

A fresh look at how to screw up a traditional design and accurately miscalculate the effect of the output impedance of the concertina drive stage. Other features not normally found in 'simple' valve designs, like a regulated supply for the 1st stage are shown but not defined. I do not know if purchasing the book, for which this article was an advert, would have explained more but I suspect the book like these typical electronics world articles from 1996 onward contained many errors and misleading comments.

The article appears to describe the amplifier pictured on the front cover of the book ( ISBN 978-0-75-062337-7) and looks as if it is based around the LEAK TL12+ output transformer but why 2 valuable mono block amplifiers were scrapped to make a stereo chassis I don't know. The amp pictured also looks like it has separate power supplies with BBC designed 120VA M188 mains transformers with the -10V tap used to reduce the heater voltage closer to 6.3V.

**In this extract from his book *Valve Amplifiers* Morgan Jones takes you through the steps of designing a valve power amplifier, and presents a prototype with a unique feature.**

# A fresh look at *valve power*



**W**hen designing your first valve amplifier, you need to be realistic.

You are not going to design a world beating amplifier overnight, and by restricting your ambitions you stand a much better chance of making something that actually works.

The design example presented here is a 10W Class AB1 push-pull 'ultra-linear' amplifier using *EL84* output valves. There are a number of reasons for this choice,

- It is cheap. If you have a 340V HT supply, this can be smoothed by 385V capacitors intended for switch-mode power supplies. In addition, the HT could be provided by a 240V isolating transformer with a silicon bridge rectifier. If any mistakes are going to be made, then it is best to make them with reasonably inexpensive components rather than expensive ones.

- There are many reasonably cheap second-hand amplifiers such as the Leak Stereo 20 or Leak TL12+, that can be cannibalised for their transformers.

- Powerful amplifiers require considerably more skill in layout and construction, and

generate bigger bangs, so it is advisable for designers to start small.

## **Bevois Valley amplifier**

This design acquired its name because the prototype was built from a pair of mono amplifiers bought for £15 – including pre-amplifier – in Bevois Valley. Once the output valves have been chosen, transformer configuration is limited, and therefore the entire output stage is fixed.

Transformer primary impedance needs to be around  $8k\Omega$  anode to anode, and with 43% taps for minimum distortion. This component might have been scavenged from a Leak, or it might even have been bought new. Either way, you will need an HT of 320V, and each valve will require  $8V_{RMS}$  for full output. Our task is to design superior driving circuitry using the following specification. Note(1)

**Low noise.** With the low noise obtainable from cd or a good vinyl recording, noise in the power amplifier needs to be undetectable. A signal-to-noise ratio of 100dB, relative to full output power, is not an unreasonable figure to aim for. This rules out pentodes and high sensitivity. Note(2)

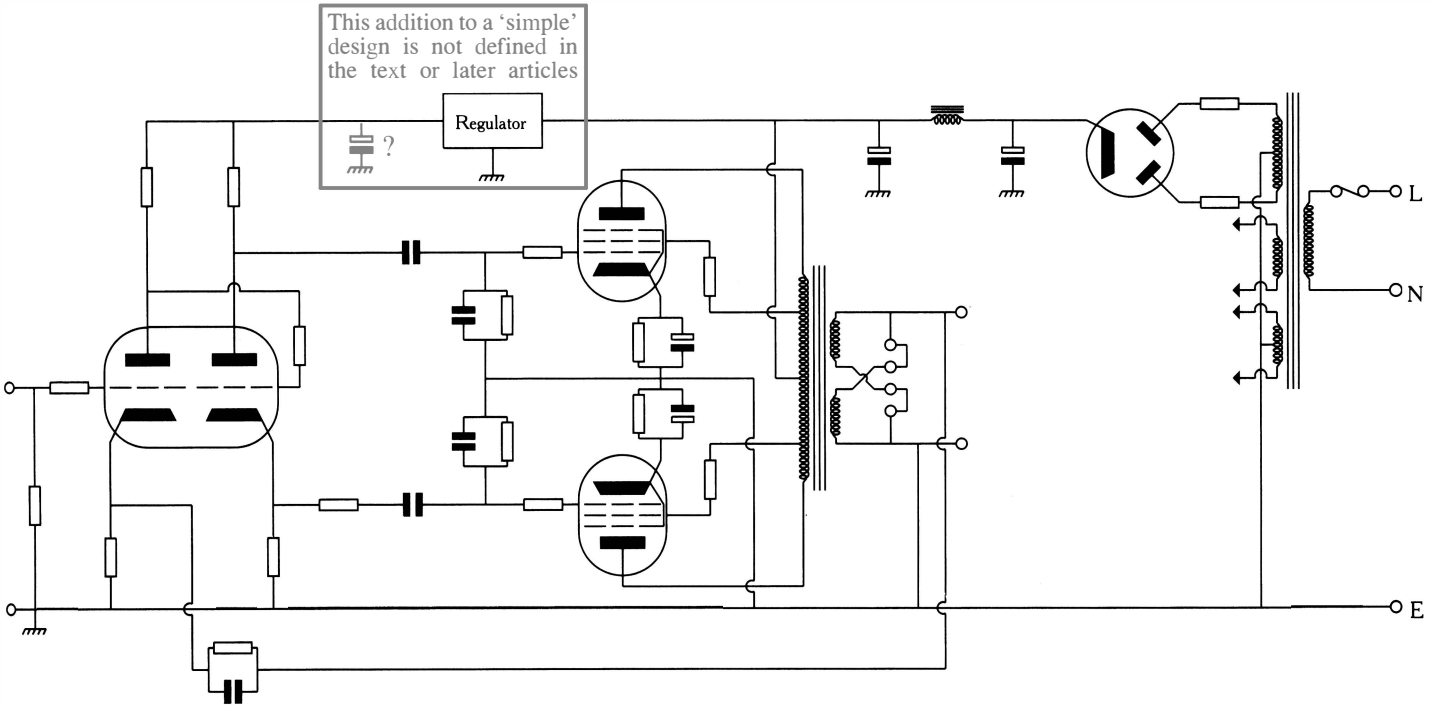


Fig. 1. Power amplifier using concertina phase splitter and featuring – possibly for the first time – cathode build-out resistor in the driver stage.

**No hum.** This implies superb standards of construction, and/or dc heaters for the input stage. Note(3)

**Stability.** To achieve good stability an absolute minimum of stages is needed. Note(4)

**Distortion.** This is a tricky topic. If you want distortion measured in parts per million, then you had better buy a decent transistor amplifier. If you think that hearing is everything and measurement is nothing, then sell the house and buy a single-ended triode amplifier. We have to be honest about this. Valve amplifiers do not measure well, but they do sound good. Presumably, we listen to music to enjoy it, so this quality is important. However, I see no reason why we should tolerate obvious engineering faults. As a result, these will be removed – although this will not imply perfection. Note(5)

**Simplicity.** Valve designs should be simple. Simple systems tend to have simple shortcomings. Additionally, they are repairable. Complex systems are built on silicon, have lots of legs, and are repeatable and disposable.

Together, these criteria demand that we use a concertina phase splitter direct coupled from the input stage without a driver stage, and we can instantly draw a circuit diagram. That this circuit is quite similar to the GEC912-Plus demonstrates that there is little new under the sun. The design rationale however is new, and to my knowledge, the cathode build-out resistor in the phase splitter is unique, Fig. 1.

Since the output valves are being driven

directly from the phase splitter, linearity of the phase splitter is paramount. The chosen phase splitter only has a gain of 1, so the input stage will also need excellent-linearity.???

Only three valves are really suitable for a concertina stage – the 6SN7, ECC82 and E88CC. We will use the E88CC. Not true

**Optimisation of dc conditions**

Because the two stages are dc coupled, the design of the two stages will be interactive. As before, the way to deal with an awkward problem like this is to garner as many facts as possible, label the drawing, and see if anything

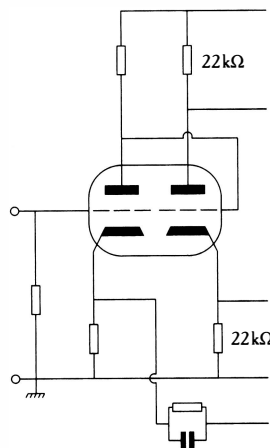


Fig. 2. Determining the operating conditions of the driver/input stage. The value 22kΩ is used because  $Z_{out}$  is approximately equal to  $R_L$ . A significantly lower value would result in excessive power dissipation.

useful appears.

Having chosen a concertina stage, we can start by labelling the anode and cathode loads as 22kΩ. This traditional value is used because  $Z_{out}$  is approximately equal to  $R_L$  and while output resistance needs to be minimised, a significantly lower value would result in excessive power dissipation, Fig. 2.

Generally, with an anode voltage of 80 to 90V, linearity of the E88CC is best when the grid is at -2.5V. Although the concertina operates under heavy feedback, it would be preferable if it were linear before feedback. As a result, it is necessary to juggle conditions such that both valves are biased with  $V_{gk} = -2.5V$ .

Since the concertina has a gain of around unity, it might be possible to arrange component values such that the signal current drawn by the concertina is equal and opposite to the signal current drawn by the input stage. This would result in zero modulation of the HT supply, and would make the HT requirements less stringent.

After much drawing of loadlines, I found that all three requirements could be met simultaneously. Additionally, they met the previously unstated requirement of being achievable with the HT available.

Balancing signal currents is the easiest requirement to satisfy. If the concertina had an  $A_V$  of unity, then for equal and opposite currents we would use an anode load in the input stage equal to the sum of the anode and cathode loads of the concertina. Since the concertina has  $A_V$  of less than one, proportionately less signal current is swung into the input stage, which means a higher value of anode load: Note(6)

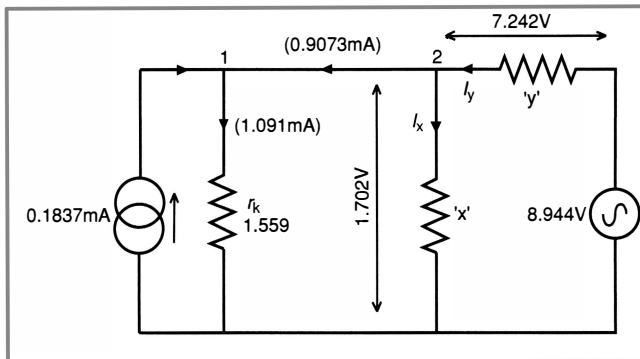


Fig. 3. Equivalent circuit of ac conditions at the valve amplifier input stage. Labelling currents and other relevant information makes calculating the cathode bias and feedback resistors easier.

Note(8)

$$R_{L(input)} = \frac{R_{anode} + R_{cathode}}{A_{V(concertina)}}$$

Note(7)

For the concertina,  $R_k=R_a=22k\Omega$  obtains, and optimum biasing is needed. Since the HT voltage is not known, gain has to be guessed. Fortunately, because the *E88CC* has such a low anode resistance, and the concertina has heavy feedback, it is possible to make a good guess at gain.

Since the *E88CC* has a low anode resistance, gain for a given value of anode load does not change greatly with HT.

As a result, an HT of 300V can be guessed at, and a loadline can be plotted to determine the gain of an *E88CC* with a 44kΩ anode load and -2.5V grid voltage. This results in an  $A_V$  of 28.75. The feedback equation can now be used to determine the gain when used as a concertina,

$$A_{V(concertina)} = \frac{28.75}{1 + 28.75} = 0.966$$

Even if had the approximation been based on  $\mu$ , the error would only have been 0.3%, indicating that the guess should be quite accurate. From this result, the value of anode load for the input stage can be calculated from,

$$R_{L(input\ stage)} = \frac{44}{0.966} = 45.53k\Omega$$

Alternatively,  $R_L$  for the input stage can be set at 47kΩ, and the concertina resistors reset to 22.7kΩ. This is a more convenient choice since the 47kΩ resistor will dissipate almost 1W, and so a 2W component is needed. This could be provided by 4×47kΩ 0.6W devices in series/parallel, or by a single 2W component.

It is inconvenient to provide non-standard values in higher ratings, whereas the concertina resistors are only dissipating around 0.33W. This can be more easily met by standard resistors.

The closest approach to 22.7kΩ is provided by 24kΩ in parallel with 430kΩ, but this

means that the 24kΩ resistor is dissipating almost all the heat, and a 0.6W component is marginal. You could use a 2W 24kΩ resistor, but the tolerance of 2W resistors is usually 2%. A better solution is to use 36kΩ in parallel with 62kΩ, which is not such an accurate approach to 22.7kΩ, but the resistors are closer tolerance. Also, the power is more evenly distributed between the components so that they are operated well within their ratings.

These choices of loads for the input/phase splitter stage will ensure equal and opposite signal currents, so we now need to arrange the correct biasing. The only way of doing this is by an iterative process.

Both stages will have an HT of less than or equal to 300V due to the voltage drop from the output stage. It is also known that each stage will have an anode voltage of 80 to 90V, for a -2.5V grid.

First draw the loadline for the concertina and find  $V_a$  for  $V_{gk}=-2.5V$ . This value is then subtracted from the HT voltage to give the voltage across  $R_k$  and  $R_a$ , and divided by 2 to give the voltage across  $R_k$ . Voltage on the grid will be 2.5V lower than this, and will equal the anode voltage of  $V_1$ .

The next job is to draw a loadline for  $V_1$  to see if the optimum anode voltage corresponds with the voltage just derived. If it doesn't, the only variable is HT voltage. Fortunately, a few iterations – by hand, not computer – found that a 285V HT voltage met all requirements, and this will be provided by a regulator.

I must say that the last determination was an incredibly tedious process. It was only carried out because in adjusting the biasing, it became obvious that it was also possible to fiddle both valves' bias voltage into balance as well. A nice computer model using real valve characteristics would solve this problem in considerably less time.

Now that the HT voltage for the two stages is known, all the ac parameters can be calculated, and the value of the build-out resistor for the concertina determined.

### Cathode bias and feedback resistance

This is easily the most complex calculation in the design of a power amplifier with negative feedback applied to the cathode of the input stage. These four factors are significant:

- Cathode bias voltage needs to be set correctly. This would normally be a trivial application of Ohm's law, but in this case the bias current flows through the cathode resistor and the feedback resistor.
- The input valve generates a feedback current through the cathode resistor, in addition to any current sourced from the output of the amplifier.
- Ratio of the two resistors needs to be set so as to obtain the desired negative feedback.
- As far as ac is concerned, the cathode resistor is shunted by  $r_k$  of the valve.

Now, with the restrictions specified, it should be possible to label a diagram and derive some equations. Since 2.5V bias on the cathode is needed, and anode current is 190V/47kΩ, the total resistance to ground from the cathode must be 618.4Ω.

Anode signal swing for full output is 8.636V rms. This means that the anode signal current must be 8.636V/47kΩ=0.1837mA rms. This current also flows in the cathode circuit and will develop a feedback voltage across any unbypassed cathode resistor.

If input sensitivity of the amplifier is to equal 2V rms, and we know that the unmodified sensitivity is 298mV rms, the feedback voltage required at the cathode will be 2-0.298V, which is 1.702V rms.

For full output of 10W, the signal at the output of the amplifier will be 8.944V rms. This means that there will be 7.242V rms across the feedback resistor. Since  $r_k$  will shunt the cathode resistor at ac, it is necessary to find  $r_k$ :

$$r_k = \frac{R_f + r_a}{\mu + 1}$$

Using this equation, you will find that  $r_k=1.559k\Omega$ .

Assume that the output of the amplifier is a true Thévenin source driving the network through the feedback resistor 'y'. The valve's own feedback current is represented as a Norton current source, and the cathode resistor 'x' is shunted with  $r_k$ . Note that Fig. 3 is an ac diagram.

Our first observation is that there is a resistor of known value  $r_k$  with a known voltage of 1.702V across it, so current through it is 1.091mA.

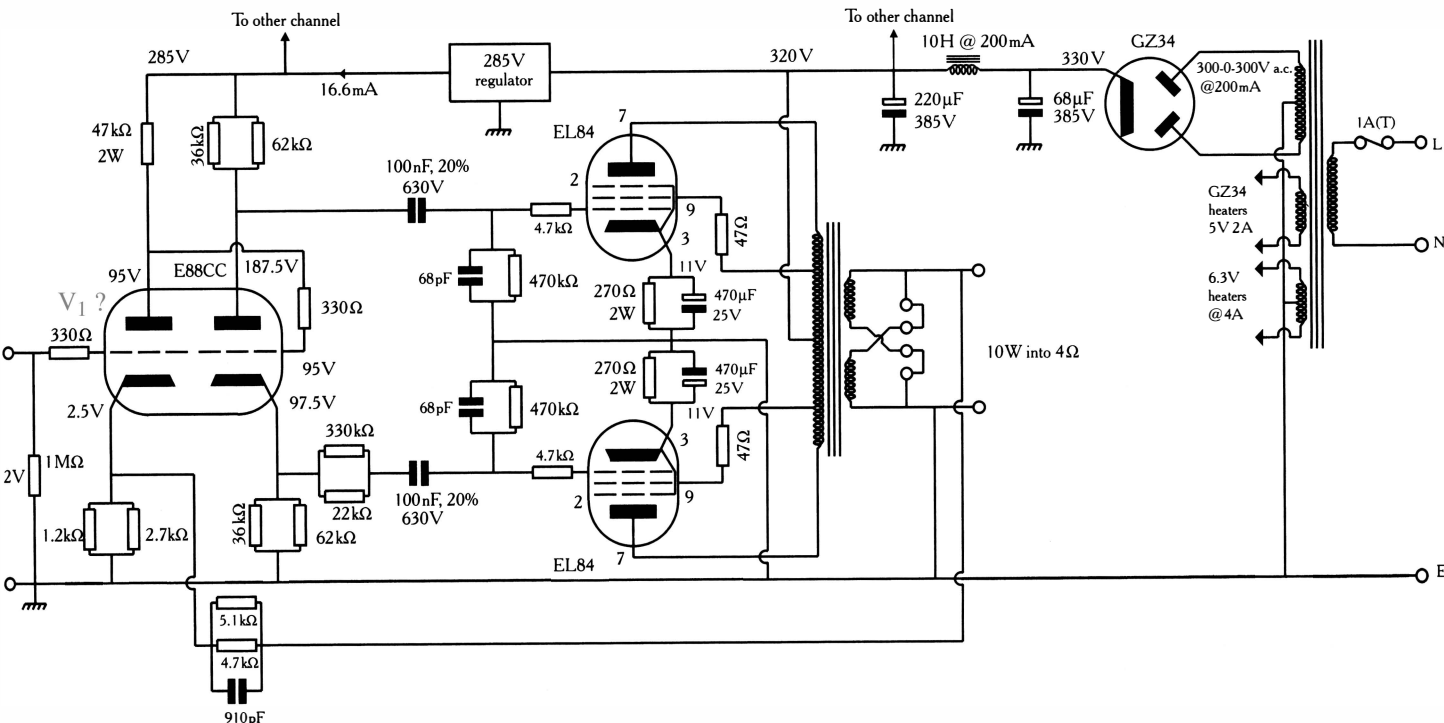


Fig. 4. Practical 10W Class AB1 push-pull amplifier using EL84 output valves and featuring the 'ultra linear' output stage.

We can now see that node 1 has two known currents flowing through it, so we can find the third, using Kirchhoff. If there is 0.1837mA flowing into the node, but 1.091mA leaving it, then 0.9073mA must be supplied by the other node. Moving to node 2, you can see that any current coming into the node must be supplied by  $I_y$  and that this splits through the resistor 'x', and to node 1. Formalising this;

$$I_x + 0.9073 = I_y \quad (\text{Eq. 1})$$

You can use Ohm's law to make statements about the currents in resistors 'x' and 'y':

$$I_x = \frac{1.702}{x} \quad (\text{Eq. 2}) \quad I_y = \frac{7.242}{y} \quad (\text{Eq. 3})$$

The final restriction is the dc restriction, which says that x and y in parallel must give 0.6184kΩ.

$$0.6184 = \frac{xy}{x+y} \quad (\text{Eq. 4})$$

The way to solve the equations is to substitute the second and third equations into the first:

$$\frac{1.702}{x} + 0.9073 = \frac{7.242}{y}$$

Rearranging, and simplifying:

$$7.982x - 1.876y = xy.$$

It is now possible to substitute this into the fourth equation, and solve it to give the ratio  $y=2.953x$ . Substituting this ratio back into the equation yields  $x=828\Omega$ . Using the ratio you therefore need 1.2kΩ in parallel with 2.7kΩ for the cathode resistor, and 4.7kΩ in parallel with 5.1kΩ for the feedback resistor.

**Distortion consideration**

Some of  $V_1$ 's cathode current is now flowing through the output transformer, and it might be thought that this would cause distortion. Assuming that dc resistance of the transformer secondary winding is negligible, the current flow will be,  $V_1$  ?

$$2.5V/2.44k\Omega \approx 1mA.$$

Now the current turns ratio of the transformer is 31.6:1 – secondary-to-primary – so 1mA of dc flowing in the secondary is equivalent to 31μA out-of-balance dc flowing in the primary. Compared to 40mA each side, this is negligible, since output valve balance is highly unlikely to be as good as this.

All component values for the driving circuitry are now known, so the values for the output stage can be determined. The EL84 is allowed a maximum grid leak resistor of 300kΩ with grid bias. As cathode bias is being

used however, this can be increased to 470kΩ. A 0.1μF coupling capacitor is necessary, which should be polycarbonate, or preferably polypropylene, with a rating of 400Vdc or more.

A value of 4.7kΩ is traditional for grid stopper resistors on the EL84. They may not be needed, but it seems sensible to fit them just in case. A resistance of 47Ω in series with  $g_2$  is alleged to reduce distortion while reducing peak power. I have not tested this, so fitting them is a matter of personal choice. The Mullard circuits included them, but the Leaks didn't.

From the data sheet, the cathode bias resistor should be 270Ω and dissipate 0.45W. Resistors rated at 2W are commonly used here, but a 15W chassis mounting metal clad type with tabs is a much better choice. This is because an electrolytic capacitor is going to be placed very close to this resistor, so it needs to be kept cool. The resistor also provides convenient tags for anchoring the capacitor.

The cathode bypass capacitor should be 2200μF for a 1Hz cut-off. But as discussed earlier, this value would cause additional problems; a good compromise is 470μF 63V. A rating of 63V may seem excessive, since it will only see around 11V, but the higher voltage component will have a lower effective

Note(9)

series resistance. This becomes significant when you are trying to bypass the 90Ω cathode resistance of the valve.

Because there is only one RC network plus the output transformer in the entire amplifier, low-frequency stability will not be a problem.

High frequency stability is not assured, and so this should be investigated.

The input stage has its basic sensitivity reduced from 298mV to 2V, which corresponds to a gain reduction of 6.71. From this, you can calculate the new  $r_a$  for the stage:

$$\frac{\mu R_L}{R_L + r_a} = 6.71 \frac{\mu R_L}{R_L + r'_a}$$

Solving this, and using  $r_a=5k\Omega$ , gives  $r'_a=302k\Omega$ , in parallel with  $R_L=47k\Omega$ ; this gives  $r_{out}=41k\Omega$ . You will find that applying global negative feedback invariably causes  $r_{out}\approx R_L$  for the input stage.

The concertina has 3.2pF of Miller capacitance. Allowing for strays, 5pF is a reasonable total value. In combination with 41kΩ, this gives a cut-off around 780kHz. The output stage will have an input capacitance that loads the 22kΩ output resistance of the concertina,

so should be determined. Although the EL84 is a pentode, it will still have Miller capacitance, albeit greatly reduced, so this should be included in the calculation.

You can find anode gain of the output stage by calculating voltage across the 8kΩ transformer primary for 11W. It is known that 16V<sub>RMS</sub> from grid to grid is needed to drive the stage. This gives a gain to the anode of 18.54. Since  $C_{ag}=0.5pF$ , this would result in a Miller capacitance of 9.8pF. Unfortunately, this value of  $C_{ag}$  is for the pure pentode connection. On the other hand, we will be using the 'ultra-linear' connection, where  $g_2$  does not stay at a constant potential. This means that allowance must be made for the Miller effect from  $C_{g2}$ . Unfortunately, the Mullard data sheet does not give a value for this, so it is probably wise to allow another 10pF.

Adding these to  $C_{in}=10.8pF$  produces a total input capacitance of around 35pF, including strays. Driven by the concertina, this gives a cut-off of 200kHz, and is the dominant pole. To achieve high-frequency stability, slug the input capacitance of the output valves, and not the concertina, as is usually done. This will

have the advantage that additional capacitance will swamp variations in the capacitance between valves, improving high-frequency balance. Shunt capacitors of 68pF across the EL84 grid leak resistors will slug this pole to 72kHz. It is now possible to draw a full circuit diagram of the amplifier, with component values, Fig. 4.

Note(10) ■

**Further reading**

Colloms, M, 'High performance loudspeakers', 3rd edn. Pentech Press, London, pp. 188-206, 1985.  
 Futterman, J, 'A practical commercial output transformerless amplifier', *Journal of the Audio Engineering Society* October 1956.  
 Hedge, L B, 'Cascade AF amplifier', *Wireless World*, 283-87, June 1956.  
 Mullard, 'Tube Circuits for Audio Amplifiers', reprinted by Audio Amateur Press, Peterborough, New Hampshire, 1993.  
 Williamson, D T N and Walker, P J, 'Amplifiers and superlatives', *Journal of the Audio Engineering Society*, 2(2), 75-80, 1954.

(1) This the project is based around finding a pair of 12W ultra linear transformers on a chassis which happen to be designed for EL84 valves. For the EL84 there are numerous transformer configurations that could be used. The task of "designing a superior driving circuit" is not achieved.

(2) If the sensitivity is about 1V-2V<sub>rms</sub> most traditional valve designs could manage >80dB and this is more than adequate. I assume the comment about not using pentodes refers to the 1st stage valve only.

(3) Low hum can be readily achieved with well balanced centre tapped heater windings and twisted wires which is more a standard practice than 'superb construction', d.c. heaters are a complication that is not needed in most valve power amps.

(4) Stability will be dictated by the output transformer and it is surprising that no consideration is given to this point. The phase shift due to the output transformer along with open loop gain and amount of negative feedback will determine the stable gain and distortion, using a LEAK transformer in a lower sensitivity amplifier makes thing simpler but other transformers may need careful consideration.

(5) Distortion will be mainly due to the output stage and amount of negative feedback that can be applied. The quote "I see no reason why we should tolerate obvious engineering faults" must refer to the perceived problem with the concertina drive stage output impedances - see 'letters' below.

(6) An E88CC is not most linear when 'the grid is at -2.5V' or at an  $I_a$  of about 4mA. Whatever valve is used, and there are more than 3 to choose from, the normal reference would be to  $I_a$  rather than  $V_g$ . There is no need to draw numerous load lines to find that the concertina resistors should be about half the value of the 1st stage anode resistor for correct biasing of these d.c. coupled stages when connected to a common HT supply.

(7) The assumption that Making the 1st stage and concertina  $I_a$  the same d.c. value will prevent modulation of the HT supply is complete rubbish, this can only be achieved when the 1st stage load resistor is the same value as each concertina resistor and the  $V_{ak}$  of both valves is the same, then the a.c. current in each valve will be the same but opposite. The complexity and accuracy even the need for these calculations is not required.

(8) Again these calculations are too complex and again wrong. There is no consideration as to how the amount of negative feedback was determined, especially with respect to the phase change in the output transformer at high frequency.

(9) A 2W resistor is more than adequate for the EL84 cathode bias. The bypass capacitor should be placed close to the cathode pin but the resistor could be almost anywhere else, the comment about heating the bypass cap is again misleading and erroneous as is the statement that a higher voltage rating capacitor will have a lower ESR is simply not true.

(10) This whole discussion about balanced dominant poles is misleading, the best practice for low intermodulation distortion, the IMD that valve amps are best know for, is to place a pole at the 1st stage and ensure that subsequent stages have increasing wider bandwidths. The output transformer usable BW determines all that precedes it. Were the concertina also the 1st stage the pole could be placed at the inputs of the EL84s and incorporate some of their capacitance but in this design that is not the case. Making a valve power amp where the even harmonics are exactly cancelled is not making a valve amp at all. www.keith-snook.info

But there is a big *but*. Recently I built a d-to-a converter using Crystal Semiconductors' latest 20 bit device, the CS4329. My power supply uses LT1085CT and LT1033CT regulators. The power supply capacitors are Rubycon Black Gate FK and NX types and Sanyo Os-Con types. These are 'exotic' components and rather expensive.

The overall sound performance with these capacitors is so much better than a LT1085/LT1033 based power supply using good quality and normal priced Elna RSH capacitors. Using Keith Jarrett's 'Köln Concert' as reference music, you can easily tell which power supply is 'playing'. The soundstage is so much improved. Jarrett's piano really 'sings'. In my opinion there is no doubt that the BGs and Os-Cons improve the sonic overall performance of a system; my ears are me tell me so.

Keep up the good work,  
**W. de Haan**  
Leiden, The Netherlands

## Agreeable distortion

'Valve sound' is essentially subjectively agreeable distortion. An analogy is the measurable sensation of travelling in a vintage Bentley rather than in a modern mid-range Ford, which is noticeably better in most respects – if not at all.

However, there is one difference – cost. Preferred output valves cost upwards of £20 each.

Morgan Jones' excellently presented article – Jan '96 – exhibits at least one flaw, however, as many of the resistance values arrived at by paralleling are within a fractional percentage of standard values. For example 330kΩ in parallel with 22kΩ is 20.625kΩ. A near value in the E96 range is 20.5kΩ – less than 1% off. If cost is no object, this is the way to do it.

I worked with valves for many years and came to accept that their characteristics varied widely from part to part. Anybody who used the EF50 will remember this. There is no point in attempting exact design where key parameters can differ by as much as 20%.

In any case, the principal feature of valve amplifiers is that they include an output transformer. If one takes a good solid-state amplifier and includes a 1:1 output transformer within the feedback loop one will achieve much the same effect.

Of course, valve and solid-state amplifiers driven near to or past saturation will sound different, but if one is any sort of a purist this is not a region in which one operates. Vast power capability overkill is an essential feature of hi-fi usage.

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Sutton  
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## Valve misunderstanding

As a designer of valve amplifiers since 1950 I have read with some disbelief the article by Morgan Jones in *Electronics World* January 1996 and the subsequent correspondence in the February and March issues. Both Morgan Jones and Frank Ogden seem to not understand the operation of the concertina phase splitter.

This circuit does not have the alleged difference in frequency response at the anode and cathode terminals. If the anode and cathode outputs are analysed separately then, of course the anode output resistance is high and the cathode output resistance relatively low as shown by Morgan Jones in his March 1996 letter. However when both outputs are loaded simultaneously with equal capacitances the output voltages remain equal throughout the audio frequency range.

This can be understood intuitively since the anode current is the same as the cathode current so when the two impedances are equal (i.e. equal resistances and equal capacitances) then the output voltages must be equal at all frequencies. It is obvious that the tendency for the anode voltage to decrease more rapidly as the frequency is raised is fully compensated by the by-passing effect of the cathode loading capacitance.

The circuit behaves as if the

output resistance at both ports is much the same as the source resistance of a cathode follower using the same valve and cathode load resistance. It can be shown that the effective output resistance used to determine the frequency response at both outputs is,

$$R_o = \frac{r_a R_L}{r_a + R_L(\mu + 2)}$$

Needless to say the 'build-out' resistor spoils this inherent wide-band balance of the concertina phase splitter.

There is another error in the Morgan Jones article in the January 1996 issue where he attempts to balance the signal currents of the input stage and the concertina. The concertina signal current is approximately grid voltage divided by cathode resistance thus the anode load of the input stage should be roughly equal to the concertina cathode resistance and not cathode resistance plus anode resistance as stated.

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Belfast

## Reference

1. 'Radio Designer's Handbook' F. Langford-Smith p.329 Fourth Edition. Published by *Wireless World* 1953.

## Shame about the error

At present I am particularly interested in the subject of valve audio amplifiers. While not having sufficient detailed information on valve characteristics at hand to check all the calculations in the January's valve power amplifier article, I was disappointed to find a clear error in the calculation of the values for the feedback resistor and the input stage cathode resistor. While the circuit diagram indicates a 4Ω output load, the calculation is based on 8Ω.

Speakers with 3Ω or 15Ω coils were common before the advent of the 8Ω speaker. This made a dual secondary winding on the output transformer popular, giving an output impedance of 4 or 16Ω. My calculations show that with a 4Ω load, a cathode resistor of 964Ω is required, and a feedback resistor of 1728Ω; with 16Ω loads they should be 753Ω and 3456Ω respectively. The method of calculating the feedback capacitor was not explained, but this should be less critical than the resistor values, and it should be adequate to adjust this proportionately.

The required values could be obtained in the case of each resistor

by using two parallel resistors of standard values as in the article, values as follows:

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## I can't hear you

For once I find myself in agreement with Ben Duncan, on the issue of the suitability of Windows (Review of Micro-CAP V, *EW+WW* Sep '95). It seems absurd that professional pc users should be saddled with a software package that appears to be a re-invention of an operating system designed in the early seventies for children. Windows is ok for the novice user, but without much doubt anybody with a modicum of experience with a standard keyboard would find it more efficient than a mouse.

Windows is, in my opinion, poorly documented, slow, cumbersome and not very logical, and a running joke among my computer literate friends. Unfortunately it is difficult to get by without it, and maintain compatibility.

To load and run Windows at an acceptable speed requires no less than a 486 – most pcs in our department are 386s – at least 8Mb of ram and a large fast hard disk. This hardware is only now becoming acceptably cheap, but Microsoft would like us to move up to Windows 95 with even greater demands on our hardware. To quote one John McCormick, "Why would anyone in their right mind use Windows for anything? You can always buy a slower computer if yours is too fast!" (from "It's not a Bug, It's a Feature!" by David Lubar).

Unfortunately that is the end of good news for Ben. In his article "Simulated attack on slew rates" (*EW+WW*, April '95) Ben boldly states on p. 307 that "...the headroom is demonstrably safer for drive units and ears alike – no matter how counter-intuitive this seems" in the course of his justification of very high slew rates and the reproduction of "...music transients above 165V...". Ben opened the piece by outlining the high frequency nature of the sound "during an Iron Maiden gig" engineered by a colleague.

*New Scientist* reports (p5, 27 Jan '96 No. 2014, Australian edition) that "rock concerts are more likely to damage your hearing than listening to a personal stereo or going clubbing", according to French hearing specialist Christian Meyer-Bisch. This conclusion is the result of a study of 1364 people, and

## Best rf article '95

Entries for this challenge are currently being evaluated. We hope to be able to make an announcement about the winner in next month's issue.