen at the default magnification. In the PSpice window, press the Zoom Area toolbar button () and drag a small rectangle in the lower left corner of the FFT graph. The graph now displays a magnified view of the selected area. Continue zooming in until the harmonic spikes at 2KHz, 3KHz, etc. are visible.

8. Harmonic spikes should exist for the output waveform only. The input is an ideal sine wave generator and has no distortion. To calculate total harmonic distortion, add up the spike values (voltages) at frequencies above 1KHz and divide by the voltage at 1KHz (the fundamental).

Additional Simulation Tips

- To change the value of any component on a schematic in the Capture program, double-click on the value and enter a new value at the prompt.
- The schematic included in the simulation files is for the optimized Morgan Jones amplifier with feedback. To simulate the non-feedback amplifiers without removing the feedback resistors, change the values of R8 to 100meg and R9 to 1 ohm (or 100 ohms as the grid-stop R_g).
- To measure the grid-cathode voltage of tubes (V_{gk}), use the Voltage Differential Marker (). Click the Voltage Differential Marker toolbar button and touch the probe to the tip of the grid pin and then cathode pin.



Note: simulations only approximate the performance of a circuit. The actual performance may vary considerably from the simulation as determined by a number of factors, including the accuracy of the component models, and layout and construction techniques.

For the latest updates, see the Project Addendum.

Questions or comments? Visit the HeadWize Discussion Forums.

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