POWER AMPLIFIERS
with valves

an approach and
a practical circuit

by
Claus Byrith

1. INTRODUCTION

A couple of years ago I met Per Lundahl of the Swedish company Lundahl Transformers and his father at an exhibition in connection with an Audio engineering Society Convention, and I tried to persuade him to make me a pair of output transformers with some additional secondaries for experiments, I wished to conduct. He was not enthusiastic, but we had a fine conversation, nevertheless, about valve amplifiers, of course. After some time, I said to him that if he would fulfil my wishes, I would supply him with a paper concerning the matter and a practical construction that could be built by amateurs, resulting in a 30 W amplifier where high quality goes hand-in-hand with simplicity and modest costs. To make a long story short he agreed, and here I am left to keep my promise, and I am beginning to realise that despite the fact that the matter is straightforward, it is not easy to pass yesterday's knowledge on to the audio enthusiast of today in a short, simple, digestible and yet satisfactory way, but I shall try my very best. I am old enough to have experienced the evolution of valve amplifiers since the mid fifties but am, at least so I am told, still in full command of my faculties.

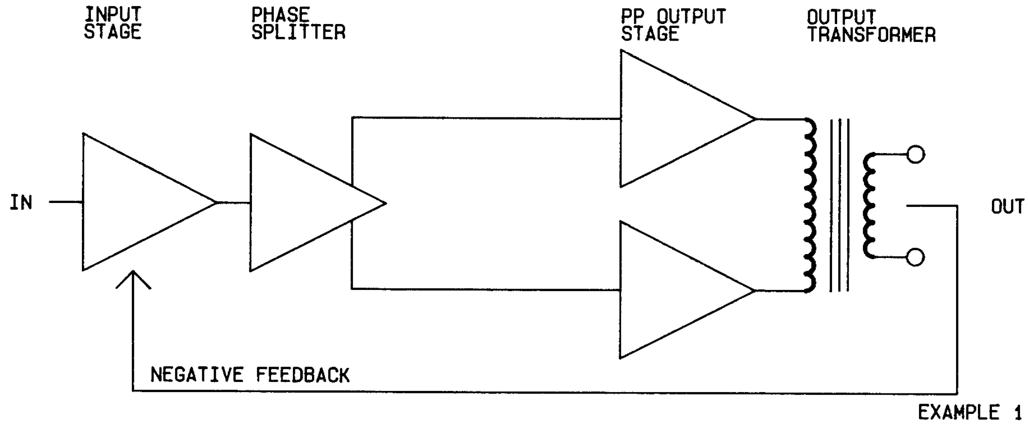
The last decade has seen a growing interest in valve amplifiers, which is not only due to nostalgia but more to the fact that a good valve amplifier sounds different from most solid state amplifiers, and despite the fact that the measured performance cannot compete with modern amplifiers, they sound very good and they do not expose you to the same degree of listening fatigue as many solid state amplifiers, truly or falsely!, are accused of. I shall not try to describe the sound. The fact that you have got so far in this paper indicates that you understand what I am talking about.

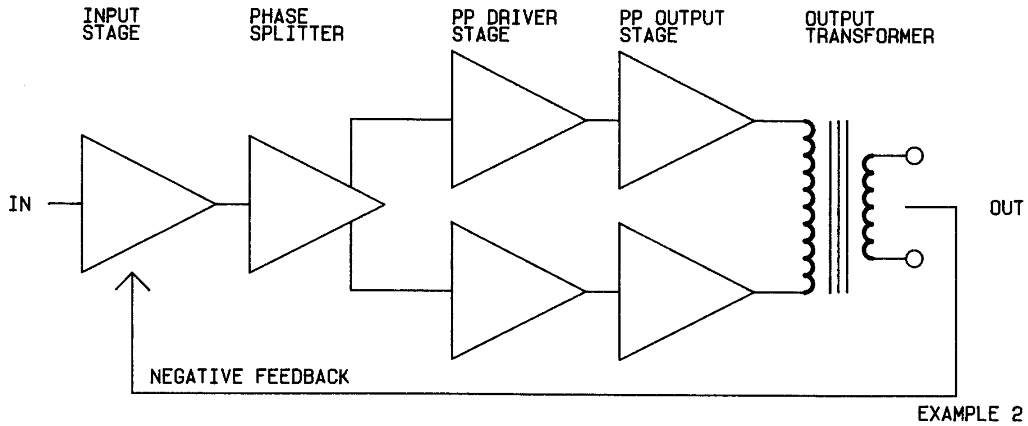
Often involved in discussions about this topic, I am forced to realise that much of the knowledge that was common goods 40 years ago has disappeared and what remains is a conglomerate of distorted facts and a complete lack of understanding of how these facts are interlinked and of their relative importance. It is as if it were seen in a distorting mirror in the hall of mirrors in an amusement park. Parts of the image are magnified out of all proportion and other parts of maybe even greater importance are simply not there!

I shall now try to improve your understanding.

2. THE LAYOUT AND THE BUILDING BLOCKS

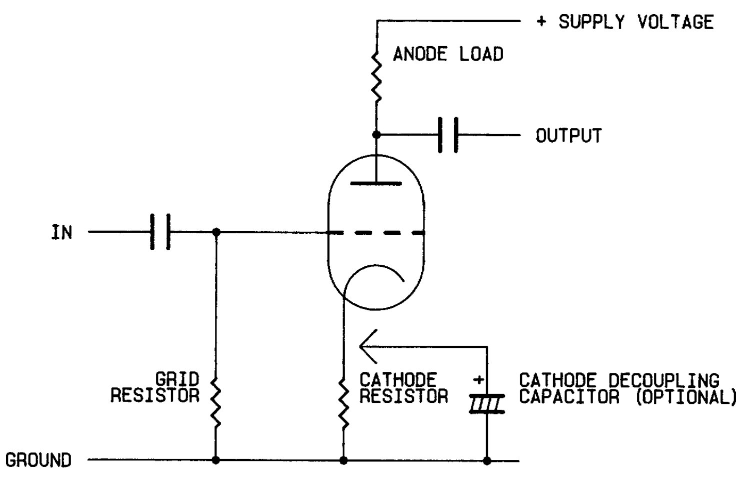
A push pull valve amplifier almost invariably follows one of the two schemes shown in block diagrams below:





Very few designs differing from these two have proven fit for real life, the Quad II being a notable exception (you will find the diagram in the appendix together with a brief explanation).

We will take a closer look at the input stage, because here we can learn a great deal about how a valve works. The input stage is normally a cathode coupled amplifier as shown below.



Electrons emitted from the heated cathode are attracted to the positive anode, passing through the grid. To obtain a working point where the valve operates in a linear way, the valve must be biased with a negative voltage on the grid with respect to the cathode. Electrons are now repelled from the grid and fewer pass. The negative biasing turns to some degree the valve off and bias is normally in small signal stages achieved by the cathode resistor. A current through the valve causes a voltage drop over the cathode resistor and so the cathode becomes positive with respect to ground. The grid resistor keeps the grid at ground potential and therefore negative with respect to the cathode. When anode voltage is applied, current starts to flow and cathode voltage rises, causing negative bias on the grid, which again lowers the current through the valve, and an equilibrium is quickly established. From this equilibrium, the working point, the current through the valve and the voltage drop over the anode load resistor can be controlled by superimposing the input signal on the grid bias. A positive pulse on the grid causes current to rise and anode voltage to drop, producing a negative pulse on the anode and vice versa – the stage inverts the signal.

What happens to the cathode voltage? Simple enough, a positive input pulse causes the voltage drop over the cathode resistor to rise, so the cathode voltage must also rise, but the rising cathode voltage counteracts to some degree the effect of the rising grid voltage. The cathode voltage tries to follow the grid voltage, and because it is the rising difference between cathode and grid voltage that causes the anode current to rise, this will be diminished by the increasing cathode voltage. The stage is said to be under influence of negative feedback (NFB). This may be wanted, but not always, and it can be avoided as shown in the diagram by connecting a condenser across the cathode resistor. The condenser should have a capacitance big enough to keep the cathode voltage constant, down to the lowest frequency of the applied signal.

We have now seen two of the main differences between a transistor and a valve: The valve is brought to its working point by being turned off to some degree by a bias voltage. – The transistor approaches the working point by being turned on by a bias current. The valve is controlled by a voltage, the transistor by a current.

As long as positive peaks in the signal does not exceed the bias voltage, the only load to the signal is the grid resistor and some capacitive loading by the electrodes plus stray capacitances. We shall in a moment return to the capacitive loading, but we have not yet seen what happens when positive peaks in the signal are of a magnitude that makes the grid positive with respect to the cathode. In this case the grid will no longer repel electrons but attract them instead and a grid current will start to flow. This current is supplied by the signal which will now be heavily loaded in its positive peaks, and if the signal source has an output resistance greater than zero – and it always has – the positive peaks will be distorted even before they are amplified. The valve may still amplify correctly the signal on the grid but the result is of course unusable. So we now know that the positive going peaks in the signal must never exceed the bias voltage.

The grid resistor is normally 0.5-2 Mw. There is a maximum value not to be exceeded. If the value is too high, electrons can pile up on the grid, making it more negative than we expect, rendering the working point different from our calculations and sometimes unstable.

The grid resistor is hardly loading our signal, but what about capacitances? In the valve tables, from which relevant pages are given in the appendix, the capacitances, grid to anode and grid to cathode are stated, and this capacitance must be charged and discharged by the signal applied to the grid. Suppose a stage amplifying 25 times. An input voltage drop of 1 volt will then cause an anode voltage rise of 25 Volts, which means that the signal charges the anode-grid capacitance not to 1 Volt but to 25 Volts. Seen from the grid the capacitance is therefore not just the anode-grid capacitance but this capacitance multiplied by the amplification of the stage.

If our valve is half a double triode ECC83, where Ca-g is 1.6pF, this capacitance acts as if it was 25 x 1.6 = 40pF + strays. The increase of the apparent capacitance with increasing gain is known as the Miller-effect and as we shall see this can have an alarming effect on the performance of an amplifier.

In pentodes, such as the EF86, we find two more grids. The second is normally held at a positive potential, but a capacitor to ground keeps it free from signal and so it screens the control grid from the anode. It is called the screen grid. The Miller capacitance of a pentode stage is normally only about 1/10 of that of a triode stage with the same gain. But pentodes are noisier than triodes, because the beam of electrons emitted from the cathode is divided between the screen-grid and the anode. This noise, known as partition noise, disappears when we connect a pentode as triode, i.e. screen-grid and anode are strapped together. Electrons are of course still parted, but the noise cancels when the two currents are added again in the anode load. By triode connection, the screening effect of the second grid is of course lost.

By these explanations, I have just scratched the surface. I have not explained how gain is calculated, nor have I explained how a suitable working point for a given valve is found. I have not looked into how output resistance of a stage is calculated or how output resistance is affected by feedback. Just how important these matters are, they are far beyond the scope of this paper. It is however necessary for you to know, even without explanation, that the output resistance of a stage is equal to the anode load in parallel with the internal resistance of the valve seen from anode, and consequently the output resistance can never exceed the anode load resistance. You must also know that the way feedback affects the output resistance depends on how the feedback is derived and on how it is injected. The negative feedback (NFB) caused by the unbypassed cathode resistor is current derived and current injected. It raises output resistance. The feedback from the output of the entire amplifier to the first stage is voltage derived and normally current injected. This NFB, known as global, lowers output resistance of the amplifier but raises output resistance of the first stage. Negative feedback decreases the amplification, as we have seen, and it affects input and output resistances but it also affects distortion and bandwidth. 20 dB (10:1) of global negative feedback reduces distortion to 1/10 and output resistance to 1/10. It also reduces any deviation from linear frequency response by a factor 10, all compared to the performance without NFB.

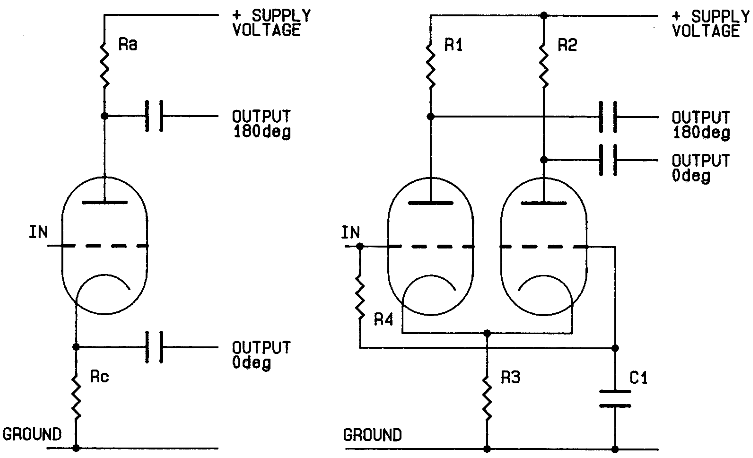
This is a very important issue and I strongly recommend further studies from relevant books.

The global NFB is often subject to discussion in certain audio circles and it is often heard that global NFB is harmful to the overall performance of an amplifier. May I remind you that this type of feedback was invented by telephone engineers in the thirties. Their objective was to keep the gain of an amplifier constant. Suppose you have an intercontinental landline with, say 30 amplifiers along the line to compensate for losses, and their gains are fluctuating maybe by 1 dB. It is obvious, that the level at the end of the line is totally unpredictable and that endless adjustments have to be made. NFB eliminate these fluctuations so the engineers could spend their time with more interesting things than gain adjustments.

The stability of the stereo image is very much dependent on an equal and constant gain in both channels. Even the smallest gain fluctuation blurrs the image, especially when it comes to reproduction of depth. That global NFB is our best tool for stabilising gain is often overlooked, despite the fact that everyone agrees on the importance of gain stability. Bear also in mind that a valve amplifier is more prone to gain fluctuations than a solid state amplifier. Small movements of the electrodes, due to the heating, change the properties of a valve to at least some degree. You will find more about NFB in the chapter about the amplifier I want to build and in the chapter about measurements.

If the two output valves are EL34^{s} each of them will need a voltage swing on their control grids of about 25 Volts$_{eff}$ and these two signals must be equal and 180^{0} out of phase. After the input stage we will need a phase-splitter. Its function is crucial for the performance of the amplifier, because perfect balance of the signals to the output valves is minimising odd harmonic distortion. I do not feel that in today's discussions enough attention is paid to this important stage. It seems just to be assumed that it does its job. As we shall see, it is not that simple.

Phase-splitting can be done in several ways. The two diagrams below show the two most commonly used (and best!).



The first is extremely simple. It is called the splitter with equal loads or the concertina splitter. It works as follows: A positive going pulse on the grid increases the current so the voltage drops on the anode but increases on the cathode. We get a negative pulse on the anode and a positive pulse on the cathode. If the two R_a and R_c resistors are equal, the pulses will have the same magnitude.

The resistors are normally between 22 Kw and 100 Kw. Such great values of the cathode resistor puts the stage under heavy NFB so gain is unity and the output resistance from the anode is close to the anode load resistor, whereas output resistance from the cathode is low, usually about I Kw. As the stage has no gain, Miller capacitance is low and so is the loading of the previous stage.

The distortion is very low too, but when the outputs are loaded by the grid resistors and the input capacitances of the next stage, the perfect balance is compromised because of the difference in output resistance, and imbalance will become more and more significant as frequency rises, because the input capacitances of the output valves halve their impedances when frequency doubles. So unless measures are taken, balance will be less perfect than we expect. In practice this load-dependent unbalance is not often compensated, but in the appendix I have given an example where at least the resistive unbalance can be compensated.

Only half the supply voltage is available to generate each of the two output signals, so output swing is limited. It must be remembered that the 25 V_{eff} we need for each of the output valves is equal to 70 V peak to peak and if we require 6 dB of overload capability, demands goes up to 140V_{pp}. Even with a 400 V supply voltage the concertina can not do this with the required linearity. The concertina is excellent for driving smaller output valves such as EL 84^s requiring 10 V_{eff} and providing 10-15 Watts, but it can not drive EL 34^s without a driver stage between phase-splitter and output stage as shown in 2. Nevertheless many constructions, also commercially available, have been seen over the years where this is ignored. In these constructions there is a great possibility that the intermediate stage becomes the weakest link of the chain, and that is to me bad engineering.

The other phase-splitter shown, known as the cathode coupled phase-splitter is more tricky in its way of working, but if we keep our heads clear it is perfectly possible to understand how it works, and this is important, because claims on balance of this stage without any foundation are often heard.

A positive pulse on the input grid will cause increasing current in the first valve. This means that the voltage on the first anode goes down, producing a negative pulse here. The voltage on the two cathodes goes up, and because the grid voltage of the second grid is kept constant at the no signal level of the first grid by means of R_4 and C_1, this grid becomes more negative with respect to cathode, and consequently the current in the second valve goes down. The anode voltage on the second anode goes up and we get a positive pulse here. It means that the AC currents superimposed on the standing anode currents are in opposite directions, and if they were of exactly the same magnitude their sum, when added in R_3 would be zero, meaning that no signal component would be present on the two cathodes. But as we have seen, there must be some signal on the cathodes to drive the second valve, and consequently the decrease of anode current in the second valve will be less than the increase in first. How much depends on the amplification in the valves.

If we want, and we certainly do, signals of equal magnitude from the two outputs, R_2 must be of a slightly higher value than R_1, so that the smaller alterations of current can produce the same voltage alterations as we find on the first anode.

In order to keep differences in output resistance from the two output as small as possible, we want the anode resistors to be of almost equal size and we therefore normally choose a valve with a high amplification factor, called m, for this type of phase-splitter. The ECC 83 with m = 100 is such a valve, and luckily for us m is one of the more long-term stable parameters of a valve.

It can as a paradox be said, that an imbalance is necessary for producing the balanced output signals, and it is often maintained, that the strapping together of the two cathodes would automatically ensure balance. As you can see, this is nonsense. It is also maintained that the cathode coupled phase-splitter is a differential pair. It is of course related to the differential pair, but with no constant current device in the tail it is not such a pair. The output swing from a cathode coupled phase-splitter is of course also limited by the supply voltage. But even when we waste 60-90 Volts over R_3, we have still got 310-340 Volts to produce each of the output signals. It means that carefully designed cathode coupled phase-splitter is able to drive a pair of EL 34^s.

Both of these phase-splitters have an elevated potential at the input grid, which makes it possible to make a direct coupling between the input stage and the splitter.

By this we approach another important matter. When the signal passes from one stage with an output resistance greater than zero to the next stage with an input capacitance greater than zero, this RC combination forms a low-pass filter with a cut-off frequency, above which the signal will decrease by 6 dB per octave. Of course we want these unavoidable filters outside the audio band, but we must realise that a filter not only affects amplitude of the signal. It introduces a phase shift too.

This phase shift is not constant. It increases with frequency from 0^0 through 45^0 at the cut-off frequency and approaches a maximum of 90^0 at 10 times the cut-off frequency, but even an octave below the cut-off, the phase shift is almost 30^0.

It is obvious that an amplifier with many stages can show considerable frequency-dependent phase shift, making global NFB very difficult, and this may perhaps account for the sometimes bad reputation of this type of feedback. If we look at the most classic of all amplifiers, the Williamson – see appendix – we see a four stage amplifier with drivers, following scheme 2. To keep this amplifier stable, Williamson had to specify his output transformer to meet extremely stringent demands (the transformer introduces phase shifts as well) and even so, his amplifier was only just stable. If you plan to build a Williamson-like amplifier with 2 ECC 82^s (the more modern descendent of the 6 SN7) and 2 EL 34^s coupled as triodes and some output transformer, scraped from an amplifier of the brand “Prince Valliant” or the like, you will face severe stability problems. You will have to design compensation networks that enable you to control phase and amplitude independently. This is no simple job and to be honest, the chances that you will succeed are extremely small!!

In my opinion four stage amplifiers according to scheme 2 are not for home constructors, and this limits accessible projects to amplifiers up to about 35 W. If you need more watts, parallel output valves is the better solution.

I admit that paralleling of output valves also makes demands on the phase-splitter more stringent, but only to load capability, not to voltage swing, and these problems can be handled.

From the phase-splitter (or the driver) the signal is passed on to the output stage consisting of two power valves as shown below. To separate the anode voltage of the phase-splitter from the grid potential of the output valves, the signal passes through blocking capacitors, which in conjunction with the grid resistors of the output valves forms a high pass filter. But we are free to choose capacitors of sufficient value to keep cut off frequency as low as we want and phase shifts in the audio band can be kept very small.

The fact that you do not see a resistor and a condenser in the transition path between stages does not necessarily mean that they are not there. The often maintained claim that no phase shift can occur when stages are direct coupled is only true at low frequencies, not at high.

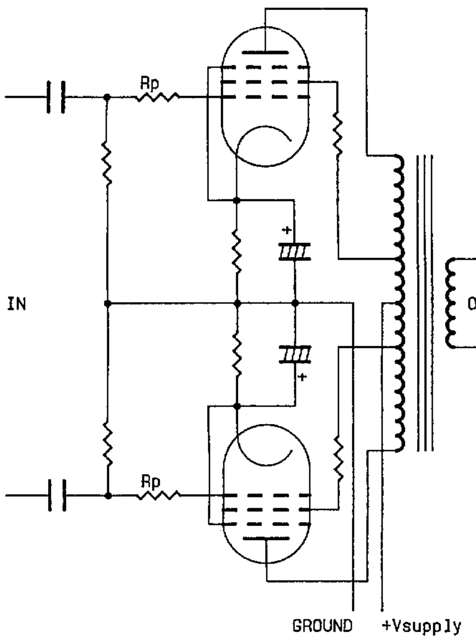


Fig. A

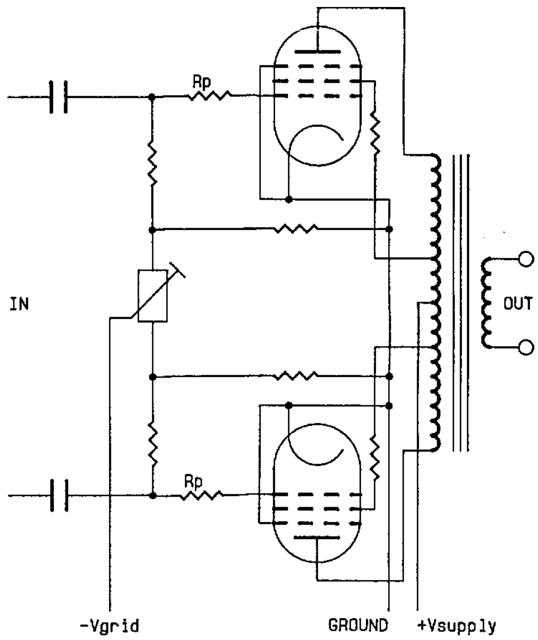


Fig. B

A suitable working point is as always achieved by negative bias on the control grids. A makes use of cathode resistors, whereas B uses an additional negative supply. Both ways shows "pros" and "cons".

It is important that the standing DC currents in the two halves of the primary of the output transformer are equal, otherwise the core will be permanently magnetized, deteriorating the transfer properties. When cathode bias is employed, the stage is to a great extent self-balancing. Two valves are never completely matched, and they will never age in exactly the same way, but if the current for some reason is greater in one valve than in the other, this greater current will cause a greater negative bias, forcing current down again. But there are two important "cons". Firstly, we are wasting 30 Volts (if the valves are EL 34^{s}) of the supply voltage, diminishing the available power and secondly, the full signal current must pass through the bypassing electrolytic condensers, and such condensers are far from ideal and should be avoided wherever possible.

Another "con" is that we are limited in our choice of working point, because cathode derived bias can only be used when the current through the valves is independent or nearly independent of signal level. If not, the working point moves up and down as current changes.

B does not suffer from these shortcomings and we can choose the working point exactly as we want, but there is no self-balancing, so DC balance has to be set and checked from time to time. As we shall se, this is an easy task to perform.

In an audio amplifier the output stage is biased to work in either Class A or Class B or somewhere in-between. Let us look into what these mystical terms really means.

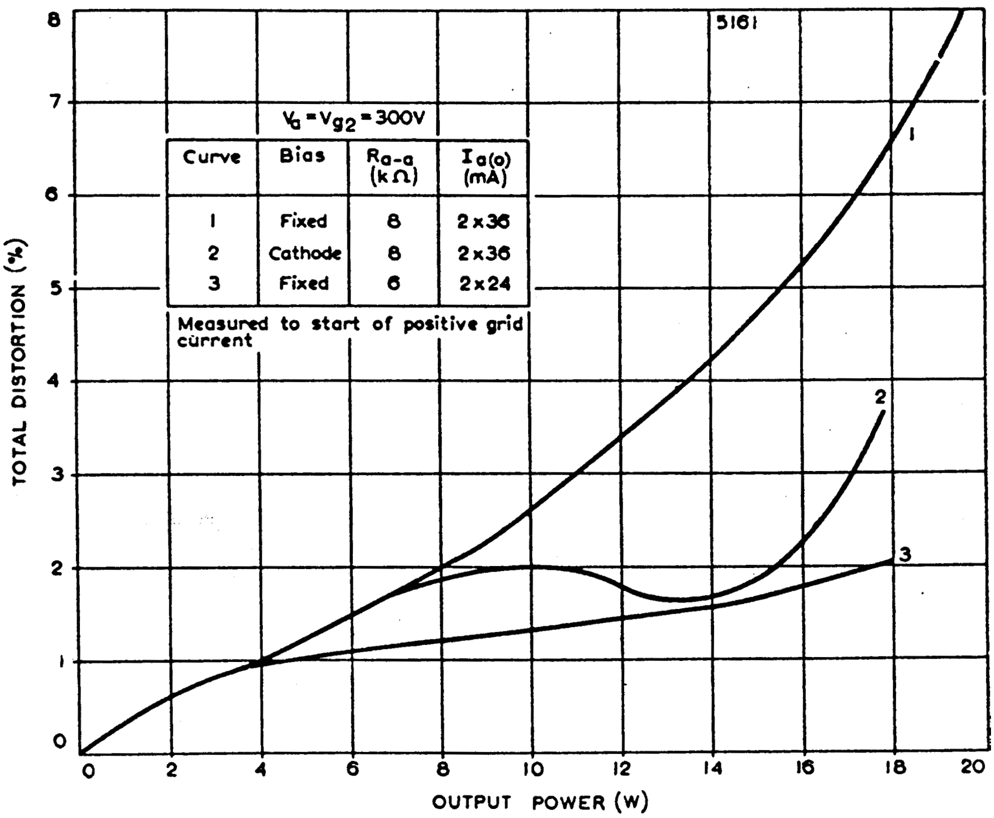
The signals from the phase-splitter are 180^0 out of phase. That means that when the grid voltage in the upper valve moves in positive direction, the grid in the lower valve becomes more negative and vice versa. The current increases in one valve, while it decreases in the other. One valve is "pushing" while the other is "pulling". This is why this type of output stage is called a push-pull stage.

In Class A, the standing current is set so that in the valve with decreasing current, the current will, even at full output, never go down to zero. An increase in one valve corresponds to exactly the same decrease in the other. We should expect no rise in the total current if the I_a/V_g curves are straight lines. This is not the case at low I_a levels, where the line becomes curved. So we will see a small increase in the total current, maybe 10-15%, when output power goes from zero to maximum.

In class B no or negligible current flows when no signal is present and only the valve receiving the positive part of the signal conducts. The anode current of the entire stage rises from almost zero to maximum along with increasing signal.

The maximum efficiency i.e. the power transferred to the load in relation to the total power dissipation is for Class A 50%, for Class B 78% (for a sine wawe).

It is obvious that Class B will show distortion in the zero crossing part of the signal, where one valve stops to conduct and the other takes over, and where the I_a/V_g lines are curved. This is the reason for amplifier freaks to claim that an amplifier can only be perfect when it is working in Class A. With good reason they base their judgement on the fact that many early solid state amplifiers, working in almost Class B, suffered from severe cross-over distortion, especially in the lower and upper end of the audio band where amplification becomes lower, and the correcting effect of the negative feedback diminishes. As often in this far from perfect world, the truth is to be found somewhere in-between. Take a look at the curves below. They were published by Mullard in the mid fifties and show distortion versus output power for two medium-sized output valves EL 84. Curve 1 is almost Class A and curve 3 is almost half the way down to Class B!



In fact I never came across a commercially available push-pull output amplifier, even the most famous and respected, working entirely in Class A.

In the good old days it was difficult to design a power-supply with an output resistance low enough to cope with great fluctuations of anode currents. Filter condensers of 8-16mF were about maximum and rectifier valves had a considerably internal resistance often further enlarged by external limiting resistors for protecting the valve against the peak current in each half-cycle and this was of course a good reason not to have an output stage working too far from Class A. To-day this is a problem of the past and we are so lucky that we can base our choice of working condition alone on sound-related considerations and personal preferences.

In my opinion a good compromise is a working point, where the power dissipated in the valves under no signal conditions is 50%-80% of the maximum permissible. The output stage will then work in Class A up to more than half the rated power and the lower dissipation will pay off in prolonged life of the valves and readjustments of balance has only to be performed at long intervals.

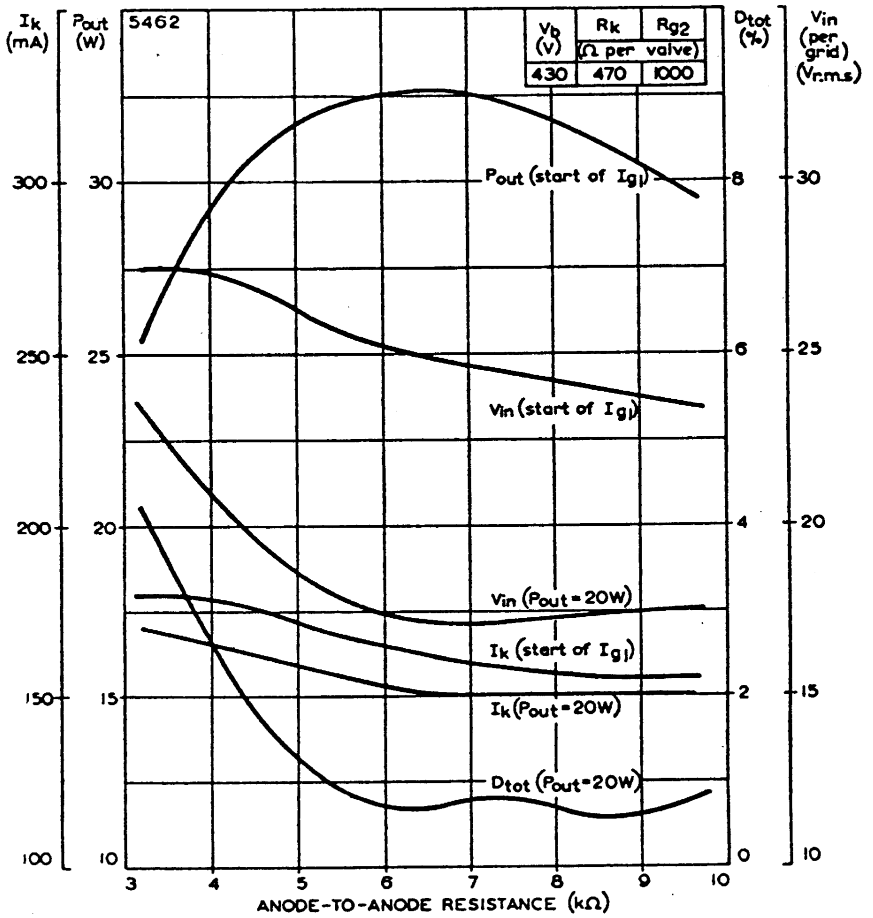
Remember that in the cross-over point the power is zero, no matter how hard the amplifier is driven, so speaking of risk of cross-over distortion at current levels like the one suggested is nonsense.

When you look at the diagrams A and B again, you will see that the screen grids are connected to taps on the primary. This way of connecting the screen grids is called ultralinear and the valves are said to be under conditions of distributed load. This deserves some explanation.

Output pentodes are intended to have their screen grids connected directly to the power supply, sometimes through a resistor 100-1000 w. Their output power is maximum and so is distortion. They can also have the screen grids strapped to the anodes, the so-called triode connection. The power that can be achieved goes down by 50-60%, but distortion is reduced dramatically. When screen grids are connected to taps as shown, distortion goes down, almost to triode level and the loss of power is almost insignificant. The tap is normally 20%-50% from the centre.

The Lundahl transformers LL 1663 and LL 1639 which I am going to use in a practical circuit described later, are wound in a way that permits the tap to be placed at 33% for normal ultralinear connection or at 67% for an "almost triode" connection. Pentode and full triode connection is always possible.

The performance curves of two EL 34s in ultralinear connection is shown below (Mullard).



— *Performance characteristics of two EL34s in pentode-connected push-pull arrangement under conditions of distributed load*

As can be seen, the ideal anode to anode resistance is about 6.5 Kw. We should remember that this resistance is a complex load, an impedance. Distortion is close to its minimum and output power is maximum. The necessary AC voltage to the control grids is minimum too.

The Lundahl transformers have an anode to anode resistance of 5 Kw only, but this is the only thing I can blame them for. In all other respects they are as perfect as transformers can be, and I can honestly say that I never saw (or heard!) better transformers.

It is always difficult to build a high quality transformer with a high primary to secondary ratio, so maybe what we loose by using a transformer with a slightly low anode to anode resistance is gained again through the somewhat lower turns ratio.

Let me conclude this chapter, which I hope in spite of its high information density, not has been too boring for you, by drawing your attention to the two resistors Rp in the diagrams A and B.Inductances and capacitances in the leads to the electrodes in the valves can sometimes form resonant circuits and amplification can therefore cause parasitic oscillations. The resistors Rp, often referred to as grid stoppers, are meant to lower Q in resonant circuits of this type so oscillations cannot occur. They must be mounted directly to the valve sockets with short leads to ensure that they perform their job.

3. A PRACTICAL CIRCUIT

Now, let us turn to a real amplifier. In the mid-fifties, Mullard in England published two amplifier constructions, the 5-10 and the 5-20, named so because they had 5 valves and were able to produce 10W and 20 W respectively. The diagrams are shown on the next two pages. As can be seen, they are nearly identical, the main differences being the output valves and the anode voltage.

The 5-20 was published in order to promote the new output pentode EL34. In these days audio amateurs and home constructors were an economical factor of some importance!

The amplifiers were built by countless enthusiasts because of their simplicity and reliability, and many manufacturers made transformers for them.

Although published by Mullard, the amplifiers cannot be said to be Mullard inventions. There was a broad consensus that medium power amplifiers (up to 25-35W) were made this way, and many commercial amplifiers from this period were very similar. Leak, Radford and Marantz amongst others used this scheme and as an example you will find the diagram of the highly respected Marantz 8B in the appendix. What Mullard really did, was to describe the way of working, and to give construction details comprehensible to the serious amateur – and they did so with great success.

We will now take a closer look at the 5-20 and try to find its weak points and see if we can, with small modifications, based on our knowledge today and modern components, improve this amplifier and adapt it to the needs of today.

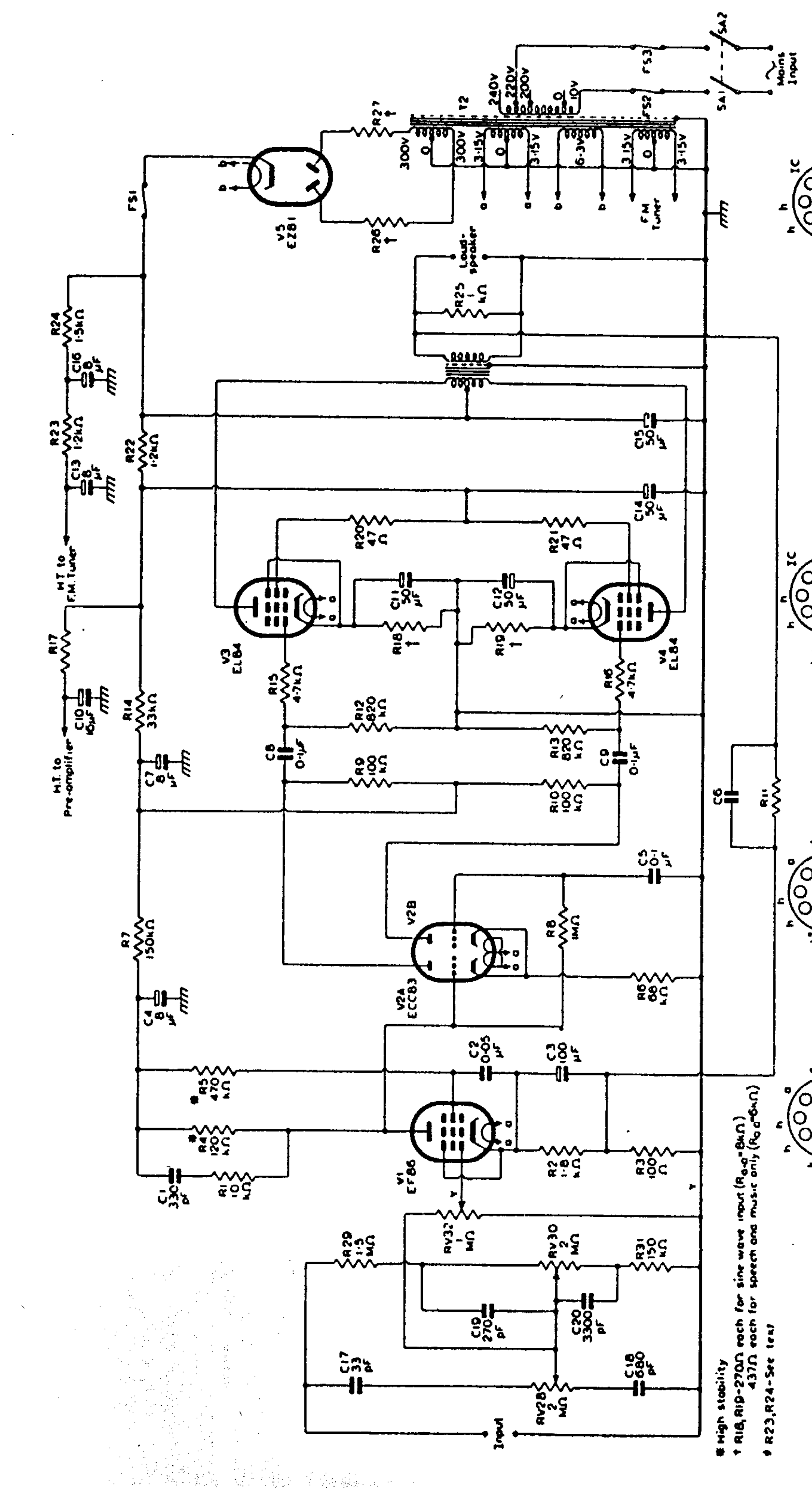
The questions we must find answers to are

1. Is the combined phase-splitter/driver capable of driving the output stage?
2. Is the pre-stage capable of driving the phase-splitter?
3. Is automatic grid bias the right way to define the working point of our output stage?
4. Is the sensitivity suitable for us today?
5. Is the amount of feedback right?

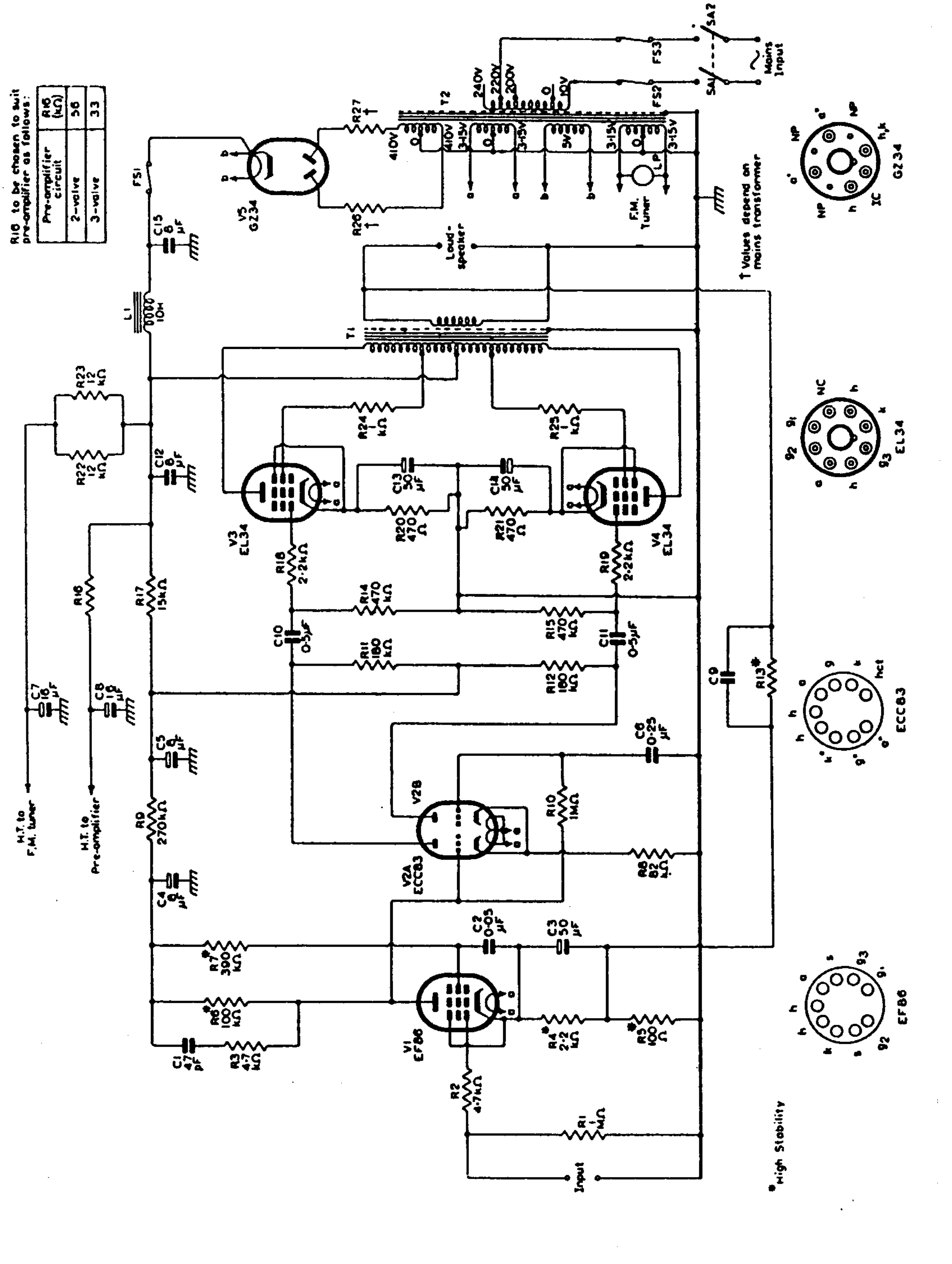
The questions are interlinked and so my answers will be.

The sensitivity is high. With the 30dB of negative feedback, prescribed by Mullard, 200 mV in will produce full output. In my opinion 30dB of feedback is too much and with the Lundahl transformer not necessary, and it could even be malign. If we decrease feedback to 20dB, we have still assured low distortion and low output resistance, but this will increase sensitivity to about 70mV for maximum output. A higher sensitivity than needed is not desirable, as it worsens the noise figures in the amplifier.

The 1st valve is the low-noise small-signal pentode EF86. Low noise is here to be understood as low by pentode standards. Pentodes are always noisier than triodes, because of the partition between screen grid and anode of the electrons emitted from the cathode. We do not need the high amplification of the pentode, so what happens if the 1st valve is connected as a triode, i.e.


* High stability
† R18, R19–270Ω each for sine wave input (R_{a-a}=8kΩ)
437Ω each for speech and music only (R_{a-a}=6kΩ)
‡ R23, R24–See text

The 5-10


R16 to be chosen to suit pre-amplifier as follows:
Pre-amplifier circuit | R16 (kΩ)
2-valve | 56
3-valve | 33
H.T. to F.M. tuner
H.T. to Pre-amplifier
V1 EF86
V2A ECC83
V2B
V3 EL34
V4 EL34
V5 GZ34
T1
T2
L1 10H
FS1
FS2
FS3
SA1
SA2
Mains Input
Loud-speaker
F.M. Tuner
† Values depend on mains transformer
*High Stability
Input
240V
220V
200V
410V
3·15V
5V
10V
EF86
ECC83
EL34
GZ34

The 5-20

the screen grid connected directly to the anode? Firstly, amplification goes down by a factor 4 from about 100 to about 25. Secondly, noise goes down by a factor 8-10, and there are no negative side effects! Mullard did not mention this possibility, because in the fifties, preamplifiers were normally very poor, and passive tone correction circuits with a high attenuation were often used in front of the output amplifier. Take a look at the 5-10! When EF86 is triode connected and 20dB of negative feedback is used, the sensitivity will be about 350mV for full output and this is still too high.

Could we in some way gain from lowering the amplification in the first stage further? This brings us to one of the other questions. Is the prestage capable of driving the phase-splitter properly?

The amplification in the phase-splitter is about 25, which means that input capacitance of the stage, due to the Miller effect, is about 40pF. The output resistance of a prestage where feedback is injected, is always close to the anode load resistor, in this case 100Kw. The 40pF of input capacitance of the next stage will load the preceding stage more and more heavy as frequency increases and the consequence is increasing distortion.

Furthermore, the output resistance in conjunction with the input capacitance forms a high frequency cut-off, starting just outside the audio band. This is to some extent compensated by the RC-network across the anode load resistor. The increasing distortion when frequency increases is however, not compensated.

If we look at the Marantz 8B in the appendix, we find no compensation network across the anode load resistor of the triode connected EF86. The explanation is that the amplification in the Marantz phase-splitter is much lower than in the Mullard design and input capacitance is therefore low too.

There are two ways of lowering the gain of the 1st stage open to us, and we will use both of them. We will reduce the anode load resistor from 100Kw to 47Kw and we will not bypass the cathode resistor any more. You might argue, that not bypassing the cathode resistor increases output resistance, but the injection of negative feedback has already increased it to its near maximum close to the value of the anode load.

These modifications make the distortion of the 1st stage less load dependent and pushes the high frequency cut-off up by one octave, and it makes the RC network across the anode load unnecessary. Sensitivity goes down to 500mV for full output and this is, I think, a fine value for today's needs.

Furthermore we get rid of a dubious compensation network and we have avoided an electrolytic condenser in the signal path.

We must recalculate the filter resistor in the power supply and the cathode resistor, so the supply voltage of 170V and the anode voltage of 85-90V are kept. This is absolutely essential, because of the direct coupling between two stages. Any change in anode voltage on the 1st valve will change the carefully chosen working point of the next stage.

The driving requirements of the output valves are 24Vrms to the grids for full output, when the valves are run with a fixed bias of ¸38V approx. and this can be considered worst case. I have chosen fixed bias instead of the automatic bias by Mullard for two reasons. First, I want to be totally free to choose the class of working, and second, I do not want any electrolytic condensers in the signal path.

The negative grid bias will, depending on class, be between ¸30 and ¸40V. The input signal can never exceed the bias voltage ¸1-2V. If it does, grid current will begin to flow in the output valves, lowering their input resistance dramatically, and by so doing, making intolerable demands on any driver.

Our phase-splitter can produce 40Vrms or 56V peak before clipping over the load 390KW in parallel with 25pF presented by the output valves. So even if the overload capability of the phase-splitter is moderate, it is completely capable of driving the output stage, and as measurements will show later, the demands to the phase-splitter do not become significantly more stringent in the high and low ends of the audio band due to an almost perfect transformer.

You might ask, why I do not inject feedback directly to the cathode. The answer is, that the feedback resistor is part of the cathode to ground path of the 1st stage and if you want to experiment with the amount of feedback, you would have to change both feedback and cathode resistor in order to keep DC conditions constant. Now as feedback is inserted over the 100W bottom resistor, the feedback resistor has a negligible influence on the cathode to ground resistance. You could even make feedback continuously variable, by changing the 100W resistor to a 100W potentiometer and inject feedback to the wiper. With the feedback resistor shown, feedback will be variable from 0 to 20dB.

In the original diagram, nothing was done to ensure low internal resistance at high frequencies of the power supply, neither for the output stage, nor for the two first stages. I have fitted 0.47mF MKT capacitors directly across the stages and HF-stop resistors are now found at every grid.

It is also now possible to adjust both DC and AC balance together with the bias adjustment of the output stage. The two 10W resistors in the cathode-to-ground path of the output valves are inserted to provide a simple and safe way to make these adjustments, and the two cathodes and a connection to ground should be accessible on measuring sockets on the amplifier chassis. Further information is to be found in the chapter concerning adjustments.

For adjustments to be accurate it is imperative that the two 10W resistors are absolutely equal and absolutely stable. See chapter concerning components.

The power supply has been modernised too. The original power supply had a fairly high internal resistance due to the vacuum tube rectifier and the very small filter condenser (8mF).

Today we will use solid state diodes, and high quality, high capacity electrolytic condensers are inexpensive in voltage ratings up to 385V. Unfortunately we need a higher rating, but if a mains transformer with a mid-point in the secondary is available (see chapter concerning components) we can, as shown use two condensers instead of one, each only carrying half the total voltage. If we use high capacity electrolytics, the choke is unnecessary, at least for a mono block.

More about this in the chapter concerning construction. The transformer must have an additional coil for grid bias, not present in the original diagram.

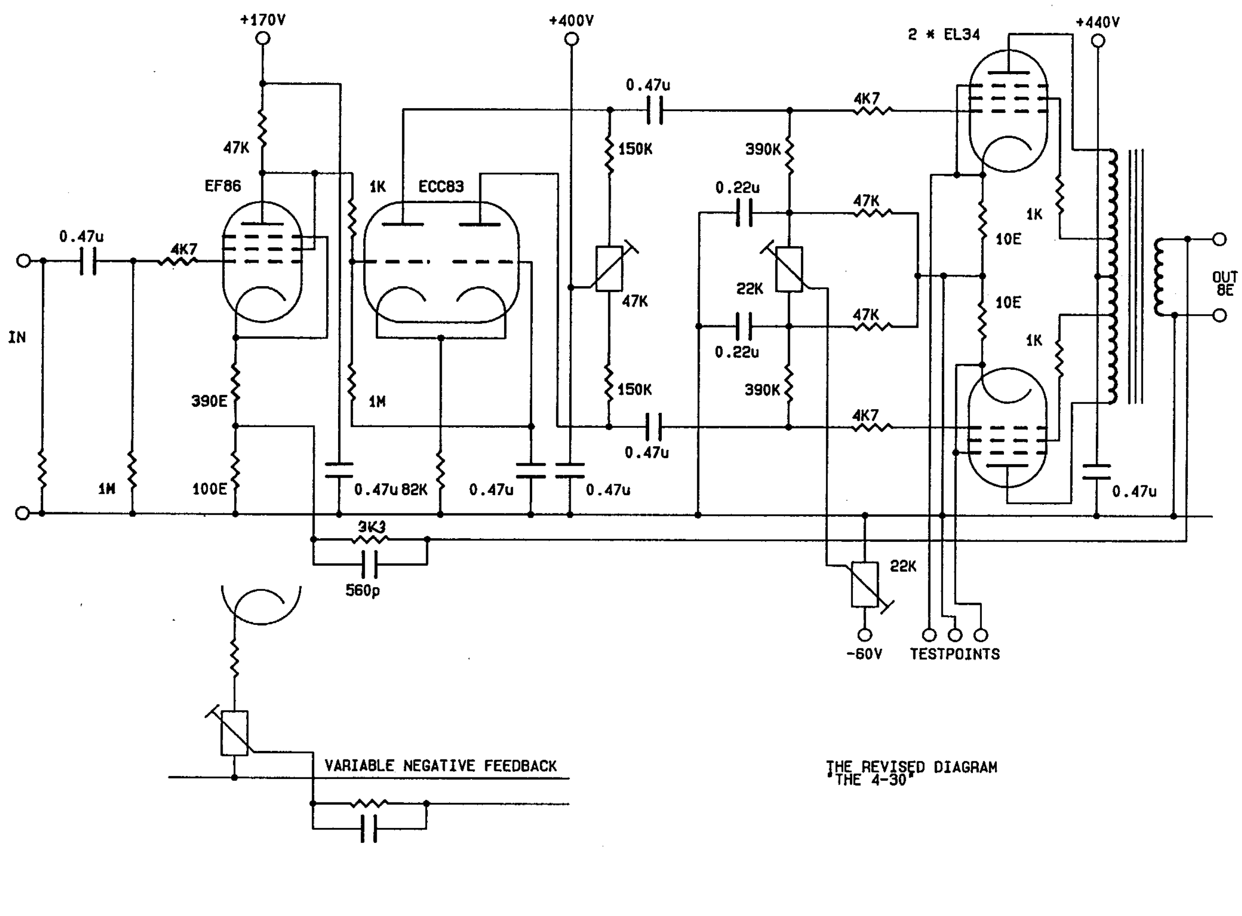
The redesigned supply will show a significantly lower output resistance than the original, thus providing improved linearity and total freedom for us to choose the working point of the output stage.

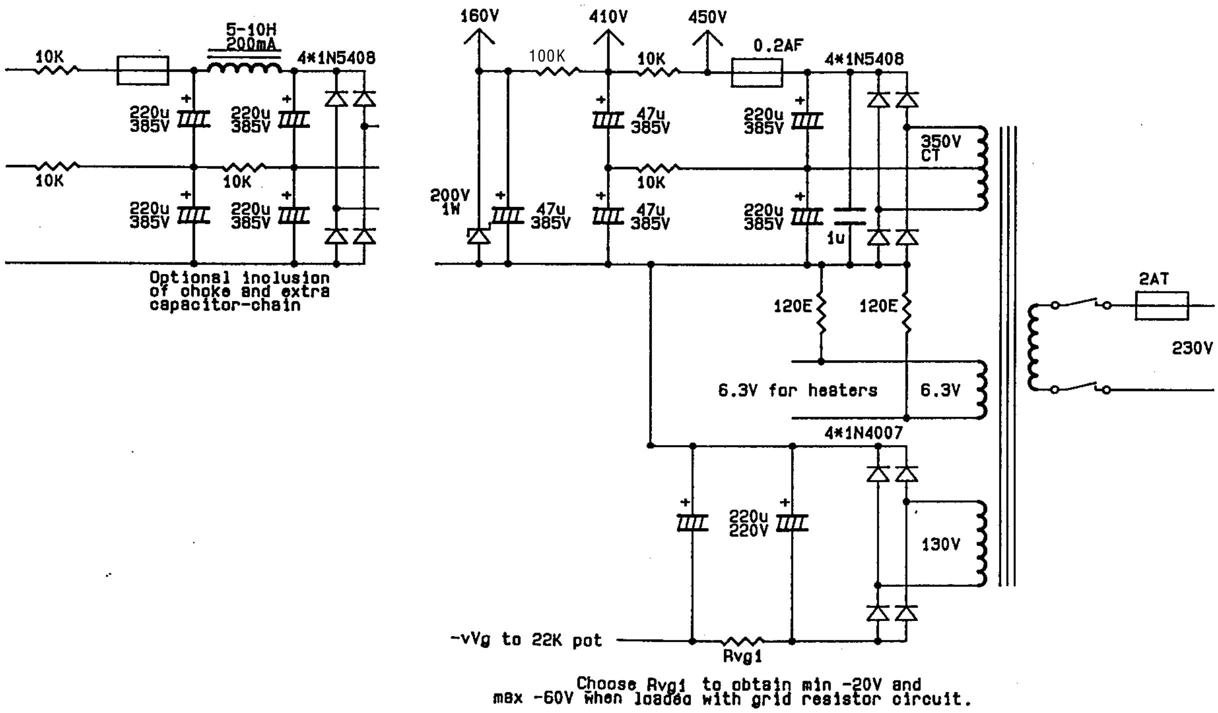
What we have got now is amplifier without electrolytic condensers in the signal path and only condenser coupled between phase-splitter and output valves.

The condenser in the input is not needed if you are absolutely sure that your signal is without DC-offset. Great care has been taken to ensure HF stability and only one compensation condenser is employed. The value of this capacitor is found by feeding the amplifier with a 1kHz squarewawe and observing the output waweform over an 8Ω resistor on the oscilloscope. The capacitor is adjusted so that no overshoot and no ringing is present but still without rounding of the front edge.

As measurements shall later reveal our efforts have paid off!

Shown here is the revised diagram. The amplifier could now be called "The 4-30"!





The values 10 Kw - 100 Kw in the supply line for 410 V and 160 V are valid for mono blocks.

4. Construction

I shall not give you a step by step building manual, only some guidelines. First of all I must confess, that I am not in favour of stereo stages – I prefer mono blocks. They are more handy, easier to build and easier to get stable, and they can be placed where they belong: close to the speakers. Demands to the power supply are also less stringent. If you insist on stereo stages, it would be advisable to use filtering with 2 chains of 220 mF capacitors in conjunction with a 5-10Hy choke.

The amplifier can be built in the old fashioned way, components being soldered between valvesockets and tagstrips. If you do so, you must remember to take all ground leads for each stage to the same point and then join these points with a thick, at least 1.5 mm^2 copperwire going from the ground point of the electrolytic capacitors in the power supply via the ground point of the output valves, the phase-splitter, the pre-stage to the ground point of the input socket and here and only here should it be connected to the chassis.

You must twist the 6.3V heater wires carefully to minimise their external magnetic field and press them into the corners of the chassis and remember to connect their artificial centerpoint to ground. You should place output transformer and mains transformer so that their coils are perpendicular and so that the coils of the one transformer follow the symmetriline of the other – see drawing. Place the input valve as far as possible from the mains transformer. The output valves generate a lot of heat, so don't place anything close to them and don't place them too close together.

To facilitate construction printed circuit boards, PCB^s have been made, both for the amplifier and for the power supply.

The amplifier PCB contains everything except:

1. the output valves,
2. the 4,7 Kw grid stoppers for the output valves and
3. the screen grid resistors for the output valves.

These resistors must be soldered directly to the grid tags of the valve sockets in order to prevent HF instability.

If you choose not to use the power supply PCB, the electrolytic condensers for anode voltage filtering for the first 2 stages can be soldered above the MKT condensers on the amplifier PCB too. In this case condensers must be 450V types.

The PCB is double sided and valve sockets, trim pots and the heater wires should be mounted on the upper side i.e. the screen side of the PCB. All other components should be mounted on the down side with the copper tracks. This is unusual, but it is practical because when the PCB is mounted below a chassis plate, the trimmers are easily accessible through three holes and heater wires are now running between the screen side of the PCB and the chassis plate. The two valves protrude through the chassis plate through 22 mm holes.

In this way every component except the trimmers are accessible from the bottom of the chassis, making experiments and service extremely simple. The dimensions of the PCB^s are 100 x 200 mm.

5. COMPONENTS

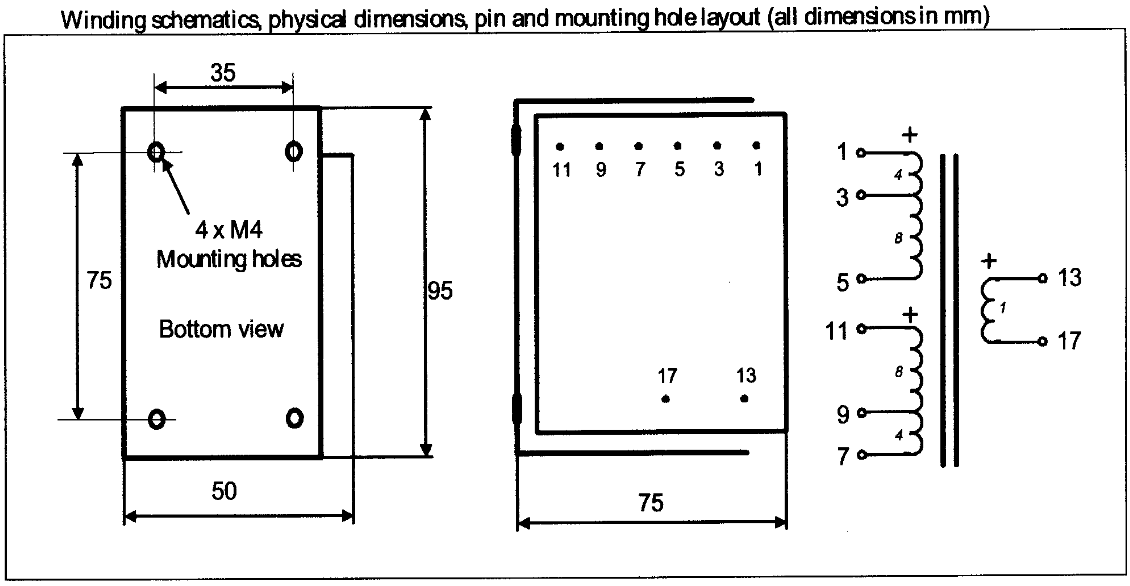
By far the most critical device in a valve amplifier is the output transformer. Not even the most sophisticated circuit in front of the transformer is able to compensate a bad component, but on the other hand a good transformer is able to pass the superior performance of well-designed circuit on to the speaker system.

Tube amplifier output transformer LL1663
5k : 8 ohms

The LL1663 is a four-sectioned dual coil C-core tube amplifier output transformer for 5 k: 8 ohms impedance ratio available in PP and SE versions.
The coil is wound using our standard high internal isolation technique with isolation foil between each copper layer. The core is an audio C-core of our own production.

Turns ratio 12+12 : 1 or (4+8)+(4+8) : 1

Winding schematics, physical dimensions, pin and mounting hole layout (all dimensions in mm)



Weight:	1.35 kg
Static resistance of each primary:	102 •
Static resistance of secondary:	0.4 •
Isolation between windings / between windings and core:	4 kV / 2 kV
Max DC current through any primary winding:	160mA

	LL1663/PP	LL1663/50mA	LL1663/100mA
Max primary signal	450V R.M.S. @ 30 Hz	200V R.M.S. @ 30 Hz	200V R.M.S. @ 30 Hz
Max output power @ 30 Hz	40W (8• spkr)	8W (8• spkr)	8W (8• spkr)

I have chosen the Lundahl LL 1663 or the LL 3910 which has proven superb. LL 1663 and LL3910 are identical except for an additional 5w secondary for feedback only in the LL3910. This secondary is not used in my design.

The primary gives you 2 options

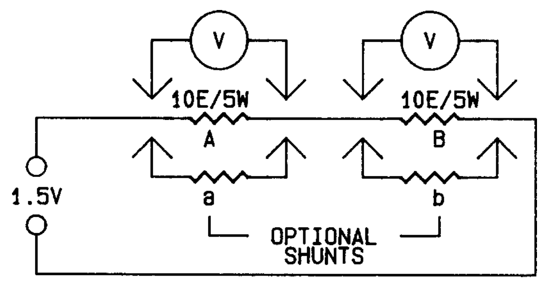
1. Supply voltage connected to 1 and 7, anodes to 5 and 11 and screen grids to 3 and 9. This is the normal ultra linear connection with taps at 33%.

2. Supply voltage to 5 and 11, anodes to 1 and 7 and screen grids to 3 and 9. This is an "almost triode" connection with taps at 67%.

My recommendation is the normal ultra linear connection. Pure triode connection is of course possible, but hardly a good idea because the EL 34s in triode connection would require a higher anode – to anode resistance, 8-10 Kw, if full advantage should be taken of the lower distortion of triodes. Pentode connection with screen grids connected to supply voltage through a common 470 w resistor is a possibility if maximum power is the objective. The valve sockets for the output valves should be ceramic to withstand the heat from the valves. Sockets for the prestages are not critical as long as good contacts are ensured.

For all resistors except the two 10 w resistors in the cathode of the EL 34s I have used 2 W metal film types. They are stable and able to withstand the sometimes high voltages across them and they are not very expensive.

The two 10 w resistors in the cathodes of the output valves are used to facilitate all adjusting of current and balance and it is imperative that they are stable and equal to a very close match. I recommend 5W wirewound types and you must match them yourself. It is done this way.



The voltages over resistor A and B are measured and compared. You must either find two resistors, where readings differ by less than 10, preferably 5 milivolts or you must shunt the resistors with resistors a and/or b until the perfect match is achieved.

The exact resulting value of the resistors is not important as long as it stays in the 10 w region. It is the match between them that counts.

For the condensers in the amplifier I use 630V MKT types except for the condenser across the feedback resistor. A 560 pF styroflex type is employed here.

Lundahl also makes a suitable mains transformer with a reference midpoint on the high voltage secondary. It is labeled LL1669A and it is capable of supplying a stereo stage too, should you wish to do so.

The PCB accepts trimpots from Bourns and Spectrol.

6. ADJUSTING AND TESTING

Before fitting the output valves you must verify that the correct bias can be set by means of the bias trimpotmeter. Adjust AC and DC balance trimpots to their middle position, apply power and measure DC voltage at the junction of the ends of the DC balance potmeter and the grid resistors. You should be able to adjust the negative bias from less than 25 V to about 50 V. If not, you will have to change the series resistor in the negative bias supply. Turn the amplifier off and remember that the reservoir electrolytes in the power supply are now charged to almost 500 Volts. Discharge them with a resistor before touching anything.

You can now place the output valves in their sockets and connect a speaker from your junk box to the output. You must now make sure that feedback from the output transformer is negative. If the amplifier starts to howl when power is applied and valves are heated, feedback is positive and you will have to reverse the ground and feedback leads to the secondary of the output transformer.

The next thing is to set bias and DC balance. Connect a voltmeter between the ground test-point and one of the cathode test-points. Adjust Bias so the reading is about 400 mV between ground and cathodes. To set balance connect the voltmeter between cathodes and adjust DC balance to 0V. Small fluctuations can occur.

The maximum permissible current in the EL 34^{s} with a supply voltage of 440 V is about 65 mA (dissipation 30 W, anode + screen grids) corresponding to 650 mV between cathodes and ground. As explained earlier, I think this is too much and will not do any good and it will definitely shorten the life of the output valves and probably of the entire amplifier due to generation of heat.

We must now set AC balance. This can be done in 3 ways:

1. If you have access to a distortion meter it is very easy to do and very accurate. Connect a 8 w load instead of the speaker, feed a 1 kHz tone to the amplifier and adjust for 14 Volt out » 25 W. Now adjust AC balance to minimum distortion. If you can hear what you are measuring, you will hear that 3^{rd} harmonic of the 1 kHz tone almost disappears (presumed of course that it was not present in the original signal!) when balance is correct.
2. A fairly accurate method, but not as good as the distortion measurement, is to measure the signal fed to the grids of the output valves with an AC voltmeter through a 0.1 mF DC blocking condenser. A normal digital multimeter can be used up to 400 Hz and you can adjust AC balance so the two readings are the same.
3. This method is very accurate and fairly easy to perform, once you have absorbed the idea. Connect the two cathode test points together and connect a headphone between the junction of the cathodes and ground. Feed a 1 kHz tone, adjust the output to 8 volts » 10 W and listen. You should now hear 3 tones: 100 Hz mains hum, the 1 kHz tone you feed to the amplifier and the 3^{rd} harmonic of that tone, 3 kHz. Now adjust the AC balance so the 1 kHz tone is minimum. You can't make it disappear totally, it will fade a little up and down. Adjust so it is as much "down" as possible. The mains hum and the 3^{rd} harmonic distortion component

are in phase in the valves and consequently cancelled in the output transformer. The 1kHz tone is out of phase in the valves and perfect balance is obtained when cancellation takes place in a common path of the current such as the one established by connecting the cathodes together. The cancellation of the odd harmonics demonstrated here is, besides higher power, one of the main advantages in the push-pull arrangement.

Although it is often said that there is nothing new under the sun when it comes to valve arrangements, I have never seen this way of setting AC balance described anywhere, so it could be new! – and it does not even rely on the match between cathode resistors as the DC adjustments do, so it is always very accurate.

As can be seen, all adjustments can be done with a normal digital multimeter and a tone generator or just a test CD with spot frequencies.

All the voltages are below 1 V when this method is employed, so there is no risk of electrical shock when you adjust the amplifier.

7. MEASUREMENTS AND RESULTS

A series of measurements were carried out on the completed amplifier built on the PCB described earlier. The conditions were: V supply, 440V current in the output valves 40 mA (per valve) load, 8 Ω and the equipment used was Audio Precision.

Output power 38W/8Ω before clipping Sensitivity 525 mV in for 30W out over 8 Ω				
Frequency response				
	At 10W	20 Hz - 20 KHz	+ 0/-0.54 dB	
	At 20W	20 Hz - 20 KHz	+ 0/-0.54 dB	
	At 30W	20 Hz - 20 KHz	+ 0/-0.6 dB	
	Above 20 KHz roll off starts. At 80 KHz response is			
	7.5 dB down at 10W			
	10.2 dB down at 20W			
	11.8 dB down at 30W			
Total harmonic + noise (THD+N)				
	At 10W	less than 0.17%		20 Hz - 5 KHz
		less than 0.32%		5 KHz - 20 KHz
	At 20W	less than 0.28%		20 Hz - 5 KHz
		less than 0.68%		5 KHz - 20 KHz
	At 30W	less than 0.38%		25 KHz - 5 KHz
		less than 1.2%		5 KHz - 20 KHz
	At 1000 Hz distortion is less than 0.12% up to 30W			
Intermodulation 60Hz/7KHz, 4:1				
	At 10W	0.22%		
	At 20W	0.26%		
	At 25W	0.27%		
	At 30W	0.29%		
	At 35W	0.51%		
Phase shift, input to output (at 10W) 20Hz 20KHz ÷ 21^0 to + 17^0				
	At 10 Hz phase shift is - 39,50^0			
	At 80 KHz (!) phase shift is +75^0			
Output resistance less than 0.6 Ω from 20 Hz - 20 KHz				
Noise on output (input terminated with 22 KΩ)				
	0.3 mV	20 Hz - 20 KHz ("Fremdspannung)		
	0.2 mV	CCIR 468		
	0.09 mV	A-weighted		

Before interpreting these measurements, we have to gain an understanding of the parameters we are measuring and their importance for the character of the sound the amplifier is able to produce. Let us look at the definitions.

1. Total harmonic distortion (THD). When a pure tone i.e. a tone only consisting of one frequency only, passes through an amplifier, the output from the amplifier will inevitably contain an amount of harmonics of the tone presented to the input. THD is defined as the percentage of these tones in the total output. Distortion with even harmonics is normally considered benign to the ear, while odd harmonic distortion is more unpleasant.
2. Intermodulation. When an amplifier is fed with 2 tones simultaneously, the output can contain not only the two tones, but also their sum and/or difference. The content of these tones in percent of the total output is called the intermodulation. This is a very unpleasant type of distortion, because the frequencies generated are mostly not harmonically related to the original frequencies.
   The main reason for the generation of intermodulation products is lack of linearity in the amplifier.
3. Linearity. An amplifier is said to be linear, when the output at any level and any frequency is equal to the input multiplied with a constant factor, the amplification. When the output approaches the upper limit, this will not be completely true. The output will be less than expected. This lack of linearity has several reasons. The three most important are: 1) the working lines of the valves are not completely straight lines and 2) output resistance of the power supply and 3) imperfections in the output transformer.

Linearity is a question of mechanics too. Hooke, the 17th century scientist made the observation that when a spring is extended by a force, the extension and the force are proportional. He expressed his observation in what we today call Hooke's law: Ut tensio, sic vis. He wrote in Latin, of course. It simply means: as extension, so force. In other words he found that a spring is a linear device. Linearity does not however extend to the extremes. When the spring is fully stretched no force can stretch it any more, but total linearity stops long before that. Our loudspeakers are such mechanical devices. We can not expect the excursion of the cones or domes to react in a completely linear way when responding to an electrical input, and the lack of linearity will increase with increasing input. We can try to make them react linearly and of course we do so, and we have achieved great results over the years, but total linearity is very unlikely to be achieved. When sound pressure is presented to our ears the timpani are forced to move inwards and outwards and these movements are transferred to the inner ear by a mechanical system, which is even at medium sound levels far from being linear. A simple experiment will prove this. If you blow two whistles with a difference in pitch simultaneously, you will not only hear the two tones blown. You will hear a third tone too. It has a frequency, which is the difference of the frequencies of the two real tones. If the interval between your whistles is a pure fifth, the third tone will be precisely one octave below the lower of your whistles. This tone is intermodulation, pure and simple. Even at moderate sound levels theese tones are easily heard and they owe their existence to the fact that doubling the force to the tympanum does not double the displacement, and the force required to give the tympanum a given displacement inwards is not the same as that required for the same displacement outwards. (This may be the reason why some people claim to be able to detect a reversing of the phase of the speakers in an audio system). The harmonic distortion and the intermodulation effects of our speakers and our ears are of much greater significance than these same effects in our amplifiers. It does of course not mean that distortion and linearity are unimportant parameters in amplifiers. I am only trying to put things into their right perspective.

I have heard people claiming that an amplifier with a vacuum tube rectifier in the power supply sounds better than the same amplifier with silicon rectifiers. The valve rectifier has a much higher output resistance than the silicon supply and consequently linearity is better for the amplifier with solid state rectifiers. Maybe too high a degree of linearity is unwanted? – at least to some of us. - Total linearity is a rare bird in the real world, at least over an extended range! Today lack of linearity at levels close to the maximum has been reintroduced as a virtue under the name of “soft clipping”!

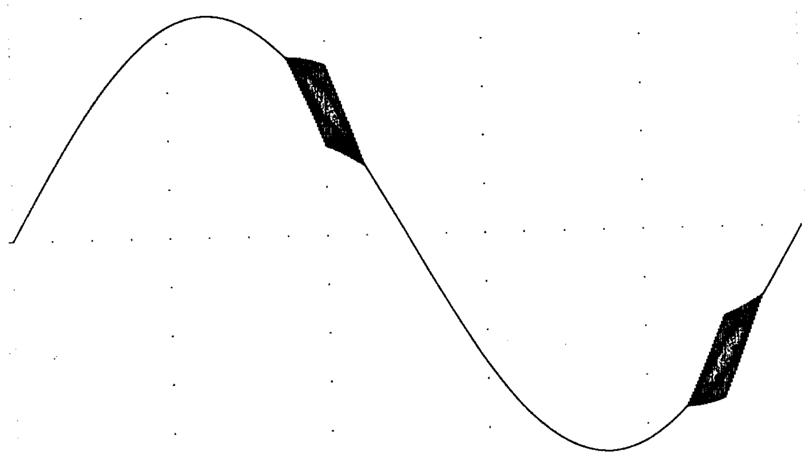
It is a fact that valves amplifiers under stress act otherwise than modern semiconductor constructions and in a way it is fair to say that they react in a more human-like way.

Let us now turn to the measurements.

The frequency response is shown at 10, 20 and 30W out with negative feedback (NFB), and at 10 and 30W out without NFB. Even at 20 kHz/30W without NFB response is only 2.5 dB down. With NFB the corresponding value is ¸ 0.60 dB. The power bandwidth of the amplifier is very good indeed. Total harmonic distortion + noise, THD + N stays below 0.12% up to 30W (input = 525 mV). This is a very good figure for a valve amplifier and you might ask: Could I use this amplifier without NFB? The answer to this is of course yes. The distortion will hardly exceed 1% up to 10W and the power bandwidth is still satisfactory. The average output power when listening to music (not rock!) is normally less than 10% of the power in the peaks. So even at high sound levels linearity is not bad and amplifier is perfectly stable without NFB. But there are three reasons for not operating the amplifier without NFB. The first is noise. The noise floor will rise, and mains hum could be a problem. This can however be counteracted to some extent by fitting a voltage divider in front of the amplifier with, say, 30dB damping and a small value of the lower resistor, maybe 5-10 kohms. Sensitivity will not be a problem since amplification will rise with 20 dB, when 20 dB of NFB is omitted. The other reason is more serious. The output resistance rises dramatically when we remove NFB. The output resistance of the amplifier with 20 dB of NFB is just below 0.6 ohms. Without NFB it goes up to several ohms and the electrical damping of your speakers is compromised. Since most speaker systems rely on amplifier damping there is every chance the sound in the bass region will change, the bass becoming “loose”, “boomy” or “woolly” – you choose the word. The 3^{rd} reason is that stability of gain is less than perfect and as explained earlier this can have serious consequences for stereoimaging.

I know that some people are against NFB. I don't think they know exactly why, but my guess is that they without their knowledge, of course, have been subject to bad engineering. The fact that an amplifier is not oscillating without signal and when terminated with a purely resistive load does in no way guarantee stability under real-world working conditions.

A picture on the scope could look this way.



showing that oscillation can occur at certain levels and this can be load and signal dependant, too. It is obvious that a behaviour like this will not contribute to clarity of the sound!

Stability of the amplifier was investigated too. It is vital to be sure that NFB remains negative under all circumstances. Phase shift input to output was measured and it turned out to stay with in +/¸ 20^0 within the audio band and was only 75^0 at 80 kHz. Together with the smooth frequency roll-off (at 80 kHz response is 15 dB down without feedback) it is ensured that feedback can never change to positive and stability is impeccable.

The good behaviour of the amplifier is also demonstrated by the way it handles square waves.

A 1 kHz square wave at 30W shows no overshoot and no ringing – at 10 kHz rounding is of course seen, due to the frequency roll off, but still no ringing.

The picture is hardly changed when the 8 w resistive loads is replaced by the complex load of a real speaker. A two-way speaker with crossover was used for this examination and this very difficult test was passed with flying colours.

If you plan to look into this yourself, you should only use bursts of very short duration, if you care for your speakers!

Measurements of THD vs. frequency at 10, 20 and 30 W out shows core saturation of the transformer below 16 Hz at 20 W and below 20 Hz at 30 W, very fine results in the mid band and an unavoidable rise in the upper end of the audio band, but when judging, bear in mind, that 3^{rd} harmonic distortion at frequencies higher than 6 kHz cannot be heard because the distortion products are moving outside our hearing range.

When you compare the results given here with the specs of your favourite amplifiers of the golden age of valve supremacy, remember that distortion was normally given by 1kHz unless otherwise stated and the manufacturer of an amplifier like this would without a twinkle of his eye have promoted it as capable of delivering 35 W (at least!) with a THD below 0.2%, an achievement in which every manufacturer would have taken pride.

The intermodulation is low up to more than 30 W but exceeds 0.5% at 35 W indicating that by this power output we are leaving the linear region.

It has once again become fashionable to run the output valves as triodes, so this was tried out too by connecting the screen grid resistors to the anodes.

The available power is about halved and the distortion graphs show, that reduced output seems to be the only benefit gained by running EL34s as triodes!

The amplification is now lower in the output stage and with no other component changed, the NFB goes down from 20 dB to about 15 dB. If we increased this again to 20 dB, we would probably see a small reduction of the THD figures. I would expect a further reduction in THD if a transformer with 10 Kw anode to anode resistance was used, but I have not access to such a transformer and have not found the prospects of further experiments in this direction promising, but the amplifier works well and is absolutely stable this way too.

The bottom line is that this conservatively built and relatively simple amplifier which makes the most out of every part used, is extremely good and listening test have confirmed this. To a large extent it owes its qualities to the very good output transformer from Lundahl. This is perhaps most clearly seen from the response graphs without NFB. To be honest, in my 45 years of experience, I have never before come across an output transformer comparable to this!

The amplifier described is maybe not the best in the world in its class, but it is very close to! – and the measurements carried out makes it one of the best documented!

The curves printed from Audio Precision are found in the appendix.

8. FURTHER IMPROVEMENTS?

In my opinion this amplifier is now at the limits of what is possible when the output power is about 30 Watts and the concept is input stage – phase-splitter – output stage.

The output resistance of 0.6W is still high by today's standard and it could be reduced with more negative feedback. A further 10 dB of NFB would lead to an output resistance of 0.2W, but I do not like this approach. NFB should be used but not abused. Output resistance could be taken down to zero ohms or even become negative when positive current derived feedback is applied. Unfortunately the sonic results are rather unpredictable. What is an improvement when used in conjunction with one speaker system and cable can be quite the opposite when used with another speaker/cable combination. The effect can be impressive and it can be disappointing. So unless you know exactly what you are doing and are able to measure and evaluate your results, I would not recommend the use of positive current feedback, but it can certainly make an improvement under the right conditions

The power supply could benefit from the inclusion of a filter choke of 5-10 Hy and an additional chain of 220 + 220mF capacitors – see diagram. These components are necessary in a stereo amplifier but no harm is done using them in mono blocks as well.

9. AKNOWLEDGEMENTS

I want to dedicate this paper to the memory of the late C. Damgaard Knudsen of Aarhus, Denmark. He was a watchmaker and a keen audio amateur. His skills as a watchmaker were seen full-blown when in about 1960 he made his own tape-recorder from scratch. This recorder was at least 10 years ahead of its time and it served him for more than 30 years. Only changing heads and amplifiers, mostly of his own design, reflected ripening technology. He was a lover of Mahler's symphonies long before Mahler came in fashion, and when I was very young he introduced me to serious high fidelity and I am thankful for many good discussions and a never failing helping hand.

I am very grateful to Mr. N.P. Petersen of Stilling, Denmark. He is the man behind the NP Mixing consoles that became the ultimate yardstick for comparing analogue mixers in our part of the world. I myself have been working behind such a console for more than 17 years. This mixer shows still no scratching potentiometers or bad switches and over all the years less than eight hours (!) of service have been required. I have always enjoyed audio discussions with him and I thank him for advice in connection with this paper and I am greatly indebted to him for making Audio Precision equipment available and for helping me with the measurements.

Mr. Clemens Johansen of the Danish Broadcasting Corporation and Mr. Ole Brøsted Sørensen of DPA Microphones (formerly, Bruel and Kjaer) have provided many good hints for me and they have without grumping read whatever I wrote, and their comments have proven extremely valuable to me.

Last but not least I am profoundly indebted to Mrs. Tove Dahl Rasmussen at The Royal Academy of Music, Aarhus, Denmark, for typing and for not complaining over my terrible handwriting (I can't type, but don't tell anybody!). Without her, this paper would not have seen the light of the day and those of you who embarked on the project described would have saved your money!

Claus Byrith
August 2000

THE AUTHOR

Claus Byrith, b. 1941, is a sound recording engineer. Since he was a boy he has designed and built all sorts of audio equipment. In 1968 he joined the team of the Department of Electronic Music and Musical Acoustics, a joined venture of the Royal Academy of Music in Aarhus and the University of Aarhus. In 1973 he became head of the studio at the Royal Academy. He has done innumerable recordings and received the Swedish "Fonogrampriset" in 1983 for recordings of Quartets by Stenhammar. He has also received an American Critics Award. In later years he has taken great interest in restoring of old recordings, and he has prepared many reissues of Danish recordings of the thirties and fourties.

His re-awakened interest in valve technology is based, as he puts it, on the fact that valve equipment is simple and easy to understand, serviceable and when well designed, extremely well sounding.

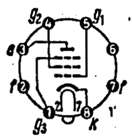
Appendix I
Data Sheets

EL 34
6 CA 7

Endpentode

Verwendung für Kraftverstärker

Power Pentode for Power Amplifier



Oktal

Kolben Nr. 23
Bulb No. 23

Allgemeine Daten / General Data

Heizung / Heating

$U_f = 6{,}3$ V
$I_f = 1{,}5$ A
Indirekt / Indirect

Kapazitäten / Capacitances

$C_{eing} = 15{,}5$ pF
$C_{ausg} = 7{,}2$ pF
$C_{ag_1} < 1{,}0$ pF
$C_{g_1f} < 1{,}0$ pF
$C_{kf} = 11$ pF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Betriebsdaten / Typical Operation

Eintakt A / Class A

U_b =	265	265 V
U_a =	250	250 V
R_{g_2} =	2	0 kΩ
U_{g_2} =	0	0 V
U_{g_1} =	−14,5	−13,5 V
I_a =	70	100 mA
I_{g_2} =	10	14,9 mA
S =	9,0	11 mA/V
R_i =	18	15 kΩ
R_a =	3,0	2,0 kΩ
$U_{g_1\sim}$ =	9,3	8,7 V_{eff}
$N_\sim$ =	8	11 W
k =	10	10 %
$U_{g_1\sim}$ $N_\sim$ (= 50 mW) =	0,65	0,5 V_{eff}
$\mu_{g_2g_1}$ =	11	11

Grenzdaten / Maximum Ratings

$U_{a\,kalt}$ = 2000 V
U_a = 800 V
Q_a ($U_{g_1\sim}$ = 0) = 25 W)
Q_a ($U_{g_1\sim}$ > 0) = 27,5 W
$U_{g_2\,kalt}$ = 800 V
U_{g_2} = 425 V
Q_{g_2} = 8 W
I_k = 150 mA
R_{g_1} = 0,7 MΩ*
R_{g_1} = 0,5 MΩ**
U_{fk} = 100 V
R_{fk} = 20 kΩ

* Kl. A und AB

** Kl. B

EL 84
6 BQ 5

Endpentode

Power Pentode

Allgemeine Daten / General Data

Heizung / Heating

$U_f = 6{,}3$ V
$I_f = 0{,}76$ A
Indirekt / Indirect

Kapazitäten / Capacitances

$C_e = 10{,}8$ pF
$C_a = 6{,}5$ pF
$C_{ag_1} < 0{,}5$ pF
$C_{g_1f} < 0{,}25$ pF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten / Characteristics

U_a = 250 V, I_{g_2} = 5,5 mA
U_{g_2} = 250 V, S = 11,3 mA/V
U_{g_1} = −7,3 V, R_i = 40 kΩ
I_a = 48 mA, $\mu_{g_2g_1}$ = 19,5

Betriebsdaten / Typical Operation

Eintakt A / Class A

U_a =		250		V
U_{g_2} =		250		V
U_{g_1} =		−7,3		V
R_k =		135		Ω
R_a =		5,2		kΩ
$U_{g_1\sim}$ =	0	0,3	4,3	V_{eff}
I_a =	48	—	49,5	mA

Grenzdaten / Maximum Ratings

U_a = 300 V
Q_a = 12 W
U_{g_2} = 300 V
Q_{g_2} = 2 W ($U_{g_1\sim}$ = 0)
Q_{g_2} = 4 W ($U_{g_1\sim}$ > 0)
$-U_{g_1}$ = 100 V
I_k = 63 mA
U_{gk} = 100 V
R_{g_1} = 1 MΩ
R_{g_1} = 0,3 MΩ*
R_{fk} = 20 kΩ

* U_{g_1} fest
Fixed grid bias

Fortsetzung ↓

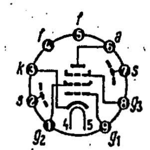
EF 86

NF-Pentode

Verwendung als NF-Verstärker

AF Pentode

AF Amplifier



Noval

Kolben Nr. 6
Bulb No. 6

Allgemeine Daten / General Data

Heizung / Heating

$U_f = 6{,}3$ V
$I_f = 0{,}2$ A
Indirekt / Indirect

Kapazitäten / Capacitances

C_{ausg} = 5,5 pF
C_{eing} = 4,0 pF
C_{ag_1} < 0,05 pF
C_{g_1f} < 0,0025 pF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten / Characteristics

U_a = 250 V
U_{g_3} = 0 V
U_{g_2} = 140 V
U_{g_1} = −2 V
I_a = 3,0 mA
I_{g_2} = 0,6 mA
S = 2,0 mA/V
R_i = 2,5 MΩ
$\mu_{g_2g_1}$ = 38

Betriebsdaten / Typical Operation

NF-Verstärker / AF Amplifier

U_b =	200	250	200	250	V
R_a =	100	100	220	220	kΩ
R_g =	1	1	1	1	MΩ
R_g' =	330	330	680	680	kΩ
R_{g_2} =	0,39	0,39	1	1	MΩ
R_k =	1	1	2,2	2,2	kΩ
I_k =	1,7	2,1	0,8	0,9	mA
v =	106	112	170	180	
$U_{a\sim}$ =	40	50	36	46	V_{eff} [1]

[1] Bei k = 5% der Klirrfaktor ist der Ausgangsspannung etwa proportional

The harmonic distorion is approx. proportional to the output voltage

Grenzdaten / Maximum Ratings

$U_{a\,kalt}$ = 550 V
U_a = 300 V
Q_a = 1 W
$U_{g_2\,kalt}$ = 550 V
U_{g_2} = 200 V
Q_{g_2} = 0,2 W
I_k = 6 mA
R_{g_1} = 10 MΩ [1]
R_{g_1} = 3 MΩ [2]
R_{g_1} = 22 MΩ [3]
U_{fk} (k pos) = 100 V
U_{fk} (k neg) = 50 V
R_{fk} = 20 kΩ [4]

[1] (Q_a < 0,2 W)
[2] (Q_a > 0,2 W)
[3] selbstanlaufend; U_{g_1} only produced by R_{g_1}
[4] In Phasenumkehrstufen max. 120 kΩ; In phase-splitter stages max. 120 kΩ

ECC 83
12 AX 7

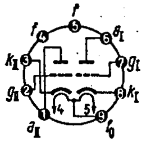
NF-Doppeltriode

Verwendung als NF-Verstärker und Phasenumkehrröhre

AF Twin Triode

AF Amplifier

Phase-splitter



Noval

Kolben Nr. 6
Bulb No. 6

Allgemeine Daten / General Data

Heizung
Heating

U_f = 6,3 V
I_f = 0,3 A
oder
U_f = 12,6 V
I_f = 0,15 A
Indirekt
Indirect

Kapazitäten
Capacitances

$C_{a_I k_I}$ = 0,33 pF
$C_{a_{II} k_{II}}$ = 0,23 pF
$C_{g_I k_I}$ = $C_{g_{II} k_{II}}$ = 1,65 pF
$C_{a_I g_I}$ = $C_{a_{II} g_{II}}$ = 1,60 pF
$C_{g_I f}$ = $C_{g_{II} f}$ < 150 mpF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten
Characteristics

U_a = 250 V S = 1,6 mA/V
U_g = −2,0 V μ = 100
I_a = 1,2 mA R_i = 62,5 kΩ

Betriebsdaten
Typical Operation
NF-Verstärker
AF Amplifier

U_b	=	200	250	300	350 V
R_a	=	220	220	220	220 kΩ
R_g	=	1	1	1	1 MΩ
R_g'	=	680	680	680	680 kΩ
R_k	=	3,3	2,7	2,2	1,5 kΩ
C_k	=	50	50	50	50 µF
I_a	=	0,36	0,48	0,63	0,85 mA
v	=	56,0	66,5	72,0	75,5
$U_{a\sim}$	=	24	28	36	37 V_{eff}
k	=	4,6	3,4	2,6	1,6 %

Grenzdaten / Maximum Ratings

je System
per section

$U_{a\,kalt}$ = 550 V
U_a = 300 V
Q_a = 1 W
I_k = 8 mA
U_g = −50 V
R_g = 2 MΩ
R_g = 22 MΩ*
U_{fk} = 180 V
R_{fk} = 20 kΩ**

* selbstanlaufend = 22 MΩ
U_{g_1} only produced by R_{g_1} = 22 MΩ

** In Phasenumkehrstufen max. 150 kΩ
In phase-splitter stages max. 150 kΩ

ECC 81
12 AT 7

HF-Doppeltriode

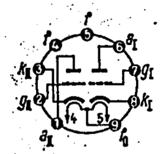
Verwendung als UKW-Oszillator, UKW-Mischröhre und HF-Verstärker

RF Twin Triode

VHF Oscillator

VHF Mixer

RF Amplifier



Noval

Kolben Nr. 6
Bulb No. 6

Allgemeine Daten / General Data

Heizung
Heating

U_f = 6,3 V
I_f = 0,3 A
oder
U_f = 12,6 V
I_f = 0,15 A
Indirekt
Indirect

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten
Characteristics

U_a = 250 V
U_{g_1} = −2 V
I_a = 10 mA
S = 5,5 mA/V
μ = 60
R_i = 11 kΩ

Grenzdaten / Maximum Ratings

je System
per section

$U_{a\,kalt}$ = 550 V
U_a = 300 V
Q_a = 2,5 W
I_k = 15 mA
U_{g_1} = −50 V
R_{g_1} = 1 MΩ
U_{fk} = 90 V
R_{fk} = 20 kΩ

Kapazitäten
Capacitances

C_{gI} = 2,3 pF
C_{aI} = 0,45 pF
$C_{a_I g_I}$ = 1,6 pF
$C_{a_I k_I}$ = 0,20 pF
$C_{k_I f}$ = 2,5 pF
$C_{k_I}/(g_I+f)$ = 4,8 pF

$C_{a_I}/(g_I+f)$ = 1,9 pF
$C_{a_I a_{II}}$ < 0,4 pF
$C_{a_I g_{II}}$ < 0,07 pF
$C_{g_{II}}$ = 2,3 pF
$C_{a_{II}}$ = 0,35 pF
$C_{a_{II} g_{II}}$ = 1,6 pF
$C_{a_{II} k_{II}}$ = 0,20 pF

$C_{k_{II} f}$ = 2,5 pF
$C_{k_{II}}/(g_{II}+f)$ = 4,7 pF
$C_{a_{II}}/(g_{II}+f)$ = 1,8 pF
$C_{g_I g_{II}}$ < 0,005 pF
$C_{a_{II} g_I}$ < 0,04 pF

ECC 85
6 AQ 8

HF-Doppeltriode

Verwendung als HF-Verstärker und selbstschwingende Mischröhre

RF Twin Triode

RF Amplifier

Self-Excited Mixer

Fortsetzung ↓

Allgemeine Daten / General Data

Heizung
Heating

U_f = 6,3 V
I_f = 0,435 A
Indirekt
Indirect

Kapazitäten
Capacitances

$C_{g_I}/(k_I+f+s)$ = 3,0 pF
$C_{a_I g_I}$ = 1,5 pF
$C_{a_I k_I}$ = 0,18 pF
$C_{g_{II}}/(k_{II}+ +s)$ = 3 pF
$C_{a_{II} g_{II}}$ = 1,5 pF
$C_{a_{II} k_{II}}$ = 0,18 pF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten
Characteristics

U_a = 250 V S = 6,0 mA/V
U_{g_1} = −2,2 V μ = 57
I_a = 10,0 mA

Betriebsdaten
Typical Operation
HF-Verstärker (System I)
RF Amplifier (Section I)

U_b = 250 V
R_{av} = 1,8 kΩ
U_a = 230 V
R_k = 200 Ω
U_{g_1} = −2 V
I_a = 10 mA
S = 6,0 mA/V
R_i = 9,7 kΩ
R_{ei} (100 MHz) = 6 kΩ
$R_{äq}$ = 0,5 kΩ

Grenzdaten / Maximum Ratings

je System
per section

$U_{a\,kalt}$ = 550 V
U_a = 300 V
Q_a = 2,5 W
$Q_{a_I}+Q_{a_{II}}$ = 4,5 W
I_k = 15 mA
U_{g_1} = −100 V
R_{g_1} = 1 MΩ
U_{fk} = 90 V
R_{fk} = 20 kΩ

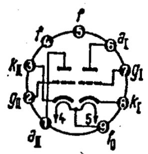
ECC 82
12 AU 7

NF-Doppeltriode

Verwendung als NF-Verstärker, Phasenumkehrröhre, Synchronisations-Trennröhre, Multivibrator und Sperrschwinger

AF Twin Triode

AF Amplifier



Noval

Kolben Nr. 6
Bulb No. 6

Allgemeine Daten / General Data

Heizung / Heating
U_f = 6,3 V
I_f = 0,3 A oder
U_f = 12,6 V
I_f = 0,15 A
Indirekt / Indirect

Kapazitäten
Capacitances

$C_{eing_{II}}$ = 1,8 pF
$C_{ausg_{II}}$ = 0,25 pF
$C_{a_{II} g_{II}}$ = 1,5 pF
$C_{g_{II} f}$ < 135 mpF
$C_{g_I g_{II}}$ < 10 pF
$C_{a_{II} g_I}$ < 60 mpF
C_{eing_I} = 1,8 pF
C_{ausg_I} = 0,37 pF
$C_{a_I g_I}$ = 1,5 pF
$C_{g_I f}$ < 135 mpF
$C_{a_I a_{II}}$ < 1,1 pF
$C_{a_I g_{II}}$ < 110 mpF

Kenn- und Betriebsdaten / Characteristics and Typical Operation

Kenndaten / Characteristics
je System / per section

U_a = 250 V
U_g = −8,5 V
I_a = 10,5 mA
S = 2,2 mA/V
μ = 17
R_i = 7,7 kΩ

Betriebsdaten / Typical Operation
NF-Verstärker / AF Amplifier

U_b	=	150	200	250	300 V
R_a	=	100	100	100	100 kΩ
R_g	=	1	1	1	1 MΩ
R_g'	=	330	330	330	330 kΩ
C_k	=	50	50	50	50 µF
R_k	=	2,2	2,2	2,2	2,2 kΩ
I_a	=	0,98	1,30	1,63	1,97 mA
v	=	14	14	14	14
$U_{a\sim}$	=	17	25	32	41 V_{eff}
k	=	5,6	5,8	5,9	6,0 %

Grenzdaten / Maximum Ratings

je System
per section

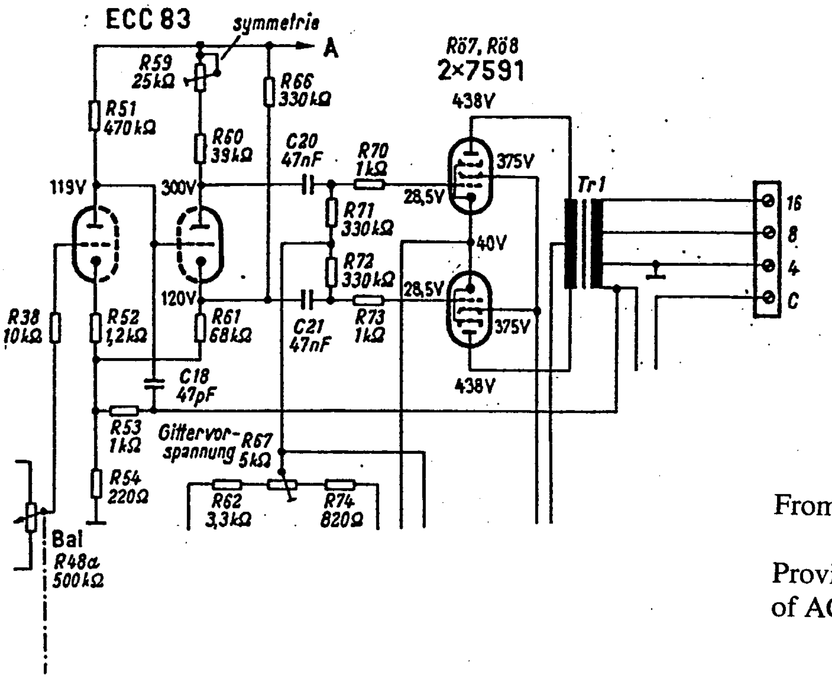
U_a = 300 V
Q_a = 2,75 W
I_k = 20 mA
$-U_g$ = 100 V
$I_{k\,sp}$ = 250 mA*
R_g = 1 MΩ
U_{fk} = 180 V
R_{fk} = 20 kΩ**

* Impulsdauer = 10% einer Periode, t_{max} = 2 ms
Pulse time = 10% per cycle, t_{max} = 2 ms

** In Phasenumkehrstufen max. 150 kΩ
In phase-splitter stages max. 150 kΩ

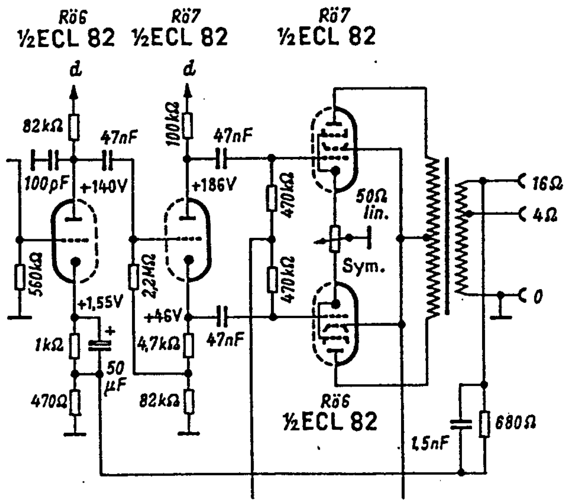
Appendix II

parts of diagrams
The Marantz 8B
The Williamson
The Quad II



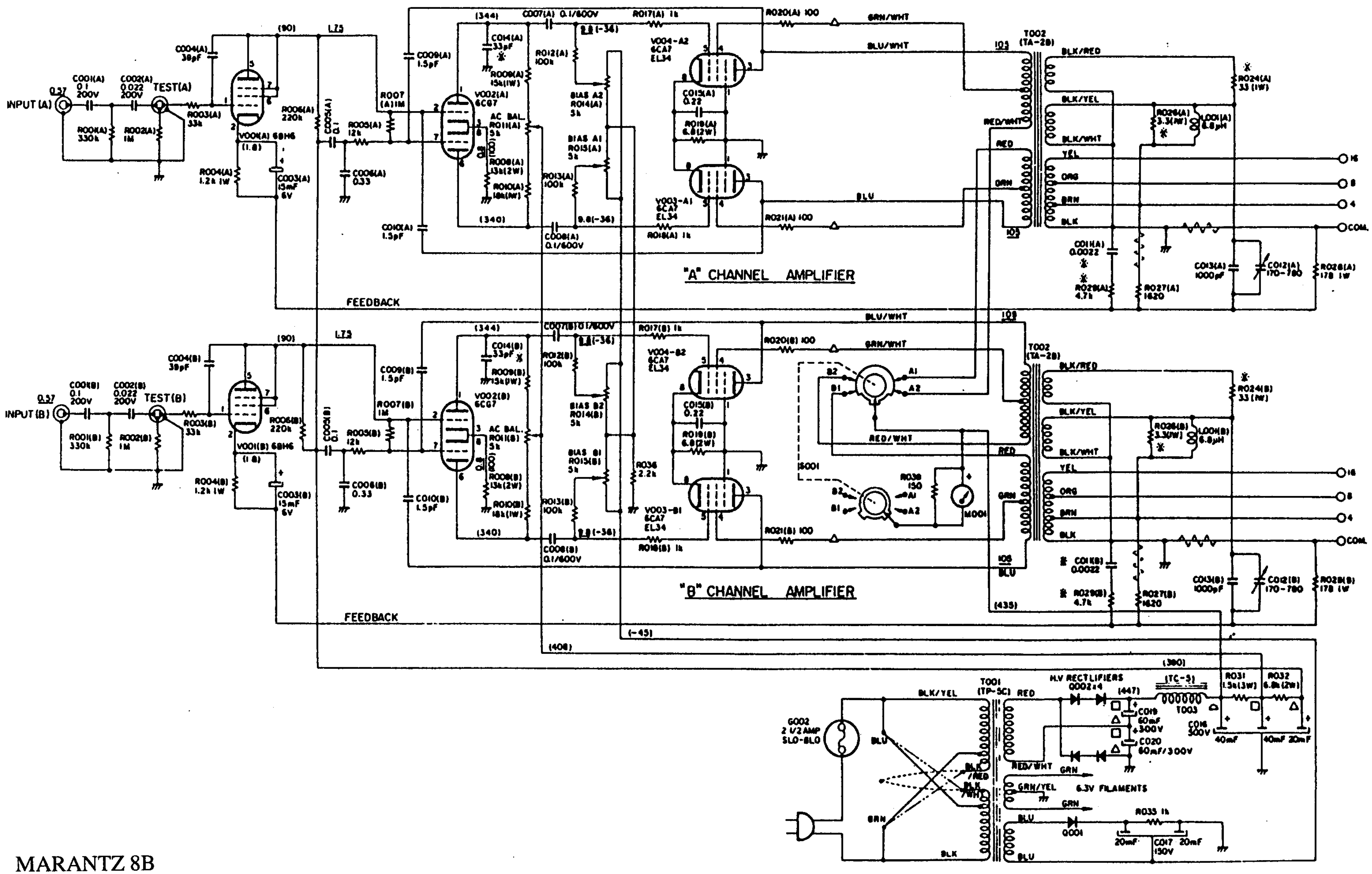
From Fisher X 101C (1964)

Provision for adjustment of AC-balance.



TELEWATT VS56 (1963)

Different load resistors in anode and cathode of the phase-splitter to compensate for difference in output resistances.



MARANTZ 8B

A very respected amplifier following a similar scheme as the amplifier described in this paper. The global feedback system is very elaborated in order to compensate for imperfections in the output transformer.

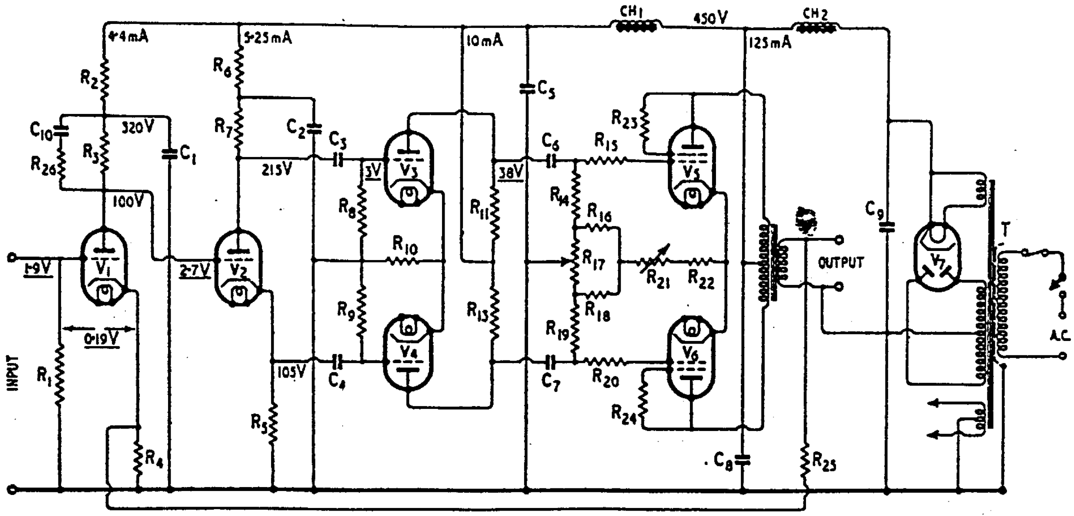


Fig. 1. Circuit diagram of complete amplifier. Voltages underlined are peak signal voltages at 15 watts output.

R_1	1MΩ	¼ watt ± 20%
R_2	33,000Ω	1 watt ± 20%
R_3	47,000Ω	1 watt ± 20%
R_4	470Ω	¼ watt ± 10%
R_5, R_7	22,000Ω	1 watt ± 5% (or matched)
R_6	22,000Ω	1 watt ± 20%
R_8, R_9	0.47MΩ	¼ watt ± 20%
R_{10}	390Ω	¼ watt ± 10%
R_{11}, R_{12}	47,000Ω	2 watt ± 5% (or matched)

R_{14}, R_{19}	0.1MΩ	¼ watt ± 10%
R_{15}, R_{20}	1,000Ω	¼ watt ± 20%
R_{16}, R_{18}	100Ω	1 watt ± 20%
R_{17}, R_{21}	100Ω	2 watt wirewound variable
R_{22}	150Ω	3 watt ± 20%
R_{23}, R_{24}	100Ω	¼ watt ± 20%
R_{25}	$1{,}200\sqrt{\text{speech coil impedance}}$	¼ watt (see table)
R_{26}	4,700Ω	¼ watt ± 20%
C_1, C_2, C_5, C_8	8μF	500V wkg.
C_3, C_4	0.05μF	350V wkg.

C_6, C_7	0.25μF	350V wkg.
C_9	8μF	600V wkg.
C_{10}	200pF	350V wkg.
CH_1	30H at 20mA	
CH_2	10H at 150mA	
T	Power transformer	

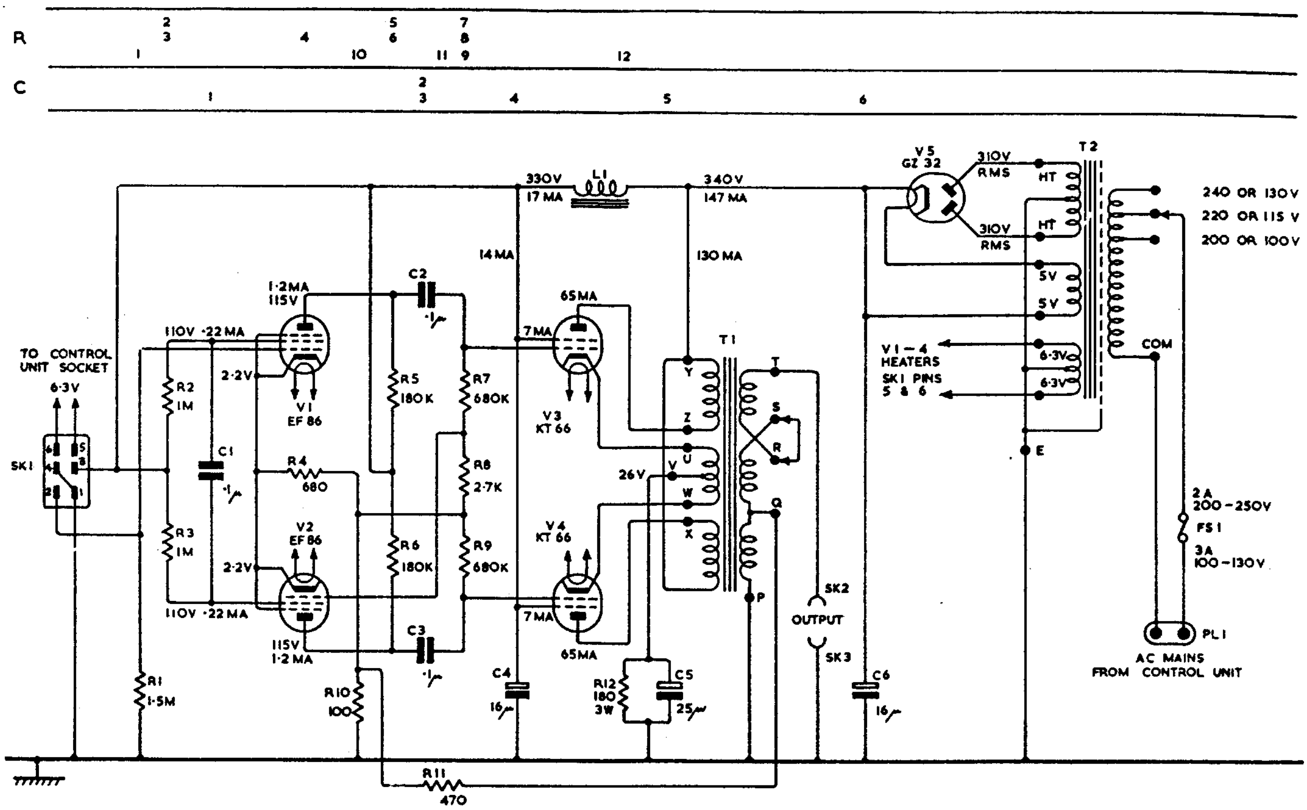
Secondary 425-0-425V 150 mA, 5V. 3A, 6.3V 4A, centre-tapped

V_1, V_2	2 × L63 or 6J5, 6SN7 or B65	
V_3, V_4	do.	do.
V_5, V_6	KT66	V_7 Cossor 53KU, 5V4

The Williamson Amplifier

D.T.N. Williamson was employed at Osram-Marconi, the company who developed and manufactured the K66 beam tetrode. His famous design was published in 1947 and it was a major step forward in terms of quality. Up to that time phase-splitting was normally performed by a transformer with a center-tapped secondary. The inclusion of two transformers in the signal path, made global feedback virtually impossible. The Williamson was able to deliver 15 Watts with a THD at 0.1% (at 1000Hz). It became a reference standard for more than a decade.

QUAD II POWER AMPLIFIER



DRG 11175. ISSUE 1.

THE VOLTAGE AND CURRENT MEASUREMENTS SHEWN ARE APPROXIMATE, AND ARE ONLY PROVIDED AS A GUIDE. ALLOWANCE SHOULD BE MADE FOR THE LOADING EFFECTS OF A VOLTMETER.

Shown here is the famous Quad II. It was a very well-made amplifier, and many of them are still in use after more than 30 years of service.

At first sight the diagram looks confusing, but after a closer look it becomes very simple.

The upper EF86 is quite a normal input stage. Signal for the lower EF86 is taken from the output of the upper via a voltage divider consisting of the 680 kw grid resistor of the upper KT66 and a 2.7 kw resistor.

The EF86^s both invert the signal so the driving voltages for the output stage are in opposite phase as required.

The voltage divider compensates for the gain of the lower EF86, where a gain of 680:2,7 » 250 apparently is expected. The actual gain is however much lower, but we find the explanation in the global feedback loop, where feedback is normal NFB to the upper EF86, but due to the inversion it is positive feedback to the lower EF86! This accounts for the high gain of that valve.

This combined input, phase splitter and driver stage is elegant, but the fact that the driving voltage to the lower EF86 has been exposed to HF cut-off and phase shift introduced by the loading of the output stage of the upper EF86 should not be overlooked.

It is unusual to tie the screen grids of the two valves together signal-wise instead of keeping them signal-free by connecting capacitors to ground from each grid, but this arrangement will keep the grids free from signal if they carry the same signal in opposite phase. If not, C_1 will help to restore AC balance from the two EF86^s.

Another unusual feature is the local feedback of the output stage by the cathode coils of the output transformer. It could be thought that this negative current feedback would raise output resistance of the entire stage. Because of the close coupling of all coils quite the opposite is the case. If output voltage decreases because of external load, the feedback voltage from the cathode coil will decrease too, which helps to restore the output signal. This local feedback of the output stage is a brilliant idea (but it raises demands for voltage swing on the grids!) and I should wish that such transformers were available. The ratio between anode – and cathode coils is approximately 10:1.

Despite its heavy weight and big transformers the Quad II was only rated as a 12W amplifier – and despite the measured performance is average, the sound is very good.

Because KT66 is hard to obtain and very expensive, it is often replaced by EL34 when Quad^{s} are restored. EL34^{s} works quite well, but remember to connect the 3rd grid to cathode as this is not done internally in EL34 (Pin 1 strapped to 8). Apart from that, the socket connections are the same. The 180w cathode resistor should be replaced by a 240w resistor (two 470w 5W resistors in parallel). As there is no provision for adjusting DC balance (and no self-balancing!) the output valves should preferably be matched. A 220w resistor in the supply line to the screen grids is recommended.

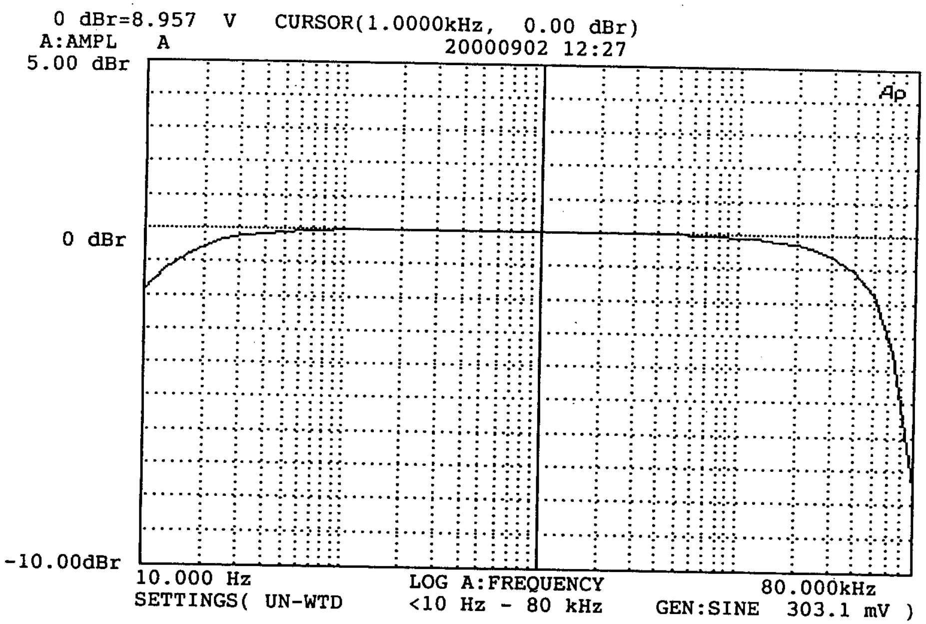
The GZ32 rectifier is often replaced by solid state diodes (1N4007), which also allows for a higher value of the surge capacitor C6, where 100m/450V would be a considerably improvement. Remember that this is not allowable with the GZ32 because it can't cope with the peak currents of a 100mF capacitor.

You cannot change the amount of feedback without changing the ratio of the voltage divider in the grid resistor of the upper KT66, and this is not advisable.

Despite reservations, I have always been fascinated by the circuit. Although it is easy to understand how it works, it has always been an enigma to me how it was conceived!

Appendix III

performance curves

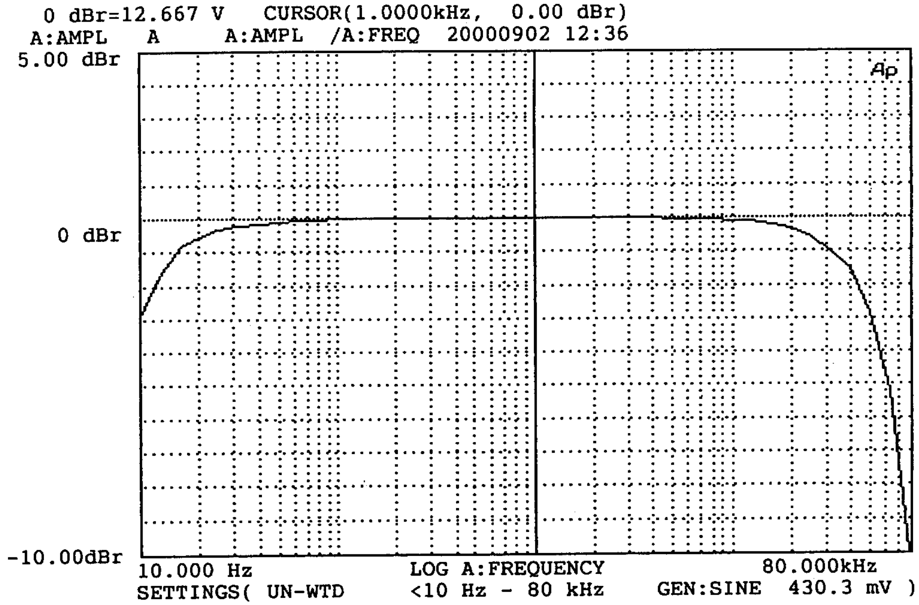


AGEN FREQ	AMPL	A
10.000 Hz	-1.88	dBr
12.500 Hz	-1.28	dBr
16.000 Hz	-0.82	dBr
20.000 Hz	-0.54	dBr
25.000 Hz	-0.36	dBr
31.500 Hz	-0.23	dBr
40.000 Hz	-0.15	dBr
50.000 Hz	-0.10	dBr
63.000 Hz	-0.06	dBr
80.000 Hz	-0.04	dBr
100.00 Hz	-0.03	dBr
125.00 Hz	-0.01	dBr
160.00 Hz	-0.01	dBr
200.00 Hz	0.00	dBr

AGEN FREQ	AMPL	A
250.00 Hz	0.00	dBr
315.00 Hz	0.00	dBr
400.00 Hz	0.00	dBr
500.00 Hz	0.00	dBr
630.00 Hz	0.00	dBr
800.00 Hz	0.00	dBr
1.0000kHz	0.00	dBr
1.2500kHz	-0.01	dBr
1.6000kHz	-0.01	dBr
2.0000kHz	-0.01	dBr
2.5000kHz	-0.01	dBr
3.1500kHz	-0.01	dBr
4.0000kHz	-0.02	dBr
5.0000kHz	-0.03	dBr

AGEN FREQ	AMPL	A	Ap
6.3000kHz	-0.05	dBr	
8.0000kHz	-0.08	dBr	
10.000kHz	-0.10	dBr	
12.500kHz	-0.14	dBr	
16.000kHz	-0.20	dBr	
20.000kHz	-0.30	dBr	
25.000kHz	-0.46	dBr	
31.500kHz	-0.72	dBr	
40.000kHz	-1.09	dBr	
50.000kHz	-1.74	dBr	
63.000kHz	-3.47	dBr	
80.000kHz	-7.48	dBr	

Frequency response measured at 10W out

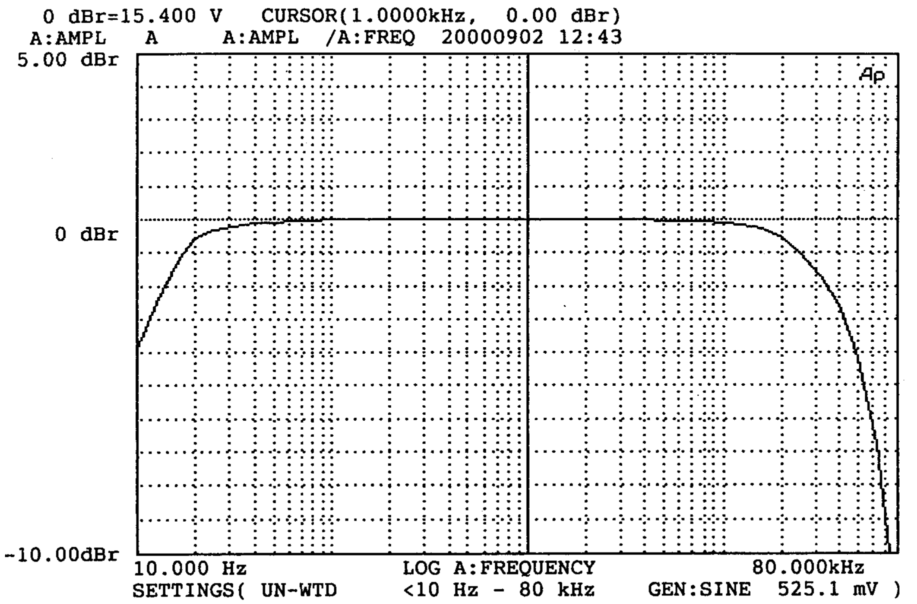


AGEN FREQ	AMPL	A
10.000 Hz	-2.90	dBr
12.500 Hz	-1.69	dBr
16.000 Hz	-0.82	dBr
20.000 Hz	-0.54	dBr
25.000 Hz	-0.36	dBr
31.500 Hz	-0.22	dBr
40.000 Hz	-0.14	dBr
50.000 Hz	-0.09	dBr
63.000 Hz	-0.06	dBr
80.000 Hz	-0.03	dBr
100.00 Hz	-0.02	dBr
125.00 Hz	-0.02	dBr
160.00 Hz	-0.01	dBr
200.00 Hz	0.00	dBr

AGEN FREQ	AMPL	A
250.00 Hz	0.00	dBr
315.00 Hz	0.00	dBr
400.00 Hz	0.00	dBr
500.00 Hz	0.00	dBr
630.00 Hz	0.00	dBr
800.00 Hz	0.00	dBr
1.0000kHz	0.00	dBr
1.2500kHz	0.00	dBr
1.6000kHz	0.00	dBr
2.0000kHz	-0.01	dBr
2.5000kHz	-0.01	dBr
3.1500kHz	-0.01	dBr
4.0000kHz	-0.02	dBr
5.0000kHz	-0.03	dBr

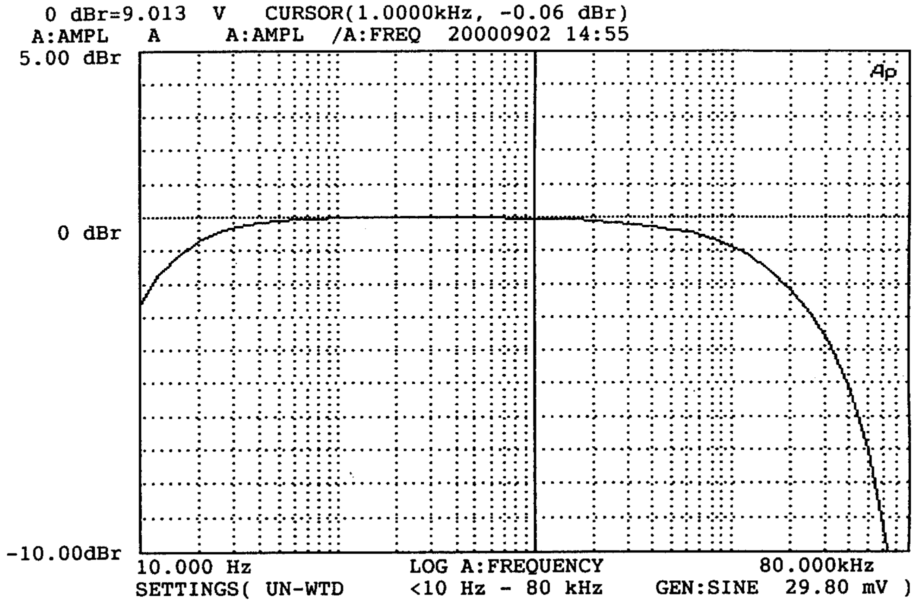
AGEN FREQ	AMPL	A	Ap
6.3000kHz	-0.05	dBr	
8.0000kHz	-0.08	dBr	
10.000kHz	-0.10	dBr	
12.500kHz	-0.14	dBr	
16.000kHz	-0.21	dBr	
20.000kHz	-0.34	dBr	
25.000kHz	-0.60	dBr	
31.500kHz	-1.02	dBr	
40.000kHz	-1.51	dBr	
50.000kHz	-2.83	dBr	
63.000kHz	-5.18	dBr	
80.000kHz	-10.15	dBr	

Frequency response measured at 20W out



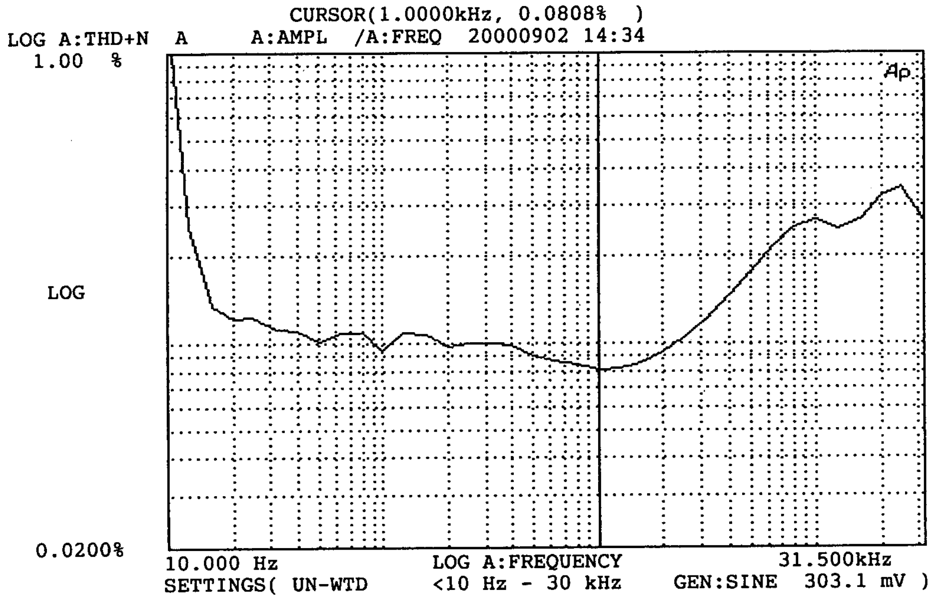
AGEN FREQ	AMPL A	AGEN FREQ	AMPL A	AGEN FREQ	AMPL A Ap
10.000 Hz	-3.87 dBr	250.00 Hz	0.00 dBr	6.3000kHz	-0.05 dBr
12.500 Hz	-2.51 dBr	315.00 Hz	0.00 dBr	8.0000kHz	-0.08 dBr
16.000 Hz	-1.25 dBr	400.00 Hz	0.00 dBr	10.000kHz	-0.11 dBr
20.000 Hz	-0.54 dBr	500.00 Hz	-0.01 dBr	12.500kHz	-0.16 dBr
25.000 Hz	-0.35 dBr	630.00 Hz	0.00 dBr	16.000kHz	-0.29 dBr
31.500 Hz	-0.22 dBr	800.00 Hz	0.00 dBr	20.000kHz	-0.55 dBr
40.000 Hz	-0.14 dBr	1.0000kHz	0.00 dBr	25.000kHz	-1.07 dBr
50.000 Hz	-0.09 dBr	1.2500kHz	0.00 dBr	31.500kHz	-1.69 dBr
63.000 Hz	-0.05 dBr	1.6000kHz	-0.01 dBr	40.000kHz	-2.58 dBr
80.000 Hz	-0.03 dBr	2.0000kHz	-0.01 dBr	50.000kHz	-4.16 dBr
100.00 Hz	-0.02 dBr	2.5000kHz	-0.01 dBr	63.000kHz	-6.80 dBr
125.00 Hz	-0.01 dBr	3.1500kHz	-0.01 dBr	80.000kHz	-11.78dBr
160.00 Hz	0.00 dBr	4.0000kHz	-0.02 dBr		
200.00 Hz	0.00 dBr	5.0000kHz	-0.04 dBr		

Frequency response measured at 30W out



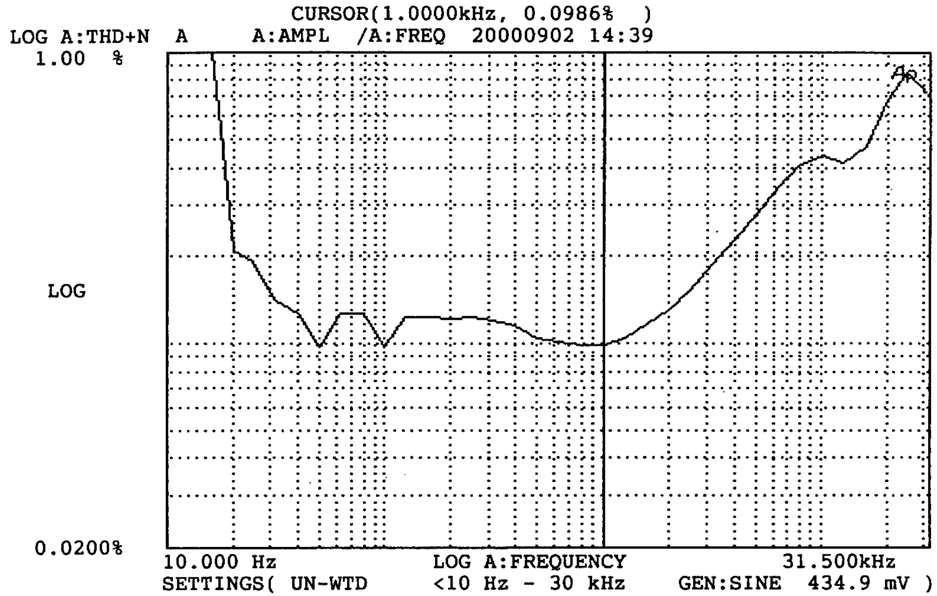
AGEN FREQ	AMPL A	AGEN FREQ	AMPL A	AGEN FREQ	AMPL A Ap
10.000 Hz	-2.61 dBr	250.00 Hz	-0.02 dBr	6.3000kHz	-0.44 dBr
12.500 Hz	-1.67 dBr	315.00 Hz	-0.03 dBr	8.0000kHz	-0.68 dBr
16.000 Hz	-1.08 dBr	400.00 Hz	-0.02 dBr	10.000kHz	-0.89 dBr
20.000 Hz	-0.71 dBr	500.00 Hz	-0.02 dBr	12.500kHz	-1.15 dBr
25.000 Hz	-0.46 dBr	630.00 Hz	-0.02 dBr	16.000kHz	-1.63 dBr
31.500 Hz	-0.30 dBr	800.00 Hz	-0.04 dBr	20.000kHz	-2.17 dBr
40.000 Hz	-0.17 dBr	1.0000kHz	-0.06 dBr	25.000kHz	-2.88 dBr
50.000 Hz	-0.09 dBr	1.2500kHz	-0.06 dBr	31.500kHz	-3.78 dBr
63.000 Hz	-0.08 dBr	1.6000kHz	-0.08 dBr	40.000kHz	-5.09 dBr
80.000 Hz	-0.04 dBr	2.0000kHz	-0.10 dBr	50.000kHz	-7.02 dBr
100.00 Hz	-0.02 dBr	2.5000kHz	-0.15 dBr	63.000kHz	-10.24dBr
125.00 Hz	0.00 dBr	3.1500kHz	-0.22 dBr	80.000kHz	-15.76dBr
160.00 Hz	-0.01 dBr	4.0000kHz	-0.29 dBr		
200.00 Hz	0.00 dBr	5.0000kHz	-0.38 dBr		

Frequency response measured at 10W out
without negative feedback



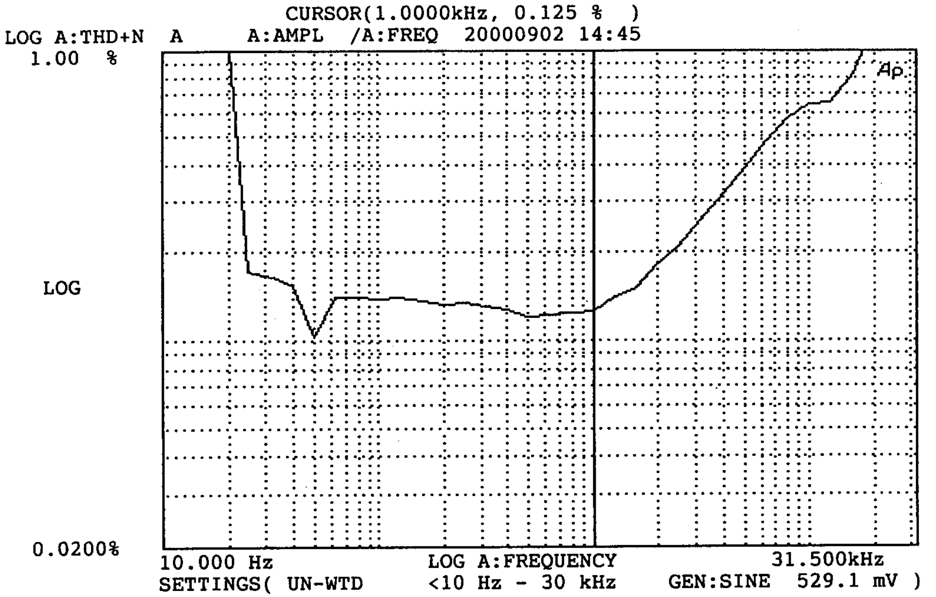
AGEN FREQ	THD+N	A	AGEN FREQ	THD+N	A	AGEN FREQ	THD+N	A	Ap
10.000 Hz	1.35	%	200.00 Hz	0.0973	%	4.0000kHz	0.147	%	
12.500 Hz	0.257	%	250.00 Hz	0.100	%	5.0000kHz	0.177	%	
16.000 Hz	0.134	%	315.00 Hz	0.100	%	6.3000kHz	0.216	%	
20.000 Hz	0.121	%	400.00 Hz	0.0985	%	8.0000kHz	0.252	%	
25.000 Hz	0.124	%	500.00 Hz	0.0910	%	10.000kHz	0.265	%	
31.500 Hz	0.111	%	630.00 Hz	0.0868	%	12.500kHz	0.249	%	
40.000 Hz	0.111	%	800.00 Hz	0.0837	%	16.000kHz	0.267	%	
50.000 Hz	0.101	%	1.0000kHz	0.0808	%	20.000kHz	0.320	%	
63.000 Hz	0.109	%	1.2500kHz	0.0823	%	25.000kHz	0.343	%	
80.000 Hz	0.109	%	1.6000kHz	0.0851	%	31.500kHz	0.262	%	
100.00 Hz	0.0950	%	2.0000kHz	0.0933	%				
125.00 Hz	0.108	%	2.5000kHz	0.103	%				
160.00 Hz	0.107	%	3.1500kHz	0.121	%				

Total harmonic distortion + noise measured at 10W out



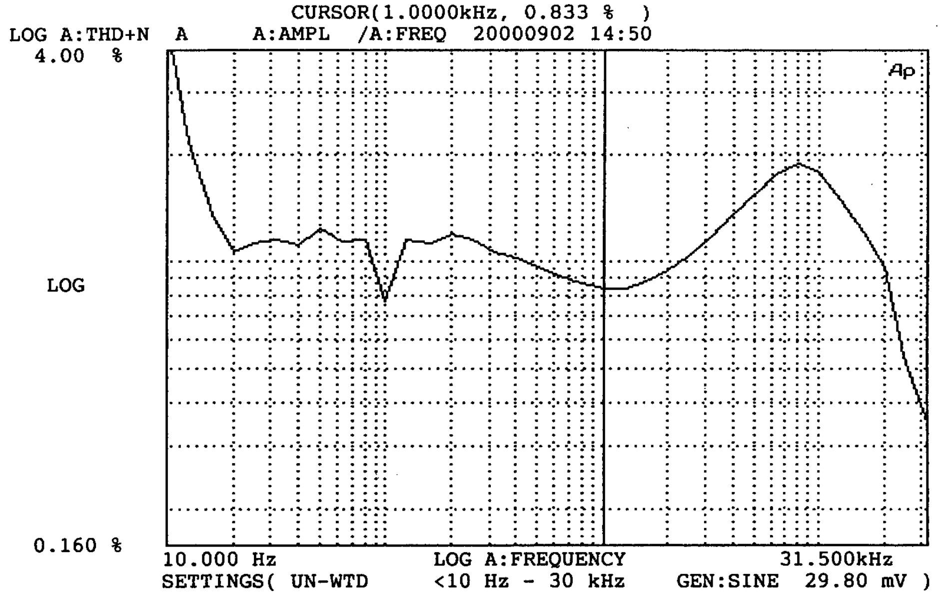
AGEN FREQ	THD+N	A	AGEN FREQ	THD+N	A	AGEN FREQ	THD+N	A	Ap
10.000 Hz	35.36	%	200.00 Hz	0.121	%	4.0000kHz	0.228	%	
12.500 Hz	21.53	%	250.00 Hz	0.123	%	5.0000kHz	0.280	%	
16.000 Hz	1.08	%	315.00 Hz	0.120	%	6.3000kHz	0.344	%	
20.000 Hz	0.209	%	400.00 Hz	0.114	%	8.0000kHz	0.412	%	
25.000 Hz	0.192	%	500.00 Hz	0.103	%	10.000kHz	0.439	%	
31.500 Hz	0.140	%	630.00 Hz	0.101	%	12.500kHz	0.416	%	
40.000 Hz	0.127	%	800.00 Hz	0.0984	%	16.000kHz	0.474	%	
50.000 Hz	0.0976	%	1.0000kHz	0.0986	%	20.000kHz	0.680	%	
63.000 Hz	0.126	%	1.2500kHz	0.104	%	25.000kHz	0.844	%	
80.000 Hz	0.127	%	1.6000kHz	0.117	%	31.500kHz	0.702	%	
100.00 Hz	0.0969	%	2.0000kHz	0.131	%				
125.00 Hz	0.123	%	2.5000kHz	0.154	%				
160.00 Hz	0.123	%	3.1500kHz	0.186	%				

Total harmonic distortion + noise measured at 20W out



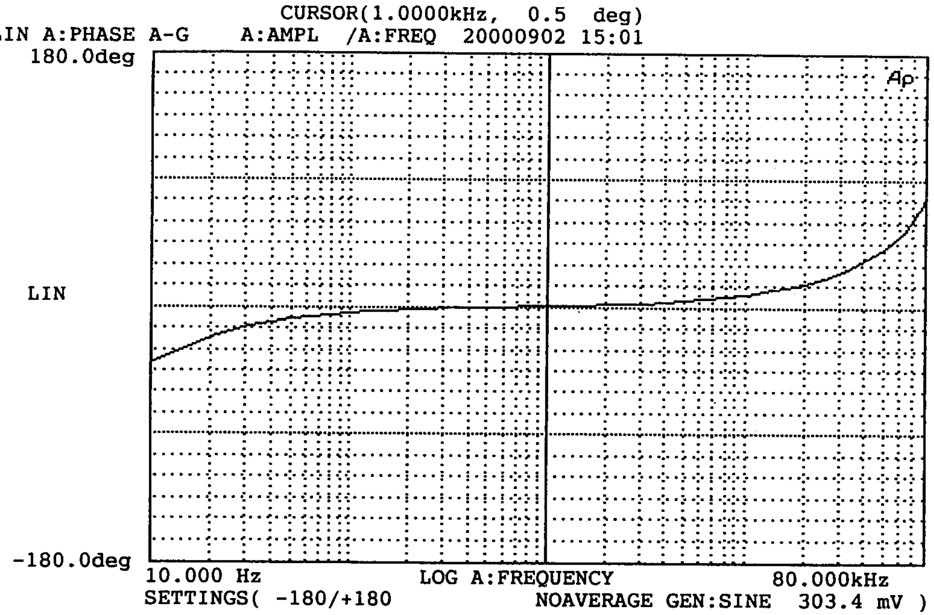
AGEN FREQ	THD+N A	AGEN FREQ	THD+N A	AGEN FREQ	THD+N A Ap
10.000 Hz	45.45 %	200.00 Hz	0.132 %	4.0000kHz	0.316 %
12.500 Hz	37.83 %	250.00 Hz	0.135 %	5.0000kHz	0.388 %
16.000 Hz	21.53 %	315.00 Hz	0.131 %	6.3000kHz	0.481 %
20.000 Hz	1.01 %	400.00 Hz	0.128 %	8.0000kHz	0.576 %
25.000 Hz	0.171 %	500.00 Hz	0.120 %	10.000kHz	0.645 %
31.500 Hz	0.165 %	630.00 Hz	0.122 %	12.500kHz	0.657 %
40.000 Hz	0.152 %	800.00 Hz	0.123 %	16.000kHz	0.827 %
50.000 Hz	0.102 %	1.0000kHz	0.125 %	20.000kHz	1.20 %
63.000 Hz	0.141 %	1.2500kHz	0.140 %	25.000kHz	1.74 %
80.000 Hz	0.140 %	1.6000kHz	0.150 %	31.500kHz	1.12 %
100.00 Hz	0.138 %	2.0000kHz	0.180 %		
125.00 Hz	0.139 %	2.5000kHz	0.209 %		
160.00 Hz	0.137 %	3.1500kHz	0.256 %		

Total harmonic distortion + noise vs. frequency at 30W out



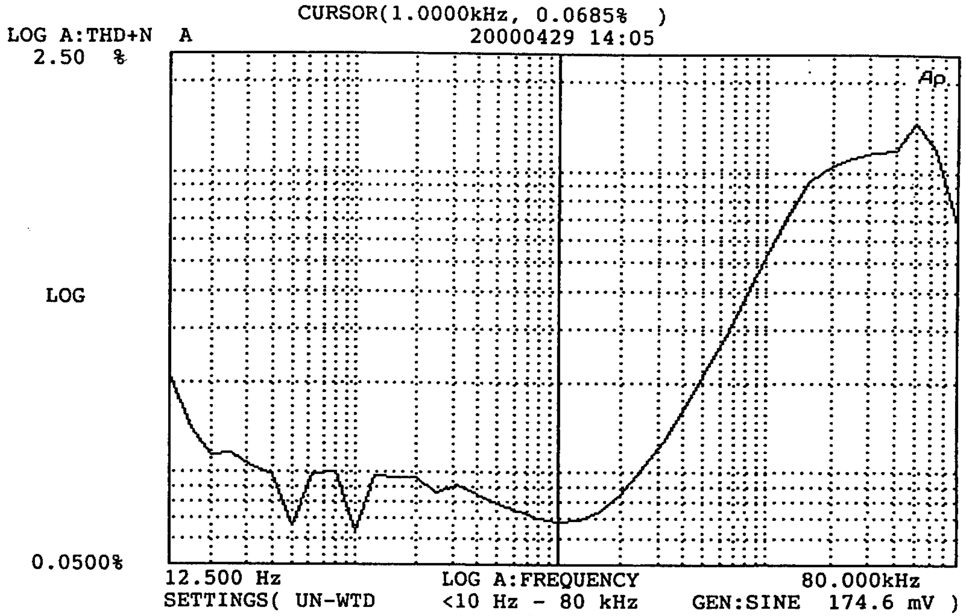
AGEN FREQ	THD+N A	AGEN FREQ	THD+N A	AGEN FREQ	THD+N A Ap
10.000 Hz	4.60 %	200.00 Hz	1.19 %	4.0000kHz	1.34 %
12.500 Hz	2.16 %	250.00 Hz	1.15 %	5.0000kHz	1.54 %
16.000 Hz	1.34 %	315.00 Hz	1.05 %	6.3000kHz	1.75 %
20.000 Hz	1.05 %	400.00 Hz	1.02 %	8.0000kHz	1.87 %
25.000 Hz	1.12 %	500.00 Hz	0.958 %	10.000kHz	1.76 %
31.500 Hz	1.14 %	630.00 Hz	0.902 %	12.500kHz	1.48 %
40.000 Hz	1.10 %	800.00 Hz	0.861 %	16.000kHz	1.20 %
50.000 Hz	1.23 %	1.0000kHz	0.833 %	20.000kHz	0.960 %
63.000 Hz	1.13 %	1.2500kHz	0.838 %	25.000kHz	0.540 %
80.000 Hz	1.14 %	1.6000kHz	0.875 %	31.500kHz	0.350 %
100.00 Hz	0.768 %	2.0000kHz	0.941 %		
125.00 Hz	1.14 %	2.5000kHz	1.02 %		
160.00 Hz	1.11 %	3.1500kHz	1.16 %		

Total harmonic distortion + noise vs. frequency
measured at 10W out *without NFB*



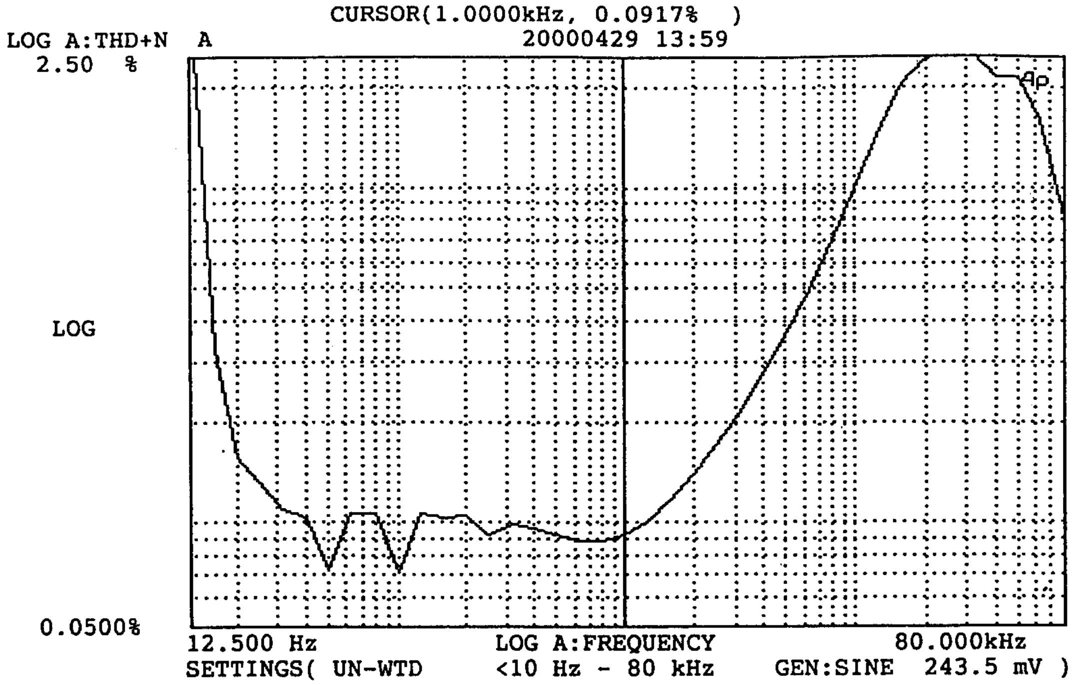
AGEN FREQ	PHASE A-G	AGEN FREQ	PHASE A-G	AGEN FREQ	PHASE A-G Ap
10.000 Hz	-39.5 deg	250.00 Hz	-1.4 deg	6.3000kHz	5.6 deg
12.500 Hz	-32.7 deg	315.00 Hz	-1.0 deg	8.0000kHz	7.0 deg
16.000 Hz	-26.3 deg	400.00 Hz	-0.6 deg	10.000kHz	8.7 deg
20.000 Hz	-21.4 deg	500.00 Hz	-0.3 deg	12.500kHz	10.9 deg
25.000 Hz	-17.2 deg	630.00 Hz	0.0 deg	16.000kHz	13.7 deg
31.500 Hz	-13.8 deg	800.00 Hz	0.2 deg	20.000kHz	17.1 deg
40.000 Hz	-10.9 deg	1.0000kHz	0.5 deg	25.000kHz	21.1 deg
50.000 Hz	-8.7 deg	1.2500kHz	0.9 deg	31.500kHz	26.4 deg
63.000 Hz	-6.9 deg	1.6000kHz	1.3 deg	40.000kHz	33.2 deg
80.000 Hz	-5.4 deg	2.0000kHz	1.7 deg	50.000kHz	41.5 deg
100.00 Hz	-4.2 deg	2.5000kHz	2.2 deg	63.000kHz	53.8 deg
125.00 Hz	-3.3 deg	3.1500kHz	2.8 deg	80.000kHz	75.0 deg
160.00 Hz	-2.5 deg	4.0000kHz	3.6 deg		
200.00 Hz	-1.9 deg	5.0000kHz	4.4 deg		

Phase shift input to output (measured at 10W out)



AGEN FREQ	THD+N A	AGEN FREQ	THD+N A	AGEN FREQ	THD+N A Ap
12.500 Hz	0.211 %	315.00 Hz	0.0908%	8.0000kHz	0.383 %
16.000 Hz	0.140 %	400.00 Hz	0.0850%	10.000kHz	0.524 %
20.000 Hz	0.114 %	500.00 Hz	0.0791%	12.500kHz	0.708 %
25.000 Hz	0.116 %	630.00 Hz	0.0748%	16.000kHz	0.935 %
31.500 Hz	0.104 %	800.00 Hz	0.0699%	20.000kHz	1.04 %
40.000 Hz	0.0982%	1.0000kHz	0.0685%	25.000kHz	1.11 %
50.000 Hz	0.0667%	1.2500kHz	0.0693%	31.500kHz	1.16 %
63.000 Hz	0.0994%	1.6000kHz	0.0749%	40.000kHz	1.18 %
80.000 Hz	0.100 %	2.0000kHz	0.0851%	50.000kHz	1.46 %
100.00 Hz	0.0633%	2.5000kHz	0.102 %	63.000kHz	1.19 %
125.00 Hz	0.0977%	3.1500kHz	0.125 %	80.000kHz	0.670 %
160.00 Hz	0.0962%	4.0000kHz	0.161 %		
200.00 Hz	0.0965%	5.0000kHz	0.209 %		
250.00 Hz	0.0859%	6.3000kHz	0.278 %		

Total harmonic distortion + noise at 8W
Output valves operating as triodes (R_{g2} connected to anode)
15dB of negative feedback

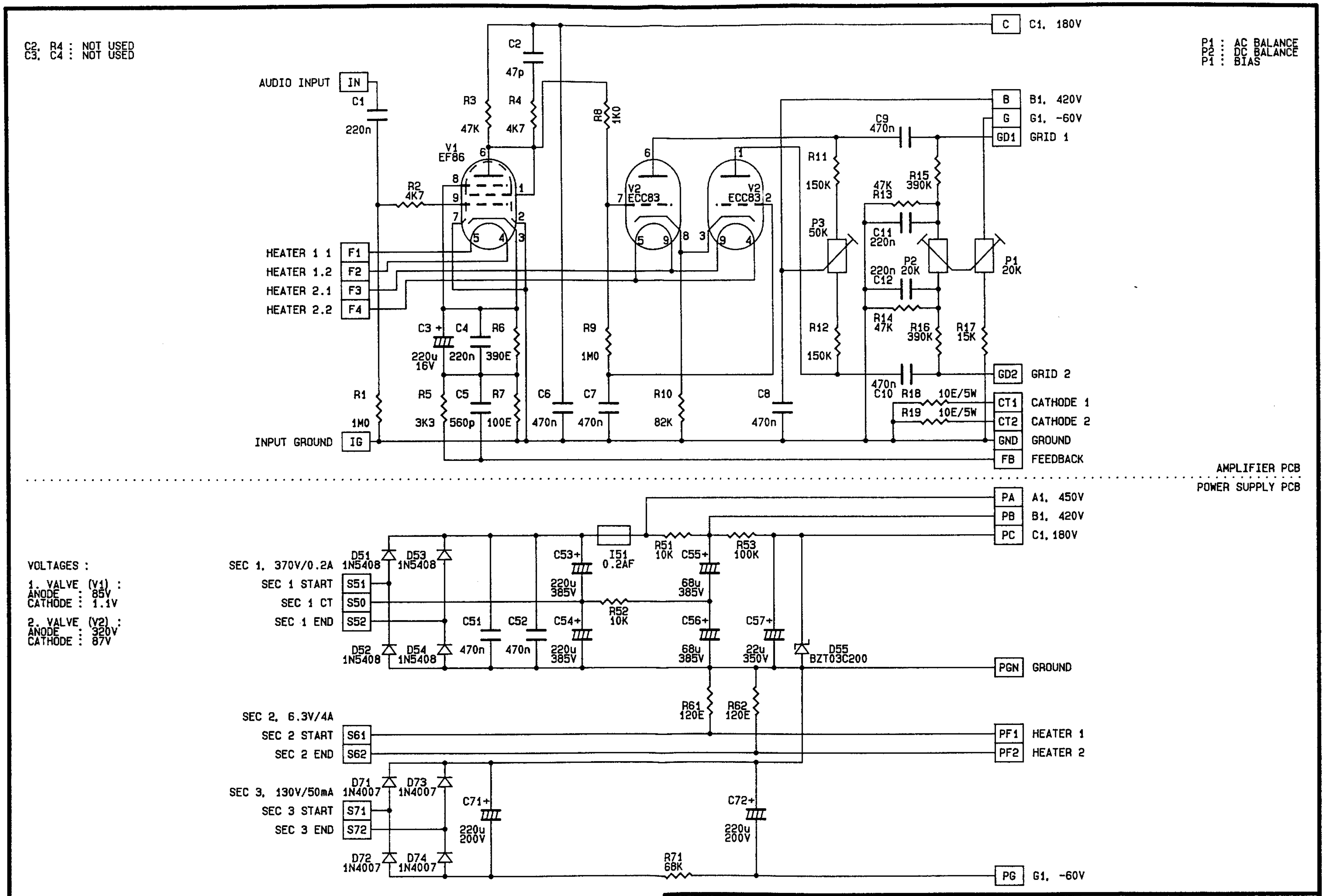


AGEN FREQ	THD+N A	AGEN FREQ	THD+N A	AGEN FREQ	THD+N A Ap
12.500 Hz	4.69 %	315.00 Hz	0.0997%	8.0000kHz	0.695 %
16.000 Hz	0.327 %	400.00 Hz	0.0958%	10.000kHz	0.991 %
20.000 Hz	0.155 %	500.00 Hz	0.0919%	12.500kHz	1.43 %
25.000 Hz	0.132 %	630.00 Hz	0.0885%	16.000kHz	2.05 %
31.500 Hz	0.110 %	800.00 Hz	0.0881%	20.000kHz	2.44 %
40.000 Hz	0.103 %	1.0000kHz	0.0917%	25.000kHz	2.55 %
50.000 Hz	0.0724%	1.2500kHz	0.100 %	31.500kHz	2.55 %
63.000 Hz	0.105 %	1.6000kHz	0.117 %	40.000kHz	2.15 %
80.000 Hz	0.106 %	2.0000kHz	0.140 %	50.000kHz	2.13 %
100.00 Hz	0.0715%	2.5000kHz	0.169 %	63.000kHz	1.58 %
125.00 Hz	0.106 %	3.1500kHz	0.212 %	80.000kHz	0.799 %
160.00 Hz	0.103 %	4.0000kHz	0.278 %		
200.00 Hz	0.104 %	5.0000kHz	0.362 %		
250.00 Hz	0.0920%	6.3000kHz	0.494 %		

Total harmonic distortion + noise at 16W
Output valves operating as triodes (R_{g2} connected to anode)
15dB of negative feedback

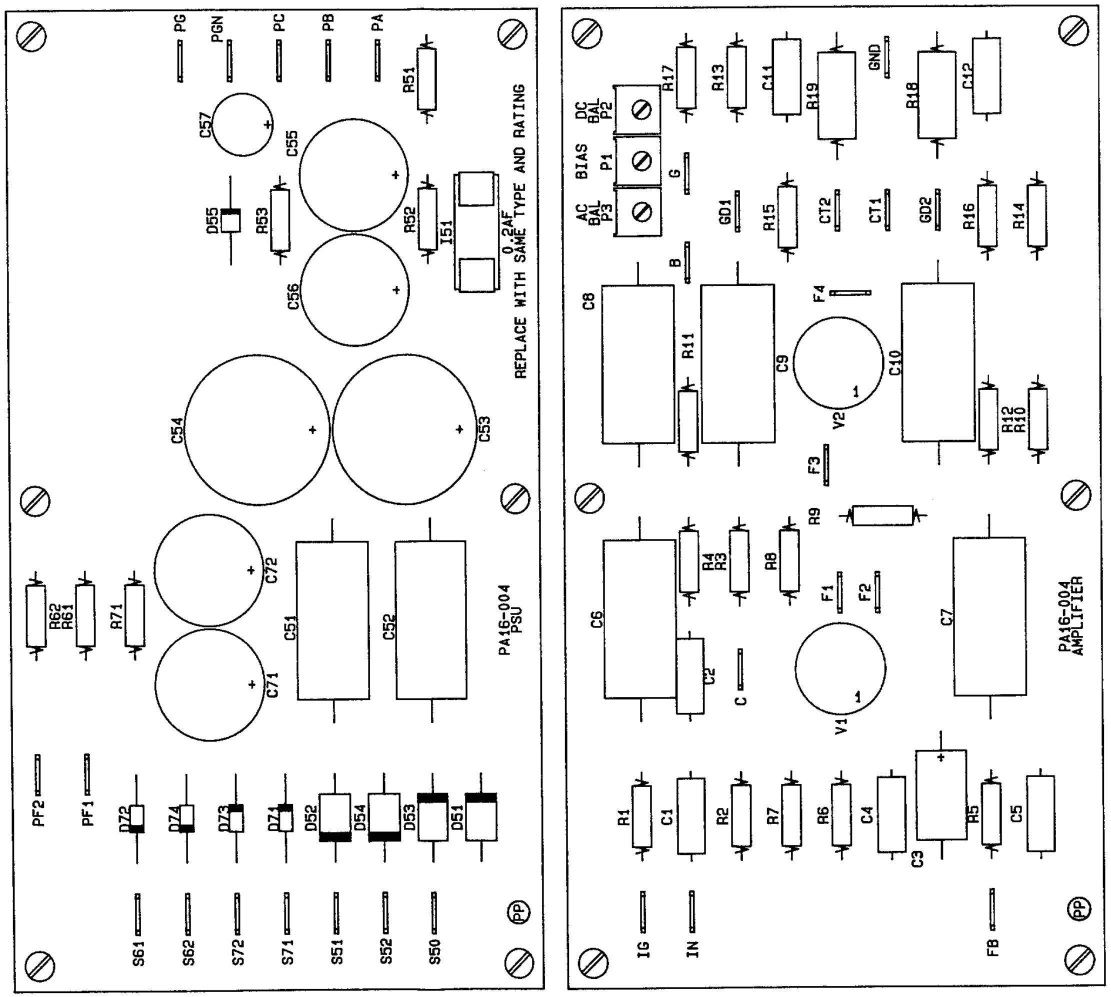
Appendix IV

PCB's and layout



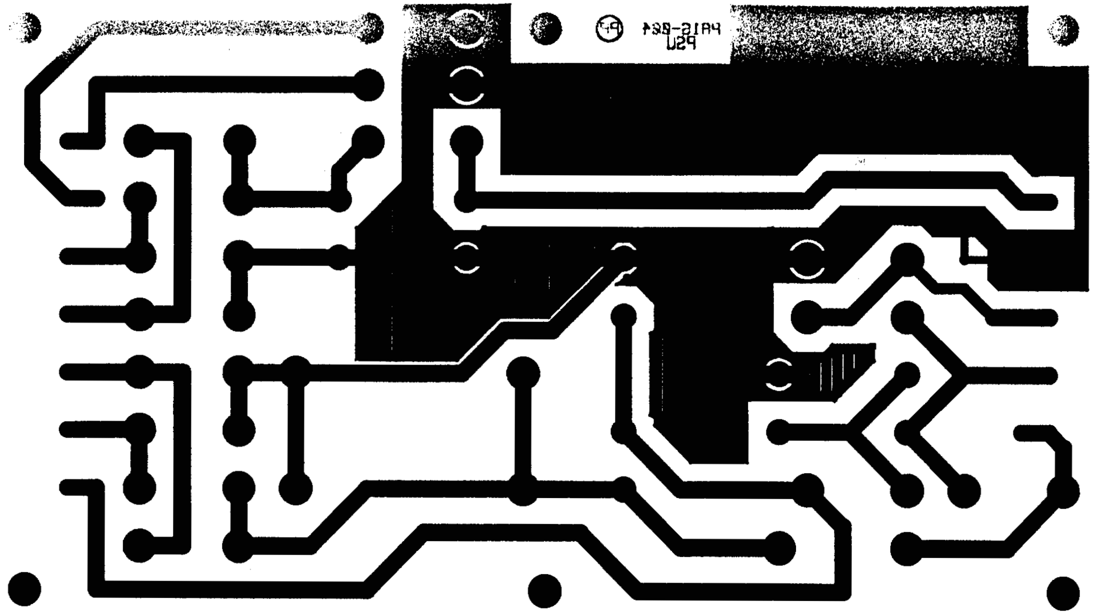
The values of R51 and R53 are valid for mono blocks

VALVEAMPLIFIER PA16	ENG: PP
	VERS: 0031.3
	REPL: 0004.2
NPN ELEKTRONIK APS	DRAWING NUMBER 00012901
SCALE	

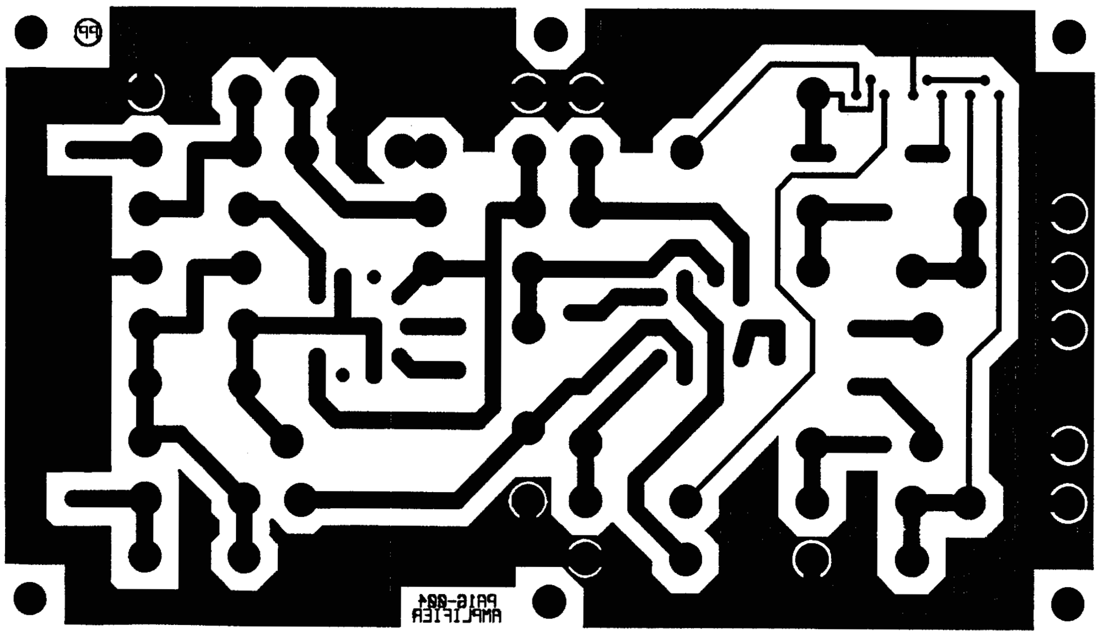


COMPONENTS D55, GND, IG, PGN, R61, R62 AND V1 MUST BE SOLDERED ON BOTH SIDES OF PCB.
OTHER COMPONENTS CONNECTED TO GROUND ON COMPONENTSIDE SHOULD BE SOLDERED ON COMPONENTSIDE IF POSSIBLE.

PA16 COMPONENTLAYOUT	ENG	PP
	VERS	0031.3
	REPL	0004.2
NPN ELEKTRONIK APS	DRAWING NUMBER 00012903	
SCALE		



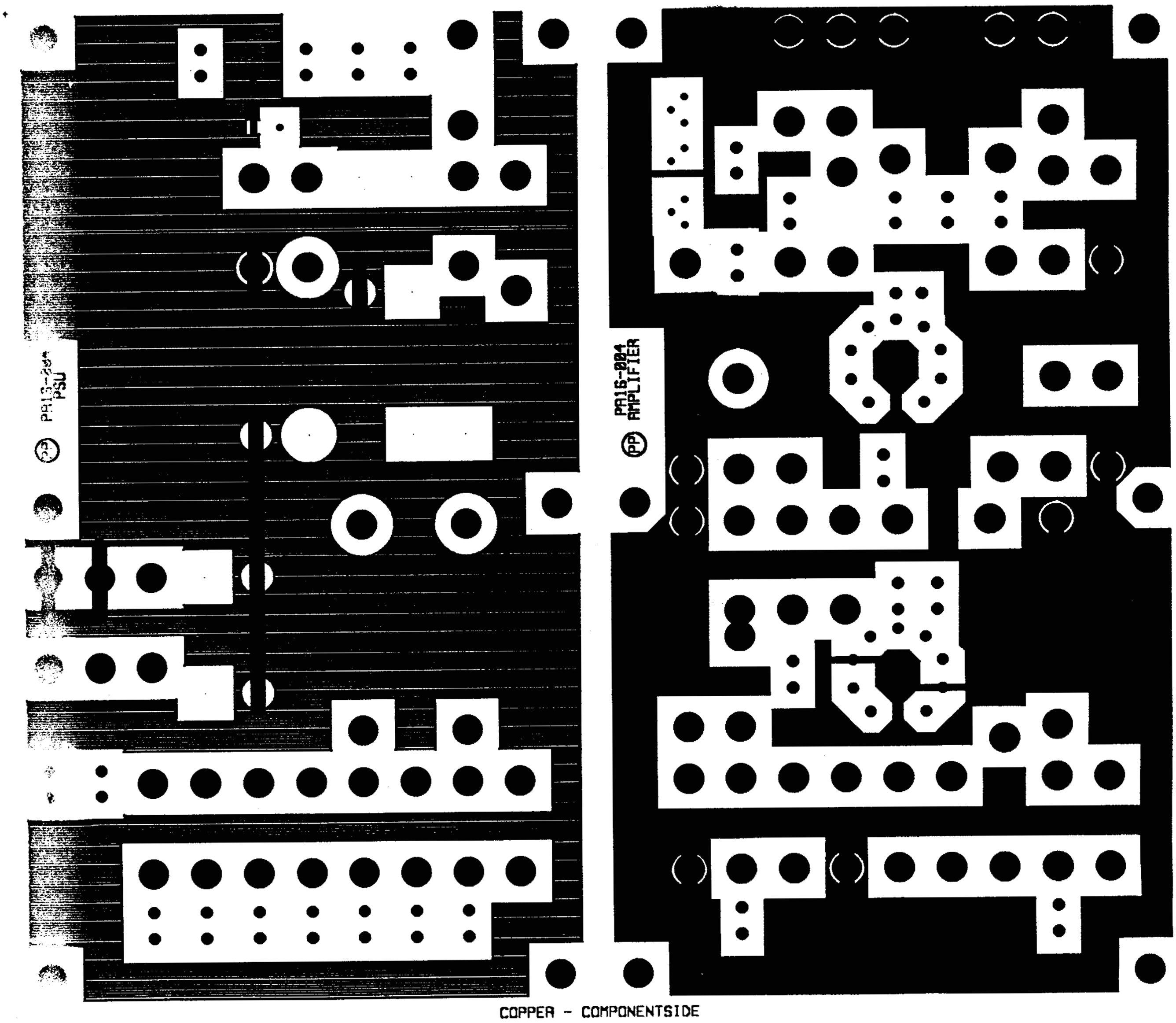
Power supply PCB
Track side



Amplifier PCB
Track side
resistors and capacitors mounted here

Power supply PCB
Component side

Amplifier PCB
Screen side
Trimpots and sockets mounted here



COPPER - COMPONENTSIDE

DET JYSKE
MUSIKKONSERVATORIUM