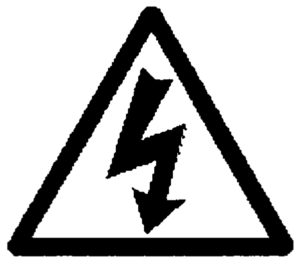
Designing Tube Preamps for Guitar and Bass

Merlin Blencowe



Caution!

Risk of electric shock.

The circuits and subjects discussed in this book operate from or involve dangerously high voltages. Do not attempt to build any of the circuits described here, or work on live equipment, without proper training. It is the reader's responsibility to ensure that any equipment constructed or modified, using the information given in this book, is safe to use.

Designing Valve Preamps for Guitar and Bass

by

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2009

Published by Merlin Blencowe

ISBN 978-0-9561545-0-7

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Preface

"This is the book I wish I'd had, when I began building amps"

This book began as a collection of design notes acquired from numerous sources by the author, as a handy personal reference. However, I soon found other amp builders asking for similar information on internet forums, or struggling with simple design problems which were nonetheless outside the expertise of 'conventional' audio circuit design, so these notes grew into a small website. The website received such an overwhelming response from thankful enthusiasts around the world that it became obvious that there must be a deplorable lack of readily-available information dedicated to valve guitar amp design. Certainly there is no shortage of wonderful, classic valve books from the 1940s and 50s, and more modern 'hifi' texts, but almost nothing written from the perspective of deliberate distortion and tone manipulation for musical instruments. Of the handful of books which *are* available, they are at best only descriptions of pre-existing or 'classic' amps, and at worst badly written, badly presented, rife with errors and heavily overpriced. After some frustration with the current state of knowledge, I have attempted to produce an accurate, succinct and comprehensive treatise on the design of guitar and bass preamps: a textbook. This required considerable research, as well as the collection of much new information on valves and circuit properties.

I have assumed some prior electronics knowledge and experience on the part of the reader. Complete beginners are forewarned that it will not be easy reading if they have never built an amp and have no basic knowledge of simple electronic principles, such as Ohm's law, simple power equations and how capacitors, inductors and valves work. Any introductory electronics textbook will provide these things.

Nevertheless, I have assumed an audience of mainly amateur hobbyists, not professionals, so I have abandoned many of the 'traditional' approaches to circuit analysis which, as a beginner, I remember being a hindrance to understanding, rather than a help. Also, the order of contents within each chapter is not consistent, as I have attempted to explain the operation of each circuit block in a manner which I hope aids the *understanding of its operation*, and this varies with different circuit arrangements. This book is really intended to be read from front to back, initially, as it introduces many terms and concepts in earlier chapters which are merely assumed or expounded upon in later chapters.

I have not given derivations for equations since I believe they will not be of much use to most readers, and will only serve to increase the word count -and cost- of this book. Instead, only relevant design formulae are given, and references to derivations are offered, for those who wish to dig deeper.

No doubt all of this will irritate electronics graduates, but then, such graduates hardly need this book!

Acknowledgements

The author would particularly like to thank, in no particular order: Stephen Keller for his infinite help in sourcing reference material, Zoe and Iain Williams-Hartney for endless inspiration, Danny Noordzy for curve tracers and general encouragement, Alex Kenis for high-gain expertise and all the members of Gothik Amps and AX84.com/bbs, for pushing boundaries and always asking questions. Also, this book could not have been accomplished without:

Chris Hurley
David Ivan James
John Lawley-Thomas
Mallory Nicholls
Holly Smith
Lee Taylor
Albert Worthington
The University of Leeds Library

Merlin Blencowe

Chapter 1: The Common Cathode, Triode Gain Stage

Fundamentals of amplification. The load line. Biasing and distortion. Limitations on bias. Applying bias. Designing a simple, triode gain stage. The valve constants. Mathematical treatment of a gain stage. Summary of formulae.

In order to understand an amplifier as a whole, we must first understand its basic principles. Most amplifiers follow the same basic arrangement or 'topology', so we can begin by examining how a simple triode valve can be used to amplify a signal voltage, and how it is used to generate distortion. Almost all guitar amplifiers have a preamp consisting of a number of ECC83 / 12AX7 triodes, while the ECC81 / 12AT7, ECC82 / 12AU7 and 12AY7 occasionally appear in some designs. The consistent use of the same type of valves is partly historical, since so many latter-day amplifiers are derivations (or merely copies!) of a few classic Fender amps. However, the ECC83 does have some properties which make it ideal for use in overdriven designs, which will become apparent. Although there are hundreds of valves which are worth experimenting with, commercial designs are bound to use the same, readily available valve types, if only to satisfy consumer expectations. Consequently the ECC83 is likely to remain the 'triode of choice' for guitar amps, indefinitely, so much of this book will focus on its use.

For readers who are not familiar with the many designation numbers it is worth mentioning that the ECC83, ECC803, CV4004, M8137, 12AX7, 7025 and 6681 are all the same valve. The different numbers indicate either different manufacturers or special quality versions, but they all have the same electrical characteristics as far as guitar amps are concerned, and all can be used in the same circuit. Additional letters (such as 12AX7A) need not concern us; they were once used to indicate a controlled heater warm-up time or some other feature particular to that iteration of the valve, but with most current-production versions any extra letters may be assumed to be a gimmick.

Some versions are reputed to have a particular tonal character, and much has been written about the apparent superiority of, say, the Mullard 'long plate' ECC83, or the RCA 'black plate' 12AX7. These subjective differences are not a consideration for the circuit designer and will not be mentioned here again. So-called 'tube rolling' and 'cork sniffing' is left to the discretion of the reader. Real tonal control comes from the choice of topology, frequency shaping and manipulation of overdrive characteristics, and from a complete understanding of the circuit's functionality, not from the particular manufacturer or vintage of the actual components used.

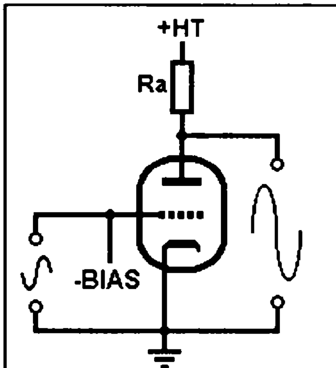

Fig 1.1: Simplified common-cathode, triode gain stage.

Fundamentals of amplification:

Fig. 1.1 shows a simplified circuit using a triode. It has two input terminals and two output terminals, and is sometimes known as **quadripole**. One input pole and one output pole are connected together at the cathode, so they are 'common' to one another, and are usually connected to ground. This is the **common-cathode** or **grounded-cathode gain stage**, and it is the fundamental

building block of all amplifiers, and we must understand this completely before we can ever hope to design a whole amp.
In hifi, a circuit like this is intended only to amplify a small input voltage into a larger output voltage capable of driving the following part of the amp, ideally with no change in the shape of the signal except an increase in amplitude. In other words, no distortion. In a guitar amp we can use this stage to do the same, but also to deliberately produce distortion and tone shaping.

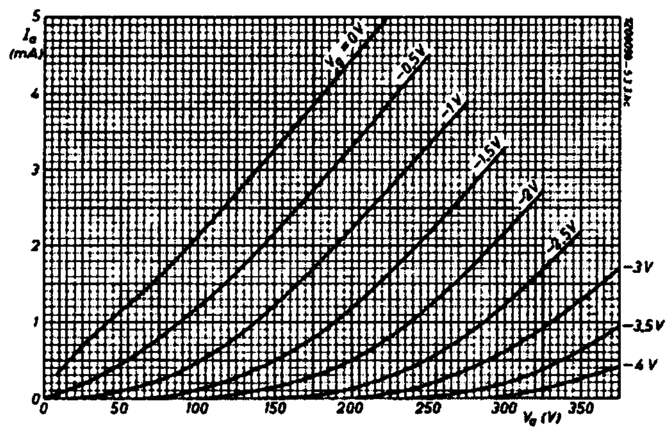

Fig. 1.2: Static anode characteristics graph of the ECC83 / 12AX7 triode, taken from the Philips data sheet.

To design such a circuit we will need some data about the valve we are going to use, and for this we consult a published data sheet, many of which can be downloaded from the internet. The most useful piece of information appears in the form of a graph known as the **static anode characteristics** (fig. 1.2), and this can be used to show how the valve will operate in a circuit. In this example we will use the venerable ECC83 / 12AX7, a high-gain valve used in practically every guitar amp ever made.

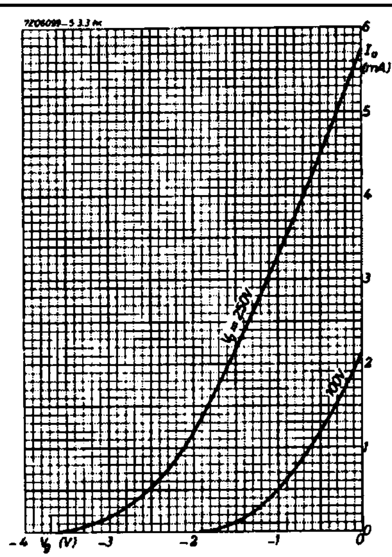

Fig. 1.3: Transfer characteristics graph of the ECC83 / 12AX7, taken from the Philips data sheet.

The static anode characteristics in fig. 1.2 show anode current, **Ia**, on the y-axis: this is simply the current flowing from anode to cathode at any time. The x-axis shows anode voltage, **Va**: this is the voltage dropped across the valve between anode and cathode. The set of curved lines shows the grid-to-cathode voltage, **Vgk**, and are known as the **grid curves**. Note that all voltages on the data sheet and graphs are measured *relative to the cathode*, so the figures given assume the cathode is at zero volts, which it usually is, more or less. Also note that the valve is intended to be operated at relatively high voltages but low currents, quite the opposite of transistors.

As well as the static anode characteristics there is also the **transfer characteristics** graph, also known as the **dynamic characteristics** or **mutual characteristics**, which shows bias voltage against anode current for different values of anode voltage, and this is good indication of the linearity of a valve. The straighter the line, the more linear the valve, and fig. 1.3 indicates that the ECC83 is remarkably linear except at very low anode currents. Although both graphs show exactly the same information, in different forms, the static anode characteristics graph is usually the most convenient to use.

A typical amplifier stage like this has a resistor connected between the HT and the anode. This is the **anode resistor** (US: **plate resistor**), Ra, and forms the load. For small-signal valves the value of the anode resistor is usually in the region of 100kΩ, although there is considerable room for variation- this is dealt with in detail later.
We also need to know what the supply voltage, or **HT*** will be, and we will usually have a rough idea based on whatever power supply we intend to use. For now we will assume the HT is 300Vdc, which is fairly typical. We need to know how the valve will operate under its 100kΩ load, and we can show this by drawing a load line.

The load line:

The anode resistor is in series with the valve. By observing Ohm's law, if no current flows through the valve then there can be no voltage dropped across the resistor, so the anode voltage must be at the same potential as the HT. We can therefore mark this point on the static anode characteristics graph in fig. 1.4, at Ia = 0mA, Va = 300V (labelled 'A').
Conversely, if enough current were to flow, we could drop *all* the available voltage across the resistor and none across the valve. Ohm's law shows that the current required to achieve this is:

$V / R = I$

$300 / 100k = 3mA$

This is the maximum possible current that could ever flow through the valve (if it were a short circuit), unless we either increased the HT or reduced the anode resistor.
We can now plot this point too at Ia = 3mA, Va = 0V (labelled 'B').
Because Ohm's law is a linear equation we can simply join these two points up with a straight line. This is the **load line**, and it is the most powerful piece of analysis which can be performed on a valve circuit.

* HT is a historical term and stands for 'high tension', which may be taken to mean 'high voltage'. In America the notation 'B+' is often used.

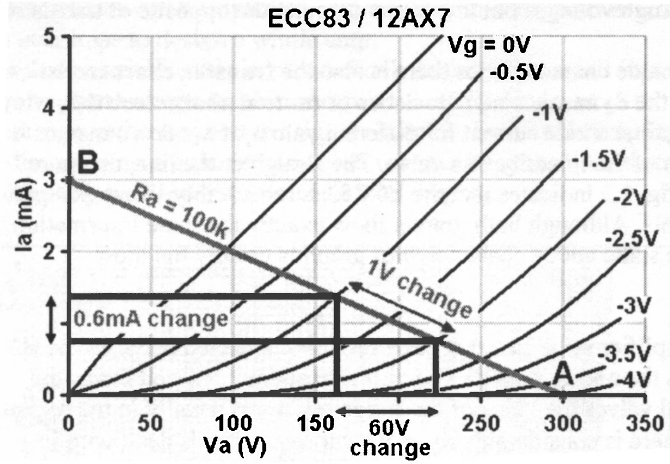

Fig. 1.4: A 100kΩ load line constructed on the ECC83 anode characteristics graph for a HT voltage of 300V. A small change in Vgk causes a change in Ia, which causes a large change in Va.

If we examine the load line we can see that it is intersected at various points by the grid curves. Each point shows us what Va and Ia will be for any given value of Vgk. Remember, the graph assumes the cathode is at zero volts, so the grid curves have been conveniently labelled with the relative grid voltage. Of course, in reality there is an infinite number of grid curves in between those shown, but drawing them all in would make the graph impossible to read!

For example; if we made the grid 2 volts negative we can deduce from the graph that 0.8mA would flow and the anode voltage would be 224V. If we then raised the grid voltage to 1 volt negative, current would increase to 1.4mA and the anode voltage would be pulled down to 164V. This is how the valve is able to amplify; we have only changed the grid voltage by 2 – 1 = 1V, but the anode voltage has changed by 224 – 164 = 60V. Clearly the grid voltage has a large degree of control over the anode current, which is why it is known as the **control grid**. Also note that the valve is controlled by the grid-to-cathode *voltage*. Valves are therefore *voltage controlled devices*, and this makes them ideal for use with electric guitar, since a guitar pickup can produce a voltage signal easily but has practically no current capability.

Biasing and distortion:

Now we have drawn the load line we must make some practical use of it. So far we have considered only the DC supply voltage, but we want to amplify AC signals.

To begin with, imagine a perfect sine wave. The wave has high peaks and low troughs. In between there must be a median point, so the wave could be said to be both rising above, and falling below, this median point. Now to speak in terms of voltage we have a sine wave that swings both positive and negative *relative* to some median voltage. This median voltage is, in a manner of speaking, our 'starting point', and it is up to us to decide where it should be on the load line in order that the valve can amplify both up-going and down-going parts of the signal, and this is known as

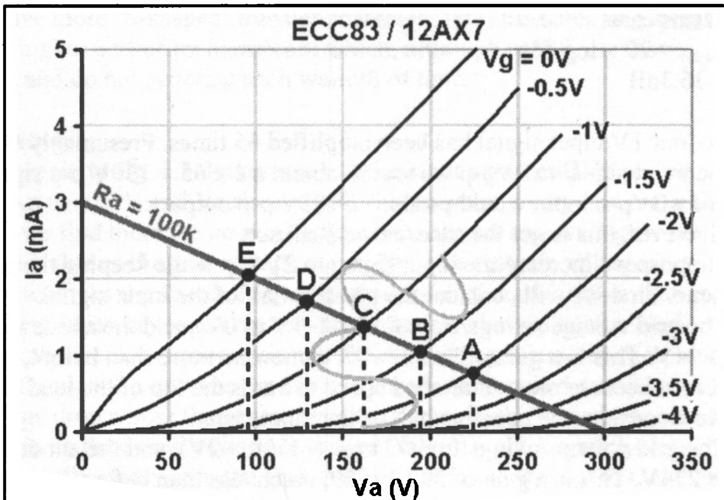

Fig. 1.5: A 100kΩ load line showing changes in Va for changes in Vgk about bias point C.

biasing the valve. For example; we could fix the grid voltage at –1V, and this is now our **bias point**, labelled 'C' in fig. 1.5. We are not yet inputting any signal, the valve is simply at rest. The valve is said to be in a state of **quiescence**, and we can see from the graph that our quiescent bias voltage of –1V causes a quiescent anode current of 1.4mA, and a quiescent anode voltage of 164V.

We will now input a 1V peak-to-peak sine wave (1Vp-p). At first the grid voltage swings more positive to D (–0.5V) and causes Va to fall to 130V. Then the grid swings more negative to B (–1.5V) causing Va to rise to 195V. You can perhaps imagine the input signal streaming down the –1V grid curve and hitting the load line, then being refracted vertically down to the abscissa,* as shown.

Positive input signals produce negative output signals, so our output signal is 'upside-down', or *180° out of phase* with respect to the input; the stage is said to be an **inverting** gain stage.

Furthermore, our 1Vp-p input signal has produced a 195 – 130 = 65Vp-p output signal between anode and cathode, so we can calculate the voltage gain of the stage:

$$A = \frac{V_{out}}{V_{in}} = \frac{-65}{1}$$

$$= -65$$

The minus sign simply indicates that the output is inverted, and is often omitted from equations. Although the human ear cannot detect changes of phase, for guitar we will be deliberately manipulating the shape of waveform, so it is important to remember which stages of the amplifier are inverting or non-inverting.

The voltage gain can also be expressed in decibels according to:

$$A_{(dB)} = 20 \log \frac{V_{out}}{V_{in}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{I}$$

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* The abscissa is the horizontal or x-axis of the graph.

In this case:
$A_{(db)} = 20 \times \log 65$
$= 36.3dB$

So, our 1V input signal has been amplified 65 times. Presumably then, if we increase the input signal to 2Vp-p we would obtain a 2 x 65 = 130V p-p signal at the output, and a 3Vp-p input would produce a 195V p-p output.
However, this is not the case, as we shall see.
Suppose we increase the input signal to 2V p-p while keeping the bias point the same. First we will consider the positive half of the input signal:
The grid voltage swings from C to E (–1V to 0V), and the anode swings from 165V to 94V. This is a gain of 71 / 1 = 71, somewhat more than before, because the grid curves become more 'stretched apart' towards the top of the load line.
Now consider the negative half of the input signal:
The grid voltage swings from C to A (–1V to –2V), and the anode swings from 165V to 224V. This is a gain of 59 / 1 = 59, much less than before, because the grid curves become 'bunched up' towards the bottom of the load line.
Thus the output signal is slightly 'squashed' or compressed on the up-going cycle. This natural compression is part of what makes triode amplification so pleasing to the ear. It produces mainly second-order harmonic distortion, that is, it introduces a new frequency which is exactly 2x the fundamental, or one octave higher than our input signal.. This colours the sound in a pleasant way, making it 'fuller' or 'warmer', the classic valve sound.

This can be also expressed numerically; the anode voltage has swung 71V more negative, but only 59V more positive of its quiescent value. The percentage 2nd harmonic distortion is given by[1]:

$$H2\% = \frac{CD - CA}{2(CD + CA)} x100 \quad \text{..........} \quad II$$

$$H2\% = \frac{71 - 59}{2 \times (71 + 59)} x100$$

= 4.6% 2nd harmonic distortion.

Using the same method on our earlier example of a 1Vp-p input signal yields a value of 3.1% 2nd harmonic distortion.
This indicates that distortion increases as the input signal is increased, since the valve is forced to operate over a wider portion of its transfer characteristic, where it becomes more nonlinear. This can be seen by the curvature near the bottom of the transfer-characteristics graph in fig. 1.3. Ordinary transistors (BJTs) are less linear

[1] Langford-Smith, F. (1957) *Radio Designer's Handbook*, p491. Iliffe and Sons ltd., London. This formula is an approximation as it assumes that *only* 2nd harmonic distortion is produced. In reality other distortion products will also be present, though to much lesser degrees.

than triodes and have more 'S-shaped' transfer characteristic. This tends to generate mainly 3rd, 5th and higher odd-order harmonics which, although useful, are less musically pleasing and do not generate such warmth of tone.

Grid-current limiting:

Referring again to fig. 1.5, let us increase the input signal to 3Vp-p. The grid can swing down to –2.5V and Va rises to 250V. But, when we try and make the grid more positive we find there are no more grid curves shown beyond Vgk = 0V. What happens here is that as the grid voltage approaches the cathode voltage, electrons being drawn from the space charge get attracted to the grid, rather than to the anode (since the grid is very much closer), rather as if we have placed a diode between grid and cathode. Conventional current flows in the opposite direction –into the grid and down through the cathode– and this is known as **forward grid current**. This causes a voltage drop across the source resistance, that is, the output impedance of the driving circuit, so the actual voltage *at the grid* falls. To put it crudely, less of our input voltage actually makes it to the grid. And the more we attempt to make the grid positive the more current flows into it to prevent us from doing so, as if some invisible volume control is being suddenly turned down. In technical terms, the input impedance of the valve quickly falls from many meg-ohms to a few kilo-ohms or less, and this effect is known as **grid-current limiting***, and it causes the output signal to appear 'chopped off' or **clipped** on the negative side. It must be fully understood though, that it is actually the *grid signal* which becomes clipped; the valve continues to amplify what appears on its grid perfectly normally.

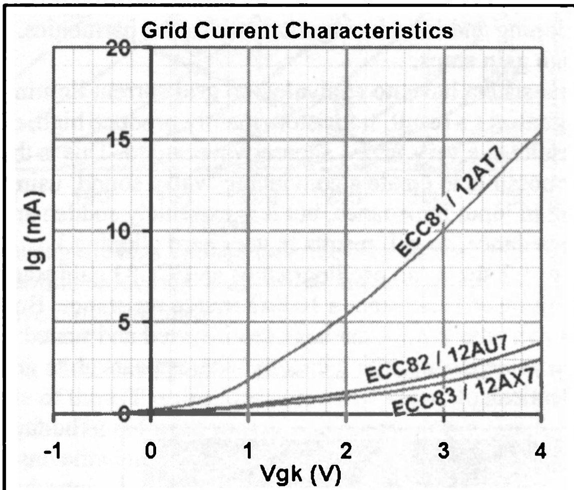


Fig 1.6: Average, forward grid current measured in three valve types operating under a 300V HT and 47kΩ anode loads. Note the much higher level and earlier onset in the ECC81, liable to cause harder clipping.

This effect does not happen instantly. Forward grid current actually begins to flow around Vgk = -1V, though this varies between

* Some modern texts may refer to this incorrectly as 'saturation', because it resembles the clipping effect produced in a saturating transistor. However, in valves 'saturation' refers to the absolute maximum current which can be emitted by the cathode, which should be well beyond the current levels ever found in a normal circuit. Truly saturating the cathode is very damaging to the cathode surface, and should never occur in an audio amplifier.

valve types. The data sheet may even quote this as '$Vgk_{(max)}$', indicating the point at which grid current will begin to exceed 0.3μA, and for the ECC83 this is -0.9V. Additionally, the level of grid current increases inversely with anode voltage, since the anode has less ability to draw electrons away from the grid as its voltage is lowered. This has an important bearing on how the tone of the distortion varies with anode load, and is discussed later.
Fig. 1.6 shows how grid current increases in three common triode types, as the grid is made more positive. Because this effect is not instantaneous the clipped signal will still have 'rounded edges' showing where the input impedance rapidly falls off, introducing plenty of 2nd and 4th order harmonics, while the a 'flat top' to the clipped wave indicates the introduction of some 3rd and 5th order harmonics, adding 'bite' to the sound. Higher harmonics will also be present, with diminishing amplitude.

The rate at which the signal is clipped also depends on the source resistance, since the input voltage is dropped across this. If the source resistance is very low (e.g., a few hundred ohms or less) it is possible to drive the grid somewhat positive, resulting in quite soft compression and very subtle clipping, although this mode of operation is not easily achieved in practice. If the source resistance is very high (e.g., hundreds of kilo-ohms) then the voltage will drop more suddenly, causing harder clipping and introducing more odd-order harmonics, and this is a common feature of high gain amps.
Transistors have no equivalent to grid current limiting and only 'hard clip' the signal. As a result, transistors mainly produce high, odd order harmonics, which results in a very fuzzy, square-wave sound. This is the main reason why it is impossible to create a convincing 'valve sound' using transistors. (JFETs do show a fall in input impedance, but it is extremely sudden and is unaffected by source impedance, so still results in very hard clipping.)
Fig. 1.7 shows an oscillogram of an ECC83 gain stage being driven into grid-current limiting at 1kHz from a 100kΩ source resistance. Both grid and anode signals are shown –the grid signal has been inverted and scaled up for comparison– but because the valve is amplifying exactly what appears at its grid, both signals are nearly identical.

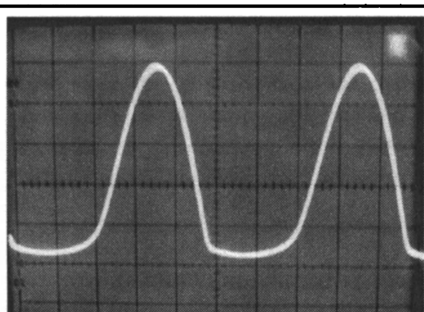


Fig. 1.7: Soft clipping due to grid-current limiting in an ECC83.

The fairly sharp, leading edge indicates the onset of grid conduction, and the introduction of a few high order harmonics, and this can be manipulated by raising or reducing the source resistance or anode voltage. As the input signal begins to fall again, grid current reduces and the valve exits clipping rather less suddenly, as indicated by the softer trailing edge. This asymmetry in the leading and trailing edges of grid-current clipping is common to all valves, though its causes are unclear.

Cut-off:

So far we have only varied the input signal amplitude. Let us now examine what happens if we biased the valve differently. Clearly, if we move the bias point further to the left, reducing bias, we would be closer to grid current limiting. In fact, we could apply no bias voltage at all, in which case the bias point would be exactly on the Vgk = 0V curve and the valve would become a half-wave rectifier; negative input voltages would be amplified while positive ones would be almost completely clipped off.

Instead, let us increase the bias to Vgk = -2.5V (point B in fig. 1.8) so the new quiescent anode voltage is 250V. We will continue to input a 3V p-p signal as before. It can be seen that the anode can now swing down to C, 165V, but cannot swing above A, the HT. Again the output signal will appear clipped, but this time on the positive side. Here, the input signal makes the grid so negative that electrons within the space charge are completely repelled by the grid, and cannot pass through. The flow of current is said to have **cut-off**, and the whole of the HT is dropped across the valve. In this condition it really *is* the output signal that is forced to clip, and not the input signal. The actual grid voltage can continue to swing more negative, but the valve will remain in cut-off.

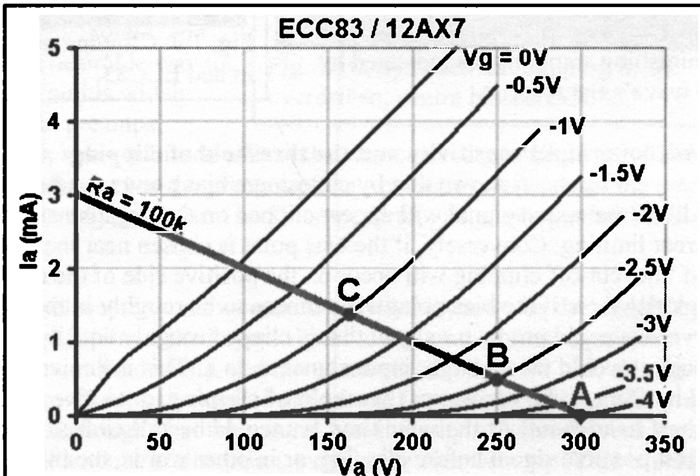


Fig. 1.8: Increasing the bias voltage increases the quiescent anode voltage while reducing the quiescent anode current, placing the valve closer to cut-off.

The grid curves become very bunched up near to cut-off, indicating a reduction in gain and compression of the signal peak. Although the graph suggests that cut-off ought to occur around Vgk = -4V, it will usually take a little more than this. This is because electrons can still pass by the extremities of the grid where the electrostatic field has less influence even when, in theory, current should have already ceased to flow. This is known as the **island effect**, and, as a result, it is in fact quite difficult to completely reduce the current flowing in the valve to zero. Because cut-off is delayed in this way, the onset of clipping is often softer than that due to grid-current limiting, although the designer cannot use source resistance to control it.

Fig. 1.9 shows an oscillogram of an ECC83 gain stage being driven beyond cut-off by a 1kHz input signal. The input signal trace has been inverted and scaled up for easy comparison with the output signal. As the valve approaches cut-off, gain begins to fall, as indicated by the rounding of the leading edge. The valve exits cut-off just as gracefully with gain rising again, so that mainly 2nd and 3rd harmonics are introduced, with many higher harmonics diminishing smoothly, as indicated by the wave's flat top.

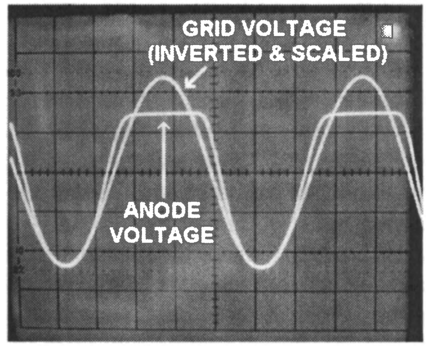


Fig. 1.9: Clipping produced by cut-off in and ECC83.

Headroom: input sensitivity and the threshold of clipping:

It has been shown that by choosing a bias point near the left-hand end of the load line the output signal will appear clipped on the negative side, due to grid-current limiting. Conversely, if the bias point is chosen near the right-hand end of the load line, cut-off clipping will occur on the positive side of the output signal. Logically then, if the bias point were chosen to be roughly in the middle of the grid curves we could produce a signal that is clipped roughly equally on both sides, though it would take a larger input signal to do it. This is known as **centre biasing**, and it offers us the maximum **threshold of clipping**, or to give it its informal term: highest **headroom**. At the same time, we would be able to feed the grid with the largest possible signal before clipping, or in other words, the **input sensitivity** of the stage is at its lowest (input sensitivity is simply the input voltage required to drive the stage to the point of first clipping). Biasing at any other point on the load line will reduce the maximum size of input signal before clipping is reached, reducing headroom and making the stage more sensitive. Increasing the HT will assist in increasing the headroom by shifting the load line to the right, although whether or not a stage actually clips depends on the amplitude of the input signal, of course. Therefore a 'clean' amp does not require a high voltage HT -we can simply attenuate the signal reaching the grid instead- though higher voltages can help.

By manipulating the bias point we can arrange for the signal to clip asymmetrically (on one side more than the other), or symmetrically (on both sides simultaneously). Asymmetric clipping tends to introduce even-order harmonics (particularly 2nd), although as the clipping becomes harder it also introduces odd-order harmonics. Symmetrical clipping produces mainly odd order harmonics and a more traditional 'driven' tone, although with valves some even order harmonics will always be present. Fig.1.10 shows an oscillogram of a centre-biased ECC83 being overdriven. Although the trace is beginning to approach a square wave, the rounded edges ensure that very high, brittle-sounding harmonics are absent, the beginnings of classic rock distortion.

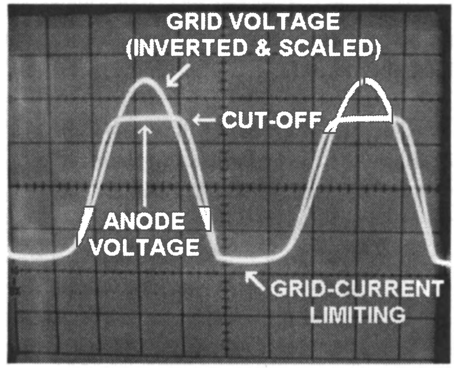


Fig. 1.10: Symmetrical clipping in an overdriven, centre biased ECC83.

When the bias voltage is small it causes more anode current to flow, making the valve run physically hot, so is known as **warm biasing** or **hot biasing**, and tends to produce a warm and 'bluesy' sounding distortion, though biasing too hot can tend to a 'raw' or even fuzzy sound. Conversely, biasing closer to cut-off is known as **cool biasing** or **cold biasing**, and tends to produce a 'harder', more 'crunchy' sound, and is generally the more desirable way to bias a preamp triode, and is widely used in very high-gain preamps.
Because the way each valve is biased affects the way each stage will distort, biasing has a large effect –perhaps even the greatest effect– on the overall tone of the amp, whether that be clean and jazzy, warm and bluesy, dirty, gritty, or even hard and fuzzy.

For the very cleanest sound we would usually bias towards the middle of the grid curves since this allows the largest signal to be input without significant distortion (i.e., maximum headroom), and this is usually the case for hifi, electro-acoustic and jazz amps. The input stage of most amps is centre biased, or nearly so. It must be remembered though, that the natural non-linearity of valves means there will *always* be a small amount of distortion, even in the cleanest amp. This is partly what makes a valve amp sound warmer and more textured than a solid state amp, even when a guitarist claims to be using a 'clean sound', there is always *some* distortion of the signal which generates new harmonics and greater texture.
Very cold biasing, where the valve is biased at or near cut-off*, is rarely used in classic guitar amps, but has become common in modern high-gain lead preamps. The Soldano *SLO100* is the classic example, using a 39kΩ cathode resistor on the third ECC83 gain stage. This causes the stage to go into cut-off long before grid-current limiting is reached, and adds some rich harmonic sustain, for a typical heavy-metal tone. Since cold biasing also reduces the gain of a stage it helps reduce the cumulative gain in the preamp from becoming unmanageable and too sensitive to parasitic oscillation, and also offers some resistance to **blocking distortion** (see chapter 2). A single, very cold biased stage is the natural tonic after a series of warm biased stages, to prevent the tone becoming too raw.
Most moderate to high-gain preamp stages are either centre biased, or bias each inverting gain stage in the opposite fashion to the preceding stage, i.e., hot, cool, hot, cool. The former will tend to develop a fairly even spectrum of odd and even harmonics for a 'classic', open sound. The latter will preferentially clip one side of the signal, developing a strong even harmonic content for rich and 'creamy'

* Older texts may refer to this as 'anode bend rectification'.

overdriven tones: a more contemporary sound. Of course, all of these conventions can be broken; ultimately the best combination of biasing can only be found by experimentation.

The attentive reader may ask then; if biasing has so much control over tone, why not make the bias of each stage variable by means of control knobs on the front of the amp? Indeed, it would be quite easy to achieve this, yet we never see amps with this option. Four possible reasons for this are:

- Nearly all preamp valve use cathode biasing (see later), and making this variable would require a pot with direct current flowing through it, which would produce an irritating, scratchy sound when turning and would also alter the frequency response of the stage at the same time.
- There is always the matter of cost; potentiometers are more expensive than simple resistors.
- There usually turns out to be only a very limited range of biasing over which the amp gives a favourable performance. Once the optimum combination has been set the amp should always sound at its best, which makes a panel full of controls rather redundant.
- Many guitarists take little care over setting the amp's controls, and are liable to simply turn every knob to maximum, which might not result in the best biasing arrangement, for which they would no doubt blame the amp and not their attitude!.

The experimenter may wish to try anyway, but will probably find that these are very good reasons why manufacturers don't offer variable preamp biasing.
It is true that the biasing of the power valves *is* sometimes user variable, but this is really to allow the user to set the amp for maximum, *safe* power output for a given set of valves, rather than for any tonal reason.

Limitations on bias:

There are some limitations on how we may bias the valve that have not been mentioned yet. Firstly, there is a maximum level of continuous power that can be dissipated by the anode, and this is published on the data sheet as $\mathbf{Pa}_{(max)}$ or $\mathbf{Wa}_{(max)}$. This is mainly governed by the physical mass of the anode and the valve's ability to radiate heat. If this maximum level is exceeded for too long the anode will glow deep red, known as **red plating,** and eventually melt, or cause the glass envelope to melt, obviously resulting in permanent destruction. If the maximum anode-dissipation curve is not already shown on the static anode characteristics graph then it may be added by hand. This is done simply by calculating the maximum allowable anode current at a selection of anode voltages. For example, the ECC83 data sheet quotes a maximum of 1W. At anode voltages of 500V, 400V 300V and 200V, the maximum allowable currents are respectively:

P / V = I
1 / 500 = 2.0mA
1 / 400 = 2.5mA
1 / 300 = 3.3mA
1 / 200 = 5.0mA

We can now draw a parabolic curve through these points, as shown in fig. 1.11. The area above the curve indicates the area in which the anode will exceed 1W dissipation (shaded grey), while the area below the curve is the safe operating region. Therefore, the bias point MUST NOT lie above this curve!

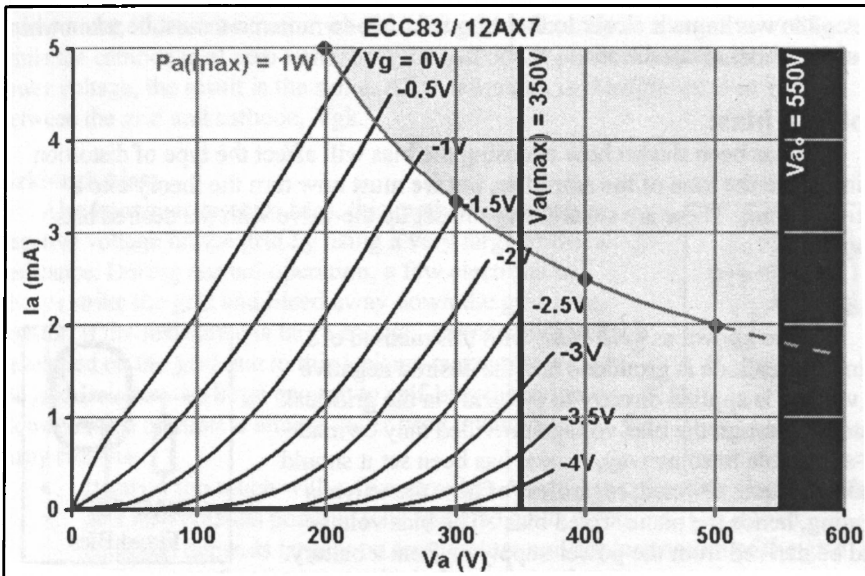


Fig. 1.11: Plotting a maximum anode dissipation curve and safe-operating area.

In the case of power valves a portion of the load line may lie above the curve, indicating that while amplifying a signal the *instantaneous* power exceeds the maximum for some of the time. This is allowable because the valve will spend more than half the time dissipating much less than the maximum, so the average heat dissipated will be less than the maximum allowable, provided the *bias point* is somewhere below the curve. However, we are unlikely to ever treat preamp valves so roughly.

Secondly, there is a maximum voltage which the anode can withstand at any time, before it will arc to other electrodes within the envelope. This is given on the datasheet as $\mathbf{Va_0}$, and for the ECC83 it is 550V. The anode voltage must NEVER exceed this value, so this is usually taken as the maximum allowable HT. This area has been shaded black in fig.1.11.

Thirdly, there is a maximum allowable *quiescent* anode voltage given on the data sheet as $\mathbf{Va_{(max)}}$, and for the ECC83 this is 350V. Therefore our bias point must MUST NOT lie to the right of Va = 350V, and this area has also been shaded grey in fig.1.11. However, while the valve is amplifying the *instantaneous* anode voltage is allowed to exceed this, provided it does not exceed Va_0 at any time.

Some valves also have a maximum allowable cathode current rating, though this usually applies to power valves.

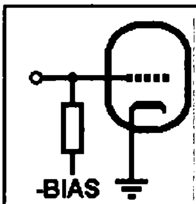
We have now fully defined the area in which it is safe to bias the valve, and in this instance the load line doesn't even fall outside any of the boundaries. Power output stages often work much closer to these boundaries, so more care must be taken when choosing operating conditions.

Applying bias:

It has been shown how choosing the bias will affect the type of distortion and influence the tone of the amplifier, but we must now turn the theory into a practical circuit. There are several ways to set up the valve with the desired bias voltage:

Fixed bias:

Also known as **grid bias**, with this method of biasing the cathode is grounded and the desired negative bias voltage is applied directly to the grid via the grid-leak resistor. Although the bias voltage provided may be made user-adjustable in some way, once it has been set it should remain constant, or fixed, regardless of how the valve is operating, hence the name 'fixed bias'. The bias voltage could be derived from the power supply or from a battery. Deriving a negative voltage from the power supply is fairly straightforward, though it would need to be very low noise indeed if we were to use it for biasing preamp valves, which adds complexity. However, this method is often used on power valves to help maximise output power.

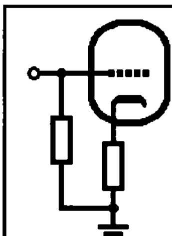


Fixed Bias

Obviously a battery has the limitation that it will eventually need replacing, although we could reasonably expect several years service from a lithium battery. Bias batteries are still sometimes used in hifi, although we will not find them in guitar amps because of the advantages of using cathode bias [see below].

Cathode bias:

Also known as **self bias** or **automatic bias**, this method of biasing involves placing a resistance in series with the cathode. Because a steady anode current flows through the valve, a voltage will be developed across this resistance, placing the cathode at a higher potential than the grid; the valve provides its own bias voltage. This has the added advantage that it is to some extent self limiting: If the average current through the valve rises, so does the bias voltage, which will oppose the increase in anode current, so there is less chance of the valve going into runaway and overdissipating in the event of some failure.

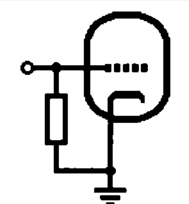

Cathode bias

Further advantages to this method are that it affords the designer some control over gain, linearity, output impedance, and most crucially, the frequency response of the

stage, by using (or not using) a cathode bypass capacitor. This will be dealt with in more detail later.
To the beginner it must be made implicit at this point; *we are only trying to make the grid more negative than the cathode, which is exactly the same as making the cathode more positive than the grid.* The *absolute* measured voltages are unimportant; it does not matter whether the grid is actually at a negative voltage while the cathode is at zero volts, or if the cathode is positive and the grid is at some lower voltage, the result is the same. All that matters is the *difference* in voltage between the grid and cathode, Vgk.

Grid-leak bias:

Also known as **contact bias**, this method develops a negative voltage on the grid by using a very large grid-leak resistance. During normal operation, a few electrons will always strike the grid and bleed away down the grid-leak resistor. If the resistance is large enough, the negative voltage developed on the grid due to this leakage current flowing in the grid-leak may be large enough to self-bias the valve. However, this method is almost never used in audio today, for many reasons:



Grid-leak bias

- It varies too much with valve samples. As well as collecting electrons, the grid also collects positive ions and also *emits* electrons. The rate of emission depends largely on temperature and contamination of the grid during manufacture, over which the designer has no control.
- It requires a very large resistance, which introduces too much unnecessary resistor noise and makes the stage very susceptible to heater hum.
- It is rather unpredictable. Choosing the resistance involves little more than guesswork, and the voltage developed tends to wander as the valve ages. As an approximation, most small signal valves will only develop -0.1V per meg-ohm when new, which severely limits our biasing range; we would need a 10MΩ grid-leak resistor just to obtain -1V bias!.
- The large grid-leak resistance is liable to promote blocking distortion (see chapter 2).
- It does not offer the frequency-shaping advantages of cathode biasing.

Designing a simple, triode gain stage:

By this point the reader should have an understanding of how a valve can amplify signals appearing on its grid, and how biasing, in combination with the load resistance and HT voltage, affects the way the valve distorts the signal. We will now examine in detail the design procedure for a simple common-cathode, cathode biased, triode gain stage. Again the ECC83 will be used, this time assuming a conservative HT voltage of 280V.

Choosing the anode load resistor:

Firstly we should choose an anode resistor. A value of 100kΩ is common, but let us consider what difference other values might make.

Fig 1.12 shows the static anode characteristics of the ECC83 with load lines drawn for three possible load resistances: 220kΩ, 100kΩ and 47kΩ, using the same process given earlier. Clearly, the steeper the load line, the lower the resistance.
For this example a bias point of –1.5V has been chosen, indicated by the dot on each load line.

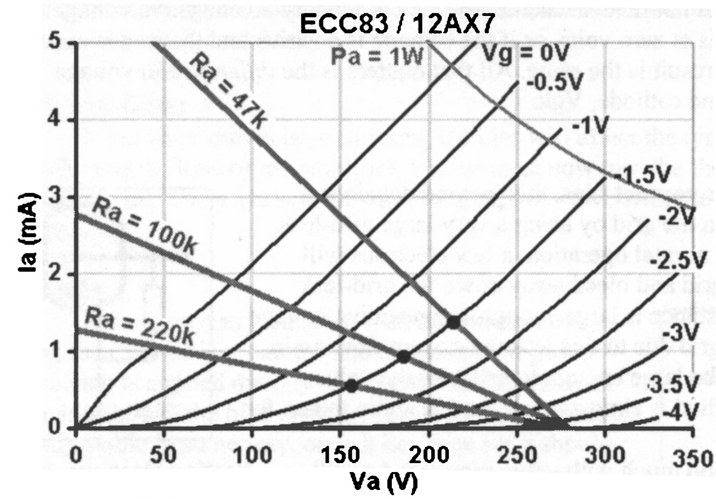


Fig 1.12: Load lines for three different values of anode resistor with an HT of 280V. Each has been given the same bias point of –1.5V.

It can be seen that the larger the load resistance, the larger the possible output voltage swing. With the 47kΩ load the maximum unclipped output signal is roughly 125Vp-p, while for the same bias voltage the 220kΩ load gives roughly 200Vp-p!

In each case the *range* of possible bias voltages is exactly the same (because the HT is the same for each), from which it can be deduced that the voltage gain of the stage must be greater when using larger load resistances. The graph suggests gains of approximately: 42, 55 and 67, for loads of 47kΩ, 100kΩ and 220kΩ respectively. However, if we make the resistance too large we push the load line down towards the abscissa, where anode current becomes very low and where the grid curves become 'bunched up'; in this region performance becomes unpredictable and gain actually begins to fall again. With most preamp valves we are unlikely ever to use anything larger than 470kΩ.
Notice also that the lower the load resistance used the more anode current can flow for a given bias voltage, and therefore more power is dissipated by the anode as heat (this can be seen by the 47kΩ line being closer to the maximum dissipation curve than the others). Sometimes we may require more current output rather than just voltage amplification, in which case we would use a relatively low load resistance, usually with a high current valve such as an ECC82 / 12AU7. With most preamp valves we are unlikely to use anything lower than about 10kΩ, to avoid unnecessary waste of power and shortening of valve life. In this case none of the load lines pass very close to the maximum dissipation curve, so we can be sure of operating well within safe limits whatever we choose.
Finally, it is worth examining how loading affects distortion. Using formula II the following results are obtained assuming a 3Vp-p input signal, and bias of -1.5V:

47kΩ: 10% 2nD harmonic
100kΩ: 7.2% ''
220kΩ: 3.8% ''
This shows clearly that a lower load resistance produces more nonlinear distortion than a higher load resistance. Using a higher load resistance is therefore preferable for hifi, except where the high resistance introduces too much noise. However, for guitar it is important to be aware that although the *total* harmonic distortion reduces as load resistance increases, what little distortion *is* produced contains an increasing proportion of odd harmonics, which can lend a bright, glassy tone, though it can sometimes tend to shrillness.
Furthermore, a larger load resistance gives a lower anode voltage for a given bias voltage. This means that when the valve is driven towards grid-current limiting the ability of the anode potential to draw electrons through the grid is reduced, allowing heavier grid current to flow instead. The result is a more sudden onset of grid-current limiting and harder clipping for a given source impedance, producing still higher harmonic content and a more hard, 'driven' tone.

Therefore, for a warm, compressive, relatively low gain amp –a bass or electro-acoustic amp say– we might use a relatively low load resistance, such as 47kΩ. A high-gain design would tend to use a larger resistance, more than 100kΩ say, for high gain and to generate a typical smooth and glassy lead tone, particularly for heavy metal. For a moderate-gain, rock and roll tone, or for an amp intended for many music styles, we are more likely to use something in between, in the region of 100kΩ.
Occasionally the designer will be forced to use a high value of anode resistor in order to obtain a relatively low quiescent anode voltage at a given bias voltage. This is not often required in guitar amps except in the case of DC coupling, which is discussed at the end of chapter 2. Alternatively, if the stage were being operated from a rather low HT (less than 200V say) we would tend towards a lower value anode resistor to avoid operating at too low an anode current.

Choosing the cathode bias resistor:

For now, we will use the traditional 100kΩ anode load. After drawing a load line we can then choose a bias voltage based on the degree of headroom and clipping we want from the stage, as discussed earlier. If this is the input stage of an amplifier we might choose to centre bias, or go a little warmer. In this case a bias of -1.5V is used. The cathode resistor may now be chosen.
Looking at the load line in fig. 1.13 it can be seen that a bias of –1.5V causes 0.9mA of quiescent anode current. Of course, this current flows right through the valve, out of the cathode and down through the cathode resistor. The grid voltage is fixed at zero, but we want it to be 1.5V more negative than the cathode, so we can place the cathode at +1.5V instead. We now know the desired voltage across, and current through, the cathode resistor so we can apply Ohm's law to find its value:
$R = V / I$
$= 1.5 / 0.0009$
$= 1667\Omega$

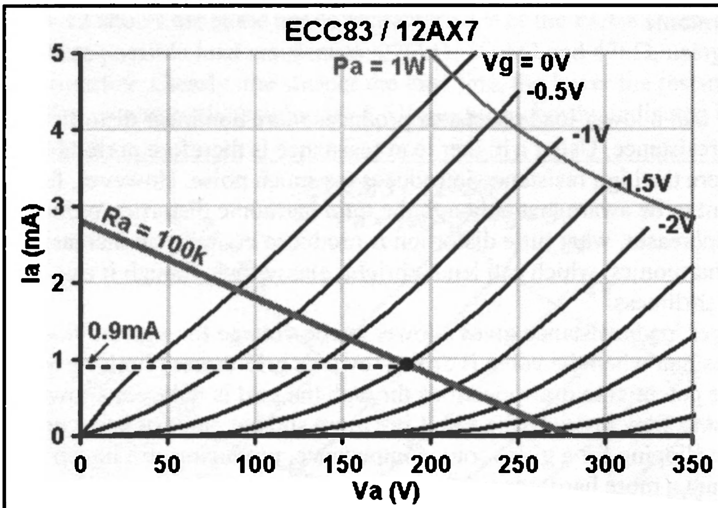


Fig. 1.13: On the 100kΩ load line a bias of −1.5V indicates that 0.9mA of quiescent current flows, and therefore the cathode resistor must be: 1.5V / 0.9mA = 1667Ω

Unfortunately, this is not a standard value. We could try and make this value up using several different resistors or we could accept that a real valve will never precisely match the datasheet anyway, and simply choose the nearest standard value. In this case the nearest standard from the E12 range is 1.5kΩ. This ought to provide a slightly lower (i.e., warmer) bias voltage, which should be fine.

The cathode load line:

It is possible to calculate the expected bias voltage if we already have an actual cathode resistor in mind, this time using a **cathode load line**. A cathode load line is drawn in a similar manner to the ordinary (anode) load line, though it will not be a perfect straight line due to the curvature of the valve characteristics. First let us suppose that 1.5mA flowed through the cathode resistor.

The voltage across it must be:

V = IR

= 0.0015 x 1500

= 2.25V

This point can now be plotted on the characteristics graph in fig. 1.14 at Ia = 1.5mA, Vgk = 2.25V (point A).

Now let us suppose that 0.5mA flowed through it:

V = 0.0005 x 1500

= 0.75V

This point can be plotted too (point B), and a straight line drawn between them; this is the cathode load line. Other points could be taken to extend the line, which would show some curvature near the bottom, and this is shown by the dashed line. Other cathode load lines could also be drawn, corresponding to different possible values o cathode resistor.

Where the cathode load line crosses the main load line indicates the expected bias voltage, and, in this case it is roughly –1.4V. So, although we have chosen a standard value resistor, the result should be very close to our initial design value of –1.5V.

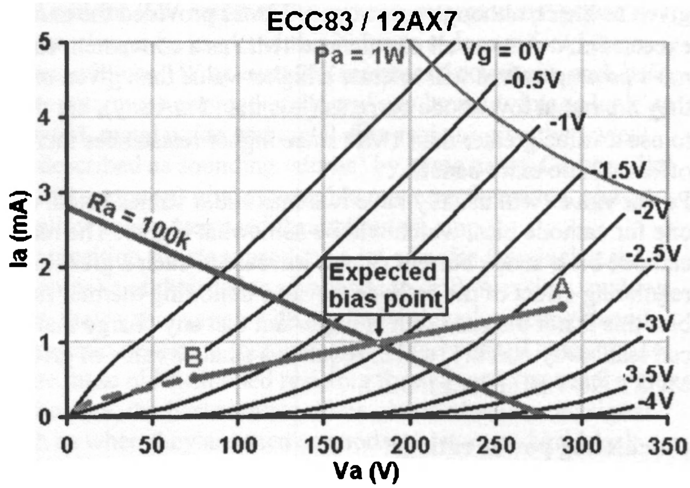


Fig 1.14: A cathode load line can be drawn for a given cathode resistance- in this case 1.5kΩ. Where the cathode load line crosses the anode load line indicates the bias voltage obtained.

Furthermore, during signal conditions the average anode current flowing through the valve will increase somewhat* due to the nonlinearity of the transfer characteristic (the 'stretching apart' of the grid lines for positive input signals), so the bias voltage during operation may actually be closer to our design value than expected!

Usually there is no need to draw a cathode load line as the nearest standard value to the calculated resistance will be quite sufficient. It may be necessary in the case of power valves, however, since the bias voltage is more critical.

Choosing the grid-leak resistor:

During normal operation the grid becomes hot (since it is close to the cathode) and will emit a few electrons, and so will become positively charged unless a leakage path is provided to replenish the charge lost. If the grid is allowed to charge positively, anode current will increase, which increases the valve's temperature, making the problem worse. Eventually the bias would drift out of control and, particularly in the case of power valves, this could easily cause red plating, thermal runnaway and destruction of the valve. The grid-leak resistor provides the necessary leakage path for electrons between grid and cathode, and holds the grid at a fixed quiescent voltage (in this case zero volts). In modern terminology it might be called a 'pull-down' resistor.
Usually we would like the grid leak resistor to be large in value, so as not to load down the preceding stage by shunting AC signals to ground. However, there is a maximum allowable value given on the datasheet. For the ECC83 this is usually

* Older texts may refer to this as the 'self rectification' effect.

given as 2MΩ, although some quote 22MΩ provided the anode current does not exceed 5mA. Almost all circuits use 1MΩ as a convenient standard value, though most preamp valves will tolerate a higher value than given on the datasheet provided they are run at low anode currents (less than 5mA say), but there is usually no need to use a value greater than 1MΩ since higher resistances increase noise while offering little extra benefit

Power valves will usually have two maximum values listed: one for fixed bias and one for cathode bias, which will be somewhat higher. The reason for this is that with cathode bias, anode current cannot increase to such a great extent due to the self-regulating effect of the cathode resistor, inhibiting thermal runnaway. With fixed bias this is not the case so it is important that any charge that does collect on the grid can leak away quickly, which requires a smaller value of grid-leak. These values MUST NOT BE EXCEEDED.

Calculating power ratings:

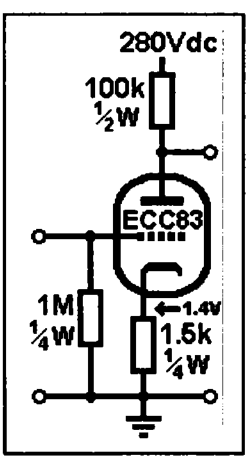
Before the circuit can be built, the necessary power ratings of the components must be calculated. A resistor having insufficient power rating will run very hot, causing it to drift in value or fail completely, sometimes by catching fire! In the case of the anode resistor, we already know from the load line that the quiescent anode current will be close to 1mA. The power dissipated in the resistor must therefore be:

$P = I^2R$

$= 1^2 \times 100$

$= 100\text{mW}$

To ensure a reliable working life it is customary to use a component having at least twice the power handling capacity, so a ¼W resistor might be quite sufficient. However, low wattage resistors usually have a low voltage rating, often 250V(that is, the maximum voltage which may be dropped across the resistor before breakdown, set by the physical construction of the resistor). When fully overdriven the peak voltage across the anode resistor will reach 200V or more, so we ought to use a ½W resistor to be safe.



The cathode resistor has only a small voltage across it, and will dissipate only:

$1^2 \times 1.5 = 1.5\text{mW}$, and even if the valve could somehow pass the maximum current of 2.8mA the power would never exceed 11.8mW, so a ¼W resistor will be quite adequate.

Under normal conditions no DC flows in the grid leak resistor (this is not the case when DC coupling). Only the signal voltage will cause any power dissipation, and even to dissipate ¼W would require an impossible $\sqrt{0.25\text{W} \times 1\text{M}\Omega}$ = 500Vrms signal! Therefore, we need only consider the voltage rating of the resistor. The maximum peak signal voltage in any part of the preamp could be as high as half the HT (140V), so a ¼ W resistor should be sufficient.

Most preamp designs will call for ½W resistors throughout, except in the power supply. Larger wattages can always be used and this helps to reduce noise. Many very high quality amplifiers will use 1W or even 2W resistors throughout, and, being physically larger, these are also more easy to handle and can lend a 'vintage look' to the construction. Wire wound, metal oxide and metal-film resistor give the lowest noise, but are sometimes described as sounding 'sterile' by some users. Carbon-film resistors (which are the most common) give reasonable noise performance, and are reputed to give a more mellow tone when used throughout an amp.
Old-fashioned carbon-composition resistors produce a lot of noise (producing the familiar 'hiss' in vintage amps) and their value also tends to drift over time, making them unreliable. However, their value varies with applied voltage[1], causing them to generate their own 3rd harmonic distortion (compression) in *very* small amounts, so some builders prefer to use these old-fashioned resistors for authenticity and for their supposed tone. If this is the case, the best results will be obtained where the signal voltages are greatest, such as when they are used as anode resistors and grid-leak resistors.

The valve constants:

By now it is clear that a small change in Vgk causes a change in anode current. This is one of the fundamental properties of an amplifying valve and is one of three 'constants', which can be used to deduce the expected performance of a valve in a given circuit. These constants are unique to each valve type and typical values will be given on the data sheet, but they can also be found from the anode characteristics graph.

Transconductance (g_m):

A valve's ability to translate a small voltage change into a current change is known as its **mutual conductance** or **transconductance**. It is given the symbol g_m and is measured in the unit of the **millisiemen**, or milliamps per volt, or **mA/V**. It can be derived from the anode characteristics graph by holding the anode voltage constant (in other words, by drawing a vertical line through the operating point, A on fig 1.15) and reading off the change in Ia for a small change in Vgk. In this case changing Vgk from –1V to –2V (a net change of 1V) causes a change in Ia of $1.7 - 0.3 = 1.4\text{mA}$. The g_m is therefore:

$$g_m = \frac{\Delta Ia}{\Delta Vgk} = \frac{1.4}{1}$$

= 1.4mA/V or 1.4mS*.

[1] Langford-Smith, F. (1957) *Radio Designer's Handbook*, p188. Iliffe and Sons ltd., London

* American data sheets may give the transconductance in 'µmhos'. 'Mho' is simply 'ohm' backwards (the opposite of resistance) and is identical to the Siemen. Therefore, 1400µmhos = 1400µS = 1.4mS or 1.4mA/V.

Amplification Factor (μ):

When a load is placed in series with the valve a small change in Vgk causes a much larger change in Va, and this reaches its maximum value when anode current is held perfectly constant. This is the maximum, theoretical voltage gain of the valve, and is given the symbol **μ** (mu). It can be derived from the anode characteristics graph by drawing a horizontal line through the operating point and reading off the change in Va for a change in Vgk. In this case a change in Vgk from –1V to –2V (a net change of 1V) results in a change in Va of 240 – 140 = 100. This gives a μ of:

$$\mu = \frac{\Delta Va}{\Delta Vgk} = \frac{100}{1} = 100$$

It has no units as it is simply a multiplication factor.

Anode Resistance (r_a):

Finally, if the bias voltage were held constant, any change in Va would cause a change in Ia. This is the resistance that the valve presents to *AC* and is known as its **anode resistance, r_a** (US: **plate resistance, r_p**). It is equal to the reciprocal of the gradient of the grid curve. Therefore, the steeper the grid curves, the lower the anode resistance of the valve. Obviously the cooler the valve is biased the less steep the grid curves are and the higher r_a will be. This is quite different from the valve's DC resistance or **beam resistance**, which is controlled entirely by the position of the operating point at any given time, and is of no use to the designer. To find r_a it is necessary to know the gradient of the grid curve which is, of course, not constant since it is curve! It can be estimated though by drawing a tangent to the

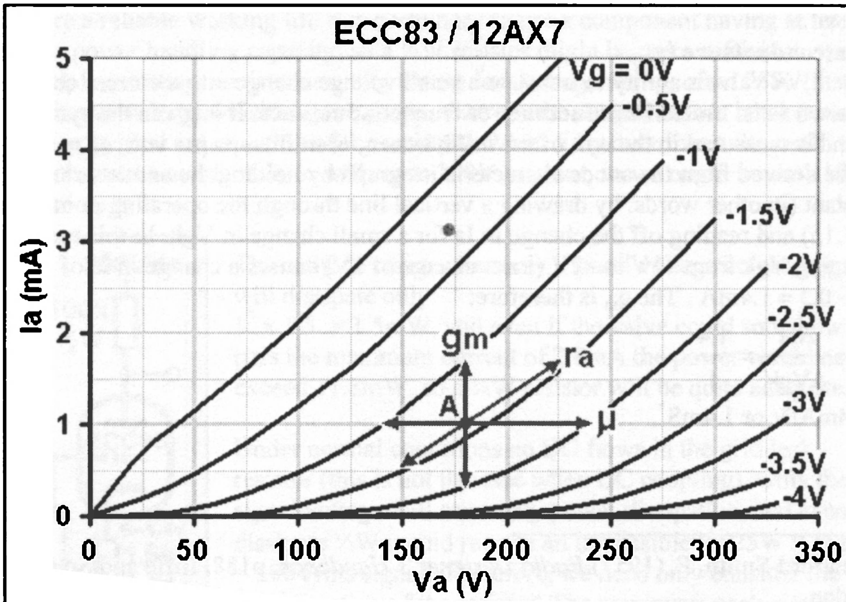


Fig. 1.15: The valve 'constants' can be derived from the operating point, which is chosen by the designer. μ is quite constant, while g_m and r_a are somewhat variable, particularly at low anode currents.

nearest grid curve to the operating point and reading off the change in Ia for a change in Va. In this case a change in Va of 225V - 150V = 75V, causes a change in Ia of 1.65 - 0.6mA = 1.05mA. This yields an anode resistance of:

$$r_a = \frac{\Delta Va}{\Delta Ia} = \frac{75}{1.05} = 71k\Omega$$

The data sheet will provide typical values of r_a, g_m and μ, but because the grid curves are not straight or evenly spaced r_a and g_m are not constant. However, for any 'typical' bias point they will usually be close to the published values, except in some very non-linear triodes such as the ECC85. The conscientious designer will, however, derive his values from the bias point he *actually intends to use*, rather than rely on these data sheet 'generic' values. Published values for the ECC83 are $\mu = 100$, $r_a = 62.5k\Omega$, $g_m = 1.6mA/V$ at an anode voltage of 250V, which are not vastly different from what we have just derived, but nonetheless indicate that our bias point is not quite the same as the data sheet assumes. It is worth noting though, that the μ of the valve *is* quite constant, and remains so throughout the life of the valve.
This process can be made simpler by observing that the valve constants are related by van der Bijl's equation[2]:

$$\mu = r_a.g_m$$

so that only g_m and μ need be found from the graph (these being the easiest to acquire accurate values for), while r_a can be calculated.

Mathematical treatment of a gain stage:

Now that we have a practical circuit and are able to find μ, g_m and r_a from the load line, we can examine its performance more closely. A method of calculating gain using the load line was given earlier, but this is not always convenient. Furthermore, by introducing the cathode resistor we alter the gain, and this is not evident from the load line. It will be convenient to have some equations to speed up the design process (though there is no substitute for the load line when deciding upon a bias voltage and cathode resistor).

The simplest way to analyse a gain stage is to imagine the valve to be a perfect amplifier whose gain is equal to μ. This amplifier forces a signal current to flow down the series combination of Ra, r_a and any other impedance which might be present, which together form a potential divider. If the signal is taken across Ra then the gain of the potential divider, β, is simply Ra / (Ra + r_a) and so the voltage gain of a simple resistance loaded, common cathode, triode gain stage is[3]:

$$A = \frac{-\mu Ra}{Ra + r_a} \quad \text{..........III}$$

[2] van der Bijl, H. J., (1919) Theory and Operating Characteristics of the Thermionic Amplifier. *Proceedings of the Institute of Radio Engineers.* 7 (2) (April). p.109.
[3] Scroggie, M.G.,(1958). *Foundations of Wireless* (7th ed.), p144. Iliffe & Sons Ltd., London.

The minus sign is usually omitted, since it merely indicates that the output is inverted. So if we were to ignore Rk, we might expect the gain of the circuit in fig 1.16 to be:

$$A = \frac{100 \times 100k}{100k + 70k}$$

$= 58.$

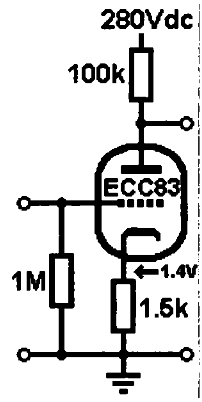


Fig. 1.16. Practical gain stage having a gain of approximately 37 (31dB).

However, we can only ignore Rk if no signal current flows in it, that is, if the cathode voltage is held constant, and in the circuit of fig 1.16 that is not true. When we input a signal to the grid the anode current will be modulated, and since this current also flows through Rk, so the cathode voltage will be modulated also. For example, if the grid rose by 1V anode current would increase and therefore the cathode voltage will also increase by some lesser amount. The *total* increase in Vgk is now not 1V, but some lesser voltage. If the grid is made more negative, the reverse happens; the cathode voltage attempts to follow the grid voltage. Since the valve *only* amplifies changes in Vgk, the output voltage will be less than what we might have expected. This phenomenon is known as **cathode-current feedback** and causes the apparent gain and non-linear distortion of the stage to be reduced, while the output impedance is increased.

Because a valve amplifiers the difference in voltage between grid and cathode, any impedance placed in series with the cathode will also appear to be amplified, by $\mu+1$, when 'looking into' the anode. Therefore the gain of the stage with the cathode resistor included is[4]:

$$A = \frac{\mu Ra}{Ra + r_a + Rk(\mu + 1)} \quad \text{.......................IV}$$

It was shown earlier that in this case μ is 100, so:

$$A = \frac{100 \times 100k}{100k + 70k + (1k \times 101)}$$

$= 36.9$

Obviously this is much lower than the figure we calculated previously. This may be desirable, particularly in a clean amp, since the stage will be less easily overdriven and will itself not overdrive the following stage quite so easily. This sort of stage is

[4] Langford-Smith, F. (1957) *Radio Designer's Handbook*, p485. Iliffe and Sons ltd., London.

not uncommon in hifi designs since the cathode feedback causes a reduction in distortion and a bandwidth that may extend well beyond audible frequencies. Generally though, we would like to achieve as much gain as possible from the stage, so we must prevent the cathode voltage from being varied by the input signal. This introduces a new and important component; the **cathode bypass capacitor**.

The Cathode Bypass Capacitor:

By placing a capacitor in parallel with the cathode bias resistor as in fig. 1.17, any instantaneous rise in cathode current will be diverted into charging the capacitor, and if cathode current falls, the capacitor will supply the deficit from its own charge. Another way of looking at it is to say that the capacitor shunts or 'bypasses' to ground any AC signals on the cathode so that signal current does not flow in the cathode resistor, while the DC bias voltage remains unchanged. With either explanation the result is the same: the cathode bypass capacitor 'smoothes out' changes in cathode voltage, helping to hold the cathode voltage constant, preventing cathode feedback and allowing full gain to be realised. The cathode bypass capacitor also affects the distortion characteristics of the stage. Without it, cathode-current feedback tends to slow down the onset of grid current, giving a smoother and more compressed or creamy sound. Adding the capacitor removes this feedback effect, result in somewhat harder clipping and a more 'brittle' or aggressive overdriven tone.

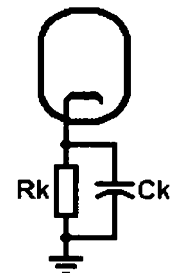


Fig. 1.17: A cathode bypass capacitor, Ck, maximises gain.

Of course, a capacitor will allow greater current flow at high frequencies than it will at low frequencies. If we want the stage to have maximum gain at all audible frequencies then the capacitor must be large enough* to smooth out the lowest frequencies of interest, and the stage could be described as being 'fully bypassed'.

If the capacitor is made relatively small then only high frequencies will be smoothed out while lower frequencies will not. Therefore the stage will have maximum gain at high frequencies and minimum gain at low frequencies, producing a treble boost, and the stage would be termed 'partially bypassed'. To the designer, this is an extremely useful consequence of using cathode bias.

If the stage has no cathode bypass capacitor it may be described as 'unbypassed' and will have minimum gain.

The exact relationship between the gain of the stage with frequency, and the size of the cathode bypass capacitor, is relatively complex. From the formulae given earlier we know what the maximum and minimum possible gain of the stage is, but

* By "large" the author is referring to the *capacitance* of the capacitor, and not its physical size.

calculating the transition from one to the other is less straightforward. The gain of the stage at any frequency is given by[5]:

$$A = \frac{-\mu Ra}{Ra + r_a} \cdot \sqrt{\frac{1 + (2\pi f Rk Ck)^2}{\left[1 + \frac{Rk(\mu + 1)}{Ra + r_a}\right]^2 + (2\pi f Rk Ck)^2}} \quad \text{.......V}$$

But this is somewhat cumbersome expression, clearly. Fortunately, the 'shape' of the frequency response plot of a triode gain stage is always predictable: it is a first-order shelving filter (see fig. 1.19). Therefore we need not calculate the gain at every frequency, but simply define the frequency at which gain is half-way between minimum and maximum, this being the 'half-boost' point. It can be shown that the half-boost frequency is given by[6]:

$$f_{(half\ boost)} = \frac{1}{2\pi Rk Ck} \cdot \sqrt{1 + \frac{Rk(\mu + 1)}{2(Ra + r_a) + \frac{1}{2} Rk(\mu + 1)}} \quad \text{.......VI}$$

Though this may be a little too lengthy for quick calculations. Fortunately, for a warm biased, high-gain stage such as this, if $Ra + r_a$ is greater than $Rk(\mu+1)$, then to a reasonable approximation;

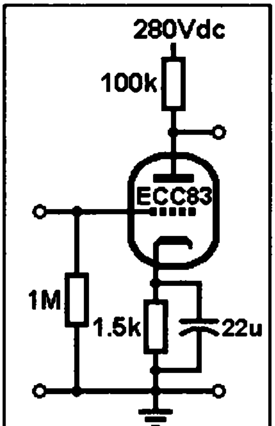


Fig. 1.18: A practical gain stage having maximum gain at all audio frequencies.

$$f_{(half\ boost)} \approx \frac{1}{2\pi Rk Ck}$$

(The true half-boost frequency will usually be a little lower.)
So for full bypassing we would choose a frequency well below the audible range, say 5Hz. Rearranging the above gives:

$$Ck = \frac{1}{2\pi \times 1.5k \times 5}$$

$$= 21\mu F$$

The nearest standard is 22µF. With this value the stage should produce a gain of 58 at all frequencies of interest. The completed schematic is shown in fig. 1.18, and some readers may recognise its similarity to the input stage of the Fender *Bassman*.

Because 22µF is a fairly large value of capacitance we will probably be forced to use an electrolytic capacitor. The cathode operates at only a couple of volts so many

[5] Langford-Smith, F. (1957) *Radio Designer's Handbook*, p484. Iliffe and Sons ltd., London.

[6] Blencowe, M & James, D. I., (2008). 'Choosing Cathode Bypass Capacitors'. *Audio Xpress*, (August), pp. 19-20.

designs use a 25V component. However, it is an unfortunate property of electrolytic capacitors that they have a limited working life, relatively high ESR*, and, when operated at polarising voltages far below their maximum working voltage or at significant AC voltages (more than a few hundred millivolts) they exhibit low frequency distortion[7]. This can often manifest as a 'wooliness' or 'sluggishness' on low notes. A further problem with using a large cathode bypass capacitor is that the stage may have *too* much gain at low frequencies. If the stage begins to overdrive the following stage, too much low frequency content can lead to blocking distortion [see chapter 2], which is *most* undesirable. This is most important in bass amps of course (it will sometimes pass unnoticed in a guitar amp), but in any amplifier using several gain stages the effects will be cumulative, so it is worth considering an alternative. There is a number of solutions:

- Use a lower voltage component, say 6.3V. This may improve bass transient response somewhat by reducing capacitor distortion, but low voltage electrolytics often have extremely poor tolerance and sometimes high ESR, which may or may not spoil a carefully designed stage.
- Use an excessively large value of electrolytic capacitor –100uF say– to cure low frequency distortion. Unfortunately this will promote blocking distortion in following stages and also prevents us from using the cathode for frequency shaping.
- Use a non-polarised electrolytic. These capacitors do not require any polarising voltage and give consistently improved results when compared with ordinary electrolytics, but they may not be available in a wide range of values.
- Use a non-electrolytic capacitor, such as a poly' or silvered-mica type. These give the best sounding results, though are not readily available in values much larger than 1µF, though for guitar this is unlikely to pose a problem.

Most early amplifiers used a fairly large cathode bypass capacitor since they used a limited number of gain stages and in the early years it was assumed that they would not be overdriven (the Marshall 50W Master Volume used 330µF in one of its channels!). As usage evolved and guitar pickups became ever more powerful (a modern high-output humbucker might deliver up to 4V p-p in the neck position!), it became important to maintain a tight and defined bass response even when overdriving, and 'bright' and 'top boost' tones became popular too. As a result, more amplifiers began using smaller values of cathode bypass capacitor. The most popular, well-designed amplifiers of recent times almost exclusively use partial bypassing in combination with small coupling capacitors [see chapter 2], after much experimentation no doubt, to encourage glassy, sustained high-note distortion with

---

* ESR is the capacitor's **Effective Series Resistance**, so-named because it acts exactly like a resistor is series with the capacitor.

[7] Self, D. (2006). *Audio Power Amplifier Design Handbook* (4th ed.), p175, Newnes, Oxford.

un-muddied bass tones. It is remarkable how little low frequency gain is actually required for a good tone in a guitar amp! Reducing over-sized cathode bypass capacitors is a cheap and safe way of experimenting with, and improving the tone and transient response of an existing amp, particularly one that sounds too bassy or muddy. What's more, smaller values of capacitance allow much higher quality, better tolerance, non-electrolytic capacitors to be used. However, where possible the *first* stage of an amplifier should always be fully bypassed so as to maximise rejection of heater and power supply hum.

Fig. 1.19 shows the frequency response of the same stage with different possible values of Ck. The stage has an obvious 'shelving' effect, boosting higher frequencies. Reducing the size of the capacitor shifts the 'shelf' up the frequency scale, without changing its basic form. The dot on each plot indicates the transition frequency for that value of Ck, which can be found using formula VI.

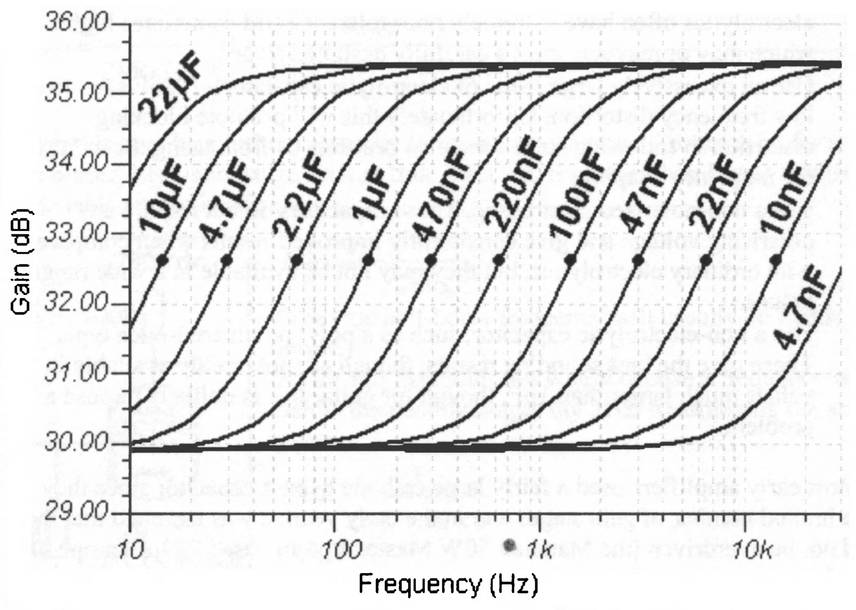


Fig. 1.19: Frequency response of the circuit in fig. 1.18 for different possible values of cathode bypass capacitor.

Other cathode bypass circuits:

Sometimes the designer may wish to use partial bypass, but with a lesser degree of boost. This is easily done by adding a boost-limiting resistor in series with the bypass capacitor, so that even the higher frequencies are never completely shunted to ground via Ck.

For example, fig 1.20a shows a typical amplifier stage where R1 is the boost-limiting resistor. At low frequencies the cathode is completely unbypassed, so the stage will have minimum gain. Using formula IV the gain will be:

$$A = \frac{100 \times 100k}{100k + 62k + (1k \times 101)}$$

$= 38$

(The value of r_a has fallen to 62kΩ since the stage is biased warmer than in fig. 1.18.) At high frequencies the 1μF capacitor bypasses the cathode resistor, but the additional 1kΩ resistor prevents *complete* bypassing. Since R1 is now in parallel with Rk, the total effective resistance in the cathode circuit is 500Ω, and the maximum gain of the stage will be:

$$A = \frac{100 \times 100k}{100k + 62k + (0.5k \times 101)}$$

$= 47$

To AC, both 1kΩ resistors are in series with the capacitor, giving 2kΩ, so to a rough approximation the half-boost frequency will be:

$$f_{(half\ boost)} \approx \frac{1}{2\pi \times 2000 \times 1 \times 10^{-6}}$$

$= 80\text{Hz}$

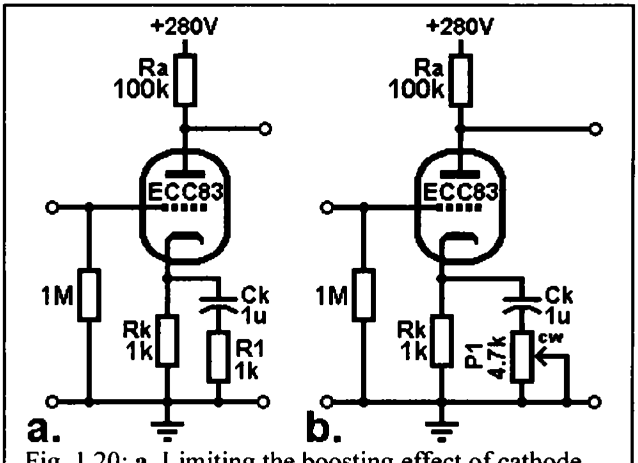


Fig. 1.20: **a.** Limiting the boosting effect of cathode bypassing. **b.** A variable boost or gain control.

An obvious modification to the circuit in fig. 1.20*a* would be to replace the boost-limiting resistor with a potentiometer (wired as a variable resistor) as shown in *b.*, and so create boost that is user variable. If the bypass capacitor is small then it becomes a treble boost or presence control (for which a linear potentiometer would normally be used), while if it is large we have a full-range gain control (for which a logarithmic potentiometer is usual). Because the capacitor will block DC from the potentiometer the control will be free from any obnoxious crackle.

Fig. 1.21 shows the actual frequency response of both circuit variations. The middle trace corresponds to 20% pot' rotation, which also corresponds to the frequency response of the circuit in *a*. Again, reducing the size of the bypass capacitor would shift the 'shelf' up the frequency scale. This clearly shows that although the stage boosts treble, the level of has been deliberately limited by the added series resistance.

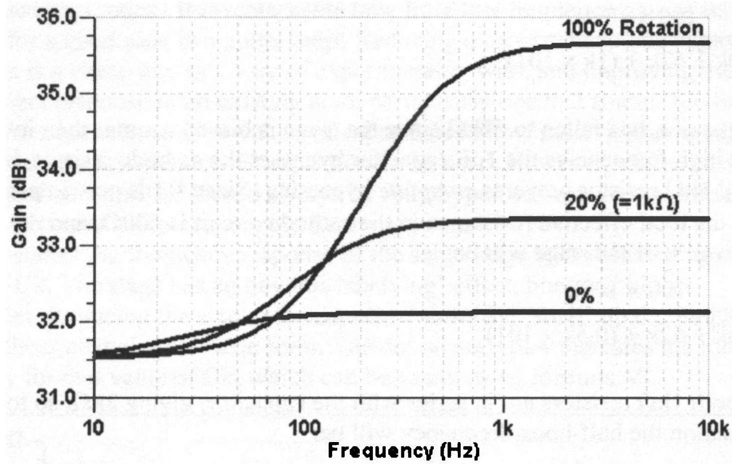


Fig. 1.21. Frequency response of the circuits in fig. 1.20.

Because the full gain of the circuit in 1.20*a* has not yet been realised, we could add a second, smaller capacitor to *completely* bypass the cathode at the highest frequencies. Fig 1.22 shows a 100nF added to the previous circuit, so that at very high frequencies the cathode is fully bypassed.
The transition frequency between partial and full boost is approximately:

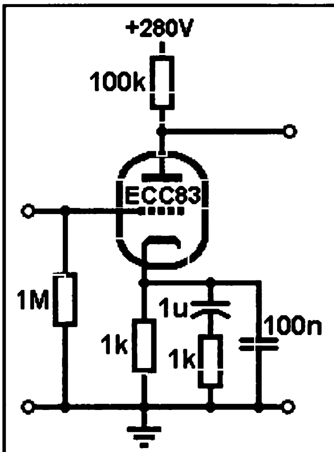


Fig. 1.22. Obtaining a 'stepped' treble boost.

$$f_{(\text{half boost})} \approx \frac{1}{2\pi \times 1000 \times 100\times10^{-9}}$$

$= 1.6\text{kHz}$

Fig 1.23 illustrates how this has modified the previous circuit's response. At very high frequencies the full gain of the stage is now realised, and the stage has a 'stepped' treble boost. For comparison, the dotted line shows the response of the previous circuit.
Even more complex bypassing arrangements could be devised, and many valve televisions used very elaborate cathode-bypass circuits indeed, to filter audio from video signals. However, for the limited bandwidth required for audio, the topology given in fig. 1.22 is probably the practical limit. Again, the boost-limiting resistor could be replaced with a potentiometer.

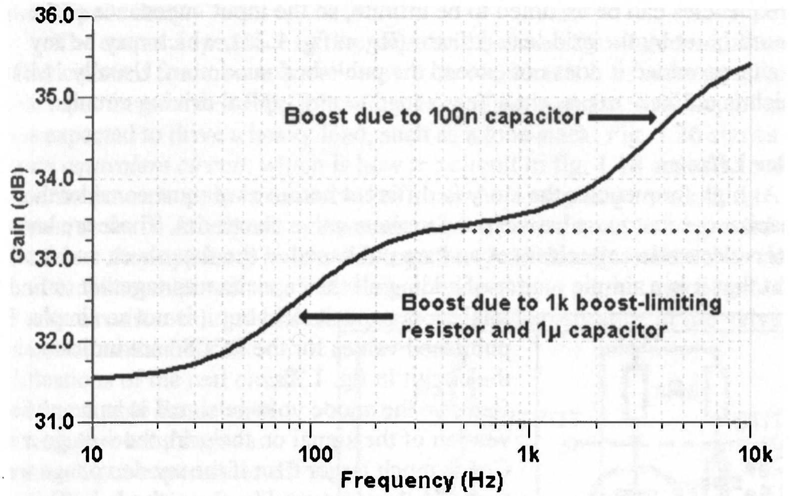


Fig. 1.23. Frequency response of the stepped treble-boost circuit in fig. 1.22

Input and Output Impedance:

Whenever one circuit is connecting –or **coupled**– to another, it is important to know what effect this will have on the circuit being driven, or on the circuit doing the driving. In a preamp we are mainly concerned with *voltage* signals. The output impedance of the driving circuit, when feeding the input impedance of the driven circuit, forms a potential divider as illustrated in fig. 1.24. If the input impedance of the driven circuit is low when compared to the previous stage's output impedance (which, as far as the driven circuit is concerned, is the **source impedance**), it will cause heavy loss of the signal voltage. In other words; less of the desired signal will actually make it to the following stage. For maximum voltage transfer we would like the input impedance to be large with respect to the source impedance (at least five times larger is usual), which is known as **impedance bridging**.

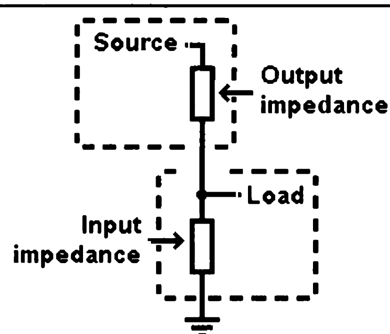


Fig. 1.24: Output or source impedance forms a potential divider with input impedance.

Input Impedance:

Because the grid is suspended in the vacuum inside the valve envelope, it should draw no current under normal circumstances, unless it is overdriven to the point of grid current of course. Therefore, the input impedance of the grid at low and middle frequencies can be assumed to be infinite, so the input impedance of the *whole* circuit is set by the grid-leak resistor (Rg in fig. 1.25), which may be any desired value provided it does not exceed the published maximum. Usually 1MΩ is used, which is unlikely to present a heavy load to any typical driving circuit.

The Miller Effect:

At high frequencies the story is different because we must consider the stray capacitances that exist between the various valve electrodes. These are known as the **inter-electrode capacitances** and are published in the data sheet, and it might be thought that it is a simple matter of adding all the capacitances together to find the total input capacitance, but it is not so simple. The published values for the ECC83 are indicated in the circuit in fig. 1.25.

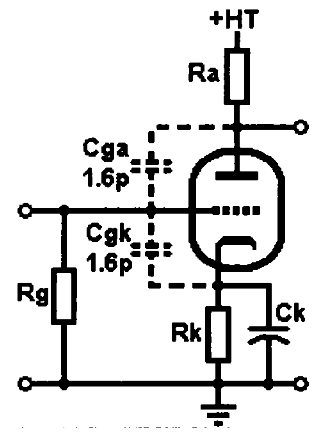


Fig. 1.25: Stray capacitance between electrodes lowers input impedance at high frequencies.

Because the anode voltage signal is an amplified version of the signal on the grid, the voltage across Cga is much larger than if the anode voltage were kept mostly constant, like the cathode is. This makes Cga appear to be multiplied by the gain of the stage, and this is known as the **Miller effect**, so the total input capacitance becomes:

Cin = Cgk + Cga.A VII

Where:

Cgk = grid-to-cathode capacitance*

Cga = grid-to-anode capacitance

A = voltage gain of the stage

In this case the gain of the circuit in fig. 1.18 was calculated to be 58, so Cin will be:

Cin = 1.6 + (1.6 x 58)

= 94.4pF

It is usual to add a few picofarads to this, to allow for additional stray capacitances when the circuit is physically built -particularly those added by the valve socket itself, so we would probably round up to 100pF. This will form a low-pass filter with the source impedance, causing high frequencies to be attenuated due to current having to charge the input capacitances. This effect is actually a useful one in a guitar amps because we can manipulate at what point the frequencies begin to roll off, which is dealt with in the chapter 2. Some high-gain designs even add a small capacitor between the anode and grid of some stages to artificially increase Cga, deliberately reducing the treble response while at the same time linearising it (since the capacitor introduces local negative feedback [see chapter 9]) which can assist is giving a smoother treble sound.

* Some datasheets may not provide Cgk but will instead give the "grid to all except anode" capacitance, which should be used instead.

Output impedance:

The output impedance of this stage will become the source impedance for whatever circuit it feeds, so it is necessary to know what it is in order to select a suitable input impedance of the following stage. This is particularly important if the stage is expected to drive a heavy load, such as a tone stack. Fig. 1.26 shows a *Thévenin equivalent circuit*, which is how the circuit in fig. 1.18 appears to *AC only*. Because the power supply is connected to ground via very large smoothing capacitors, as far as AC signals are concerned, the power supply and ground are one and the same: they appear to be shorted together. Additionally, rather than show a valve with unknown properties, it has been replaced by a perfect signal generator in series with a resistance, r_a, which represents the anode resistance of the valve. Such diagrams are mathematical simplifications of the real circuit and are invaluable for solving AC circuit problems such as input and output impedances. If the cathode resistor is bypassed it is effectively short circuited as far as AC is concerned, and so it is missing from circuit *a*. There it can be seen that the total resistance through which current has travelled to reach the output, is the parallel combination of Ra and r_a, so the output impedance is:

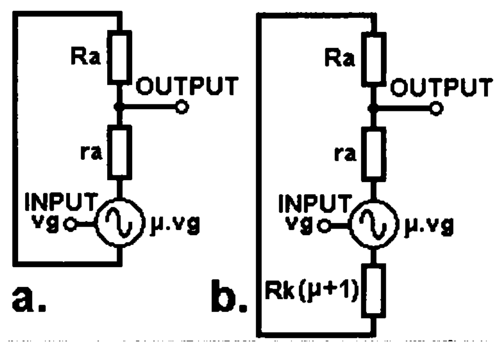


Fig. 1.26: Thévenin equivalent circuits of the circuit in fig. 1.18, at **a.** middle and high frequencies and **b.** low frequencies.

$Z_{out\,(Rk\ bypassed)} = Ra \parallel ra$ VIII

It was calculated earlier that r_a is in this case about 70kΩ, so the output impedance is:

$Zout = 100k \parallel 70k$

$$Zout = \frac{100k \times 70k}{100k + 70k}$$

$= 41k\Omega$

If the cathode resistor is unbypassed, which it will be at very low frequencies, then the circuit appears as in fig. 1.26*b*, and Rk is now in series with ra. Because Rk is below the cathode its value appears multiplied by μ+1. Therefore the output impedance when the cathode is unbypassed becomes:

$Z_{out\,(Rk\ unbypassed)} = Ra \parallel (ra + Rk(\mu+1))$ IX

If Rk is 1.5kΩ and μ is 100 then:

$$Zout_{(Rk\ unbypassed)} = 100k \parallel (70k + 1.5k \times (100 + 1))$$

$$Zout_{(Rk\ unbypassed)} = 100k \parallel 251.5k$$

$$Zout_{(Rk\ unbypassed)} = \frac{100k \times 251.5k}{100k + 251.5k}$$

= 72kΩ

This clearly indicates that an unbypassed cathode resistor considerably increases the output impedance. If the stage is only partially bypassed it will therefore have a higher output impedance at low frequencies than at high frequencies, and this will somewhat increase the degree of bass attenuation caused by the output coupling capacitor (see chapter 2).

Cathode biasing with diodes:

Cathode biasing using a resistor and bypass capacitor is certainly the most common (and traditional) method of biasing a preamp stage. However, there is an alternative method which is well worth mentioning before ending this chapter on basic stage design.

In this arrangement the cathode resistor is replaced with a diode. This may be an ordinary signal diode, a light emitting diode (LED), a zener diode or even a valve diode.

A solid-state diode has a voltage drop that is quite constant once current through it exceeds a certain threshold (usually much less that 100μA). This ensures that the cathode voltage will remain constant regardless of anode current, or even what valve is plugged in, so there is no need for a bypass capacitor. Thus the stage will always achieve full gain without any of the low-frequency sluggishness associated with large electrolytic capacitors, because the diode presents practically no resistance to AC. This is particularly useful for the first stage of an amplifier, where rejection of

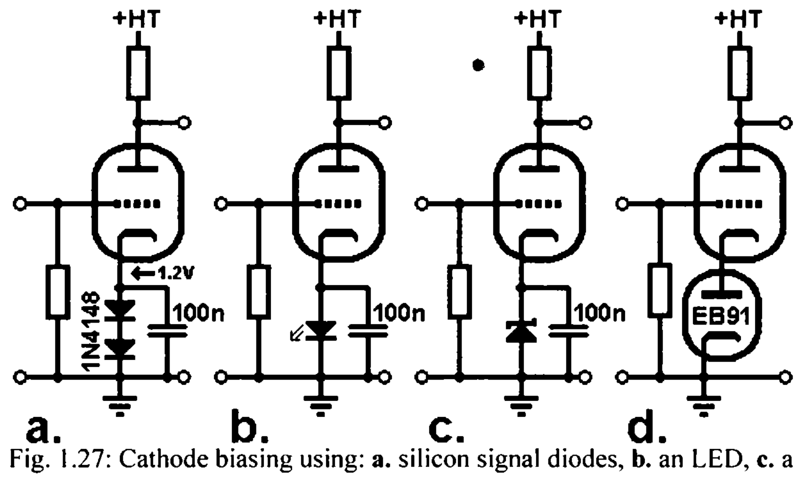


Fig. 1.27: Cathode biasing using: **a.** silicon signal diodes, **b.** an LED, **c.** a zener diode, **d.** a valve diode.

heater and power supply hum is most important. What's more, because no bypass capacitor is used there will be no phase shift in anode current, which may be important if the stage is within a feedback loop (which makes diode biasing very popular in hifi amplifiers). However, if the stage is likely to be overdriven, a 'snubbing' capacitor of around 100nF may be added in parallel with the diodes to reduce any noise created as they switch on and off with the valve. If a zener diode is used then this may also be given a capacitor to shunt any noise voltage it might produce, whether the valve is overdriven or not.
The actual bias voltage developed will depend on the type of diode used, and several diodes can be connected in series to give the desired voltage; some typical arrangements are shown in fig. 1.27. Ordinary rectifier diodes or signal diodes have a voltage drop in the region of 0.6V each, while the drop across LEDs depends on their colour and efficiency. Table 1.1 gives the typical voltage drop across some diodes.

Diode Type	Typical Voltage Drop at 5mA Anode Current
Signal / Rectifier Diode	0.6V
Red LED	1.6V
Bright Red LED	2V
Yellow LED	2.1V
Green LED	2.2V
Blue LED	3V
Red (Super Bright) LED	1.8V
Blue (Super Bright) LED	4.5V
EB91 / 6AL5 Valve	2 to 4V

Table 1.1: Typical diode voltage drop.

An attractive side effect of LED biasing is that the LED will of course light up, and could be mounted on the amp's front panel as a 'power on' indicator, or similar. If the valve is very heavily overdriven then its average anode current will tend to increase, and grid current also flows down through the LED, causing it to glow brighter, so it can serve as a rudimentary 'overdrive-level' indicator.
A valve diode will give slightly less predictable performance since it has a much higher AC resistance (r_a) and its transfer characteristic is not perfectly linear, though it will produce hardly any switching noise. The voltage drop it produces can be quite variable with manufacturer and age, but may be manipulated by altering the heater voltage, so some experimentation will be required to produce the desired results. Fig. 1.27*d* shows an EB91 / 6AL5, which is a double-diode with a 7-pin (B7G) base. It is not in current production, but was so common that it is still readily available as NOS very cheaply.

Summary of formulae:

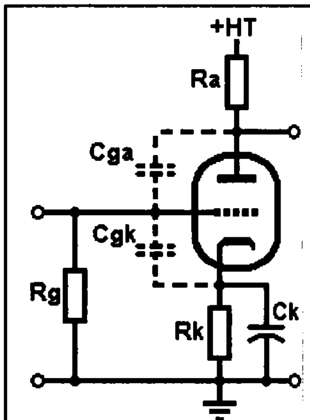


Fig. 1.28

I; Voltage gain in decibels:

$$A_{(dB)} = 20\log\frac{V_{out}}{V_{in}}$$

III; Voltage gain of a resistance loaded triode (cathode resistor fully bypassed):

$$A = \frac{-\mu Ra}{Ra + r_a}$$

IV; Voltage gain of a resistance loaded triode (cathode resistor unbypassed):

$$A = \frac{-\mu Ra}{Ra + r_a + Rk(\mu + 1)}$$

V; Voltage gain at any frequency:

$$A = \frac{-\mu Ra}{Ra + r_a}\cdot\sqrt{\frac{1 + (2\pi f Rk Ck)^2}{\left[1 + \frac{Rk(\mu + 1)}{Ra + r_a}\right]^2 + (2\pi f Rk Ck)^2}}$$

VI; Half-boost frequency due to cathode bypass capacitor:

$$f_{(half\ boost)} = \frac{1}{2\pi Rk Ck}\cdot\sqrt{1 + \frac{Rk(\mu + 1)}{2(Ra + r_a) + \frac{1}{2}Rk(\mu + 1)}}$$

Solving for Ck gives:

$$Ck = \frac{1}{2\pi f Rk}\cdot\sqrt{1 + \frac{Rk(\mu + 1)}{2(Ra + r_a) + \frac{1}{2}Rk(\mu + 1)}}$$

Where:
f = the desired half-boost frequency.

If Ra + ra > Rk(μ+1) then to a reasonable approximation:

$$f_{(half\ boost)} \approx \frac{1}{2\pi Rk Ck}$$

Input impedance (before grid current):

$$Zin = Rg$$

VII; Total input capacitance:

$$Cin = Cgk + Cga.A$$

VIII; Anode output impedance (cathode resistor bypassed):

$$Zout = Ra \parallel r_a = \frac{Ra.r_a}{Ra + r_a}$$

IX; Anode output impedance (cathode resistor unbypassed):

$$Zout = Ra \parallel r_a + Rk(\mu + 1) = \frac{Ra\left[r_a + Rk(\mu + 1)\right]}{Ra + r_a + Rk(\mu + 1)}$$

Where in all cases all notations are as in fig. 1.28 and;
r_a = internal anode resistance of valve at operating point.
μ = amplification factor of valve.
All resistances in ohms. All capacitances in farads.

Chapter 2: Coupling

The amplifier's input. Fundamentals of interstage AC coupling. The AC load line. Blocking distortion and grid stoppers. Practical AC coupling networks. DC coupling.

In the previous chapter it was shown how a triode gain stage is designed, and what tone-shaping possibilities it offered. By itself though, a single stage is of no particular use, and its tonal character is limited. Usually we would like the amp to offer various features and controls, and most importantly to generate the desired levels of distortion. For this, multiple valve stages are required, each feeding the signal to the next in a manner known as **cascading** for obvious reasons. A low to medium-gain amp might use two or three stages before the output stage (plus a phase inverter if the amp is push-pull). Modern high-gain amplifiers generally rely on plenty of preamp distortion, which usually requires four or more stages. Connecting individual stages together is known as **coupling** (in the same way as a train of wagons are coupled to one another by means of linkages), and provides innumerate tone-shaping opportunities, only a few of which can be discussed here.

The amplifier's input:

Perhaps the most obvious place to start is at the input of an amplifier, where the guitar –or some other piece of audio equipment– is coupled to the very first gain stage. The valve requires a grid-leak resistor of course, and this also defines the input impedance of the amp. This raises the question; what is the ideal input impedance for a guitar? The output impedance of a typical guitar varies wildly. At full volume it is defined mainly by the guitar pickups, and could be as low as a couple of kilo-ohms, but if the guitar's volume pot is turned down somewhat then the output impedance can quickly rise to several hundred kilo-ohms, depending on the pot's resistance (usually 250kΩ with single coils or 500kΩ with humbuckers). Therefore, it is wise to make the input impedance of the amp at least 470kΩ or the guitar signal could be too heavily attenuated even before it reaches the first gain stage. This is important; noise is being generated in the guitar cable all the time by its own resistance and by interference from external sources. The amplifier itself also generates a fairly steady level of noise, known as the **noise floor**, so it is most important that the guitar signal reaches the amp at the highest amplitude so it may be amplified immediately by the first stage, so as to preserve a good **signal-to-noise ratio**. The most common value is 1MΩ and there is no particular reason to alter this. Higher values increase noise while offering little further benefit.

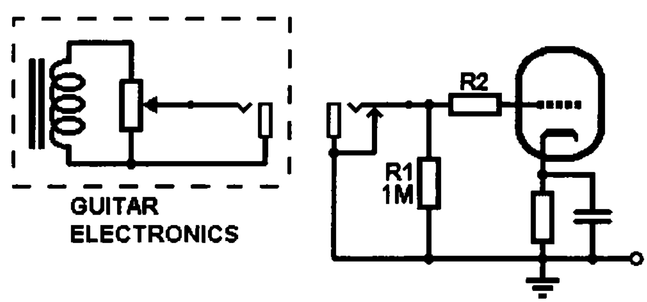


Fig. 2.1: A simple guitar output circuit and amplifier input circuit.

As mentioned in the previous chapter, the preamp stages used in guitar amplifiers are cathode biased, with the grid referenced to ground (zero volts) by the grid-leak

(R1 in fig. 2.1). Because the output from a guitar or any properly designed piece of audio equipment is pure AC, it can be injected directly into the grid of the first stage. The input jack socket is always of the switched, shorting type, so that when no cable is plugged in the amp's input is grounded via the jack socket to prevent it from picking up noise and producing an annoying hum in the speaker.

The input grid-stopper:

There is one other essential component though: the **grid-stopper**. A grid-stopper is any resistor that is placed directly in series with the grid of a valve, such as R2 in fig. 2.1. They have several uses, but at the input of the amp its main purpose is to eliminate or 'stop' radio frequency interference by forming an RC (low-pass) filter with the input capacitance of the valve, as shown in fig. 2.2. In the previous chapter it was shown that that the input capacitance, Cin, of a typical ECC83 gain stage is about 100pF, so if we wished to attenuate all non-audible frequencies above 20kHz then we might calculate the required grid-stopper to be:

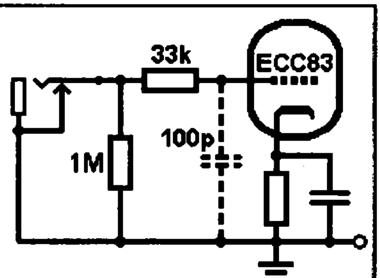


Fig. 2.2: A grid-stopper forms a low-pass filter with the valve's input capacitance (dashed).

$$R = \frac{1}{2\pi f C_{in}}$$

$$R = \frac{1}{2\pi \times 20000 \times 100\times10^{-12}}$$

$= 79.5\text{k}\Omega$

The nearest lower standard value is 68kΩ, and this is the value used in countless vintage amps. However, the guitar's output impedance *also* contributes to series grid resistance. This does not much matter if the guitar's controls are at maximum, but turning down a 500kΩ volume pot slightly could easily raise the output impedance to 200kΩ. Together with the 68kΩ grid-stopper this will then attenuate all frequencies above:

$$f = \frac{1}{2\pi \times 268000 \times 100\times10^{-12}}$$

$= 5.9\text{kHz}$

which is low enough to be audible. This is why a guitar will often sound dull or lacking in brightness at lower volume settings. Fortunately, radio interference as low as 20kHz is very unlikely and so the grid-stopper can usually be made much smaller in value, which will alleviate this problem and reduce noise too. The author generally uses 33kΩ, but values as low as 10kΩ are usually sufficient. Some builders may be tempted to remove the grid-stopper altogether, but this is *not recommended* as it invites high frequency oscillation due to cable inductance. A value of at least 100Ω should *always* be used, and is low enough to be completely tonally transparent.

In classic designs the grid stopper is often positioned before the grid-leak resistor. This forms a potential divider and will attenuate the guitar signal somewhat, worsening the signal-to-noise ratio, so most modern designs place the grid-stopper after the grid-leak to prevent this, as in fig. 2.2.

A typical input circuit is shown in fig. 2.3, where two input jacks are provided. J2 is switched so that if only J1 is used, R1 and R2 form a potential divider that attenuates the guitar signal by half (-6dB), giving a 'low gain' option, as this is preferable to turning down the guitar's own volume control for reasons stated earlier. (Of course, the low-gain function is lost if both jacks are used simultaneously.)

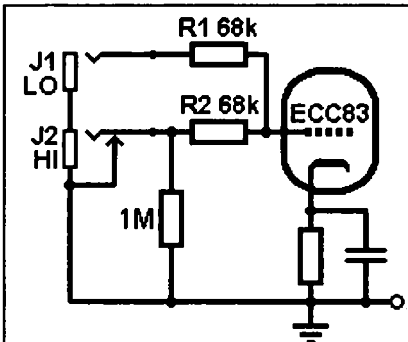


Fig. 2.3: A typical dual-input circuit.

A few obscure amplifiers have used more elaborate, frequency-dependant input circuits*. This is to be frowned upon in current design, where the emphasis is on preserving as much of the guitar signal as possible until it has been amplified above the noise floor by the first stage, with one possible exception: very high-gain, lead amps. In such amplifiers it is sometimes beneficial to immediately attenuate high-treble frequencies at the input in order to reduce the effects of 'fret squeal', which can be an annoyance on high-gain solos. This could be achieved by using a large grid-stopper, but a better, low-noise solution is to use a shunt capacitance after the grid stopper to shunt treble frequencies to ground. A suitable roll-off point is around 4kHz to 6kHz. Assuming a 33kΩ grid stopper and a 4kHz roll-off, and neglecting guitar output impedance:

$$f = \frac{1}{2\pi \times 33k \times 4k}$$

= 1.2nF

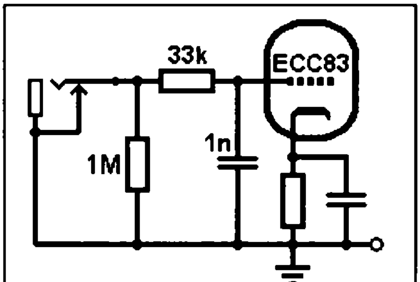


Fig. 2.4: In very high gain amps it may be beneficial to attenuate high-treble 'fret-squeal' at the input.

100pF is already provided by the input capacitance of the valve of course, so 1nF would be quite suitable, as shown in fig. 2.4. This might seem extreme, but treble frequencies can always be preferentially boosted (and therefore more heavily distorted) later in the preamp to compensate for the loss, which works well for a very high-gain, lead tone.

The physical position of the grid-stopper is also worth considering. Since all wiring can act as an antenna for radio

* Such as the Ampeg B12X and B22X.

frequencies, it should be placed as close to the input valve as possible, and preferably soldered directly to the valve socket for maximum effect. Input wires from the jack socket should be kept as short as possible and tightly twisted, or even shielded: this is of utmost importance in high-gain amplifiers!

Fundamentals of interstage AC coupling:

The output level from a typical guitar pickup is usually between about 0.5Vp-p and 1Vp-p, though this can vary wildly depending on type. Even if the input valve is a sensitive one such as the ECC83, or even a pentode, it will not be driven hard enough by the guitar to generate significant distortion. Even active pickups or a booster pedal is unlikely to deliver more than about 4Vp-p, which is barely enough to cause anything but mild clipping in a typical ECC83 stage. The job of the first stage is mainly just to amplify the signal quite ordinarily, to lift it above the noise floor and bring it to a level where it may overdrive the following stage.
For an input signal amplitude of 1Vp-p, if V1 has a gain of 60, say, then the signal appearing at its anode will be 60Vp-p, but this is superimposed on the normal DC anode voltage. In other words, the signal appearing at the anode is not a pure AC signal but a varying DC signal. If we wish to pass the signal on to a second, similar gain stage, then the DC component must be filtered out and only the AC signal passed on to V2. This is easily done with a **coupling capacitor** (sometimes known as a **blocking capacitor** since it 'blocks' DC). This is illustrated in fig. 2.5. The AC signal then appears across V2's grid-leak, Rg, and may be amplified again. This is known as **C-R coupling, capacitor coupling** or (usually) **AC coupling**, since only the AC component is passed between stages.
The coupling capacitors used in an amp are *very* important to overall tone, because not only do they block DC but they can also be used to deliberately attenuate bass frequencies since C1 and Rg form a high-pass filter. The value of the coupling capacitor, C1, is given by:

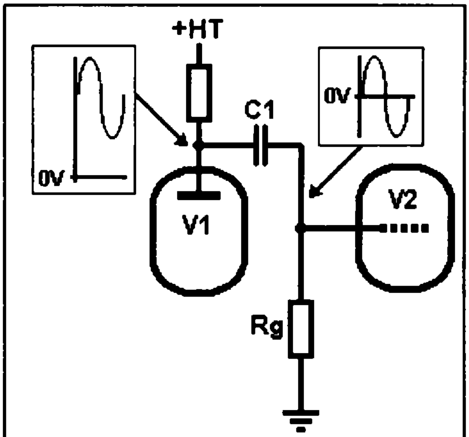


Fig. 2.5: Coupling capacitor C1 'blocks' DC at V1's anode from reaching V2, allowing only AC signals to appear across R1.

$$C = \frac{1}{2\pi f R'} \quad \text{.......... X}$$

Where:
f = the desired -3dB roll-off frequency.
R' = Rg + Z_{out}. The total resistance formed by R1 *plus the output impedance of the first triode* (calculated using formula VIII or IX) since these are both effectively in series with the capacitor, as indicated by the Thévenin equivalent circuit in fig. 2.6.

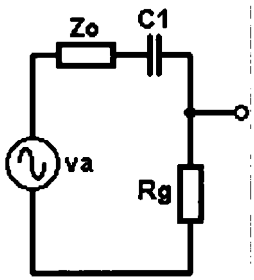


Fig. 2.6: Thévenin equivalent of the coupling circuit in fig. 2.5.

With triodes, Rg will usually be very large compared to Z_{out}, in which case Z_{out} may be safely ignored.

In a hifi amp we would normally set the roll-off frequency very low, say 10Hz. If grid-leak Rg is 1MΩ, and the output impedance is 40kΩ, then:

$$C = \frac{1}{2\pi fR'} = \frac{1}{2\pi \times 10k \times 1040k}$$

= 15.3nF

A common value found in classic amplifiers is 22nF giving a roll-off of about 7Hz, betraying its hifi origins!

In modern designs, even bass amps, there is no need for such a low frequency roll-off. It is much more desirable to use a smaller capacitor to attenuate very low frequencies –since these are almost entirely unwanted noise signals– and because it will reduce blocking distortion (discussed later) and promote a 'tighter', more controlled bass tone. Having too much bass response in an amplifier nearly always leads to 'woofiness' or 'muddiness', *even in a bass amplifier*. This is especially true in live situations where careful bass attenuation is desirable to give distinction and clarity to the instruments, and this becomes more important the louder the band is playing, something which is not always appreciated by amateur players.

Immediately following the input stage it is quite beneficial to set the roll-off point of the coupling capacitor well above 50Hz since this will assist is reducing the mains frequency heater hum, which is usually most prevalent in the first valve.

Keeping bass frequencies at bay is also useful from the point of amplifier stability as it places less strain on the power supply, output transformer and speaker. Remember, the narrower the bandwidth of an amplifier the easier it is to make stable; there is no benefit to making the amplifier capable of reproducing frequencies that we do not actually need, and nowhere is this truer than in a high gain, overdriven and generally abused guitar amplifier. You have been warned.

As a design rule-of-thumb, it is better to attenuate low frequencies the most, early in the preamp, since this is where most low frequency noise (hum) is picked up. They can then be boosted later on if desired, and it is worth noting that the traditional FMV tone stack used in most amps has a significant bass emphasis.

The DC voltage rating of any coupling capacitor should be *greater* than the highest possible HT supply voltage, which is usually equal to the *peak* AC voltage supplied by the power transformer, even though the working voltages may be lower. The reason for this is that when the amp is first switched on the cathodes will be cold and no anode current will flow. The HT voltages –and therefore the anode voltages on the valves– may rise higher than their normal working levels, and it is imperative that the coupling capacitors are capable of withstanding this. Fortunately, high-voltage capacitors rated for 400Vand higher are readily available, quite cheaply.

Foil type capacitors of the PTFE (Teflon), polystyrene, polypropylene, polycarbonate, and polyester composition generally give the smoothest and most favourable subjective tone.
Metallised polypropylene, polycarbonate and polyester types also very good, and are usually indistinguishable from the former type in terms of tone. Incidentally, the famous Mallory 150s and Mullard 'mustard' capacitors are both metallised polyester types, while the Vishay Sprague 'orange drop' capacitors are metallised polypropylene.
All these types may be lumped under the title of "poly' capacitors", and nearly all capacitors up to about 1μF fall into this category.
Silvered-mica capacitors give the most 'theoretically perfect' performance, but are very expensive, and some builders have suggested that they can sound too 'sterile'.
Commonly available ceramic capacitors are well known for having a 'grainy' sound at high voltages, since their capacitance varies with applied voltage, causing unpleasant distortion. They are also rather microphonic. However, they can be useful when used in select positions within the amp, in order to add texture to the tone, and one famous manufacturer is reputed to use them in parallel with poly' types for this reason!

The AC load line:

Before continuing with AC coupling in more detail is would be prudent to explain how to select a suitable value of grid-leak for V2.
In the previous chapter we saw how the anode resistor forms the load on a valve, and how to draw a load line to indicate this. However, we must now consider that, *as far as AC is concerned*, Rg (or any impedance following the coupling capacitor) is also loading the valve since AC is free to pass via C1. Fig. 2.7 shows a Thévenin equivalent circuit, remembering that, to AC, the power supply and ground are one and the same (see fig. 1.26). Ra and Rg therefore appear in parallel, so the total load impedance on the valve must be lower for AC than for DC.

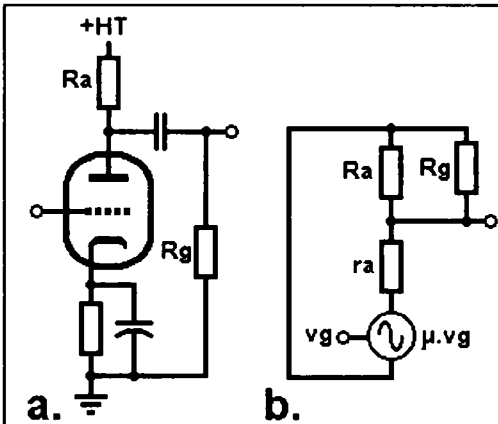


Fig. 2.7: A simple gain stage, **a.**, and its Thévenin equivalent, **b.**, showing how Ra and Rg appear in parallel to AC.

The AC load may be analysed using an **AC load line**. This is done by first drawing the DC load line as described in chapter 1.
For example if Ra = 100kΩ and Rg = 220kΩ, and the HT is 260V, then the DC load line will pass between Va = 260V, and Ia = 260 / 100k = 2.6mA. The bias point is set by the quiescent current, which is DC, so it must lie on this line. In this case, suppose the bias is −1V.

The AC load is:

$Ra_{(ac)} = Ra||Rg$

$$Ra_{(ac)} = \frac{100k \times 220k}{100k + 220k}$$

$= 69k\Omega$

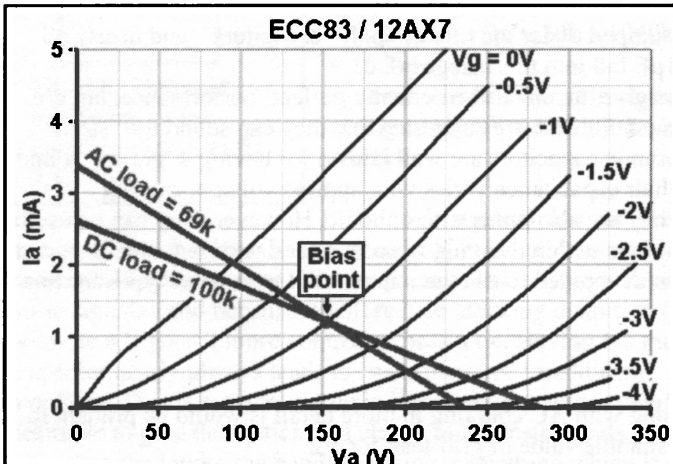


Fig 2.8. DC and AC loadlines for the circuit in fig. 2.8. The addition of the AC load has reduced the possible gain and output swing of the stage.

Since the DC bias point cannot be affected, the new AC load line must pass through it. Noting the quiescent anode voltage of 150V and quiescent anode current of 1.1mA, the AC load line must be drawn between the bias point and: 1.1 + (150 / 68k) = 3.3mA, and can also be extended down to the abscissa of course, as shown in fig. 2.8. The initial load line has a gradient of 1/Ra, but is forced to rotate about the bias point to a new gradient of $1/ Ra_{(ac)}$.

This has several implications:

- The gain has been reduced (from about 60 to 50).
- The maximum output signal swing has been reduced (from about 180Vp-p to 130Vp-p).
- The input sensitivity has increased (from about 3.5Vp-p to 3Vp-p).
- The harmonic distortion has increased.

The attentive reader may wonder why it now appears that more anode current can flow in the valve than can actually be supplied by the 100kΩ anode-load resistor? The answer to this is that when the anode voltage rises, the coupling capacitor will charge slightly, effectively 'stealing' current from the anode. As the anode voltage falls again this stored charge is supplied back to the anode, and this is indicated by the arrows in fig. 2.9.

From this treatment we may deduce that the gain formula, III, may be modified to give a more accurate result:

$$A = \frac{\mu(Ra \parallel Rg)}{(Ra \parallel Rg) + r_a} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{XI}$$

Where Rg is the following load resistance.

In fact, all equations given in the previous chapter should have Ra replaced by Ra||Rg if accuracy is to be maintained when a stage is heavily loaded.
However, provided that Rg is greater than about five times Ra, its effect on the stage will be negligible and the formulae given earlier may be used without significant inaccuracy. Therefore, in most preamp designs we will find that Rg is in the region of 470kΩ to 1MΩ. Occasionally we might deliberately load a stage heavily in order to obtain the characteristics of a low gain stage, but without the extra current demands on the HT that would come with using a low value anode resistor.

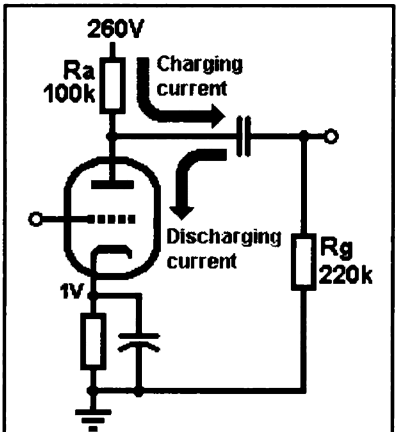


Fig 2.9: A gain stage which has been heavily loaded by the following grid-leak resistor, causing a reduction in expected gain.

Nevertheless, the most common place to find a heavily loaded stage is where a gain stage feeds a power valve, which cannot tolerate such large values of grid-leak.
For reference, the graph in fig. 2.10 shows the gain of a typical ECC83, calculated using formula XI, for different values of anode resistor when loaded by various values of loading resistance (Rg).

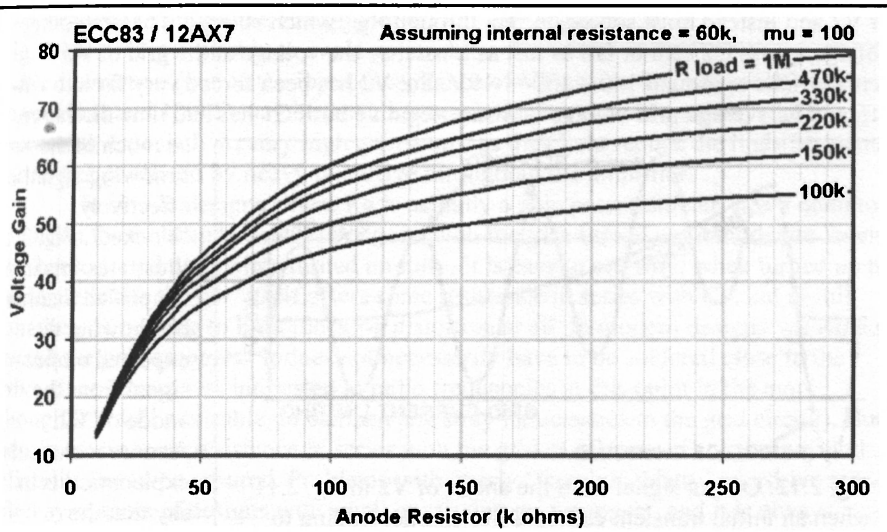


Fig. 2.10: Variation in gain with anode resistor, for various values of following load resistance.

Blocking distortion and grid stoppers:

It was shown earlier how a grid-stopper may be used to attenuate high frequencies. However, they also serve another purpose when stages are AC coupled: they prevent **blocking distortion**.

Blocking distortion occurs when an AC coupled valve is overdriven.

Consider the circuit in fig 2.11. At quiescence the voltage at the anode of V1 is 150Vdc, and the voltage on the grid of V2 is 0V, therefore the coupling capacitor C1 has 150V across it. Under signal conditions, if the anode voltage of V1 rises by 1V, a tiny amount of current will flow into C1, and out of it, down the grid-leak Rg. The voltage across R2 therefore rises by 1V and this is how the AC signal appears at the grid of V2, as expected. The voltage across C1 is still 150V. We will suppose that this is the maximum signal which can be fed to V1 before overdriving.

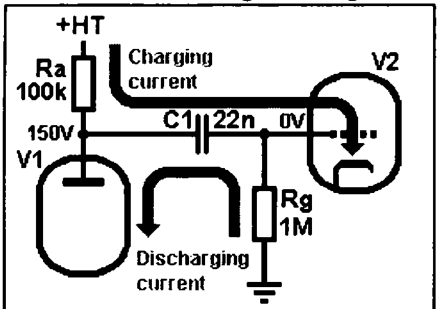


Fig. 2.11: Current paths in an overdriven, AC coupled stage. The absence of any grid-stopper leads to blocking distortion.

Now, let's suppose that the anode of V1 continues to rise by a further 9V. Current continues to flow into C1, and out of it, but now V2 is drawing grid current which prevents the grid voltage from rising any further above 1V. The voltage across C1 must therefore increase to 159V, so considerable charge is being stored in it.

Now we will assume that the anode voltage of V1 swings down to 140V for the other half of the waveform. Normally we would expect the voltage across Rg to also fall to -10V (putting V2 into cut-off), but C2 is now unable to discharge via the grid of V2 and instead must source current through Rg , which is large. Therefore, the voltage across C2 cannot fall as fast as it rose, so the voltage at the grid of V2 actually falls to around 140 – 159 = -19V. So, V2 has been forced very far into cut-off, and its average grid voltage will stay negative until C1 has had time discharge, so much of the following signal is effectively 'blocked'. Fig. 2.12 illustrates this in a simplified manner, by showing the expected *output* signal from the anode of V2.

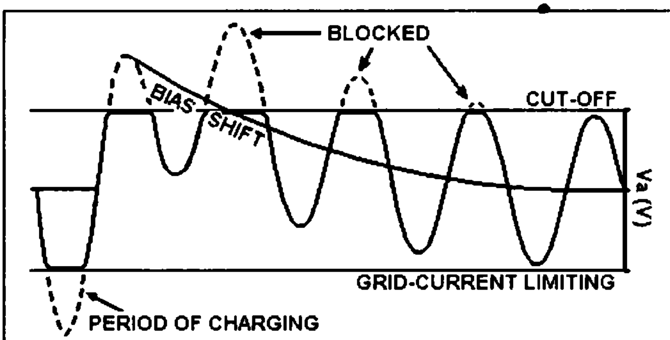


Fig. 2.12: Output signal from the anode of V2 in fig. 2.11 when an initial transient causes C1 to charge, leading to the 'blocking' of the following signal.

A more technical explanation is to say that the *charging* time-constant of

$Ra + Rg \times C1 = 0.02s$. But when V2 is overdriven this falls dramatically to a value determined by the input impedance of the grid, which is typically a few hundred ohms, to around 0.002s, allowing C1 to charge up significantly. When V2 stops being overdriven the *discharging* time-constant returns to $Rg + ra \times C1 \approx 0.02s$, meaning it takes about ten times longer to discharge to its normal state again, which causes an unwanted bias shift that places V2 into cut-off for a period of time, which 'blocks' any forthcoming signals.

Obviously the problem is worsened at low frequencies since C1 will have a greater period of time in which to charge. Blocking is worsened further if V2 has a cathode bypass capacitor since this will be charged up somewhat by the grid-current, but may take longer to discharge via the cathode resistor, placing V2 even further into cut-off. These are even more reasons why bass frequencies should be carefully attenuated and controlled, and coupling and cathode-bypass capacitors kept as small as possible in a guitar amp.

Depending on severity, the sound produced by blocking distortion may vary from an ugly, 'blatting' sounding overdrive, to a very unpleasant raspberry sound known affectionately as 'farting out'. Preamp valves are the most susceptible since they are the most easily overdriven, and many amps suffer from the problem; remember, older designs were never intended to be heavily overdriven, but many contemporary amps are simply copies of these.

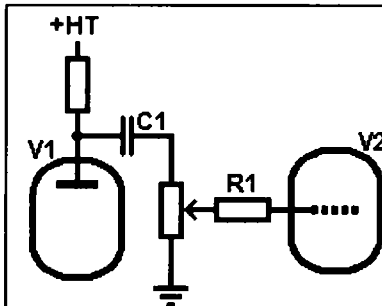


Fig. 2.13: A grid-stopper (R1) of greater than 10kΩ is essential to reduce blocking distortion.

The solution to this problem is not only to curtail bass frequencies, but also to add a grid-stopper to V2 to inhibit the charging of C1 [fig. 2.13]. If it is desirable to avoid noticeable treble attenuation (due to the input capacitance of V2) then a low value between about 10kΩ and 100kΩ may be used, but much larger values are usually quite successful, especially in very high gain amps, as they help reduce the high-frequency harmonics generated by heavy overdrive throughout the amplifier.

In simple designs where Rg is actually a gain potentiometer, it is a common condition for the amp to sound very good with the gain turned up to moderate levels, say, but to sound poor when turned up fully. It is easy to see why; when turned up to moderate levels the pot' itself offers some resistance in series with C1, but at full gain there is nothing to limit blocking distortion at all. In modern designs: ***all valves must have grid-stoppers!*** It does not necessarily have to be soldered close to the valve since we are less interested in radio frequencies at this point in the amp (though it is still advisable, to dampen any stray inductances in the grid circuit). But there must be *some* resistance in series with the grid *at all times* if consistency and reliability are to be ensured. Problems with overly 'fizzy' or 'blatty' overdrive are often symptoms of designs with small or missing grid stoppers, and it is always advisable to add grid-stoppers to any existing amplifier that does not already have them. A value of 10kΩ would be suitable on preamp valves (usually less on power valves) to avoid altering the existing tone, but many users will find they actually like the tone when values of up to 1MΩ are used. This is one of the most valuable pieces of information regarding tone-tweaking.

Practical AC coupling networks:

Now that the reader appreciates how coupling networks act upon the previous stage, by loading it, and how they can lead to blocking distortion in the following stage, we can examine some of the common coupling networks found in guitar amps. In most modern amps, and especially in high-gain amps, considerable frequency shaping and attenuation takes place between stages. Often this is made user-variable by means of tone and gain controls, though not always (too many controls become a hindrance rather than a benefit).

Wide-band attenuation:

In amplifiers which will be driven to anything more than mild levels of overdrive, it will usually be necessary to attenuate the signal level between gain stages. Typical triode stages are capable of generating peak-to-peak signal levels of around 2/3 HT, or even more for pentodes, which is far too much to feed to other preamp valves without resulting in harsh clipping and susceptibility to blocking distortion. Also, if the cumulative gain of the preamp is extremely high the amp will usually lack touch sensitivity, and may even end up sounding like a fuzzy, solid-state amp.

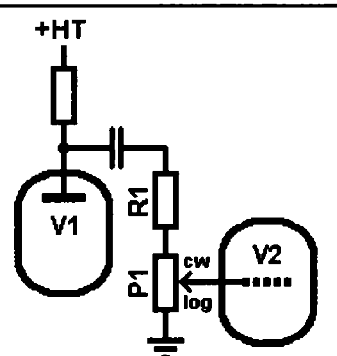


Fig. 2.14: It is usually necessary to limit the maximum signal passed between stages.

A gain control may well be included between the stages, but a further resistor will usually be added to limit the absolute maximum signal level. This is shown in fig. 2.14. R1 is the attenuating resistor, and serves the dual purpose of acting as a grid-stopper for V2. Preamps which don't use this form of attenuation usually resort to using very large grid stoppers to subdue harsh sounding distortion, which is a less elegant (and noisier) solution.

Of course, an obvious question is: why not redesign the preceding stage for less gain, rather than generate too much and then 'throw gain away' between stages? Indeed, in a clean amp there is a good argument for operating at higher anode currents and less gain as this usually promotes 2nd order harmonic distortion and a very 'warm' and even 'shimmering' sound. Indeed, very good high-gain tones, with good character and touch sensitivity can be obtained by alternately using high-gain and low-gain stages, either by using a mixture of valve types (particularly the ECC83 / 12AX7 with the ECC82 / 12AU7 or 6SN7) or by heavily loading some stages with low-value grid leak resistors, a method adopted in the Trainwreck *Express*.

However, it is often the case that the sharper tonal quality of an ECC83 operating with a large load impedance is preferable to the rounder, less 'focused' overdriven tone produced under lower load impedances. Also, even low-gain valves still produce very high amplitude output signals, so some form of interstage attenuation will inevitably be required. Furthermore, attenuation between stages offers a natural opportunity for frequency shaping and tone control, as will be discussed.

Unfortunately, there is no particular rule governing the degree of attenuation required between stages. Knowing how much to attenuate the signal comes partly from experience, but mainly by experimentation. The normal approach would be to select P1, often 1MΩ, and then try different values for R1 (or even tack in a variable resistor) and adjust to taste before soldering in the final value. Referring to fig. 2.15, the gain of the interstage divider (actually a loss), β, is given by:

$$\beta = \frac{R2}{R1 + R2} \quad \text{.......................XII}$$

and will usually be around 0.5 to 0.9 (in other words, the ratio of R1:R2 could be 1:1 to 1:10), though this is merely a guide.
However, once the perfect amount has been found, it is worth considering whether the values of resistance can be scaled down, and capacitances scaled up, to reduce noise while retaining roughly the same circuit performance.

For example, fig 2.15a shows a network which appears in both the Mesa *Dual Rectifier* and Soldano *SLO-100*, but circuit *b.* offers almost identical performance with less than half the resistor noise. The higher resistances in *a.* will produce more high-frequency attenuation due to the input capacitance of V2, but this could easily be duplicated in *b.* by adding C1 (which should be roughly equal to the dynamic input capacitance of V2, in this case), though we probably wouldn't find it necessary. Scaling the resistances down too far would begin to load V1 heavily, of course, though it might not turn out to be detrimental!

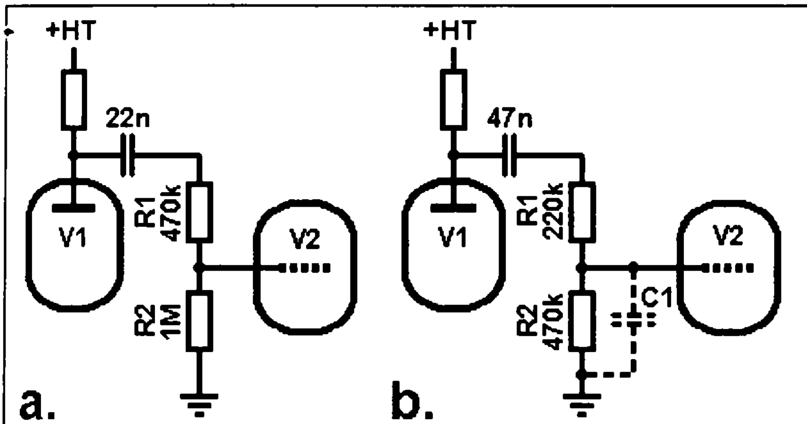


Fig. 2.15: Circuit ***a.*** shows a typical attenuation network, while circuit ***b.*** offers the same attenuation but with less resistor noise.

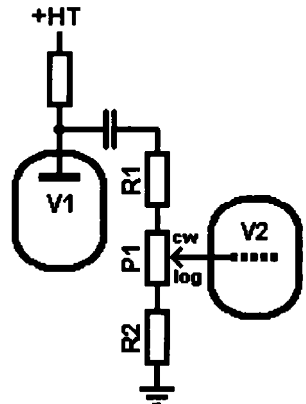


Fig. 2.16: R2 prevents the amplifier from being completely silenced if P1 is turned fully down.

Any gain controls will usually be of the logarithmic type, since they usually affect the overall volume of the amp rather then just the perceivable distortion. However, guitarists who are more familiar with solid-state amps may be confused by controls which are labelled 'gain' but which can be turned completely down, silencing the amplifier like a regular volume control. Some designs will therefore place a resistor between the gain control and ground –R2 in fig 2.16– to limit the minimum degree of attenuation. Its value will usually be in the region of 10kΩ.

Treble boost (bass cut) and 'bright' capacitors.

Treble boost or bass cut is an important feature of guitar preamps. Not only does cutting bass reduce hum, blocking distortion and avoid muddiness, but enhancing treble frequencies also increases harmonic distortion, because the harmonic distortion components generated by amplifiers are nearly always integer multiples of the fundamental. As a simple example; if we were to pass a 1kHz sine wave through a triode gain stage producing 10% 2nd harmonic, the output will contain 90% 1kHz signal, plus 10% 2kHz distortion. If we then pass that resultant signal through a high-pass filter causing 6dB attenuation at 1kHz, but no attenuation at 2kHz, then the output will now contain 45% 1kHz and 55% 2kHz. In this way the relative proportions of harmonics is increased by treble boosting. What's more, the boosted frequencies will then overdrive the following stages to a greater degree than the cut frequencies, increasing their apparent sustain, a significant feature of high-gain lead tones.

There are many ways to enhance higher frequencies, including using partial cathode bypassing [chapter 1], but interstage coupling offers even greater scope for tailoring the frequency response of the preamp.

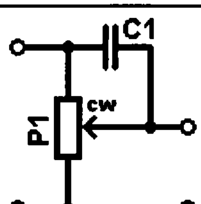


Fig. 2.17: A 'bright capacitor' may be placed across gain controls to avoid loss of treble at low settings.

It was mentioned at the start of this chapter that a problem caused by the input capacitance of a triode is that when a preceding gain or volume control is turned down, the resistance in series with the grid increases, causing more treble attenuation than when the control is fully up. To combat this, fig 2.17 shows a small capacitor connected across the potentiometer, to allow high frequencies to bypass it. This is often referred to as a **bright capacitor**, after the many Fender amplifiers having a 'bright channel' which was different from the 'normal channel' only by the addition of this capacitor. Adding a bright capacitor to an amplifier's 'normal' channel is one of the most popular and simple of amp modifications. High frequencies are 'shelved' the more the potentiometer is turned down.

For example, fig 2.18 shows a simple coupling circuit similar to that found in the Marshall *50W Master Volume*. Fig. 2.19 shows the frequency response of this circuit without the bright capacitor (faint), illustrating the increasing loss of treble as the gain control is turned down. The bold traces show how adding the capacitor causes a treble boost, which helps maintain good tone at low volumes. At full gain both traces are identical of course, since the bright capacitor is short circuited by the wiper of the potentiometer. In many amplifiers a switch is placed in series with the capacitor*, to allow selection between 'normal' or 'bright' modes.

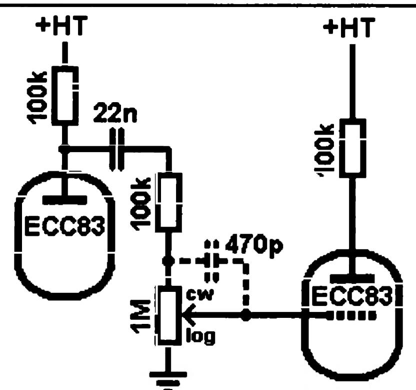


Fig. 2.18: A coupling circuit similar to that found in the Marshall *50W*

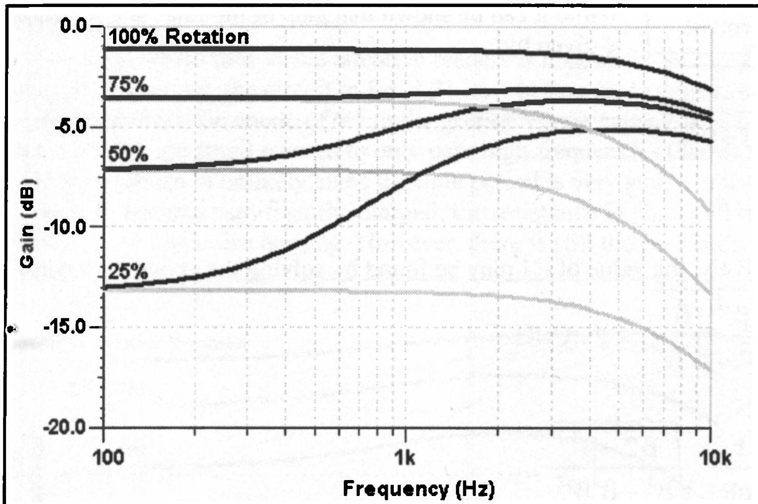


Fig. 2.19: The frequency response of the circuit in fig. 2.20 without the bright capacitor (faint), and with the capacitor added (bold).

Again, this component is usually selected to taste. A good starting point is to note that in fig.1.17, at 50% rotation, and neglecting any other circuit impedances, the

* e.g., the 'ultra high' switch on many Ampegs, the 'bright' switch on many Fender *Bassmans*, and the 'brilliance' switch on the Sound City *Concord*, among many others.

value of bright capacitor is given by:

$$C = \frac{2.74}{2\pi fR}$$

Where:
C = the capacitance of the bright capacitor.
f = the +3dB or half-boost frequency (typically between 800Hz and 5kHz).
R = the resistance of the potentiometer.

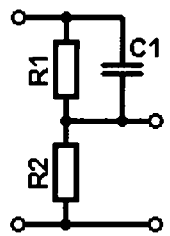


Fig. 2.20: A common treble boosting arrangement.

A similar treble boost (bass cut) function may be achieved by bypassing the upper resistor in a potential divider, as in fig. 2.20. (It should be obvious that this is really just a fixed version of the circuit in fig. 2.17). This compensates for a loss in treble caused by the presence of a large attenuating resistance, such as that in fig. 2.15a.
Assuming the output impedance of the preceding stage is small compared to R2, then at very low frequencies the gain of the network, β, is simply:

$$\beta = \frac{R2}{R1 + R2}$$

While it can be shown that gain of the filter at any frequency, β', is given by[1]:

$$\beta' = \beta\sqrt{\frac{1+k^2}{1+\beta^2k^2}} \quad \text{.......XIII}$$

Where:
$k = 2\pi fC1R1$

Alternatively, the value of C1 may be found by solving the above for k, yielding:

$$k = \sqrt{\frac{\beta'^2 - \beta^2}{\beta^2 - \beta^2\beta'^2}} = 2\pi fC1R1$$

so,

$$C1 = \frac{1}{2\pi fR1}\cdot\sqrt{\frac{\beta'^2 - \beta^2}{\beta^2 - \beta^2\beta'^2}} \quad \text{.......XIV}$$

Where β' is the desired boost at frequency f, and must lie between β and 1. To make this the half boost frequency take β' to equal:

$$\beta'_{(\text{half boost})} = \frac{\beta + 1}{2}$$

[1] Voorhoeve, N.A. J. (1953) *Low Frequency Amplification*, p203. Cleaver Hume Press Ltd., London.

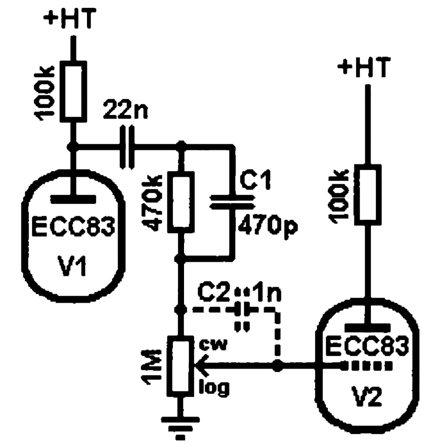


Fig. 2.21: A treble boosting circuit found in the revised Marshall 50W Master Volume.

Fig. 2.21 shows an example taken from the 'revised' Marshall *50W Master Volume* though similar networks can be found in most medium to high-gain amplifiers. The frequency response of this circuit is shown in fig. 2.22. The bold traces show how the addition of the treble boost capacitor, C1, counters any loss due to the input capacitance of V2, though the resistance offered by the potentiometer to V2 begins to slightly reduce frequencies above about 5kHz at lower volume settings. The revised Marshall *50W Master Volume* also uses the bright capacitor, C2. The faint traces in fig. 2.22 show the response with both capacitors used; frequencies above about 2kHz are hardly reduced even at low gain settings, and it is worth noting that this arrangement has so much treble boost that it often produces a 'thin' or 'peaky' sound at lower volume levels, so many player prefer to remove C2 from the Marshall circuit.

One thing which may strike attentive readers is that, as far as higher frequencies are concerned, the circuit in fig. 2.21 now has no grid stopper; there is an uninterrupted path from the anode of V1 to the grid of V2 via capacitors C1 and C2. Provided C1 and C2 are small enough to only pass high frequencies (above 500Hz say) blocking distortion in unlikely since the time period is very small, and even if C1 or C2 were to become significantly charged, the resistance in parallel with each one will allow fairly rapid discharging. However, there is still the possibility of high

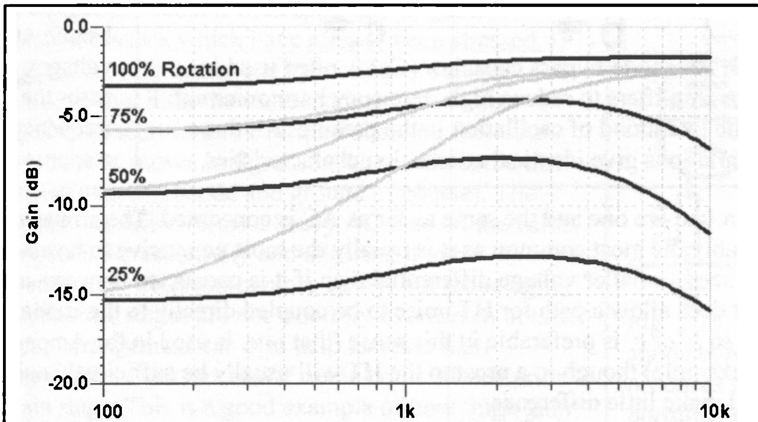


Fig. 2.22: The frequency response of the circuit in fig. 2.23 with treble boost capacitor C1 (bold), and with both C1 and C2 (faint).

frequency oscillation due to stray circuit impedances, so a small grid stopper, less than 10kΩ say, should still be added for peace of mind, even though most commercial amps fail to do so.

Treble cut (bass boost) and anode-bypass capacitors:

In very high-gain amps, severe clipping is liable to generate high-order harmonics which may make the tone too harsh, buzzy or shrill, particularly when large amounts of treble boost have been used between gain stages. High-gain preamps are also prone to oscillation at high frequencies due to positive feedback via parasitic capacitances between various wires and components. To alleviate either problem it is common to find a small treble-shunt capacitor immediately following one or more gain stages. Such a capacitor is sometimes referred to (misleadingly) as a **snubbing capacitor**, though **anode bypass capacitor** is more accurate. In combination with the output impedance of the anode this forms a simple low-pass filter, reducing the bandwidth of the amplifier and so improving stability by reducing unpleasantly high frequencies.

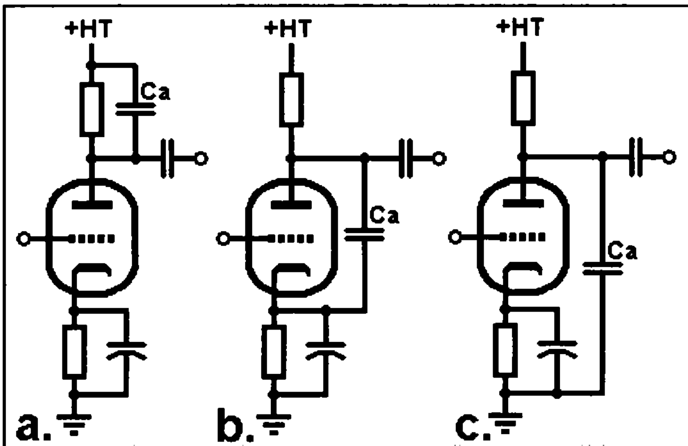


Fig. 2.23: An anode-bypass capacitor (Ca) is often used in high-gain amplifiers to reduce high frequency harmonics and reduce the likelihood of oscillation in the preamp. All three circuit variations give identical attenuation characteristics.

This is illustrated in fig. 2.23, though the anode-bypass capacitor (Ca) could just as easily be connected *after* the coupling capacitor, which would be useful since the bypass capacitor then only has to handle AC voltages. In all cases the frequency response is the same, since the HT and ground are one and the same as far as AC is concerned. The arrangement in *a.* is probably the most common as it is usually the most conducive to layout and the capacitor sees a smaller voltage differential than if it is connected between anode and ground. It does allow a path for HT noise to be coupled directly to the anode, however, so *b.* or *c.* is preferable in this sense (that in *c.* is used in the Ampeg *GS-12R,* for example) though in a preamp the HT will usually be sufficiently noise-free that it will make little difference.

For example, fig. 2.24 shows a typical gain stage, and Ca has been added to reduce high frequency harshness. Assuming any following loading resistances are large with respect to Z_{out}, then its value may be found according to:

$$Ca = \frac{1}{2\pi f Z_{out}}$$

Where f is the desired -3dB roll-off frequency.
In this case the output impedance of the stage, Z_{out}, is 73kΩ, found using formula IX. A typical roll-off frequency is in the region of 5kHz, yielding:

$$Ca = \frac{1}{2\pi \times 5k \times 73k}$$
$$= 442pF$$

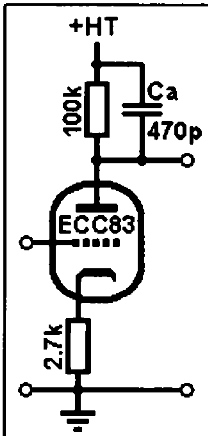


Fig. 2.24: A typical anode bypass circuit.

So values between about 330pF and 470pF might be suitable. In reality, calculation is rarely used. Such capacitors are normally added after the amplifier is mostly completed and are adjusted to taste, usually being in the range of 100pF to 1nF. Values of less than about 100pF are normally used to prevent parasitic oscillation within the preamp, rather than to reduce treble, since they will attenuate frequencies above 20kHz. Some amateur builders are reluctant to use anode bypass capacitors for stability on the assumption that they might alter the tone of the amp, or rob it of 'airiness'. This is often due to confusion with hifi literature which places more emphasis on slew rate and ultra-sonic bandwidth, neither of which are significant in a guitar amp. If such capacitors are chosen properly, oscillation can be avoided while leaving the tone unaffected.

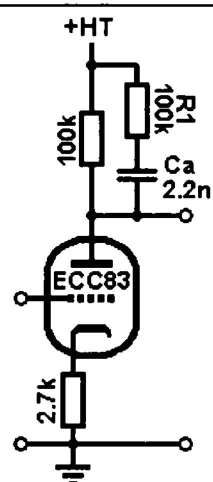


Fig. 2.25: Limited anode bypassing produces a bass boosting, shelving filter.

Fixed bass boost is used sparingly in guitar amplifiers, for reasons which have already been stressed. If it is required, then limited anode bypassing is a simple means of achieving it. In the same manner as the treble boosting circuits seen so far, it forms a shelving filter. An example circuit is shown in fig. 2.25 (R1 and Ca could also be connected between anode and ground of course). The anode bypass capacitor forms a heavy AC load on the valve, increasing its input sensitivity and allowing it to be overdriven at high frequencies more easily, but reducing its gain so following stages are not woefully overdriven in turn This circuit arrangement can also help increase high frequency sustain-even more so if treble boost is used prior to this gain stage. This is a good example of how 'high gain' in the musical sense does not necessarily demand 'high gain' in the technical sense; input sensitivity is what counts.

At very high frequencies the gain of the filter, β, is simply:

$$\beta = \frac{R1}{R1 + Z_{out}}$$

While it can be shown that the degree of attenuation, β', at any frequency is given by[1]:

$$\beta' = \beta\sqrt{\frac{1+k^2}{\beta^2+k^2}} \quad \text{.........XV}$$

Where:

$k = 2\pi fCaR1$

Solving for Ca yields:

$$Ca = \frac{1}{2\pi fR1}\cdot\sqrt{\frac{\beta^2(\beta'^2-1)}{\beta^2-\beta'^2}} \quad \text{.........XVI}$$

Where β' is the desired boost at frequency f, and must lie between β and 1. Again, to make this the half boost frequency take β' to equal:

$$\beta'_{(half\ boost)} = \frac{\beta+1}{2}$$

Fig. 2.26 shows the frequency response of the circuit in fig. 2.24 (faint), illustrating the low-pass filter response formed by Z_{out} and Ca. The bold trace is the response of the circuit in fig. 2.25, showing how the addition of series resistor R1 produces a bass-boosting, shelving filter.

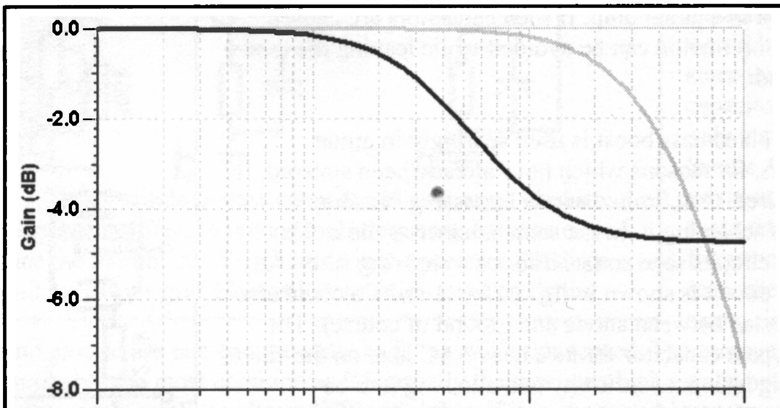


Fig. 2.26: Frequency response of anode bypass circuits in fig. 2.24 (faint), and fig. 2.25 (bold).

[1] Voorhoeve, N.A. J. (1953) *Low Frequency Amplification*, p202. Cleaver Hume Press Ltd., London.

DC coupling:

The type of interstage coupling most commonly used in guitar amplifiers is AC coupling, where the DC component is blocked while AC signals are allowed to pass to the next stage. However, with care it is also possible to couple both DC and AC component simultaneously, so that the input of the following stage exactly tracks the output of the preceding stage.

Direct coupling:

The simplest form of DC coupling is a direct connection between the output (usually the anode) of one stage, and the input (usually the control grid) of the next. This is often referred to simply as direct coupling, and a simple example is illustrated in fig. 2.27 (cathode bypass capacitors are omitted for clarity). It has the obvious advantage that it eliminates blocking distortion, since no coupling capacitors are used, and DC coupled stages have a reputation for generating smooth, strong overdrive even at low frequencies, and this may be why. The avoidance of coupling capacitors is also particularly favoured in hifi as it removes any possibility of capacitor-induced distortion, reduces resistor noise (since no grid leak is required), and phase shift between stages cannot occur, which is important if the stages are enclosed within a feedback loop. The absence of intervening resistors makes interstage frequency shaping is more awkward than with AC coupling, however.

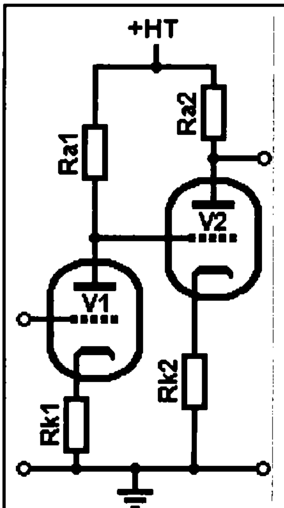


Fig. 2.27: DC / Direct coupling between stages.

It is clear from fig. 2.27 that V1 is a fairly ordinary gain stage, except that its quiescent anode voltage also defines the quiescent grid voltage of V2. For normal operation we know that the cathode voltage of V2 must be somewhat higher than its grid voltage, so its cathode resistor must be unusually large to provide sufficient voltage drop to achieve this. This must also mean that the useful voltage across V2 is much less than that across V1 (unless we were to supply it from a separate, higher voltage HT) so its maximum output signal swing may be limited. A further concern is that because the cathode voltage of V2 is necessarily high, there may be risk of exceeding the valve's **maximum allowable heater-to-cathode voltage, $Vhk_{(max)}$**. This will be quoted on the datasheet, and it indicates the maximum voltage (positive or negative) which the heater insulation can withstand before there is risk of it breaking down. For most valves it is +/-90V, although some datasheets give rather optimistic values up to 180V, and this is the case for the ECC83. Symptoms of breakdown include excessively loud heater hum or intermittent pops and crackles. If the circuit designs demands a high cathode voltage, then it may be necessary to elevate the heater supply so that $Vhk_{(max)}$ is not exceeded [see chapter 12]. Some datasheets also quote a **maximum allowable heater-to-cathode resistance, $Rhk_{(max)}$**, which puts an upper limit on the allowable value of

Rk before heater-to-cathode leakage current becomes significant. For the ECC83 this is 150kΩ, but, admittedly, many designs exceed this value without any apparent problems, provided Vhk is well within limits.

DC coupling is easier to achieve with pentodes or triodes having a low r_a, both of which can be easily biased to fairly low anode voltages. We will consider an example using the ECC82 / 12AU7 ($r_a \approx 8k\Omega$) with an HT of 350V. Choosing component values is a somewhat circular process, and usually requires some juggling of values.
The design of the first stage, V1, follows normal conventions (and could also have a cathode bypass capacitor, of course). In this case a 33kΩ load has been chosen, and a -4.2V bias results in a quiescent anode voltage of 145V, which is now also the grid voltage of V2. We therefore know that the cathode voltage of V2 must be at least equal to this, so the maximum useful *anode-to-cathode* voltage of V2 must now be:
$HT - Vg = 350 - 145 = 205V$.

There is more than one way to proceed, but one method is now to choose an approximate value for anode current for V2 so as to discover roughly how large the cathode resistor will need to be. For the ECC82 we might begin with 5mA, meaning the cathode resistor would have to be at least 145V / 5mA = 29kΩ. A close standard is 33kΩ, so we will take this as a practical value.
We may now choose the anode load resistor, Ra2. In the previous chapter when we designed ordinary gain stages we were able to ignore the value of cathode resistor when drawing a load line, since it is normally very small compared to the anode resistor. This is now not the case; we must consider the total resistance in series with the valve. If we select 22kΩ for Ra2 then the total DC load is 55kΩ, which is not unreasonable for this valve. The DC load line has been drawn in fig. 2.28.

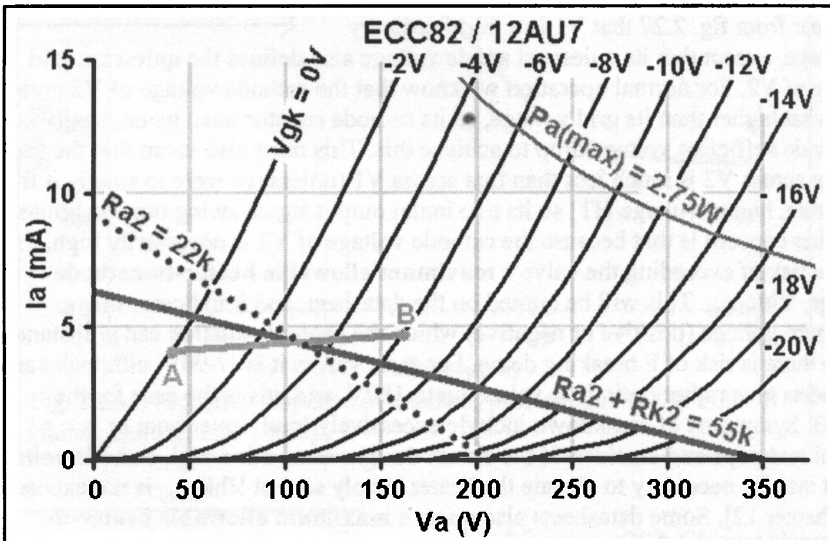


Fig. 2.28: AC, DC and cathode load lines drawn for V2 in fig 2.29.

Since we have already chosen the cathode resistor, Rk2, we may draw a cathode load line to find out if we have a sensible bias point. If the bias were 0V then the cathode voltage must be the same as the grid voltage, which is 145V. The current flowing in Rk2 must therefore be 145V / 33kΩ = 4.4mA, so this point is plotted at A. If the bias were -6V, then the cathode voltage must be 6V higher than the grid, making 145 + 6 = 151V. The current in Rk2 must be 151V / 33kΩ = 4.6mA, and this is plotted at B and the cathode load line drawn between them. The crossing point is the bias point, and in this case is around Vgk = -3V. We can also draw an AC load line for Ra2 only, to indicate the useful range over which the valve will operate, assuming the cathode is bypassed, because when anode current tries to increase it flows into the cathode bypass capacitor, and as it tries to decrease, current is returned from the capacitor into the cathode resistor, increasing Vgk. This is shown by the dotted line passing through the bias point in fig. 2.28.

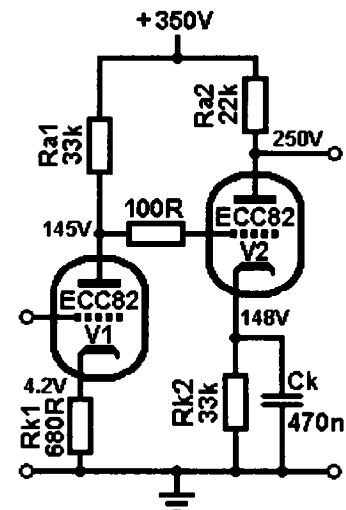


Fig. 2.29: A practical, direct - coupled pair of gain stages.

Because the cathode resistor is so large, cathode current feedback will be very great, and formula IV gives suggests a gain of only 0.6! Therefore, a cathode bypass capacitor, Ck, will be required to achieve full gain from V2, and may be found using formula VI. In this case a 470nF capacitor produces a half-boost frequency of 14Hz. A useful gain control could be implemented by adding a variable resistor in series with Ck (as in fig. 1.20*b*).
The final schematic is shown in fig. 2.29, and an obligatory grid stopper has been added to protect against HF oscillation, and does not otherwise affect normal operation, though a larger value could be used to attenuate high frequencies in the usual manner, of course.

Because each triode draws over 4mA, Ra1, Ra2 and Rk2 must be at least 2W rated. $Vhk_{(max)}$ for the ECC82 is 180V, and $Rhk_{(max)}$ is 150kΩ, so we are within limits in this case. However, there is a less obvious concern: When the power supply is first switched on the cathodes will usually be cold, meaning no anode current can flow. The anode of V1, and therefore the grid of V2, will rise to the full HT potential (350V in this case) while the cathode of V2 is still at zero volts. With such a high potential between grid and cathode there is a significant chance of arcing between the two, which would be extremely damaging to the cathode coating; this *does* happen in amplifiers having DC coupled stages. The traditional solution is to use a standby switch to allow the heaters to warm up before the HT is applied, but there is no guarantee that it will be used, and in any case it would not protect the valve if the heater supply were to fail. A perfect solution is to add the protection components shown in fig. 2.30. At startup, current is allowed to flow from the HT to ground via R1 and D1, keeping the grid and cathode of V2 safely within a few tens

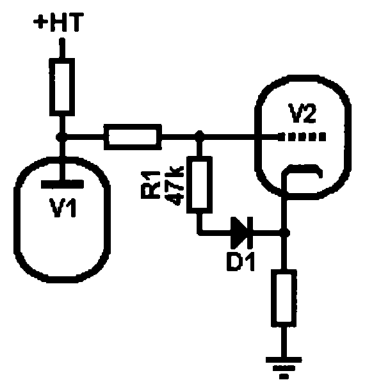


Fig. 2.30: R1 and D1 prevent arcing between grid and cathode at startup.

of volts of each other. When the cathode reaches normal temperature D1 will be reversed biased, off, and plays no further part in circuit operation. R1 is included simply to ensure that D1 does not affect clipping when V2 is overdriven (R1 would not be required in a hifi amp). This will also act as a bleeder path for the power supply capacitors when the amp is switched off. D1 must be rated to withstand the full HT, and will normally be of the same type used in the power supply, such as a 1N4005 or better.

It should now be clear that DC coupling is somewhat more difficult to implement than AC coupling, particularly in guitar amplifiers where the biasing of stages for a particular tone is quite specific. It would probably be necessary to replace Rk2 with a variable resistor to allow more precise bias setting (and high wattage variable resistors are very expensive), though it is worth noting that once set, DC coupled stages are very resilient to fluctuations in the HT and to changing valve characteristics by virtue of the heavy cathode-current feedback which tends to keep the normal bias conditions under control. Nevertheless, it is fair to say that the necessity for quite high HT voltages, the difficulty in setting bias and the limitations on frequency shaping usually outweigh the advantage of freedom from blocking distortion, so it is rarely used.

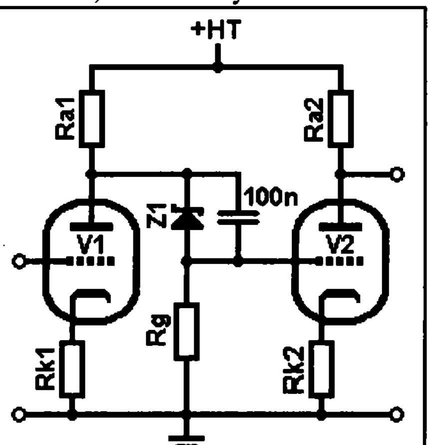


Fig. 2.31: A zener diode can be used to 'level shift' a DC voltage down to the desired level, though it may be difficult to arrange in practice.

Level shifting:

The main problem with DC coupling is trying to marry the naturally high anode voltage of V1 to the grid of the following valve without exceeding any voltage ratings or restricting output swing, but it is also possible to lower that voltage to a more practical level (less than 100V say) in a manner known as **level shifting**. This places less strain on the heater-cathode insulation of V2 and allows a higher voltage to appear across it (and therefore keep output signal swing good), and reduces the chances of arcing at startup whilst still retaining immunity from blocking distortion.

One method is to use a zener diode to drop the desired voltage, and this has the advantage that AC signals are not affected, they are passed directly to V2. This is shown in fig. 2.31. The 100nF capacitor simply shunts zener noise, and has no effect on frequency response. Unfortunately this system is difficult to arrange because enough current must permanently flow down through Z1 to maintain reliable operation, 2mA to 5mA say, and this current is sourced through Ra1. Therefore, the anode current of V1 will need to be significantly greater than this, or its anode voltage will be pulled down too far by the stolen current. Rg must also be around five times larger than Ra1 to avoid undue loading, which itself limits the amount the current we can feed to the zener! Such a circular argument makes this an impractical circuit for a guitar amp.

A much older[1] and more useful arrangement is to level-shift using a normal resistor divider. This is shown in fig. 2.32. The immediate disadvantage is that AC signals are also attenuated by the divider, but capacitor C1 may be added to allow these to pass. Clearly this is the same arrangement as the treble boosting circuit in fig 2.20, and the same formulae apply. Of course, the presence of a capacitor invites blocking distortion again, but since it is effectively in parallel with R1 *and* r_a+Rg, discharging time is greatly reduced. A small grid stopper still ought to be added though, as usual.

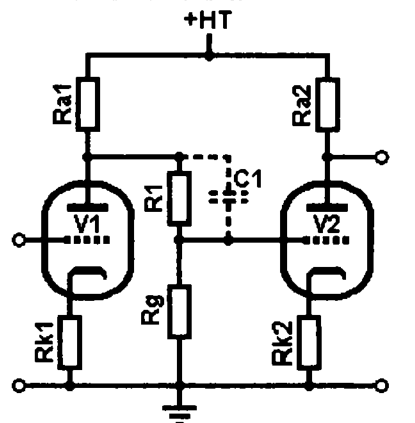


Fig. 2.32: Level shifting using a resistor divider. C1 can be added to avoid attenuating the audio signal, or for treble boost.

Design procedure is much the same as before. Since we now have more freedom over the setting of DC levels we shall consider the case of the ECC83: Again we will assume that V1 has been designed normally, and that its quiescent anode voltage, Va_0, is 210V. We should now decide what level to shift this DC voltage down to. There is no particular value we need aim for; our only provisos are that the divider should not steal too much current from V1, and that Rg will normally be at least five times Ra1, to avoid undue AC loading of V1 (because R1 will be bypassed). These conditions are easily met by setting R1 = 1MΩ, Rg = 470kΩ. Applying the usual formula for a potential divider, XII, this will shift the DC voltage down to Va x β;

$$\beta = \frac{Rg}{R1 + Rg} = \frac{1M}{470k + 1M}$$

$$\beta = 0.32$$

[1] White. E. L. C. (1936). *Improvements in and Relative to Coupling Means for Thermionic Valve Circuits.* British Patent 456450.

So the new DC level at the grid of V2 will be; 210V x 0.32 = 67V.
We could omit CI, in which case we could achieve a flat response down to DC, which can be useful for developing a solid, overdriven bass tone. Alternatively we might make CI large so as to achieve very high-gain levels, or even take advantage of the opportunity for some treble boost.
Applying formula XIV;

$$CI = \frac{1}{2\pi f R1} \cdot \sqrt{\frac{\beta'^2 - \beta^2}{\beta^2 - \beta^2\beta'^2}}$$

To solve for the half-boost frequency, take β' to equal:

$$\beta'_{\text{(half boost)}} = \frac{\beta + 1}{2} = \frac{0.32 + 1}{2}$$

= 0.66

In this case we will choose a half-boost frequency of 30Hz, to develop high gain across the whole guitar range and strong bass overdrive, which could be tailored later in the preamp. Solve XIV:

$$CI = \frac{1}{2\pi \times 30 \times 1M} \cdot \sqrt{\frac{0.66^2 - 0.32^2}{0.32^2 - (0.32^2 \times 0.66^2)}} = 5.3\text{x}10^{-9} \times \sqrt{\frac{0.33}{0.058}}$$

= 13nF
The nearest standard is 10nF and results in a half-boost frequency of 38Hz, which is close enough.
Following the same procedure as we did for fig. 2.29 we first estimate a value for Rk2. If we take a typical anode current of 1mA, this yields 67V / 1mA= 67kΩ, so we

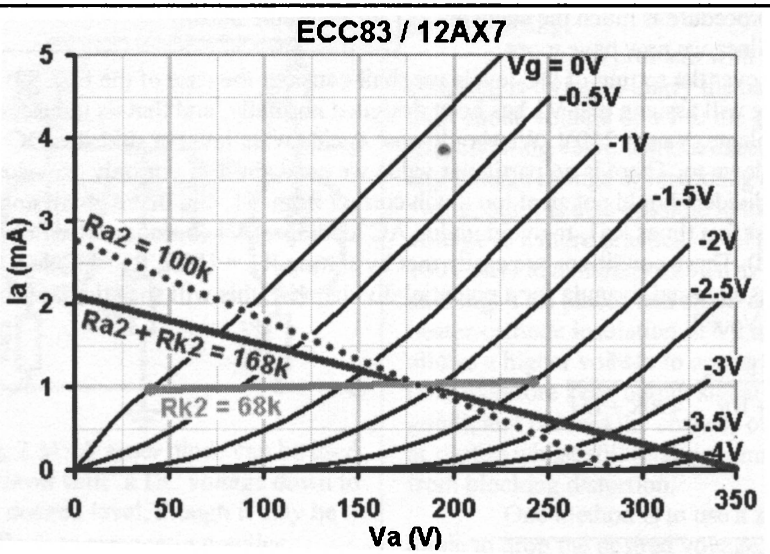


Fig. 2.33: Load lines drawn for V2 in fig 2.34.

would naturally select a 68kΩ resistor. The power dissipated in it will be approximately 67^2 / 68k = 0.06W, so a ½W resistor would be sufficient.

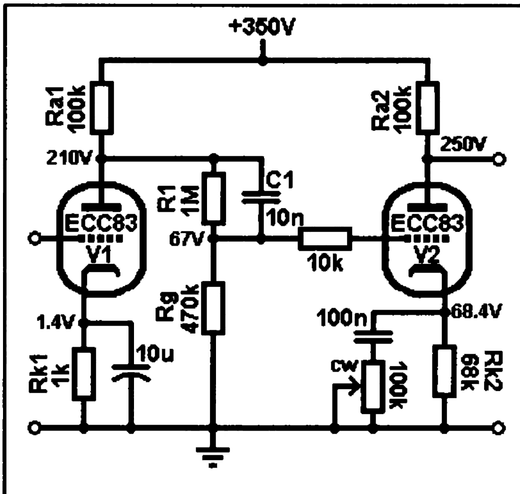


Fig. 2.34: A practical level-shifted circuit offering good distortion on bass notes with freedom from blocking distortion.

Since we are not restricted by a low HT as before, we are free to use a typical 100kΩ anode resistor. The DC, AC and cathode load lines are shown in fig. 2.33, and indicate a bias of about -1.4V.
Again, leaving Rk2 unbypassed would result in very low gain indeed (roughly 0.9!) so a bypass capacitor has been added as before, and a linear variable resistor in series with this could serve as a gain control. The final schematic is shown in fig. 2.34, and a modest 10k grid-stopper has been included to guarantee no blocking, without affecting the treble response. At startup the grid of V2 cannot rise above 90V, which is unlikely to cause arcing during the warm-up period so no protection diode is required.
The frequency response of the circuit, measured at the anode of V2 relative to the anode of V1, is shown in fig. 2.35. Note that the response now extends all the way down to DC. The expected half-boost frequency (indicated by the dot) appears slightly offset due to the rising reactance of Ck2 at low frequencies, though this probably wouldn't worry us.

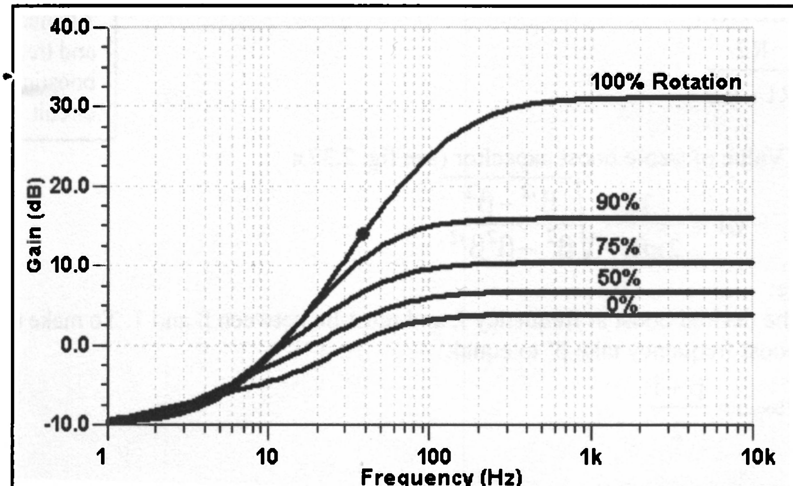


Fig. 2.35: Frequency response of the circuit in fig. 2.34.

Level shifting is almost unheard of in guitar amps, the Hiwatt *STA* series, Roost *Reverb* series and Sound City *LB* amplifiers being notable exceptions. But if it is used sparingly, with attention paid to not being over zealous with bass response, it may prove useful, particularly for bass amplifiers where lower frequency harmonics are otherwise difficult to generate without ugly, capacitor-related distortions.

Summary of formulae:

X; -3dB roll-off due to coupling capacitor (see fig. 2.36):

$$Co = \frac{1}{2\pi f R'}$$

Where $R' = Z_{out} + Rg$

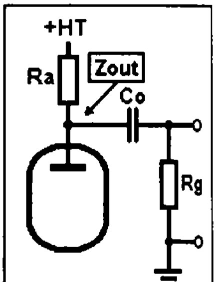


Fig. 2.36: Basic coupling.

XI; Gain of a triode when loaded (see fig. 2.36):

$$A = \frac{\mu(Ra \parallel Rg)}{(Ra \parallel Rg) + r_a}$$

XII; Gain of a potential divider (see fig. 2.37):

$$\beta = \frac{R2}{R1 + R2}$$

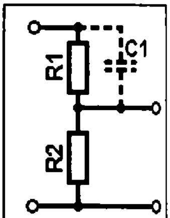


Fig. 2.37: Attenuation and treble-boosting circuit.

XIII; Gain of a treble boosting shelving filter (see fig. 2.37):

$$\beta' = \beta\sqrt{\frac{1 + k^2}{1 + \beta^2 k^2}}$$

Where:

$k = 2\pi f C1 R1$

$$\beta = \frac{R2}{R1 + R2}$$

XIV; Value of treble boost capacitor (see fig. 2.37):

$$C1 = \frac{1}{2\pi f R1} \cdot \sqrt{\frac{\beta'^2 - \beta^2}{\beta^2 - \beta^2 \beta'^2}}$$

Where:

β' = the desired boost at frequency f, and must lie between β and 1. To make this the half boost frequency take β' to equal:

$$\beta'_{(\text{half boost})} = \frac{\beta + 1}{2}$$

XV: Gain of a bass boosting, shelving filter (see fig. 2.38):

$$\beta' = \beta\sqrt{\frac{1+k^2}{\beta^2+k^2}}$$

Where:

$k = 2\pi fC1R2$

$$\beta = \frac{R2}{R1+R2}$$

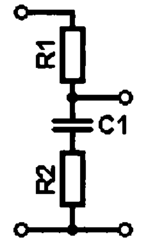


Fig. 2.38: Bass-boosting circuit.

XVI; Value of capacitor for bass boosting (see fig. 2.38):

$$C2 = \frac{1}{2\pi fR1}\cdot\sqrt{\frac{\beta^2(\beta'^2-1)}{\beta^2-\beta'^2}}$$

Where:

β' = the desired boost at frequency f, and must lie between β and 1. To make this the half boost frequency take β' to equal:

$$\beta'_{(\text{half boost})} = \frac{\beta+1}{2}$$

Where in all cases resistances are in ohms and capacitances in farads.

Chapter 3: The Small-Signal Pentode

The suppressor grid (g3). The screen grid (g2). Pentodes connected as triodes. Noise in pentodes. Designing pentode gain stages. Mathematical treatment of the pentode gain stage. Switchable and variable pentode designs.
Summary of formulae.

In chapter 1 it was shown how to design triode gain stages, which are the main building blocks of guitar preamplifiers. Of course, most readers will know that small-signal pentodes (as opposed to the *power pentodes* found in power-output stages) also appear in some amplifiers, and this chapter will discuss how they are used. Interestingly, it seems pentodes were once more popular with manufacturers than they are now. New pentodes are usually more expensive than the popular double-triodes, and usually contain only one valve per envelope, which may account for their lack of use in commercial amplifiers. However, the independent builder is not bound by such constraints. The most popular current-production pentode is the EF86 / 6267, but there are also *many* unfashionable types, available very cheaply as NOS, which can be put to good use by the inventive designer. Useful pentodes include the EF91, EF93 / 6BA6, EF94 / 6AU6, 6AM6, 6BR7, 6SJ7 and 5879. There are also many triode-pentode valves such as the ECF80 (in current production), ECF82 / 6U8A, 6GH8, 7199 and 6BL8.
The design of such stages is slightly more involved than for triodes, but not difficult, provided we understand exactly how pentodes *work*.

The suppressor grid (g3):

In addition to the cathode, control grid and anode found in the triode, the pentode contains two more electrodes: the screen grid and the **suppressor grid**. We shall dispense with a lengthy explanation of its invention save to say that the screen grid is normally held at a high voltage and accelerates electrons toward the anode at greater velocity than could be achieved with the control-grid alone. The electrons hit the anode with such force that other electrons may dislodge from its surface, and these are known as **secondary electrons**. At low anode voltages it is possible for more electrons to be knocked out of the anode than are actually received by it, and without the suppressor grid the secondary electrons would be collected by the nearest positive electrode –the screen grid– meaning total anode current would appear to be flowing backwards! This would result in the valve having negative resistance over a part of its operating characteristics, rendering it prone to oscillation and making it

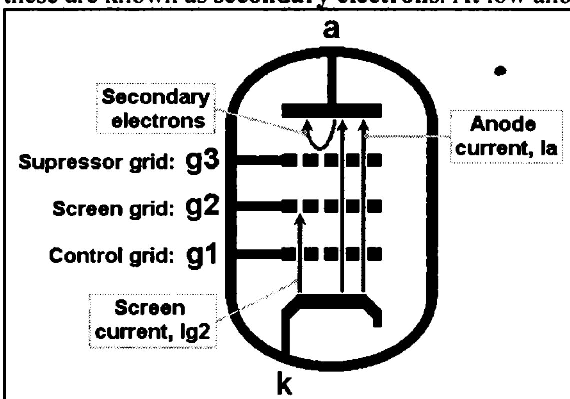


Fig. 3.1: Simplified operation of a pentode. The screen-grid accelerates electrons towards the anode. The suppressor-grid repels secondary electrons back to the anode.

useless for audio amplification. To prevent this, the suppressor grid is interposed between the screen grid and the anode, and is made considerably more negative than the anode. Secondary electrons are thereby prevented from reaching the screen grid since they are repelled by the suppressor grid, back to the anode, as illustrated in fig. 3.1.
The suppressor grid is made more negative than the anode by connecting it directly to the cathode, or to ground. In many pentodes it is already connected to the cathode internally. If there is any impedance in the cathode (such as an unbypassed cathode bias resistor) then connecting the suppressor to ground will dramatically increase the 2nd and 3rd harmonic distortion[1]. Obviously this is very detrimental for hifi designs, but might be useful in a guitar amp.
Apart from this, the suppressor grid is regarded as benign and plays no other part in the design of the stage. If in doubt, connect it directly to the cathode. Some pentodes, including the EF86, have an internal screen around the whole valve to shield against external electromagnetic (EM) interference. This should be connected to ground or to the cathode, whichever is most convenient.

The screen grid (g2):

The **screen grid** is the most important electrode in the pentode, and has the greatest influence over its operation, and we must understand it functioning fully before proceeding. Many designers do not, in fact, understand the screen grid, and this inevitably leads to unreliable, unpredictable or noisy circuits. This may be because there is no solid state equivalent to the pentode, so its operation can seem alien to those with only solid-state experience.

The screen grid gets it name from the fact that it shields or 'screens' the control grid and cathode from the anode. In a triode the only thing which attracts electrons towards the anode is the anode's own EM field. In a pentode the screen grid is able to attract (or impede) electrons itself, depending on the strength of its EM field.
In a triode, capacitances between control grid and anode appear to be multiplied due to the large voltage swings at the anode –the Miller effect– but in a pentode the screen grid is usually held at a constant voltage, or nearly so, so the control grid does not 'see' the voltage swing at the anode, greatly reducing the Miller effect and allowing a pentode to have a very wide bandwidth indeed.
Furthermore, changes in anode voltage will not greatly affect anode current, since the screen-grid's fixed voltage continues to accelerate electrons towards the anode regardless of changes in anode voltage. This makes the valve more efficient, since it effectively magnifies the controlling element of the control grid, allowing the anode to swing much lower than is possible in a triode. The effect is obvious when looking at the static anode-characteristics graph. The grid curves are nearly horizontal (and not dissimilar to a transistor's characteristics), indicating that for a given bias voltage, anode current will remain substantially constant even though the anode voltage may vary wildly. This also indicates that the pentode has a very high anode

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[1] Shepard, W. G., (1953). Suppressor Grid Frequency Doubler. *Electronics*. October, p200.

resistance, r_a, and so the pentode is sometimes referred to as a **constant-current device**.

The grids in a multi-grid valve are arranged concentrically. The control-grid wire is nearest the cathode and is wound with close spacing so that its electromagnetic (EM) field is dense, giving excellent control over the electron flow. The screen grid is wound more coarsely but its voltage is usually high. Most electrons pass between its wires and are further accelerated by its strong EM field, towards the anode. Some electrons do strike the screen grid though, and form the **screen current, Ig2**, though it is much less than the anode current. The suppressor grid is wound even more coarsely and is so negative compared to the anode and screen grid that it collects no electrons. The total current flowing down the cathode is therefore the sum of the anode current and screen current, and this may need to be taken in account when choosing a cathode resistor.

The effect of screen voltage:

The characteristics of a transistor are fixed when the device is manufactured, and the circuit designer has no control over this. Similarly, the characteristics of a triode are assumed to be fixed. In theory we could alter them somewhat by changing the heater voltage, though it is not recommended since it may adversely affect the valve's working life, and is not a very useful or predictable method of control. **The characteristics of pentodes, on the other hand, are controlled by the screen voltage.** By manipulating the screen voltage we can 'morph' the characteristics from those of high current, low input sensitivity and high gain, to those of low current, high input sensitivity and lower gain, and we can even turn the pentode into a triode. This is best understood by looking at the change in the anode characteristics as the screen voltage is varied from a high value to a low value. The oscillograms in fig. 3.2 show the measured characteristics of a real Mullard EF86, probably the most popular small-signal, audio pentode ever produced. For ease of comparison, the anode current and anode voltage scales, and the grid-curve divisions have all been kept the same in each graph.

•

The top-left graph shows the static anode characteristics when the screen voltage is fixed at 130V, which is relatively high for an EF86. The higher the screen voltage the stronger its EM field will be, and therefore the more powerfully it will attract electrons, so the greatest anode current is able to flow in this situation.
The input sensitivity of the pentode is lowest in this situation since the control-grid voltage must be varied by a large amount in order to override the *constant* attraction of the screen voltage, so the pentode would be able to handle the largest input signals before clipping. Note that the grid curves are virtually horizontal, indicating an extremely high r_a and μ, much too high to calculate from the graph.
The grid curves are also the most widely spaced when the screen voltage is high, indicating that the transconductance –and therefore the possible gain– is greatest in this situation. It is also clear that because the grid curves are very nearly parallel above about Va = 50V, the g_m is almost perfectly constant *for a given bias voltage.*

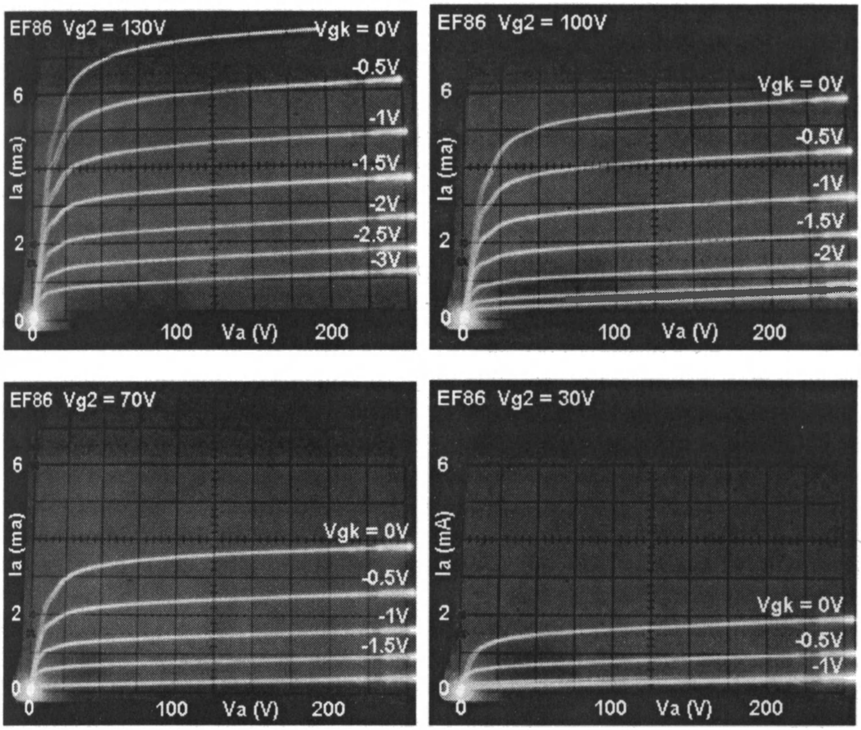

Fig. 3.2: Lowering the screen voltage 'squashes' the grid-curves down, reducing g_m and increasing r_a somewhat.

The value of g_m varies with bias voltage of course, being greater at lesser bias voltages, which make the pentode rather non-linear for larger input signals.
The top-right graph shows how lowering the screen voltage to 100V 'squashes' the grid curves down, since the strength of the screen's EM field has reduced, so for a given bias voltage the anode current is reduced. The transconductance has reduced somewhat, so the possible gain is less, while the input sensitivity has increased. Lowering the screen voltage to 70V compresses the curves down even further, continuing the trend.
When the screen voltage is very low then its EM field is minimal, and it will have a weak influence on electrons. In the bottom-right graph the screen voltage is reduced to 30V, and the transconductance is now very low. The input sensitivity, however, is very high, since small changes in control-grid voltage now have a greater effect on what little anode current can flow. Reducing the screen voltage to zero would almost completely cut-off all anode current.
The relationship between screen voltage and anode current is non-linear, depending mainly on the physical construction of the particular pentode. There is no single law which describes how the screen voltage will affect anode current in any given pentode; the relationship must be measured, and most datasheets will give enough

information that anode-characteristics graphs can be drawn by hand for any screen voltage. The method will be described in detail later.

The effect of screen current:

When designing a pentode gain stage it is usually necessary to know the screen current, since this facilitates the choosing of the cathode bias resistor and screen supply circuitry. The data sheet should give at least one example value of screen current under certain operating conditions, and it may also include a graph of screen current on the static anode characteristics graph.

An important observation of the pentode is that the ratio of anode current to screen current, Ia:Ig2, is quite constant, *for a given anode voltage*. Therefore, if we know the anode and screen current at one particular voltage, we may reliably estimate the screen current for any other operating point, as will be described later.

To see the relationship between currents in the valve, consider fig. 3.3. This shows the 0V grid curve (the others have been removed for clarity) at a screen voltage of 100V. The lower curve shows what the screen current, Ig2, will be at any anode voltage *when the bias is 0V*. Anode current and screen current both flow down the cathode, so the total current flowing in the valve is shown by the top line, and it can be seen that it is almost perfectly constant, which is why pentodes may be regarded as constant-current devices.

We may also note that for anode voltages greater than the screen voltage (100V), the screen and anode currents are both more-or-less constant. However, when the anode voltage drops below the screen voltage the screen-grid becomes the dominant electrode and will begin to collect more electrons, so the screen current increases, 'stealing' current from the anode. Once we approach the knee of the grid curve

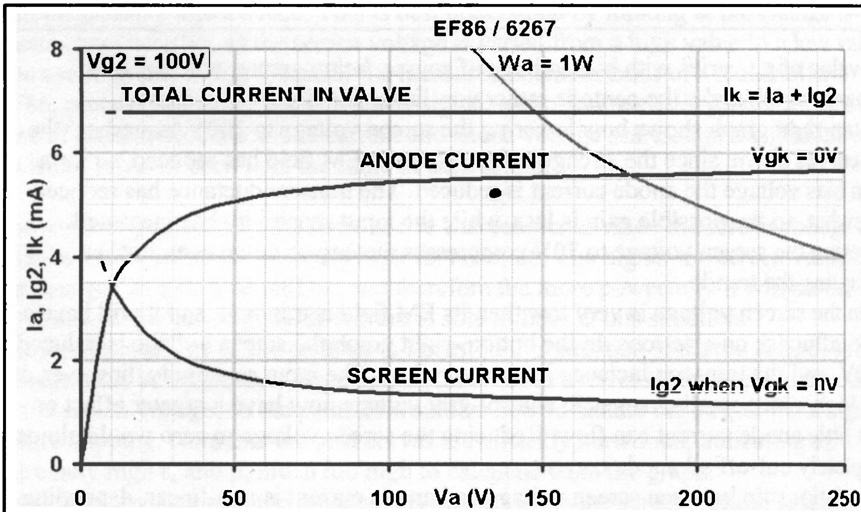

Fig. 3.3: For a given screen and bias voltage, cathode current, Ik, remains substantially constant at any anode voltage, since anode current and screen current are 'mirror images' of one another.

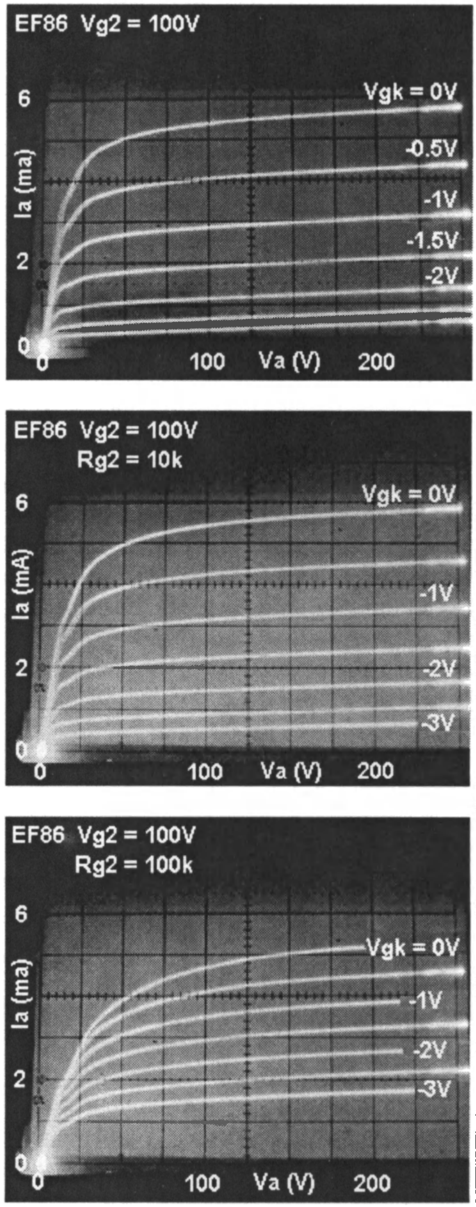

Fig. 3.4: Upper: Static anode characteristics of the EF86 with a screen voltage of 100V. Middle and lower: Dynamic characteristics produced by adding a screen-stopper resistor, Rg2. This increases linearity but reduces g_m and 'softens' the knee of the curves.

around Va = 50V, anode current begins to fall rapidly and screen current rises equally rapidly. (At extremely low anode voltages, screen current falls again, but in an unpredictable manner, but this region is of no interest and is not shown.) The grid curve and the screen-current curve are 'mirror images' of one another.
More curves could be drawn, corresponding to other bias voltages, and some power-valve data sheets will provide these curves. These are useful because if the operating point reaches some point much below the 'knee' of any grid curve, where the grid curve (and therefore anode current) rapidly drops down, the screen current must have risen by the same amount. Since the anode current drops very quickly below the knee, the screen voltage must rise equally rapidly. This rapid increase of screen current can cause serious damage to the screen grid by over dissipation, and great care must be taken to ensure that the screen grid does not exceed its maximum dissipation rating, which should be given on the datasheet (0.2W for the EF86). Of course, this is of greatest concern in power valves where voltage and current levels are higher. In small-signal pentodes the current levels are usually low enough that over dissipation is unlikely in any normal circuit.

Although the increase in screen current at low anode voltages is a cause for concern, it may also be taken advantage of. If a resistance is placed in series with

the screen grid then the variation in screen current with anode current will cause the voltage drop across the resistance to vary in sympathy with the control grid. This has much the same effect as an unbypassed cathode resistor: it introduces negative feedback since the varying screen voltage tends to oppose the change in anode current, which reduces the g_m (and gain) but increases the linearity of the stage. However, if the input signal is large enough that the anode voltage is pulled below the screen voltage then the screen current will increase more rapidly, relative to the anode current, increasing the voltage drop across the series resistance and so dynamically reducing the screen voltage and 'squashing' the grid curves down, particularly near the 'knee'. For example, in fig. 3.4 the top oscillogram is reproduced from earlier and shows the EF86 with the screen voltage *fixed* at 100V, and g_m is quite constant above about Va = 50V.
The middle graph is for the same screen voltage but a screen-stopper, Rg2, of 10kΩ has been placed in series with the screen grid. Firstly, the spacing between curves is more even, indicating the improvement in linearity but a reduction in g_m, due to the screen-current feedback. The 'knee' of the curves is somewhat softer due to the greater proportion of screen current at lower anode voltages.
Increasing the screen-stopper to 100kΩ produces the lower graph*. The spacing between the grid curves is now quite consistent, indicating excellent linearity, but the transconductance, and therefore the possible gain, is greatly reduced. The knee is now very soft, so that g_m begins to fall gradually below about Va = 150V. If the operating point enters this region (during the positive half-cycle of the input signal) then the curves will begin to compress, gently and dynamically compressing the audio signal with them. Because this compression is asymmetric it tends to generate a strong spectrum of even-order harmonics and a warm, 'fat' tone, an effect which cannot be duplicated with an unbypassed cathode resistor. This 'screen compression' effect is unique to pentodes and tetrodes and is highly desirable for guitar (particularly bass guitar), and will be explored further, later in this chapter.

Pentodes connected as triodes:

A pentode may also be connected for triode operation by connecting the screen-grid directly to the anode via a screen-stopper resistor (fig. 3.5). Under this condition the screen-grid and anode effectively become one single electrode, and the valve is said to be **triode strapped**. The suppressor grid is usually connected to the cathode, though it may also be connected to the screen grid, there is no significant difference in performance (connecting the suppressor grid to the control grid tends to result in instability).

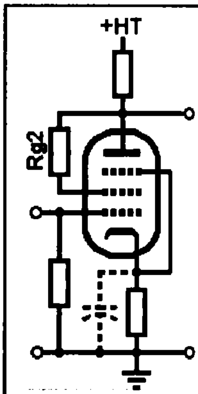

Fig. 3.5: A pentode connected for triode operation.

* It must be appreciated that the middle and lower graphs are *dynamic* characteristics; they only apply to AC signals. For biasing we would still use the upper, static characteristics.

The screen stopper serves a dual purpose of damping any stray impedances which might cause oscillation, and it also ensures that the screen voltage is kept slightly below that of the anode at all times, and so limits the screen current. A value of 100Ω is usual, although the characteristics of the triode may be manipulated by using larger values. A very large screen stopper will place the screen voltage much lower than the anode voltage, so that the screen's EM field has a less attractive force on the electrons, which reduces the g_m of the valve.

For example, the upper oscillogram in fig. 3.6 shows the static anode characteristics of the EF86 when triode strapped using a standard screen stopper of 100Ω (suppressor grid connected to cathode). Looking at the curves: $\mu \approx 35$, $g_m \approx 2$mA/V, $r_a \approx 17.5$kΩ (which is not dissimilar to the old ECC32). Increasing the value of Rg2 to 100kΩ produces the lower characteristics in fig 3.6, yielding the values: $\mu \approx 35$, $g_m \approx 1.2$mA/V, $r_a \approx 29$kΩ . Notice that μ has not changed since this is dependant on the control grid. Of course, there is not a great deal of incentive to use triode-strapped pentodes in a guitar amp, since we may just as easily use ordinary triodes. However, it is a simple matter to provide a switch to toggle between pentode and triode operation, which is just one of many tonal possibilities offered by pentodes, which will be addressed in due course.

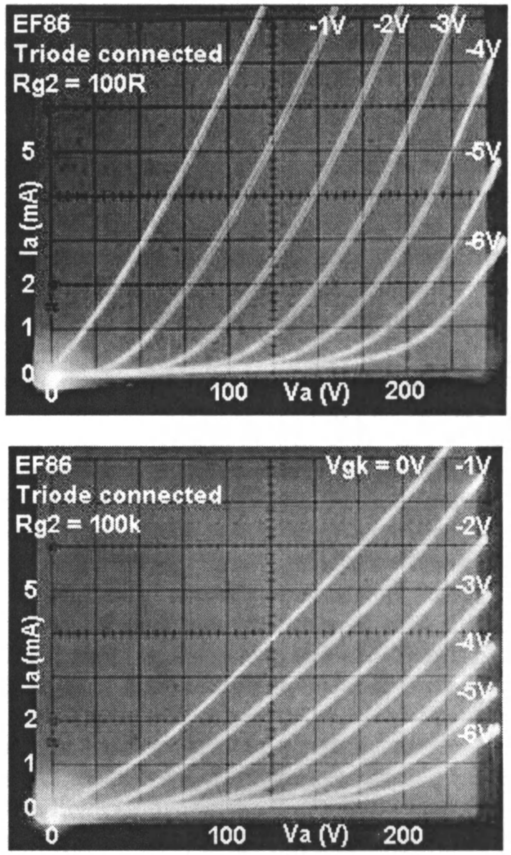

Fig. 3.6: Characteristics of a triode-strapped EF86 with different values of screen stopper.

The triode connected curves are also useful because they tell us what the total cathode current is when Va = Vg2, and can be used as a reference when choosing a cathode resistor.

Noise in pentodes:

Although pentodes can offer a wide bandwidth and high gain, their main shortcoming is noise. All valves suffer from noise induced by the heater (which may

be avoided by reducing the gain of the stage and/or the cathode resistor, or through the use of a *clean* DC heater supply) and shot noise. **Shot noise** is caused by the random variation in arrival time of electrons striking the anode, which constitutes minor fluctuations in anode current. This continual fluctuation in anode current generates a noise voltage across the anode resistor, and the larger the anode resistor (i.e., the higher the gain of the stage) the greater the amplitude of the noise appearing at the output. Shot noise is easily expressed as an **equivalent noise resistance**, that is, a pure resistance which would generate the same amount of noise as the valve. For triodes the equivalent noise resistance, R_{eq}, may be approximated as[2]:
$R_{eq} \approx 2.5/g_m$
This indicates that for low noise preamplifiers, high g_m triodes with healthy values of quiescent anode current are preferable, and good hifi preamplifiers are designed with this in mind. This is the main reason why ECC83s are less favoured for modern hifi preamps than in older designs. For guitar amplifiers this is usually only a minor concern; the large resistances used for grid-stoppers, inter-stage attenuation, tone stacks, gain controls and so on, are much more significant noise sources.

Pentodes, on the other hand, also suffer from **partition noise**. This is caused by the total electron stream flowing in the valve having to split, or part, to go either to the screen grid or to the anode. Although the ratio between anode current and screen current is fairly constant on average, its *instantaneous* value is continually fluctuating, so that the anode current in a pentode will vary by an even greater degree than that caused by shot noise alone. Furthermore, partition noise is inversely proportional to frequency, known as **1/f noise**, so that its amplitude is greater at low frequencies, which tends to be more noticeable to the ear. In general, this will make a pentode from three to five times more noisy than a triode of the same amplification. This makes small-signal pentodes fairly undesirable for serious hifi, and they are not found in modern, high-spec' designs, except when they are connected as triodes. Again, for guitar this source of noise is of minimal concern. Of greater concern is the pentode's very large, internal anode resistance, which results in a very poor power-supply-ripple-rejection factor, or **PSRR**. In a typical pentode gain stage any noise on the power supply will appear at the anode, and so a well filtered HT is always necessary when using pentodes in a preamp.

By far the greatest complaint about pentodes in guitar amps, however, is their susceptibility to **microphonics**. Microphonics are spurious noises caused when a component is physically moved in some way, so that electrode spacing, interelectrode capacitances and EM fields fluctuate. It should be pointed out, of course, that *all electronic components are microphonic*, to some extent. Problems only arise when those components are subjected to strong mechanical vibrations, or when they are used in high-gain circuits, or both. Unlike hifi amps, guitar amps are commonly subjected to heavy, continuous vibrations from nearby loudspeakers, drums and so on, particularly in the case of combo amps where the loudspeaker is

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[2] Langford-Smith, F (1957). *Radio Designer's Handbook* (4th ed.), p783. Illiffe and Sons Ltd., London.

mounted very close to the amplifier chassis. What's more, guitar amps usually have high-gain circuits, and tapping a preamp valve with a pencil will usually produce a sound like a bell in the speaker.
Because pentodes have more grids than triodes they are slightly more prone to microphonics but, which is worse, they are often set up with inappropriately-high gain levels of well over 100, so that what microphonic effects they do possess will be much more greatly amplified than in the case of triodes. This has lead to some guitarists making sweeping statements such as "all EF86s are microphonic". This is nonsense of course, they are only slightly more microphonic than an ECC83, but since it may have a gain four times grater than an ECC83, the microphonics will seem four times worse. The problem is that circuits using the EF86 are often lifted directly from datasheets –high-gain circuits which were intended for use in early hifi preamps– and are not appropriate for modern guitar amps. Most of these reported problems with pentodes can be allayed or even eliminated simply by reducing their gain to more sensible levels, and this will be expounded upon later.

Designing pentode gain stages:

The connections to the anode, control grid and cathode are the same as for a triode, and normal conventions apply. The grid-leak resistor, heater-to-cathode voltage ($Vhk_{(max)}$), quiescent anode voltage ($Va_{(max)}$), peak anode voltage (Va_0) and anode dissipation ($Wa_{(max)}$ or $Pa_{(max)}$), should not exceed the rated maximum values, which will be of a similar order to those of preamp triodes. In addition, the maximum, quiescent screen-grid voltage, $\mathbf{Vg2_{(max)}}$, maximum peak screen voltage, $Vg2_0$, and maximum screen-grid dissipation, $\mathbf{Pg2_{(max)}}$, must not be exceeded either.

Choosing the anode resistor:

When selecting the value of anode resistor, similar considerations apply to pentodes as were discussed for triodes in chapter 1: larger values will result in higher gain but also a higher output impedance. The tonal variation with a pentode, however, is greater than for most triodes. Consider fig. 3.7 which shows the static anode characteristics of the EF86 at a screen voltage of 130V. Three load lines are shown for different possible values of anode load resistor with an HT of 250V.

The 100kΩ load passes *below* the knee. It is clear that with this load the possible output signal swing is greatest, being almost equal to the HT! The gain is high, at around 110 near the centre of the load line. The grid curves are more 'bunched together' near cut off, as was the case with a triode, but in this case they are also bunched together near the top of the load line. In fact, above Vgk = -1V the anode voltage cannot swing down much further before grid-current limiting occurs. Therefore, if the bias point were roughly central, around Vgk = -2.5V say, large input signals would be compressed near the bottom of the load line in the usual fashion, but also rapidly and prematurely clipped near the top. Therefore this mode of operation is generally avoided in traditional circuits, since the risk of distortion due to premature clipping is great, unless the input signal is guaranteed never to become too large. For guitar, of course, this marked difference between the

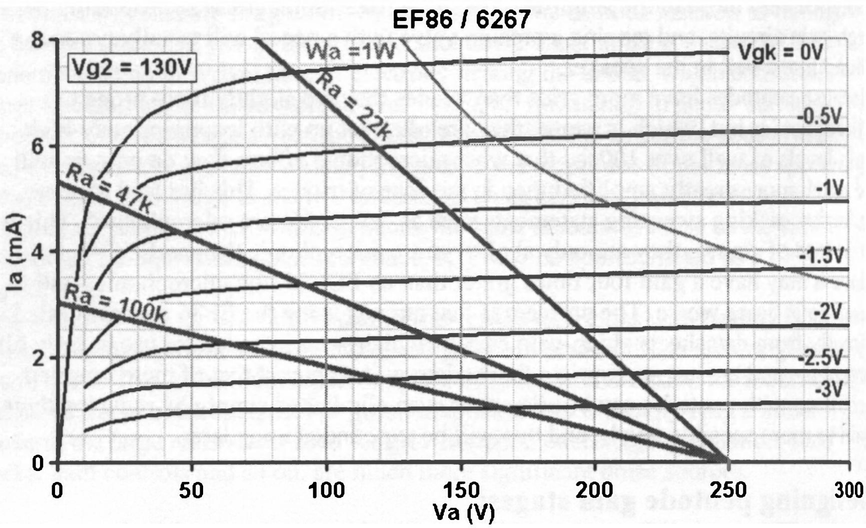

Fig. 3.7: Static anode characteristics of the EF86 with three possible values of load. Normally we would choose the value passing through the 'knee'.

compressive clipping of cut-off, and the sharp clipping due to grid current, can be very useful, developing a complex array of odd and even harmonics.

The 47kΩ load line passes *through* the knee. The gain is slightly lower, at around 100 for small signals.
The grid curves near the top of the load line are not bunched quite so closely together, so the onset of clipping will be more compressive, less sudden, and more like those of cut-off clipping. This roughly symmetrical compression gives rise to a spectrum of mainly odd-order, diminishing harmonics (3^{rd}, 5^{th}, 7^{th} etc.), which is classic of pentode non-linearity. These tend to add 'bite' and 'roughness' to the tone, and are a pleasant complement to a triode's 'warmth'. What's more, this symmetrical compression only happens on larger input signals; low-level signals will tend to be compressed only on one side –as with a triode– so the harmonic spectrum of a pentode with a large anode load resistance tends to vary dynamically with signal level, and is highly dependent on the bias voltage. If screen compression is also used then the tonal texture can become very complex indeed, and highly touch sensitive. This is the traditional mode of operation for a pentode since it offers the highest gain before the risk of gross distortion due to premature clipping, and is also the condition for maximum output *power*; when the load line passes through the knee, optimum anode current and voltage swing are obtained.

The 22kΩ load passes *above* the knee. The most significant change is that the grid curves no longer appear bunched near the top of the load line: the spacing is more like that of a triode. This will result in a triode-like tone, richer in 2^{nd} and 4^{th} harmonics. Furthermore, the operating point never approaches the knee of the Vgk = 0V grid curve, so there will be no rapid increase in screen current on signal peaks.

The gain is lower, at around 50, but this is still about as high as a typical ECC83 gain stage. When pentodes are used in modern hifi this is the preferred operating condition, since it still offers high gain (compared with most triodes) but less high-order harmonic distortion, and less risk of non-linearity due to screen compression. For guitar use it is a less logical operating condition, since the usual reason for using a pentode in the first place is its unique tone and compressive properties. Therefore we would probably select a 47kΩ load, or greater.

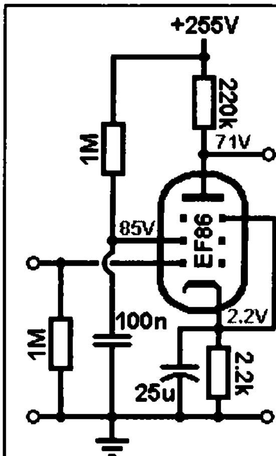

Fig. 3.8: The early Vox *AC15* input circuit is lifted directly from high-gain, data sheet circuits. It has a gain of around 200 and is notorious for microphonics.

Some readers may be familiar with circuits using the EF86 with a very high value of anode load (greater than 100kΩ say) and a low screen voltage, resulting in extremely-high gain and very-low input sensitivity. Such circuits were common in audio products like tape recorders and microphone preamps of the 1950s and 1960s, where high gain from a single, low-current stage was a new luxury. The same arrangement was famously used as the input stage of the first Vox *AC15* combo (fig. 3.8). Unfortunately, as mentioned previously, the very high gain is a hindrance in a guitar amp since the problems of microphonics become very great if the volume of the amp or its surroundings is loud. This was a familiar problem of the AC15 and some effort was required to specially select fortuitously low-microphonic EF86s, and it was removed from later models. However, the modern reverence for vintage amps such as the AC15 has prompted some manufacturers to begin using small-signal pentodes again. Unfortunately these commercial designs are usually close copies of the original, and have the same problems with microphonics. From a tonal perspective it also makes little sense to use a pentode at the input of a guitar amp since it is unlikely to ever be overdriven by the guitar's low output level, so that the full character of the pentode is unlikely to be heard; a transistor could perform the same task of providing high gain but without the problems of microphonics. Therefore, the author strongly recommends that the *conscientious* designer does not use a total anode load (including AC loading) greater than about 100kΩ with a small signal pentode; extremely high gain is not essential for 'the pentode tone'. Much improved results, and more versatile tones are possible if the pentode is operated at sensible gain but is preceded by at least one other gain stage, so that it may be overdriven itself, exposing its full character. The Matchless *Clubman* is one of the few commercial amplifiers to adopt such a topology.

Choosing the screen voltage:

Although we now have an appreciation for the effects of the anode load we need to consider the effect of the screen voltage. The load lines drawn in fig. 3.8

were at a screen voltage of 130V. This is actually fairly high for a small signal pentode; most circuits use screen voltages in the range of 50V to 100V since this results in slightly improved linearity (in hifi terms) and less anode current, meaning a power saving. Operating at lower screen voltage also implies using larger resistances in the anode, cathode and screen circuits, which allows smaller capacitors to be used for coupling and decoupling, which was an important factor when large, high-voltage capacitors were expensive. Nowadays these things are not particularly advantageous, as high-current power supplies and large, high-quality capacitors are quite inexpensive, and we are not interested in good linearity.
To avoid exacerbating the problems of microphonics, the best approach is to mutually select a screen voltage and anode load which together result in a sensible gain, not exceeding 100 say, with the load-line passing through the desired part of the anode characteristics.
It should be remembered from the last chapter that even an ECC83 usually offers more gain than is really required for most preamps, so that interstage attenuation must be employed. A pentode practically guarantees high gain whatever we do, so our primary aim should be with generating the desired *tone*, and not some arbitrary gain figure.

To fully appreciate the interaction between anode current, load, and screen voltage, it is worth considering a couple of practical situations. In general we would like the load line to pass through, or slightly below the knee (remembering from chapter 2 that any AC load we also apply will cause the load line to rotate clockwise somewhat). Allowing it to pass much above the knee will tend towards triode operation, which has less appeal for guitar, although is still of use for clean tones and for bass guitar.
First, from Fig. 3.7 it was clear that the 47kΩ load line passed through the knee, so

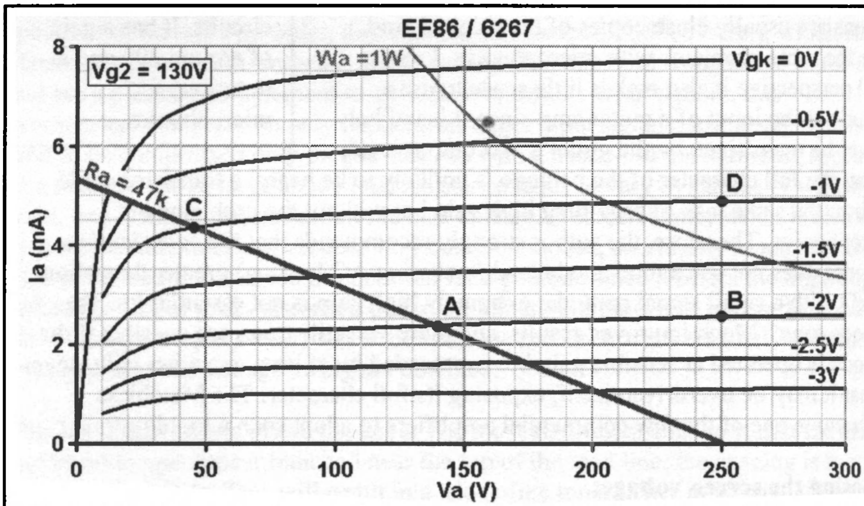

Fig. 3.9: Static anode characteristics reproduced from fig. 3.7, in which a 47kΩ load has been selected.

the same load line is reproduced in fig. 3.9. Alternatively we could have chosen the 100kΩ load but lowered the screen voltage to around 60V to 70V so that the 'knee' moves down to meet the load line (as in fig. 3.13), and it is this interplay of load line and screen voltage that is the essence of good pentode-circuit design.

We are most concerned with ensuring the screen grid is not at risk of over dissipation, and this varies somewhat depending on the bias point chosen.
Since $Pg2_{(max)} = 0.2W$, the maximum quiescent screen current we could allow in this case is:
$I = P/V$
$I = 0.2 / 130$
$= 1.5mA$.
It was mentioned earlier that the ratio of anode current to screen current, Ia:Ig2, is relatively constant for a given anode voltage. The EF86 data sheet gives a set of 'typical characteristics'* which are reproduced in fig. 3.10, and show that at when Ia = 3mA , Ig2 = 0.6mA, at an anode voltage of 250V. This is a ratio of 3 : 0.6, or in other words, when the anode voltage is 250V we can expect the screen current to be approximately: m = 3 / 0.6 = 5 times less than the anode current, even though we are not using the same screen or bias voltage as in the data sheet example. We can use this knowledge to extrapolate the screen current for any given operating point.
For example, we might select a bias of around -2V (roughly central) resulting in a quiescent anode voltage, of 140V and a quiescent anode current of about 2.4mA (point A).

TYPICAL CHARACTERISTICS

Anode voltage	V_a	250	V
Grid No.3 voltage	V_{g_3}	0	V
Grid No.2 voltage	V_{g_2}	140	V
• Grid No.1 voltage	V_{g_1}	-2.2	V
Anode current	I_a	3.0	mA
Grid No.2 current	I_{g_2}	0.6	mA
Transconductance	S	2.2	mA/V
Amplification factor	$\mu_{g_2g_1}$	38	-
Internal resistance	R_i	2.5	MΩ

Fig. 3.10: Example operating conditions for the EF86 / 6267, taken from the Mullard data sheet.

* Note that these are *not* recommended operating conditions, they are simply test conditions.

To find the screen current we first calculate what the screen current would be at the *same bias voltage*, but at Va = 250V, since this corresponds to the data sheet information.
At Va = 250V and Vgk = -2V (point B), the anode current would be 2.5mA, so applying the Ia:Ig2 ratio we expect the screen current to be 2.5 / 5 = 0.5mA.
At our chosen bias point the anode current is actually 2.4mA, which is 0.1mA less. Therefore, the screen current must increase by the same amount, giving:
0.5 + 0.1 = 0.6mA.
And the quiescent screen dissipation will be: P = IV = 0.6 x 130 = 78mW, which is well within limits.

An important observation to make of this is that because of the near-horizontal grid curves, the difference between the anode current at Va = 250Vand at Va = 140V is very small: only 0.1mA. For this reason it is usually safe to assume that the Ia:Ig2 ratio will work at anode voltages other than the data sheet value, *providing it is greater than the screen voltage*, with only a small inaccuracy. In this case we could have approximated the screen current to be 2.4 / 5 = 0.48mA, which is only 20% lower than the true figure. This is sufficiently accurate for most purposes since the tolerance of the actual valve used will usually be worse than this. And the higher the anode voltage is, relative to the screen voltage, the more this error shrinks.

Let us now consider a different bias point, say -1V, which is very warm (point C). At this point Va = 45V, and Ia = 4.4mA. If we tried to apply the Ia:Ig2 ratio directly we would predict a screen current of 4.4 / 5 = 0.88mA. However, at this bias point the anode voltage is much lower than the screen voltage, so we must use the correct approach. Again, we first find the screen current at the same bias voltage when Va = 250V (point D). At this point Ia = 4.9mA, so we expect the screen current to be 4.9 / 5 = 0.98mA. At our actual bias point the anode current is 4.9 – 4.4 = 0.5mA less, so the screen current must be 0.5mA more, giving 0.98 + 0.5 = 1.48mA . This is 60% higher than the approximation suggested. This is extremely important, because we now find the quiescent screen dissipation to be 1.48 x 130 = 192mW, rather than 114mW if the approximate figure had been used. This is far too close to the published dissipation limit of 0.2W, so we find that this bias point is not acceptable, particularly given the poor tolerance of real valves. Even if the valve did yield the expected currents, when it is operated the average screen current will increase slightly so screen failure would be extremely likely, particularly if the valve were being overdriven.

Designing a stage in which the screen voltage is higher than the anode voltage is somewhat less predictable, and some experimentation is usually required to obtain the desired results safely (it is worth noting that the Vox input circuit in fig. 3.8 does yield an anode voltage which is slightly lower than the screen voltage, due its warm bias). Consequently, nearly all pentode designs operate the screen at a lower voltage than the anode, and all further discussion will pertain to this mode of operation Lowering the screen voltage not only reduces screen dissipation but also gives us a greater range of bias points over which the anode voltage is greater than the screen voltage, so design of the stage becomes more predictable.

Adapting the anode characteristics for any screen voltage:

From previous discussion it was shown that the anode characteristics of a pentode change with screen voltage, 'squashing down' at lower screen voltages. This allows us to force the characteristics to suit a particular load resistance, or we can begin with a known screen voltage and choose an appropriate load. The data sheet, however, will not show anode characteristics at every screen voltage, usually it will give only one or two graphs at example screen voltages. We could resign ourselves to using the example curves at a particular screen voltage, and design a stage around this, but suppose we wish to use some other screen voltage, how do we derive a new set of anode characteristics?

The data sheet should provide the mutual characteristics or transfer characteristics graph for the pentode*, which shows anode current against bias voltage, for different screen voltages at a fixed anode voltage. Fig. 3.11 shows those of the EF86 at Va = 250V.

Suppose we wish to draw curves for a screen voltage of 80V, we can expect the 80V screen curve to pass half-way between the 60V and 100V screen curves. The curves are redrawn in fig. 3.12*a* for clarity, and the 80V curve has is shown by the dashed line.

This curve now shows us what the anode current will be, for a given bias voltage, at Va = 250V.

For example, when Vgk = 0V the anode current is shown to be 4mA (point A in fig. 3.12). At -1V it is 2mA (point B) and at -2V it is 0.6mA (point C). These points can now be plotted at Va = 250V on the usual axes of anode voltage against anode current, as has been done in fig. 3.12*b*. These give the 'starting points' for three grid curves at Vgk = 0V, -1V and -2V.

At this point we could assume the pentode is perfect and infer ideal, horizontal grid curves (shown dotted), which is quick and easy if we are simply estimating input sensitivity, or choosing a suitable

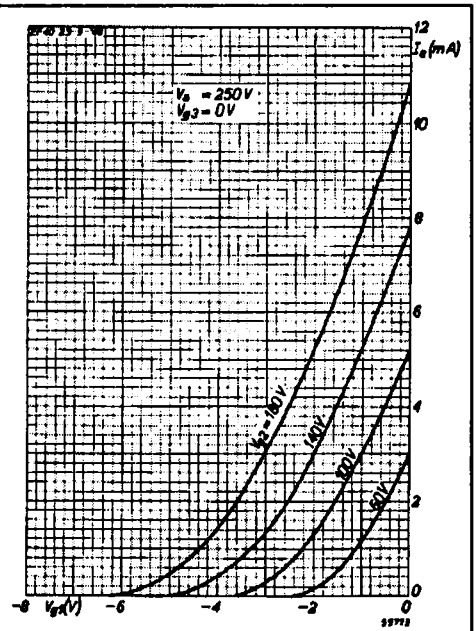

Fig. 3.11: Transfer characteristics of the EF86 / 6267, taken from the Philips data sheet.

* Some data sheets give very little graphical information, in which case it is better to find a more comprehensive data sheet from another manufacturer.

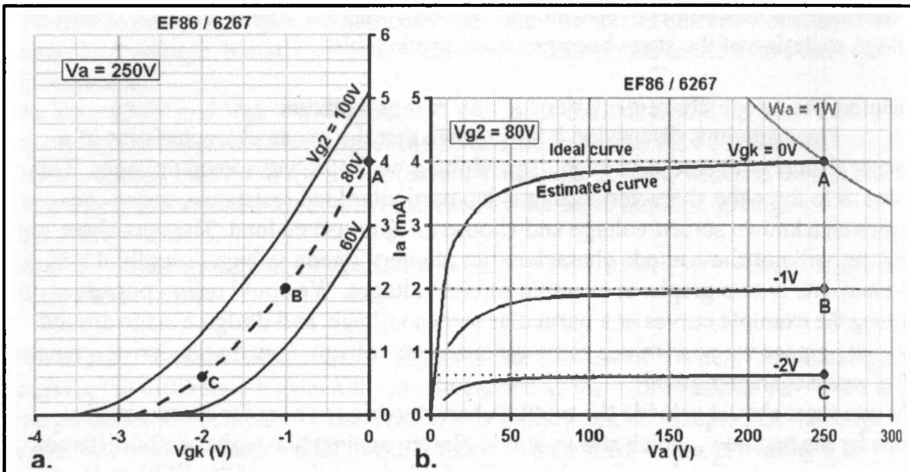

Fig. 3.12: **a.** transfer characteristics redrawn from fig. 3.11, with the Vg2 = 80V curve inferred (dashed). **b.** estimated anode characteristics Vg2 = 80V, derived from the transfer characteristics.

load resistance since it shows where the 'knee' effectively occurs (at Va = 0V, Ia = 4mA in this case). Or we can use the characteristics which *are* provided in the data sheet to estimate more realistic curves, from which we can estimate the screen current, and so on. Some data sheets may even give two sets of transfer characteristics corresponding to different anode voltages, allowing us to plot other points through which the grid curves can be drawn. The maximum anode dissipation curve has also been drawn, using the method described in chapter 1.

Choosing the cathode bias resistor:

It might be thought that choosing the cathode bias resistor would be somewhat complicated by the presence of screen current, but in fact it is quite straightforward if we recall from fig. 3.3 that cathode current is substantially constant for a given screen and bias voltage.

For example, consider the load line in fig. 3.13 which operates the EF86 at a screen voltage of 60V with a 100kΩ anode resistor, which is about as high in gain and input sensitivity as we would dare use in a guitar amp.
We might choose a bias of -1V (point A) giving a quiescent anode current of about 1.2mA. This bias point is roughly central, although grid-current limiting will occur slightly before cut-off so we would expect a slight dominance of even-order harmonics when overdriven.
To calculate the cathode current we simply make use of the Ia:Ig2 ratio as we did when calculating screen current earlier. From fig. 3.10 we determined the Ia:Ig2 ratio to be 3 : 0.6 (m = 5) at Va = 250V.
Point B shows that at our desired bias of -1V, *at an anode voltage of 250V*, give an anode current of about 1.25mA. To find the total cathode current we must add the screen current which is 1.25 / 5 = 0.25mA, making a total cathode current of:

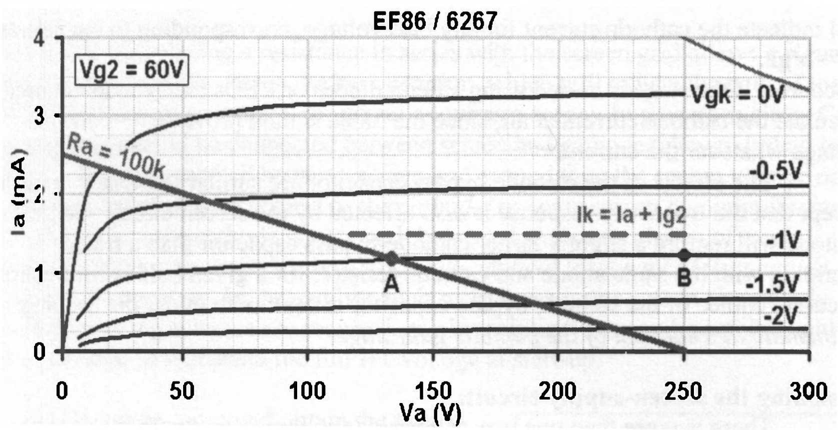

Fig. 3.13: At a screen voltage of 60V, and a chosen bias voltage of 1V, cathode current may assumed to be constant at 1.5mA.

Ik = Ia + Ig2
Ik = 1.25 + 0.25
= 1.5mA
This is shown by the dashed line to remind the reader that it does not change with anode voltage.
Applying Ohm's law we find the cathode resistor to be:
1 / 0.0015 = 667Ω
So we would probably choose a standard value of 680Ω, and this is shown in fig. 3.14.

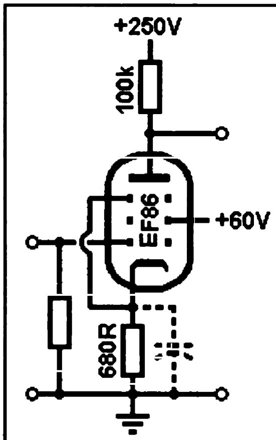

Fig. 3.14: Component values derived from fig. 3.13.

If we had ignored the screen current completely and used only the anode current to calculate the cathode resistor, in exactly the same way as for a triode we would have obtained a value of 1 / 0.00125 = 800Ω. We may have then used a standard value of 820Ω and found the bias to be rather more than predicted, when the circuit was built. Of course, often the difference will be too small to be of concern, especially since we will probably be adjusting the bias on test anyway when tweaking the tone of the amp, but in a power pentode the error could be more important.
Even if we are using anode characteristics which we have estimated from the transfer characteristics, our value of cathode resistor will still be accurate since the 'typical characteristics' (fig. 3.10) normally correspond to the same anode voltage as the transfer characteristics (fig. 3.11)! And because the cathode current remains constant for a given bias voltage and screen voltage, it does not even matter if our quiescent anode voltage is less than the screen voltage; the above method will still yield the necessary value of cathode resistor to the same degree of accuracy.
In the rare case where the quiescent anode voltage is expected to be equal to the screen voltage then we may simply refer to the triode-strapped characteristics, which

will indicate the cathode current for any bias voltage, corresponding to our selected Va = Vg2.
Of course, if the stage is biased using silicon diodes or LEDs then we do not need to calculate the cathode current at all, since the diode should provide the same bias voltage whatever the current.

The effects of the cathode bypass capacitor are similar to those in a triode, except that the frequency response is *also* affected by the screen circuit. Generally, a pentode will require a slightly larger cathode-bypass capacitor than a triode operating with the same anode and cathode resistor, for a given half-boost frequency. Accurate choice of the cathode bypass capacitor is dealt with under the heading: *mathematical treatment of the pentode gain stage.*

Designing the screen-supply circuit:

There is more than one way to provide the screen grid with the necessary voltage, though the commonest method is via a single **screen dropping resistance, Rg2**. Because the screen draws a (relatively) constant current, a resistance can be connected between the HT and the screen grid so that the resulting voltage drop provides the desired screen voltage.
To continue with the circuit in fig. 3.14, we require a screen voltage of 60V. We already know from the load line in fig. 3.13 that the quiescent anode current will be ~1.2mA. By using the Ia:Ig2 ratio we also know that the total cathode current should be 1.5mA. Therefore at our selected bias point the screen current should be:
Ig2 = Ik – Ia
Ig2 = 1.5 – 1.2
= 0.3mA

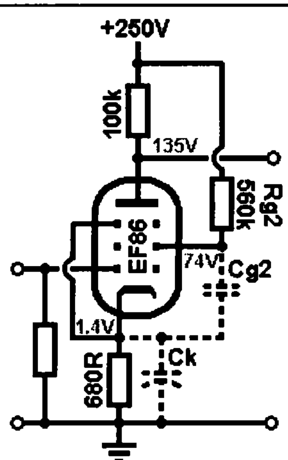

Fig. 3.15: Adding a screen dropping resistor to obtain the desired screen voltage.

Since the HT is 250V we must drop 250 – 60 = 190V across the screen dropping resistor. Applying Ohm's law produces the value:
R = V / I
Rg2 = 190 / 0.0003
= 633kΩ.
The nearest standard is 680kΩ. Alternatively, in a guitar amp we might decide to take the nearest *lower* value, which is 560kΩ, which should result in a marginally higher screen voltage, thereby putting the load line slightly below the knee which is generally more interesting, tonally, than if the load line passes above the knee. The circuit was built and tested with a Mullard EF86, and measured voltages are shown in fig. 3.15, from which it can be seen that using a 560kΩ screen dropping resistor has raised the screen voltage to 74V, which is a reasonable value. This method of providing the screen voltage is also largely self-compensating; the operating point does not vary much as the pentode ages.

Like the cathode, placing a resistance in series with the screen grid allows a signal voltage to appear at the screen during operation, which will reduce the gain of the stage. For maximum gain then, the screen requires a **screen bypass capacitor, Cg2**. Normally this should be connected between screen and cathode (shown dashed in fig. 3.15). If it is connected between screen and ground then the screen voltage no longer stays constant with respect to the *cathode*, so some screen compression may still occur, unless the cathode is very well bypassed. Accurate choice of this capacitor is dealt with under the heading: *mathematical treatment of the pentode gain stage.* Fortunately, because of the high resistances in the screen circuit only small values of capacitance are required, so electrolytic types can be safely avoided. It must be rated to withstand the full HT voltage at start-up.

Of course, we could obtain the screen voltage from other sources, even a separate power supply if we so wished! However, the conventional dropping resistor described above is the simplest, and provides the greatest screen compression effect if it is only partially bypassed. Of course, if we wanted a cleaner, more 'hifi' sound then we might choose to provide the screen grid with a better regulated, less variant or 'stiff' supply voltage. This tends to produce a more 'open' sound when not overdriven, and a 'harder' distortion sound with a faster transient response when the stage is overdriven, though with less touch sensitivity. A traditional way to improve the 'stiffness' of the screen supply is by the use of a potential divider as shown in fig. 3.16, rather than a single dropping resistor,

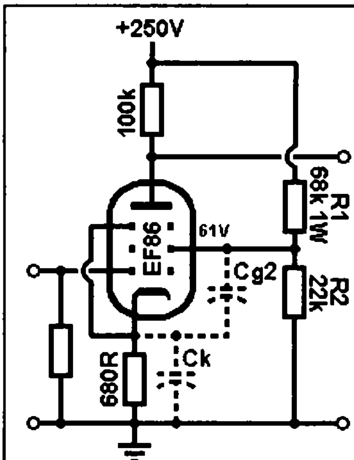

Fig. 3.16: Supply the screen grid from a potential divider.

R2 acts as a primitive shunt regulator, since any variation in screen current now only constitutes a small part of the current flowing in the dropping resistance R1. If we allowed at least ten times more current to flow down R2 than is taken by the screen grid then we may safely neglect screen current when calculating resistances. In this case, Ig2 x 10 = 3mA. Applying Ohm's law a suitable value for R2 is therefore:

R2 = Vg2 / I = 60 / 0.003

= 20kΩ.

The nearest standard is 22kΩ and the power dissipated will be only 60^2 / 22k = 0.16W. Ignoring screen current, R1 must be:

$$R1 = \frac{R2.HT - R2.Vg2}{Vg2} = \frac{(22k \times 250) - (22k \times 60)}{60}$$

= 70kΩ.

The nearest standard is 68kΩ, and the power dissipated will be 190^2 / 68k = 0.5W, so a 1W or better resistor is required.

The total source impedance presented to the screen grid is now only R1||R2 = 17kΩ, rather than 680kΩ as in fig. 3.15, so even without a screen bypass capacitor this will provide a reasonably constant screen voltage, though mild screen compression is still possible if left unbypassed. A small advantage of this method is that the screen bypass capacitor need only be rated for the maximum voltage appearing at the junction of the two resistors, rather than for the full HT.

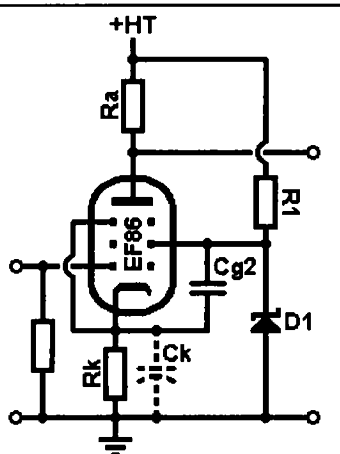

Fig. 3.17: A zener diode will provide a very constant screen voltage.

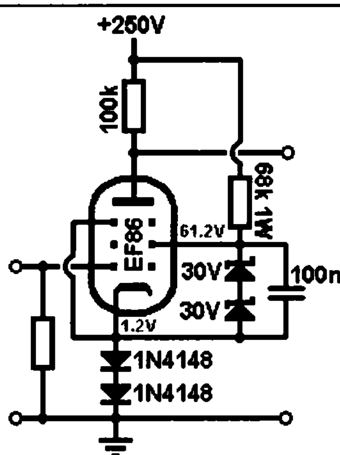

Fig. 3.18: A highly regulated pentode amplifier.

Taking this to the logical extreme, if we desired absolutely constant gain at all frequencies, and the most transparent sound, we could properly regulate the screen supply using zener diodes as in fig. 3.17. R1 must be small enough to provide enough current to the zener for proper regulation. Again, if we allow 10x the screen current, 3mA, then we find a value for R1 of:

$$Rg2 = \frac{HT - Vg2}{I} = \frac{250 - 60}{0.003}$$

=63kΩ

So 68kΩ 1W or 56kΩ 1W would do.

Since 60V is not a common zener voltage we could stack several diodes to obtain the correct value. Two 30V zeners would do, and since each passes about 3mA they will each dissipate 0.003 x 30 = 0.09W, so small, 500mW devices can be used. However, because zener diodes produce spurious noise a bypass capacitor is still required, and a value of 100nF is typical.

We might even go so far as to use LED or diode cathode biasing, for a truly fixed-voltage design. This is shown in fig. 3.18, in which the zener diodes have been returned to the cathode rather than to ground, so that their current also flows in the cathode biasing diodes. This ensures that even if the pentode is driven to cut-off, the biasing diodes never switch off, thereby avoiding any switching noise. The same principle could be applied to the potential divider method described above, of course. If a cathode *resistor* is used instead, then returning the divider to the cathode will cause the cathode voltage to rise, unless Rk is suitably reduced, but the bias voltage will be made more constant this way.

The obvious drawback of the potential divider and zener regulator methods is that it consumes more current from the HT, so that good HT filtering will be required.

Consequently, these methods will rarely be found outside of hifi designs, but have been described here for the sake of completeness.

Mathematical treatment of the pentode gain stage:

As may be expected, treatment of the pentode can be made more complex that that of the triode due to the presence of the screen grid.

Gain:

Because the internal anode resistance of the pentode is extremely high, making the grid curves virtually horizontal, if the cathode and screen are fully bypassed then the gain of pentode is simply:

$A = g_m Ra$ XVII

For example, referring to the load line in fig. 3.13 we determine the g_m at the bias point to be around 1.2mA/V. This gives the gain of the corresponding circuit in fig. 3.29 to be:

$A = 0.0012 \times 100k$

$= 120$

which is very high indeed, although it will be reduced somewhat if the stage is loaded by a following impedance.

Of course, it was noted earlier that g_m tends to vary a lot depending on bias, so these formulae are really only accurate for small signal levels. This circuit was tested with a Mullard EF86 and found to have a gain of 120 with a 50mVp-p input signal, falling to 100 with a 400mVp-p input, illustrating the strong non-linearity of the pentode at higher signal levels.

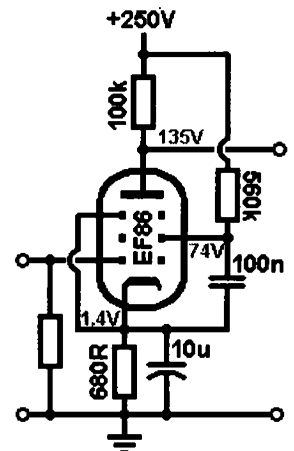

Fig. 3.19: A stage having fully bypassed screen and cathode, giving a gain of around 120 (42dB).

If either screen or cathode is *not* bypassed then this gain is reduced to[3]:

$A' = g_m Ra\beta\gamma$ XVIII

Where:

β = loss of gain due to an impedance at the screen grid.

γ = loss in gain due to an impedance at the cathode.

The effect of screen bypassing:

In order to find the attenuation due to the screen circuit, β, we must first calculate the **internal screen resitance, r_{g2}**. This is analogous to anode resistance, r_a, as it is the effective resistance seen 'looking into' the screen grid. Unfortunately its value is not normally given on data sheets. However, it may be closely estimated

[3] Terman, E. T., Hewlett, W. R., Palmer, C. W., Pan, W. (1940). Calculation and Design of Resistance Coupled Amplifiers using Pentode Tubes. *Trans. A.I.E.E.*, **59**. pp.879–884.

from the triode-strapped characteristics at the same anode current as the pentode circuit, according to[3]:

$r_{g2} = r_{a(triode)}(1+m)$ XIX

Where:

$r_{a(triode)}$ = the anode resistance of the same valve when triode strapped, at the corresponding anode current.

m = the Ia:Ig2 ratio.

For example, in fig. 3.6 (upper) the anode resistance of the EF86 when triode strapped, at 1.2mA, is approximately 25kΩ. The Ia:Ig2 ratio, m, was found from fig. 3.10 to be equal to 5. Therefore the internal screen resistance is approximately:

$r_{g2} = 25k \times (1 + 5)$

$= 150k\Omega$.

The effect of an unbypassed screen dropping resistance in reducing the possible gain of the stage is:

$$\beta = 1 - \frac{Rg2}{Rg2 + r_{g2}} \quad \text{......XX}$$

For example, the circuit in fig. 3.20 is the same as that in fig. 3.19 except that the screen is no longer bypassed. This reduces the gain by a factor of:

$$\beta = 1 - \frac{560k}{560k + 150k}$$

$= 0.21$

So the gain becomes:

$A = g_m Ra\beta = 0.0012 \times 100k \times 0.21$

$= 25$

which is a considerable reduction. The improvement in linearity due to this screen-current feedback will be large, though not of great use in a guitar amp.

An interesting observation is that if Rg2 = r_{g2}, the gain is reduced by exactly half, or -6dB*, when the screen is unbypassed.

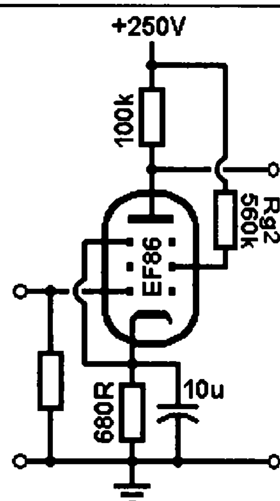

Fig. 3.20: A stage having an unbypassed screen, reducing the gain to around 25 (28dB).

If the screen is only partially bypassed then we obtain a treble boosting effect, as with partial cathode bypassing. Provided Rg is greater than r_{g2}, which is nearly always true, it is convenient to define the frequency at which the gain is reduced by -3dB (rather than

* This assumes that anode current is independent of anode voltage, which is essentially true if the load line does not pass above the knee.

the half-boost frequency used for cathode bypassing). To a very close approximation this is given by:

$$f = \frac{1}{2\pi Cg2Rg2'} \quad \text{.....................................XXI}$$

Where:
Cg2 = the screen bypass capacitor.
Rg2' = $Rg2||r_{g2} = Rg2.r_{g2} / (Rg2 + r_{g2})$

For example, suppose we wish to add treble boost to the circuit in fig. 3.20. First we find the parallel combination of Rg2 and r_{g2} to be:

$$Rg2' = Rg2 \,||\, r_{g2} = \frac{Rg2.r_{g2}}{Rg2 + r_{g2}} = \frac{560k \times 150k}{560k + 150k}$$

= 118kΩ

For a really bright sound we might select the -3dB frequency to be 400Hz or so. Rearrange and solve XXI for Cg2:

$$Cg2 = \frac{1}{2\pi f Rg2'} = \frac{1}{2\pi \times 400 \times 118k}$$

= 3.4nF

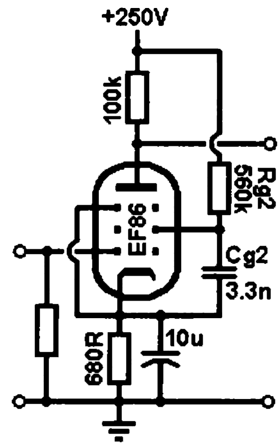

Fig. 3.21: A stage having partial screen bypassing, giving a -3dB frequency of around 409Hz.

So we would probably use the nearest standard of 3.3nF. This is shown in fig. 3.21, and the frequency response of this circuit is shown in fig. 3.22.

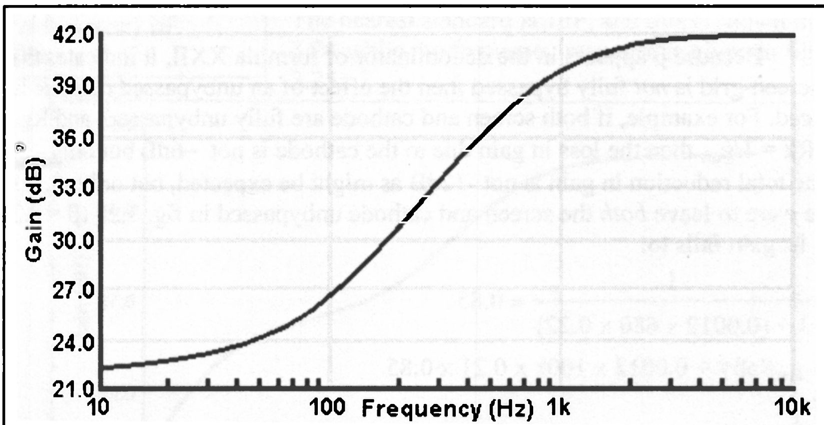

Fig. 3.22: Frequency response of the circuit in fig. 3.21.

The effect of cathode bypassing:

The effect of the cathode resistor in reducing the gain is given by:

$$\gamma = \frac{1}{1 + g_m R k \beta} \quad \text{...XXII}$$

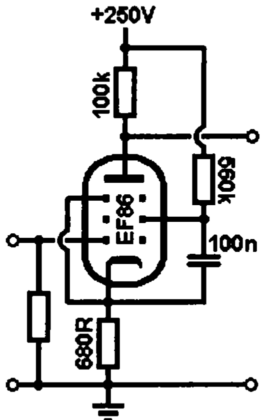

Fig. 3.23: A stage having an unbypassed cathode, but a bypassed screen, giving a gain of around 66 (36dB).

Assuming the screen is fully bypassed, then if $Rk = 1 / g_m$, then the gain is reduced by exactly half, or -6dB.

If the screen grid is fully bypassed ($\beta = 1$), and Rk is completely *unbypassed*, then combining XVII and XXII shows the voltage gain to be:

$$A_{(Rk\ unbypassed)} = \frac{g_m Ra}{1 + g_m Rk} \quad \text{...XXIII}$$

For example, the circuit in 3.23 is the same as that in fig. 3.19, except that the cathode has been left unbypassed. The screen is fully bypassed and has no effect on gain. The unbypassed cathode reduces the gain to:

$$A_{(Rk\ unbypassed)} = \frac{g_m Ra}{1 + g_m Rk} = \frac{0.0012 \times 100k}{1 + (0.0012 \times 680)}$$

$= 66$

which is not dissimilar to the gain of an ECC83 *with* a bypassed cathode!

Because β appears in the denominator of formula XXII, it indicates that if the screen-grid is *not* fully bypassed then the effect of an unbypassed *cathode* is reduced. For example, if both screen and cathode are fully unbypassed, and $Rg2 = r_{g2}$ and $Rk = 1/g_m$, then the loss in gain due to the cathode is not –6dB but only -3.5dB. So the total reduction in gain is not -12dB as might be expected, but only -9.5dB. If we were to leave *both* the screen and cathode unbypassed in fig. 3.23 ($\beta = 0.21$), then its gain falls to:

$$\gamma = \frac{1}{1 + (0.0012 \times 680 \times 0.22)} = 0.85$$

$A' = g_m Ra \beta \gamma = 0.0012 \times 100k \times 0.21 \times 0.85$
$= 21$

Which is not much less than the gain with the cathode bypass capacitor included (see fig. 3.20). In other words, screen bypassing has much more effect on the gain of the pentode than cathode bypassing does.

If the cathode is only partially bypassed, then the gain at any frequency is given by[4]:

$$A = g_m R_a \sqrt{\frac{1+(2\pi f R_k C_k)^2}{(1+g_m R_k)^2 + (2\pi f R_k C_k)^2}} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{XXIV}$$

But, as with the triode, it is more convenient to define the half-boost point, which may be found using:

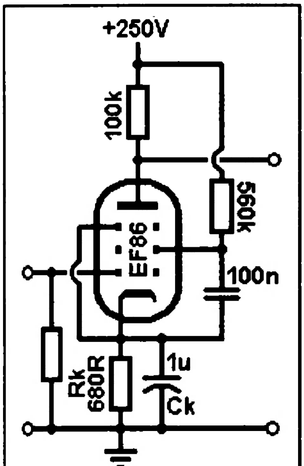

Fig. 3.24: A stage having partial cathode bypassing, giving a half-boost frequency of 234Hz.

$$f_{(\text{half boost})} = \frac{1}{2\pi R_k C_k} \cdot \sqrt{1 + \frac{g_m R_k}{2 + \frac{1}{2} g_m R_k}} \quad \ldots\ldots \text{XXV}$$

However, if the screen-grid is fully bypassed and provided $g_m R_k$ is not greater than about 3, which is true for most small-signal pentode circuits, then to a useful approximation the half-boost frequency is:

$$f_{(\text{half boost})} \approx \frac{1}{2\pi f R_k C_k}$$

For example, if we wanted to add treble boost to the circuit in fig. 3.23 we might select a half-boost frequency of 200Hz. Solving the above for Ck gives:

$$C_k \approx \frac{1}{2\pi f R_k} = \frac{1}{2\pi \times 200 \times 680}$$

$= 1.2\mu F$

The nearest standard is $1\mu F$, and this is shown in fig. 3.24 while the frequency response is given in fig. 3.25.

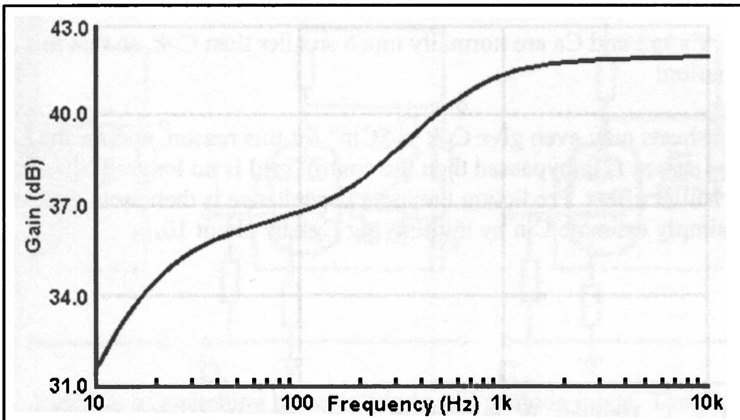

Fig. 3.25: Frequency response of the circuit in fig. 3.24.

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[4] Langford-Smith, F. (1957). *Radio Designer's Handbook* (4th ed.). p500.

If *both* the screen grid and the cathode are partially bypassed then the exact gain becomes more difficult to calculate, since γ is somewhat dependant on β. The method is too lengthy to discuss here; the reader may wish to refer to a semi-graphical technique described by Terman[5]. Otherwise, approximation using XXI and XXV may be used to begin with, and values adjusted once the circuit is built.

Output impedance:

The output impedance of a pentode is given by the same formula as used for a triode, VIII:

$$Zout = Ra \parallel r_a = \frac{Ra.r_a}{Ra + r_a}$$

However, because r_a is very large in pentode -typically at least 1MΩ- then Ra will normally be much less than r_a, so the above simplifies to:

$Zout \approx Ra$..XXVI

This is still true whether or not the screen and cathode are bypassed, since they only serve to increase r_a. Consequently, the output impedance of a pentode is normally much greater than that of a triode, so it is not well suited to driving heavy loads such as tone stacks. It is not unusual to find a cathode follower immediately following a pentode, so as to isolate it from heavy loads (see chapter 5).

Input capacitance:

Because the screen grid serves to shield the control grid from the anode, the Miller effect is virtually eliminated. The input capacitance is then given by:

$Cin = Cgk + Cg1g2 + Cga$

Where:

Cgk = grid to cathode capacitance.

Cg1g2 = control grid to screen grid capacitance.

Cga = grid to anode capacitance.

However, Cg1g2 and Ca are normally much smaller than Cgk, so that to a close approximation:

$Cin \approx Cgk$..XXVII

Some datasheets may even give Cgk as "Cin" for this reason, and for the EF86 it is 4pF. If the screen is unbypassed then the control grid is no longer fully screened from the Miller effect. Predicting the input capacitance is then more difficult, so it is usual to simply estimate Cin by multiplying Cgk by about 10.

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5 Terman, E. T., Hewlett, W. R., Palmer, C. W., Pan, W. (1940). Calculation and Design of Resistance Coupled Amplifiers using Pentode Tubes. *Trans. A.I.E.E.*, 59. pp.879–884.
See also: Langford-Smith, F. (1957). *Radio Designer's Handbook* (4th ed.). p500.

Switchable and variable pentode designs:

Because of the wide range of tones offered by pentodes it is useful to be able to switch or blend between different modes of operation. The most common of these is a pentode/triode switch, offering either high-gain, high sensitivity pentode operation or lower gain triode operation, for a warmer, cleaner sound.

Triode/pentode switches:

A triode/pentode switch is easily implemented by providing a switch to connect the screen grid either to the screen-supply circuit for normal pentode operation, or to the anode for triode-strapped operation. The Vox *AC30HH* 'Heritage' reissue incorporates such a switch, for example. Fig. 3.26 shows some switching arrangements (cathode bypass capacitors are omitted and screen bypass capacitors are shown dashed, for clarity).

Beginners often make the mistake of connecting the switch as in fig. 3.26*a*. This circuit must *not* be used, for two reasons: Firstly the screen grid must have a stopper resistor between it and the anode, to limit screen current and to dampen any stray impedances which might otherwise lead to oscillation. The second, less obvious problem with this arrangement is that when the switch is thrown the screen grid will momentarily be completely disconnected from anything. This is very worrisome as any 'floating' electrode can build up a static charge which may lead to arcing within the valve, and is made worse by switch bounce. Although many circuits have been built this way with no apparent problems, there is the further possibility of the switch failing to make proper contact as it ages or becomes dirty.

The necessary corrections to this circuit are shown in *b*. A resistor, R2, is connected across the switch so that at no point is the screen left floating. This will also help reduce audible 'pop' sounds if the switch is thrown while the amp is on. Its value should be large enough that it cause negligible loading on the valve during triode mode, and a value of 1MΩ to 2.2MΩ should be suitable for any circuit. During pentode mode it is shorted out by the switch, so does not interfere with normal

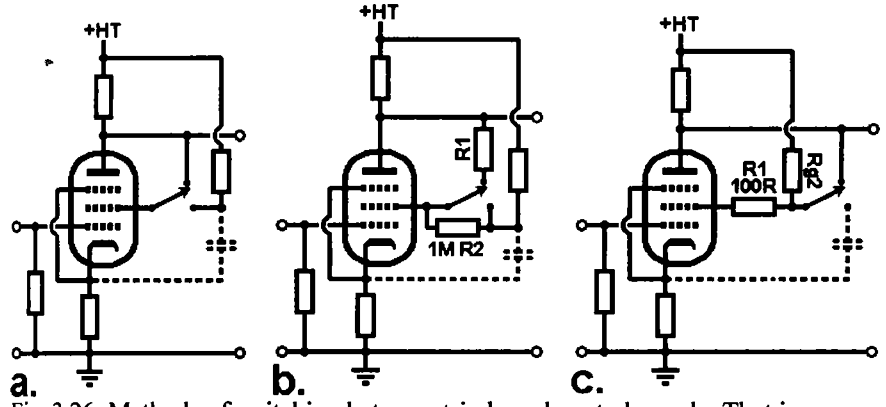

Fig. 3.26: Methods of switching between triode and pentode mode. That in **a.** should not be used as it could lead to arcing when the switch is thrown, while **b.** and *c.* avoid this. Screen and cathode bypass capacitors may or may not be used, as desired.

operation. The screen-stopper resistor, R1, is switched in during triode mode. A value of 100Ω is normally used for 'standard' triode operation, but larger values could be used to lower the gain and tailor the triode characteristics, as was shown in fig. 3.6.

The circuit in fig. 3.26*c* shows an alternative arrangement. A 100Ω screen-stopper is permanently connect to the screen, but is small enough that it has no significant effect during pentode operation. For triode operation the screen-dropping resistor, Rg2, is connected to the anode and therefore appears directly in parallel with the anode resistor. This reduces the load during triode mode, for reduced gain, more headroom and a warmer tone. Again, at no time is the screen left floating.
In each case we do not change the cathode bias resistor, so the valve will normally tend to bias warmer in triode mode. We could switch in a new value of cathode resistor, using a double-pole switch, but this might be taking a simple switching feature unnecessarily far.

Variable screen bypassing:

A triode/pentode switch is easy to implement in an existing amp since a miniature toggle-switch takes up very little space. However, since the screen grid wields such power over the operation and tone of the pentode, it makes sense to take even greater advantage of it, if we have sufficient space for a potentiometer. An obvious possibility would be to replace the screen-dropping resistor with a variable resistor so that the screen voltage could be varied, but this would place both high voltage and DC on the pot', which would be very noisy (though the use of an LDR and LED is one noise free alternative). A more straightforward approach is to make the screen-bypassing variable instead, in the same way that variable cathode bypassing was used in figs. 1.20b and 2.34. Two example circuits are shown in fig. 3.27. If the screen bypass capacitor is large, as in fig. 3.27*a*, then P1 operates as a gain control. However, at lower resistance settings it also acts as a variable screen-compression control since it alters the source resistance of the screen supply circuit, which gives powerful control over sustain when the pentode is overdriven. The frequency response of this circuit is shown in fig. 3.28.

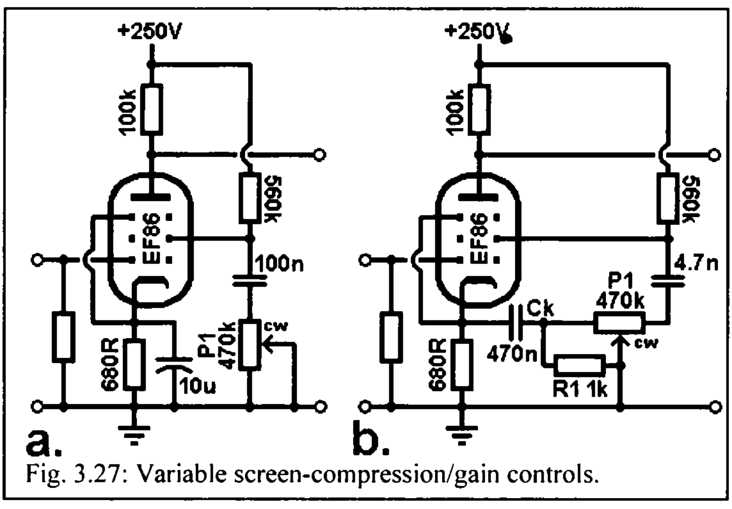

Fig. 3.27: Variable screen-compression/gain controls.

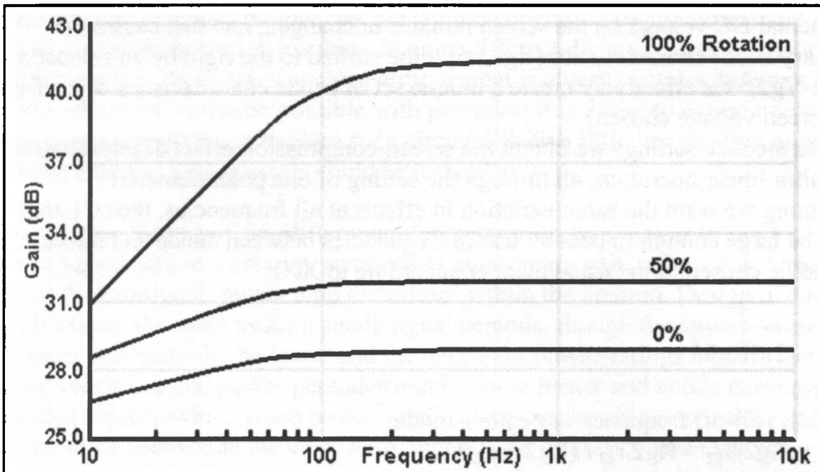

Fig. 3.28: Frequency response of the circuit in fig. 3.27a.

The circuit in fig. 3.27*b* shows a modified arrangement in which at one extreme the screen is partially bypassed but the cathode is unbypassed, and at the other extreme the opposite is true. P1 now operates as an active treble boost control, as well as affecting the screen compression on treble frequencies.
The value of the potentiometer is not critical, as even a modest resistance of 100kΩ is sufficient to heavily reduce the effectiveness of the screen bypass capacitor. R1 is included to alter the value of the variable resistance as far as the cathode is concerned, and help give a more even variation in tone as the pot' is rotated.

Triode/pentode 'morph' control:

Taking this even further it is not hard to combine both pentode/triode switching *and* variable screen bypass in a single control, if we appreciate that for triode mode the screen only has to be connected to the anode *as far as AC is concerned.* With the addition of one capacitor we obtain the circuit in fig. 3.29. We now find that with the control fully clockwise, the screen bypass capacitor, Cg2, is connected directly to ground and we have normal pentode operation. P1 is large enough that is does not heavily load the pentode.
When turned fully anticlockwise we see that C1 and Cg2 appear in series and effectively connect the screen directly to the anode, as far as AC is concerned, resulting in triode operation. (C1 is included simply to keep DC off the pot'.)

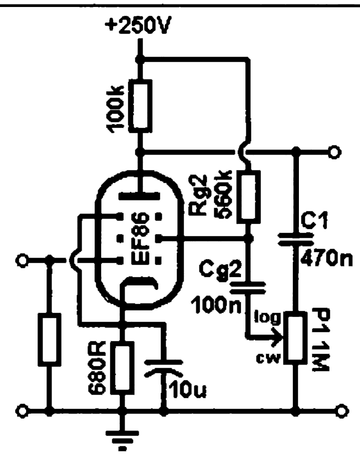

Fig. 3.29: Fully morphable pentode/ultra-linear/triode control.

The actual DC voltage on the screen remains unchanged, and this causes the ordinary triode characteristics [fig. 3.6] to be shifted to the right by an amount equal to Va-Vg2;[6] we effectively create a unique set of triode characteristics depending on the screen voltage chosen!
At intermediate settings we obtain the screen-compression effect described earlier *and* ultra-linear operation, all through the setting of one potentiometer!
Assuming we want the same variation in effects at all frequencies, then C1 and Cg2 must be large enough to pass all useful frequencies between anode and screen. Cg2 should be chosen in the usual manner according to XX1:

$$Cg2 = \frac{1}{2\pi f Rg2'}$$

Where:
f = -3dB roll-off frequency in pentode mode.
$Rg2' = Rg2||r_{g2} = Rg2.r_{g2} / (Rg2 + r_{g2})$

C1 must be large enough that the reactance of the series combination of C1 and Cg2 is less than $Ra||r_{a(triode)}$ at the lowest desired frequency. Therefore C1 should be at least five times larger than Cg2, if possible, for triode operation down to around 100Hz. The frequency response of the circuit in fig. 3.29 is shown in fig. 3.30, and it is interesting to note the stage has the lowest gain at an intermediate (ultra-linear) setting of P1.
Unusual results are obtained if C1 is made small, 1nF to 22nF say, since this causes

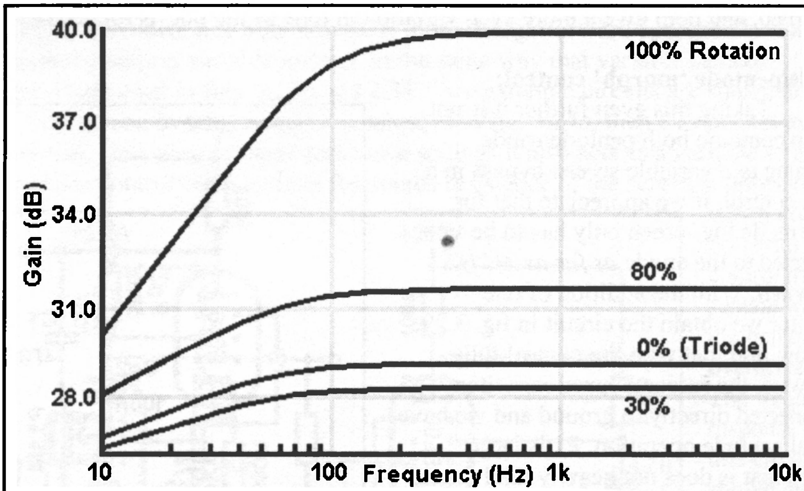

Fig. 3.30: Frequency response of the circuit in fig. 3.29. Note that the circuit actually has the least gain at 30% rotation.

[6] Cuthbert, D. (2003). Get More Power with a Boosted Triode. *Electronics Design Strategy News*. (June 12), pp90-91.

triode operation at higher frequencies but pentode operation (with unbypassed screen) at lower frequencies, resulting in quite different distortion characteristics at the two extremes. Analysis of such an arrangement is a very complex testament to the enormous tonal variation possible with pentodes; it is easier to experiment and use computer simulation to explore such circuits! With a little imagination quite advanced tone and gain controls could be arrived at.

Some 'boutique' amplifier manufacturers even go so far as to use a power pentode somewhere in a preamp, arranged as an ordinary gain stage, so as to capture some of the traditional 'power amp distortion' within the preamp. Design of such a stage is exactly the same as for a small-signal pentode, though the appropriate resistances will normally be lower and the necessary power-ratings higher. The obvious concern is that power pentodes require more heater and anode current than small-signal types -which some power supplies may not be able to provide- and will dissipate more heat inside the amp. Also, without the frequency and load varying characteristics of a real output transformer and speaker, such circuits rarely come very close to modelling the sound of a real power amp. Nevertheless, very small power pentodes commonly used in late-generation valve radios and record players are a reasonably efficient possibility. Examples include triode power pentodes such as the ECL80 / 6AB8 which contains a low gain triode similar to the ECC82 and 3.5W pentode, and requires only 300mA, 6.3Vfor the heater, though the triode and pentode share the same cathode connection. The ECL86 / 6GW8 contains half an ECC83 triode and a 9W pentode, but requires 660mA, 6.3V for the heater. The EL91 / 6AM5, EL95 / 6DL5 (all 7-pin miniature types) and EL85 power pentodes require only 200mA, 6.3V for their heaters, and can often be bought as NOS very cheaply.

Summary of formulae:

XVII; Voltage gain with screen and cathode fully bypassed:

$$A = g_m Ra$$

XVIII; Voltage gain with screen and cathode not fully bypassed:

$$A' = g_m Ra \beta \gamma$$

Where:
β = loss of gain due to an impedance at the screen grid.
γ = loss in gain due to an impedance at the cathode.

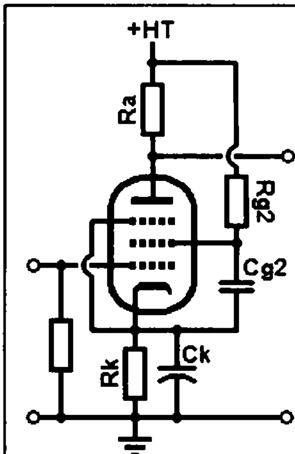

Fig. 3.31.

XIX; Internal screen resistance:

$$r_{g2} = r_{a(triode)}(1+m)$$

Where:
$r_{a(triode)}$ = the anode resistance of the same valve when triode strapped, at the corresponding anode current.
m = the Ia:Ig2 ratio.

XX; Reduction in gain due to an unbypassed screen dropping resistor:

$$\beta = 1 - \frac{Rg2}{Rg2 + r_{g2}}$$

XXI; -3dB roll-off frequency due to a partially bypassed, screen dropping resistor:

$$f = \frac{1}{2\pi Cg2Rg2'}$$

Where:
Cg2 = the screen bypass capacitor.
$Rg2' = Rg2 \| r_{g2} = Rg2.r_{g2} / (Rg2 + r_{g2})$

XXII; Reduction in gain due to an unbypassed cathode-bias resistor:

$$\gamma = \frac{1}{1 + g_m Rk\beta}$$

XXIII; Voltage gain with screen fully bypassed and cathode unbypassed:

$$A_{(Rk\ unbypassed)} = \frac{g_m Ra}{1 + g_m Rk}$$

XXIV; Voltage gain at any frequency with screen fully bypassed:

$$A = g_m Ra\sqrt{\frac{1 + (2\pi fRkCk)^2}{(1 + g_m Rk)^2 + (2\pi fRkCk)^2}}$$

XXV; Half-boost frequency due to cathode bypass capacitor;

$$f_{(half\ boost)} = \frac{1}{2\pi RkCk}.\sqrt{1 + \frac{g_m Rk}{2 + \frac{1}{2}g_m Rk}}$$

Solving for Ck gives:

$$Ck = \frac{1}{2\pi fRk}.\sqrt{1 + \frac{g_m Rk}{2 + \frac{1}{2}g_m Rk}}$$

Where f is the desired half-boost frequency. However, if the screen-grid is fully bypassed and provided $g_m Rk < 3$ then to a reasonable approximation:

$$f_{(half\ boost)} \approx \frac{1}{2\pi fRkCk}$$

XXVI; Anode output impedance assuming $r_a >> Ra$:

$$Zout \approx Ra$$

XXVII; Total input capacitance:

$$Cin \approx Cgk$$

Where Cgk is the grid-to-cathode capacitance, and may be quoted on the data sheet simply as 'Cin'. If the screen is unbypassed then the control grid is no longer fully screened from the Miller effect. Predicting the input capacitance is then more difficult, so it is usual to simply estimate Cin by multiplying Cgk by about 10.

Where in all cases all notations are as in fig. 3.31 and;
g_m = transconductance of pentode at operating point.
All resistances in ohms. All capacitances in farads.

Chapter 4: Valves in Parallel

The Matchless input stage. The common-anode mixer. Switchable designs.

It is obvious when looking at existing amplifier schematics that the majority of preamps are made up of single valve stages cascading into one another, with various controls and coupling networks in between. However, there are instances where it can be useful to use two or more valves in parallel, rather than in series, and this will be discussed in this chapter.

When valves are placed in parallel we may either look upon it as a number of individual valves working into the same load or, which is more straightforward, as a single, new, 'bigger' valve. Normally the valves will be of the same type and in the same envelope, but they do not have to be, and this opens up even more tonal possibilities, such as triodes in parallel with pentodes!

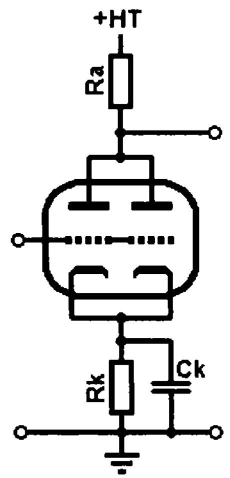

Fig. 4.1: Two triodes connected in parallel.

Fig. 4.1 shows a typical arrangement of parallel triodes in the same envelope. The grids are shown connected to each other inside the envelope to keep the diagram simple, in reality the connection would be external of course. The easiest way to go about selecting component values is to remember that a valve can be regarded as a signal generator in series with a resistance, r_a, and the same rules which apply to paralleled resistances also apply to valves: the total resistance must be less than either resistance alone, total current increases while voltage relationships remain the same. If the valves are connected as in fig. 4.1, with all electrodes connected together and sharing the same loads, we may regard the whole as a single valve with twice the current handling ability. We can easily represent this on the anode characteristics graph by simply doubling the anode current scale (y-axis). This is illustrated in fig. 4.2. If we had connected three identical triodes we would triple the current scale, and if four triodes then we would quadruple the current scale, and so on.

A single ECC83 triode has a maximum anode dissipation limit of 1W, so two in parallel should have a total limit of 2W* since currents are doubled, and this is also shown.

Notice that because the anode voltage scale (x-axis) and grid curves are unchanged, μ is also therefore also unchanged. However, for a given change in grid voltage the change in anode current is now doubled, so g_m must be double its original value. Since $\mu/g_m = r_a$, it also follows that r_a is halved, and this is to be expected since two identical resistances placed in parallel equal a total resistance of one half their individual values.

---

* When the valves are in the same envelope the total dissipation limit may not always be doubled, since there is a maximum safe heat limit which can be radiated by the envelope, and this will be given on the datasheet. For example, the 6SN7GT dual-triode quotes a limit of 5W for each triode, but only 7.5W for both triodes simultaneously.

If we had used three valves in parallel we would find g_m tripled and r_a would be on third its original value, while μ is still unchanged, and so on. Exactly the same principles apply to pentodes.

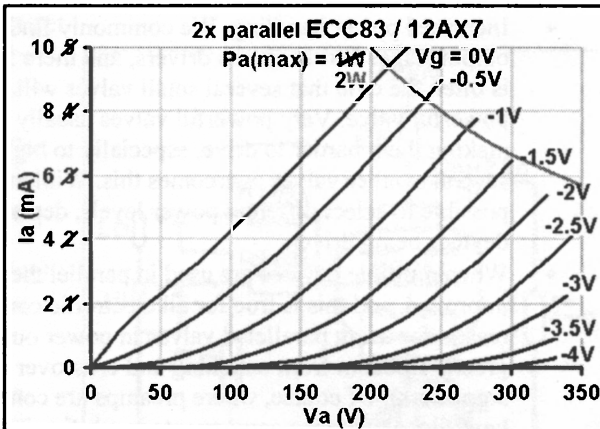

Fig. 4.2: Static anode characteristics of two ECC83 triodes in parallel. Current and g_m are doubled, while r_a is half the value for a single triode.

To get an idea of how all this affects gain we could apply formula III to some typical values. For quick reference we will take datasheet values for a *single* triode of $\mu = 100$, $r_a = 63k\Omega$. For a usual $100k\Omega$ anode load;

$$A = \frac{\mu.Ra}{Ra + r_a} = \frac{100 \times 100k}{100k + 63k}$$

= 61 (35.7dB).

Now, noting that for a pair of triodes in parallel, r_a is now $63k\Omega / 2 = 31.5k\Omega$, if we were to use the *same* $100k\Omega$ anode resistor:

$$A = \frac{\mu.Ra}{Ra + r_a} = \frac{100 \times 100k}{100k + 31.5k}$$

= 76 (37.6dB).

Because g_m has doubled, the change in current in the load has increased for a given input signal, resulting in a 23.5dB increase in gain! In fact, to achieve the *same* gain as the single triode we would need only half the value of anode resistor, or $50k\Omega$. We can also note that for the *same* bias point we would only need half the value of cathode resistor, Rk, since twice the anode current flows, while the cathode bypass capacitor, Ck, would need to be doubled to obtain the same frequency response.

Why use valves in parallel?

With the previous discussion in mind it is easy to see the advantages of using valves in parallel:

- Anode resistance is halved, and other circuit resistances will usually be lower than for a single valve too, meaning reduced noise. This is the main reason for using parallel preamp valves in hifi. This also makes them most useful as an input stage in a guitar amp, where the signal-to-noise ratio is very important.
- Lower resistances will mean a lower output impedance, which may be useful for driving heavy loads such as tone stacks.

- Increased power handling. We commonly find paralleled valves in power output stages and as reverb drivers, and there are several reasons for this. It is often the case that several small valves will cost less than one, more powerful valve. Very powerful valves usually have low input sensitivity, making them harder to drive, especially to high distortion levels, and using several smaller valves overcomes this. With multiple valves it is also possible to select different power levels, depending on the number of devices being driven.
- When multiple devices are used in parallel the tolerance of the whole is improved, and this is true for all electronic components. This is another reason for using paralleled valves in power output stages since it gives greater freedom from matching and crossover distortion (in push-pull amplifiers). Of course, where preamps are concerned it is more important to have close tolerance components in a hifi amp that in a guitar amp, but it is still useful.

The obvious disadvantages to paralleled valves are the loss of a valve which might be more useful elsewhere in the circuit, and a possible increase in current demanded from the power supply, particularly where power valves are concerned, since several valves in parallel may need more heater and/or screen current than one, more powerful valve. It could also be argued that there are more possibilities for tone shaping and control with two cascades stages, but this needn't always be the case, as discussed later. Paralleled valves will also draw greater grid current than one valve of the same type. This is less of a concern with power pentodes as their grid current is usually quite low. For triodes, however, using too many in parallel may result in a harsh sounding overdrive, and increased risk of blocking distortion due to too much grid current being demanded. Admittedly, this may be somewhat offset by the increased input capacitance (interelectrode capacitances are doubled when two valves are used) which will help to attenuate harsh, high-frequency harmonics, and in any case this is not a major concern for low grid-current valves such as the ECC82 and ECC83; we are unlikely ever to use more than two in parallel at once, and at such levels the overdrive should be pleasantly 'gravelly'. Many users have described the sound of a paralleled ECC83 as being 'fatter' than a single one, and this may indeed be partly due to the increased input capacitance generating a slightly softer tone.

It should be clear from the above discussion that apart from altering the anode current scale on the static anode characteristics graph, and factoring in increases in interelectrode capacitance, the design of simple paralleled gain stages will follow exactly the same procedure as a single-valve gain stage.

The Matchless input stage:

At this point it may be worth examining a real example taken from a well regarded amplifier, the Matchless *Spitfire*. Fig. 4.3 shows the circuit, which appears at the input of the amp (the same input circuit is also used in the Matchless *Lightning*). The first things to note are that the anode resistor is large considering it is a paralleled stage, implying very high gain, and that the input grid-stoppers are also quite large considering the input capacitance will be greater than usual.

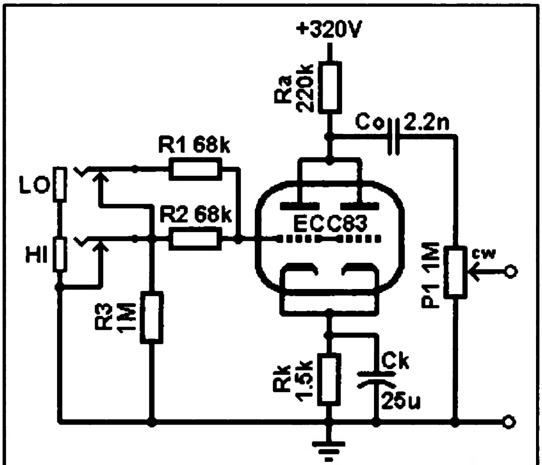

Fig. 4.3: Very high gain, parallel input stage featured in the Matchless *Spitfire* and *Lightning*.

Fig. 4.4 shows the anode and cathode load lines. The load line appears very low down due to the increased anode current scale, so interpreting the graph will be less accurate than usual, but the bias point appears to occur at about Vgk = -1.3V, which is very warm. Judging the gradient of the grid curves at the bias point suggests an anode resistance, r_a, of about 62kΩ, which is about the same as a single ECC83 triode under more typical conditions. Obviously this circuit does not take advantage of the opportunity for less noise.

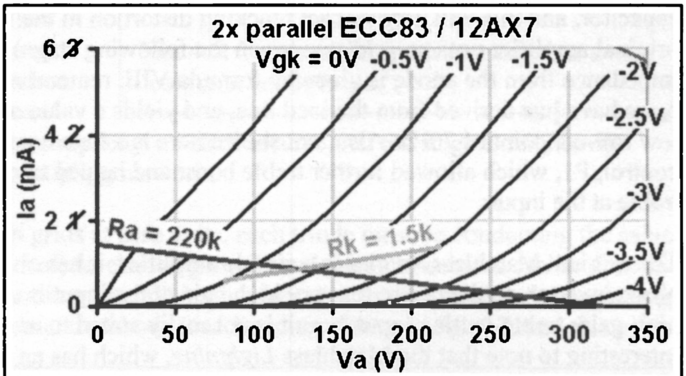

Fig. 4.4: Anode and cathode load lines for the circuit in fig. 4.3.

The maximum output signal swing is very large at about 295Vp-p, while the gain is about 76 (37.6dB)!

We can now calculate the input capacitance. The datasheet quotes the following values; Cga = 1.6pF, Cgk = 1.6pF, but since we have two triodes both these values are doubled to 3.2pF. Applying formula VII:

Cin = Cgk + (Cga.A)

Cin = 3.2 + (3.2 x 76)

= 246pF

This is more than twice the value we find in most preamp stages.

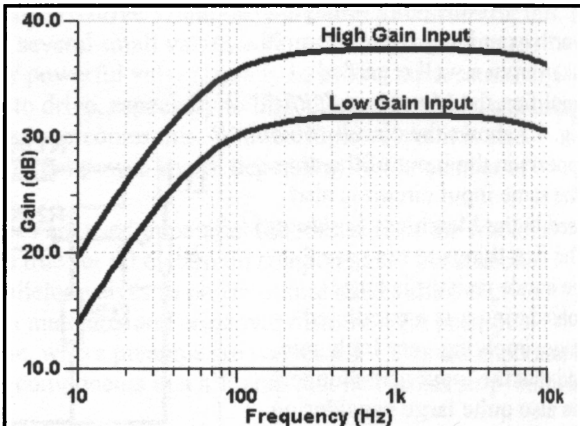

Fig.4.5: Frequency response of the circuit in fig. 4.3.

The switching jack socket arrangement means that when the high-gain input is used, R1 and R2 appear in parallel giving an input roll-off point of around 19kHz, which is conservatively low (the low gain input gives a similar figure, since R2 is shorted to ground by the other switched jack). The frequency response for each input is shown in fig. 4.5.
However, the high roll-off frequency quickly falls if the guitar's volume control is turned down even a small amount, causing a noticeable loss of 'chime'.
There is considerable bass attenuation introduced by the small output coupling capacitor, and this will help restrict blocking distortion in the following stage (the original amplifier uses no grid stopper on the following stage). The output impedance from the anode is given by formula VIII, remembering to use the value of r_a we have just derived from the load line, and yields a value of about 49kΩ, giving a low roll-off point of 70Hz. Also, not shown here is a tone control following the gain control, P1, which allowed further treble boost and helped to compensate for loss of treble at the input.

The original Matchless *Spitfire* is a simple amplifier; it has only an input stage and phase inverter, so the desire for very high gain at the input is understandable, but this high-gain, treble-cutting input circuit is not really suited to multi-stage preamps. It is interesting to note that the Matchless *Lightning*, which has an additional two stages after the input stage, includes considerable measures to reduce bass response and avoid blocking distortion. A more elegant solution would have been simply to reduce the input's anode resistor to say, 100kΩ, or even a little lower, thereby reducing the gain and input capacitance to more manageable levels without going so far as to rob the amp of its ability to produce high-gain overdrive. The Matchless *Clubman*, on the other hand, used a high-gain 6SH7 pentode for the second stage, so to avoid extremely high overall gain levels the parallel input stage used a mere 10kΩ anode load with a 2.2kΩ *unbypassed* cathode resistor, producing a gain of only about 4! The more logical approach would have been to reduce the gain of the pentode so that it could handle a greater dynamic range, and increase the gain of the input stage to a more usual level, so as to amplify the input signal above the noise floor more.

It is probably fair to say that very few players have ever valued the traditional 'low gain' input provided on many amplifiers, as it does not offer a great deal more than could be achieved by simply turning down the guitar's volume control. Fortunately, parallel input stages, in particular, offer many possibilities for differently voiced inputs without compromising the natural chime of the guitar. With this in mind we shall explore some more advanced designs.

The common-anode mixer:

Fig.4.6 shows two triodes in an arrangement commonly referred to as a **common-anode mixer**. This does not look much different from the parallel input stage discussed previously, except that the grids are *not* connected to each other. Instead, each grid is an individual input so both triodes can amplify independently, but because they share a common anode resistor the signals are mixed together at the output. The anode resistance, r_a, of each valve forms a further load on the other, which produces some unusual results. For simplicity we shall assume that the stage is not heavily loaded by the following coupling network.

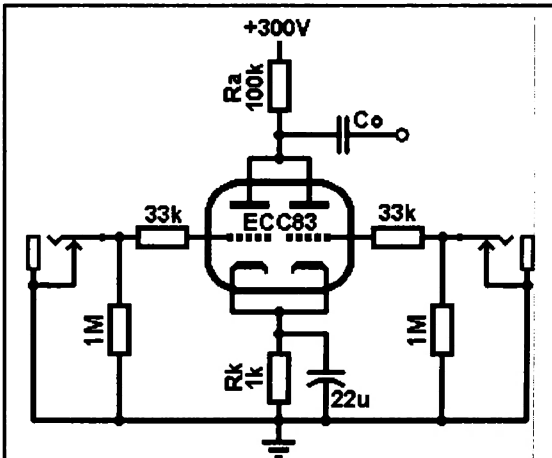

Fig. 4.6: Two parallel triodes arranged as a 'common anode mixer'.

At quiescence, with both grids at zero volts, each triode must be conducting the same (assuming they are identical) so choosing anode and cathode resistors can be done in the same way as for an ordinary parallel pair of triodes, as discussed at the beginning of this chapter. The anode and cathode load lines are shown in fig. 4.7, and this gives us our quiescent bias point of just under $Vgk = -1.4V$. Quiescent anode current is about 1.4mA, so each valve must be conducting 0.7mA. However, the gain of each triode is still the about same as that of a single

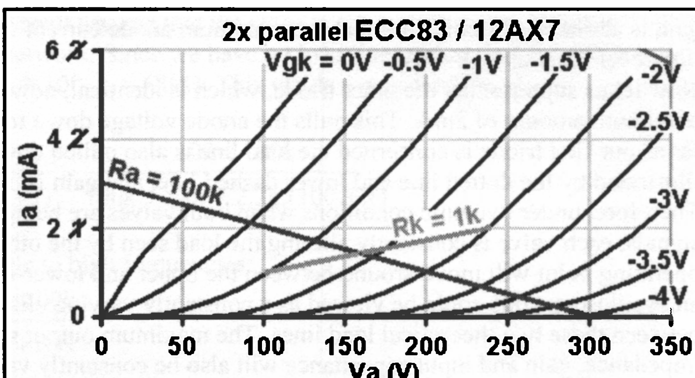

Fig. 4.7: *Quiescent* anode and cathode load lines for the circuit in fig. 4.6, indicating a bias of just under -1.5V.

stage since both triodes work independently (unless they received identical signals, resulting in normal parallel operation).

The dynamic operation of each valve is somewhat unexpected:
We already know what the bias point is from fig. 4.7 and this cannot change (apart from the small increase we normally see under dynamic conditions), and is indicated by the dot in fig. 4.8.

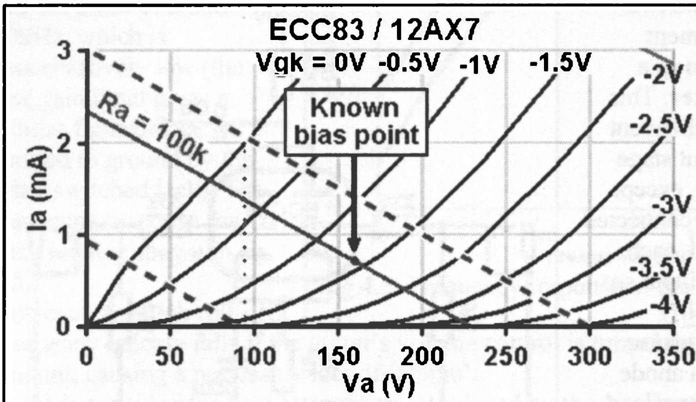

Fig. 4.8: Dynamic operating area for *one* triode in the circuit in fig. 4.7. The operating point may exist anywhere between the two, dotted load lines.

If one valve remains idle, with no input signal, then it will continue to conduct 0.7mA, pulling the anode voltage permanently down by 70V. The load seen by the remaining valve can then be represented as an ordinary 100kΩ load line beginning at 300 – 70 = 230V, and passing through the bias point as shown by the solid line in fig. 4.8. If only one input is used then this load line suggests a gain of about 52 (34dB) and a maximum output signal swing of about 155Vp-p. Note that we have not drawn an AC load line; the load presented by the other triode is a *DC* load, since no coupling capacitor is involved.
However, if we suppose that one triode is driven fully to cut-off then it no longer pulls the anode voltage down –it is effectively removed it from the circuit– and the remaining triode sees only the 100kΩ anode resistor and a load line beginning at the 300V HT. This is indicated by the upper dashed line, from which we can see that the gain is about 60 (35.6dB) and that the maximum anode current which can flow is a little over 2mA at Vgk = 0V.
Now let us suppose that the other triode, which is identical, now draws this maximum amount of 2mA. This pulls the anode voltage down to about 95V, so as far as our first triode is concerned the load line is also pulled down to this level, as illustrated by the dotted line and lower dashed line, and gain is only about 45 (33dB). Therefore, under dynamic conditions when both valves are being driven by different signals, each valve is constantly altering the load seen by the other, and each valve's operating point will move around between the upper and lower-most load lines. On an oscilloscope this could be viewed as a constantly varying ellipse, existing between these two theoretical load lines. The maximum output signal swing, output impedance, gain and input capacitance will also be constantly varying, and all this leads to some interesting harmonic generation, which is well suited to a guitar amp.

The circuit shown in fig. 4.6 is for an input stage, and is very similar to that found in the Fender *'TV front' Bassman*, but this type of mixer could be used anywhere in a preamp, to mix two separate channels, say. The only point of note is that if the two triodes receive input signals which are identical but 180° out of phase with one another then each triode will conduct in exactly the opposite manner to the other, so the two will cancel out and we will have no output signal. Of course, we are unlikely to ever encounter this situation.

In the last case the cathodes shared the same cathode resistor, which was fully bypassed to maximise gain. However, we could use *separate* cathode-bias circuits, which introduces further possibilities, and fig. 4.9 shows one such example. We now have two triodes with very different gain characteristics; V1a has a 2.2kΩ, unbypassed cathode resistor while V1b has an 820Ω, partially bypassed cathode resistor, implying warmer bias and higher gain than V1a. Trying to draw load lines for this sort of arrangement would be very time consuming, so we would normally estimate suitable values for cathode resistor and, if necessary, adjust them once the circuit is built.

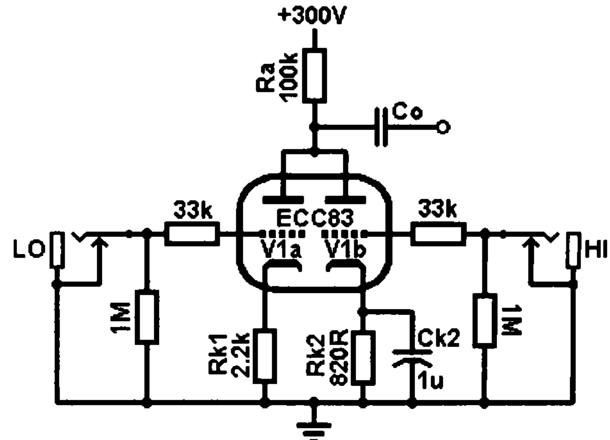

Fig. 4.9: Common anode mixer with dissimilar cathode biasing.

We can estimate the gain of each individual triode in the usual way, using formulae III and IV. Again we shall assume that the stage is not heavily loaded by the following coupling network. Since we have no load line to examine we will take datasheet values of $\mu = 100$, $r_a = 63k\Omega$. This yields a gain for V1a of:

$$A = \frac{\mu.Ra}{Ra + r_a + Rk(\mu + 1)} = \frac{100 \times 100k}{100k + 63k + 2.2k.(100 + 1)}$$

= 26 (28dB).

And for V1b at middle to high frequencies:

$$A = \frac{\mu.Ra}{Ra + r_a} = \frac{100 \times 100k}{100k + 63k}$$

= 61 (35.7dB).

In practice we will find that these figures are somewhat lower due to the loading effect of each valve, but immediately we see that we have a 'low gain' and 'high gain' input.

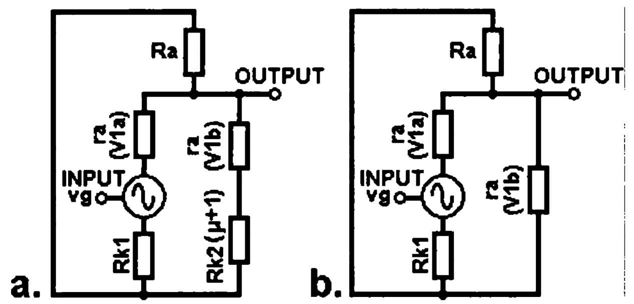

Fig. 4.10: Thévenin equivalent circuits for V1a in fig. 4.9 assuming V1b is in a quiescent state, for **a.** low frequencies and **b.** high frequencies.

The high gain input, V1b, is partially bypassed by Ck2, giving a half boost frequency of around 190Hz. However, this has implications for V1a: In chapter 1 it was mentioned that impedances in the cathode appear multiplied by μ+1 when looking into the anode, and so at low frequencies when Ck2 is effectively an open circuit, the resistance presented to V1a by V1b must be $r_a + Rk2(\mu + 1) \approx 146k\Omega$. At high frequencies, however, Ck2 bypasses Rk2 and the impedance seen by V1a is simply $r_a = 63k\Omega$. This is indicated by the Thévenin equivalent circuit in fig. 4.10. Therefore, even when V1b is *not* being driven, V1a still works into a dynamic load, so its gain and maximum output swing will both be reduced at high frequencies due to the heavier DC load formed by V1b's anode resistance. V1a's cathode is unbypassed, so it presents a constant resistance to V1b of $r_a + Rk1(\mu + 1) \approx 285k\Omega$. So, if V1a is not driven, V2b sees a constant load impedance and its gain and other parameters will be unchanging. If both triodes are driven then we have a very complex, dynamic load, as in our earlier example.

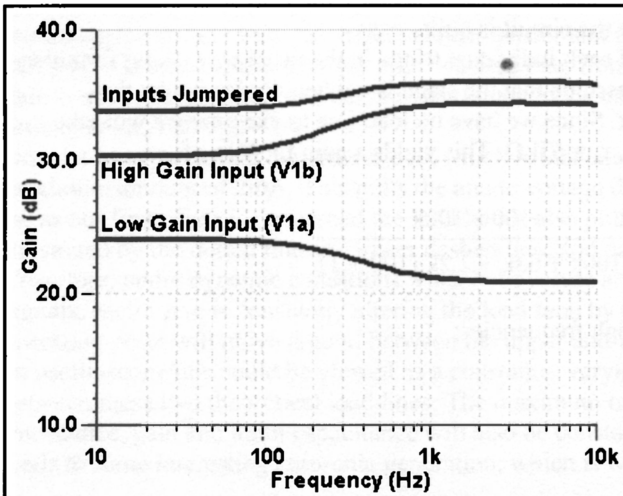

Fig. 4.11: Frequency response of the circuit in fig. 4.9.

Fig. 4.11 shows the frequency response when either input is used, and with both inputs jumpered, and the gain of each stage is indeed slightly lower than the estimated values above. The lower line clearly shows the reduction in gain at high frequencies caused by V1b loading V1a. The loss of treble is only about -3dB, not enough to dull the tone, but could be described as 'mellow' sounding,

while the small treble boost at the high-gain input adds 'clarity'.
Again, although the circuit shown in fig. 4.9 is an input circuit, the same idea could be applied elsewhere in the preamp. For example, when mixing a high amplitude 'dry' signal within the amp with a low amplitude 'wet' signal received from an effects loop, it may be advantageous to use a mixer with low and high gain inputs, such as this one, so that the signals being mixed at the output are then of similar level [see fig. 11.19]. Alternatively, if we connect both grids together as in the standard parallel arrangement we will obtain some very textured harmonics, and it is well worth experimenting with the cathode bias arrangements. Such a circuit lends itself especially well to clean or medium gain acoustic, jazz, blues and bass amplifiers, where this type of complex harmonic generation may help add 'shimmer' and 'warmth' to the tone, and could even be regarded as a simple means of replicating some of the features of power-amp distortion.

A further variation on the common-anode mixer is shown in fig. 4.12 and shows how a mix or 'blend' control can be incorporated. The potentiometer, P1, allows the cathode bypassing on each triode to be varied from negligible to fully bypassed, and Ck1 and Ck2 also keep DC off the pot' to avoid scratching sounds. In this way the gain of each triode can be varied from minimum (unbypassed) to maximum (fully bypassed) in the opposite fashion to the other, so that the relative proportions of each input signal can be blended together. This too is useful for mixing wet effects with dry signals, or blending between a clean and high-gain channel. The example shown uses an ECC83, but the same principle could be applied with any valves and with any combination of anode and cathode resistors and cathode bypass capacitors, of course.

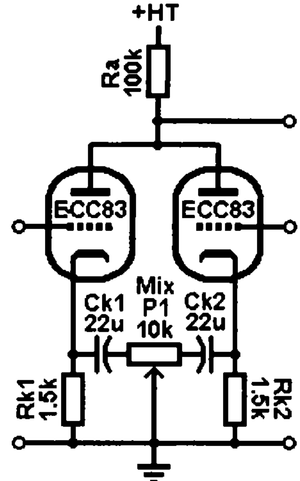

Fig. 4.12: A common-anode mixer with a mix or 'blend' control.

Switchable designs:

By this point the reader should appreciate that a paralleled pair of valves can have almost as many tonal features to offer as a pair of cascaded gain stages or even a pentode. But we need not limit ourselves to one arrangement or the other. With a little ingenuity we could devise switchable circuits. Just a few possibilities are suggested here.

Fig. 4.13 is essentially the same circuit as in fig. 4.6 except that the switched jack sockets have been arranged so that if we use the 'clean' input only, one triode is used while the other input remains grounded. If only the 'fat' input is used then both triodes are driven in parallel for slightly more gain, a fuller tone and lower noise. If both inputs are used simultaneously then each triode works independently, of course.

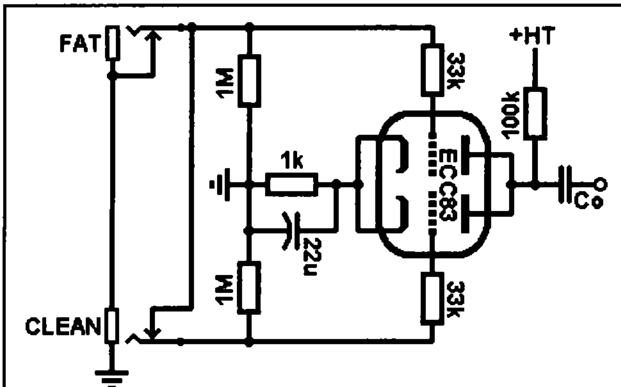

Fig. 4.13: An input stage offering either single or paralleled triode operation.

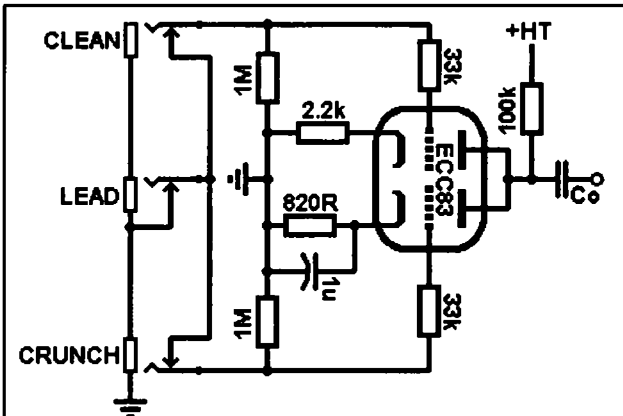

Fig. 4.14: An input stage offering low-gain single triode, high-gain single triode, or parallel triode operation.

When nothing is plugged in then both triodes are grounded via the 'fat' input jack, ensuring quiet operation. This arrangement is a good alternative to the Matchless *Spitfire* input discussed earlier.
Fig. 4.14 takes this idea further, and is a variation of the circuit in fig. 4.9, but with a third input jack provided. When the 'clean' input is used then only the low gain triode is driven, the other being grounded via the 'lead' jack. The 'crunch' input drives the higher gain input, while the other is again grounded. When the 'lead' input is used then both triodes are driven in parallel, for slightly higher gain and an even more textured sound (see also fig. 4.11).

Finally, fig. 4.15 shows a more complex arrangement of switching. This circuit is closely based on the Marshall *Master Volume Preamp*, which appeared in the 1970s JMP and 1980s JCM800 series of amplifiers and makes clever use of switching jack sockets. A double-pole switch, S1, has been added to the original design to allow toggling between parallel or cascaded (serial) triodes, for which some explanation may be required:

As shown, S1 is open and the circuit operates as the original did: When the LO input is used both triodes receive an input signal (because the low input jack is connected to the HI input jack's switch). V1b has an unbypassed cathode and very low gain. The output of V1a, however, is grounded via R1 to the HI input jack's ground switch and contributes nothing to the sound, only V1b is actually used.
When the HI input is used then R1 is disconnected from ground and remains connected to the LO input jack's switch. The output of V1a is then free to pass to

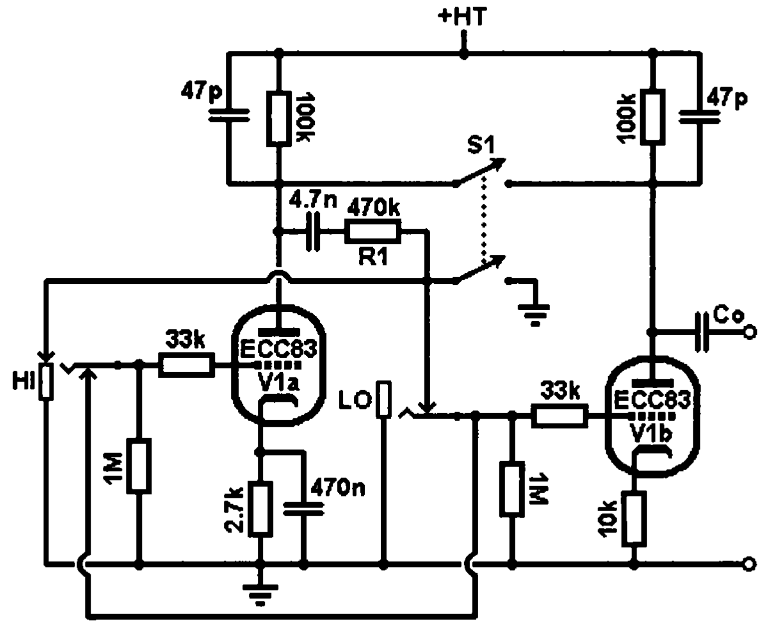

Fig. 4.15: A complex input arrangement based on the Marshall *Master Volume Preamp*. S1 has been added to allow switching between parallel and cascaded triodes.

V1b via the coupling capacitor and attenuating resistor R1. In this mode we therefore have cascaded operation and a conventional high-gain input circuit. V1a is partially bypassed with a half-boost frequency of around 120Hz.

Now consider when S1 is closed. The anodes of the valves are connected together for parallel operation (the paralleled anode resistors form a total load of 50kΩ of course). Now when the HI input is used only V1a receives an input signal, V1b's input is grounded. V1b is still loading V1 though, and we have a situation similar to that in fig.4.14 when the 'crunch' input was used. S1 also ensures that R1 is permanently connected to ground to avoid feeding the output of V1a to V1b's grid. The value of R1 is large enough that it does not cause significant loading in this mode.

When the LO input is used both valves again receive the input signal and we have full parallel operation, as in fig. 4.14 when the 'lead' input was used. In this sense the 'low' and 'high' roles are reversed.

In this way we obtain four possible tonal variations, depending on the input used and the setting of S1. The main drawback to this circuit is that there may be a loud 'pop' when S1 is actuated and the DC levels at the anodes instantly adjust. Anode bypass capacitors have been added to help reduce this slightly, and to reduce the risk of parasitic oscillation when in cascade mode. Fig. 4.16 shows the frequency response for each of the four input variations, measured from jack socket to the output of V1b.

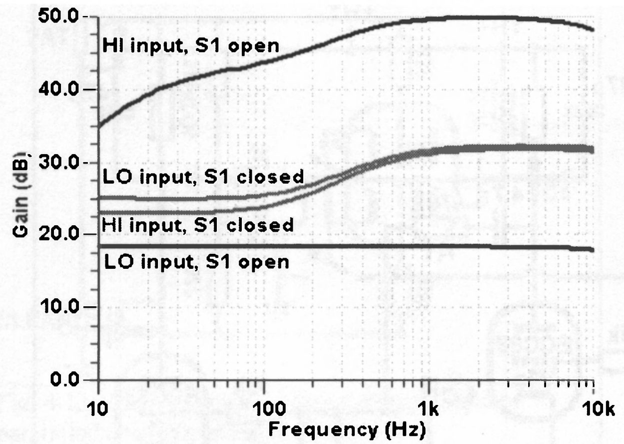

Fig. 4.16: Frequency response of the circuit in fig. 4.15.

Note that in parallel mode the difference between inputs appears only slight in terms of gain, the greater difference is in the timbre of the tone. Wildly different results can be obtained by altering the values of the cathode bias resistors and/or bypass capacitors; it is always worth experimenting. Because of the relative complexity of this circuit, a suggested layout is provided in fig. 4.17 for ease of construction (heater wiring is omitted for clarity). Stereo jack sockets are shown, though only mono jack sockets are really required. S1 is shown as a conventional double-pole, double-throw (DPDT) switch, though only a single-pole, double-throw (SPDT) switch is really required.

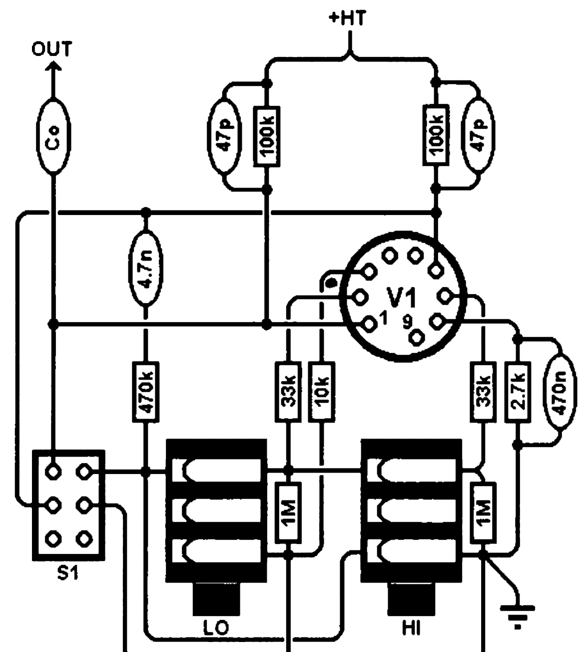

Fig. 4.17: Suggested point-to-point layout for the circuit in fig. 4.15 (heater wiring omitted).

Chapter 5: The Cathode Follower

Fundamental parameters of the cathode follower. Graphical analysis. Biasing arrangements for AC-coupled cathode followers. DC-coupled cathode followers. The pentode as a cathode follower. Cathode followers in guitar amps. Bootstrapping for more gain. Summary of formulae.

In an ordinary gain stage we use the control grid for the input and the output is taken from the anode. However, we can also take the output from the cathode, in which case we derive a circuit known as the **cathode follower** and, as will be shown, it has some unique properties. Many textbooks give the cathode follower only cursory treatment, describing it simply as an impedance buffer. This is broadly true in hifi circuits, but in guitar circuits a cathode follower may contribute as much to the tone of the amp as *any* other stage, and is particularly useful in high-gain designs. Cathode followers are also commonly used to drive tone stacks and effects loops. This chapter begins by describing the cathode follower in conventional, 'textbook' terms, before explaining its typical role in a guitar amp.

Fundamental parameters of the cathode follower:

Fig. 5.1 shows a simplified cathode follower circuit. The bias might be derived from a fixed or cathode-biased source, both will be described later. The first point to note is that if the grid swings more positively, current through the valve increases and this also flows in the cathode resistor. This causes the voltage across the cathode resistor to increase also, and so the output from the cathode must be *in phase* with the input, so the stage is said to be **non-inverting**.

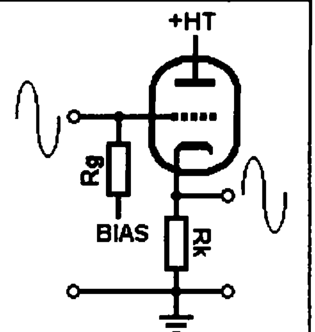

Fig. 5.1: A simplified cathode-follower circuit.

Gain:

The cathode follower is a special case of an amplifier with 100% negative (cathode current) feedback [see chapter 9]; all of the output is returned to the input. The feedback fraction, β, is therefore equal to 1 and the gain of the stage becomes equal to Ao / (1 + Ao). The gain of an ordinary stage with an unbypassed cathode resistor is given by formula IV:

$$A = \frac{\mu Ra}{Ra + r_a + Rk(\mu + 1)}$$

But in this case there is no anode resistor, and the signal is taken across Rk, yielding:

$$A = \frac{\mu . Rk}{r_a + Rk(\mu + 1)} \quad \text{.......................XXVIII}$$

$$A = \frac{\mu}{\mu + 1} . \left(\frac{Rk}{Rk + r_a / (\mu + 1)} \right)$$

But for most valves μ is much greater than 1, so to a close approximation:

$$A \approx \frac{\mu}{\mu + 1} \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{XXIX}$$

From which it is clear that the voltage gain of a cathode follower is always somewhat less than 1.

Because of the very high level of feedback, harmonic distortion and noise are reduced by a factor of $\mu + 1$ while bandwidth is extended by the same factor. Since a single-ended triode produces very little harmonic distortion even without feedback, a cathode follower becomes an extremely linear amplifier indeed, with a very wide bandwidth, making it invaluable for hifi applications. What's more, it retains this extremely linear amplification even when the valve is very worn out.
Because the output is in-phase with, and nearly identical to the input signal, the cathode is said to follow the grid, from which we get the name 'cathode follower'.
At first glace we might assume the cathode follower is of little use for guitar since its gain cannot exceed unity. However, it is wrong to say the cathode follower has 'no gain', *all* circuits have gain, even if it is technically a loss (which is simply a gain of less than 1). Instead, the cathode follower offers *current* gain and many other useful properties, as will be shown.

Input capacitance:

To see why bandwidth is increased we might consider the input capacitance. In a normal gain stage the Miller effect results in a significant increase in input capacitance by multiplying Cga by the gain of the stage. In a cathode follower, however, the anode is fixed at HT potential while Cgk is effectively *divided* by the action of feedback, giving:

$$Cin = Cga + Cgk(1 - A) \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \text{XXX}$$

But since Cgk $(1 - A)$ is extremely small, for audio applications this may be simplified to:

$$Cin \approx Cga$$

For an ECC83 this is only 1.6pF, so we can expect the frequency response to extend very far indeed.

Input impedance:

In the circuit in fig. 4.1 the input impedance is simply equal to the grid-leak resistor, which could be any convenient value, provided it does not exceed the valve's ratings. In other circuit arrangements, which will be discussed, the input impedance can be made very great indeed, or even infinite if it is directly coupled to the previous stage.

Output impedance:

It was noted in chapter 1 that when looking into the anode, impedances below the cathode appear multiplied by μ+1. However, when looking into the cathode, impedances *above* the cathode appear *divided* by μ+1. From this is it easy to show that the output impedance of the cathode itself, known as the cathode impedance, r_k, is:

$$r_k = \frac{Ra + r_a}{\mu + 1} \quad \text{.........XXXI}$$

Usually there is no anode resistor, Ra, so the cathode impedance is simply $r_a / (\mu + 1)$. For an ECC83 this would be in the region of 640Ω. As far as AC is concerned, this appears in parallel with the cathode resistor, Rk, so the output impedance of the whole circuit is:

$$Zout = \frac{r_a}{\mu + 1} \parallel Rk \quad \text{.........XXXII}$$

$$Zout = r_k \parallel Rk$$

But in most cases Rk is much larger than r_k, so may be ignored. Furthermore, for all normal valves μ is much greater than 1, and since $\mu / r_a = g_m$, then $r_a / \mu = 1 / g_m$ and so the output impedance may be approximated as:

$$Zo \approx r_k \approx 1/g_m \quad \text{.........XXXIII}$$

From which it is clear that the higher the transconductance, the lower the output impedance.
This low output impedance suggests the cathode follower should be capable of sourcing significant current from the power supply, if need be, to dump into the following circuit. This is one of the most useful properties of the cathode follower since it allows it to drive relatively heavy and varying loads without gross attenuation of the signal, known as **insertion loss**. Additionally, the low output impedance is ideal for driving long cables since any noise induced in the cable will be shunted by the low output impedance of the cathode follower, maintaining a good signal-to-noise ratio.
This is why the cathode follower is often described as an **impedance buffer**; it can offer a very high input impedance to the previous stage and a low source impedance to the following stage. It may act as an intermediate stage between a high impedance circuit and a low impedance or noise-sensitive circuit, without adding any further significant distortion or noise itself.

Graphical Analysis:

Before investigating the sort of cathode followers which are used in most guitar amps, it is important to understand first how 'textbook' cathode followers are designed.
If we are most interested in having a very low output impedance and good current-driving capability then we would naturally choose a high-current, high-gm valve. Fig. 5.2 shows a simple arrangement using an ECC82 / 12AU7, which has good current handling for a preamp

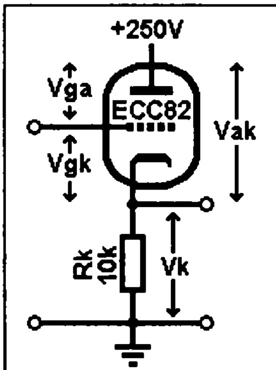

Fig. 5.2: Relative voltage notations.

triode. In this case the HT is 250V, though the cathode follower is fairly immune to wildly varying supply voltages.
Our main load resistor is the cathode resistor, Rk, so to choose it we draw a load line in a similar manner to choosing an anode resistor for an 'ordinary' gain stage. Since a cathode follower is primarily a voltage-to-current amplifier, or **transconductance amplifier**, we would usually use a relatively low value of resistance so that we can select a high quiescent anode current. In this case we might choose 10kΩ, and the load line is drawn in fig. 5.3.
However, we cannot read values such as gain and signal swing directly from this graph since all values on the graph are given *relative to the cathode*, but in this case the cathode voltage is not constant, since it follows the grid, so anything measured relative to it must also be varying. Instead, we can re-draw the curves to represent a new valve with the same characteristics as the cathode follower[1]. Since the anode voltage is fixed at the HT it makes sense to measure all voltages relative to the anode, so the grid curves will be re-drawn to equal Vga, rather than Vgk. In practice it is not necessary to draw new curves when designing a circuit, but they do aid understanding.
Fig. 5.2 indicates that Vga = Vak – Vgk , therefore Vak = Vga + Vgk.
So, if Vgk = 0 then $Vak_{(cathode\ follower)}$ must be equal to Vak on the original graph also.

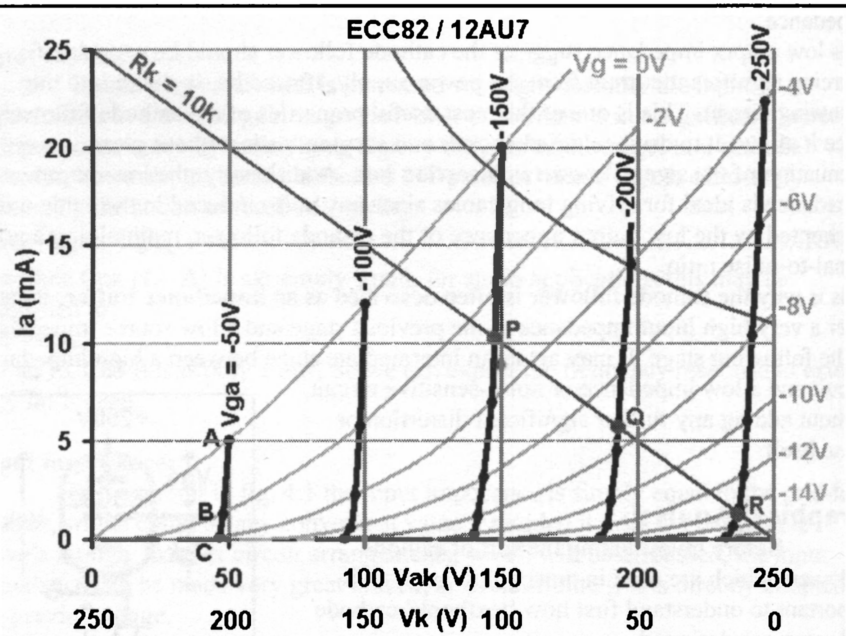

Fig. 5.3: Static anode characteristics graph, modified to show the operation of the cathode follower in fig. 5.2.

[1] 'Cathode Ray' (1955). Cathode Followers – with Particular Reference to Grid Bias Arrangements. *Wireless World*, (June). pp292–296.

For example, if Vak = 50V and Vgk = 0V, then:
$Vak_{(cathode\ follower)} = 50 - 0$
$= 50V$
And this point may be plotted on the original 0V grid curve, giving point A in fig. 5.3.

If Vak = 50V and Vgk = -2V, then:
$Vak_{(cathode\ follower)} = 50 - 2$
$= 48V$
and we plot point B on the original -2V grid curve.

If Vak = 50V and Vgk = -4V, then:
$Vak_{(cathode\ follower)} = 50 - 4$
$= 46V$
And we plot point C. Joining these points up we obtain a new Vak = -50V grid curve for the cathode follower. Other curves can be plotted in exactly the same fashion, as shown.

The most obvious change is that the new curves are almost vertical, indicating the effective anode resistance of the valve has been heavily reduced to roughly $1/g_m$, or about 330Ω in this case. Secondly, the curves are very evenly spaced, even near the bottom of the load line, and are quite straight, suggesting very low distortion, which is exactly what we expect.
Additionally, it is useful know what the *actual* voltage at the cathode will be, so a second voltage scale has been added to the abscissa. If the full HT appears across the valve (anode to cathode) then the cathode must be at zero volts. If the valve were a short circuit then the cathode voltage would be equal to the HT, so the new voltage scale for Vk is simply 0V to HT, but reversed.
With this new graph we can 'see' how the cathode follower works more clearly. For the sake of simplicity we will choose a bias point which is exactly on the Vga = 200V curve, labeled Q. This indicates that our quiescent grid voltage must be 200V more negative than the anode (HT), or 250 – 200 = 50V. The *bias* voltage, Vgk, is easily read off from the old grid curves, and is about -7V. Alternatively we may simply read off the cathode voltage directly from the scale which was added to the bottom of the graph, which indicates 57V, so the two methods are in agreement. The voltage relationships are shown in fig. 5.4, to make them obvious.
If we input a positive signal then the grid moves positive, and so the grid-to-anode voltage must be reduced. If we input a 100Vp-p signal, say, then the operating point swings along the load line between points P and R. The cathode voltage swings between about 147V and 236V, a total swing of 89V. The gain is therefore:
Vout/Vin = 89/100 = 8.9, so the gain is slightly less than

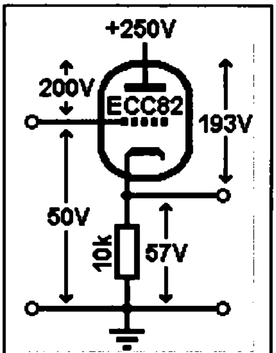

Fig. 5.4: Voltage relationships indicated by the load line in fig. 4.3.

unity, as we would expect. Increasing the load resistance would increase the gain so that it approaches $\mu/(\mu+1)$, or about 0.95 for the ECC82, though in practice such trifling figures are not usually important.
In this case the position of the bias point, Q, shows that the valve will reach cut-off before grid-current limiting, and that the maximum unclipped output signal is about 115Vp-p or, in other words, the input sensitivity of the stage is 115/0.89 = 129Vp-p. Of course, maximum, unclipped output swing is obtain when the bias point is half way between grid-current limiting and cutoff, at which point the maximum theoretical output swing becomes[2]:

$$V_{O(max\ p\text{-}p)} = \frac{HT}{1 + r_a / R_k} \qquad \text{XXXIV}$$

If the cathode follower is very heavily loaded by an impedance, Rl, then Rk in the above formula should be replaced by Rk||Rl, which indicates that its low output impedance does not necessarily mean it can drive large signal swings into heavy loads.
Provided it is not heavily loaded, however, the input sensitivity of a cathode follower is very large, making it very difficult to overdrive. Also, although the cut-off characteristics of the cathode follower are broadly similar to those of a normal gain stage, distortion due to grid current is quite different: As we make the grid more positive the cathode also moves more positive, since it always tries to follow the grid, so that the grid must 'catch up' with the cathode before grid current will flow. What's more, once we reach this threshold, grid current flows down through the cathode resistor, and so contributes to the output signal's positive swing! The result is that the positive peaks of the output signal begin to compress softly, long before they can be made to 'clip', or flatten off. In a guitar amp this compressive feature can be used to 'soften' or 'fatten up' the tone, and this will be discussed in detail later.

Biasing arrangements for AC-coupled cathode followers:

So far we have analysed the performance of the valve, but we have yet to complete the circuit. We know from the load line in fig. 5.3 that the quiescent grid voltage is 50V, and the cathode voltage is 57V, so all that remains is to set up these bias conditions.

Fixed bias:

The simplest arrangement is to apply 50V directly to the grid via a potential divider from the HT, as shown in fig. 5.5*a*. Since it is usually desirable to keep the input impedance high, large resistors must be used. In this case R2 was chosen to be 470kΩ, so for a grid voltage, Vg, of 50V, R1 must be:

$$R1 = \frac{HT - Vg}{Vg / R2} = \frac{250 - 50}{50 / 470k} = 1880k\Omega$$

[2] Valley, E. G. and Wallman, H. (1948) *Vacuum Tube Amplifiers*, p349. McGraw-Hill Book Company, Inc., London.

1.8MΩ is a very close standard, yielding an actual grid voltage of:
250 x [470k / (1800k + 470k)] = 52V, which is sufficiently close.

There are two obvious drawbacks to circuit *a*. The first is that the input impedance is equal to R1||R2, or 373kΩ in this case, which is fairly low, though could be overcome by using larger values for R1 and R2. The second problem is that any noise on the HT is coupled directly to the grid (albeit attenuated). Of course, if the HT is well filtered or the audio signal level is always very large, then this may not matter.

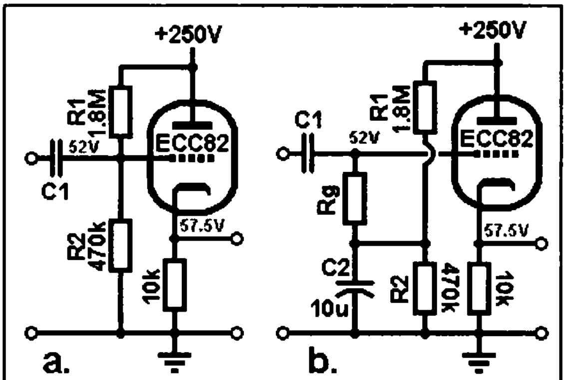

Fig. 5.5: Fixed-biased cathode followers. Circuit **b.** uses C2 to remove and power supply noise.

An input coupling capacitor, C1, is of course required to 'block' the bias voltage from interfering with the previous stage, and its value is chosen in conjunction with the input impedance, in the same way as for any coupling capacitor.
If HT noise *is* a concern then the circuit could be modified to ensure that the 52V bias is itself very well filtered. The circuit in fig. 5.5*b* shows this. The bias voltage provided by the potential divider is heavily filtered by C2, and this bias voltage is applied to the grid via the grid-leak, Rg. The grid-leak could be made any desired value, providing it does not exceed the datasheet maximum, so we again have control over the input impedance of the stage. The minimum required value of C2 (to obtain good filtering down to 1Hz) is given by:

$$C = \frac{1}{2\pi R}$$

Where:
$R = R1||R2||Rg$

But an arbitrarily large value of 10μF will normally suffice.
Grid-stoppers have been omitted for clarity, but are always recommended. In this case the grid voltage is only 52V so there is little risk of arcing when the circuits are first switched on, but if the design calls for a grid voltage in excess of 100V then a protection diode and resistor ought to be added [see chapter 2, *DC coupling*].

Cathode bias:

The fixed-bias arrangements described above are not often used, since cathode biasing usually requires fewer components and has the advantage that it will self-adjust as the valve ages. Cathode biasing also offers a *much* higher input impedance, and avoids the possibility of arcing between grid and cathode when the circuit is first switched on, since the grid voltage will rise with the cathode voltage as the cathode warms up.

Cathode biasing a cathode follower is easily done by 'tapping off' the necessary grid voltage from the cathode resistor, as shown in fig. 5.6, thereby placing the grid at a lower potential than the cathode by the desired amount. The potentiometer is shown for illustrative purposes; in reality a bias resistor is used, as shown in fig. 5.7. The bias resistor, Rb, may be found in exactly the same way as for a normal gain stage, either by drawing a cathode load line or simply by calculating its value from the bias point.
Referring to fig. 5.3, in this case the bias was found to be around 7V, and the quiescent anode current is about 6mA. The bias resistor would therefore be:
R = V/I = 7 / 0.006 = 1.17kΩ.
The nearest standard is 1kΩ. Of course, the *total* cathode load resistance is increased by the addition of this resistor, which will alter the load line somewhat, so we might choose to make Rk equal to 9kΩ so that the total resistance is 10kΩ again, but in practice the difference will be negligible.

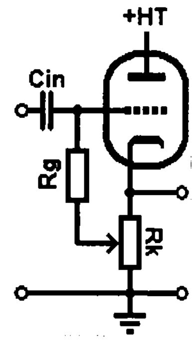

Fig. 5.6: The necessary grid voltage may be 'tapped off' the cathode resistor.

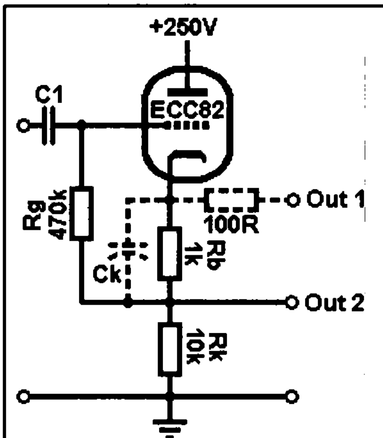

Fig. 5.7: A practical cathode biased, cathode follower.

Fig. 5.7 also shows how the output signal might be taken directly from the cathode (output 1) or from the junction of Rb and Rk (output 2). Taking the output directly from the cathode is the traditional arrangement, but it must be impressed that because of the very high level of negative feedback present within the cathode follower it is susceptible to ringing or even oscillation if loaded by an unfortunate combination of stray capacitance and inductance (this is the principle of the cathode-follower oscillator). If Output 1 is used then it is highly recommended that a **build-out resistor** of 100Ω or more be added in series with the output as shown, soldered close to the valve socket, to isolate the cathode from stray impedances.*

* It might be thought that a cathode follower would be immune to stray impedances since its gain is assumed to be less than unity. However, with the right (or rather, wrong!) combination of inductance and capacitance in the grid or cathode circuits it is entirely possible to make a cathode-follower oscillate[3]. Consequently, a grid stopper and possibly a build-out resistor are still necessary to protect against HF instability.
[3]Schlesinger, K. (1945). Cathode -Follower Circuits. *Proceedings of the I.R.E*, (December), pp843 – 855.

Output 2 has the disadvantage that the signal will be slightly attenuated, since Rb and Rk form a potential divider, though Rk will normally be much larger than Rb so the degree of attenuation will only be small. However, Output 2 has the advantage that Rb does the same job as a build-out resistor, isolating the cathode from stray impedances. Because this eliminates the need for an extra component, output 2 is to be preferred.

It might be thought that the input impedance of this circuit would be equal to the grid-leak, Rg, but in fact it is much greater. The reason for this is that the AC signal at the cathode is identical, in phase, and only fractionally lower in amplitude than the input signal. Obviously, the same signal also then appears at the junction of Rb and Rl, albeit slightly attenuated due to the potential divider formed by Rb and Rk. What this means is that at the bottom of the Rg there is an AC signal, identical in phase, and only slightly lower in amplitude than the input signal which is at the top of Rg. Therefore the difference in voltage between the top and bottom of Rg is always very small, much smaller than if it were connected directly to ground. As a result, negligible signal current is lost through the grid-leak resistor, thereby increasing the effective input impedance of the stage hugely. This effect is known as **bootstrapping**; the value of the resistor appears to be much greater at AC than its true DC resistance, the circuit is improving its own performance, 'pulling itself up by its bootstraps'. The input impedance thus becomes equal to:

$$Zin = \frac{Rg}{1 - \frac{A.Rk}{Rk + Rb}} \quad \text{..........XXXV}$$

A further advantage of this is that we are free to use a smaller value for the grid-leak than we normally would, without loading the previous stage heavily, and thereby reduce resistor noise. In fig. 5.7 a value of 470kΩ is used, and taking the gain of 0.89 indicated by the load line in fig. 5.3, the input impedance becomes:

$$Zin = \frac{Rg}{1 - \frac{A.Rk}{Rk + Rb}} = \frac{470k}{1 - \frac{0.89 \times 10k}{1k + 10k}}$$

= 2.5MΩ.

This value is so large that the input capacitor, C1, need only be small to pass all audible frequencies. However, this type of biasing suffers a problem as it assumes that all of the signal voltage appearing at the junction of Rb and Rk actually reaches the grid. In reality the output impedance of the previous stage forms a potential divider with Rg, so that only a portion of the cathode voltage actually reaches the grid, reducing the feedback factor and therefore the 'ideal' values for output impedance and distortion may not be achieved. Fortunately this is of little concern for guitar purposes, except that it is advisable to make C1 larger than might be thought necessary, to avoid unexpected loss of bass. Usually it will be whatever convenient value is already to hand, say 10nF. Bass frequencies may be more accurately controlled using the output coupling capacitor instead.

In old-fashioned designs it is not unusual to see a bypass capacitor, Ck, (shown dashed) connected in parallel with Rb. The reason for this is that by bypassing Rb the signal is no longer attenuated by it, and the input impedance is increased to:

$$Zin = \frac{Rg}{1-A} \quad \text{.......................XXXVI}$$

or about 4.3MΩ in this case. However, the capacitor would need to be large, implying an electrolytic, which is nearly always detrimental to tone, so this component is not usually found in modern circuits. If the input impedance needs to be higher (which is unlikely) then Rg could be increased in value. In a hifi design we might replace Rb with an LED (or several diodes in series to obtain the correct bias voltage) and so obtain the high input impedance without the need for a capacitor. An example of this is shown in fig. 5.8. An ECC83 is used in this case since it requires only a small bias voltage, which is easily provided by one red LED (~1.6V). The LED could even be mounted on the front panel of the amplifier to serve as power-on indicator. Since the LED is effectively a short circuit at AC, a build-out resistor will be required at either output, and a grid stopper has also been added to complete the circuit. The gain of this follower is about 0.92 and the input impedance is roughly 6MΩ. This circuit is particularly useful as an input circuit for peizo pickups, used in many electro-acoustic guitars, which can sound rather bright and 'metallic' unless they are married with a very high input impedance.

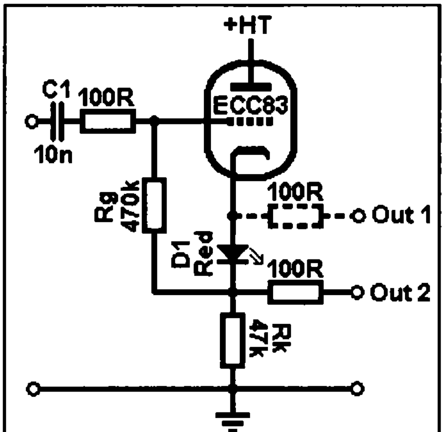

Fig. 5.8: An LED-biased cathode follower with very-high input impedance.

DC-coupled cathode followers:

A cathode follower is an obvious candidate for DC coupling, since its grid is naturally required to be at a high voltage. Similar design procedure considerations apply as were described in chapter 2, *DC Coupling*, and need not be repeated. Again, some juggling of valves and values may be required to produce a suitable circuit. For example, the ECC82 with 10kΩ cathode load shown earlier requires a grid voltage of around 50V. Obviously, the anode of the preceding stage is unlikely to rest at such a low voltage, even if it is a pentode, but level shifting is a possibility. Fig. 5.9 shows a theoretical example in which the preceding stage has a quiescent anode voltage of 100V. This is level shifted down to 50V and a treble-boosting capacitor could also be added (shown dashed), and may be found using formula XIV. In this case a value of 4.7nF would give a half-boost frequency of about 200Hz. A very similar circuit is found in the Sound City *LB* series of amplifiers.

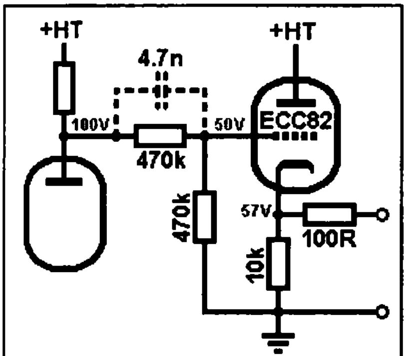

Fig. 5.9: A level-shifted, DC-coupled cathode follower.

Of course, direct coupling of the grid to the previous stage's anode is desirable from the point of view of simplicity, and in a guitar amp this is the most common arrangement, and can produce some unexpected results.

For 'textbook' direct coupling we design the cathode follower so that the necessary grid voltage matches the anode voltage of the previous stage. This may require a valve with a lower r_a or a large value for Rk, or both. The task is made easier if the previous stage naturally has a low anode voltage of course, which implies either a low r_a triode or a pentode.

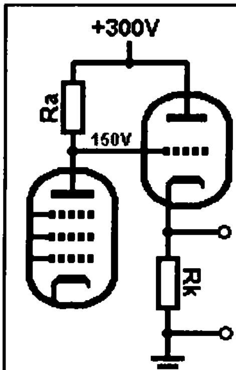

Fig. 5.10: The low anode voltage of a pentode is well suited to DC coupling.

For example, fig. 5.10 shows a circuit in which the preceding valve is a pentode whose quiescent anode voltage is ½HT, or 150V. To show the advantages of using a low r_a triode for the cathode follower, the static anode characteristics of the ECC83 and ECC82 are shown in fig. 5.11. The cathode-voltage scale has been added and note that the only 'new grid curve' which needs to be drawn is the one corresponding to the known grid voltage of 150V. It is immediately clear that the ECC83 would require a high value for Rk in order to bias properly. Even with a 100kΩ load, it is still very warm biased at Vgk = -0.75V (indicated by the dot on the load line) and the implications of this are discussed later. The ECC82, on the other hand, has a much lower r_a and would be roughly centre biased with a 100kΩ load, maximizing output signal swing, but would also operate well under a 22kΩ load, if maximum signal swing were not required. Clearly the ECC82 is the obvious choice. The completed cathode follower is shown in fig. 5.12, and a compromise value of 47kΩ has been used for Rk, to allow both reasonable transient response and output signal swing. A grid-stopper and build-out resistor have been added to ensure HF stability, and a diode and series resistor are added to protect against arcing at switch-on.

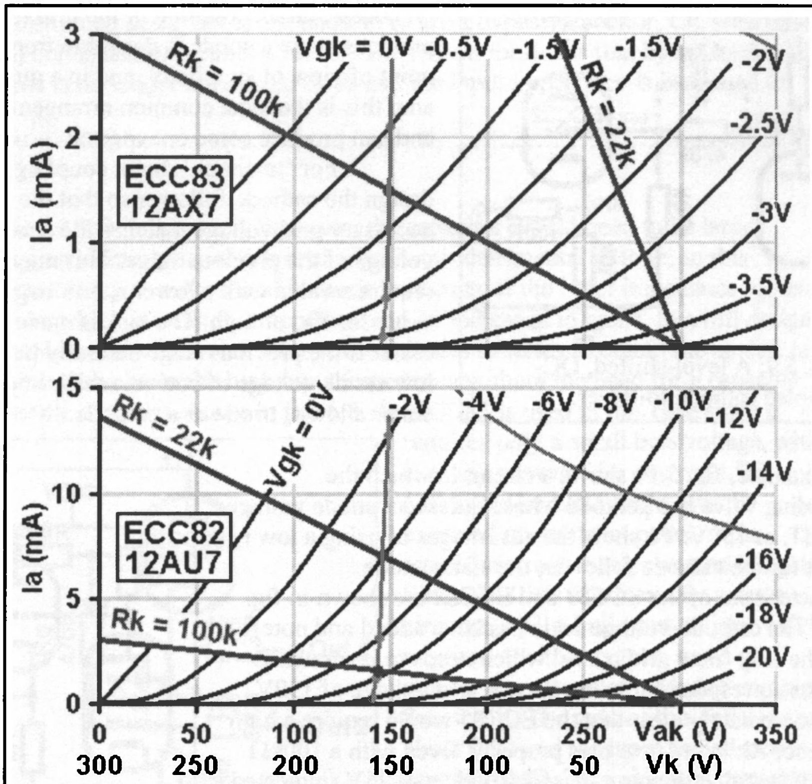

Fig. 5.11: Comparison of the ECC82 and ECC83 when used in the circuit of fig. 5.10.

The combination of pentode and cathode follower is very common in valve audio designs, since a pentode offers high gain but has a high output impedance, while the cathode follower offers a near infinite input impedance and so buffers the pentode from the following circuit. There are also several valves containing one small-signal pentode and one triode in a single envelope, such as the ECF80 / 6BL8 (which is in current production), ECF82 / 6U8, ECF802, 7199, 6EA8, 6AW8, 6LN8 to name but a few. It is not uncommon to see such valves connected in the manner described, and fig. 5.13 shows a practical example using the ECF80 (the input grid-leak and associated components are omitted since they would be chosen to suit the preceding circuitry). In a guitar amp the cathode follower would probably be used to drive a tone stack.

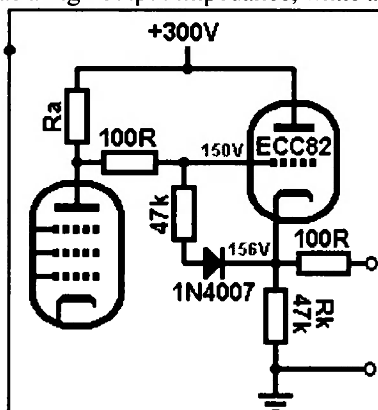

Fig. 5.12: A cathode follower with large output signal swing and arc protection.

It should also be remembered from chapter 2 that if the cathode voltage is high, it may be necessary to elevate the heater supply to avoid exceeding the valve's maximum heater-to-cathode voltage rating [see chapter 12]. For the ECF80, $Vhk_{(max)}$ is 100V, so in the circuit of fig. 5.13 the heater should be elevating by around 30V to 50V, so that it lies well within 100V of the cathodes of *both* the triode and the pentode section of the valve.

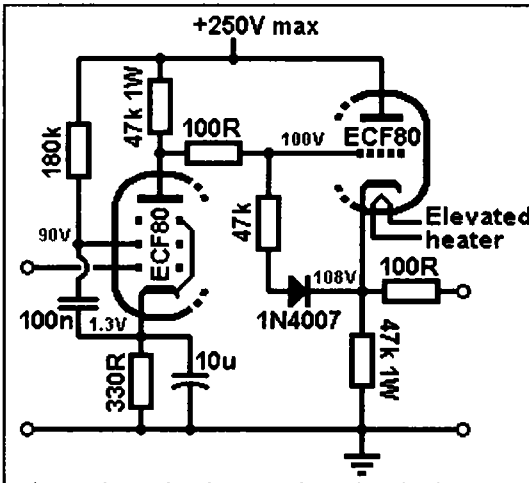

Fig. 5.13: A classic pentode and cathode-follower arrangement using the ECF80.

The pentode as a cathode follower:

A pentode can also be connected a as cathode follower, if necessary. The main advantage of a pentode is that, as with a normal gain stage, a pentode cathode follower is capable of delivering an output signal swing which is almost from HT to ground. The high μ of a pentode should also result in a more linear circuit, although the improvement when compared with a triode is largely academic.

Design follows exactly the same procedure as for a triode, except that it must be remembered that screen current also flows in the cathode resistor, and that the screen voltage is measured *relative to the cathode.* Since ground and the HT are the same as far as AC is concerned, if the screen-bypass capacitor were connected to ground the screen voltage would be fixed relative to ground, so it would appear to be connected directly to the anode, as far as AC is concerned. Thus the pentode would be effectively triode-strapped. One advantage to this arrangement, however, would be that the smooth screen voltage would shield the cathode from HT noise on the anode although, again, the improvement is usually immaterial since signals at the anode will appear divided by μ+1 at the cathode anyway.

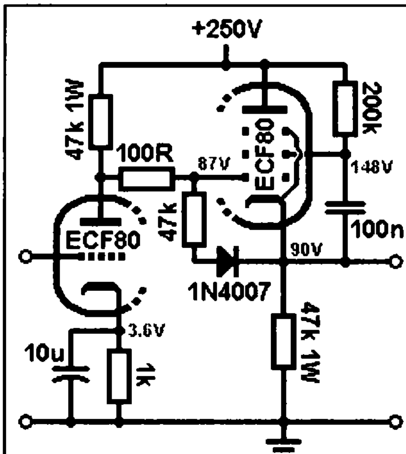

Fig. 5.14: A pentode cathode follower, using the ECF80.

To give true pentode operation the screen-to-cathode voltage must be held constant, so the screen-bypass capacitor must be connected directly to the cathode. The screen dropping resistor must always be included, otherwise the output signal from

the cathode would be shunted to the HT via the screen bypass capacitor.
True pentode cathode followers are very rarely found in valve audio, since the triode is simpler and gives almost equally good performance. Pentode cathode followers were more typical of early valve computers, where optimum linearity down to DC was essential, a circumstance we are unlikely to meet in a guitar amp! Fig. 5.14 shows an example circuit using the ECF80, in which the pentode section now serves as the cathode follower.

Cathode followers in guitar amps:

Now that we a have a proper understanding of the operation and design of cathode followers it is worth examining the role of the cathode follower in a guitar amp.

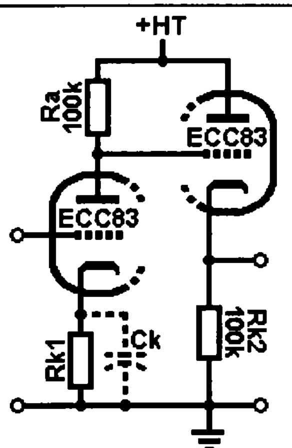

Fig. 5.15: The most common cathode-follower arrangement found in guitar amps.

In some cases a cathode follower is used simply to drive an effects loop, its sole purpose being to offer a low output impedance capable of driving the (relatively) low input impedances of solid-state effects pedals, mixers, PA equipment and so on, and to shunt noise which might be picked up on the long interconnecting cables. In such cases the cathode follower will often be of the capacitor coupled, self-biased type, and is not expected to generate any tonal effects itself.
However, anyone familiar with even a few commercial guitar amp designs will know that the most common occurrence of a cathode follower in a guitar amp is as a driver for a tone stack, and many (perhaps even most) famous designs use exactly the same arrangement, which is shown in fig. 5.15. The circuit universally consists of an ECC83 / 12AX7 gain stage, directly coupled to an ECC83 / 12AX7 cathode follower (usually in the same envelope), each with 100kΩ load resistors. The HT voltage and biasing of the gain stage may vary between designs, but the overall topology is nearly always the same. It might be though that the cathode follower's purpose is simply to act as a buffer between the gain stage and the relatively heavy load of the tone stack, while offering no tonal colouration itself and, indeed, this is almost certainly why the circuit first appeared in very early guitar amps, notably the 1959 Fender *5F6 Bassman* and its successors. However, by accident of design this particular topology is as important to the tone of the amps that use it as *any other stage,* and is partly responsible for the huge success of the revised Fender *Bassmans* and early Marshall and Vox amplifiers which also used the same arrangement, from which almost all modern amps have been derived.

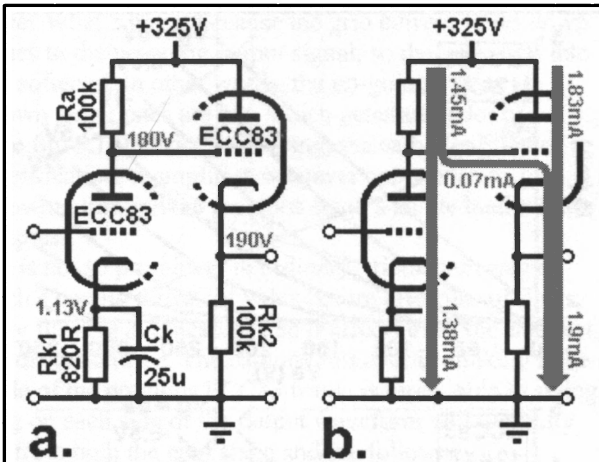

Fig. 5.16: **a.** The Fender *5F6 Bassman* circuit arrangement with measured voltages and **b.**, current paths showing quiescent grid current.

But why is this apparently innocuous circuit so tonally important? To see why it is worth examining a circuit with known voltages. The circuit in fig. 5.16*a* is taken directly from the Fender *5F6 Bassman*, with measured voltages for a Mullard ECC83. (The original circuit used a 7025 / ECC803, which is simply a high-quality version of the ECC83 / 12AX7.) From the voltages shown we discover an interesting fact: The quiescent current through the anode resistor of the gain stage, Ra, is:

$$\frac{325 - 180}{100k} = 1.45mA$$

But the current through its cathode resistor, Rk1, is:

$$\frac{1.13}{0.82k} = 1.38mA$$

We would expect the current through Ra, the first triode and Rk1 to be identical, but we find an additional 0.07mA of current is apparently missing!

Also, at a glace we can see that the bias on the cathode follower is 180 - 190 = -10V which is very large for an ECC83 (in previous chapters we have been dealing with bias points in the range of not more than -4V for this valve), so we might expect the cathode follower to be close to cut-off, but since there is 190V across its cathode load resistor this is clearly not the case.

The only place for the 'missing current' to flow to is into the grid of the cathode follower and down Rk2, increasing the voltage across it by $0.07 \times 100k = 7V$, as shown by the current paths in fig 5.16*b*.

A similar conclusion can be derived from load lines for the two triodes: The upper graph in fig. 5.17 shows the anode and cathode load lines for the gain stage. The expected, quiescent anode voltage is close to 180V (actually it is a little higher), and this is therefore the grid voltage of the cathode follower. If we initially assume that the cathode follower's grid and cathode voltage are identical we may plot this voltage on the lower graph, which has a reversed scale to show the cathode voltage, and thereby derive a rough bias point for the cathode follower. This is shown by the dashed line and reveals a bias point of around -0.55V.

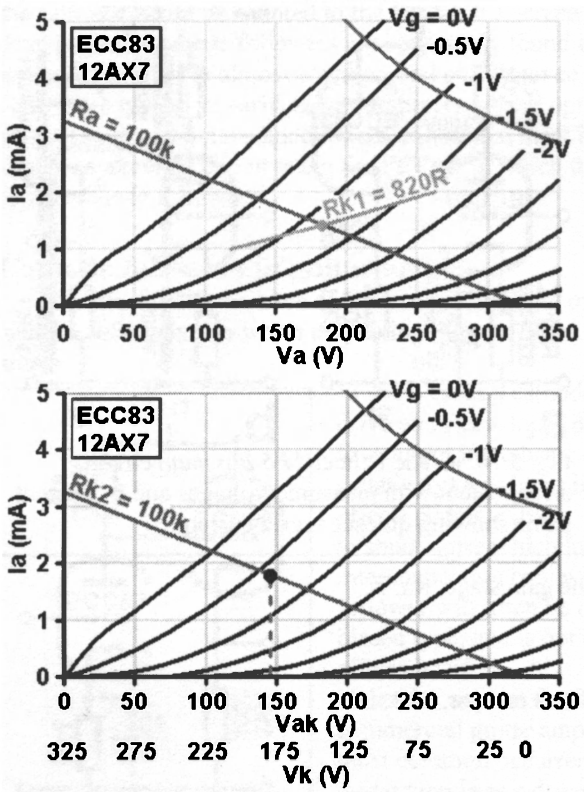

Fig. 5.17: Load lines for the gain stage and cathode follower in fig. 5.16.

However, this does not agree with the measured cathode voltage of 190V and bias of -10V. To account for this discrepancy we recall from chapter 1 that grid current begins to flow in a valve even when the bias is still slightly negative, around -0.9V, and we see from fig. 5.17 that our predicted bias point for the cathode follower is already well within this grid-current region, and considering we know that the cathode voltage must actually be a little higher than the grid, this simply pushes the bias point even further up the load line. In a normal gain stage the level of grid current in this region is too small to be of any concern when biasing, but in this case even a small amount causes a large change across the cathode load resistor, which is itself large. The current flowing into the grid and down Rk2 increases the cathode voltage and simultaneously lowers the anode voltage of the previous stage (whose bias point also adjusts slightly), effectively 'stealing' anode current from it. But since the rising cathode voltage and falling grid voltage oppose the flow of grid current, an equilibrium point will be reached where the grid current generates enough bias to prevent any further increase in grid current. In other words, the valves will come to rest with grid current *permanently* flowing into the cathode follower, and previous calculation showed this to be 0.07mA.

Having found that the cathode follower is permanently 'stealing' current from the previous stage, what is the result? If a down-going signal appears at the gain stage's anode, the grid voltage of the follower is pushed down, back into the area of "normal operation", and grid current stops flowing more or less immediately, for the duration of that cycle. But when the incoming signal is positive going the grid is pushed more positive, which induces even more grid current to flow into the cathode follower as it tries to maintain bias, which in turn 'drags down' the anode voltage of the gain stage

which is trying to move positive! What's more, because the grid current flows down the cathode resistor it contributes to the up-going output signal, so the transition into grid-current clipping is greatly softened. In other words, the up-going cycles are heavily compressed, but the down going ones are not, which generates a lot of second harmonic distortion [see fig. 5.19]. Strictly speaking, the cathode follower still gives extremely linear amplification, it amplifies whatever appears on its grid exactly, it is the grid current flowing through the previous stage's anode load resistor which causes the signal to compress.
This 'softness' of compression is not so prominent in ordinary, triode gain stages, since the signal level must reach a certain threshold value before grid current flows, whereas in this case it is already flowing at quiescence so it affects even the smallest of signals and is also blocking-distortion free. The cathode follower is unlikely to be driven to cut-off since the anode of the previous ECC83 triode is rarely able to swing much below 1/3HT, so clipping on each side of the output waveform will normally be due to grid-current limiting from both the gain stage and the follower.

This natural compression is one of the most important contributors to smooth, rich, overdriven tones, and was even identified as being 'part of the electric guitar sound' by the Audio Engineering Society[4], though they failed to identify the *quiescent* grid current in the cathode follower as the controlling factor. Indeed, very few authorities on guitar amps appreciate the importance of this topology, and it is serendipitous that its existence is entirely thanks to 'bad design': Because the ECC83 is a high r_a valve it is difficult to bias it properly if the grid voltage is high and the cathode load resistor is not sufficiently large, as any classic textbook will indicate. If the supply voltages in the Fender *Bassman* had been much lower, or if the cathode load resistor had been 'properly chosen', this effect might not have been realised, and the amp may not have become the icon it is today. Similarly, if a lower r_a valve were plugged into this same circuit we probably wouldn't generate any quiescent grid current and again we would not obtain the compression effect (of course, the effect *can* be duplicated with any valve, if resistor values are suitably chosen.)
In a hifi amp we would desperately avoid this grid current by lowering the grid voltage; by reducing the HT, increasing the anode resistor of the previous stage or by biasing the previous stage hotter. Alternatively we could increase the cathode load resistor, which would make the load line steeper and bring the bias point out of the grid-current region. For guitar purposes, on the other hand, it is desirable to take advantage of this compression effect, and even increase it. This could be done by increasing the grid voltage by some means. However, it must be remembered that if the grid is passing current then it must be dissipating power, which the grid is not designed for. The maximum dissipation rating of the grid in a conventional preamp triode is rarely quoted, but is unlikely to exceed 0.2W, so increasing the grid voltage too much could cause it to overdissipate and melt, and this may be a contributing factor in the failure of DC coupled stages (see chapter 2, *DC coupling*). Using a high

[4] Rutt, T. E., (1984). Vacuum Tube Nonlinearity as Part of the Electric Guitar Sound. Presented at the Convention of the Audio Engineering Society, 8/11 October 1984, New York

grid voltage also implies a high cathode voltage, which puts greater strain on the heater-to-cathode insulation. The Mesa Boogie *Dual Rectifier*, for example, operates its cathode followers with cathode voltages in excess of 200V, well above their optimistic 180V rating, which greatly increases noise due to leakage between heater and cathode.

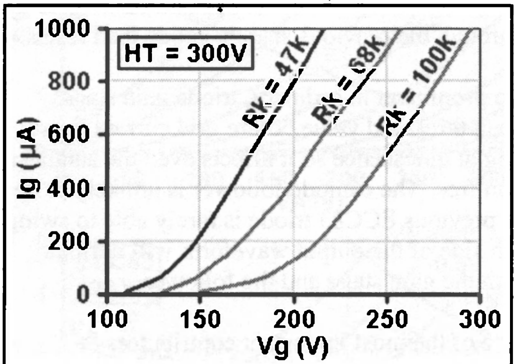

Fig. 5.18: Grid-current measured in a Mullard ECC83 cathode followerwith different values of cathode load with a 300V HT.

The safer method of increasing the grid current is to reduce the cathode load resistor, Rk2, and this has the added benefit of reducing the output impedance of the follower.

To illustrate this, fig. 5.18 shows the quiescent grid current measured in a DC coupled, ECC83 cathode follower, operating from an HT of 300V with different values of cathode load resistor. This clearly shows that for a given grid voltage the quiescent grid current can be increased by reducing Rk, since more current must flow into the grid to reach the equilibrium point. The result of this increase is shown in fig. 5.19, which shows oscillograms of the output signal from the cathode follower at 1kHz, approximately 55Vp-p, and the corresponding input signal to the gain stage's grid, which has been inverted and scaled for comparison. The upper trace is for a cathode load resistance of 100kΩ, and the compression of the up-going part of the waveform is clearly visible. The lower trace shows that by reducing the cathode load to 47kΩ the compression is increased. Although visually subtle, this actually has a dramatic effect on 'smoothing out' any brittle sounding distortion which may have been generated earlier in the preamp. In an existing amp having this cathode follower arrangement, lowering the cathode load resistor to around 47kΩ to 82kΩ is a remarkably useful way of increasing the sustain and overall 'crunchiness' of the overdriven tone, and is highly recommended.

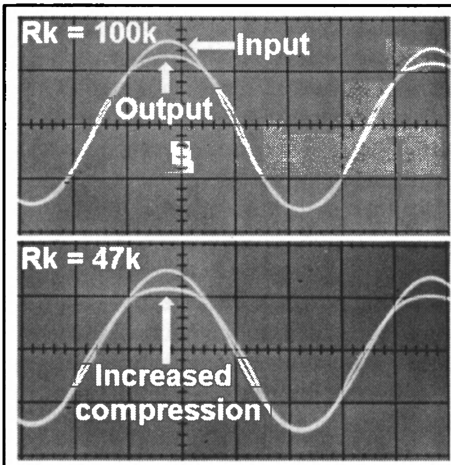

Fig. 5.19: Oscillograms showing the increase in compression caused by reducing the cathode load resistor.

It should always be remembered that the heater may need to be elevated to avoid exceeding the valve's maximum heater-to-cathode voltage rating, and a

100Ω grid-stopper and arc protection [fig. 5.12] is also advised if long life is to be expected, especially if the HT is greater than about 300V (the arc protection circuit has no detrimental effect on the tone).

This topology of an ECC83 gain stage, DC coupled to a cathode follower, is most important in high-gain designs, and sometimes appears multiple times in a high-gain circuits. For example, the Soldano *SLO100* uses two cascaded (with a serial effects loop inserted between), while the Marshall 30th anniversary model uses one in its 'crunch' channel and *three* cascaded in its lead channel! It is also interesting to note that many Vox amps, including the AC30, used this topology with a cathode load resistor of 56kΩ and a grid voltage of 180V, resulting in considerable compression. This is another good example of how numerical gain, in itself, is not necessary for a 'high gain' sound. Even the humble cathode follower, with less than unity gain, can contribute heavily to the overdriven tone of an amp.

Bootstrapping for more gain:

It has been illustrated how the DC coupled cathode follower can assist in generating a good overdriven tone, but there is yet more performance we can extract from it. By taking advantage of its very low impedance output, we can *increase* the gain of the previous stage! This can be done by connecting the output of the cathode follower to a tapping point on the previous stage's anode resistor, via a coupling capacitor. This form of bootstrapping is commonplace in transistor amplifiers, but is surprisingly under-used in valve designs, despite its obvious advantages.

Fig. 5.20*a* shows the usual arrangement with relative AC signal voltages. The AC signal is greatest at the anode of V1 and zero at the power supply. The signal 'half way along' the anode resistor must therefore have half the amplitude of the signal at the anode. The circuit in fig. 5.20*b* shows the bootstrapped arrangement. The anode resistor of V1 has now been split into two parts, R1, R2, and the output of the cathode follower is coupled to the junction via C1. Assuming C1 is large, and the cathode follower is perfect and has unity gain,

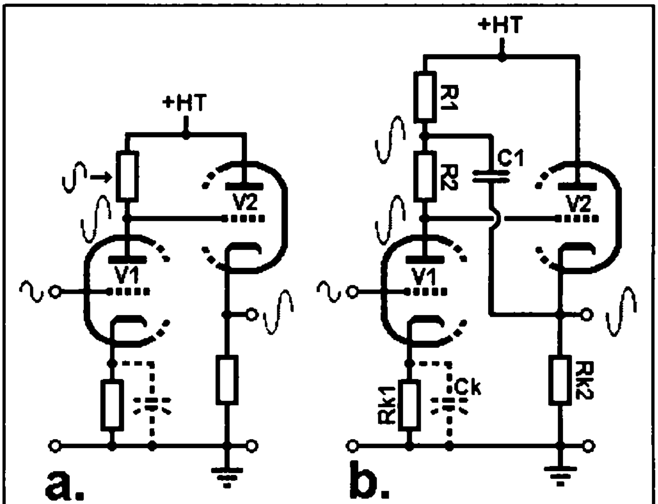

Fig. 5.20: **a.** the usual DC coupled cathode follower arrangement. **b.** bootstrapping the previous stage's anode resistor to increase its gain.

then the signal at the anode of V1 being fed to the cathode follower is immediately buffered and passed back to the junction of R1, R2. Therefore the same signal voltage appears at both ends of R2. Since there is now no difference in AC voltage across this resistor there can be no change in AC current through it, and it appears to have infinite resistance! We now see that R2 has been bootstrapped, and from our understanding of load lines, if V1 operates into an infinitely large load then its gain becomes equal to μ.
R1, however, is not bootstrapped. If the anode resistor has been split equally then R1 now sees twice the difference in AC voltage across it, and therefore twice the AC current through it, compared with circuit *a*. This current must be sunk by the cathode follower, which is why we require its low output impedance.

In reality, cathode followers are not perfect, so we will not quite obtain a gain of μ from the preceding stage. At AC, the value of R1 is effectively increased by bootstrapping to a value of:

$$r = \frac{R1}{1 - A}$$

Where:
A = the gain of the cathode follower.

So the total anode load on V1 is r + R1. If the cathode resistor Rk1 is bypassed, we achieve a gain for V1 of:

$$A = \frac{\mu\left(R1 + \frac{R2}{1-A}\right)}{r_a + R1 + \frac{R2}{1-A}}$$

If Rk1 is not bypassed:

$$A = \frac{\mu\left(R1 + \frac{R2}{1-A}\right)}{r_a + R1 + \frac{R2}{1-A} + Rk(\mu + 1)}$$

Of course, the same principles can be applied with any type of cathode follower, though the DC coupled arrangement gives the best (and simplest) performance. V1 could be any valve, even a pentode. However, bootstrapping a pentode may result in so much gain as to be unusable!

Practical designs:

Assuming most of the circuit has already been designed in the normal way, adding the bootstrapping is trivial, and easily added to existing circuits too. Fig. 5.21 shows the familiar arrangement using an ECC83, which is to be bootstrapped.

When the load of V1 is split into two parts, R1 and R2, the ratio of these two resistors is not critical. There *will* be an optimum ratio which results in the most gain, but deriving the necessary values is unnecessarily lengthy. Instead it is simpler to make R1 equal to R2, since this normally yields almost as much gain as the 'optimum' values anyway. It must be realised, however, that the act of bootstrapping places additional load on the cathode follower, the total cathode load being equal to:
$Rk2 \| R1 \| (R2+r_a)$
(Assuming Rk1 is fully bypassed).

This does not take into account the further load added by the following circuit, which also appears in parallel with Rk2! If we make the load on the cathode follower too great, any increase in the gain of V1 will be useless since the gain and output signal swing of the cathode follower will decrease accordingly. To avoid this, as a rule of thumb R1+R2 should be at least equal to Rk2, and R1 should be at least 47kΩ.

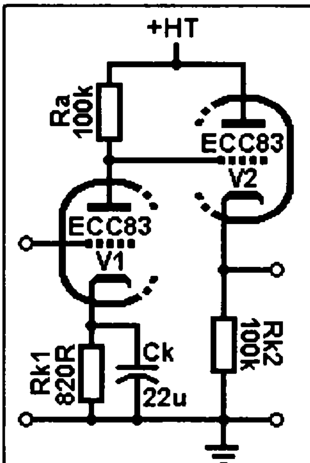

Fig. 5.21: The conventional ECC83 gain stage and cathode-follower arrangement has a total gain of around 60 (35.6dB).

In this case Ra is 100kΩ, which may easily be divided into two equal parts of 50kΩ. The standard value of 47kΩ is close enough to this, and should be large enough to avoid very heavy loading of the cathode follower. If Rk1 is bypassed, then taking a gain for the cathode follower of 0.95, the gain of V1 will be:

$$A = \frac{\mu\left(R1 + \frac{R2}{1-A}\right)}{r_a + R1 + \frac{R2}{1-A}} = \frac{100\left(47k + \frac{47k}{1-0.95}\right)}{65k + 47k + \frac{47k}{1-0.95}}$$

$= 94.$

But if Rk is left *unbypassed* then we would obtain a gain of:

$$A = \frac{\mu\left(R1 + \frac{R2}{1-A}\right)}{r_a + R1 + \frac{R2}{1-A} + Rk(\mu+1)} = \frac{100\left(47k + \frac{47k}{1-0.95}\right)}{65k + 47k + \frac{47k}{1-0.95} + (0.82 \text{ x } (100+1))}$$

$= 87.$

This shows that the cathode bypass capacitor now has much less effect on the gain of the circuit, thanks to the bootstrapping. In many cases this can be highly advantageous since it allows us to omit the cathode bypass capacitor for a 'crunchier', less fuzzy-sounding overdrive from V1 while still obtaining a high level of gain with which to overdrive following stages.

The coupling capacitor, C1, is required to block DC. In conventional circuits its value is made large so that R2 is bootstrapped at all audio frequencies, although for guitar purposes this doesn't have to be the case. Ignoring the output impedance of the cathode follower which is negligible, the resistance in series with C1 is R1||(R2+r_a). To a reasonable approximation, the -3dB roll-off frequency due to the bootstrapping will occur at:

$$f \approx \frac{1}{2\pi RC}$$

Where:
R = R1||(R2+r_a).
C = coupling capacitor C1.

To reduce the gain at bass frequencies we might take a frequency of 100Hz:

$$R1 \,||\, (R2 + r_a) = \frac{R1(R2 + r_a)}{R1 + R2 + r_a} = \frac{47k \times (47k + 65k)}{47k + 47k + 65k}$$

= 33kΩ.

$$C \approx \frac{1}{2\pi \times 33k \times 100}$$

= 48nF.

So 47nF is a close standard, and should be rated for the full HT voltage. The completed circuit is shown in fig. 5.22 (the necessary grid-stopper and arc protection are omitted for clarity). The circuit was tested with a Mullard ECC83 and found to have a gain of between 90 and 95, depending on the biasing of V1. This is a considerable increase in gain at very little cost! The frequency response of the whole circuit, with different values of C1 and with Rk1 fully bypassed, is shown in fig. 5.23. This clearly shows that the 'un-boosted' gain is about 60, rising to about 90 when the bootstrapping is fully effective.

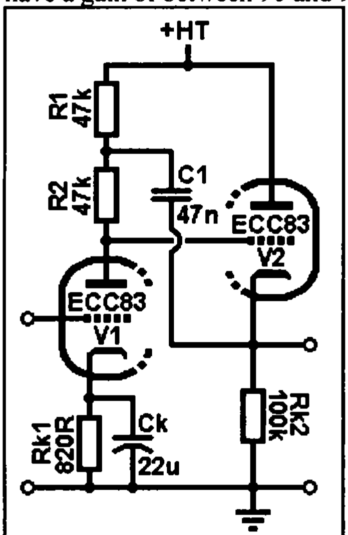

Fig. 5.22: Bootstrapping V1 increases the total gain to around 90 (39dB)!

The output signal swing of V1 is increased due to the increase in load impedance. However, in this particular circuit the positive-going output swing is restricted by grid-current limiting in the cathode follower, so that the total output signal swing of this circuit is hardly changed by bootstrapping, at about 130Vp-p. If the cathode follower were a lower r_a triode, this would be greater.
There are, it seems, no detrimental tonal effects caused by this modification and it is well worth considering for high-gain designs and modifications. In fact, because the bootstrapping is effectively a type of positive feedback (albeit a

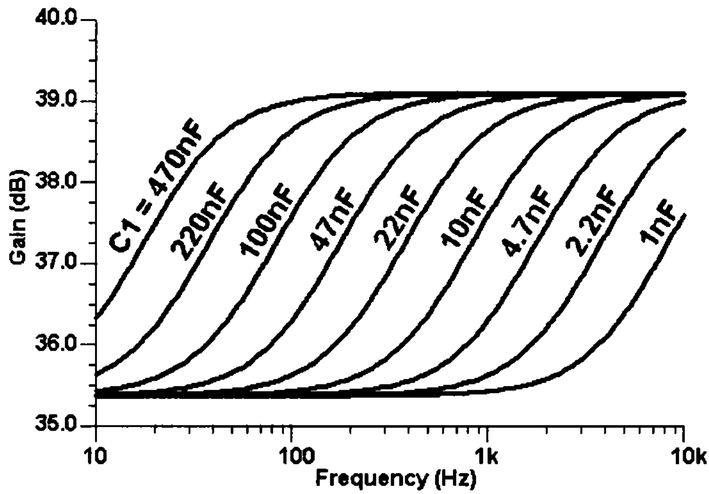

Fig. 5.23: Frequency response of the circuit in fig. 5.22 for different values of C1.

stable one since the cathode follower always has less than unity gain) the compression effect due to grid-current is actually increased, helping to develop even more sustain. In an existing amp, C1 could be connected to a switch for a boost or 'vintage / modern' function.

A further advantage of this circuit is that it is fairly immune to aging valves, since a cathode follower shows almost no reduction in gain with age, and the μ of an aging triode also changes very little. Since the gain of V1 is forced to approach μ, the total gain of the circuit does not decay much even with very worn-out valves.

Summary of formulae:

XXVIII; Voltage gain of a cathode follower.

$$A = \frac{\mu.Rk}{r_a + Rk(\mu + 1)}$$

XXIX; Voltage gain to a close approximation, if μ is much greater than 1:

$$A \approx \frac{\mu}{\mu + 1}$$

XXX; Input capacitance:

$$Cin = Cga + Cgk(1 - A)$$

But since Cgk (1 – A) is extremely small, for audio applications this may be simplified to: Cin ≈ Cga

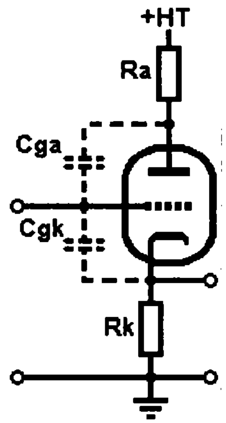

Fig. 5.24

XXXI; Internal cathode impedance:

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

(Ra is normally zero)

XXXII; Output impedance:

$$Zout = \frac{r_a}{\mu + 1} \parallel Rk$$

So:

$$Zout = r_k \parallel Rk$$

XXXIII; Output impedance to a close approximation if Rk is much greater than r_k and μ is much greater than 1:

$$Zout \approx r_k \approx 1/g_m$$

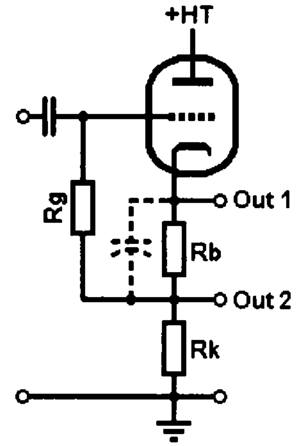

Fig. 5.25

XXXIV; Absolute maximum output signal swing (class A1 conduction):

$$VO_{(max\ p\text{-}p)} = \frac{HT}{1 + r_a / Rk}$$

XXXV; Input impedance of the arrangement in fig. 5.25 when Rb is unbypassed:

$$Zin = \frac{Rg}{1 - \frac{A.Rk}{Rk + Rb}}$$

XXXVI; Input impedance of the arrangement in fig. 5.25 if Rb is bypassed:

$$Zin = \frac{Rg}{1 - A}$$

•

Where in all cases all notations are as in fig. 5.24 and 5.25;
g_m = transconductance of valve at operating point.
μ = amplification factor of valve
r_a = internal anode resistance of valve at operating point.
HT = supply voltage in volts.
All resistances in ohms. All capacitances in farads.

Chapter 6: The Cascode

Basic operation of the cascode. The effect of the screen-grid voltage. Deriving anode-characteristics graphs. Fundamental parameters of the cascode. Designing a cascode. Current-boosted cascodes. Summary of formulae.

The cascode is one of a family of amplifiers known affectionately as **totem-pole** circuits, because they stack multiple devices on top of one another. The cascode uses two triodes, and although its creation can be traces back to the 1930s, it did not become popular until the widespread use of high-frequency circuits in television, radar and so forth, because of its very wide bandwidth. In audio it is often regarded as a substitute for a pentode, offering most the advantages of high gain, wide bandwidth and similar distortion characteristics, but without the inherent drawbacks of a real pentode such as partition noise, microphonics and higher cost. There were several late-generation, dual triodes which were specifically designed for cascode and totem-pole operation, such as the ECC84 / 6CW7, 6BK7, 6BQ7, 6BZ7 and ECC88 / 6DJ8, the last of which is enormously popular for modern hifi due to its excellent linearity. In guitar amplifiers, however, there is a strong compulsion to use the ECC83 / 12AX7 due to its universal availability and low cost. However, as will be seen, the ECC83 is probably the worst of all triodes to use in a cascode! The ECC82 / 12AU7 works well[1], though any triodes could be used if care is taken over design, and the author recommends experimentation.

Basic operation of the cascode:

The name 'cascode' may be assumed to be an abbreviation of '**cascaded-triodes** having similar characteristics to a pent**ode**'[2].
A simplified cascode circuit is shown in fig. 6.1 and it should be obvious that it resembles a pentode, since the grid of the upper triode looks like the screen grid in a pentode, and it operates in a similar (but not identical) manner.
It can be seen that the lower triode, V1, operates into the load presented by the cathode impedance of the upper triode, V2, which is very low (a few hundred ohms typically). Therefore V1 operates as a **transconductance amplifier** or **voltage to current converter**, and serves to control the current flowing through V2. Consequently V1 offers very little voltage gain, so the Miller effect is considerably reduced. This is why the circuit is well suited to high-frequency operation. Of course, for guitar this extended frequency response is well beyond what we actually require.

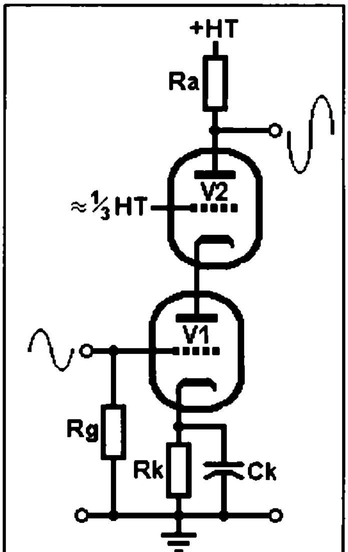

Fig. 6.1: Simplified cascode circuit.

[1] Price, L. R. (1954). The Cascode as a Low Noise Audio Amplifier. *Trans. I.R.E.*, **2**, (March), p64.
[2] 'Cathode Ray' (1955). The Cascode and its advantages for Band-III Reception. *Wireless World*, 61, 8 (Aug), p397.

The grid of V2 is normally held at a fixed voltage, and for maximum gain it will be bypassed with a capacitor, in the same manner as the screen grid in a pentode. In other words, we are injecting the input signal at the cathode while the grid is effectively grounded, as far as AC is concerned. Thus, V2 is arranged as a **common grid** or **grounded grid** amplifier, and the grid shields the cathode from voltage changes at the anode, so the input capacitance of the cathode is also very low, which maintains the wide bandwidth. Unlike a pentode, however, the grid of V2 does not normally draw grid current so there is no screen-compression effect, which is a benefit for high fidelity amplification but something of a disappointment as far as guitar is concerned.

The whole circuit offers high gain because it produces a nearly constant transconductance, like a pentode, allowing very high gains to be realised with most triodes. However, the gain of a cascode is not as high as if the same two triodes were cascaded. Fortunately, the overdrive characteristics of the cascode are useful enough that this sacrifice is a worthy one.

Since V1 operates like an ordinary common-cathode gain stage, albeit with very low voltage gain, it is inverting. Since V2 receives its input at its cathode it is non-inverting. Overall then, the circuit is an inverting gain stage, the same as a pentode. The triodes are connected in series as far as the power supply is concerned, but because the output signal of the lower triode is passed onto the second triode they may be said to be cascaded, as far as AC is concerned (rather than, say, push-pull).

The effect of the 'screen-grid' voltage:

The main controlling factor in a cascode is the voltage on the grid of the upper triode, which we will call the screen grid, g2, for brevity. Since the upper triode must be biased normally for proper operation its grid will be at a slightly lower voltage than its cathode, which is also the anode voltage of the lower triode. But since this difference is only a few volts it is usual to assume that the screen-grid voltage is equal to (and defines) the anode voltage of the lower triode, when initially choosing operating points. Generally this voltage will be in the region of 80V to 100V, or around ⅓HT. Below about 80V* many triodes will give poor or unpredictable performance in a cascode, unless they are specifically designed for the purpose (though the ECC82 / 12AU7 is a useful exception), so this is usually taken to be the lower limit for ordinary triodes. Higher voltages will increase the g_m, and therefore the possible gain but, as will be seen, it is at the expense of output signal swing.

The best way to appreciate the effect of the screen-grid voltage is to examine the static anode characteristics of the cascode for different values of Vg2, as we did for the pentode in chapter 3. The oscillograms in fig. 6.2 show the measured characteristics of an ECC82 / 12AU7, which operates well at lower anode voltages. For ease of comparison, the anode current and anode voltage scales, and the grid-curve divisions, have all been kept the same in each graph. Note that the anode

* With most triodes, Vg2 = 80V seems to be something of a 'magic number' for the cascode, above which the transconductance rapidly increases.

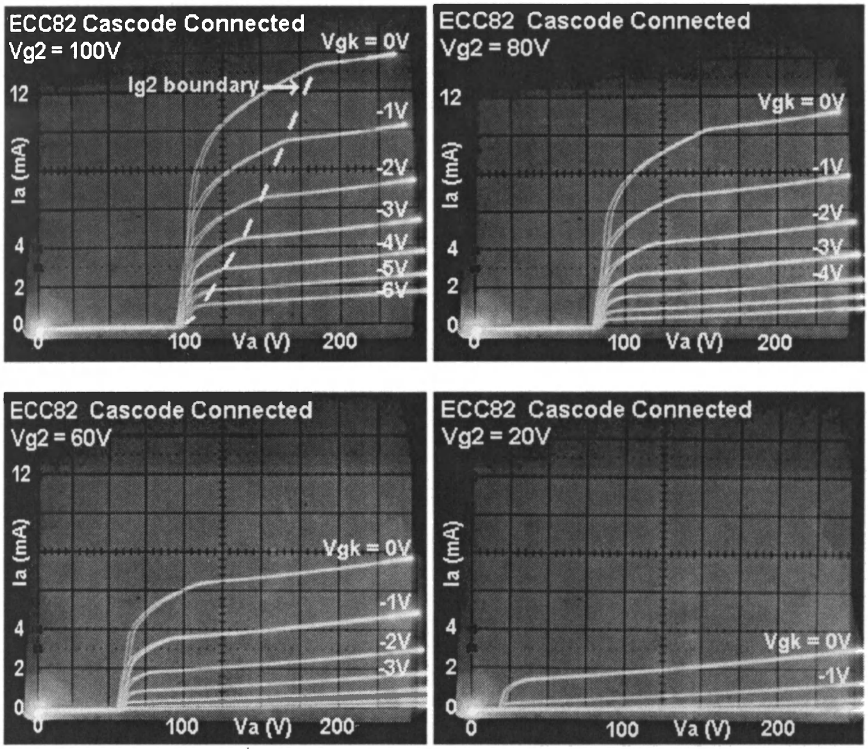

Fig. 6.2: Lowering the screen-grid voltage 'squashes' the grid-curves down, reducing g_m, and also shifts the curves to the left.

voltage scale indicates the voltage between the anode of the *upper* valve and the cathode of the *lower* valve. Thus the stage is treated as one, single valve.

The most obvious feature of the characteristics is that they resemble those of a pentode, except that they are shifted to the right by an amount equal to the screen-grid voltage. The reason for this should be obvious: if the anode voltage falls below the screen-grid voltage then there is effectively no voltage across the upper triode and it cannot conduct any anode current. So the absolute maximum, peak-to-peak, output signal swing of a cascode is always limited to HT-Vg2.

Another feature to note is the distinctive inflection in the curves near the 'knee', which marks the point at which the upper triode begins to draw grid current (indicated by the dashed line in the upper graph.)

Lowering the screen-grid voltage 'squashes' the curves down, since lowering the anode voltage of V1 reduces its g_m, and therefore its ability to force current in V2 to change. This reduces the possible gain, but the curves are also shifted further to the left, so the maximum output swing increases. Unlike the pentode, the anode resistance, r_a, of the cascode does not increase appreciably as the screen-grid voltage is lowered, nor does its linearity improve much.

For comparison, the cascode characteristics of some other common triodes are shown in fig. 6.3, from which it can be seen that the ECC83 / 12AX7 gives

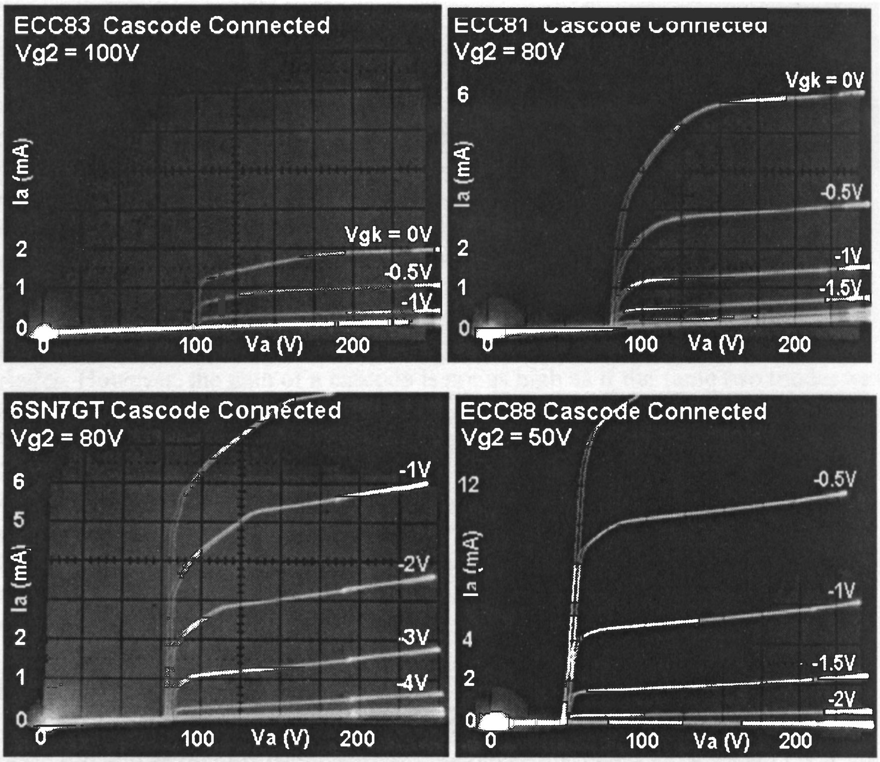

Fig. 6.3: Cascode characteristics of some common triodes.

extremely poor performance, even with a high screen-grid voltage of 100V. This is due to its low transconductance. The ECC81 / 12AT7 provides some useful characteristics, and might be a good choice for hifi. However, its high grid current tends to produce a brittle tone when overdriven, making it unpopular for guitar amp use* (its high grid-current is also the reason for the very soft, 'well rounded' knee to the curves). The 6SN7GT gives similar performance to the ECC82, which is to be expected since they are very similar valves, and both are very popular for guitar and hifi. The ECC88 / 6DJ8 give technically the best performance of all, giving a total g_m of over 5mA/V, depending on bias, and could achieve very high gain indeed. However, it is a relatively expensive valve and has a different heater arrangement from the ECC83, so it is unlikely to appear in guitar amplifiers. Consequently, most of this chapter will explore the use of the ECC82 / 12AU7 as a cascode.

Deriving anode-characteristics graphs:

For those who do not have the luxury of a curve tracer, the static anode characteristics of a cascode can be derived from the ordinary triode characteristics

---

* Although the ECC81 / 12AT7 does give a good tone when used as a long-tailed pair (chapter 8).

given in the data sheet[3]. What's more, we can use this method to estimate the performance of *dissimilar* triodes arranged as a cascode. The accuracy of this technique is not perfect, but should be sufficient for most purposes.
Let us take the example of a 12AY7 being used as the lower triode, V1, while an ECC82 / 12AU7 is used for the upper triode, V2. Exactly the same process applies to any combination of triodes.

- Firstly we take the ordinary, static anode characteristics of the *lower* triode, the 12AY7. To these we add the Vgk = 0V grid curve of the *upper* triode, the ECC82, but we shift it to the right by an amount equal to the chosen screen voltage. 80V has been selected in this case (fig. 6.4*a*).
- We then note the anode currents when the anode voltage across V1 is equal to Vg2, at each value of Vgk, and project these across to the new grid curve. These points are marked by the large dots in fig. 6.4*b*, corresponding to Vgk = 0V, -1V and -2V.
- Now we plot the next grid curve for the upper valve, except that we shift it to the right by an amount equal to $Vg2 - Vgk_{(upper)}$. In this case the -4V curve is drawn, so it must be shifted by:
  80 – (–4) = 84V.
  We now note the anode current when the anode voltage across V1 is equal to $Vg2 - Vgk_{(upper)}$, or 84V, and project these across to this new grid curve, again marking the spots as in fig. 6.4*c*.

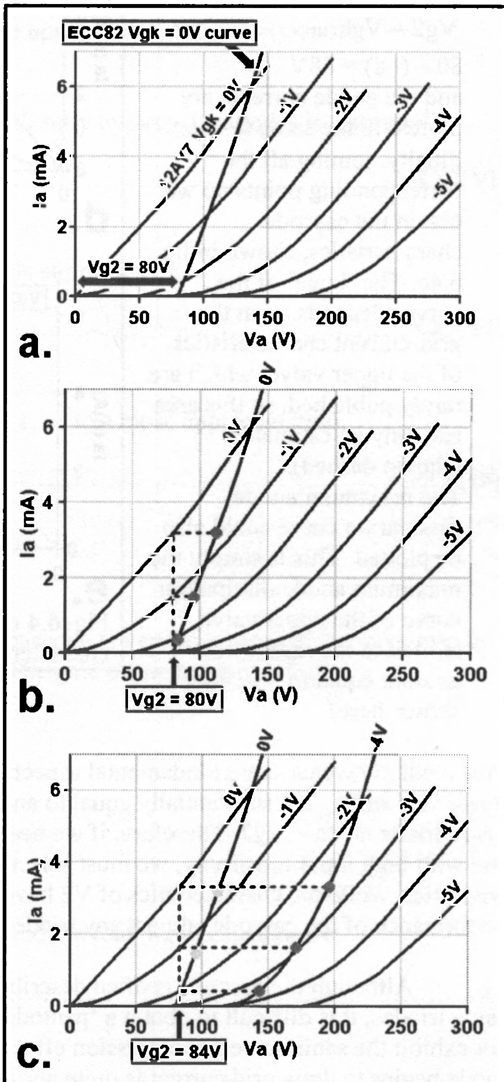

Fig. 6.4: Deriving cascode characteristics.

[3] Grant, W. (1957). Cascode Characteristics. *Wireless World*, 1, (January), pp33-36.

- More curves are plotted in exactly the same way. In fig. 6.4*d* the ECC82 Vgk = -8V curve is added, again shifted to the right by
  Vg2 – Vgk(upper) :
  80 – (–8) = 88V
  and the anode currents are plotted in the same fashion.
- Finally, joining all the corresponding points up we obtain the cascode characteristics, shown in fig. 6.4*e*. The 'knee' of the curves depends upon the grid-current characteristics of the upper valve, which are rarely published, so this area can only be estimated (shown dashed).
  The maximum anode dissipation curve could also be plotted. This is simply the maximum anode dissipation curve of the upper valve, shifted to the right by an amount equal to Vg2 (not shown here).

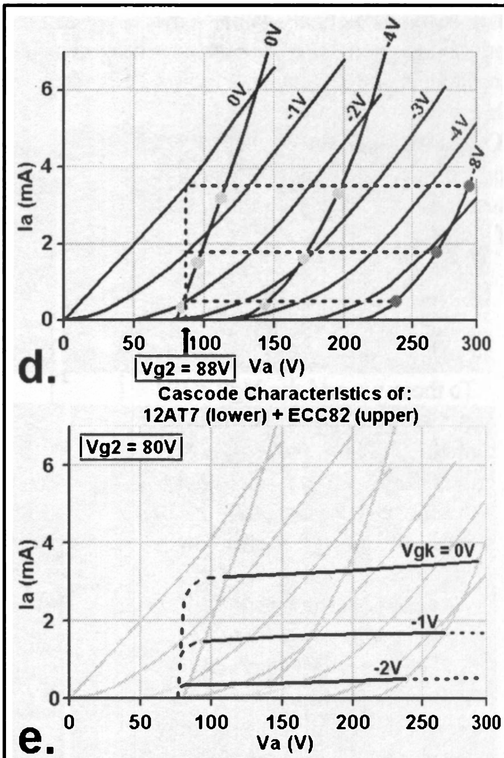

Fig. 6.4 cont': Deriving cascode characteristics.

This method reveals some fundamental aspects of the cascode: We find that its input sensitivity and g_m are substantially equal to and almost entirely defined by that of the *lower* triode at Va = Vg2. Therefore, if we need a high transconductance cascode or one with high input sensitivity, we must select a triode for V1 having those properties, while the characteristics of V2 have only a slight influence on the final performance of the cascode, almost any triode could be used.

Although the cascode is often described as a way of 'simulating a pentode' using triodes, it is difficult to obtain a 'pentode tone' from the cascode, since it does not exhibit the same screen-compression effects. Also, the region where the upper triode begins to draw grid current is quite abrupt, and it can be difficult to enter this region of operation without hard-clipping the signal first. On the other hand, the cascode makes a good input stage due to its high gain and wide bandwidth, although to take advantage of it distortion characteristics it must be overdriven of course, which implies using it further on in the preamp. In essence, the cascode may be regarded as lying somewhere between a true pentode and a triode, in terms of tone and usefulness for guitar purposes.

Fundamental parameters of the cascode:

A valve cascode does not lend itself well to mathematical treatment, since so many of its parameters are dependant on specific values of r_a and g_m, which are themselves highly influenced by the operating point, particularly at low anode voltages and currents. In the following formulae, subscript 1 indicates the lower triode V1, and subscript 2 indicates the upper triode, V2.

Gain:

The gain of a conventional cascode with fully bypassed cathodes is given by[4]:

$$A = \frac{\mu_1(\mu_2 + 1)Ra}{Ra + r_{a2} + r_{a1}(\mu_2 + 1)} \quad \text{.......XXXVII}$$

But if both triodes are identical then this simplifies to:

$$A = \frac{\mu(\mu + 1)Ra}{Ra + r_a(\mu + 2)} \quad \text{.......XXXVIII}$$

For most triodes μ is much larger than 1, so to a close approximation:

$$A \approx \frac{g_m.Ra}{1 + \frac{Ra + r_a}{\mu.r_a}} \quad \text{.......XXXIX}$$

Input capacitance:

The input capacitance of the cascode is very low because the lower triode operates with very little voltage gain. The gain of the lower triode is:

$$A_{lower} = \frac{\mu_1.r_{k2}}{r_{a1} + r_{k2}}$$

Where:

$$r_{k2} = \frac{Ra + r_{a2}}{\mu_2 + 1}$$

Substituting this into formula VII we find the total input capacitance:

$$Cin = Cgk + Cga\left(\frac{\mu_1.r_{k2}}{r_{a1} + r_{k2}}\right) \quad \text{.......XL}$$

But since the voltage gain of the lower triode is not much greater than unity, this can be approximated as: $Cin \approx Cgk + Cga$, with only a modest error. As a result, quite

[4] Attree, V. H. (1955). A Cascode Amplifier Degenerative Stabilizer. *Electronic Engineering*, (April), pp174-177.

large grid stoppers may be used without affecting the treble response of the stage, as with the pentode.

Output impedance:

Assuming all cathode resistors are bypassed, the output impedance of the cascode is given by:

$Zout = Ra \parallel r_{a2} + r_{a1}(\mu_2 + 1)$ L

For most triodes $r_a + r_a(\mu + 1)$ is large with respect to Ra, so the output impedance may be approximated as: $Zout \approx Ra$, which is the same as for a pentode.

If the cathodes are unbypassed then the output impedance is increased to:

$Zout = Ra \parallel [Rk1(\mu_1 + 1) + r_{a1} + Rk2].(\mu_2 + 1) + r_{a2}$

But since the right hand expression is normally extremely large with respect to Ra, the output impedance may still be approximated as: $Zout \approx Ra$. Hence, cathode bypassing has minimal effect on the output impedance of a cascode.

Designing a cascode:

When designing a cascode for guitar use, many of the design considerations are the same as for a small-signal pentode [chapter 3], except that it tends to be even more iterative. Of all stages, cascodes are probably the least likely to perform exactly as expected on paper, due to the various non-linearities and inconsistencies of real valves. Nevertheless, the methods described here should provide a good starting point for modification.

Ideally we would like the load line to pass through or slightly below the 'knee' of the grid curves, for the most 'pentode like' sound. Allowing it to pass much above the knee will tend to result in more triode-like distortion characteristics. Of course, if we actually wanted a high-gain triode stage then this would be a suitable option.

Although the cascode is capable of very high gain, in a guitar amp too much gain in a single stage can be a hindrance, robbing the amp of touch sensitivity as well as increasing the risk of microphonics and parasitic oscillation. With the small-signal pentode it was recommended that the gain did not exceed more than about 100. With the cascode this can reasonably be increased to 200 say, if necessary.

The easiest way to begin is to choose the anode load resistor, Ra, and make the cascode characteristic fit the load line. Having discovered that most of the cascode's characteristics are defined by the lower triode the design process is made simpler; we begin by drawing the load line on the static anode characteristics graph of the lower triode. In this case the HT is 300V and an ECC82 / 12AU7 is used. A 47kΩ load is chosen as this is large enough to ensure a large output signal swing, but not so large that operation is constrained to low anode currents, where performance may be poor. Fig. 6.5 shows the load line. It must be appreciated though, that this is *not* the actual load line that the lower triode operates on (which is near vertical), it is simply used for selecting electrode voltages.

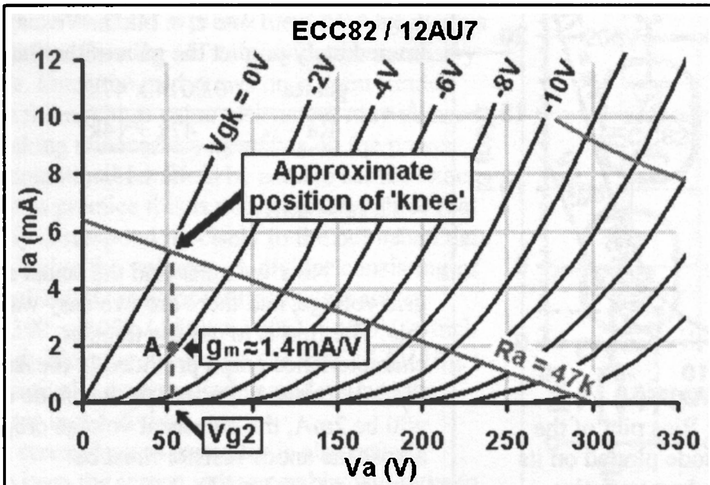

Fig. 6.5: Drawing the load line on the lower triode's characteristic gives an indication of the screen voltage which should be used.

If we would like the load line to pass through the knee of the final characteristics, the approximate screen voltage we require corresponds to anode voltage where the load line meets the Vgk = 0V grid curve, as indicated by the dashed line. In this case it is around 55V, which is convenient since it is well below the maximum heater-to-cathode voltage rating of the upper triode, so no heater elevation will be required.
The lower triode operates into the internal cathode impedance of the upper triode, which is low enough that we may assume it is zero (a vertical load line). We may therefore take the input sensitivity as being that which is indicated by the dashed line in fig. 6.5. This suggests about 4Vp-p, although the grid curves are very bunched up near cut-off, so the real value will be a little larger. For roughly centre biasing a value -2V is required, indicated by the bias point, 'A'. The anode current at this point is approximately 2mA, so apply Ohm's law we find the necessary cathode resistor:

$R = V/I = 2 / 0.002$

$= 1k\Omega$

(We could have chosen a high screen grid voltage so that the load line passes below the knee. However, the chosen bias point on the dashed line can only exist *below* the load line.)

In practice we will usually find the quiescent current (and therefore the bias voltage) to be a little higher than predicted, because the knee does not actually drop down absolutely vertically towards the abscissa, though the difference is not usually very great.

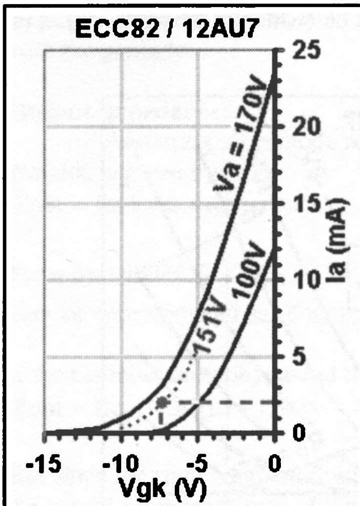

Fig. 6.6: Bias pint of the upper triode plotted on its transfer characteristics.

The g_m indicated at the bias point gives us the approximate g_m of the finished cascode. In fig. 6.5, at an anode voltage of 55V the g_m is roughly 1.4mA/V, μ = 19 so r_a = 14kΩ. We may immediately predict the gain of the final circuit:

$$A \approx \frac{gm.Ra}{1+\frac{Ra + r_a}{\mu.r_a}} = \frac{0.0014 \times 47k}{1+\frac{47k + 14k}{19 \times 14k}}$$

= 53.5 (34.6dB).

Now we must find the actual screen-grid voltage, and there are two easy ways to do this. The first is to use the transfer characteristics graph provided in the data sheet. Since we already know that the anode current will be 2mA, the quiescent voltage dropped across the anode resistor must be:

V = IR = 0.002 x 47k
= 94V.

We have also determined the anode voltage of the lower triode, and therefore the cathode voltage of the upper triode, to be 55V. The quiescent *anode-to-cathode* voltage across the upper triode must therefore be: 300 – 94 – 55 = 151V . The transfer characteristics graph is shown in fig. 6.6 and the Va = 151V curve has been extrapolated (shown dotted). The bias point for 2mA anode current is plotted and indicates a bias voltage of -7.5V. Since we know the cathode voltage to be 55V, the actual screen grid voltage must be: 55 – 7.5 = 47.5V .

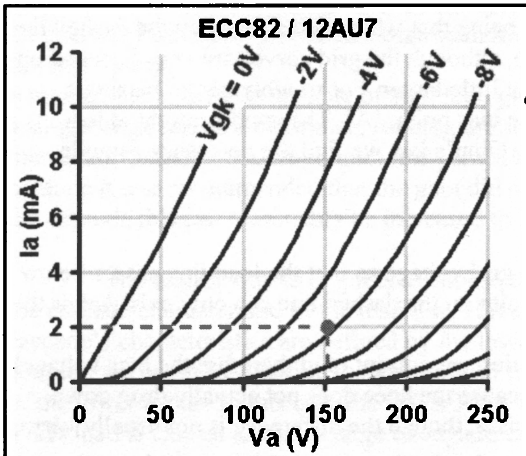

Fig. 6.7: Bias point of the upper triode plotted on its static anode characteristics.

The second method is to again calculate the anode-to-cathode voltage, but to plot the bias point on the static anode characteristics graph instead. This is shown in fig. 6.7 and the bias point also indicates a bias voltage of about -7.5V and therefore a screen voltage of 47.5V. Either method can be used to derive the screen voltage, although the static anode characteristics graph is often easier to interpret than the transfer characteristics graph. The partially completed circuit is shown in fig. 6.8, with expected voltages.

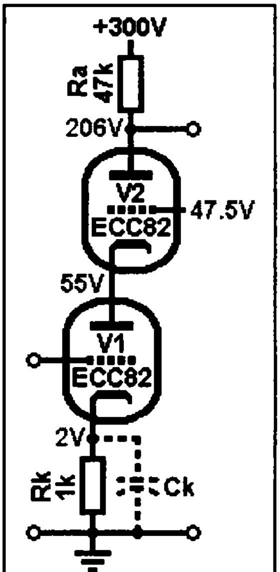

Fig. 6.8: Partially completed cascode.

Applying fixed bias to the upper triode:

Now that the necessary voltages are known, we must decide *how* the upper triode will be biased. The 'traditional' method is to use fixed bias, by using a potential divider from the HT to provide the necessary grid voltage. Since the grid draws no current (under normal conditions) high value resistances may be used, to avoid making unnecessary demands on the power supply. A potentiometer could be used to set the exact voltage, but in practice this is not necessary, since real valves rarely correspond precisely to the published data we used to derive the voltages. A divider consisting of a 560kΩ and 100kΩ resistor will provide a screen voltage of 300 x (100k / 560k + 100k) = 45.5V, which is sufficiently close to our design value of 47.5V [fig. 6.9]. Alternatively, a zener diode could replace the lower resistor in the divider.

A screen bypass capacitor is also usually included to keep the screen voltage stable, and to keep power supply noise off the grid. This capacitor does not affect the gain of the circuit (because the grid does not draw grid current in the same way a pentode does) but it does affect the overdrive characteristics. During up-going input cycles the cathode voltage of the upper triode is pulled down and the upper grid will eventually begin to draw grid current. If the screen bypass capacitor is not included, these pulses of current will cause the screen voltage to fluctuate in sympathy, producing a roughly half-wave rectified signal voltage to appear on the screen grid. This signal being mixed with the wanted signal can produce some very bizarre output waveforms, with heavy high-order harmonics and intermodulation distortion. This generally produces a harsh and unpleasant sound, although it is highly dependant on the screen voltage. (Interestingly, the effect is worsened if the load line passes above the knee.) The screen bypass capacitor will supply these pulses of grid current while keeping the grid voltage substantially constant, and the tone is usually more compressive and pleasant with this capacitor included. A suitable value of capacitor may be chosen according to:

$$C = \frac{1}{2\pi f(R1 \parallel R2)}$$

Were:

f = the lowest frequency at which the grid voltage will be held substantially constant.

$$R1 \parallel R2 = \frac{R1.R2}{R1 + R2}$$

In this circuit R1||R2 = 85kΩ. To keep the grid voltage steady down to 20Hz:

$$C = \frac{1}{2\pi \times 20 \times 85k}$$

= 93.6nF

So a 100nF capacitor would be suitable. Its voltage rating should be sufficient to handle the normal operating voltage, although it need not be rated for the full HT voltage,

Choosing the cathode bypass capacitor:

The cathode bypass capacitor, Ck, is normally included to maximise gain (it has very little effect on the output impedance of the stage). Omitting it will reduce the gain of the stage since it reduces the transconductance of the lower triode to an effective value of:

$$g_{m(eff)} = \frac{gm}{1 + gmRk} = \frac{0.0014}{1 + (0.0014 \times 2000)}$$

= 0.37mA/V

The gain of this cascode would therefore be reduced to:

$$A \approx \frac{gm(eff).Ra}{1 + \dfrac{Ra + r_a}{\mu . r_a}} = \frac{0.00037 \times 47k}{1 + \dfrac{47k + 15k}{19 \times 15k}}$$

= 14.5 (23dB).

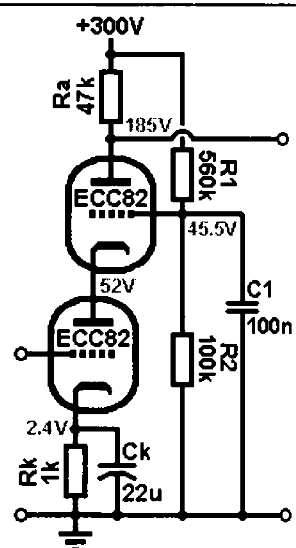

Fig. 6.9: A cascode having a gain of around 50 (34dB) in which the upper triode is fixed biased

Ck may be chosen according to formula VI, except that the load on the lower triode is formed by the internal cathode resistance of the upper triode. This modifies the formula to:

$$Ck = \frac{1}{2\pi fRk} . \sqrt{1 + \frac{Rk(\mu_l + 1)}{2\left(\dfrac{Ra + r_{a2}}{\mu_l + 1} + r_{al}\right) + \frac{1}{2}Rk(\mu_l + 1)}}$$

But since Ra + $r_{a2}/(\mu + 1)$ is much smaller than r_{al}, this may be simplified to:

$$Ck \approx \frac{1}{2\pi fRk} . \sqrt{1 + \frac{Rk(\mu_l + 1)}{2r_{al} + \frac{1}{2}Rk(\mu_l + 1)}}$$

Where:

f = the half-boost frequency.

μ_l = amplification factor of the lower triode.

r_{al} = anode resistance of the lower triode.

For full bypassing we might take a half-boost frequency of 10Hz, which gives a value of 19µF, so a 22uF capacitor is suitable.
The completed circuit is shown in fig. 6.9. This circuit was tested and found to have a gain of 50, close to the calculated value of 55, and a maximum output signal swing of 200Vp-p. The output signal swing is less than might have been predicted from HT-Vg2, and this is due to the inability to supply enough grid current to the upper triode to enter the region to the left of the Ig2 boundary [see fig. 6.2 upper].

The gain could be raised by increasing the load resistor, Ra. However, if it is made too large then the load line will be pushed further down into the region of low g_m and gain will fall again. Generally the optimum value can only be found by experimentation. In the circuit of fig. 6.9 a maximum gain of 80 is achieved into a 90kΩ load. However, the load line for this value passes well below the knee, so that clipping tends to become harder and more asymmetric, making the overdriven tone more aggressive (sometimes too aggressive). For most playing styles, a better tone is obtained when the load line passes through or very close to the knee, even though the gain is not as high as it could be.
Fig. 6.10 shows a similar circuit optimised for the ECC83 / 12AX7. Its measured gain is 140, which might sound better than the ECC82 example designed previously, but it should be noted that the anode load resistor and screen voltage are greater, which limits the output signal swing to around 110Vp-p.

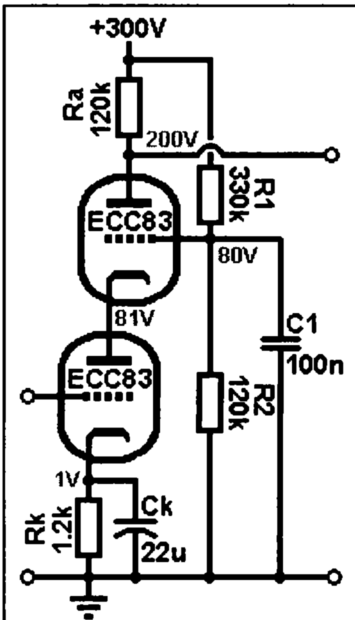

Fig. 6.10: A cascode having a gain of around 140 (43dB) but a 110Vp-p maximum output swing.

An interesting difference is observed between the clipping characteristics of the ECC83 cascode, when compared with most other triodes. The left-hand oscillogram in fig. 6.11 shows the 200Vp-p output signal of the ECC82 cascode in fig. 6.11 when heavily overdriven with an input signal of about 12Vp-p. The source impedance was 100kΩ, and the grid signal is visibly clamped on the upper portion due to grid-current limiting. The output signal shows clipping on the up-going cycle due to cut-off, and on the down-going cycle due to a combination of grid-current limiting in *both* triodes, softening the clip and giving a good pentode-like sound.
The right-hand oscillogram shows the output of the ECC83 cascode in fig. 6.10 when driven by the same input signal. The grid-current limiting and cut-off characteristics in the lower triode are broadly the same as those of the ECC82 [see also chapter 1, fig. 1.6]. But the 110Vp-p output signal shows that the clipping due to the upper triode is quite different, producing a very soft and rounded waveform on the down-going cycle. This produces a tone that is less like a pentode and more like

that of an ordinary cold-biased, triode gain stage. The same phenomenon is witnessed in all forms of cascode using the ECC83, and the causes remain unclear.

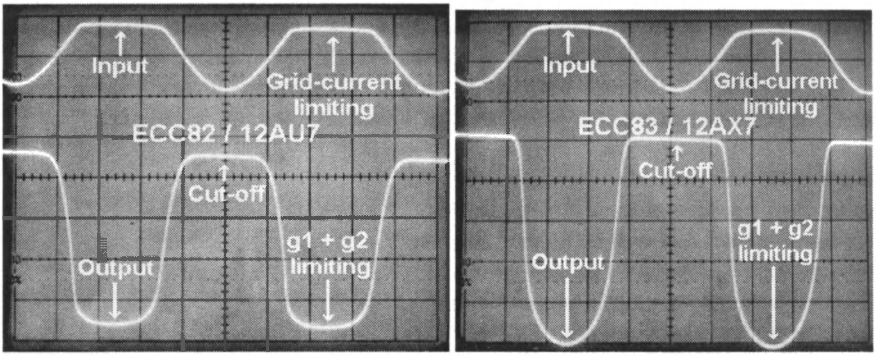

Fig. 6.11: Striking differences are observed between the clipping characteristics of the ECC83 when compared with other types of triode.

Applying cathode bias to the upper triode:

Although most textbook cascodes use fixed bias, cathode bias can also be used, as shown in fig. 6.12. An additional bias resistor, Rk2, is inserted between the two triodes. The bias voltage is applied to the screen grid via a grid-leak resistor, Rg2, and a decoupling capacitor ensures the grid voltage is held constant. In this arrangement, if the capacitor were omitted the signal voltage appearing at the anode of V1 would be passed directly to the grid of V2. With almost no change in voltage between its grid and cathode the upper triode would resist changes in its anode current, and gain would be substantially reduced.

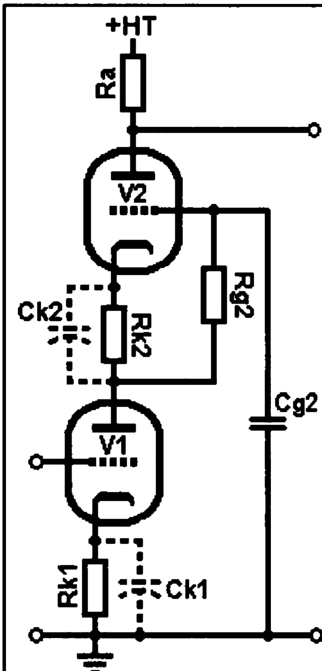

Fig. 6.12: Cascode in which both triodes are cathode biased.

In the earlier example using the ECC82, the bias for the upper triode was determined to be -7.5V. The quiescent current was predicted to be 2mA. Applying Ohm's law we find the value of Rk2:
$R = V/I = 7.5 / 0.002$
$=3750\Omega$.
So a standard value of $3.3k\Omega$ would normally be suitable.

Of course, the presence of the resistor will itself reduce the quiescent current in the cascode slightly. However, since we have already observed that the quiescent current will usually be slightly higher than predicted anyway, the presence of Rk2 will actually work in our favour.

The value of the grid leak resistor, Rg2, is not critical. A value of 1MΩ is common, although a much lower value of say 100kΩ could be used for less noise, without any change in performance.

The screen bypass capacitor, Cg2, may be chosen according to:

$$Cg2 = \frac{1}{2\pi fRg2}$$

Where:
f = the lowest frequency at which the grid voltage is to be held substantially constant.

For a frequency of 10Hz:

$$C = \frac{1}{2\pi \times 10 \times 1000k}$$

15.9nF

So a standard value of 22nF could be used. Normally we would not even bother to calculate its value, and simply use an arbitrarily large value, 100nF say.

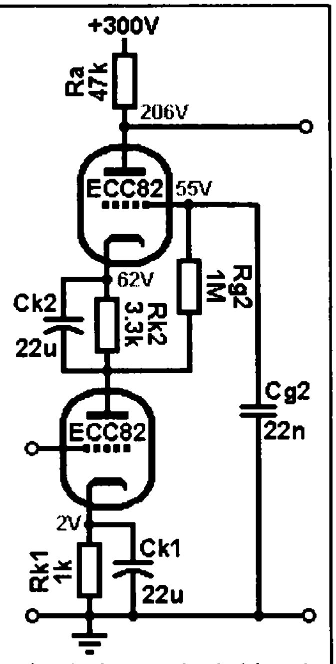

Fig. 6.13: A cathode biased cascode having a gain of around 45 (33dB).

Because Rk2 increases the load into which V1 operates, it reduces its effective g_m and therefore reduces the gain of the cascode. However, because its value is normally not very large the reduction in gain is not great. If maximum gain is to be realised, though, then a bypass capacitor, Ck2, can be added. To a reasonable approximation, the half-boost frequency created between Ck2 being unbypassed and fully bypassed is:

$$f \approx \frac{1}{2\pi fCk2Rk2}$$

A more complex treatise is not necessary, since the difference in gain between the unbypassed and fully bypassed case is very small. In this example, for a frequency of 10Hz a 4.8μF capacitor is required. A 4.7μF capacitor could be used, although it may be convenient to use the same value as used to bypass Ck1. Much smaller values can be used for a very subtle amount of treble boost, of course. On the other hand, it may be

desirable to leave Rk2 unbypassed since this avoids the use of another electrolytic capacitor, with only a slight sacrifice in gain.
Alternatively, one or both bias resistors could be replaced with zener diodes or LEDs to provide the necessary bias, which would negate the need for bypass capacitors.

The completed cascode circuit is shown in fig. 6.13 with measured voltages, and it can be seen that the quiescent current is indeed lower than for the fixed-bias version. It was found to have a gain of 45, which is somewhat less than the fixed-bias version and this can be ascribed to the lower quiescent current causing a reduction in g_m. A maximum gain of 64 was achievable into a load of 180kΩ. This appears to be less than the figure measured in the fixed-bias version, but this is because the screen-grid voltage is now proportional to the anode current, which is itself reduced by the increased value of anode resistor.
The tone of this form of cascode tends to be slightly 'warmer' and 'smoother' than the fixed-bias version. It also cannot suffer from grid-cathode arcing in the upper triode.

Applying grid-leak bias to the upper triode:

Probably the simplest cascode topology is one in which the upper triode is grid-leak biased, as shown in fig. 6.14. This is actually more easily done than for a regular gain stage, because of the presence of the screen bypass capacitor, Cg2. Unlike regular grid-leak biasing, which relies on a leakage current flowing into the grid, this arrangement does not require a very high value grid-leak resistor.

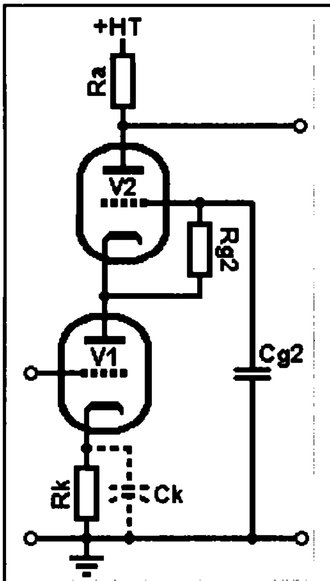

Fig. 6.14: Simple cascode in which the upper triode is grid-leak biased.

During quiescence there is zero bias on the upper triode. But when a signal is applied and the cathode voltage rises, a charging current flows into Cg2 via the grid-leak, Rg2, which develops a small negative voltage across Rg2, relative to the cathode. The greater the signal level, the greater the bias voltage developed. This has a small compressive effect since the average current, and therefore the g_m of the lower triode, is reduced when the cascode is driven hard. Since there is also a time delay while the bias voltage develops, transients are hardly affected; only if a high signal level is maintained is the output signal compressed, in a very similar manner to that caused by power supply 'sag' in push-pull amplifiers using a valve rectifier! This effect, and the simplicity of this cascode topology, make it probably the most useful for guitar purposes. It is also useful because it reduces the number of components which can be adjusted on test.

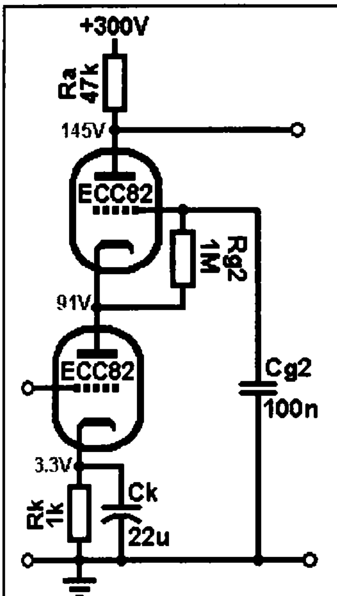

Fig. 6.15: A simple grid-leak biased cascode having a gain of roughly 50 (34dB).

The value of the grid leak resistor, Rg2, is not critical, and a value of 470kΩ or 1MΩ would be typical. In order for the gain and distortion characteristics of the cascode to remain the same across the full bandwidth, the time constant produced by Rg2 and Cg2 should be at least equal to the time period of the lowest frequency which is to be amplified (although interesting results are obtained if the capacitor is made small, since this introduces unusual distortion characteristics at low frequencies).
Since the time period of a sine wave is: $\tau = 1 / f$, and the time constant is Rg2 x Cg2 , the value of Cg2 may be chosen according to:

$$Cg2 = \frac{1}{f.Rg2}$$

Where:
f = the lowest frequency to be amplified 'cleanly'.

For example, if Rg2 = 1MΩ, and we take an arbitrarily low frequency of 10Hz:

$$Cg2 = \frac{1}{10 \times 1M}$$

= 100nF

This value is shown in fig. 6.15, with measured voltages. Note that the quiescent current is much higher than in the previous circuits, because the upper triode is effectively zero biased while there is not signal input. When fully overdriven the average current falls and the anode voltage of the lower triode reduces to close to the design value of ~51V.
Fig. 6.16 shows a similar arrangement optimised for the ECC83 / 12AX7. Its measured gain is 100, but its maximum output signal swing is again only about 110Vp-p.

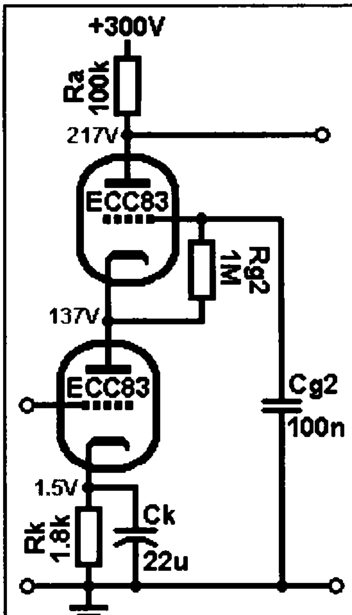

Fig. 6.16: A simple grid-leak biased cascode having a gain of roughly 100 (40dB).

Current-boosted cascodes:

It should now be clear that the gain of the cascode is highly dependant on the transconductance of the lower triode. Therefore, it may be desirable to increase its transconductance to achieve higher gain without resorting to a high screen-grid voltage (which reduces the output signal swing and places more stress on the heater-to-cathode insulation) and without using a completely different valve type. This might seem like an impossible task unless we recall that the g_m can be increased by operating at a higher quiescent current. However, if we simply increase the current through the whole cascode, while maintaining the same screen-grid and anode voltage, we shall be forced to bias the upper triode hotter, which places the bias point closer to the region of screen-grid current, resulting in very limited head room indeed.

The more subtle way to increase the gain of the cascode without these drawbacks is to supply the lower triode with the additional current via a separate resistor, R1, from the HT, as shown in fig. 6.17.

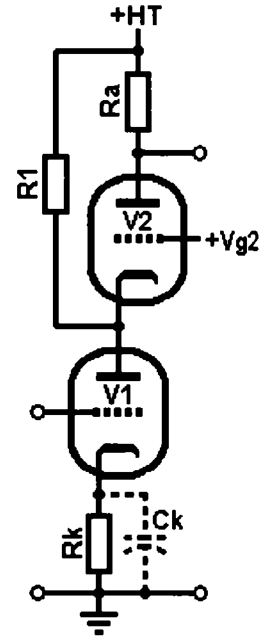

Fig. 6.17: Boosting the g_m of the lower triode by feeding it additional current via R1.

Design of such a stage is much the same as before. First a load line is drawn on the lower triode's characteristics, as before (fig. 6.18). After selecting the anode voltage we again decide upon the quiescent current for the lower triode. This time, however, we wish to maximise g_m by selecting a high value of current without reducing the bias to a uselessly low level. For this reason, a high anode voltage of 75V has been chosen, so that the load line will now pass below the knee. Also, because we can supply any amount of current to the triode from the HT the bias point can be anywhere on the dashed line, even above the load line.

A suitable bias point has been labelled 'A', and results in a bias of -2V and a quiescent current of about 3.5mA. The cathode bias resistor must be:

$R = V/I = 2 / 0.0035$

$= 571\Omega$.

So a value of 560Ω would be suitable. Choosing the cathode bypass capacitor follows exactly the same method as earlier.

At this point g_m appears to be around 2.2mA/V, $\mu = 19$ so $r_a = 8.6k\Omega$, sand we may predict a gain of:

$$A \approx \frac{gm.Ra}{1+\frac{Ra + r_a}{\mu.r_a}} = \frac{0.0022 \times 47k}{1+\frac{47k + 8.6k}{19 \times 8.6k}} = 77 \text{ (37.7dB)}$$

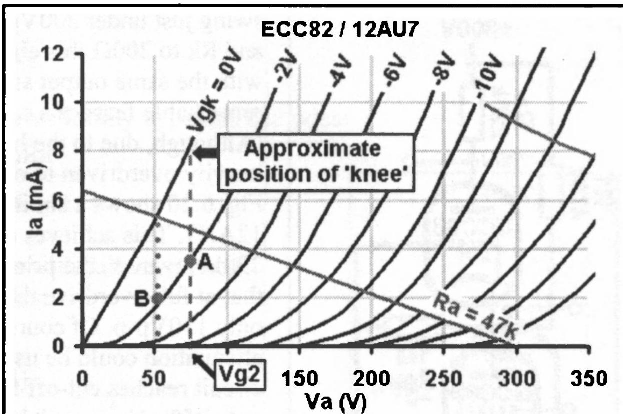

Fig. 6.18: By supplying the lower triode with additional current, the g_m and gain is increased. The quiescent anode currents of each triode are no longer equal, of course, so two points are plotted.

We may now select the quiescent current for the upper triode. This is limited by the load line, and the range of possible current is shown by the dotted line. For roughly centre biasing a current of 2mA is appropriate, as indicated by point 'B'. The biasing of the upper triode can be done in any of the ways described earlier, since this part of the circuit is unaffected by the additional current flowing in the lower triode.

Finally we must select the value of the current-feed resistor, R1. We have already selected the total current in the lower triode to be 3.5mA. Since the anode current of the upper triode has been selected as 2mA, the remainder must be supplied by R1. The additional current required is 3.2 – 2 = 1.5mA.
We also chose the anode voltage of the lower triode to be 75V. The voltage dropped across R1 must therefore be:
HT – Va1 = 300 - 75
= 225V.

Ohm's law yields the required resistance:
R = V / I = 225 / 0.002
= 112.5kΩ.
So a value of 120kΩ would probably be close enough. The power dissipated should be close to:
$P = I^2R = 0.002^2 \times 120k$
= 480mW
So a 1W resistor should be sufficient.

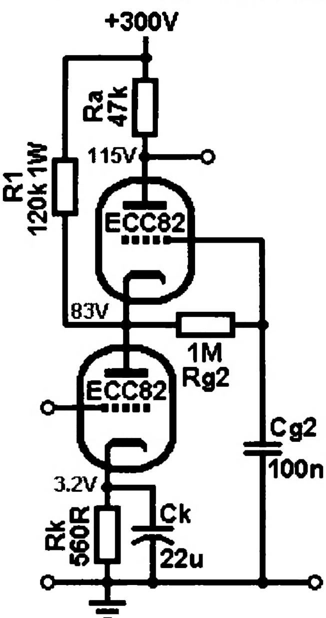

Fig. 6.19: A current-boosted cascode having a gain of 80 (38dB).

The completed schematic is shown in fig. 6.19.
Because grid-leak bias has been used for the upper triode the quiescent current is higher than predicted, but the average current falls to the expected value when the stage is overdriven. The measured gain was 80, and the maximum output signal

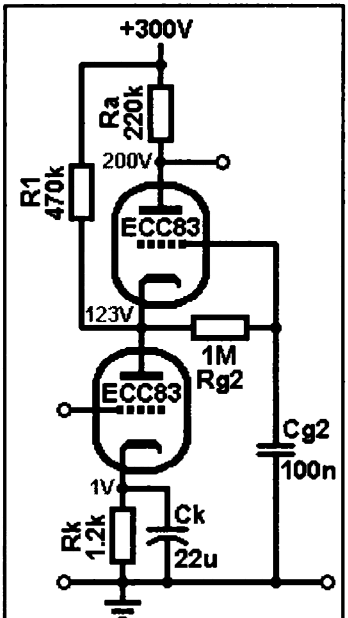

Fig. 6.20: A current-boosted cascode having a gain of 230 (47dB) but very high input-sensitivity and limited output signal swing.

swing just under 200Vp-p. Adjusting Ra to 100kΩ and Rk to 200Ω this circuit achieves a gain of 115 with the same output signal swing, which is a remarkable feat for a so-called 'low gain' valve! (Although, due to the hot bias, the stage must be heavily overdriven to obtain the full output swing.) Fig. 6.20 shows a similar circuit using the ECC83 / 12AX7. This achieves an extraordinary gain of 230! However, the price paid for such high gain is that at full overdrive the output signal swing is only 110Vp-p. Of course, less interstage attenuation could be used after this cascode. This circuit reaches cut-off first, at an input level of only 350mVp-p, so it best placed at the input of the preamp. It makes a useful alternative to the Vox style EF86 pentode input [see chapter 3, fig. 3.8] since, by using triodes, it should be less prone to microphonics, although at such high levels of gain even some triodes will give trouble. If necessary, R1 may be removed or switched out, reducing the gain to around 150.

A final point worth mentioning about the current-boosted cascode is that cut-off occurs more abruptly, when overdriven, than with the other forms of cascode (this is true for any triode type). This is illustrated by the oscillogram in fig. 6.21, which shows the 200Vp-p output from the cascode in fig. 6.19 when overdriven by a 12Vp-p input signal (100kΩ source impedance).

The reason for this very sharp cut-off is unclear, but the author suggests that when an 'ordinary' cascode approaches cut-off the reduction in current in the upper triode causes its r_a to rapidly increase, reaching hundreds of meg-ohms at cut-off, causing the lower triode's load line to become less steep as cut-off is approached, which 'smears' the transition into cut-off. With the current-boosted version the load formed by the upper triode is in parallel with R1, which sets an upper limit above which the load impedance cannot rise, reducing this smearing effect.

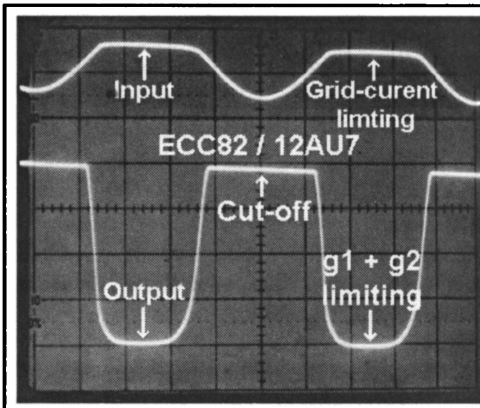

Fig. 6.21: Cut-off in the current-boosted cascode happens very sharply (compare with fig. 6.11).

Summary of formulae:

XXXVII; Gain of a cascode with fully bypassed cathodes:

$$A = \frac{\mu_1(\mu_2+1)Ra}{Ra + r_{a2} + r_{a1}(\mu_2+1)}$$

XXXVII; Gain of a cascode with fully bypassed cathodes and identical triodes:

$$A = \frac{\mu(\mu+1)Ra}{Ra + r_a(\mu+2)}$$

XXXIX ; Gain of a cascode with fully bypassed cathodes and identical triodes when $\mu >> 1$:

$$A \approx \frac{g_m.Ra}{1 + \frac{Ra + r_a}{\mu.r_a}}$$

If lower Rk is unbypassed then g_m should be replaced by:

$$g_{m(eff)} = \frac{gm}{1 + gmRk}$$

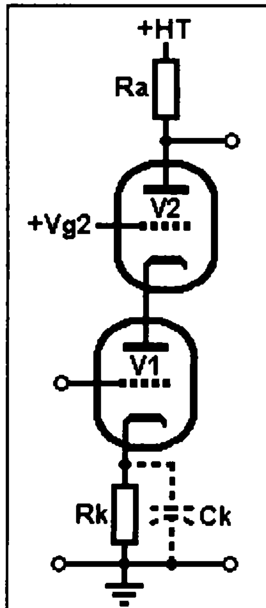

Fig. 6.22.

Half-boost frequency due to cathode-bypass capacitor:

$$f_{(half-boost)} \approx \frac{1}{2\pi CkRk}.\sqrt{1 + \frac{Rk(\mu_l+1)}{2r_{al} + \frac{1}{2}Rk(\mu_l+1)}}$$

Where:
f = the half-boost frequency.
μ_l = amplification factor of the lower triode.
r_{al} = anode resistance of the lower triode.

XL; Total input capacitance:

$$Cin = Cgk + Cga\left(\frac{\mu_1.r_{k2}}{r_{a1} + r_{k2}}\right)$$

Where:

$$r_{k2} = \frac{Ra + r_{a2}}{\mu_2 + 1}$$

But since $\mu_1.r_{k2} / (r_{a1} + r_{k2})$ is rarely much greater than unity, Cin may be approximated as:

$$Cin \approx Cgk + Cga$$

L; Output impedance:

$$Zout = Ra \parallel r_{a2} + r_{a1}(\mu_2 + 1)$$

For most triodes $r_a + r_a(\mu + 1) >> Ra$, so Zout may be approximated as:

$$Zout \approx Ra$$

Where in all cases notations are as in fig. 6.22. Subscript 1 indicates the lower triode, V1, and subscript 2 indicates the upper triode, V2.

Chapter 7: The Cathodyne Phase Inverter

Fundamental parameters of the cathodyne phase inverter. Designing a cathodyne. Biasing arrangements for AC-coupled cathodynes. DC-coupled cathodynes. Overload characteristics of the cathodyne. Bootstrapping for gain. Summary of formulae.

Small, low-powered amplifiers –less than about 10W say– normally have a single power valve (such as an EL84 or 6V6GT) in the output stage. The preamp signal will be coupled to the grid of this power valve in the conventional fashion. More powerful amplifiers (which probably constitute the majority of valve guitar amps) have a push-pull, power output stage, since this is a more efficient and cheaper method of generating audio power. Therefore there will usually be a need to split the preamp's audio signal into two equal but 180° out-of-phase signals, each of which drives one half of the power output stage. Devices for generating these two signals from one are known as **phase splitters** or **phase inverters**, since each output-signal waveform will appear 'upside-down', or inverted, when compared with the other. Such a stage makes the transition from the normally single-ended, or **unbalanced** operation of the preamp stages to the push-pull, or **balanced** operation of the power amp.

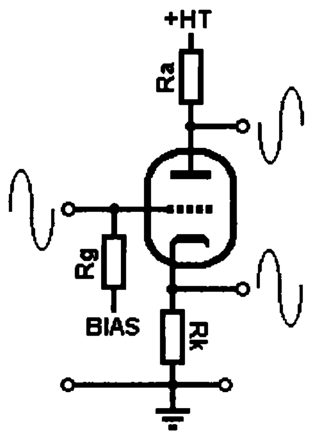

Fig. 7.1: A simplified cathodyne phase-inverter circuit.

The cathodyne phase inverter is perhaps the simplest and most efficient type of phase inverter found in valve amplifiers, and is ideal for driving more sensitive power output stages, such as those using EL84s or 6V6GTs, although it may be used with less sensitive valves too, if large amounts of power-valve distortion are not required. However, the cathodyne is remarkably uncommon in guitar amps, being found in many Ampeg amplifiers and famously in the Fender *5E3 Deluxe*, but was dropped in later models in favour of the long-tailed pair [chapter 8]. It is true that in many circuits its overdriven tone can be quite unpleasant, but this needn't be the case, if designed properly.
Fig. 7.1 shows a simplified cathodyne arrangement and we may immediately see that outputs are taken from both the cathode and the anode, so it appears to be a combination of cathode follower with an anode load, and an ordinary gain stage with a very large cathode resistor. Each of the outputs is 180° out of phase with the other, since when current through the valve falls, voltage drop across the anode load also falls and anode voltage therefore rises. At the same time, voltage drop across the cathode load falls and cathode voltage therefore also falls. This is to be expected of course, since we already know that a cathode follower is non-inverting, whereas a common-cathode gain stage is inverting, and this circuit may be regarded as a special combination of the two. Normally, the anode and cathode load resistors should be equal in value so that the signals generated across them are equal in amplitude, yielding balanced outputs. If they are not equal in value then the output signals will not be balanced (unequal in

amplitude). It is easy to see why the circuit is also known as the **split load** or **concertina** inverter, because the total load resistance is shared between anode and cathode, and the shape of the mirror-image output sine waves resembles a concertina's bellows.

Fundamental parameters of the cathodyne phase inverter:

Referring to fig. 7.1, the bias might be derived from a fixed or cathode-biased source, both will be described later.

Gain:

The gain to the anode is the same as that of an ordinary stage with an unbypassed cathode resistor, which is given by formula IV:

$$-A = \frac{\mu Ra}{Ra + r_a + Rk(\mu + 1)}$$

While the gain to the cathode is the same as a cathode follower, XXVIII:

$$A = \frac{\mu . Rk}{Ra + r_a + Rk(\mu + 1)}$$

But for normal purposes Ra = Rk. Hence, if we call the load at the anode or cathode R, the gain to anode *or* cathode is given by[1]:

$$A = \frac{\mu R}{r_a + R(\mu + 2)} \quad \text{...LI}$$

(Remembering that the gain to the anode should actually be –A, since it is inverting.)

From this it is clear that the gain to each output is identical, *provided both outputs are equally loaded*, remembering to take into account any additional AC load appearing in parallel with Ra and Rk. It is also clear that when equally loaded the gain to each output is always somewhat less than 1 (unity), the same as a cathode follower. However, the gain to both outputs combined, that is, the **differential gain**, is twice this value. In other words, we can expect the *difference* in AC voltage between anode and cathode to be about twice the input-signal amplitude. Rearranging and dividing by μ+1:

$$A = \frac{\mu}{\mu + 1} . \left(\frac{R}{\frac{R + ra}{(\mu - 1)} + R} \right)$$

[1] Jones. G. E, (1951). Analysis of Split-Load Phase Inverter. *Audio Engineering*, **35** (12), p16.

But for most valves μ is much greater than 1, so to a close approximation:

$$A \approx \frac{\mu}{\mu + 1}$$

Which is the same as the gain of a cathode follower, formula XXIX.
Also like the cathode follower, the cathodyne has 100% negative feedback between cathode and grid, so harmonic distortion and noise are reduced by a factor of $\mu + 1$ while bandwidth is extended by the same factor. Since a single-ended triode produces very little harmonic distortion even without feedback, the cathodyne becomes an extremely linear amplifier with a very wide bandwidth, making it invaluable for hifi applications.

Input capacitance:

The input capacitance is quite low, since the gain to the anode is very low, while Cgk is effectively divided by the action of feedback, giving:

$$C_{in} \approx C_{gk}(1 - A) + C_{ga}(1+A) \qquad \text{LII}$$

But since Cgk (1 – A) is extremely small, and A is normally close to unity, for audio applications this may be simplified to:

$$C_{in} \approx 2C_{ga}$$

For an ECC83 this is only 3.2pF, so large grid stopper may be used without treble attenuation.

Input impedance:

In the circuit in fig. 4.1 the input impedance is simply equal to the grid-leak resistor, which could be any convenient value, provided it does not exceed the valve's ratings. Alternatively the input impedance can be made very great if cathode-biasing is used.

Output impedance:

The output impedance of the cathodyne is a troublesome subject, as it varies hugely depending on whether or not it is equally loaded. Assuming it is equally loaded, the output impedance from either output is[2]:

$$Z_{out} = \frac{R.r_a}{r_a + R(\mu + 2)} \qquad \text{LIII}$$

But since μ is normally much larger than 1, this may be approximated as $1/g_m$, which is the same as for the cathode follower. The fact that not only is the output impedance equal at both outputs, but also very low in value (usually less than 1kΩ) has puzzled many authors, yet it is true. The theory behind this is beyond the scope of this book; interested readers are referred to Jones[3] and Preisman[4]. In fact, the

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[2] Jones, M. (2003).Valve Amplifiers (2nD ed.), p280. Newnes, Oxford.
[3] Jones. G. E. (1951). Analysis of Split-Load Phase Inverter. *Audio Engineering*, **35** (12), pp.16+40-41.

cathodyne gives guaranteed, perfect balance, *only* when both outputs are loaded identically! As a result it is a very useful circuit for hifi, since it requires no special effort on the part of the designer to achieve perfect balance.

However, if the outputs are *not* equally loaded then the output impedance rises. From the anode it reaches a value the same as that of a conventional gain stage with an unbypassed cathode resistor, being given by formula IX:

$$Z_{out(a)} = R_a \parallel r_a + R_k(\mu + 1)$$

But since $R_k(\mu+1)$ is normally very large compared with Ra, this may be approximated as:

$$Z_{out(a)} \approx R_a$$

While the output impedance from the cathode becomes the same as that of a cathode follower:

$$Z_{out(k)} = R_k \parallel \frac{R_a + r_a}{\mu + 1}$$

But since Rk is normally much larger than $R_a + r_a/(\mu+1)$ it may usually be ignored.

When not equally loaded the output impedance from the anode rises to a high value, whereas that from the cathode is hardly affected, and is still typically less than 1kΩ. This has significant implications for the overdriving characteristics of this circuit, which will be described later.

Output bandwidth:

There is a persistent myth that cathodyne must become unbalanced at high frequencies since the interelectrode capacitances, Cga and Cgk, are unequal. However, the stray wiring capacitance and the input capacitance of the following valves will normally be roughly equal, and so large as to swamp these interelectrode capacitances, so that the balance normally remains excellent to well above 100kHz. The –3dB frequency at either output, caused by *equal* capacitative loading of each output, is given by[1]:

$$f_{-3dB} = \frac{1}{2\pi RC.\frac{r_a}{\mu+2} \Big/ \left(R.\frac{r_a}{\mu+2}\right)} \quad ....LIV$$

Where:
R = Ra = Rk
C = Ca = Ck, as shown in fig. 7.2.

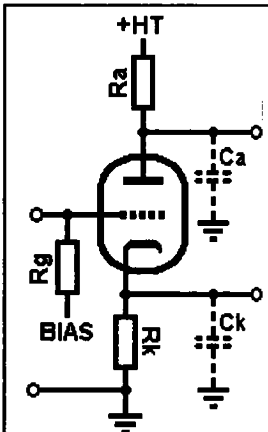

Fig. 7.2: With equal loading capacitances Ca and Ck, the cathodyne maintains perfect balance.

[4] Preisman, A. (1960). Notes on the Cathodyne Phase-Splitter. *Audio,* (April), pp22-23.

Since the output impedance of the cathodyne when equally loaded is very low, quite large loading capacitances may be applied without affecting audible frequencies. With most circuits, to obtain a -3dB frequency of around 20kHz would require about 8nF of capacitance on both outputs!

Designing a cathodyne:

Many of the design considerations for the cathodyne are similar to those of the cathode follower. Lower load resistances increase its ability to drive heavier loads, but reduce its linearity and output signal swing. However, a phase inverter is not normally expected to drive very heavy loads. Instead it only has to drive the power valves, whose input impedance is normally determined by grid-leak resistors, which will be several hundred kilo-ohms in value. These are labeled 'load' in fig. 7.3. Instead, output-signal swing is the more significant factor when choosing Ra and Rk.

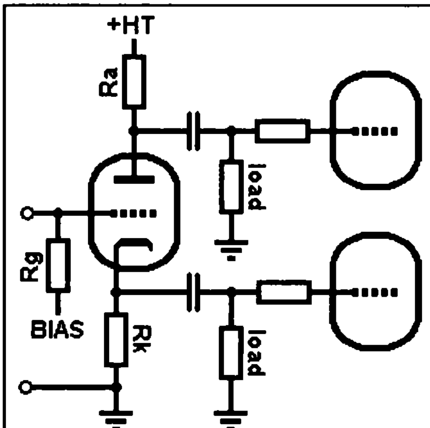

Fig. 7.3: The cathodyne is normally loaded by the grid-leak resistors of the power valves.

A load line can be drawn for the cathodyne in exactly the same way as for any common-cathode, triode gain stage, except that we must remember to take into account both Ra *and* Rk. From chapter 1 we may recall that the output signal swing of a triode is greatest when the load resistance is large, although making it too large will push the load line down into the region of unpredictable operation. Usually there is little to be gained by making Ra and Rk much larger than about 1/5th the value of the following loading resistance.
Additionally, a suitable value for Ra and Rk depends on whether or not we want to heavily overdrive the power valves. For a relatively clean amp less signal swing is required, and so much lower values may be used.

For example, if the power valves are EL84s or 6V6GTs, a typical value of grid leak might be 470kΩ. For a clean amp we might settle for values of Ra and Rk around 1/10th of this, say 47kΩ if we are using an ECC83 / 12AX7, or even less with a lower r_a triode, since these are naturally capable of greater output signal swing for a given load. With an ECC82 / 12AU7, for example, values of 33kΩ to 22kΩ would not be inappropriate. For higher levels of overdrive we might increase this to 100kΩ with an ECC83, or around 47kΩ to 68kΩ with a lower r_a triode.

Suppose we choose to use an ECC83 with an HT of 300V, and we have decided to make Ra and Rk equal to 100kΩ. Remembering that the *total* load on the cathodyne is Ra + Rk, fig. 7.4 shows a load line of 100kΩ + 100kΩ = 200kΩ.

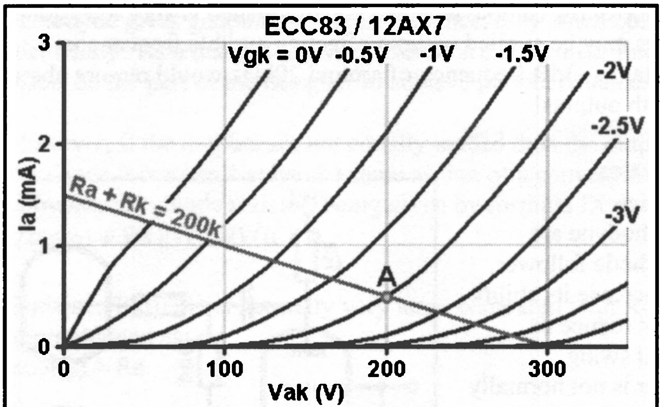

Fig. 7.4: A load line drawn for Ra = Rk = 100kΩ.

The maximum output signal swing is about 250Vp-p. However, this is shared equally between anode and cathode, so that the maximum output signal swing from anode *or* cathode is half this value, about 125Vp-p. This could also be deduced by noting that the maximum anode *current* swing is about 1.25mA. Ohm's law indicates the voltage across a 100kΩ resistor to be V = IR = 0.00125 x 100k = 125V , which supports our observation. This is still more than enough to overdrive an EL84 or 6V6GT to quite extreme levels, and is enough to drive larger valves such as EL34s, 6L6GCs and KT88s to full output, though not into such heavy overdrive.

Of course, since the cathodyne has roughly unity gain to either output, the input signal must therefore be 125Vp-p (actually slightly more in practice) to achieve maximum output swing, assuming the stage is centre biased. It should also be obvious that a greater output signal swing may be achieved by using a higher HT voltage.

A suitable bias voltage might be Vgk = –2V, since this is roughly central (noting that the valves appears to reach cut-off around –4V) and is labeled 'A' in fig. 7.4. The anode current at this bias point is 0.5mA and the anode-to-cathode voltage is 200V. We may then calculate the voltage dropped across each resistor to be (HT – Vak) / 2 , or by knowing the anode current we could calculate the voltage dropped to be V = IR = 0.0005 x 100k = 50V . The cathode voltage must therefore be 50V, and the anode voltage is HT – 50 = 250V. This is not unexpected, since in chapter 1 it was mentioned that for centre biasing the anode voltage will normally be around $\frac{1}{2}$ to $\frac{2}{3}$ of the HT, but for the cathodyne this is shared between anode and cathode so the quiescent cathode voltage will normally be around $\frac{1}{4}$ to $\frac{1}{3}$ of the HT, and the anode voltage lower than the HT by the same amount. This is normally low enough that it will not exceed the maximum allowable heater-to-cathode voltage, so heater elevation is not essential in most cases.

Biasing arrangements for AC-coupled cathodynes:

When AC coupling, the cathodyne may be treated in the same was as a cathode follower, and the usual biasing arrangements are the same.

Fixed bias:

Using the previous example, the cathode voltage was found to be 50V and the bias voltage -2V, so the grid voltage must be 48V, although, simply setting the grid voltage at 50V would probably be close enough, since the heavy cathode-current feedback will keep the bias within a very narrow range.

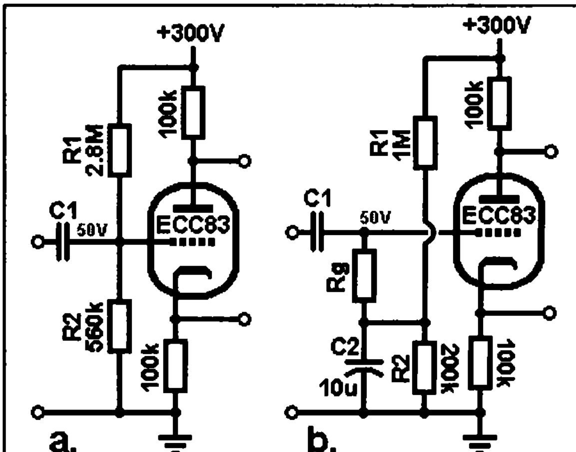

Fig. 7.5: Fixed-biased cathodynes. Circuit **b.** affords a greater input impedance more easily.

Fig. 7.5 shows the same biasing arrangements as were shown in chapter 5, fig. 5.5.

The circuit in *a.* applies the bias directly to the grid via a potential divider from the HT.

Unfortunately, because we would normally like to obtain a reasonably high input impedance, which implies making R2 large, R1 must be *very* large since the voltage dropped across it is very great. If R2 is chosen to be 560kΩ, then for a grid voltage of 48V, R1 must be:

$$R1 = \frac{HT - Vg}{Vg / R2} = \frac{300 - 48}{48 / 560k}$$

= 2940kΩ.

This could be closely approximated using three 1MΩ resistors in series, although 2.8MΩ is a reasonably close standard, yielding an actual grid voltage of:

300 x [560k / (2800k + 560k)] = 50V , which is sufficiently close.

The input impedance is equal to R1||R2, or 467kΩ in this case, which is reasonable.

C1 is required to 'block' the bias voltage from interfering with the previous stage, and its value is chosen in conjunction with the input impedance, in the same way as for any coupling capacitor.

If a higher input impedance is required then the arrangement in *b.* may be used. The bias voltage provided by the potential divider is heavily filtered by C2, and this bias voltage is applied to the grid via the grid-leak, Rg, which could be made any desired value, providing it does not exceed the datasheet maximum.

The minimum required value of C2 (to obtain good filtering down to 1Hz) is given by:

$$C = \frac{1}{2\pi R}$$

Where:

$R = R1||R2||Rg$

Although an arbitrarily large value of about 10µF will normally suffice.

Cathode bias:

As with the cathode follower, cathode biasing is useful since it bootstraps the grid-leak resistor, allowing a very high input impedance to be obtained. The bias is provided by 'tapping off' the necessary grid voltage from the cathode resistor, as shown in fig. 7.6, thereby placing the grid at a lower potential than the cathode by the desired amount. The potentiometer is shown for illustrative purposes; in reality a bias resistor is used, as shown in fig. 7.7. The bias resistor, Rb, may be found in exactly the same way as for a normal gain stage, either by drawing a cathode load line, or simply by calculating its value from the bias point.

Referring to the load line in fig. 7.4 the bias was found to be around -2V, and the quiescent anode current is about 0.5mA. The bias resistor would therefore be:

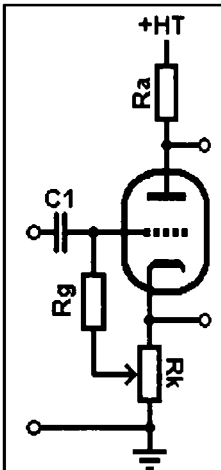

Fig. 7.6: The necessary grid voltage may be 'tapped off' the cathode resistor.

$R = V/I$
$= 2 / 0.0005$
$= 4k\Omega$.

The nearest standard value from the E12 range is 3.9kΩ. Of course, the *total* cathode load resistance is increased by the addition of this resistor, so that the anode and cathode loads are now not perfectly equal. This will unbalance the stage slightly. However, for guitar purposes this small unbalance is not important; in fact, it will promote some even-order harmonic distortion in the power output stage, giving a slightly more complex tone. It should perhaps be pointed out that almost none of the valve guitar amps ever produced have perfectly balanced phase inverters, and most players prefer the more 'colourful' quality of power amp distortion this produces. This is not

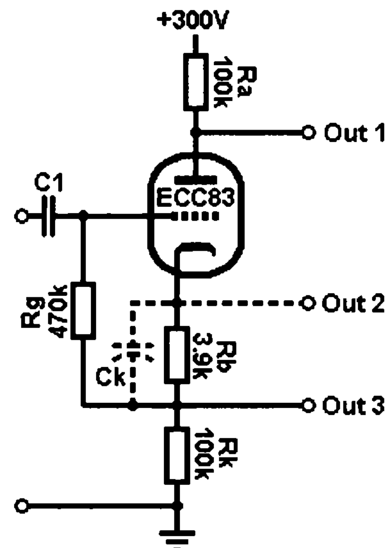

Fig. 7.7: A practical cathode-biased cathodyne.

true of hifi amps of course, in which it is common to see a bypass capacitor in parallel with Rb (Ck, shown dashed) so that the AC loads are once again equal. The modern hifi approach would be to use diode biasing.

Fig. 7.7 also shows how the cathode output signal might be taken directly from the cathode (output 2) or from the junction of Rb and Rk (output 3). There is little to differentiate the two, except that output 3 is slightly more preferable for guitar use, since Rb thereby appears in series with the output signal, which assists in maintaining a good overdriven tone. This will become clear later.

Because Rg is bootstrapped, the input impedance is increased. The formula is the same as that for the similar cathode follower circuit, XXXV:

$$Zin = \frac{Rg}{1 - \frac{A.Rk}{Rk + Rb}}$$

For the load line in fig. 7.4 we find $r_a = 80k\Omega$, $\mu = 100$. Applying formula LI we find the gain to one output to be:

$$A = \frac{\mu R}{r_a + R(\mu + 2)} = \frac{100 \times 100k}{80k + 100k\,(100+2)}$$
$$= 0.97$$

So for the circuit in fig. 7.7 the input impedance is:

$$Zin = \frac{Rg}{1 - \frac{A.Rk}{Rk + Rb}} = \frac{470k}{1 - \frac{0.97 \times 3.9k}{3.9k + 100k}}$$
$$= 7078k\Omega$$

This value is so large than the input capacitor, C1, need only be small to pass all audible frequencies. However, this formula assumes that all of the signal voltage appearing at the junction of Rb and Rk actually reaches the grid. In reality the output impedance of the previous stage forms a potential divider with Rg, so that only a portion of the cathode voltage actually reaches the grid, reducing the feedback factor and therefore the 'ideal' value of input impedance may not be achieved. Fortunately this is of little concern for guitar purposes, except that it is advisable to make C1 larger than might be thought necessary, to avoid unexpected loss of bass. Usually it will be whatever convenient value is already to hand, say 10nF. Bass frequencies may be more accurately controlled using output coupling capacitors instead.

If Rb is bypassed (or diode biasing is used) the input impedance is given by formula XXXVI:

$$Zin = \frac{Rg}{1 - A}$$

or about 15.6MΩ in this case! Such a high value is well beyond the requirements of a conventional preamp.

DC-coupled cathodynes:

A cathodyne is an obvious candidate for DC coupling, since its grid is naturally required to be at a high voltage. Similar design considerations apply as were described in chapter 2, *DC Coupling*, and need not be repeated. Again, some juggling of valves and values may be required to produce a suitable circuit, particularly since the required grid voltage is normally rather lower than that of, say, a cathode follower. For example, the ECC83 with 100kΩ load resistors shown earlier requires a grid voltage of around 48V. Obviously the anode of the preceding stage is unlikely to rest at such a low voltage, even if it is a pentode, but level shifting is a possibility. Fig. 7.8 shows a theoretical example in which the preceding stage has a quiescent anode voltage of 150V. This is level shifted down to 48V and a treble-boosting capacitor could also be added (shown dashed), and may be found using formula XIV. In this case a value of 10nF would give a half-boost frequency of about 30Hz.

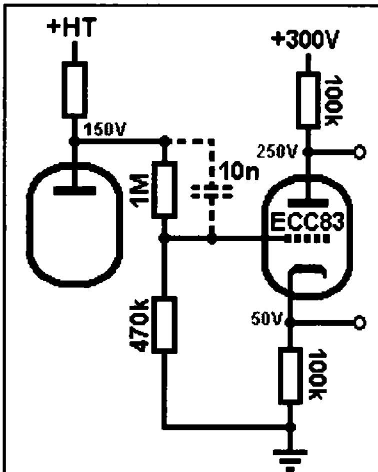

Fig. 7.8: A level-shifted, DC-coupled cathodyne.

It is very common to find a cathodyne directly coupled to the anode of the previous stage, and this will be made easier if both triodes have a low r_a, since the quiescent anode voltage of the previous stage will normally be reasonably low, and the required grid voltage of the cathodyne will be somewhat higher than is possible with the ECC83. Biasing the cathodyne hotter will permit a higher grid voltage, although clipping due to grid-current limiting will happen much sooner than cut-off, as a result. In an ordinary gain stage this would not be out of the ordinary but, as will be learnt later, a cathodyne can suffer overload problems and usually benefits from being roughly centre biased.
For example, fig. 7.9 shows a direct-coupled cathodyne using an ECC82 with arc

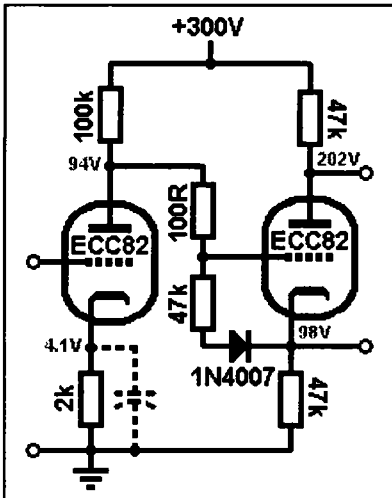

Fig. 7.9: Effective direct coupling is facilitated by using low r_a triodes.

protection. Even with this low r_a, valve the cathodyne is still warm biased, and the output signal swing is limited to about 65Vp-p before clipping due to grid-current limiting occurs, even though an output amplitude of 130Vp-p is theoretically available. Of course, this is not a problem in a hifi amp which only needs enough signal swing to drive the power valves to full power and no further, but in a guitar circuit we can expect any or all stages to be routinely overdriven, and this is where the cathodyne's performance can suffer if we continue to treat it as a textbook, hifi circuit.

Overload characteristics of the cathodyne:

Probably the main reason why the cathodyne is not found in guitar amps more often is its peculiar overload characteristics, which can produce a very ugly tone, both when the power valves are overdriven and when the cathodyne is itself overdriven. But this *is* avoidable, if we are willing to abandon some 'traditional' design attitudes.

The 'gain-spike' effect:

Firstly, let us consider what happens when the cathodyne begins to overdrive the power valves, since this will usually be the first to happen. Fig. 7.10 shows a conventional circuit arrangement, except that the usual grid stoppers have been removed, and waveforms are shown. As the input signal moves positively, the signal at the cathode of the cathodyne, V1, follows it. Eventually the grid of the power valve, V3, will be driven sufficiently positive that it begins to draw grid-current. At this point its grid signal quickly becomes 'clamped', and cannot rise any further; attempting to drive the grid even more positive simply increases the current flowing into the grid (and the coupling capacitor, C2, begins to charge up). The input signal may continue to rise, but the grid of V3 –and therefore the cathode of V1– cannot rise because it is clamped. *If the cathode voltage is no longer changing, this is effectively the same as if Rk were bypassed by a large cathode bypass capacitor*, and we know from chapter 1 that holding the cathode voltage constant allows the valve to achieve maximum gain to the anode. The gain to the anode is normally about unity, but as soon as V3 begins to draw grid current the gain to the anode leaps to its maximum level. At this point the anode signal is moving negatively of course, so a pronounced spike occurs when the gain increases.

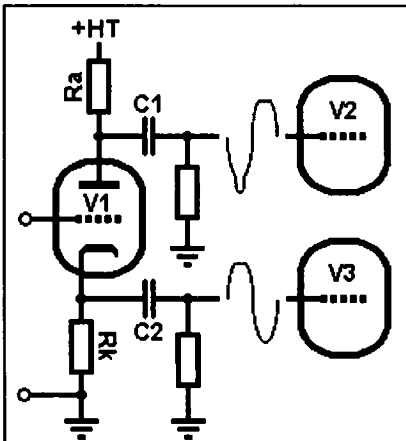

Fig. 7.10: Overdriving V3 causes a 'gain spike' to appear at V2.

When the anode output moves positively and the anode beings to overdrive V2, the anode voltage becomes clamped in the same way, and Ra effectively becomes bypassed and V1 becomes a pure cathode follower. However, the gain and performance of a

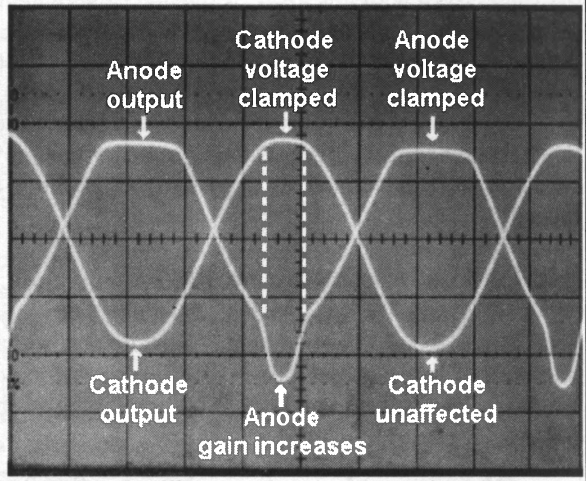

Fig. 7.11: Oscillogram showing the 'gain spike' effect when the power valves are overdriven. 1kHz, 5V/div.

cathode follower is hardly any different from the cathodyne, so there is no obvious change in the cathode output signal. In other words, the anode output is heavily influenced by the way in which the cathode is loaded, but the cathode is almost completely immune to any loads placed on the anode. This is clearly seen in the oscillogram in fig. 7.11 which was produced from the circuit in fig. 7.9 when overdriving a pair of EL84s with 1kΩ grid stoppers.

If the power output stage is class AB, as most push-pull guitar amps are, then this negative 'gain spike' is not always a concern because the corresponding power valve (V2 in fig. 7.10) is already in cut-off when the negative spike occurs. However, if the power valves are poorly matched, or if the power output stage is class A, then this spike may be amplified by the corresponding power valve and will introduce high-order intermodulation distortion products, and is rarely pleasant.

The frequency-doubling effect:

Now let us examine what happens when the cathodyne is itself overdriven. Cut-off is quite normal, and needs no special discussion. In any case, in most circuits the cathodyne will be biased sufficiently warm that cut-off is never reached. When the cathodyne is overdriven positively, however, trouble arises.

As the grid voltage rises, eventually the cathodyne begins to draw grid current. The grid current flows into the grid and down the cathode resistor, thereby contributing to the cathode output signal's positive-going cycle. As the grid is driven more and more positive, eventually a point will be reached where the valve is conducting the maximum amount of anode current it can, and driving the grid even more positive cannot increase this any further. The anode-to-cathode voltage therefore reaches a minimum value, and can go no lower. But as the grid is driven more positive, increasing grid-current continues to flow into the cathode and down Rk, so that the voltage across Rk must increase as a result, according to Ohm's law. Since the cathode voltage is continuing to rise, but the anode-to-cathode voltage cannot fall any further, the anode voltage must also rise; the anode voltage is effectively 'jacked up' by an amount equal to Ig x Rk. This causes an inverted 'copy' of the cathode output signal peak to appear at the anode, resulting in a kind of frequency doubling, although because it is rectified it is a very 'raspy' sound. The process is illustrated in

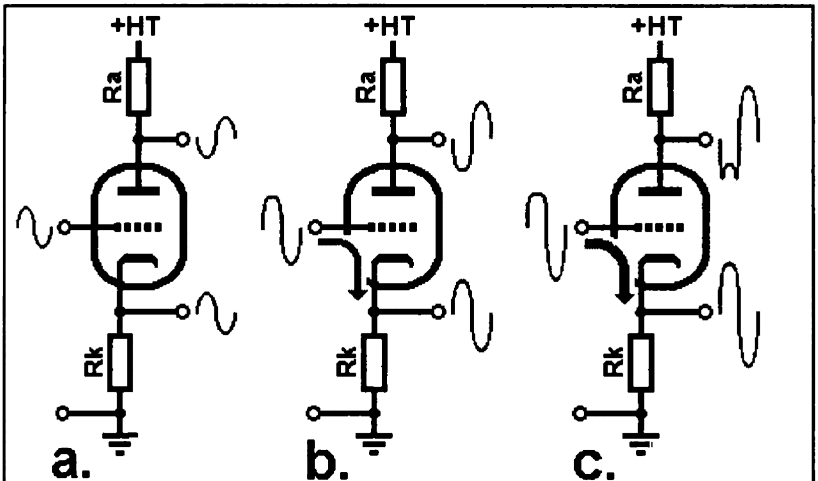

Fig. 7.12: The frequency-doubling effect is due to the cathodyne being heavily overdriven. The arrows indicate the flow of grid-current. **a.** before grid-current, **b.** mild grid current, **c.** heavy grid current and frequency doubling at the anode.

fig. 7.12, and the oscillogram in fig. 7.13 was produced from the circuit in fig. 7.9 under heavy overdrive (without power valves; the visible clipping is actually due to the previous gain stage and not to cut-off in the cathodyne).

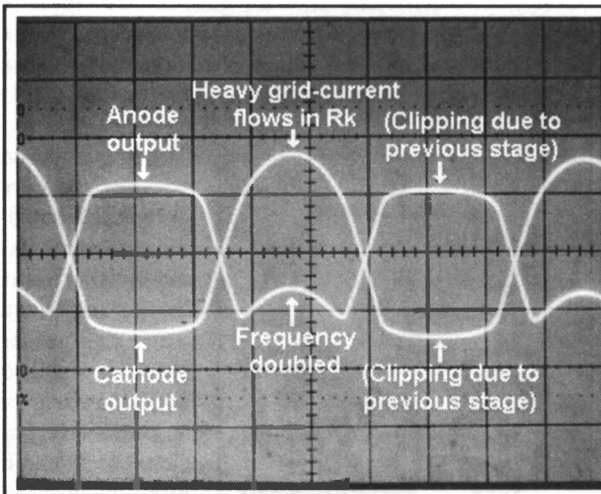

Fig. 7.13: Oscillogram showing the frequency-doubling effect when a cathodyne is heavily overdriven. 1kHz, 25V/div.

Depending on the specific circuit and the degree of overdrive, it is possible for the frequency doubling effect to cause V3 (fig. 7.10) to come out of cut-off for part of the half-cycle, and even for the anode output signal to become almost completely frequency doubled. In very lucky situations the effect can be quite pleasant, introducing a phaser-like sound sometimes known as **swirl**, as the dominant frequency of overdrive varies with the audio signal. However, it is *extremely* difficult to control this effect; the intermodulation products nearly always result in a very ugly, intermitted fuzz sound, rather like that of digital 'garble' interfering with the true tone of the amp. It also encourages very problematic blocking distortion, and for most builders is to be avoided at all costs.

Avoiding unpleasant distortion effects:

Since all the problematic effects of the cathodyne are related to too much current drawn from it, or into it, the logical way to avoid them is to limit this current. This is easily done with large grid stoppers, and this might be regarded as the 'secret' to obtaining a good, consistent tone with the cathodyne, whether AC or DC coupled.

Using larger grid stoppers on the power valves will help prevent the cathode voltage ever becoming clamped by grid current, since the load on the cathode cannot fall below the value of the grid stopper, which will avoid the 'gain spike' effect. In fact, using larger-than-usual grid stoppers on power valves is a useful way to obtain a consistent, crunchy sound, regardless of what is driving them.

Because the power valves are nearly always pentodes or beam tetrodes, which have very little input capacitance, large grid-stoppers may be used without affecting the audible treble response. The 'traditional' grid stoppers used on power valves are normally in the region of 1kΩ to 4.7kΩ, but these can easily be increased, the only proviso being that the total grid-leak resistance must not exceed the maximum allowable value, since the grid stopper is usually part of this resistance. A value of 100kΩ will virtually eliminate the gain-spike effect in most circuits. Large grid-stoppers will greatly reduce the chances of blocking distortion too, and no doubt they would be more commonly found in modern amps were it not for a lack of real valve-circuit knowledge within modern amp companies.

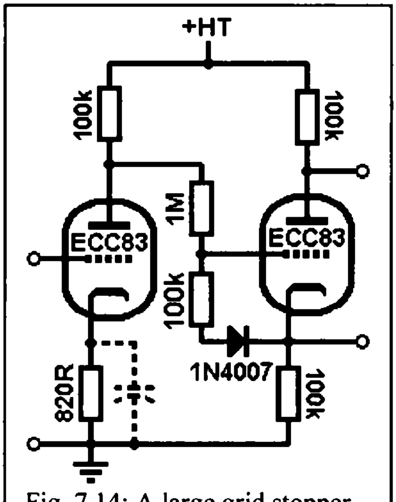

Fig. 7.14: A large grid stopper prevents the frequency doubling effect and helps high r_a valves to be DC coupled to the cathodyne.

The frequency-doubling effect may be alleviated to a great extent by using a large grid stopper on the cathodyne. Again, because the cathodyne has roughly unity gain its input capacitance is extremely low, so a large grid stopper may be used without affecting the treble response, and will also help prevent blocking distortion if AC coupling is used, of course. Values as large as 1MΩ may be required in high-gain designs, and a practical circuit is shown in fig 7.14.

From the discussion about quiescent grid current in cathode followers (chapter 5, fig. 5.16), under similar circumstances we can expect an ECC83 cathodyne to draw *even more* quiescent grid current, since it naturally requires a lower grid voltage. This greatly increases the frequency-doubling effect and also results in *extreme* unbalance of the output signals at all but the lowest signal levels. However, in fig. 7.14 this quiescent grid-current flows through the 1MΩ grid stopper, causing a voltage drop which lowers the grid voltage to a more appropriate value, as well as suppressing the frequency doubling effect. No siginificant loss of treble is suffered, thanks to the cathodyne's low input capacitance. The maximum, peak-to-peak output signal swing from the cathodyne is roughly ⅓HT, which is only slightly less than the theoretical maximum.

Bootstrapping for gain:

Because the cathodyne offers a low-impedance output from its cathode, it is possible to bootstrap the anode resistor of the previous stage to increase its gain[5], in the same manner as explained in chapter 5. Such circuits have appeared in a few hifi amps, though the author knows of no commercial guitar amp ever to employ it.

The anode resistor of the preceding gain stage is split roughly equally into two parts and the cathode output of the cathodyne is coupled to the junction via a coupling capacitor, C1, as shown in fig. 7.15. The same principles apply here as for the cathode follower and need not be repeated. It should be pointed out though, that as far as AC is concerned the total *cathode* load on the cathodyne is now formed by

$Rk2||R1||(R2+r_a)$

(Assuming Rk1 is fully bypassed).

So for reasonable balance from the cathodyne we should attempt to make Ra equal to this.

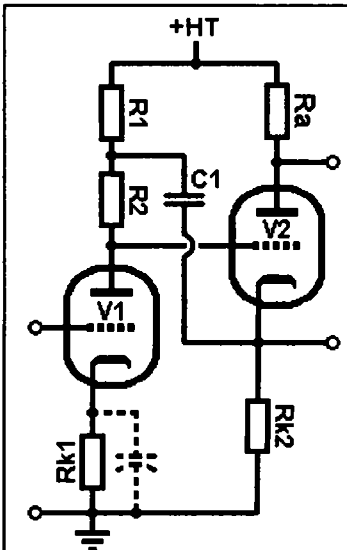

Fig. 7.15: Bootstrapping V1 to increase its gain.

A practical circuit is shown in fig. 7.16 (arc-protection components are omitted for clarity). The total AC load on the cathode is close to:

$Rk2||R1||(R2+Zout) = 100k||100k||(100k+65k)$

$= 38k\Omega$

Which is easily matched by a 39kΩ anode resistor, so balance of the circuit is quite good. The half-boost frequency caused by bootstrapping is approximately:

$$f \approx \frac{1}{2\pi RC}$$

Where:

$R = R1||(R2+r_a)$

C = coupling capacitor C1 (fig. 7.15).

So a 100nF capacitor gives a frequency of 26Hz.

[5] Jeffery, E. (1947). Push-Pull Phase-Splitter: New High Gain Circuit. *Wireless World*, (August), pp274-277.

The gain of V1 should equal:

$$A = \frac{\mu\left(R1 + \frac{R2}{1-A}\right)}{r_a + R1 + \frac{R2}{1-A}} = \frac{100\left(100k + \frac{100k}{1-0.97}\right)}{65k + 100k + \frac{100k}{1-0.97}}$$

= 98

While the gain to each output should then be 98 x 0.97 = 95.

The circuit was tested with a Mullard ECC83 and found to have a gain of just over 95 to each output, which agrees with the predicted value, although the maximum, peak-to-peak output signal swings are unequal when overdriven, being about ¼HT for the anode and ⅓HT for the cathode, due to the large difference in Ra and Rk2. This imbalance only arises during heavy overdrive conditions, however, and is not a great concern. Because the anode load of V1 is fairly large, resulting in a fairly low quiescent anode voltage, the cathodyne biases reasonably well and only a modest 100kΩ grid stopper is required to prevent frequency doubling.

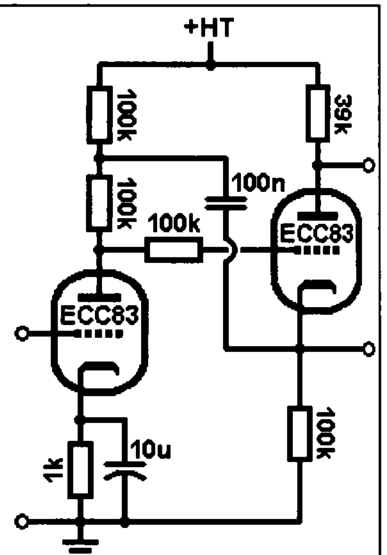

Fig. 7.16: A practical circuit having a gain of over 95 (40dB) to each output.

Summary of formulae:

LI; Gain to either output when equally loaded:

$$A = \frac{\mu R}{r_a + R(\mu + 2)}$$

But for most valves $\mu >> 1$, so this may be approximated as:

$$A \approx \frac{\mu}{\mu + 1}$$

LII; Total input capacitance:

$$C_{in} \approx Cgk\,(1 - A) + Cga(1+A)$$

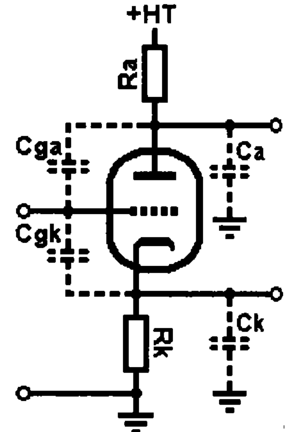

Fig. 7.17

But since Cgk (1 – A) is extremely small and A is normally close to unity, for audio applications this may be simplified to:

$$\text{Cin} \approx 2\text{Cga}$$

LIII; Output impedance from either output (when equally loaded):

$$\text{Zout} = \frac{R.r_a}{r_a + R(\mu + 2)}$$

But for most valves $\mu >> 1$, so this may be approximated as:

$$\text{Zout} \approx 1/g_m.$$

LIV; -3dB roll-off frequency due to equal loading capacitances, Ca = Ck = C:

$$f_{-3dB} = \frac{1}{2\pi RC.\frac{r_a}{\mu+2}\Big/\left(R.\frac{r_a}{\mu+2}\right)}$$

Where in all cases notations are as in fig. 7.17 and Ra = Rk = R.

Chapter 8: The Long-Tailed-Pair Phase Inverter

Basic operation of the long-tailed pair. Fundamental parameters of the long-tailed pair. Designing a long-tailed pair. DC-coupled long-tailed pairs. AC-coupled long-tailed pairs. Global negative feedback and the long-tailed pair. Presence control without global negative feedback. Scale control. Overload characteristics of the long-tailed pair. Summary of formulae.

The long-tailed pair is the most popular phase inverter circuit found in guitar amps. It is also known as the **differential amplifier, cathode-coupled inverter** or **Schmitt inverter** after Otto Schmitt described its operation in 1938[1], although the conventional form of the circuit can be traced at least as far as 1934[2]. The cathodyne phase inverted described in chapter 7 is well suited to driving relatively sensitive power valves to high levels of overdrive, but if we wish to drive larger valves to similar levels of distortion then we require large signal amplitudes. This could be obtained from a cathodyne operating under a high HT voltage, but this is not always available. The long-tailed-pair, however, offers a greater output signal swing for a given HT, although it requires two valves, which will usually be in the same envelope. It also has a useful level of gain, although it should be pointed out that in combination with an additional gain stage, a cathodyne offers more gain with the same two valves. It also seems to be the case that almost any type of triode gives a good tone when used in a long-tailed pair, even the ECC81 / 12AT7! The 12AY7 is also a popular choice for low to moderate gain designs, and has a particularly rich tone. The ECC82 / 12AU7 is also useful for very clean or bass amplifiers.

Basic operation of the long-tailed pair:

A simplified diagram of a long-tailed pair is shown in fig. 8.1, from which it can be seen that the circuit has two inputs (plus ground) and two outputs (plus ground), making it a six-pole, entirely balanced or push-pull circuit. It is clear than the anode current of each valve flows into the shared cathode resistor, Rk, which forms the 'tail' of the circuit.

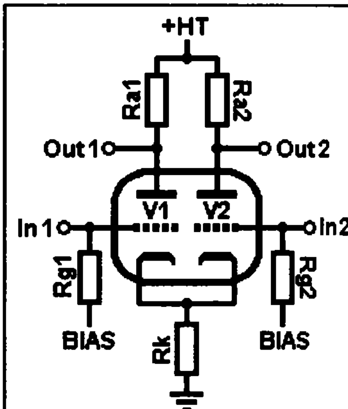

Fig. 8.1: Simplified diagram of the long-tailed pair.

Suppose we input a positive signal to V1, so its anode current increases. This current flows in Rk causing the voltage across it to increase, which is effectively the same as if the grid voltage of V2 had decreased, so the anode current through V2 decreases. In other words, V2 did the opposite of V1, so antiphase signals appear at each output. An alternative way to view it is to say that V1 acts as both an ordinary gain stage and as a cathode follower. The output from its cathode is

[1] Schmitt, O. H. (1938). Cathode Phase Inversion, *Journal of Scientific Instruments.* **15** (March), pp 100-101.

[2] Matthews, B. H. C. (1934). A Special Purpose Amplifier. *The Journal of Physiology*, 81, 28P-29P.

passed to the cathode of V2, which amplifies it but in a non-inverting fashion. Therefore the output from V1 is *out-of-phase* with the input of V1, while the output of V2 is *in phase* with the input of V1, so the circuit acts a phase inverter. The same arguments follow if we had input the signal to V2. In a perfectly balanced circuit the output signal amplitudes will be equal, which is only possible if each valve sees one half of the total input signal, from which we deduce that the signal appearing at the cathode must be exactly one half the amplitude of the input signal.

However, suppose we input the *same* positive signal to both valves simultaneously. Both valves attempt to increase their anode current. This flows in Rk, increasing the voltage across it; each cathode attempts to follow the corresponding grid. Under ideal circumstances the cathode voltage rises by the same amount as the voltage on each grid, so the total change in the *grid-to-cathode* voltage of each valve is zero. Therefore, there is no signal for the valves to amplify and we obtain no output. From these thought experiments we discover that the long-tailed-pair will only amplify the *difference* in voltage between its inputs, which is why it is also known as a differential amplifier. When we input the same signal to both grids the difference between the two is zero, so nothing is amplified. This property is known as **Common-Mode Ripple Rejection** or **CMRR**. It is one of the most desirable features of balanced amplifiers, since it discriminates between the wanted signal(s) and 'common-mode' hum and noise, that is, noise which is the same or 'common' to both inputs.

We could also input two completely different signals. The circuit would mix the two by amplifying the difference between them, and provide two out-of-phase output signals. Hence the long-tailed pair can be used as a phase inverter and as a phase-inverter/mixer, as shown in fig. 8.2.

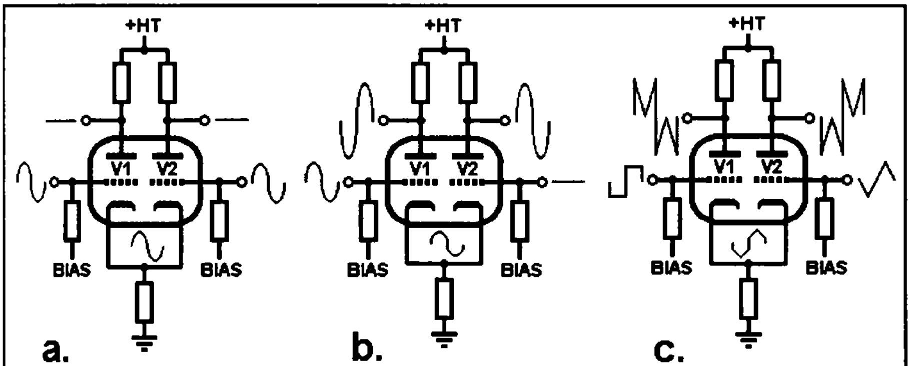

Fig. 8.2: The long-tailed pair only amplifies the difference between its inputs. In **a.** there is no difference between inputs so no output signal is produced. **b.** the circuit as a simple phase inverter. **c.** the circuit as a differential mixer and phase inverter.

Fundamental parameters of the long-tailed pair:

In most guitar amps the stage is used simply as a phase inverter in which the second grid is grounded -as far as AC is concerned- by a decoupling or bypass capacitor, Cg2, and the input signal is applied to the first grid.

This is shown in simplified form in fig. 8.3, and referring also to fig. 8.2*b* we see that the output from V1 is inverted, while that from V2 is non-inverting. If the second grid were not bypassed then the input signal may find its way onto the second grid and we would have no difference between the inputs and therefore obtain either no output at all, or very unbalanced output signals, depending on the circuit arrangement used.

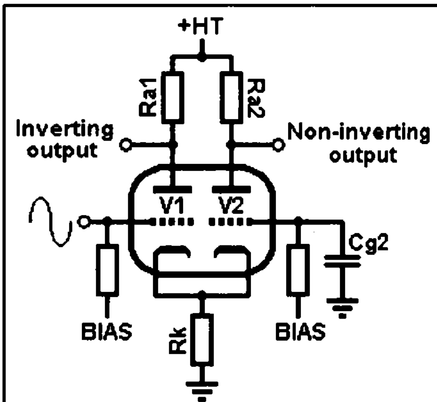

Fig. 8.3: Simplified diagram of the long-tailed pair as a phase inverter.

Gain:

Because V1 must also act as a cathode follower which has somewhat less than unity gain, not all of the input signal is passed on to V2, so the gain to each output may not be identical and the stage will be unbalanced. It can be shown that for identical triodes, if Ra1 = Ra2 = Ra, the gain to the non-inverting output is[3]:

$$A2 = \frac{\mu Ra}{(Ra + ra).\left(2 + \frac{Ra + ra}{Rk(\mu + 1)}\right)} \quad \text{.......LV}$$

But the gain to the inverting output will be greater than this by a factor of:

$$\frac{A1}{A2} = 1 + \frac{Ra + ra}{Rk(\mu + 1)} \quad \text{.......LVI}$$

From which we see that if the anode resistors are equal, the overall balance depends on the value of Rk, and perfect balance is achieved when Rk is infinite. If Rk is not infinite then balance is improved by using high-μ valves. We could also use mis-matched anode resistors to bring about equal gain to each output, which will be described later.

Alternatively we might measure the output signal *between* both outputs, giving us the *differential gain.* Assuming perfect balance this equal to:

$$A = \frac{\mu Ra}{Ra + ra}$$

Which is exactly the same as formula III, for a common-cathode gain stage. In other words, the gain to either output of a long-tailed pair is roughly half that which would

[3] Valley, G. E. & Wallman, H. (eds.) (1948). Vacuum Tube Amplifiers (1st ed.), p447. McGraw-Hill, New York.

be obtained from the same, single triode, if used as an ordinary gain stage with the same anode resistor. If a cathodyne were coupled to the same, conventional gain stage then that gain would be measured at *both* outputs, which demonstrates why it actually offers more gain from the same two valves.

Input capacitance:

The total input capacitance is given by:

$C_{in} \approx C_{gk}(1 - A_k) + C_{ga}.A_a$ LVII

Where:

A_k = the gain from grid to cathode, being approximately 0.5

A_a = gain to the anode, which will be approximately $A_a \approx \dfrac{\mu R_a}{2(R_a + r_a)}$

And since $C_{gk}(1 - A_k)$ is very small, this may be approximated to:

$C_{in} \approx C_{ga}.A_a$

Input impedance:

In the circuit in fig. 8.3 the input impedance is simply equal to the grid-leak resistor, which could be any convenient value, provided it does not exceed the valve's ratings. In other circuit arrangements, which will be discussed, the input impedance can be made much greater.

Output impedance:

If the circuit is perfectly balanced then the output impedance is the same as that of an ordinary gain stage, given by formula VIII:

$$Z_{out} = R_a \,||\, r_a = \frac{R_a.r_a}{R_a + r_a}$$

However, if the stage is unbalanced then the output impedance at the highest gain output will increase. However, the change is not very great, reaching a maximum of[4]:

$Z_{out} = R_a \,||\, r_a + R$

Where:

$$R = \frac{R_a + r_a}{1 + \dfrac{R_a + r_a}{R_k(\mu + 1)}}$$

[4] Jones, M. (1999). Valve Amplifiers (2nd ed.), p116. Newnes, Oxford.

If Rk and μ are large then $R \approx Ra + ra$ and the output impedance may be simplified to:

$$Zout \approx Ra \parallel 2ra + Ra = \frac{Ra}{2} \qquad \text{LVIII}$$

Input sensitivity:

Because half the input signal appears at the cathode of a long-tailed pair, the input sensitivity is halved when compared to the same triode is used as a conventional gain stage with a fully bypassed cathode, under similar conditions. For example, a typical ECC83 / 12AX7 gain stage has in input sensitivity of around 4Vp-p when centre biased. But when connected as a long-tailed pair this is halved to around 8Vp-p, making the long tailed pair less easy to overdrive (i.e. it has more headroom). Of course, this is still a good deal more sensitive than a cathodyne.

Designing a long-tailed pair:

For guitar purposes the design of the long-tailed pair is quite straight forward, since precise balance is not required. In fact, the small amount of imbalance which is inherent in most designs encourages even-harmonic distortion in both the phase inverter and the power output stage, which lends greater texture to the tone (this is in contrast to the misleading "technical advice" given in some guitar magazines that wrongly state that perfect balance is 'essential'). Normally it is in the nature of push-pull circuits to cancel out even-harmonic distortion, since asymmetrical non-linearity in one device is matched by the same non-linearity in the opposing device, but with reversed phase, so that the distortion products are equal but opposite, and cancel out when they are ultimately mixed together in the output transformer. Distortion in a perfectly balance long-tailed pair using triodes is dominated by third-order *intermodulation* distortion, with harmonic distortion appearing at a lower level[5]. Depending on the amount of imbalance, however, odd and even harmonic distortion can be increased relative to the intermodulation products, helping to develop the characteristic 'growly' sound of the long-tailed pair. On the other hand, making to the phase inverter *too* unbalanced can result in the opposite effect, making the tone too 'bland'.

The easiest way to design a long-tailed pair is to begin by deciding how much voltage can be spared for the tail resistor, Rk. A greater voltage implies a larger value for Rk, which improves the inherent balance of the circuit but also reduces the available output signal swing. Obviously, the greater the HT then the more voltage drop can be tolerated across Rk. A clean amplifier can spare the most voltage of course –say 0.2HT to 0.5HT– since a large output signal swing is not necessary and it is also more likely to benefit from a reasonably well balanced phase inverter, whereas a moderate to high-gain amp might allow around 0.1HT to 0.3HT.

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[5] Abuelma'atti, M. (2005). Large Signal Analysis of Differential Triode Tube Amp, *Audio Xpress* [online]. Available from: http://www.audioxpress.com/magsdirx/ax/addenda/media/muhammad.pdf

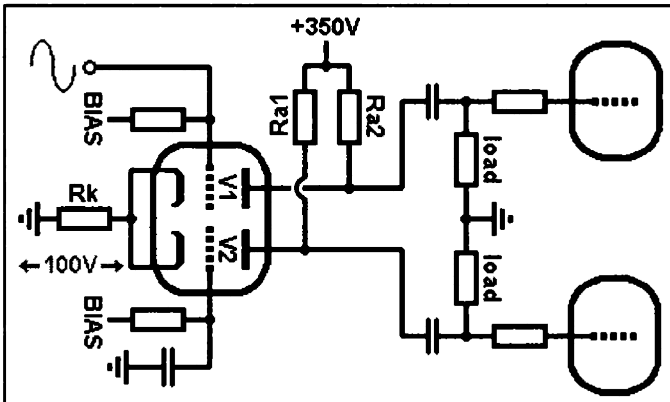

Fig. 8.4: A simplified circuit arrangement with 100V allowed across Rk.

For example, if the HT is 350V we might allow 100V for Rk, leaving 250V of 'effective HT' for the valves themselves, as shown in fig. 8.4. Since we know the output signal swing of a triode is typically around 2/3HT, we can predict about 170Vp-p output. This is certainly enough to overdrive almost any power valve.

We can now select a suitable load, and draw a load line using this 'effective HT' of 250V. For example, if we were to use an ECC83 / 12AX7 we might take a typical load of 100kΩ. The load line drawn in fig. 8.5 suggests about 170Vp-p maximum output signal swing, however, it is worth considering the additional load placed on the circuit by the power valves' grid-leak resistors (labeled 'load' in fig. 8.4). Normally we are free to ignore such following loads, but the maximum value of grid-leak for a power valve is much less than for a preamp valve. If the grid-leaks were 220kΩ each, which would be typical if we were driving EL34s or 6L6GCs, then the AC load on each triode is 100k||220k = 69kΩ. Choosing a typical bias point of Vgk = –1V the AC load line is shown in fig. 8.5, from which it can be seen that the actual output signal swing is reduced to about 140Vp-p. Fortunately this is still enough for most purposes.

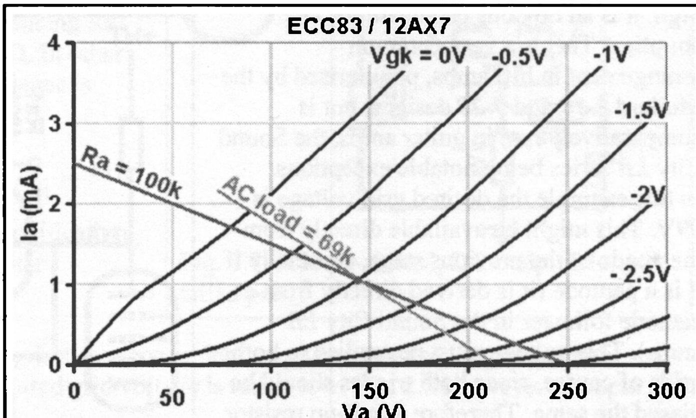

Fig. 8.5: DC, AC and cathode load lines drawn for an effective HT of 250V with a 100kΩ anode resistor and 220kΩ following loading resistance.

Cold biasing will encourage cut-off clipping in the phase inverter, but at the expense of output signal swing. The result is that the output valves may be driven less hard themselves, but more phase-inverter overdrive can be heard, shifting the tonal balance in favour of preamp distortion rather than power amp distortion. Warm biasing has the opposite effect; even though grid-current clipping will happen sooner in the phase inverter, this produces a down-going output swing which will not usually be amplified by the corresponding power valve since it will already be in cut-off (unless the output stage is Class A).

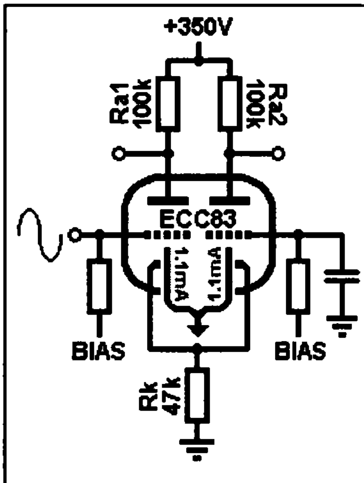

Fig. 8.6: A partially completed design showing anode current paths.

Having selected a bias point, the quiescent anode current can be found. In this case, fig. 8.5 indicates about 1.1mA. However, the current flowing in Rk is the sum of the currents in both triodes, giving twice this value, or 2.2mA. Since we initially allowed 100V across Rk, and we now know the current though it to be 2.2mA, applying Ohm's law yields its value:
$R = V/I = 100 / 0.0022$
$= 45.5k\Omega$.
So a typical value of 47kΩ is very close. Fig. 8.6 shows the circuit in a state of partial competition. Since the required bias is Vgk = –1V and the cathode voltage is 100V, we require a grid voltage of 99V. All that remains is to bias the circuit, by some means.

DC-coupled long-tailed pairs:

Because the required grid voltage of a long-tailed pair is often relatively high, it is an obvious candidate for DC coupling. This is a very common arrangement in hifi amps, popularised by the Mullard *5-10* and *5-20* designs, but is comparatively rare in guitar amps, the Sound City *LB* series being notable exceptions.
In this example the desired grid voltage is 99V. This might be available directly from the anode of the previous stage, especially if it is a pentode (it is derived directly from a cathode follower in the Sound City *LB* amps). This voltage must be applied to both grids of course, since both triodes should be biased the same. Therefore a pull-up resistor, Rg, is connected between the grids, ensuring the second grid is 'pulled up' to the same DC voltage as the first. Since the second grid is to be bypassed, this resistor will form an

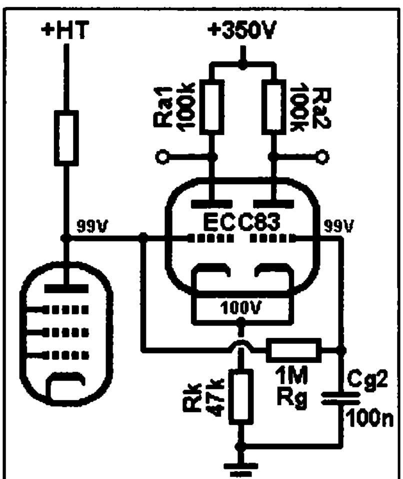

Fig. 8.7: A direct-coupled, long-tailed pair, typical of hifi designs.

AC Load on the previous stage, so its value should be large (but should not exceed the maximum allowable value of grid-leak of course) and 1MΩ is usual. Assuming we wish the second grid to be fully bypassed at all frequencies down to say, 1Hz, then we may find the value of Cg2:

$$C = \frac{1}{2\pi fR} = \frac{1}{2\pi \times 1 \times 1M}$$

= 159nF

A value of 150nF might be chosen, although a convenient value of 100nF will decouple down to 1.6Hz and is the usual choice. This is shown in fig. 8.7 (the obligatory grid-stopper and arc-protection diode are omitted for clarity [see chapter 2, *DC coupling*]). At switch-on, Cg2 may be exposed to the full HT voltage so should have a voltage rating greater than the HT.

It was noted earlier that unless Rk is infinite (or nearly so) the gain to the inverting output will be greater than the gain to the non-inverting output by a factor of:

$$\frac{A1}{A2} = 1 + \frac{Ra + ra}{Rk(\mu + 1)}$$

In this case the difference in gain is*:

$$\frac{A1}{A2} = 1 + \frac{100k + 65k}{47k \times (100+1)}$$

= 1.03

Or in other words, the inverting triode has 3% more gain than its partner, which is not a very great difference. However, if we were worried about this imbalance we might correct it by reducing Ra1 to: 100k / 1.03 = 97kΩ. In other words, perfect AC balance is obtained when[6]:

$$\frac{Ra2}{Ra1} = 1 + \frac{Ra2 + ra}{Rk(\mu + 1)} \quad \text{.........LIX}$$

But in practice this modification is unnecessary for guitar purposes.

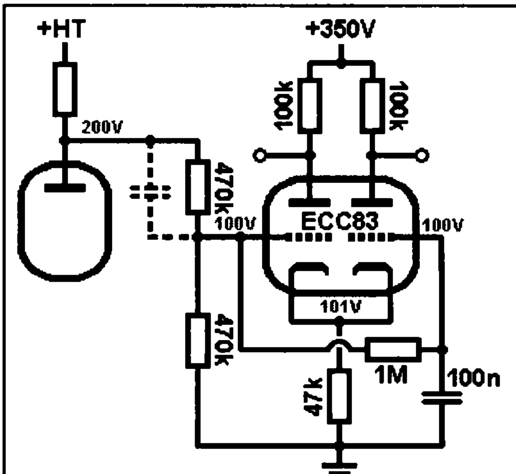

Fig. 8.8: A level-shifted, DC-coupled long-tailed pair.

Level shifting is also a convenient way to couple the long-tailed pair, particularly if the required grid voltage is rather low. Fig. 8.8 shows an example in

* Strictly we should replace Ra with the value of Ra||Rl, where Rl is the following loading resistance. In practice however, the error is negligible.

[6] Clare, J. D. (1947). The Twin Triode Phase-Splitting Amplifier. *Electronics Engineering,* (February), pp62-63.

which the previous stage has an anode voltage of 200V. This is level-shifted down to 100V, which is sufficiently close to the design value of 99V. If a treble boost capacitor is used (shown dashed), then an additional grid-stopper is advisable.

DC-coupled, long-tailed pairs are not often found in a guitar amp, and this is usually because it is preceded by a tone stack or volume control which requires AC coupling to keep DC off the potentiometers, which would otherwise cause loud scratching noises when turned.

AC-coupled long-tailed pairs:

Designing an AC-coupled long-tailed-pair follows essentially the same process as described earlier, except that the input is isolated by a coupling capacitor, so we must apply the necessary grid voltage by some other means.

Fixed bias:

The required grid voltage could be applied to the first (input) grid, but since the second grid is bypassed anyway it is more convenient to apply the bias here, as illustrated in fig. 8.9. Attentive readers may note its similarity to the fixed-biased cathodyne in fig. 7.5*b*.

R1 and R2 form a potential divider which provides 99V to the second grid, while Rg 'pulls up' the first grid to the same voltage. The bypass capacitor Cg2 must be made large, however, since the effective resistance supplying the second grid has been reduced to R1||R2||Rg, or:

$$R = \frac{1}{\frac{1}{R1} + \frac{1}{R2} + \frac{1}{Rg}} = \frac{1}{\frac{1}{560k} + \frac{1}{220k} + \frac{1}{1000k}} = 136k\Omega$$

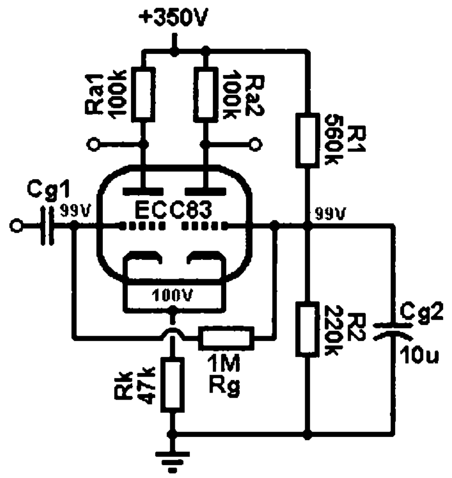

Fig. 8.9: A fixed-biased long-tailed pair.

Now, Cg2 not only decouples the grid at audio frequencies but must also filter out any HT noise. If we wish to filter out frequencies as low as 1Hz then:

$$C = \frac{1}{2\pi fR} = \frac{1}{2\pi \times 1 \times 136k}$$

$= 1.2\mu F$

So an arbitrarily large value of 10μF would ensure excellent filtering, and need only have a voltage rating somewhat greater than the grid voltage, say 160V, so will not be a very bulky component.

The input coupling capacitor may be chosen in the usual way, in combination with the input impedance which is equal to Rg.

This form of fixed-biased, long-tailed pair is rarely seen in audio amplifiers, but is included here for completeness. Its lack of popularity may be because the cathode-biased version (see below) offers a higher input impedance with the same number of components, and does not require a large (and therefore electrolytic) grid bypass capacitor.

Cathode bias:

This is the most common form of long-tailed pair. The bias is applied in the same way as for the cathode-biased cathode follower and cathodyne circuits. The necessary grid voltage is tapped off the tail resistance by splitting it into two parts and connecting the grids to the junction, via grid-leak resistors, as shown in fig. 8.10. We could go to the trouble of ensuring that Rk + Rb equal the total tail resistance that was initially specified, but in practice the bias resistor, Rb, is small enough that there is no need, it can simply be stacked on top of Rk.

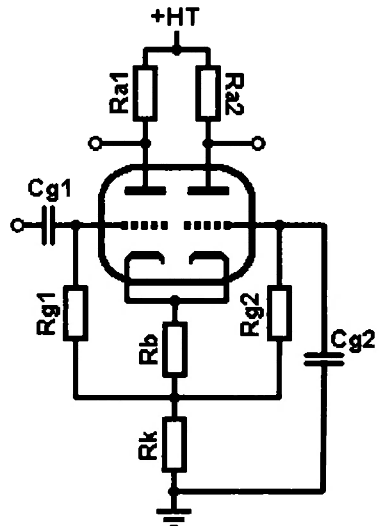

Fig. 8.10: The cathode-biased long-tailed pair is the most common phase-inverter arrangement found in guitar amplifiers.

The load line for the earlier design is reproduced in fig. 8.11 for convenience. Taking the same bias point of Vgk = –1V, the anode current is seen to be around 1.1mA. However, remembering that the current from both triodes flows in the tail resistance, this must be doubled, giving 2.2mA. The necessary bias resistor is then:

R = V/I
R = 1 / 0.0022
= 455Ω.
The nearest standard is 470Ω.

Remembering that half the input signal also appears at the cathode, most of this also appears at the junction of Rb and Rk (the bottom of the

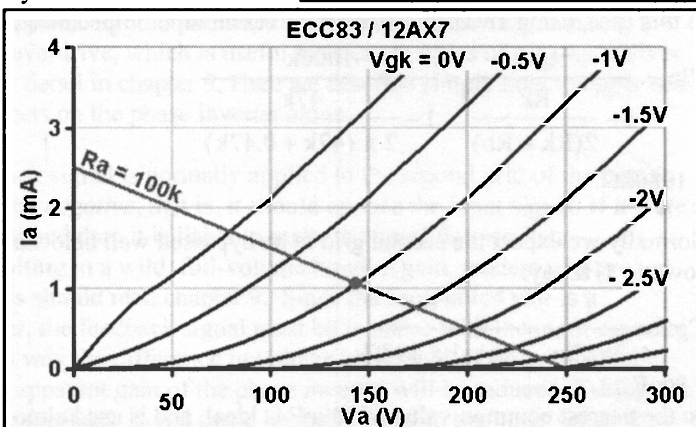

Fig. 8.11: A 100kΩ load line reproduced from fig. 8.5. Taking a bias point of -1V the anode current is about 1.1mA.

grid-leak resistors), so Rg1 and Rg2 are both bootstrapped.
Assuming good balance, and Rg1 = Rg2 = Rg, the input impedance to each grid becomes:

$$Zin = \frac{Rg}{1 - \dfrac{Rk}{2(Rk + Rb)}} \qquad \text{.......LX}$$

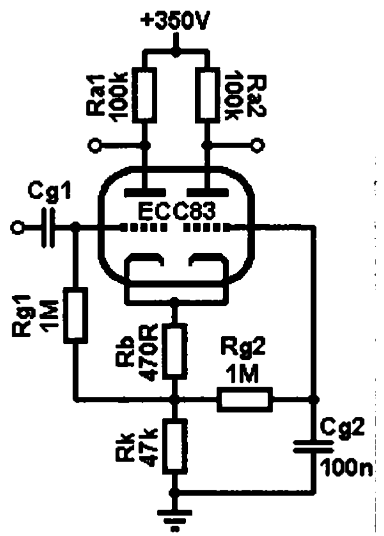

Fig. 8.12: A completed cathode biased, long-tailed pair phase-inverter.

But usually Rk/(Rk+Rb) ≈ 1, so this may be simplified to:
$Zin \approx 2Rg$

So using conventional 1MΩ grid-leaks gives an input impedance close to 2MΩ, or 470kΩ resistors could be used for less noise while retaining an input impedance of close to 940kΩ.

The input coupling capacitor, Cg1, may be chosen in the usual way, using the calculated input impedance. Often, however, it is made arbitrarily large, with frequency shaping being performed by an earlier interstage coupling network or tone stack.

The bypass capacitor for the second grid, Cg2, is chosen according to:

$$Cg2 = \frac{1}{2\pi fZin}$$

In this case, using 1MΩ grid stoppers gives an input impedance of:

$$Zin = \frac{Rg}{1 - \dfrac{Rk}{2(Rk + Rb)}} = \frac{1000k}{1 - \dfrac{47k}{2 \text{ x } (47k + 0.47k)}}$$

= 1980kΩ.

Normally we expect the second grid to be bypassed well below audible frequencies, down to 1Hz say:

$$Cg2 = \frac{1}{2\pi fZin} = \frac{1}{2\pi \text{ x } 1 \text{ x } 1980k}$$

= 80nF

So the nearest common value of 100nF is ideal, and is used almost universally. The completed circuit is shown in fig. 8.12. This is very similar to the phase inverter used

in the Vox *AC30*, except that it was arranged as a mixer / inverter for mixing the normal and tremolo channels, as shown in fig. 8.13.
Observant readers may comment that when feeding a signal into only *one* input, if the volume control of the opposite input is not turned down then the opposite grid will not be properly decoupled and the balance may suffer. In practice, however, the imbalance is not much more than +/-1dB, because the potential divider formed by the opposite grid-leak and corresponding volume control attenuates the AC signal actually appearing on opposite grid.

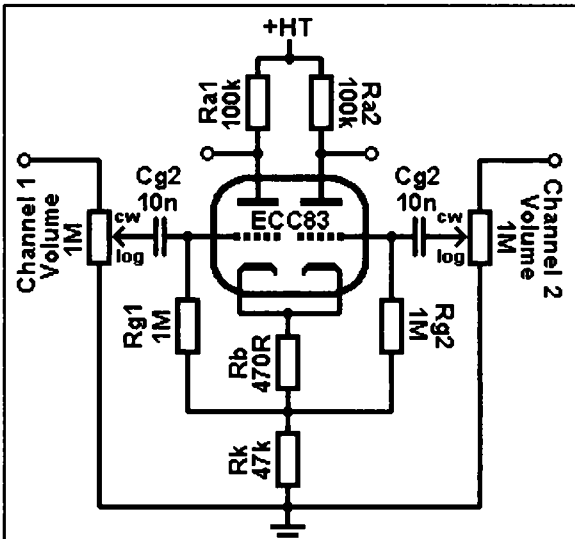

Fig. 8.13: The cathode-biased long-tailed pair as a mixer / inverter, similar to that used in the Vox *AC30* and Marshall *1974 18W*.

Global negative feedback and the long-tailed pair:

The previous sections have shown how to design the essential parts of a long-tailed pair. However, in guitar amplifiers it is so common also to find global **negative feedback** (NFB) applied to it that the subject deserves special attention. This feedback is usually taken from the secondary of the output transformer and serves to linearise the power output stage, increasing its bandwidth and 'flattening' the frequency response. It also reduces the effective input sensitivity of the stage to which feedback is applied. In other words, it increases the headroom, making the amplifier harder to overdrive, which is useful for cleaner styles of playing. This is dealt with in greater detail in chapter 9. Here we describe simply how to *apply* this feedback, and it effects on the phase inverter alone.

The feedback signal is normally applied to the second grid of the circuit. The feedback must be *negative*, that is, it should oppose the input signal. If it were to reinforce the input signal then it is liable to make the amplifier oscillate uncontrollably, resulting in a wild, full-volume howl! Again, readers who are not already aware of this should read chapter 9. Since the long-tailed pair is a differential amplifier, the feedback signal must be *in phase* with the input signal at the first grid. In this way the difference in voltage between the two grids will be reduced, and so the apparent gain of the phase inverter will be reduced. Taking this to an extreme, if the feedback signal could be made equal to the input signal then there would be no difference between the two grids at all and we would obtain no

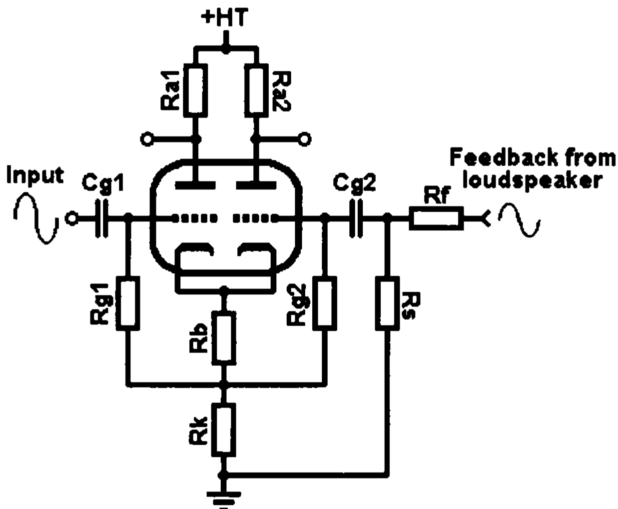

Fig. 8.14: Applying negative feedback to the long-tailed pair via a simple potential divider formed by a feedback resistor, Rf, and shunt resistor, Rs.

output signal! In reality this is impossible since no output would also mean no feedback signal, but it illustrates the phase relationship between the two grids.

The simplest arrangement is to take the feedback from the secondary of the output transformer / loudspeaker and apply it to the second grid via a potential divider Rf and Rs (fig. 8.14), which determines the amount of feedback being applied. Although this is a perfectly acceptable circuit, anyone familiar with existing guitar amp designs will immediately notice that this does not yet look like the 'usual' circuit arrangement.

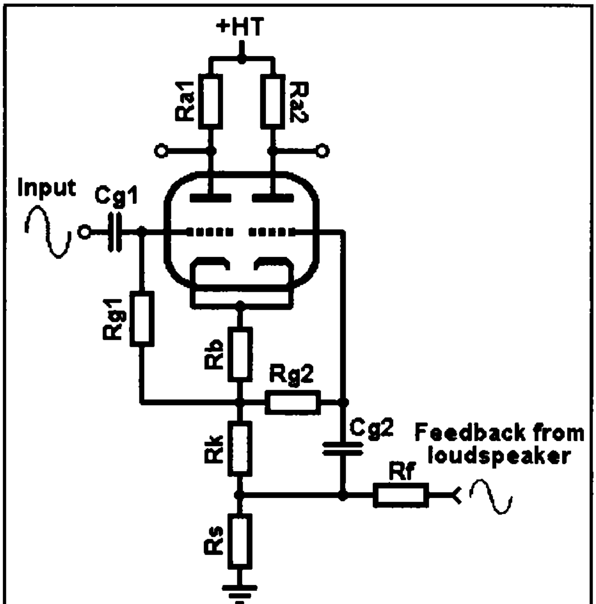

Fig. 8.15: Applying negative feedback to the second grid *and* tail, bootstraps the tail resistance, and is the most common form of long-tailed pair found in guitar amplifiers.

By moving Rs so that it forms part of the tail resistance, as shown in fig. 8.15, the circuit performance can be improved slightly. Since the input and feedback signals are assumed to be substantially similar except in amplitude, the signal appearing at the cathode will also have the same form. Since the feedback signal is also being applied to part of the tail resistance we now have substantially similar signals both at the top and at the bottom of the tail; Rb and Rk have been bootstrapped. The effective tail resistance is increased from Rb + Rk + Rs to a new value of:

$$Rt = (Rk + Rb)\frac{Vin}{Vin - Vfb} + Rs$$

or:

$$Rt = (Rk + Rb)\times 10^{dB/20} + Rs$$

Where:
dB = the amount of feedback in decibels.

In theory, increasing the effective tail resistance helps improve the balance of the phase inverter. The feedback signal delivered to the tail resistor can also help in *maintaining* balance when the anodes are unequally loaded (e.g, when the power valves are being overdriven)[7], but the circuits used in guitar amps are never actually optimised for this. In fact, this form of the phase inverter often gives worse balance than that in fig. 8.14!
Nevertheless, this has become the most common form of long-tailed pair used in guitar amplifiers, and was first adopted by Fender in 1958 for the new versions of the *Bassman* and *Twin Amp*, and was also popularised in classic Marshall amplifiers; it has become something of an 'industry standard' despite its basic design flaws. Of course, these are 'flaws' only in the conventional 'hifi' sense. For musical instruments, expectations are relaxed, and the circuit still gives good tonal results.

In practice, around 4dB to 10dB of global negative feedback is used, which is to say that the gain of the phase inverter is reduced by 4dB to 10dB, or 0.6 to 0.3 times its full level before feedback was added. The feedback signal should therefore be between $1 - 0.6 = 0.4$ to $1 - 0.3 = 0.7$ times the input signal. This must be set by the potential divider formed by Rf and Rs, which means we must also know the amplitude of the signal appearing at the loudspeaker, for a given input signal to the long-tailed pair. This could be measured directly of course, or it may be calculated.

A worked example; the Fender *5F6-A Twin Amp*:

In the Fender *5F8-A Twin Amp*, only 27V was allowed for the tail resistor. The HT was 355V, giving an effective HT of 328V, and 100kΩ loads were chosen for use with an ECC83. Fig. 8.16 shows the load lines. A 470Ω bias resistor was used, but since twice the current flows in this resistor, a cathode load line of $2\times470 = 940\Omega$ is drawn. The expected bias point shows a quiescent current of 1.35mA. Since twice this, or 2.7mA, flows in the tail resistor, a value of $27 / 0.0027 = 10k\Omega$ was used. This is relatively small, so the imbalance of the circuit is more significant. The g_m at the bias point is 1.6mA/V and r_a is around 62.5kΩ, so the difference in gain between the two triodes is:

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[7] Birt, D. R, (1960). Self-Balancing Push-Pull Circuits. *Wireless World,* (May), pp223–227. Also: Balanced Output Amplifiers of Highly Stable and Accurate Balance. *Electronic Engineering*. June 1946, p189.

$$\frac{A1}{A2} = 1 + \frac{Ra + ra}{Rk(\mu + 1)} = 1 + \frac{100k + 62.5k}{10k \times (100 + 1)}$$

= 1.16.

In other words, the gain of the first triode is 16% greater than that of the second. This could be corrected by reducing Ra1 to: 100k / 1.16 = 86.2kΩ. The nearest standard is 82kΩ, which is what Fender used. The grid-leak resistors are set at a conventional value of 1MΩ and Cg2 is made the usual, large value of 100nF.

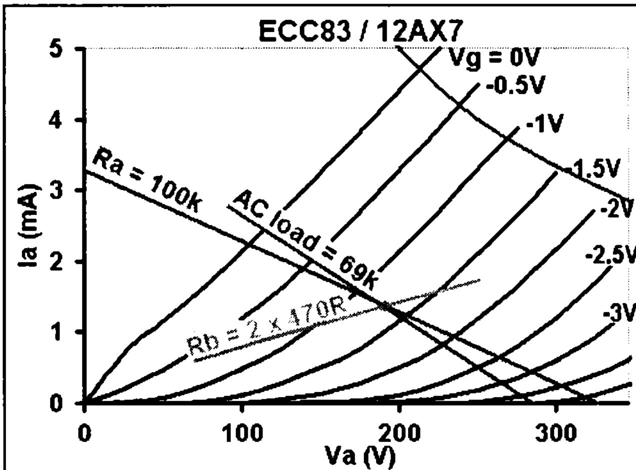

Fig. 8.16: Anode and cathode load lines drawn for an effective HT of 328V.

All that remains is to apply the negative feedback, but this requires knowing the gain of the amplifier from the phase inverter to speaker. This could be measured directly or, if we know (or closely estimate) the maximum output-power before clipping, it may be calculated. The Fender *5F8-A Twin Amp* used four 5881 power valves and quoted an output power of 80W. The 5881s were biased at around –50V, so require 50 / √2 = 35Vrms of drive signal for full output. The gain to *one output* of the phase inverter is half that indicated by the AC load line, or about 26. For maximum, unclipped output power, the input signal to the phase inverter must therefore be: 35 / 26 = 1.35Vrms .

If this develops 80W into a 2Ω speaker (the original amp used two 4Ω speakers in parallel), then the voltage at the speaker must be:

$P = V^2/R$

$V = \sqrt{PR} = \sqrt{80 \times 2}$

= 12.6Vrms.

We now have sufficient information to choose the feedback components. Fender applied 12.5dB of feedback. This is a relatively large amount, but this is unsurprising since it was designed for pedal-steel players who required a very clean tone.

The ratio of input signal to feedback signal is given by:

$$\frac{v_{in}}{v_{fb}} = 1 - \frac{1}{10^{dB/20}}$$

Where:
v_{in} = the input signal amplitude.
v_{fb} = the feedback signal amplitude.
dB = the amount of feedback in decibels.

So for 12.5dB of feedback:

$$\frac{v_{in}}{v_{fb}} = 1 - \frac{1}{10^{12.5/20}}$$

= 0.76

In other words, for every 1V of input signal, 0.76V of feedback signal is returned to the second grid.
It has already been calculated that for full power, 1.35Vrms of input signal is required. The feedback signal at this level must therefore be: 1.35 x 0.76 = 1.03V. This is applied from the speaker via a potential divider formed by the feedback resistor and shunt resistor, Rf and Rs respectively. Since the voltage at the speaker is 12.6V and the required feedback signal is 1.03V, the gain of this divider, β, must be:
β = 1.03 / 12.6 = 0.082

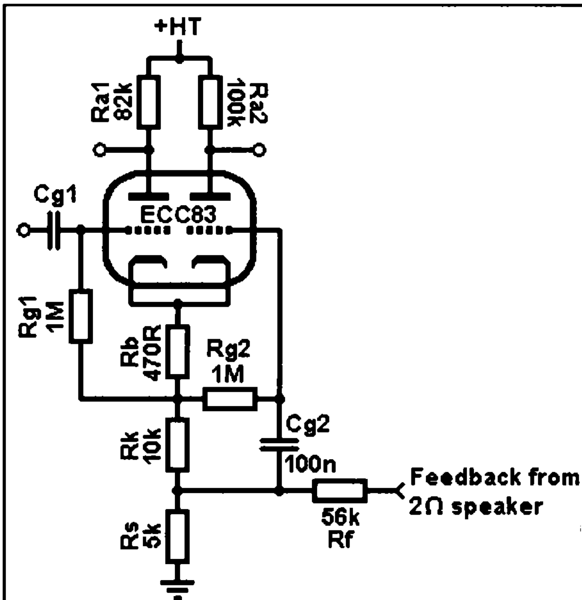

Fig. 8.17: The basic, long-tailed pair circuit, first used in the Fender *5F8-A Twin Amp.*

The absolute value of these resistors is not important, except that they must be large with respect to the speaker impedance. Rs is usually made relatively small with respect to Rk, so that it does not significantly alter the previously calculated tail resistance. However, Fender selected a rather large value of 5kΩ for Rs. The value of Rf must therefore be:

$$Rf = \frac{Rs - Rs.\beta}{\beta} = \frac{5k - (5k \times 0.082)}{0.082}$$

= 56kΩ.

Note that the value of the feedback resistor depends on the power output and speaker tap from which feedback is taken; we cannot simply 'shoehorn' this circuit into a different amplifier without altering Rf or Rs to suit the new conditions. If in doubt, do not use any feedback at all, it is not essential!
The completed circuit is shown in fig. 8.17. The rather high values of Rf and Rs were heavily reduced in most later designs, which also reduces noise and improves the stability of the feedback loop since it lessens the influence of stray capacitances.

The most significant change in the operation of the phase inverter is a decrease in input sensitivity due to the addition of feedback, because the feedback signal effectively 'cancels out' part of the input signal. By adding 12.5dB of feedback, the input sensitivity is decreased by the same degree, or 4.2 times. The load line in fig. 8.16 indicated an initial input sensitivity of around 3.5Vp-p, but remembering that for a long-tailed pair the headroom is twice that indicated, this gives 7Vp-p. But the addition of feedback has reduced this to: 7 x 4.2 = 29.4Vp-p, so the phase inverter is now much harder to overdrive.
This additional headroom is one of the main reasons for using negative feedback, but of course, placing a potential divider before the phase inverter (without feedback) can produce the same amount of effective headroom without any problems of instability. Since good fidelity is rarely a requirement for modern musicians, global negative feedback in guitar amplifiers is less useful today than it was in the 1950s.

Another interesting point to note about this original circuit is the use of the 82kΩ and 100kΩ anode resistors. Because a relatively low tail resistance was used, this mis-match did serve to improve the balance of the circuit before feedback was added. However, if the tail resistance is increased to just over 16kΩ the difference in gain between the two triodes becomes small enough that better balance is actually obtained with identical 100kΩ anode resistors. But almost all versions of this circuit produced today are descended from this original design, so that the 82kΩ:100kΩ combination is still seen in designs even when much larger tail resistances are used, betraying the fact that most modern manufactures only copy parts of existing designs, rather than design from scratch! This historical curiosity appears to have begun with the CBS-Fender designs, and has perpetuated ever since.

The presence control:
Keen readers will have noticed that the circuit in fig. 8.17 is missing the ubiquitous presence control. This simple control effectively removes upper middle and treble frequencies from the feedback signal by shunting them to ground via a capacitor. By removing the feedback at these frequencies the gain of the phase inverter returns to its normal level, giving an active boost in the upper middle and treble frequencies. The earliest arrangement was simply to replace Rs with a potentiometer, with the shunt capacitor connected to the wiper, as shown in fig. 8.18*a*. However, since DC current flows in this resistor, when operated the control *always* develops a loud scratching sound over time, so the modified version in fig. 8.18*b* should be used instead, where the capacitor C1 blocks DC from reaching the potentiometer.

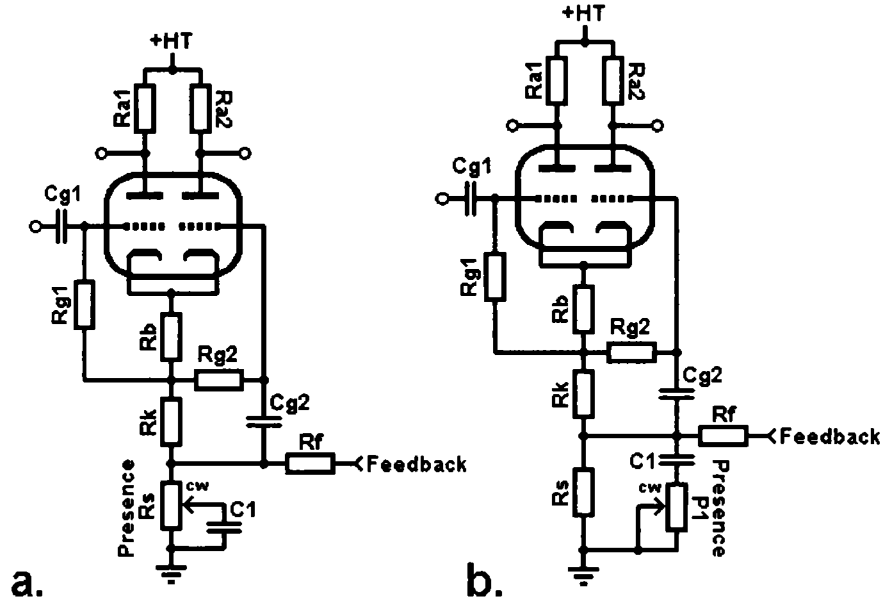

Fig. 8.18: Long-tailed pairs with 'presence' control. The old-fasioned arrangement in **a.** produces an obnoxious scratching sound due to DC on the potentiometer, so the circuit in **b.** is preferred for new designs.

The potentiometer, P3, is normally made larger than Rs so that it does not substantially alter the calculated amount of feedback (this is not very critical however, since this resistance will only serve to reduce feedback slightly, which is much safer than increasing it).
For example, if Rs is 100Ω then a convenient value for P1 might be 4.7kΩ. If the presence control is turned fully up (zero resistance) then the +3dB transition frequency between nominally flat response and the boosted level is given by*:

$$f = \frac{1}{2\pi C1(Rf \parallel Rs)} = \frac{1}{2\pi C1\left(\frac{Rf.Rs}{Rf + Rs}\right)}$$

In the case of the Fender 5F8-A example, Rf = 56kΩ, Rs = 5kΩ and a value of 100nF was used for C1, giving a boost frequency of:

$$f = \frac{1}{2\pi \times 100\text{x}10^{-9} \times \left(\frac{56k \times 5k}{56k + 5k}\right)} = 347\text{Hz}$$

This covers almost all the middle range frequencies and above. Other manufacturers design for frequencies in the range of 350Hz to 600Hz. Interestingly, some guitarists

* This formula breaks down if less than 6dB of feedback is used, although it will still be sufficiently accurate for most practical purposes.

find the presence control invaluable, and go to great lengths to add one to amplifiers which don't already have one, while other players find the presence control to be of no practical use at all. This may be due to some amplifiers using so little feedback that the effective boost has little audible effect, of course.

Variable feedback:

The effects of negative feedback are discussed at length in chapter 9. Because these effects are not to everyone's tastes, a useful option is to make the feedback variable between its full level and a lesser amount (or zero). For users who find the presence control of limited use, it may be converted into a feedback control by increasing the value of the presence capacitor, C1 (fig. 8.18), so that it shunts virtually all the feedback signal to ground. This is shown in fig. 8.19*a*. Alternatively, variable feedback may be implemented by connecting a potentiometer (wired as a variable resistor) in series with the feedback resistor, as shown in fig. 8.19*b*, effectively increasing the feedback resistance. However, because of the relatively small amount of feedback used in guitar amplifiers, often there is little audible difference in the tone over much of the control's range. For many users, the ability to switch between full feedback and (virtually) no feedback is quite sufficient, and this is shown in fig. 9.19*c*. Note that R1 is included so that the feedback loop is never actually broken, which reduces popping sounds. Its value is made sufficiently large as to reduce the feedback to almost nothing.

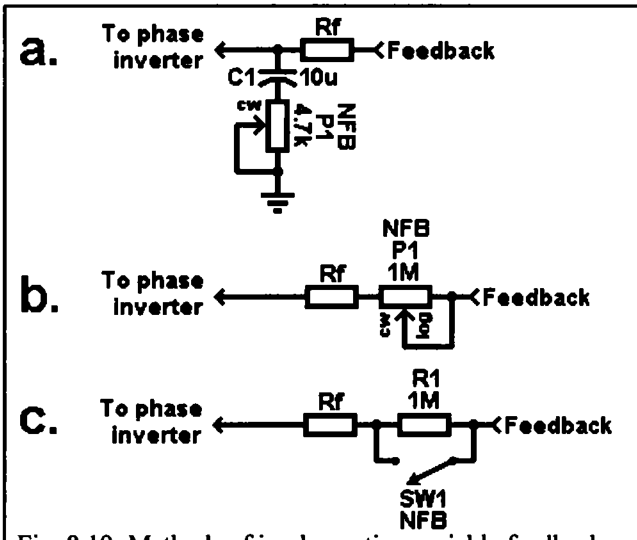

Fig. 8.19: Methods of implementing variable feedback. **a.**: modifying an existing presence control by increasing C1. **b.**: adding a variable resistance in series with Rf. **c.**: switching between two levels of feedback.

Presence control without global negative feedback:

Global negative feedback may not suit all playing styles, and can also raise problems of instability (see chapter 9).
For builders who are only interested in adding a presence control, the use of feedback can seem like an unfortunate necessity. Thankfully, the long-tailed pair phase inverter allows the traditional, feedback presence control to be 'faked'. By manipulating the size of the second grid's bypass capacitor we can alter the gain of the circuit to some extent, enough that it can be turned into a presence control, as shown in fig. 8.20.
Here we see that the grid-bypass capacitor, Cg2, is made quite small so that only middle and treble frequencies are bypassed, so these frequencies achieve full gain. A larger capacitor, C1, is connected in parallel so that all frequencies can be bypassed, while the potentiometer in series with this allows the degree of bypass at low frequencies to be controlled. Really this is a bass-cut control, but the tonal effect is the same as a conventional presence control.
The frequency response is given in fig. 8.21.
In this case the tail resistance is greater than 16kΩ so identical 100kΩ anode resistors have been used, although some players may prefer the tonal effect of *greater* imbalance caused by using the 'historic' values of 82kΩ and 100kΩ. In any case, the tail resistance and anode load resistances are not specific to the presence control. The maximum degree of bass cut is determined by the potential divider formed by Rg2 and P1. Rg2 is bootstrapped to an effective value of:

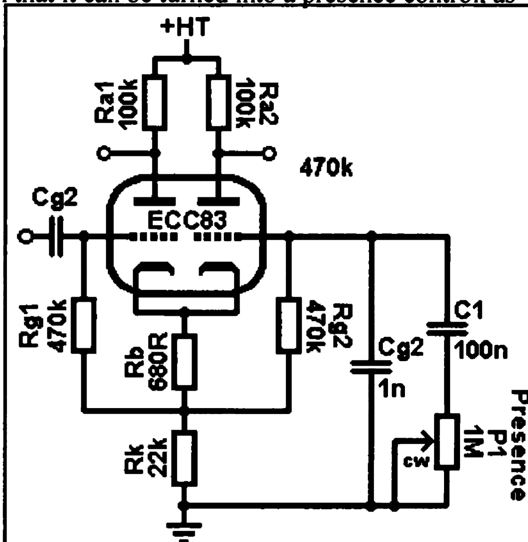

Fig. 8.20: A presence control for an amplifier without global negative feedback.

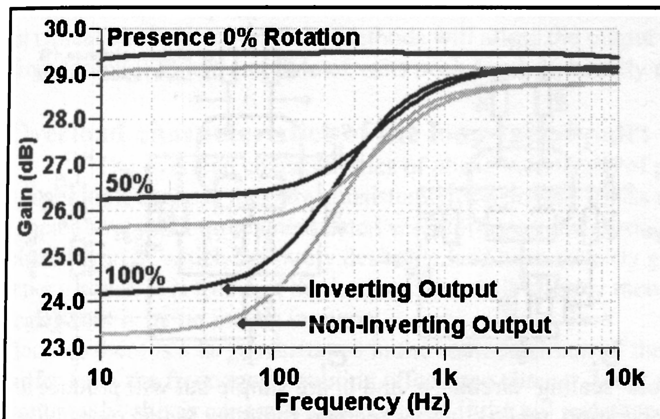

Fig. 8.21: Frequency response of the circuit in fig. 8.20.

$$Zin = \frac{Rg}{1 - \frac{Rk}{2(Rk + Rb)}} = \frac{470k}{1 - \frac{22k}{2 \times (22k + 0.68k)}}$$

= 913kΩ.

Since P1 is made 1MΩ, to a close approximation the maximum degree of bass cut is simply:

$$\beta \approx 1 - \frac{P1}{P1 + Zin} = 1 - \frac{1000k}{1000k + 913k}$$

= 0.47 or -6.4dB.

This shows that using smaller grid-leak resistors increases the maximum degree of cut. We may also note that the degree of cut does not depend upon the type of valves used, but only on the tail and grid-leak resistances.
This control could easily be applied to any existing design where only one input of a long-tailed pair is used, and no global negative feedback is applied.

'Scale' control:

It is not often pointed out that the gain and output-signal swing of the long-tailed pair can be controlled quite successfully by varying its bias. It might be thought that the tail resistance would have more effect on the latter, but in practice this does not seem to be the case; varying Rk has very little effect on the characteristics and tone of the circuit, provided it is not smaller than about 10kΩ. By varying Rb, however, operation can be controlled from its normal mode of moderate gain and high output signal swing, to less gain and, crucially, much less output signal swing. This can serve as a pseudo-power-scaling effect. By increasing the bias the output from the phase inverter can be reduced to a few volts, while the phase inverter itself can then be overdriven more heavily, effectively reducing the volume of the amp while retaining broadly the same amount of perceived overdrive

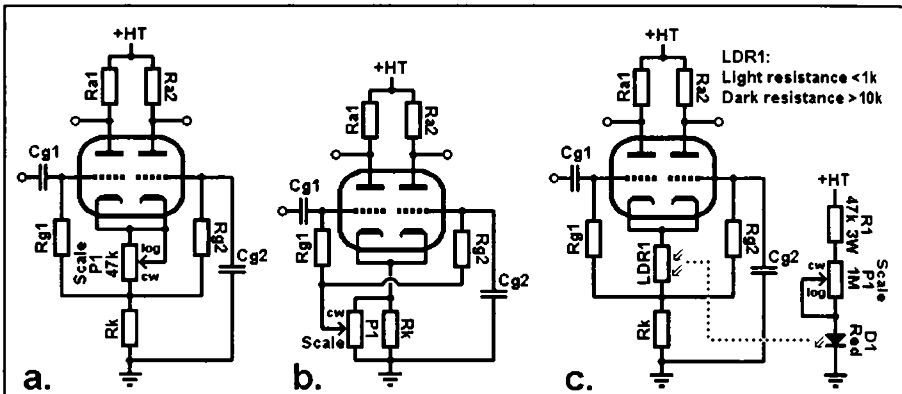

Fig. 8.22: Three possible 'scaling' circuits. **a.** and **b.** are simple but will produce a scratchy sound when operating, while **c.** varies the bias remotely via an opto-isolator.

levels. Three possible circuits for achieving this are shown in fig. 8.22.

The circuit in fig. 8.21*a* is perhaps the most obvious and easy to install in an existing circuit, in which the bias resistor is replaced with a large variable resistance in the region of 10kΩ to 47kΩ. This allows the bias to be varied from zero (high gain and output) to a high value (low gain and output). The rest of the phase inverter is designed in the usual way.

The circuit in *b.* removes the bias resistance altogether and instead places a high-value potentiometer in parallel with Rk (the combined resistance of the two should be the same as the usual tail resistance of course). The grid bias voltage is now literally tapped off the tail resistance. This method has the advantage that the total tail resistance no longer varies with the bias setting, which is preferable if global negative feedback is also applied. This is the same circuit as the 'limiting' control used in the Carlsbro *60TC*.
A high value potentiometer (100kΩ to 470kΩ say) should be used so that most of the tail current flows only in Rk, with little power being dissipated in P1.

Both the previous circuits have DC on a potentiometer which will cause scratching sounds when operated, although many users may be willing to suffer this, considering how useful this control is. For a new design though, it may be preferable to vary the bias remotely. One such example is shown in fig. 8.21*c*. This is essentially the same as the circuit in *a.* except that the bias resistance is replaced with a light-dependant resistor (LDR) whose resistance is varied by an LED. This greatly 'smoothes out' the DC changes in the circuit as P1 is rotated, reducing spurious noises. R1 determines the maximum current –and therefore brightness– for the LED, and this can be reduced to almost nothing with P1. Most LEDs will be sufficiently bright at around 5mA to 8mA current. As shown, the LED is powered from the HT, but any convenient DC supply will do of course, if R1 is suitably chosen. The LED and LDR should be mounted facing one another and sealed from external light with heatshrink tubing. Alternatively, LED/LDR combinations can be bought in a single package called an opto-isolator.
In typical circuits, any of these methods will allow the output swing to be varied from 'normal' to just a couple of volts peak-to-peak, heavily overdriven.

Overload characteristics of the long-tailed pair:

The overload characteristics of phase inverters is of particular importance in guitar amps, because they are consistently overdriven. It was shown in the previous chapter how some quite unexpected waveforms can be produced by a cathodyne phase inverter, which are rarely desirable, and have unfairly given it a bad reputation among some. It is true that the long-tailed pair is slightly more predictable in this regard, but is by no means immune.
Because there is a large resistance in the cathode circuit of the long-tailed pair, it can suffer from the frequency-doubling effect [see chapter 7, fig. 7.12] and swirl. Fortunately, this is generally only an issue if the tail resistance is very large, more

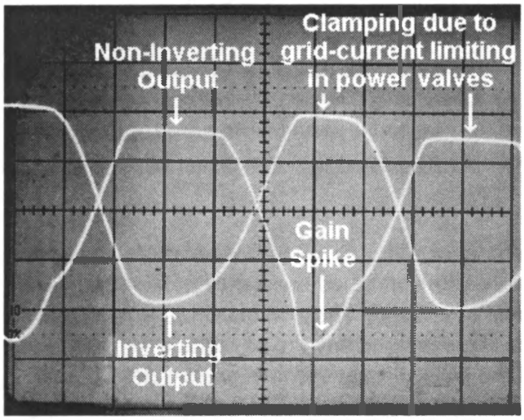

Fig. 8.23: Oscillogram showing the 'gain spike' effect when the power valves are overdriven. 1kHz, 5V/div

than 100kΩ say, or if it is replaced with a constant-current sink, which is rarely the case in a guitar amp. However, it may surprise some readers to learn that the long-tailed pair *does* suffer from the gain-spike effect first described in chapter 7 [fig. 7.11], though for different reasons.

This spike occurs at the *non-inverting* output when *the inverting* output voltage is clamped, that is, when the power valves are overdriven. When the inverting-output signal is large enough, it will begin to overdrive the corresponding power valve. As the power valve begins to draw grid current it will clamp the anode voltage of the inverting anode, so that it cannot rise any further even though the input signal to the long-tailed pair is still moving negatively. Since the anode voltage is no longer changing it appears grounded as far as AC is concerned or, in other words, the input triode now acts as a pure cathode follower. Normally only half of the input signal appears at the cathode and is passed to the second triode, but now that the first triode is acting as a proper cathode follower with near-unity gain, almost all of the input signal appears at the cathode. This is passed to the second triode and amplified, causing its output signal to suddenly increase, creating the gain spike. This is seen in the oscillogram in fig. 8.23 which was produced form the circuit in fig. 8.12 when overdriving a pair of EL84s without grid stoppers.

If the power output stage is class AB, as most push-pull guitar amps are, then this negative 'gain spike' is not always a concern because the corresponding power valve is already in cut-off when the negative spike occurs. However, if the power valves are poorly matched, or if the power output stage is class A, then this spike may be amplified by the corresponding power valve and will introduce high-order intermodulation distortion products, and is rarely pleasant.

As with the cathodyne, this effect may be subdued or even eliminated by using larger grid stoppers on the power valves. Since the power valves are nearly always pentodes or beam tetrodes, which have very little input capacitance, large grid-stoppers may be used without affecting the audible treble response, and values of around 100kΩ will virtually eliminate the gain-spike effect in most cases, as used on the Marshall *Studio 15*. Larger grid-stoppers will greatly reduce the chances of blocking distortion too, and can also be used to assist in stabilising any global negative feedback [see chapter 9, fig. 9.11]. Even if the amp is class AB, larger-than-usual grid stoppers of around 10kΩ are well recommended and will help maintain consistency if the power valves do not age equally.

Summary of formulae:

LV; Gain to non-inverting output assuming a symmetrical circuit where Ra1 = Ra2 = Ra:

$$A2 = \frac{\mu Ra}{(Ra + r_a).\left(2 + \frac{Ra + r_a}{Rk(\mu + 1)}\right)}$$

LVI; Difference in gain between outputs:

$$\frac{A1}{A2} = 1 + \frac{Ra + r_a}{Rk(\mu + 1)}$$

Differential gain assuming perfect balance:

$$A = \frac{\mu Ra}{Ra + r_a}$$

LVII; Total input capacitance:
$Cin \approx Cgk\,(1 - A_k) + Cga.A_a$

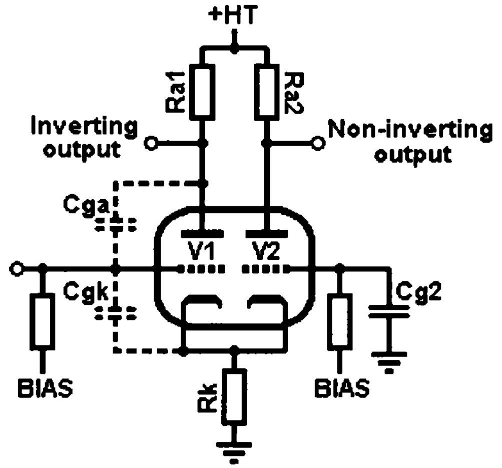

Fig. 8.24

Where:
A_k = the gain from grid to cathode, being approximately 0.5.

A_a = gain to the anode, which will be approximately: $Aa \approx \dfrac{\mu Ra}{2(Ra + r_a)}$

But since $Cgk\,(1 - A_k)$ is very small, this may be approximated as:
$Cin \approx Cga.A_a$

LVIII; Output impedance assuming *unequal* loading:

$$Zout \approx Ra \parallel 2r_a + Ra = \frac{Ra}{2}$$

Output impedance for equal loading:

$$Zout = Ra \parallel r_a = \frac{Ra.r_a}{Ra + r_a}$$

LIX; Perfect AC balance is obtained when:

$$\frac{Ra2}{Ra1} = 1 + \frac{Ra2 + ra}{Rk(\mu + 1)}$$

LX; Input impedance of the circuit in fig. 8.25 where Rg1 = Rg2 = Rg:

$$Zin = \frac{Rg}{1 - \frac{Rk}{2(Rk + Rb)}}$$

But usually Rk/(Rk+Rb) ≈ 1, so this may be simplified to:

$$Zin \approx 2Rg$$

Where in all cases notations are as in figs. 8.24 and 8.25.

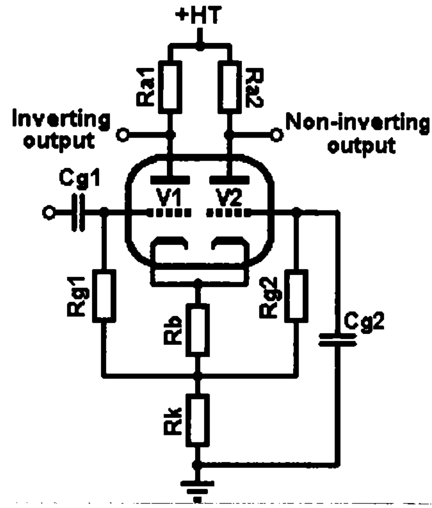

Fig. 8.25

Chapter 9: Feedback Theory

The universal feedback equation. Practical, local-feedback amplifiers. Effect on linearity, noise and frequency response. Effect on input and output impedance. Practical, virtual-earth amplifiers. Effects on overdrive. Stability and limitations on feedback. Global feedback in practice. Summary of formulae.

Feedback, as the name suggests, involves the feeding of a portion of the output signal back to the input of an amplifier. It may be intentional, or it may occur by accident via stray capacitances and inductances when the amplifier is physically built. Unfortunately, feedback is an advanced topic in electronics, and involves somewhat more theory than most of the chapters in this book. The first part of this chapter deals with the simple case of feedback around a single valve, while the latter half explores feedback when applied around more stages.

The universal feedback equation:

If the voltage signal fed back is proportional to the output current, it is known as **current feedback**; if it is proportional to the output voltage then it is **voltage feedback**. As far as we are concerned, however, these terms are largely academic since both types are used to achieve the same thing: to manipulate the input signal voltage.

If the voltage fed back is in phase with the input voltage it will serve to increase it; this is known as **positive feedback** (older texts may use the term **regeneration**). If the feedback signal is large enough it will increase the input voltage enough to increase the feedback signal, which will increase the input voltage, which will increase the signal fed back and so on. This will cause the amplifier to self oscillate and is generally undesirable!

If the voltage fed back is 180° out of phase with the input voltage it will serve to decrease it; this is known as **negative feedback** and provides some useful results (older texts may use the term **degeneration**).

Fig. 9.1 shows a block diagram of an *inverting* amplifier with feedback applied to the input via the feedback resistor, Rf. The amplifier section could consist of a single gain stage, or several, or it may be an entire amplifier circuit. If it is just one stage then the feedback is known as **local feedback**. If there is more than one gain stage within the loop then it is termed **global feedback**.

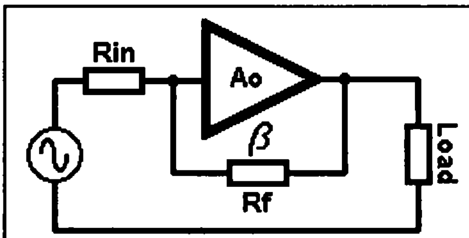

Fig. 9.1: Block diagram of a simplified (voltage) feedback amplifier.

Because the feedback path forms a 'loop' around the amplifier, we can describe the amplifier either as **open loop** with no feedback, or **closed loop** with feedback applied. If Ao is the open-loop gain (the gain before feedback is added) then a 1V input will produce –Ao volts at the output.

A portion of the output signal is now fed back to the input. If β is the fraction of output signal fed back, then for every volt applied to the input –Ao volts appear at the output and $-Ao\beta$ volts are fed back. The closed-loop amplifier therefore requires not a 1V input, but a $1+Ao\beta$ volts input to produce the same output as before feedback was added. From this we obtain the **universal feedback equation** for closed-loop gain (the gain after feedback is added):

$$Acl = \frac{Ao}{1 + Ao\beta} \qquad \text{LXI}$$

Where:
Acl = closed-loop gain
Ao = open-loop gain
β = the feedback fraction

The function $Ao\beta$ is known as the **loop gain**. It is the gain of the amplifier and feedback loop combined, that is, the gain that would be seen if the loop were broken at some point and the gain of the circuit measured between the two broken ends. Normally we want the loop gain to be a negative value so the feedback is negative.

If the open-loop gain, Ao, or **feedback fraction**, β, is very large or infinite, then $Ao\beta$ will be a very large positive number so that $Ao\beta \approx 1+A_o\beta$ and the closed-loop gain equation may be simplified to:
$Acl = 1 / \beta$

In such cases Acl is set solely by the feedback components and Ao no longer affects performance. Thus there will be no change in the amplifiers performance even as the active components age or are replaced with one of different characteristics. This rule usually applies to solid-state amplifiers where extremely high open-loop gains are easily possible, whereas there are very few instances in valve amps where this is true, unless we cascade several gain stages, or use very high feedback fractions. The cathode follower and cathodyne phase inverter are rare examples of this.

A final point to note is that in cases where Ao and/or β is large, the input of the stage will behave as a **virtual earth**. For example, suppose we have an amplifier with a large open-loop gain of 1000 and we apply 50% feedback, reducing the closed-loop gain to:

$$Acl = \frac{Ao}{1 + Ao\beta} = \frac{1000}{1 + (1000 \times 0.5)} = 2$$

If our source signal is 1V then we will receive a 2V signal at the output. But we know that the true gain of the amplifier inside the loop is 1000, so the signal present at its input must be 2/1000 = 2mV. This is so small that it might be regarded as zero; the input appears to be close to zero volts, or grounded, from which we get the name 'virtual earth'. Whatever our source voltage is, virtually all of it appears across the input resistor and input current flows into the virtual earth. We could therefore have several input resistors feeding in different signals without the worry of them interacting with one another, and this is known as a **virtual-earth amplifier** or **virtual-earth mixer**.

Practical, local-feedback amplifiers:

Let us return to a practical example using valves. Suppose an ECC83 with a typical load of 100kΩ, as shown in fig. 9.2, has an open loop gain of –60 (though we normally omit the negative sign for simplicity). Anode and cathode resistors are chosen in the usual way. Negative feedback is applied from anode to grid via coupling capacitor C1 (to block DC) and feedback resistor Rf. This forms a potential divider with the input series resistor, Rin. Since the feedback components also load the valve they must be relatively large. Rf will usually made at least three times larger than the anode resistor, preferably more, to avoid dragging down the open-loop gain. However, excessively large values, greater than 1MΩ say, will introduce too much noise and may degrade performance. The cathode is fully bypassed and C1 is large enough to pass all intended frequencies: this is discussed later.

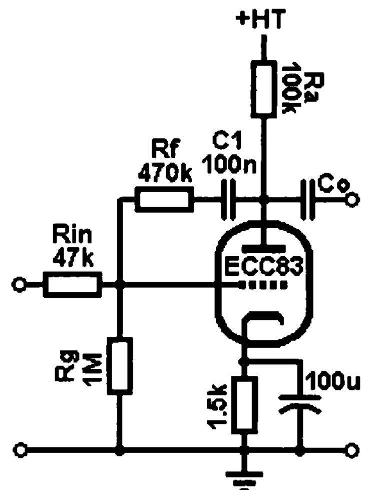

Fig 9.2: A practical circuit of a triode with local negative feedback.

Firstly we note the ratio of Rf to Rin, which we will term B (not to be confused with β)

B = Rf / Rin

B = 470k / 47k

= 10

It can be shown that the gain of a circuit such as that shown in fig. 9.2 is[1]:

$$Acl = \frac{1}{\frac{Rin}{Rf} + \frac{1}{Ao}\left(\frac{Rin}{Rg} + \frac{Ri}{Rf} + 1\right)}$$

This is a modification of the universal feedback equation, applied to the circuit arrangement in fig. 9.2, to account for the shunting effect of the grid-leak, Rg. By rearranging and substituting B we obtain:

$$Acl = \frac{Ao.B}{\frac{Rf}{Rg} + Ao + B + 1} \qquad \text{LXII}$$

[1] Boegli, C. P., (1960). The Anode Follower. *Audio*, (December). p19-22.

In this case:

$$Acl = \frac{60 \times 10}{\frac{470k}{100k} + 60 + 10 + 1}$$

= 8.4 or 18.5dB

Although in reality it may be somewhat less than this since the source impedance of the preceding stage is unlikely to be zero, in which case the source impedance should be added to Rin when performing calculations, if greater accuracy is required.

So the stage requires an input signal that is $1+A_o\beta$, or 60 / 8.4 = 7.1 times larger to produce the same output as before feedback was applied, so the input sensitivity has been reduced, or in other words: *headroom appears to have increased.* The overall impression is of a valve with completely different characteristics from the one actually being used, although we are really only altering the input signal of course.

Additionally, it is usual to express the amount of feedback ($1+A_o\beta$), known as the **feedback factor,** in terms of decibels gain reduction, in this case $20\text{x} \log(60/8.4) = 17\text{dB}$. However, a reduction in gain is not the only effect we observe.

Effect on linearity, noise and frequency response:

Any distortions or noise *produced* by the stage are fed back to the input where they are attenuated, and since the input signal did not contain them to begin with, they will to some extent cancel themselves out. The overall effect is to reduce total harmonic distortion and noise by the same amount as the feedback factor (17dB in the previous example), increasing the linearity of the amplifier.

Frequency distortion is also reduced by the same factor. In this case the Miller effect causes frequencies at 34kHz to be attenuated by –3dB before feedback is added. The open-loop gain at this frequency has fallen by a factor of:

$\log^{-1}(-3 / 20) = 0.71$, indicating an open-loop gain of 60 x 0.71 = 42.

Thus the closed-loop gain at this frequency becomes:

$$Acl = \frac{42 \times 10}{\frac{470k}{1000k} + 10 + 1}$$

= 7.9 or 18dB

which is only 6% less than our earlier calculation for closed-loop gain, so frequencies at 34kHz will be down by roughly -0.5dB and not –3dB, after feedback.

This illustrates how frequencies that are lacking from the output will also be lacking in the feedback signal, and therefore their overall attenuation is reduced. Likewise, frequencies that are accentuated in the output signal will cause greater feedback and will therefore be attenuated more. In this way the bandwidth of the amplifier is increased by the feedback factor, and the frequency response is made 'flatter'. For every 6dB of feedback added, the bandwidth extends by one octave above its original upper limit, and one octave below its original lower limit (assuming the open-loop gain falls at a first-order rate, which it should do for stability reasons). However, this assumes that the feedback loop itself does not contain any frequency dependant impedances, and in fig. 9.2 we find C1 is part of the loop. At low frequencies the reactance of C1 rises, reducing the feedback fraction. The gain will therefore begin to rise again, back towards its open-loop level.

The exact relationship between C1 and the feedback components is actually quite complex, but to a reasonable approximation: when the reactance of the C1 is equal to Rf then the feedback fraction will be lower by roughly -3dB, so we can expect the closed-loop gain to have increased by 3dB at a frequency of:

$$f_{+3dB} = \frac{1}{2\pi C1Rf} = \frac{1}{2\pi \times 100\times10^{-9} \times 470k}$$

$= 3.4Hz$

Fig. 9.3 shows the actual frequency response of the circuit. The open-loop response begins to decrease at high frequencies due to input capacitance, and falls off asymptotically with a first-order slope (-6dB per octave). At very low frequencies the open-loop response again begins to fall due to inadequate cathode bypassing, though this could be avoided with LED biasing.

The closed-loop response remains flat to much higher frequencies, extended by the feedback factor as described previously. At very low frequencies, however, the reactance of C1 causes a reduction in feedback and the response rises to meet the open-loop response again, at a first-order rate. Although we predicted the response would be up by +3dB at 3.4Hz, this actually occurs a little lower around 2Hz, a modest error.

Using smaller values for C1 shifts the response to the right as more low frequencies

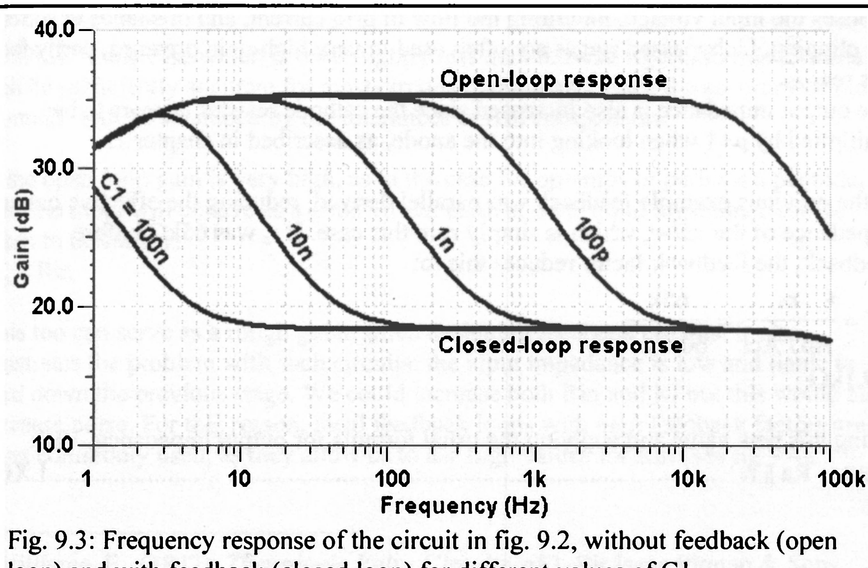

Fig. 9.3: Frequency response of the circuit in fig. 9.2, without feedback (open loop) and with feedback (closed loop) for different values of C1.

are prevented from being fed back. Note that making C1 ten times smaller shifts the response to the right by exactly one decade. Halving its value would shift it by exactly one octave. This is the principle behind active tone controls, which alter the amount of feedback at different frequencies to create boost and cut.

Note that if the cathode were only partially bypassed then the open-loop response would be lower at low frequencies, showing the expected 'shelved' response. The feedback would attempt to compensate for this, showing a more flat response (but still with a subdued shelf). However, combining partial cathode bypassing *and* a small value for C1 is not recommended as the phase shifts contributed by the two can cause ringing if they coincide, and sounds very unpleasant. Where feedback is used, the cathode should always be fully bypassed, or completely unbypassed. Of course, leaving it unbypassed would reduce the open-loop gain, spoiling some of the benefits of the added feedback loop, so this not a common arrangement.

Effect on input and output impedance:

The way in which feedback is derived and applied will affect output and input impedance. If the connection is a parallel or shunt one then *voltage feedback* is being used, and impedances are reduced; they are divided by the feedback factor. If we make a series connection then *current feedback* is being used, and impedances are increased; they are multiplied by the feedback factor.

For example, wherever we have an unbypassed cathode resistor then we have series-derived, current feedback. This serves to increase the input and output impedance of the valve. When the valve is not being overdriven then this makes little difference since the input impedance of valves is already extremely high even without the feedback. However, if we overdrive the valve to the point of grid current then this flows to the cathode and down the cathode resistor, developing a voltage which opposes the input voltage, inhibiting the flow of grid current, and this tends to soften the clipping. Unbypassed stages are often used in very high-gain preamps, partly for this reason.

The output impedance is also increased since the cathode resistor appears to be multiplied by $\mu+1$ when looking into the anode, as described in chapter 1.

In the previous example feedback was parallel derived, reducing the effective output impedance of the *valve*, which is simply r_a in this case. If r_a was 65kΩ before feedback, the feedback factor reduces this to:

$$r_a' = \frac{r_a}{Ao / Acl} = \frac{65k}{60 / 8.4}$$

$= 9.1\text{k}\Omega$

Using this new value and applying the usual formula for output impedance, VIII:

$Zout = Ra \parallel r_a'$ LXIII

In this case:

$$Zout = \frac{100k \times 9.1k}{100k + 9.1k}$$

$= 8.3k\Omega$

and this may need to be taken into account when choosing the output coupling capacitor, Co, when coupling the circuit to a following stage.

If global negative feedback is derived from the secondary side of an output transformer then the output impedance of the whole amplifier will be reduced, so the response of the amplifier will be less affected by differing loudspeaker characteristics (known as speaker damping). However, this is only of minor concern in hifi, and is completely irrelevant as far as guitar amplifiers are concerned since the output impedance is very low even before feedback.

The input impedance of the circuit in fig. 9.2 is less easily calculated. It can be shown that the true input impedance is given by[2]:

$$Zin = \frac{r_a(Rin + Rf) + Ra(r_a + Rin + Rf + \mu Rin)}{r_a + Ra + \mu Ra} \quad \text{..................................LXIV}$$

Estimating r_a to be 65kΩ this gives a value for Zin of about 55kΩ in this case (this value appears in parallel with the grid leak, making it even lower).
Note: if the stage drives a relatively heavy load, Rl, then Ra should be substituted with the value Ra||Rl = Ra x Rl / (Ra + Rl).

If Ra is large with respect to r_a, then the above formula may be simplified to:

$$Zin = Rin + \frac{Rf}{\mu + 1} \quad \text{..LXV}$$

or 52kΩ in this case which is only slightly less than the true figure, so this formula will be sufficiently accurate for most circuits, particularly since the accuracy of our formulae will be swamped by the tolerances of real components.

If the open-loop gain is very high, as in the case for op-amps or perhaps a pentode, then the input will behave as a good virtual earth and the input impedance can be taken to be simply:

$Z_{in} \approx Rin$

This too can serve as a rough guide when designing feedback circuits. It also illustrates the problem with such circuits: the input impedance is low and liable to load down the previous stage. We could increase both Rin and Rf but this would also increase noise. For this reason, local feedback loops with *high* feedback factors are most commonly used, as they allow us to use high values for Rin (see fig 9.4).

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[2] Williams, E., (1947). *Thermionic Valve Circuits*, p83. Sir Isaac Pitman & Sons, London.

Practical, virtual-earth amplifiers:

As a further example we will increase the feedback factor by choosing new values for Rin and Rf. The circuit is shown in fig. 9.4.
Although operationally identical to the earlier example, some changes have been made to the basic design. Firstly, the grid-leak resistor has been moved to place it outside the loop, so it may now be ignored, simplifying our mathematical treatment. Secondly, the feedback is now taken from the output coupling capacitor. This eliminates one capacitor, though we cannot now use it to tailor bass response in the usual fashion ,when coupling to the following circuit.

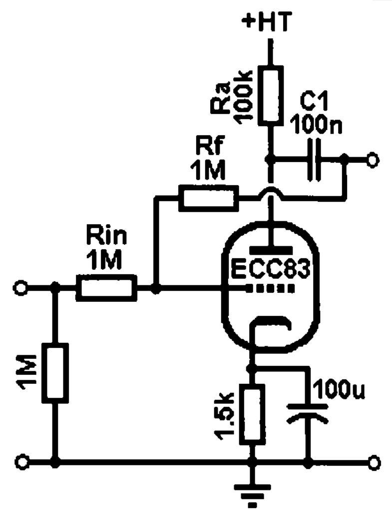


Fig. 9.4: A virtual-earth amplifier; a valve with a high feedback factor, sometimes known as an 'anode follower'.

The ratio B = Rf / Rin is now equal to 1, and we are feeding back almost all of the output signal. Rg is no longer part of the loop, so formula LXII simplifies to:

$$Acl = \frac{AoB}{Ao + B + 1} = \frac{60 \times 1}{60 + 1 + 1}$$
$= 0.97$

The anode resistance is reduced to an effective value of:

$$ra' = \frac{ra}{Ao / Acl} = \frac{65k}{60 / 0.97}$$
$= 1.05k\Omega$

Giving an output impedance of:
Zout = Ra || ra'

$$Zout = \frac{100k \times 1.05k}{100k + 1.05k}$$
$= 1.04k\Omega$

So we have created a stage with roughly unity gain and a very low output impedance, as well as a 60/0.97 = 62 or 36dB reduction in harmonic distortion and noise. Ideal for driving a tone stack or effects loop, perhaps. The input sensitivity has also been reduced by 60 times, so the stage will be very difficult to overdrive.
The circuit performance is very similar to a cathode follower, and indeed this circuit is sometimes known as an **anode follower**, although this name is misleading because the output signal is inverted. The disadvantage of this circuit when compared to the

true cathode follower is that its output impedance is marginally higher and its input impedance is lower (about 1M in this case, compared to several meg-ohms usually), but the advantage is that the cathode is not placed at high potential so heater-cathode insulation is not stressed.

Because of the high feedback factor the grid will act as a reasonably good virtual earth, so this circuit would make a good mixer, and an example of this is shown in fig. 9.5. This good virtual earth also allows the grid leak to be omitted; the grid instead gets its reference via the feedback resistor, provided there is no DC level present after the coupling capacitor (a 1M load resistor is shown, to illustrate how the valve gets its grid reference).
Inputs 1 and 2 have the same value input resistor, Rin1 and Rin2, while the third input has a smaller resistor, Rin3, and so should experience more gain. However, to calculate the gain for one input we need to take into account the shunting effect of the other two input resistors, which effectively replace Rg in formula LXII.

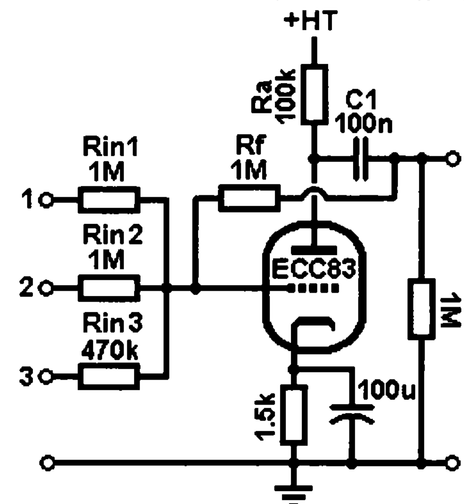


Fig. 9.5. A virtual-earth amplifier as a mixer offers good isolation between channels. Input 3 has higher gain than inputs 1 and 2.

For input 1, Rin2||Rin3 = 320kΩ, and this value is used for Rg:

$$Acl = \frac{AoB}{\frac{Rf}{Rg} + Ao + B + 1} = \frac{60 \times 1}{\frac{1000k}{320k} + 60 + 1 + 1}$$
$$= 0.92$$

So the gain is slightly lower than the previous example because of the shunting effect of the other two input resistors, though probably not enough to worry us. Input 2 sees exactly the same combination of resistances so will have the same gain.

Input 3 has a 470kΩ input resistor, so B = Rf / Rin3 = 2.1. It also sees a shunting resistance of Rg = Rin1||Rin2 = 500kΩ:

$$Acl = \frac{AoB}{\frac{Rf}{Rg} + Ao + B + 1} = \frac{60 \times 2.1}{\frac{1000k}{500k} + 60 + 2.1 + 1}$$
$$= 1.98$$

So this input has roughly double the gain of the other two and could be used for lower-level input signals, such as reverb or effects-loop recovery. These conclusions could also have been arrived at by applying the approximation of $A = Rf / Rin$, with only a small error.

The input impedance may be estimated to be:

$$Zin = Rin + \frac{Rf}{\mu + 1} = 1000k + \frac{1000k}{100 + 1}$$

$= 1010k\Omega$ for inputs 1 and 2.

And for input 3:

$$Zin = 470k + \frac{1000k}{100 + 1}$$

$= 480k\Omega$.

With a perfect virtual-earth mixer, such as an op-amp, dozens of inputs could be added, but with a valve the practical limit is probably around five.

Effects on overdrive:

So far we have observed that negative feedback *reduces* gain, noise and distortion, but *increases* the apparent headroom and bandwidth of the circuit within the loop. It also reduces the tonal changes that occur as valves age or are replaced with ones with slightly different characteristics, and it alters the input and output impedance of the amplifier. The use of negative feedback is therefore enormously advantageous to hifi design since it makes the amplifier more linear, and offers some independency from varying valve characteristics and loading circuits. Of course, a *guitar amp* does not need to be very linear, nor does it need a wide bandwidth. And if we desire less gain and more headroom it would be much easier to simply use a valve with lower gain and lower input sensitivity, such as the ECC82 / 12AU7. This would seem to make the usefulness of feedback in a guitar-amp limited to virtual-earth amplifiers and the occasional active tone controls, were it not for an effect that will now be discussed.

It was stated earlier that the stage required an input signal that was $1+A\beta$ times larger than normal to produce the same output as before feedback was applied. Suppose the circuit in fig. 9.2 needs a 1Vp-p signal to drive it to the point of clipping. Adding 10% negative feedback increases this to about 7Vp-p. What happens if the input signal is increased beyond this level to say, 8Vp-p? We expect 0.8Vp-p to be fed back, but this is not so: The stage is now being overdriven so the output signal will appear clipped, and so too is the feedback signal. The feedback signal is not 0.8Vp-p but 0.7Vp-p clipped, so the feedback factor has been suddenly reduced. The feedback signal cannot reduce the grid voltage by the normal amount, so the grid voltage is allowed to increase, which increases the degree of clipping, which reduces the feedback factor still further, which increases the grid voltage, and so on, until the feedback factor reaches some minimum value. Of course this description simply illustrates the point, in reality the process happens instantly. At the moment the grid voltage reaches the threshold of clipping the feedback loop collapses, and the stage effectively become open loop, and bandwidth and distortion revert to their full open-loop levels until the grid voltage falls below the threshold of clipping again. This creates a sudden transition between clean amplification and severe overdrive; there is no natural compression and gradual increase in distortion levels as there would be if the stage had no feedback. Additionally, when the input level falls back below the threshold of clipping, the instant reactivation of normal feedback conditions may sometimes cause a spike or some other spurious transient, which may give a different character to the overdrive when compared with a feedback-free amplifier.

If an amplifier has negative feedback a player can move between a clean and a very overdriven sound, but with very little middle ground, just by varying how hard he plays, provided the average input signal level is close enough to the threshold of clipping. When playing clean, the broader and flatter frequency response will also tend to smooth out the tone and give plenty of deep bass response, particularly if the feedback is of the global type taken from the speaker, which is the most common arrangement. This is especially useful for rhythm playing and is typical of most of the larger, classic Fender amps.
An amp using no negative feedback, however, will have high input sensitivity and therefore will generally only sound clean at very low volume settings, overdrive levels will then gradually increase as the volume is increased, so the player has more control over the actual amount and tone of the overdrive, but may have more difficulty swapping between a clean and overdriven sound. This is useful for a more expressive lead sound, since overdrive levels are controlled by pick attack. The Vox *AC15* and *AC30* amps are famous examples of amps having no global negative feedback, and are well known for their favourable lead tones but weaker rhythm tones. Classic Marshall amps, although very similar in design to Fender amps, generally use less negative feedback so fall somewhere between the two extremes. Whether or not negative feedback should be used therefore depends on the taste of the individual player, ignoring other design factors such as active tone controls. A jazz, blues or bass player requiring a smooth, clean sound will probably benefit from it, whereas a hard-rock or metal player who generally uses an overdriven sound may prefer to use very little or no feedback, and thereby have full control over the distortion. With global feedback a practical option is to make the feedback switchable between two or more preset levels, or to make it fully variable by means of a potentiometer (such a control is sometimes given the gimmicky name "variable damping"), as was illustrated in chapter 8, fig 8.19.

Stability and limitations on feedback:

Because negative feedback is a powerful tool it must be used cautiously. When local feedback is applied over a single gain stage, large feedback factors can be used without too much worry. However, when global feedback is applied over two or more gain stages we must be sure that the phase shift of the feedback signal does not deviate from the desired value, which in all the previous circuits was 180°*.

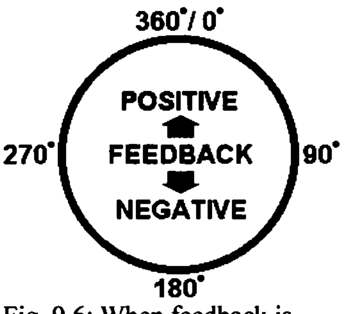


Fig. 9.6: When feedback is applied to the same grid as the audio input signal, feedback must be between 90° and 270° out of phase with the input to be negative, 180° being the ideal.

To illustrate the point we will continue to suppose that our feedback signal must be 180° out of phase with the input signal for normal operation. If the phase shift of the feedback signal deviates from 180° then the effect of feedback is reduced and the closed-loop gain will begin to rise. If the phase shift equals precisely +/– 90° then the feedback and input signals effectively do not 'see' one another, and we have a null point where amplifier will behave as if there is no feedback at all. If the phase shift exceeds +/– 90° then the feedback begins to reinforce the input signal –the feedback becomes positive– and the closed-loop gain will increase above the level of open-loop gain, and at precisely +/-180° the input and feedback signals are *in phase* and the gain reaches a maximum.
If the phase shift is +/–180° *and* the loop gain $A\beta = +1$ (unity) or greater, then the feedback signal is equal-to or greater than the input signal; the amp achieves *infinite* gain and will oscillate. If the loop gain is only slightly less than +1 then it will **ring**, that is, it will produce unsustained oscillations when an input signal of the right frequency triggers it. Even if the phase shift is not quite +/–180° the amp will still ring or oscillate if the loop gain is high enough. Problems with oscillation and ringing normally occur at high frequencies, caused by stray or unexpected capacitances, but low frequency instability is also possible and leads to effects such as **motorboating** (low frequency oscillation) or **breathing** (very low frequency oscillation, less than 1Hz).

Without an oscilloscope, ringing is a most difficult condition to diagnose because the amplifier may appear to work normally, since the bursts of oscillation will masked by the audio signal. The oscillations may also occur at ultrasonic frequencies and, although these are inaudible, they create intermodulation distortion products that *are* audible, and very unpleasant sounding indeed. This is, in fact, worryingly common in guitar amplifiers. Most manufacturers make no effort to ensure their amplifiers are perfectly stable, unlike hifi manufacturers, and the poorest designs the author has seen are almost on the verge of permanent oscillation. The usual symptom is that the

* With some topologies, such as when feedback is delivered to the unused input of a long-tailed-pair [see fig. 8.15] or to the cathode of an ordinary gain stage, then the feedback signal must be *in phase* with the audio input signal, because the feedback is being injected into an input which opposes the input where the audio is injected; the two are subtracted, rather than summed.

amp may sound normal most of the time, but on particular notes or at particular settings –often when the treble control is set high– the tone sounds 'dry', 'brittle', overly 'peaky' or even painfully high pitched. The last is sometimes known as the "ice-pick in the forehead" effect, which is a good description! Many of the most revered, 'classic' guitar amps exhibit oscillation or ringing at high settings, either in the preamp or poweramp, and this is sometimes exacerbated by component values drifting over time. Additionally, some popular DIY amplifiers are only stable when built in a 'particular way', and any small deviations in lead dress can tip the amplifier into ringing or oscillation. Needless to say, a proper design should be stable regardless of variations in layout and lead dress, within reason.

When designing an amplifier, we know which gain stages will be inverting or non-inverting, but there will usually be CR coupling networks between them. The phase shift contributed by each network will be 45° at the –3dB roll-off frequency, and increases asymptotically below this frequency, so any feedback loop which encloses such a network is immediately subject to phase shift of the signal being fed back. The more CR networks the loop encloses the greater the phase shift and the chance of oscillation, so it is advisable that no feedback loop enclose more than two CR networks, or any other type of phase shifting network for that matter, such as the output transformer, which has the most unpredictable phase-shift properties.
There will also be stray capacitances within the loop when the circuit is physically built, which will cause further unpredictable phase shifts at high frequencies (stray inductances are usually too small to be of any concern, except perhaps in the case of the cathode follower and cathodyne phase inverter) and the more feedback is added the more likely it is that these unknown phase shifts will become a significant problem. Clearly, *the greater the level of feedback the more unstable the amp is liable to be.*
Many hifi designs, particularly older ones in the style of the Williamson amplifier, apply feedback around the output transformer *and two* CR coupling stages. Such designs are *notorious* for ringing and are only really possible if the output transformer is of exceptional quality, though such designs are still popular among hobbyists in America even today. Later designs, notably those by Mullard and most guitar amps, wrap feedback around the output transformer and only *one* CR coupling stage (the one coupling the driver stage to the power valves), which is easier to make stable and is typical of contemporary European hifi amps. In cathode biased amps there may also be one or more cathode bypass capacitors enclosed within the feedback loop, but provided they are large (fully bypassing) then the phase shift introduced by these will be negligible, and even at the half-boost frequency they are unlikely ever to exceed 30°[3].
The cavalier attitude to feedback shown by so many guitar amp manufactures is old-fashioned and completely unacceptable. What's worse, it is not even difficult to make most amps stable; nearly all use the same arrangement, where feedback is taken from the secondary of the output transformer to the phase inverter, across one

[3] Langford-Smith, F. (1957). *Radio Designer's Handbook* (4th ed.), p 486. Illife & Sons, Ltd., London.

CR coupling stage. As will be shown, this topology requires only two small capacitors to make stable in almost all cases, and this should be done as a matter of course.

Stabilising a feedback amplifier:

Fig. 9.7 shows a theoretical amplifier consisting of three identical gain stages. We will suppose that each stage has a gain of 60, an output impedance of 40kΩ and an input capacitance of 100pF (which is shown dashed). For simplicity we will also assume that the cathodes are all fully bypassed. We wish to apply global feedback from output to input, but how can we be sure the amplifier will not ring or oscillate? Traditionally we would recourse to a Nyquist diagram, but a quicker and more intuitive approach is provided by Cherry[4] and is based on Bode's[5] observation that a frequency-response graph indirectly contains phase information too.

First we will consider the situation at high frequencies. The mid-band gain of the whole amplifier is simply 60 x 60 x 60, or 107dB. We also see that the loop will enclose three low-pass RC filters formed by the output impedance of each triode plus the following grid stopper and each triode's input capacitance. Each -3dB roll-off is known as a **pole**, and introduces 45° of phase shift, which increases with increasing frequency. By calculating the frequency of each pole we can produce an idealised frequency response graph.

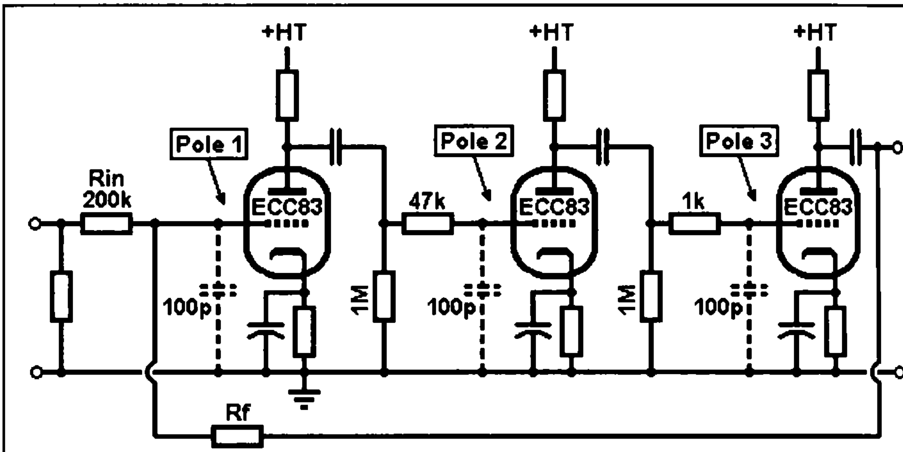


Fig. 9.7: Applying feedback around three time constants requires careful consideration of stability.

[4] Cherry, E. M., (1983). Designing NDFL Amps. *Electronics Today*, p46 – 51. April ed.

[5] Bode, H. W., (1945). *Network Analysis and Feedback Amplifier Design*. van Nostrand, Princeton, New Jersey.

Pole 1 has a roll-off frequency of:

$$f = \frac{1}{2\pi CRin} = \frac{1}{2\pi \times 100\times10^{-12} \times 200k}$$

= 8kHz.

Talking into account the 40kΩ output impedance of the first stage, Pole 2's roll-off occurs at:

$$f = \frac{1}{2\pi \times 100\times10^{-12} \times (40k + 47k)}$$

= 18.3kHz.

And again with Pole 3:

$$f = \frac{1}{2\pi \times 100\times10^{-12} \times (40k + 1k)}$$

= 39kHz.

The frequency response is plotted in fig. 9.8. The open-loop gain begins to fall off at a first-order rate (-6dB / octave) at 8kHz and therefore phase shift begins to exceed 45°. At 18.3kHz, pole 2 contributes a further -6dB / octave attenuation, resulting in a

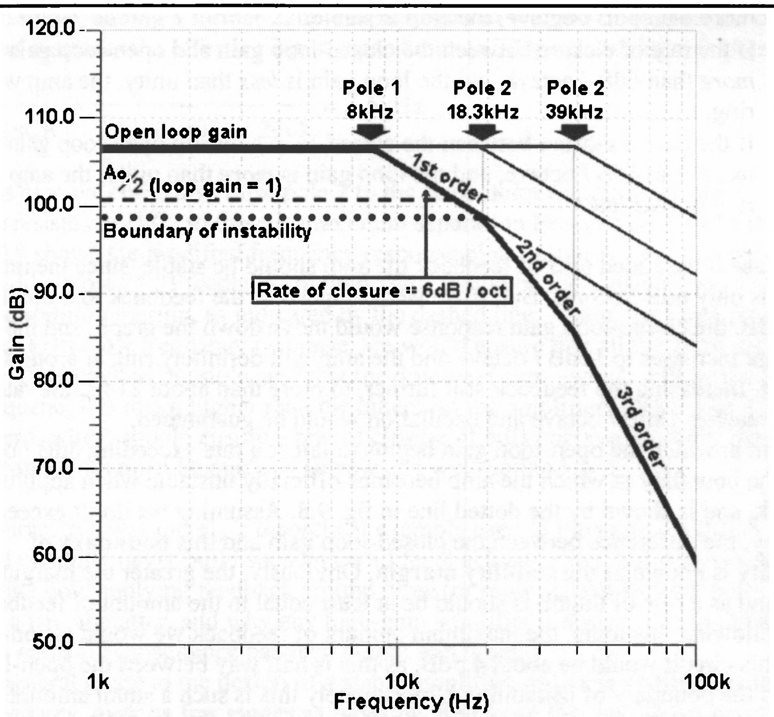


Fig. 9.8: An idealised frequency response graph can be used to assess the stability of a feedback amplifier.

second-order slope, and phase shift increases at a greater rate (and must be greater than 90° by definition), and at 39kHz pole 3 adds a further -6dB / octave, producing a 3rd order-slope and yet more phase shift (which must be greater than 135° by definition).

If 6dB of feedback is applied then the gain of the amp will be reduced by half, and the loop gain, Aoβ, will be equal to unity, since applying the universal feedback equation yields:

$$Acl = \frac{Ao}{1 + Ao\beta} = \frac{Ao}{1+1} = \frac{Ao}{2}$$

This is shown by the dashed line in fig. 9.8, and represents the closed-loop gain of the amp if 6dB of feedback were applied. If less feedback were applied then it would be impossible for the amp to oscillate since the loop gain would be less than unity, though it might still ring. We can see that the dashed line is horizontal and meets the open loop gain, which is falling at a rate of -6dB / octave. Therefore the **rate of closure** of the two lines is 6dB / octave, that is, the rate at which the two lines converge is *first order*.

The simplicity of Cherry's method is this:

- If the rate of closure between the closed-loop gain and open-loop gain is not more than 6dB / octave, the amp is stable.
- If the rate of closure between the closed-loop gain and open-loop gain is *more* than 6dB / octave, but the loop gain is *less* than unity, the amp will ring.
- If the rate of closure between the closed-loop gain and open-loop gain is *more* than 6dB / octave, and the loop gain is *more* than unity, the amp will ring or oscillate.

In this case if we added 6dB of feedback the amp should be stable, since the rate of closure is only 6dB / octave. But, if we were to increase the feedback to more than about 9dB, the closed-loop gain response would move down the graph, and the rate of closure increases to 12dB / octave and the amp will definitely ring at around 18.3kHz. Increasing the feedback still further, to more than about 21dB, the rate of closure reaches 18dB / octave and oscillation would be guaranteed.

The point at which the open-loop gain begins to fall at a rate exceeding 6dB / octave marks the boundary at which the amp becomes officially unstable when applying feedback, and is shown by the dotted line in fig. 9.8. Assuming we don't exceed this boundary, the difference between the closed-loop gain and this **boundary of instability** is known as the **stability margin**. Obviously, the greater the margin the better, and as a rule of thumb is should be at least equal to the amount of feedback used. Following this logic, the maximum amount of feedback we would rationally add to this circuit would be about 4.5dB, as this is half way between the open-loop gain and the boundary of instability. Unfortunately this is such a small amount of feedback that the feedback resistor would have to be unfeasibly large (hundreds of meg-ohms) because the open-loop gain is so high.

Even if we did not *deliberately* apply feedback, stray capacitance between output and input could be enough to introduce unwanted feedback, which is why high-gain preamps are prone to parasitic oscillation unless their gain at high-frequencies is reduced by installing anode bypass capacitors.

Exactly the same arguments follow for the low-frequency response. The loop contains three CR high-pass filters formed by the coupling capacitors and each valve's source resistance plus the following grid-leak resistances. Each -3dB roll-off is known as a **zero**, and introduces 45° of phase shift. One of these zeros is also determined by the value of feedback resistor, which we cannot choose, so our analysis need go no further.

Suppose we *do* want to add more feedback, how can we make this possible? To ensure an amplifier's stability we must force the rate of closure to be 6dB / octave, and this requires **slugging the dominant pole**. The dominant pole is the lowest-frequency pole within the loop, which is pole 1 in this case, at 8kHz. To slug the pole means to make it even lower, ensuring it is far removed from all the other, higher poles, so the amplifier 'looks' like it has only one HF roll-off frequency, as much as possible. We could increase the input resistor, but this adds noise. The best way to lower the pole frequency is to slug the input capacitance with yet more capacitance.
For example, adding a further 220pF capacitor from the grid of the first triode to ground increases the effective input capacitance to 320pF, lowering this pole to:

$$f = \frac{1}{2\pi CR} = \frac{1}{2\pi\, 320\times 10^{-12} \times 200k} = 2.5\text{kHz}$$

In this case we could also raise pole 2 to the same point as pole 3, by reducing the 47kΩ resistor to 1kΩ, so there is maximum separation between pole 1 and the others. Fig. 9.9 shows the modified frequency response plots, and it is clear that we could now add almost 28dB feedback, though 14dB would be the sensible limit to maintain a good stability margin, as indicated by the dashed line. Again, the same process applies to the low frequency response; we would ensure that all the zeros were as low in frequency as possible, and that the dominant zero –which will be the highest in frequency– is much higher than the others, so the amplifier 'looks' like it has only one zero, which should ensure a rate of closure of 6dB / octave. In short, the open-loop bandwidth must be made as narrow and well-defined as possible.

Of course, this was merely a theoretical example used to illustrate the theory; we would certainly never build such an awful feedback amplifier, but it serves to illustrate how applying feedback around three (or more) phase-shifting poles or zeros is not a trivial matter, and why any high-gain amplifier is susceptible to instability due to parasitic capacitance between output and input.
As a general guide to the design of a stable amplifier: **only use as little feedback as is necessary, over as few stages as possible, and keep the open-loop bandwidth of the amplifier as narrow as possible. Better still, use no feedback at all.** Luckily, guitar amps are more suited to these provisos than hifi amps.

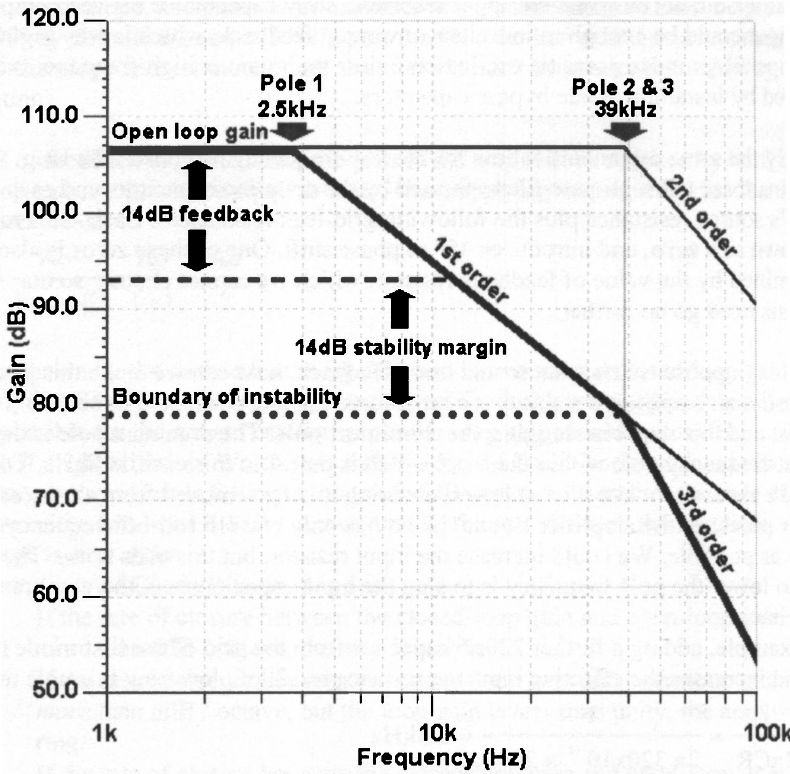


Fig. 9.9: Slugging the dominant pole increases stability and allows more feedback to be added safely.

Global feedback in practice:

Considering the previous discussion, it might be thought that applying feedback around an output transformer and *one* CR coupling stage would not pose a problem, but this overlooks a very important fact: *At high frequencies an output transformer has a second-order attenuation characteristic, due to its leakage capacitance and leakage inductance.* In other words, when the high-frequency response of the output transformer is -3dB, then the phase shift is already 90°, and not 45° as might be expected. At low frequencies an output transformer has a first-order attenuation characteristic, due to its primary inductance in parallel with its primary (reflected) impedance. Admittedly, even this is a simplification. Real transformers also have a bandwidth that varies with flux density, as well as a resonant frequency (which should be very high), which complicates matter. However, in a conventional amplifier covering only the audible range of 20Hz to 20kHz, the simplified model usually suffices.

Fig. 9.10 shows an example of an output stage and is similar in design to most push-pull guitar amps. An ECC83 long-tailed-pair is CR coupled to a pair of power pentodes (multiple pairs of output devices could be used, the analysis remains the same). The output stage is fixed biased, though cathode biased amps are usually fully cathode bypassed so treatment is also much the same.
We will take the gain of the ECC83 to *one* output as 30, and its output impedance as 40kΩ. The gain of each EL34 pentode is 12. The output transformer's primary impedance is 3.6kΩ anode-to-anode.

We need to know the voltage gain of the output transformer so we can calculate the open-loop gain of the whole circuit from phase inverter to speaker tap. Supposing we take feedback from the transformer's 8Ω speaker tap (it does not matter which tap is used) then the transformer's impedance ratio is 3.6kΩ:8Ω. Because the impedance is the *square* of the turns ratio, we find the turns ratio to be $\sqrt{3.6k / 8} = 21.1{:}1$. In other words, the input voltage signal to the transformer is divided by a factor of 21.1 (the voltage gain of the transformer is simply the reciprocal of this value). Since we are considering only one 'side' of the circuit we take half this value, or 10.6.

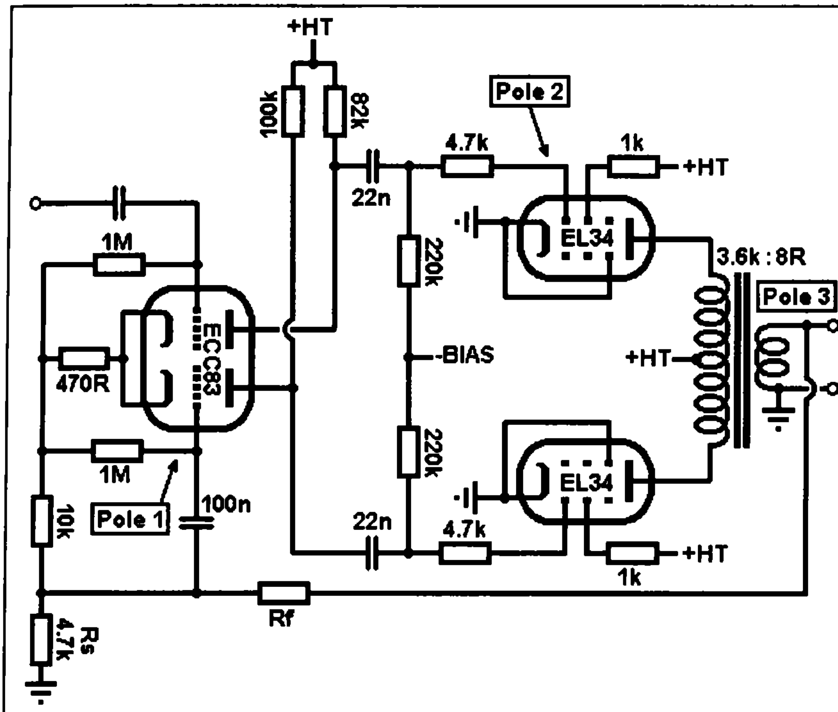


Fig. 9.10: A typical, unstabilised output stage with global feedback.

The total open-loop gain of the amplifier is therefore:

$$Ao = \frac{30 \times 12}{10.6}$$

= 34 or 30.6dB
Global feedback is then applied from the 8Ω tap to the phase inverter.

There are three poles within the loop:

- Pole 1 is due to the input capacitance of the phase inverter, where feedback is applied.

The gain of the triode is 30 and Cga = 1.6pF, Cgk = 1.6pF. Applying formula LVII the total input capacitance is:

$$Cin = Cgk(1 - A_k) + Cg1.A_a$$

$$Cin = 1.6(1 - 0.5) + (1.6 \times 30)$$

= 49pF

The pole frequency is determined by the input capacitance and the feedback and shunt resistors in parallel, Rf||Rs. Rf has not yet been chosen, but in a worst-case situation the frequency cannot be lower than:

$$f = \frac{1}{2\pi CinRs} = \frac{1}{2\pi \times 49\times10^{-12} \times 4.7k}$$

= 691kHz
which is very high and unlikely to be a problem!

- Pole 2 is due to the input capacitance of the power valves.

Each EL34 has a gain of 12 and the datasheet quotes Cg1 to all except anode as being 15.2pF*, and Cg1a = 1.1pF. If the screen grid is held at a fixed potential then the Miller effect is eliminated, and the input capacitance would simply be 16.3pF. However, there is a 1kΩ screen-grid stopper which will spoil some of the screen's shielding effect, so we will make an estimation of the total input capacitance to be 25pF.
The output impedance of the phase inverter is 40kΩ, which appears in parallel with the 220kΩ grid-leak resistors making 33.8kΩ, and the pentodes' grid stoppers are each 4.7kΩ giving a total source resistance of 38.5kΩ. This yields a pole frequency of:

$$f = \frac{1}{2\pi CinR} = \frac{1}{2\pi \times 25\times10^{-12} \times 38.5k}$$

= 165kHz.

* The Mullard data sheet gives this figure simply as 'Cin' as it is assumed that the screen is held at a fixed voltage.

- Pole 3 is due to the transformer's own high-frequency loss.

It might be possible to measure the transformer directly, although it must be driven at high-power levels to get a true figure for bandwidth, or we might rely on the manufacturer's data. We will suppose the upper limit quoted is -1dB at 30kHz. Because this is a *2nd order* roll-off the -3dB pole is actually about one octave higher than this figure, making 60kHz. If the manufacturer quoted a -3dB roll-off figure then we would use that instead, of course.

Immediately we have a worrying situation: the dominant pole is the output transformer, with its second-order (and somewhat unpredictable) attenuation characteristic, so the rate of closure is guaranteed to be at least 12dB / octave, and this is why so many amps of this type suffer from ultrasonic ringing, unless they use exceptional-quality, wide-bandwidth output transformers.
We cannot alter the output transformer, and we cannot slug pole 3 because it is in series with the feedback loop only; it causes a phase shift in the feedback signal but it does not affect the open-loop response in the same way. Instead we must slug pole 2, the input capacitance of the power valves, to force it to become the dominant pole, and there is more than one way to do this.

One method adopted in Marshall designs was to connect a 47pF capacitor between the anodes of the phase inverter. Because the circuit is balanced at this point, its value appears doubled when considering only one side of the circuit, since twice the signal voltage actually appears across it. If we ignore the grid stopper, whose value is fairly negligible, this slugs pole 2 to around:

$$f = \frac{1}{2\pi CR} = \frac{1}{2\pi \times (25\times10^{-12} + 94\times10^{-12}) \times 33.8k}$$

$= 38.9kHz.$

Although this is some improvement on older, unstabilised designs, it is still too high for good stability, being barely one octave below the output transformer's pole. It also relies on the amplifier being well balanced, which is an optimistic assumption. In reality the balance will probably be mediocre, and will also vary considerably when the power valves are overdriven. In short, it is a half-heated and cheap remedy, and allows less than 6dB of feedback to be added with dubious stability (this may be why early Marshall amplifiers used less feedback than Fenders). The Fender *Showman 6G14-A* used a 100pF capacitor in this position, which is slightly better. Some Mesa Boogie amplifiers use an even larger value of up to 220pF, which is a further improvement, slugging the pole to around 10kHz.
The Mesa Boogie *Dual Rectifier* uses a more elaborate arrangement. A 75pF capacitor is connected between the anodes of the phase inverter, and a 150pF capacitor is placed in parallel with each anode resistor (which is exactly the same as connecting it between anode and ground, of course). The amps use four 6L6GC beam tetrodes, but their input capacitance is roughly the same as an EL34, making about 50pF for two valves in parallel. The total shunt capacitance acting on each of

the phase inverter's outputs is increased to 150 + (2 x 75) + 50 = 350pF. This lowers the pole to around:

$$f = \frac{1}{2\pi \times 350 \times 10^{-12} \times 33.8k}$$

= 13.5kHz.

This is low enough to ensure stability even with the poorest output transformer, and is a conscientious design choice. However, this arrangement is actually more complicated than it really needs to be; the capacitor connected between the anodes is superfluous and is almost certainly a relic of older designs, and the presence of the grid stoppers introduces some error (though in this case it would be swamped by such large anode bypass capacitors).

The proper way to stabilise a circuit like this is to slug the inputs of the pentodes directly, as this ensures the most predictable operation. Adding a 220pF capacitor from grid to ground on each EL34 lowers the pole to:

$$f = \frac{1}{2\pi \times (25 \times 10^{-12} + 220 \times 10^{-12}) \times 38.5k}$$

= 16.9kHz.

This is not low enough to affect the amplifier audibly, but low enough to allow less than about 10dB of feedback to be added with a reasonable stability margin. Since this is a guitar amp with no need for an outstanding slew rate, the author would recommend slugging this pole even further to around 10kHz say, as some Mesa Boogie designs do, for maximum stability.

Some amateur builders may feel uncomfortable in adding such capacitances on the assumption that it will surely remove some audible treble or 'airiness', and spoil the tone, but this shows a misunderstanding of feedback, the differences between triodes and pentodes, and the limitations of guitar speakers *and* human hearing!
Firstly, there is very little useful harmonic presence above about 10kHz, and guitar speakers produce almost nothing above this range anyway. Also, most guitar amps of this type apply around 4dB to 10dB of feedback, so the closed-loop bandwidth will still be extended to well above audible frequencies, with no detriment to tone. The author *always* modifies existing feedback amplifiers in this way, as it ensures stability even with the most dubious output transformer and lead dress, and *should* be an industry standard. Heavily slugging the dominant pole also has the unexpected advantage that when the amp is overdriven and the feedback loop collapses, the HF response will resort to its lower, open-loop level, helping to reduce unpleasant high-order harmonics from being generated.

Choosing the feedback resistor:

In this case the boundary of instability is around 10dB down (see fig. 9.12) so we might choose around 6dB of feedback, reducing the gain by half from 34 to 17, or 30.6dB to 24.6dB.

To choose a value for the feedback resistor, Rf, we first find the feedback fraction, β, by rearranging the universal feedback equation:

$$Acl = \frac{Ao}{1 + Ao\beta}$$

So:

$$\beta = \frac{Ao - Acl}{Ao.Acl} = \frac{34 - 17}{34 \times 17}$$
$$= 0.03$$

And since Rf and Rs form a potential divider then to a close approximation for this circuit:

$$\beta \approx \frac{Rs}{Rs + Rf}$$

Solving for Rf gives:

$$Rf \approx \frac{Rs}{\beta - Rs} = \frac{4.7k}{0.03 - 4.7k}$$
$$= 152k\Omega$$

So we could choose 150kΩ as the nearest standard, or we might risk using 100kΩ for a little more feedback (around 8dB), which is what most Marshall designs use. The modified schematic is shown in fig. 9.11. For reference, the Marshall / Mesa stability capacitors are also shown, dashed.
If we had taken feedback from a 4Ω tap on the output transformer then we would have derived a turns ratio of $\sqrt{3.6k / 4} = 30 : 1$, so the open-loop gain would have been less by a factor of $1/\sqrt{2}$. Therefore, to achieve the same feedback factor would have required a feedback resistor, Rf, that is $1/\sqrt{2} = 0.71$ times smaller than the value that was calculated for the 8Ω tap. If taken from a 16Ω tap it would have been $2^2 = 4$ times *larger* than the value calculated for the 8Ω tap. Hence, it does not matter which tap feedback is derived from, only the feedback resistor need be scaled accordingly.

Fig. 9.11 also shows a capacitor placed in parallel with the feedback resistor (dotted). Its purpose is subtle: At high frequencies not only is there falling open-loop gain but also phase shift (the reader should be well aware of this by now!). The falling gain reduces the feedback factor which helps to compensate for the loss is gain, so the feedback extends and flattens the frequency response as described earlier in this chapter. However, at the highest frequencies, the input and feedback signals are separated by increasing phase lag, approaching -90°. To put it crudely, they do not "see" each other as much, and the effect is to reduce the feedback factor even further. This causes the closed-loop gain to rise somewhat at high frequencies, causing a 'hump' in the HF response. Some feedback amplifiers correct this by adding a small capacitance in parallel with the feedback resistor to allow these

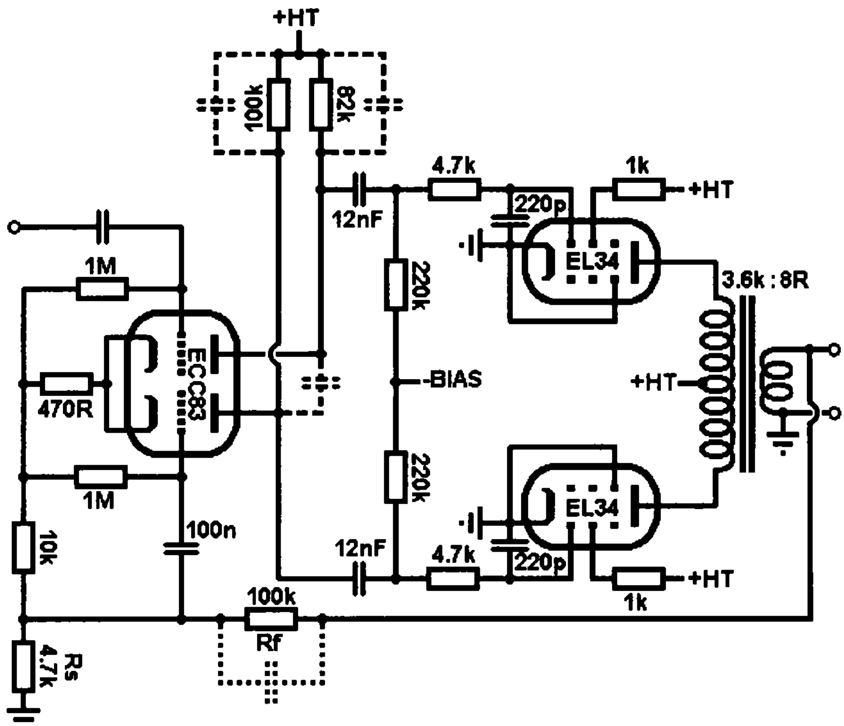


Fig. 9.11: Smaller coupling capacitors and the addition of 220pF grid-to-ground capacitors ensures stability when feedback is applied.

lagging frequencies to regain some phase shift and 'speed up', so it is sometimes known as a **speed-up capacitor**, and helps give a properly flat frequency response. There is no particular way to calculate this component as it is highly dependant on the real amplifier's parasitic capacitances when built, but as a rule of thumb the value is usually in the region of:

$$C \approx 2000.\frac{1}{Rf}$$

Where:
C = the speed-up capacitance in pico-Farads.
Rf = the feedback resistor in kilo-Ohms.

This component is almost never found in guitar amplifiers as there is no requirement for a perfectly flat frequency response, but if an otherwise stable amplifier does exhibit some unpleasant 'peakiness' then it may be worth adding. The usual method is to input a square wave (10kHz or so) into the amp and monitor the output across a resistive dummy load, on an oscilloscope. If the output wave shows some spiking of the rising edge then different values of speed-up capacitor may be added until the output is a reasonably good square wave.

Finally, we should also consider low frequency stability of our initial circuit in fig. 9.10. There are three zeros:
Zero 1 occurs where feedback is injected into the phase inverter via the 100nF capacitor. The input of the phase inverter is bootstrapped to around 2MΩ, so even if the feedback resistors are very small this zero is unlikely to be higher than:

$$f = \frac{1}{2\pi CR} = \frac{1}{2\pi \times 100\times10^{-9} \times 2000k}$$

= 0.8Hz
which is unlikely to be a problem.

Zero 2 is due to the high-pass CR coupling network between phase inverter and power valves. The roll-off due to the coupling capacitors is given by rearranging formula X:

$$Co = \frac{1}{2\pi f(Zout + Rg)}$$

So:

$$f = \frac{1}{2\pi C(Zout + Rg)} = \frac{1}{2\pi \times 22\times10^{-9} \times (40k + 220k)}$$

= 27.8Hz.

Zero 3 is due to the transformer's own low-frequency loss roll. Its primary inductance, L_{pri}, may be measured directly or may be given by the manufacturer. In this case we shall assume it is 30Henrys. The transformer's -3dB low frequency roll-off should occur at*:

$$f_{-3dB} = \frac{r_a \,||\, Z_{pri}}{2\pi L_{pri}}$$

Where:
r_a = the anode resistance of the power valves in parallel with one another.
Z_{pri} = the transformer's primary impedance (anode-to-anode).
L_{pri} = the transformer's primary inductance (anode-to-anode).

The EL34 datasheet gives a valve of anode resistance of 15kΩ, or 7.5kΩ for the two in parallel, yielding:

$$f = \frac{r_a \,||\, Z_{pri}}{2\pi L_{pri}} = \frac{7.5k||3.6k}{2\pi \times 30} = 12.9\text{Hz}$$

Again, this is concerning, because the two most dominant zeros are barely one octave apart, meaning no more than 6dB of feedback may be added without risk of motorboating, and illustrates why high-inductance output transformers are always

* Note that we do not use the low-frequency limit quoted by the manufacturer, since this figure is meaningless unless the test impedance (which is rarely stated) happens to match our source impedance, r_a.

desirable for power amps. Again, we cannot alter the transformer and we are unlikely to try using lower impedance output valves, so we are forced to alter the coupling capacitors. We could either make them much, much larger (1µF say) so their roll-off is below 1Hz and the transformer becomes the dominant pole (remembering that its low-roll off is first-order), or we can make them smaller, and this does have the added benefit of reducing the chances of blocking distortion. In a hifi amp we would already be concerned by a roll-off of 27.8Hz, but for guitar and even bass guitar we can safely raise this to around 50Hz or even higher, which could be met by using 12nF coupling capacitors, although 10nF would not be unreasonable, remembering that feedback will extend the closed-loop bandwidth below this frequency. In practice the falling load impedance at low frequencies causes the output power to also fall, reducing the feedback factor, which usually prevents motorboating. Nonetheless, the author strongly urges the use of properly chosen coupling capacitors. In an amplifier using multiple valves in parallel, the primary impedance and effective anode resistances will be greatly reduced, so the transformer's low roll-off frequency will usually be much lower also. This allows a lower roll-off frequency for the coupling capacitors also, and is one of many good reasons to use paralleled output valves in a bass or hifi amp.

The final frequency response is shown in fig. 9.12. The stability margin is 2dB, which is less than we would hope, though a higher-quality output transformer would improve this of course, or we could slug the dominant pole and zero even further. The stability of an amplifier may be tested by inputting a square wave (the test being performed at various frequencies across the audio band) while monitoring the output across a resistive dummy load, on an oscilloscope. There should be no spiking witnessed on the rising edge of the wave. In a guitar amp the rising edge may even

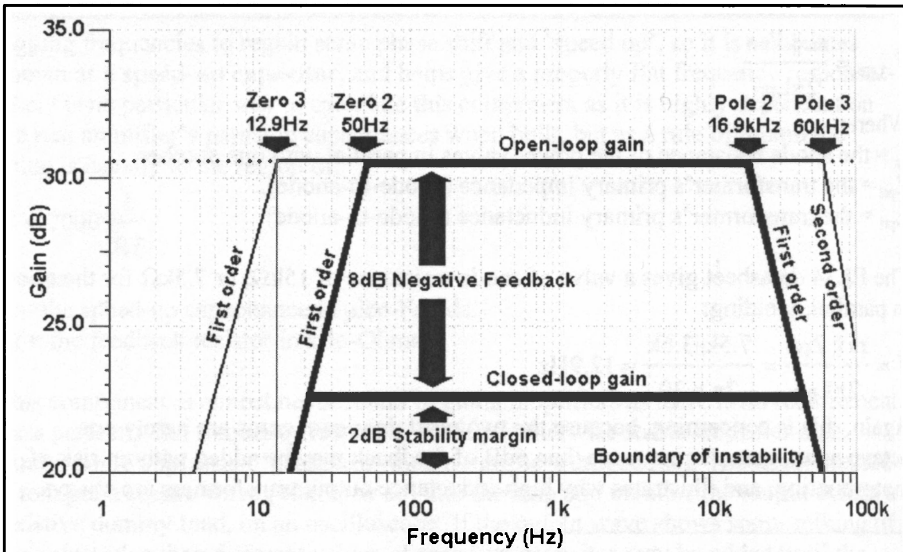


Fig. 9.12: The idealised frequency response of the circuit in fig. 9.11.

be well 'rounded' indicating that very high frequencies are attenuated: this is perfectly acceptable. Placing a 100nF capacitor in parallel with the load resistor gives an indication of the stability margin. If the square wave shows only modest peaking of the leading edge when the capacitor is added, then the stability margin is acceptable. If it does not peak at all with the capacitive load then the stability should be excellent!

Summary of formulae:

LXI; Universal feedback equation:

$$Acl = \frac{Ao}{1 + Ao\beta}$$

Where:
Acl = closed-loop gain
Ao = open-loop gain
β = feedback fraction

But if Ao and/or β is very large or infinite then $Ao\beta \approx 1 + A_o\beta$ and the above may be simplified to:

$$Acl = 1 / \beta$$

LXII; Gain of a triode with local feedback (fig. 9.13):

$$Acl = \frac{Ao.B}{\frac{Rf}{Rg} + Ao + B + 1}$$

Where: B = Rf / Rin

But if Rg is negligible or outside the feedback loop (shown dashed in fig. 9.13) then this simplifies to:

$$Acl = \frac{Ao.B}{Ao + B + 1}$$

LXIII; Output impedance of a triode with local feedback (fig. 9.13):

$$Zout = Ra \parallel r_a'$$

Where:

$$r_a' = \frac{r_a}{Ao / Acl}$$

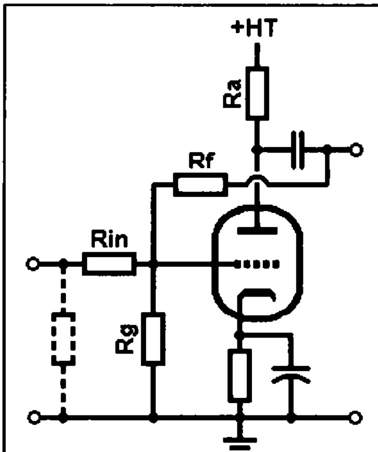


Fig. 9.13

LXIV; Input impedance of a triode with local feedback (fig. 9.3):

$$Zin = \frac{ra(Rin + Rf) + Ra(ra + Rin + Rf + \mu Rin)}{ra + Ra + \mu Ra}$$

LXV; If Ra is large with respect to r_a, then the above may be simplified to:

$$Zin = Rin + \frac{Rf}{\mu + 1}$$

But if μ is much greater than 1 then;

$$Z_{in} \approx Rin$$

Approximate value of a speed up capacitor;

$$C \approx 2000.\frac{1}{Rf}$$

Turns ratio of an output transformer;

$$n = \sqrt{\frac{Z_{pri}}{Z_{sec}}}$$

Where:
Zpri = the rated primary impedance.
Zsec = the rated secondary (speaker) impedance.
Voltage gain from primary to secondary is simply 1/n.

Output transformer -3dB low roll-off frequency;

$$f = \frac{ra \parallel Zpri}{2\pi Lpri}$$

Chapter 10: Tone Controls and Tone Stacks

Basic 'one knob' tone controls. Tone stacks. Designing an FMV tone stack. Expanding the FMV tone stack.

Although a large amount of tone shaping is permanently 'built in' to the preamp by means of partial cathode and anode bypassing, interstage coupling and so on, almost all amps will provide some user-adjustable tone controls. Some tone controls have already been described in earlier chapters (see figs. 1.20, 3.27, 8.18, 8.19) and might be regarded as being 'integral' to a particular gain stage. In this chapter we explore tone controls in their own right, as another 'building block' of the preamp.

Of course, an infinite number of tone control circuits is possible, limited only by the will and inventiveness of the designer. However, for guitar purposes there is a handful of well-used circuits and arrangements which are sufficient for most designs, and may be expanded upon (or simplified) to suit new designs, and these will be discussed.

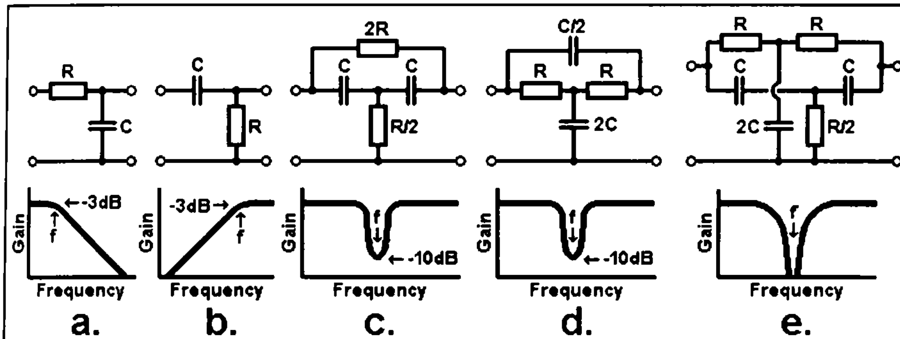


Fig. 10.1: Fundamental filters where $f = 1/(2\pi RC)$. **a.** low-pass filter (treble cut). **b.** high-pass filter (bass-cut). **c.** band-stop, bridged-T filter (mid scoop). **d.** band-stop, bridged-T filter (mid scoop). **e.** band-stop, twin-T filter (mid-scoop).

The fundamental RC filters used to construct tone controls are shown in fig. 10.1, and should already be familiar to readers. Virtually all the passive tone controls used in guitar amplifiers are simply variations and combinations of these filters, and most are simply variable versions of the interstage coupling networks described in chapter 2.

Inductors (chokes) are very rarely used for audio tone controls since very large values of inductance are normally required for use at such low frequencies, which tend to be difficult and expensive to obtain, and very bulky. Large inductors are also extremely sensitive to picking up interference and will inevitably give rise to unwanted hum in a practical amplifier. For these reasons, only resistance-capacitance tone controls are considered here.

Full mathematical treatment of filter circuits can easily consume an entire book, and is beyond the scope of this text. Instead, only basic formulae and suggested-circuits are given, which are easily modelled and altered using circuit simulation software, if necessary.

Notes on impedance bridging:

Tone-control circuits typically have rather low input impedances, so that if they are driven from a high source impedance, there will be significant attenuation of the signal, which is known as **insertion loss**. This may not be a concern in a low-gain, clean preamp, but for higer-gain circuits it is usually desirable to drive a tone circuit from a low impedance source such as a cathode follower, and this arrangement is found in coutless amplifiers. However, most of the tone circuits described in this chapter (unless otherwise stated) have been designed to have a relatively high input impedance, so that they may be driven from a typical ECC83 / 12AX7 gain stage (Zout $\leq 40k\Omega$) with hardly any more loss that if they are driven by a cathode follower. This should allow maximum flexibility when approaching a new design, since we are not immediately 'forced' to use a cathode follower to drive a tone circuit. Nevertheless, a low source impedance is certainly desirable, and a useful compromise is to drive the tone controls from a low anode resistance triode, such as an ECC82 / 12AU7 (Zout $\approx 8k\Omega$).

Additionally, most tone circuits have a high *output* impedance, which varies considerably with the actual setting of the control potentiometers. Consequently we should not heavily load a tone stack or further loss will be incurred, and the range and frequency response of the controls may be inhibited too. Most tone circuits will be followed by a $1M\Omega$ gain control or grid-leak resistor, or they may simply be connected directly to the grid of the following valve, if the tone-circuit provides the grid-leak path itself. For the least loading it is desirable to place a tone stack immediately before a stage having a high input impedance, such as a cathode follower, cathodyne or long-tailed pair

Notes on design:

For guitar purposes, 'treble frequencies' may be estimated to be frequencies of about 1kHz and above. 'Bass frequencies' may be taken to be around 200Hz and below. Middle frequencies then lie between 200Hz and 1kHz, and 600Hz to 800Hz is often taken as a 'centre' frequency when designing tone controls for guitar. Note that this is somewhat lower than the standard 1kHz used in hifi tone controls. For bass guitar all these figures may be roughly halved.

Strictly speaking, 'presence' applies to the upper-middle range of frequencies around 600Hz to 2kHz, in the region where the human ear is quite sensitive (this remains the same for both guitar and bass guitar). However, the 'traditional' presence control found in many guitar amplifiers actually acts on all frequencies above about 600Hz, and is really just an extended treble control.

For reference, a chart showing the frequencies of notes on the fretboard is given at the very end of this chapter.

Basic 'one-knob' tone controls:

Small practice amps often have only a single 'tone' control, which usually controls the treble content. This is usually required to compensate for the naturally bright (small diameter) loud speakers used in such amplifiers. 'One-knob' tone controls are also useful when it is desirable to distribute the frequency shaping throughout the preamplifier, so that progressive tone shaping is possible, rather than performing all the equalisation 'in one go'. This works particularly well in high-gain preamps. Generally it is desirable to place the bass control early in the preamp, so that later stages are less likely to be overwhelmed by low frequency content, which may otherwise lead to 'flubbiness' or blocking distortion. Treble control is well placed towards the end of the preamp, so that all of the harmonics generated in earlier overdriven stages can finally be adjusted to taste, to avoid shrillness. Alto or middle-frequency control can usually be placed almost anywhere in an amp, with good results.

Treble:

The very simplest tone control is based on the filter in fig. 10.1*a*, and is simply a treble-cut control, the same as the 'tone' control found on most guitars. This is shown in fig. 10.2 and it can be seen that it is simply a shunt capacitor in series with a potentiometer, allowing the degree of treble cut to be varied from negligible to infinity, with a -3dB roll-off frequency of:

$$f = \frac{1}{2\pi CR}$$

Where:
C = the treble shunt capacitance.
R = the source resistance in series with the capacitor when the pot' is turned to zero resistance, which is simply Zout in this case.

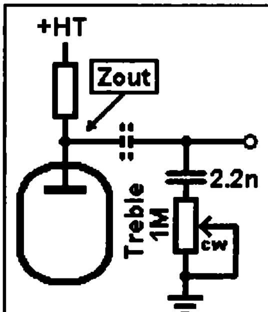


Fig. 10.2: The most simple 'tone' control (treble cut).

In this example the control is placed after the main coupling capacitor (shown dashed) so that the shunt capacitor need only be rated for the maximum AC signal voltage, which is rarely more than 250Vp-p. This is really just a variable version of the anode bypass capacitor explained in chapter 2, fig. 2.23.

The previous, primitive control can be improved somewhat by combining it with some other variable, such as a gain control or partial cathode bypassing. For example, fig. 10.3*a* shows the tone control used in the Fender *Princeton,* and several Gibson *GA* amplifiers, among many others. It should be obvious that when P2 is turned fully clockwise, C2 is effectively the 'bright

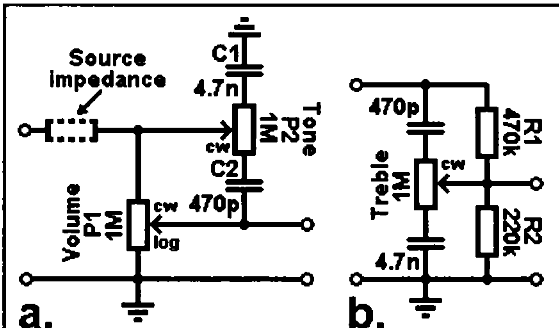


Fig. 10.3: **a.** the treble cut/boost control used in the Fender *Princeton.* **b.** an alternative circuit without volume control.

capacitor' which is often attached to gain controls [see fig. 2.17] and introduces an effective treble boost (unless the volume pot' is turned up fully). C1 is simply a treble-shunt capacitor, so P2 offers both treble cut and boost, although the degree of boost does depend on the setting of the volume pot'. Some interaction between control settings is common to most guitar-amp tone control circuits. Despite its simplicity, this circuit gives a very useful tonal range.

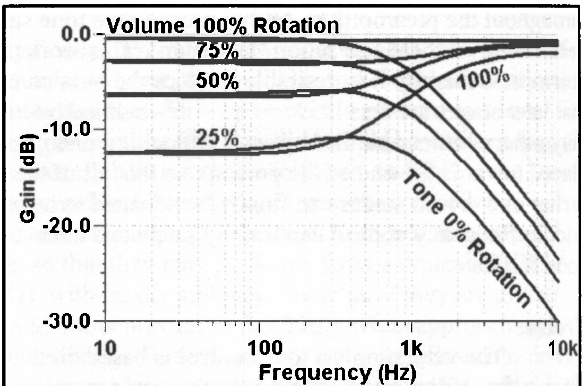


Fig. 10.4: Frequency response of the circuit in fig. 10.3*a* assuming a 40kΩ source impedance.

The +3dB boost due to C2 when the volume control is at 50% rotation was given in chapter 2:

$$f = \frac{2.74}{2\pi \, C2 \, P1}$$

The -3dB roll-off frequency cause by C1 when the tone control is turned fully down is as before:

$$f = \frac{1}{2\pi \, C1 \, R}$$

Where:
R = the source impedance.

Fig. 10.4 shows the frequency response of the circuit assuming a source impedance of 40kΩ, which is a typical output impedance of an ECC83 / 12AX7 gain stage. Note that an additional coupling capacitor will be needed before the circuit shown, in order to keep DC off the potentiometers, and may be chosen in the usual way.

If no volume control is used then it could be replaced by fixed resistor, R1 and R2 in fig. 10.3*b*, whose values may be large enough not to load down the previous stage. These attenuate the whole signal by a factor of R2/(R1+R2), or 0.3 in this case (-10dB). The apparent available boost is then +10dB when the treble control is turned fully up. When the treble control is at 50% rotation the frequency response is approximately flat, while below this point the treble is shunted to ground by the 4.7nF capacitor in the usual way.

Fig. 10.5 shows the treble control used in the Gibson *GA30RVT*, and is a combination of the simple treble-cut control [fig 10.2] and partial cathode bypassing [fig. 1.20].

When the pot' is turned fully up the cathode bypass capacitor is grounded and provides treble boost, while at the other extreme the cathode bypass is effectively removed and the treble-shunt capacitor is connected to ground instead. In this way the control may be regarded as semi-active. Fig. 10.6 shows the frequency response when used with an ECC83 with Ra = 100kΩ and Rk = 1kΩ.

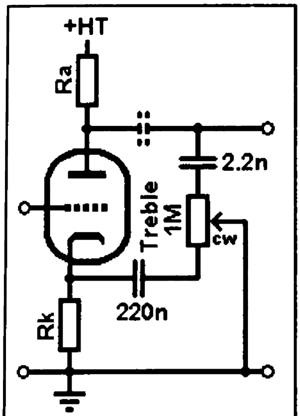


Fig. 10.5: A treble control providing active boost and passive cut.

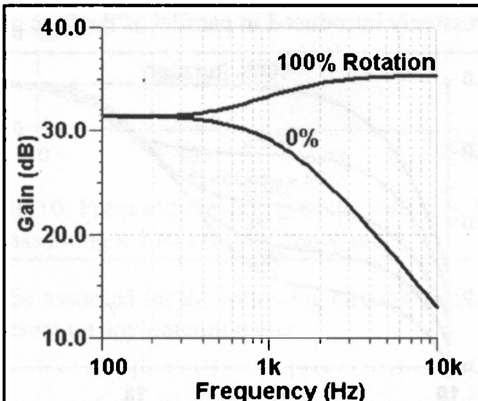


Fig. 10.6: Frequency response of the circuit in fig. 10.5 using an ECC83 with Ra = 100kΩ, Rk = 1kΩ.

Bass:

Most bass-cut controls rely on using a small coupling capacitor, so that lower frequencies are 'blocked', and this too may be made variable. One option is to switch between different values of coupling capacitor, as shown in fig. 10.7*a*. This is a characteristic feature of the Orange *Graphic* amplifiers and offers infinite, first-order bass attenuation. The -3dB roll-off frequency caused by each combination of capacitors depends on the loading resistance, which is often a 1MΩ volume pot', according to formula X:

$$f = \frac{1}{2\pi CR'}$$

Where:
C = the coupling capacitance.
R' = the total effective resistance in series with the capacitor, or Zout + load in this case.

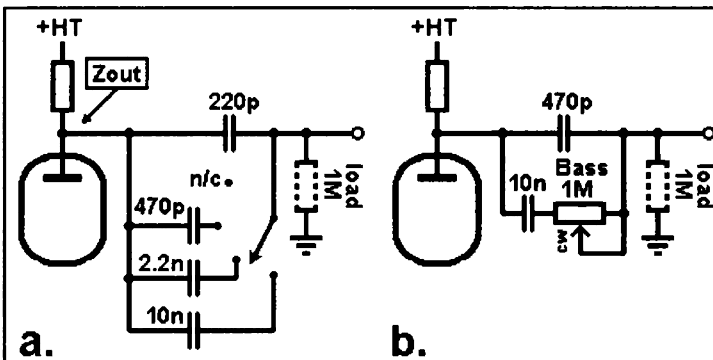


Fig. 10.7: **a.** selectable coupling capacitors for varying bass attenuation. **b.** bass-cut shelving control.

In this case the 220pF coupling capacitor is permanently connected while larger values can be switched in parallel with it. This avoids breaking the circuit completely while switching, which helps to reduce popping sounds.

Fig. 10.7*b* shows a bass-cut shelving filter, which works on a similar principle to that in *a*. A small value coupling capacitor is permanent while a larger value can be progressively introduced in parallel as the bass pot' is turned to zero resistance. The maximum amount of useful cut is determined by the potential divider formed by the pot' and the following load resistance. The frequency response of this circuit is shown in fig.10.8, where it can be seen that because both the pot' and the load resistance are equal, the maximum useful cut is -6dB (half). Reducing the load resistance would increase the amount of available cut, but would also load down the valve more, so some compromise must be come to with respect to available gain and the useful range of tone control.

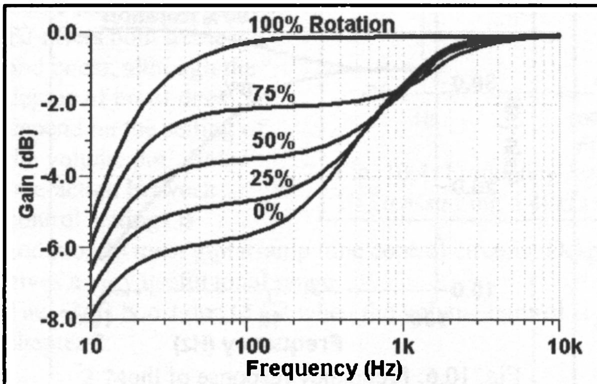


Fig. 10.8: Frequency response of the bass-cut shelving control in fig. 10.7*b*.

A more advanced bass control is shown in fig. 10.9, though it requires more components. At a glance it is clear that C1 resemble a 'bright' capacitor while C2 is a treble shunt capacitor. However, because the circuit is symmetrical about the wiper of P1, at 50% rotation the degree of boost and cut is equal, so the circuit offers a flat response, although the whole signal is attenuated by half (–6dB). When P1 is turned fully clockwise C1 is shorted, so any treble emphasis is removed, or in other words, a bass boost is effectively produced. At the other extreme C2 is shorted so that treble is now more emphasized and a bass-cut is produced. Assuming a symmetrical circuit such as this, the +/-3dB frequency at maximum and minimum cut is given approximately by:

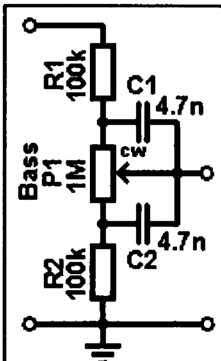


Fig. 10.9: A wide-range bass control.

$$f \approx \frac{1}{2\pi C1\,R'}$$

Where:
R' = R1 + the source impedance.

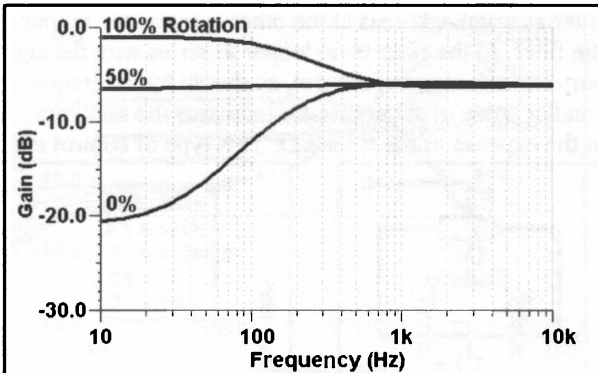


Fig. 10.10: Frequency response of the circuit in fig. 10.9 assuming a 40kΩ source impedance.

The frequency response of this control is shown in fig. 10.10. A variation on this circuit (combined with a bridged-T middle control) was used in several Silvertone amplifiers, the Fender *Bantam Bass* and Danelectro *Twin 15*. Again, an additional coupling capacitor is required, prior to the circuit shown, and may be an arbitrarily large value (47nF to 100nF say). This will also be assumed for the following circuits discussed, if it is obvious that DC would be present on any potentiometer.

Middle:

Treble cut and bass-cut filters are easily implemented almost anywhere in a preamp, but a middle cut or band-stop filter is slightly more elaborate, since it is a combination of the other two types. The Gibson *GA30RVT* used a bridged-T filter (fig. 10.1*d*.) with a potentiometer added to allow the amount of cut to be varied. A version of this circuit is shown in fig. 10.11. By using an unusually-large value of shunt capacitor the centre frequency is lowered and the maximum amount of cut is increased, but it is at the expense of frequencies adjacent to the centre frequency, that is, the Q of the filter is reduced. The frequency response is shown in fig. 10.12. Many Gibson amplifiers used bridged-T filters as fixed, interstage frequency shaping networks for a characteristic 'scooped mid' tone.

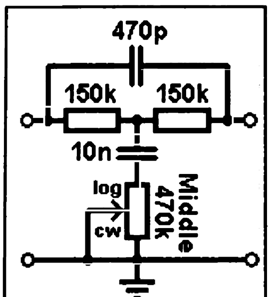


Fig. 10.11: A middle-cut control.

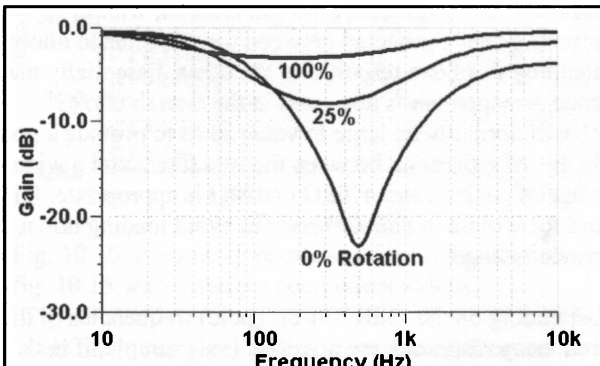


Fig. 10.12: Frequency response of the circuit in fig. 10.11 assuming a 40kΩ source impedance.

Fig. 10.13 shows a middle control using the bridged-T filter of fig. 10.1*c*. The bridge resistor is replaced by a potentiometer so that at one extreme the output is taken from the right side of the filter as usual, whereas at the other extreme the output is taken from the input side of the filter, so the filter is no longer in series with the signal path at all. This gives a very smooth range of control, as shown by the frequency response in fig. 10.14. A smaller value of shunt resistor increases the maximum amount of cut but, again, it is at the expense of the filter's Q. This type of control is favoured in Framus amplifiers.

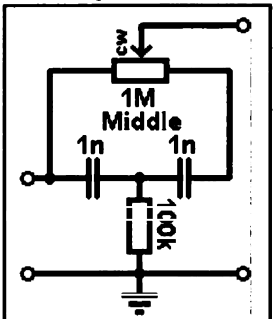


Fig. 10.13: A middle-cut control.

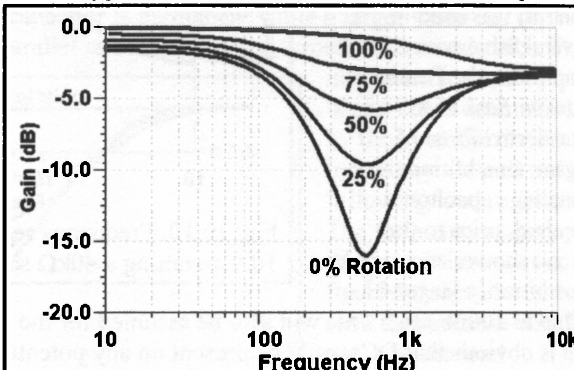


Fig. 10.14: Frequency response of the circuit in fig. 10.14 assuming a 40kΩ source impedance.

Tilt contol:

By combining complementary filters it is possible to produce a control which gives a substantially flat response in the centre position, but which simultaneously boosts treble and cuts bass when rotated in one direction, and does the opposite when rotated in the other direction. In this way the frequency response can be 'tilted' from one extreme to the other. Hence one **tilt** control can take the place of two individual treble and bass controls.

Fig. 10.15 shows the basic arrangement of the tone control used in the Electro-Harmonix *Big Muff-π* fuzz pedal. It should be obvious that it is a simply an RC (low-pass) and CR (high-pass) filter in parallel, with a potentiometer connected between their outputs to allow blending from one response to the other. Essentially the same arrangement is also used in the Gretch *G6162*.

PI will normally be large in value so as to provide a good degree of separation between the two filters and a wide range of control, and 470kΩ or 1MΩ is appropriate. R1 and R2 will be of similar order, to avoid loading down the previous stage.

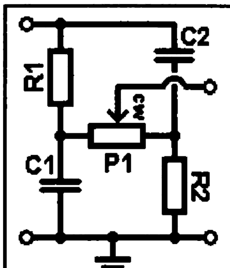


Fig. 10.15: The Big Muff-π tilt control.

Depending on the -3dB roll-off (pole) frequencies of the two filters, various mid-frequency responses are possible. For example, if both filter are assigned the same -3dB frequencies (that is, C1 = C2 and R1 = R2) then a substantially flat response is obtained at 50% rotation. This would be well suited for jazz or electro-acoustic use.

However, if the –3dB frequency of the low-pass filter is set *higher* than that of the other filter then a mid-hump is produced at intermediate settings, which might be useful for blues players using humbuckers.
Conversely, if the –3dB frequency of the low-pass filter is set *lower* than that of the other filter then a mid-scoop is produced at intermediate settings, which is useful for higher-gain playing styles. The further apart the poles are made, the more pronounced the hump or scoop becomes, although the range of control from bass boost to treble boost is reduced. If the poles are set one octave apart then the hump / scoop will be +/-3dB at the centre setting, and usually they should not be more than a decade apart if a wide range of control is to be retained. The frequency response graphs in fig. 10.16 show three example conditions.

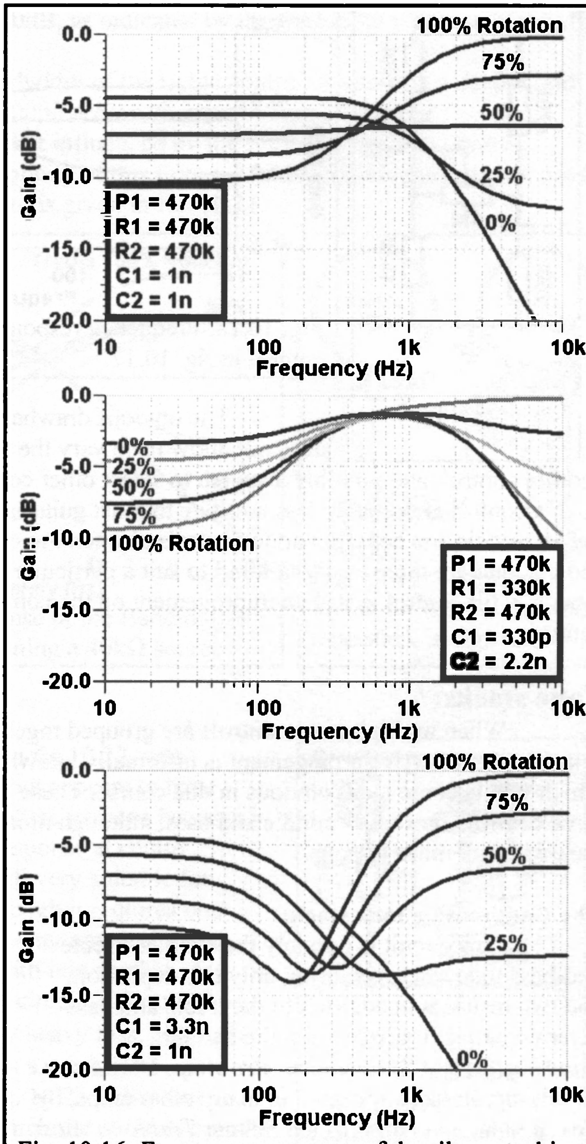


Fig. 10.16: Frequency response of the tilt control in fig. 10.15 with different component values.

With the addition of two more resistors we obtain the circuit in fig. 10.17. Assuming the same component values are used for each filter, this gives a more symmetrical frequency response which is appropriate for hifi use, as indicated by the frequency response plot in fig. 10.18. Indeed, an active version of this circuit was used in the Quad *34* preamplifier. For guitar purposes this tilt control has little advantages over the Big Muff-π type control, but is included here for completeness.

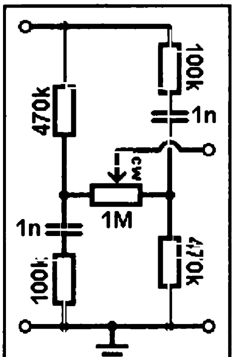


Fig. 10.17: The Quad-34 tilt control.

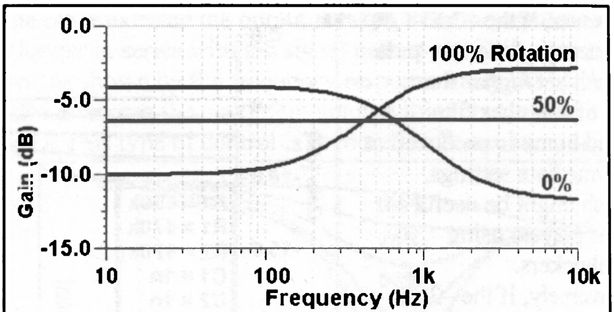


Fig. 10.18: Frequency response of the Quad-34 tilt control in fig. 10.17.

The obvious drawback of a tilt control is that it does not allow us to vary the mid boost or scoop, unless another control, and possibly a switch to select other component values, is provided. Such a control is probably less intuitive to most guitarists than separate knobs for treble, middle and bass. Separate controls also tend to offer a greater degree of boost and cut, and are more easily tailored to suit a particular tone. For simple amplifiers though, a tilt control is still an improvement on the conventional 'tone' (treble cut) control.

Tone stacks:

When multiple tone controls are grouped together into a single circuit 'block', the collective arrangement is informally known as a **tone stack**, a name which will become more obvious in due course. These generally consist of two or three controls for treble, middle and bass, although more advanced stacks may include many more options.

The Bandmaster tone stack:

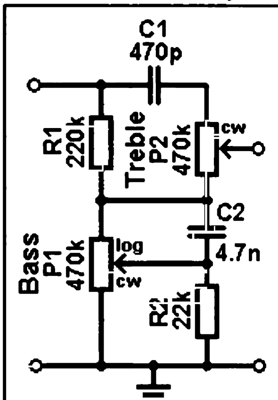


Fig. 10.19: An extremely simple treble-and-bass tone stack based on the Fender *Bandmaster* circuit.

This circuit is possibly the simplest treble-and-bass tone stack, requiring only two capacitors and two resistors in addition to the treble and bass potentiometers. It appeared in the Fender *Bandmaster*, and 'brown face' *Pro-Amp*. Variations on this circuit have appeared in many other amps, by Vox, and later versions of the Selmer *Treble 'n' Bass 50W* amplifiers. The circuit is shown in fig. 10.19 and it can be seen that the treble pot' appears to be 'stacked' on top of the bass control, and it is classic circuits like this and the FMV circuit which give rise to the term 'tone *stack*', even though many tone circuits do not actually have this stacked appearance. The component values shown in fig. 10.19 differ from the original and have been chosen to yield less loss when driven from a 40kΩ source impedance, the

'average' loss being about –10dB, as indicated by the frequency response plot in fig. 10.20.
Reducing C1 reduces the bandwidth of the treble control, generating a deeper mid' scoop, and values as low as 100pF are worth experimenting with.
The treble response is somewhat influenced by the setting of the bass control, although considering the simplicity of this circuit it is remarkable that it is not used more often. A suggested layout is given in fig. 10.21.

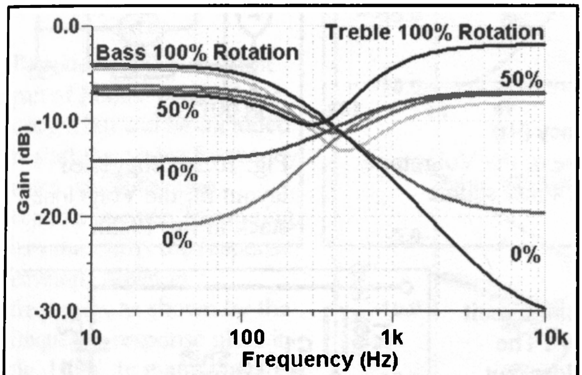


Fig. 10.20: Frequency response of the Bandmaster tone stack in fig. 10.19 assuming a 40kΩ source

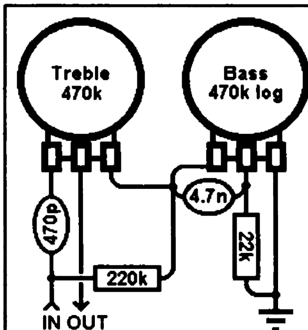


Fig. 10.21: Suggested layout for the Bandmaster tone stack in fig. 10.19.

The Voigt tone stack:

This circuit, shown in fig.10.22, was described by Voigt in 1940[1], but appears to have been almost completely forgotten, perhaps because it did not give a perfectly flat response at centre settings. Nevertheless, it is still very suitable for musical instruments, and although it requires one more resistor than the Bandmaster tone stack, the controls are less interactive with one another. At centre settings the gain of the circuit is about 0.2 (–14dB) which is a relatively heavy loss, although the range of control is large as a result. The frequency response plot in fig. 10.23 indicates that the centre frequency about which the controls operate is around 200Hz, all of which makes this circuit well suited for use with clean, bass amps.

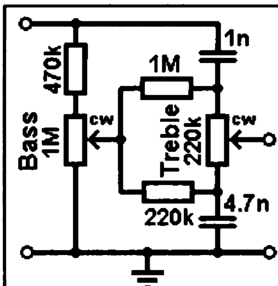


Fig. 10.22: The Voigt tone stack has been almost forgotten, but is particularly suited to bass guitar use.

[1] Voigt, G. A. H., (1940). Getting the Best from Records: Part III. *Wireless World*, (April), pp210-212.

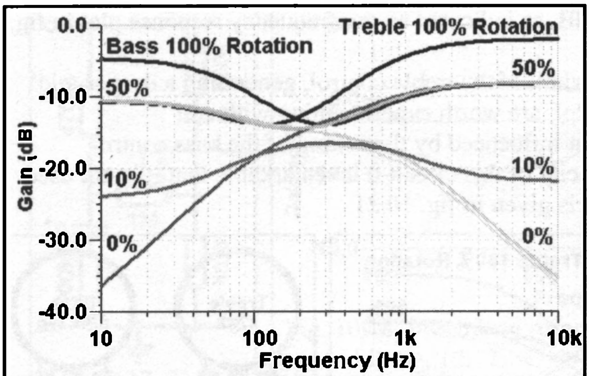


Fig. 10.23: Frequency response of the Voigt tone stack in fig. 10.22 assuming a 40kΩ source impedance.

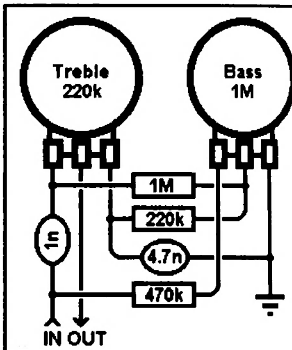


Fig. 10.24: Suggested layout for the Voigt tone stack in fig. 10.22.

The James tone stack:

This enormously popular circuit was described by James in 1949[2]. The circuit itself is actually much older, but James' name has become permanently attached. Unfortunately some modern texts incorrectly refer to this circuit as a 'Baxandall' tone control, which may lead to some confusion as this term actually applies to a subtly different circuit arrangement, which is described next. The James circuit is shown in fig. 10.25; component values have been altered from the original, to suit guitar use and to reduce the inherent loss of the circuit to around –7dB. Some readers will notice that it is really a combination of the 'one knob' treble and bass controls already described in figs. 10.3*b* and 10.9. The only difference is that, in this circuit, the bass control doubles as the potential-divider originally required by the treble control in fig. 10.3*b*.

Numerous amplifiers have used this stack, or subtle variations on it,

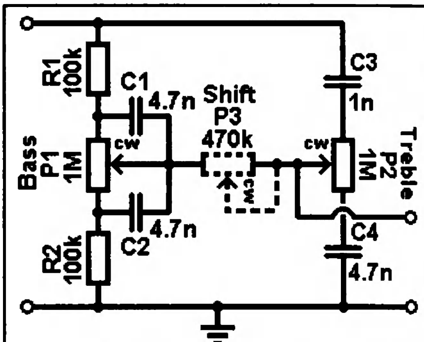


Fig. 10.25: The James tone stack with optional shift control.

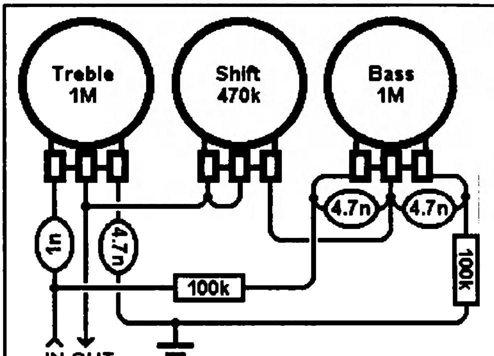


Fig. 10.26: Suggested layout for the James tone stack in fig. 10.25.

[2] James, E. J. (1949). Simple Tone Control Circuit. *Wireless World*, (February) pp48–50.

including amps by Ampeg, Dynaco, Garnet, Magnatone, Orange, and the Fender 'blonde' *Twin Amp*, making this perhaps the second most popular tone stack, after the FMV circuit (see later).

Potentiometer P3 was not part of James' original design, but can be included to shift the centre frequency about which the two controls operate. Increasing its value shifts the response envelope down in frequency, as shown by the frequency response plots in fig. 10.27. In many ways this is more useful than a simple middle control, particularly for bass guitar. A very similar circuit is used in the Carlsbro *60TC* amp.

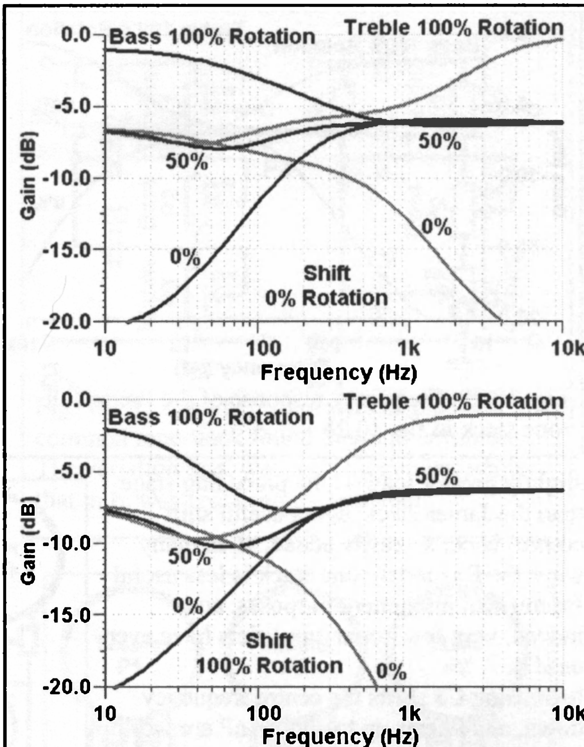


Fig. 10.27: Frequency response of the James tone stack in fig. 10.25.

The passive Baxandall tone stack:

This highly influential circuit is based on a design by Peter Baxandall, published in 1952[3]. The paper actually described an active circuit, which is why this version is referred to as the 'passive Baxandall'. The original also required a tapped potentiometer, but this is normally omitted since they are not readily obtainable.

As seen in fig. 10.28, it incorporates the same bass control as the James tone stack, but the treble arrangement is different. Resistor R3 is required to separate the two controls (without it the treble control is

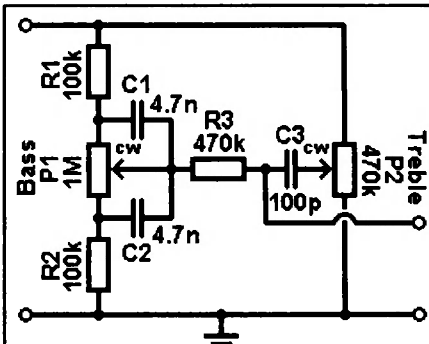


Fig. 10.28: The passive Baxandall tone stack is somewhat less versatile than the James stack.

[3] Baxandall, P. J. (1952). Negative-Feedback Tone Control. *Wireless World*, (October) pp.402-405.

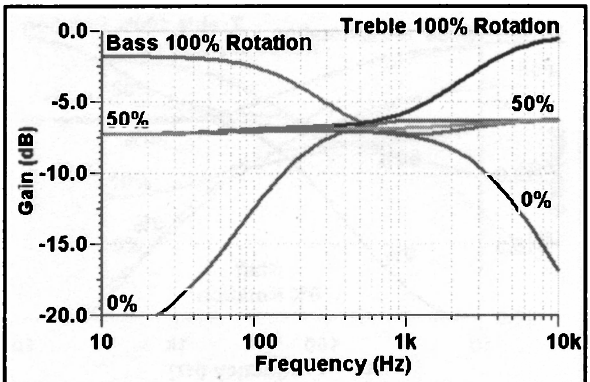


Fig. 10.29: Frequency response of the Baxandall tone stack in fig. 10.28.

rendered useless). This circuit arrangement is capable of giving a near-perfectly flat response at centre setting (see fig. 10.29), good isolation between controls and a very predictable, though subtle, range of control; unsurprisingly it became very common in hifi amplifiers.
However, for equivalent component values the Baxandall stack places a slightly greater load on the preceding stage than the James stack, and a useful shift control is not so easily added, so in many ways the Baxandall tone stack is less useful for musical instrument purposes, and indeed, very few guitar amplifiers have ever used it.
Increasing C3 shifts the centre frequency down, and values up to about 1nF are worth experimenting with. An obvious addition would be to place a capacitor (470pF say) and switch in parallel with C3 to allow the centre frequency to be shifted down by about one octave.

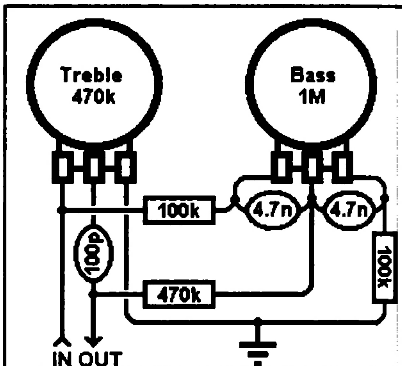


Fig. 10.30: Suggested layout for the passive Baxandall tone stack shown in fig. 10.28.

The FMV tone stack:

This circuit is the most common tone stack found in guitar amps. Originally devised by Fender, it was famously copied by Marshall and Vox, and used almost exclusively of any other tone stack, from which the abbreviation **FMV** is obtained. It has since become regarded as an 'industry standard' for guitar amps, and has been used by almost every other guitar amplifier company, at one time or another. The attraction of the circuit is obvious; it requires only three capacitors and one resistor in addition to the treble, middle and bass potentiometers. However, the circuit is quite lossy, the controls are highly interactive and the middle control leaves something to be desired (this will be expounded upon in due course) and readers should not look upon it as the 'ideal' tone stack, but merely as the most economical. In spite of its shortcomings, it is so well-known that it deserves special attention.

Two variations on the FMV tone stack are shown in fig. 10.31, the only difference being with the connection of the middle control. This makes very little difference to the functioning of the circuit, except at minimum settings. With the connection in *a.*, used by Fender, when all controls are set to minimum the signal is completely shunted to ground; the amp will be silenced. This may confuse some players and so the arrangement in *b.*, favoured by Marshall, may be used to avoid this.

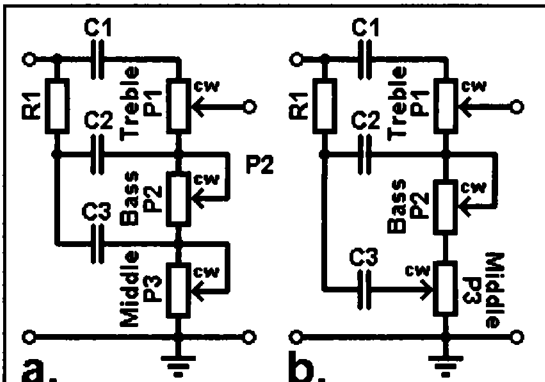


Fig. 10.31: The FMV tone stack is the most common tone stack found in guitar amplifiers. The variation in **a.** was used by Fender, while that in **b.** was preferred by Marshall.

R1 is known (inexplicably) as the **slope resistor**, and for the most part it determines the minimum input impedance of the circuit, and should be made sufficiently large so as not to load down the previous stage too grossly. If the stack is driven by a cathode follower then values between about 33kΩ and 100kΩ are usual, whereas if it is an ECC83 / 12AX7 gain stage then values of around 100kΩ to 220kΩ will be necessary. Typical frequency response plots are given in the next section.

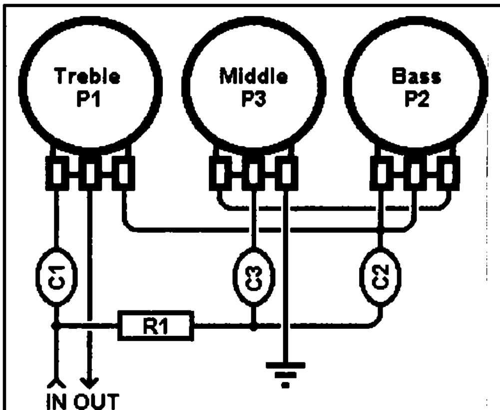


Fig. 10.32: Suggested layout for the FMV tone stack in fig. 10.31b.

Designing an FMV tone stack:

The easiest way to design this type of tone stack is to begin at the bottom, with the middle control. The following method is sufficiently accurate for both versions of the circuit shown in fig. 10.31.

Let us assume that the middle control, P3, is set to maximum, while the bass and treble controls, P1 and P2, are set to minimum, since this yields an approximately flat response. This is shown in fig. 10.33*a*. At middle frequencies, C1 is small enough that it may be ignored while C2 and C3 appear in parallel and are large enough that we may assume they are short-circuits. The circuit may then be reduced to the equivalent circuit in fig. 10.33*b*.

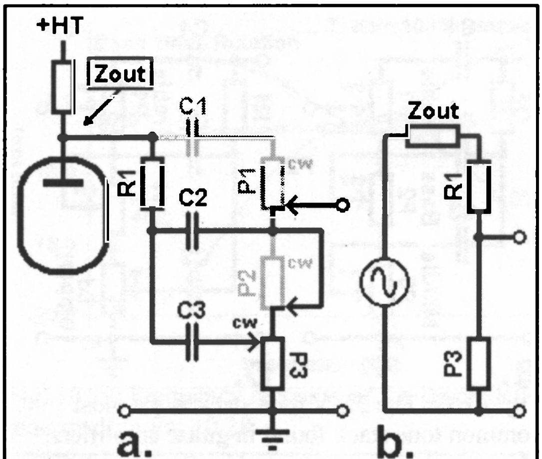


Fig. 10.33: At middle frequencies, with P3 at maximum and P2 at minimum (**a.**), C2 and C3 appear in parallel and the circuit may be reduced to the Thévenin equivalent in **b**.

We will suppose the stage is to be driven from a typical gain stage with an output impedance of 40kΩ, and that we take the slope resistor, R1, to be to be 100kΩ.
We may now choose P3.
Since Zout + R1 and P3 form a potential divider, this determines the degree of attenuation at the flat setting. The greater the attenuation, β, then the greater the amount of apparent treble and bass boost we can 'add back' with the controls at maximum. However, this also means the greatest reduction in middle-frequency signal level, where most of the guitar's range lies. The classic Fender amps use around –23dB of attenuation giving a wide tonal range but a low-gain, clean preamp. Classic Marshall amps, reputed for their higher levels of gain and cathode-follower driving stage, use only about –7dB attenuation.
In this case we might choose an intermediate value of –10dB.

$$\beta = 10^{dB/20} = \frac{P3}{Zout + R1 + P3}$$

In this case:

$$\beta = 10^{-10/20}$$

= 0.32
Solving the earlier equation for P3 gives:

$$P3 = \frac{-\beta(Zout + R1)}{\beta - 1} = \frac{-0.32 \text{ x } (40k + 100k)}{0.32 - 1}$$

= 66kΩ.
The nearest standard potentiometer value is 47kΩ, which actually gives an attenuation of:

$$\beta = \frac{47k}{40k + 100k + 47k}$$

= 0.25 or -12dB
Which is reasonable. The weakness of the middle control also becomes obvious: whereas treble and bass can be boosted above this 'flat level', the middle control is already at maximum at this point, so it can only *cut* middle frequencies further. Its range of cut is also rather limited and very interactive with the treble control.

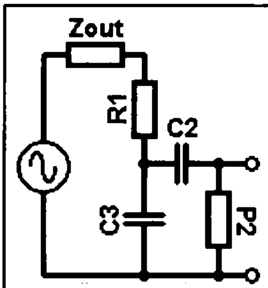


Fig. 10.34: Low-frequency Thévenin equivalent circuit,with the middle control is set to minimum and the bass control set to maximum.

We may now choose C3 by assuming that the bass control is set to maximum, and the middle control is set to minimum, giving the equivalent circuit in fig. 10.34. It can be seen that C3 shunts upper frequencies to ground, while the remaining lower frequencies are passed to the bass control P2. Therefore, C3 determines the 'upper end' of the bass control's range. To a very close approximation, the +3dB transition frequency between the middle and bass bands is then:

$$f \approx \frac{1}{2\pi C3(Zout + R1)}$$

The Fender *5F6 Bassman* uses a frequency of 140Hz (for use with Bass guitar) while the Marshall *JMP50* redesigned this to 213Hz.
Suppose we chose a frequency of 200Hz:

$$C3 = \frac{1}{2\pi f(Zout + R1)} = \frac{1}{2\pi\, 200\, (40k+100k)}$$

= 5nF
The nearest standard value is 4.7nF giving a frequency of 242Hz, which is a reasonable figure for a design with more mid-range focus. A value of 6.8nF would give 167Hz, which would be more suitable for use with bass guitar. Given the limited range of capacitor values it is often easier to adjust the value of the slope resistor at this point.

We may now choose the bass components P2 and C2.
The larger the value of P2, the greater the degree of bass-boost possible, since the theoretical, *minimum* amount of bass cut is then:

$$\beta_{(bass)} = \frac{P2 + P3}{P2 + P3 + R1 + Zout}$$

In practice this value is only achieved at very low frequencies, and only if C2 is made very large, so a high value of 1MΩ is normally used. However, this can make the tone stack rather 'bass-heavy', except at the very lowest bass settings. This may be suitable for rhythm playing with single-coil pickups, but can make the control of bass content somewhat fiddly when using hum-buckers or playing lead, and the author prefers to use a lower value of around 470kΩ. This may be a linear or logarithmic type, depending on taste.
C2 determines the 'lower end' of the bass control's range, but this is normally set below audio frequencies and has the least tonal impact of all the components in the circuit. In fact, C2 could be omitted completely, and replaced with a wire link, were it not for the necessity to keep DC off the potentiometer.

With the controls at maximum, the lowest, –3dB transition frequency is then:

$$f = \frac{1}{2\pi C2(Zout + R1 + P2 + P3)}$$

Although this will rise somewhat as the bass control is turned down.
A typical value is arbitrarily low, at around 10Hz. In this case a value of 10nF for C3 would yield a frequency of:

$$f = \frac{1}{2\pi C2(Zout + R1 + P2 + P3)} = \frac{1}{2\pi \times 10\times10^{-9} \times (40k + 100k + 470k + 47k)}$$

= 13Hz.

We may now choose the treble components P1 and C1. The value of P1 should be large enough that C1 can be made small, otherwise middle frequencies will be passed via C1 straight to the output, and the range of the treble control will be poor. In other words, the other two controls control will become exceedingly interactive with the treble control. However, if P1 is made too large then not only will it contribute a good deal of noise but C1will need to be so small that stray wiring capacitance becomes a significant factor. A value of between 220kΩ and 470kΩ is normally ideal.
C1 now determines the 'lower end' of the treble control's range, the -3dB transition frequency being given approximately by:

$$f \approx \frac{1}{2\pi C1(P1 + Zout)}$$

Setting it too low will rob the middle control of range, so a typical frequency is around 2kHz. If we use a 470kΩ potentiometer then:

$$C1 \approx \frac{1}{2\pi f(P1 + Zout)} = \frac{1}{2\pi \times 2000 \times (470k + 40k)}$$

= 156pF.

Values of 150pF and 120pF are readily available. Often this will be a ceramic capacitor although, as suggested in chapter 2, a poly' type is preferred by most players.
The final schematic is shown in fig. 10.35, together with the frequency response in fig. 10.36. If the circuit were driven by a cathode follower then Zout will normally be small enough that it drops out of the previous equations, but it will be clear to most readers that with this method the FMV tone stack can be designed to give the same response with *any* source impedance, the only limitation being how large we are willing (or able) to make the resistances.

Because all three capacitors block DC from reaching the potentiometers they will normally need to be rated for the full HT voltage. However, commercial manufacturers may save money by using an additional, high-voltage coupling capacitor before the tone stack (having a value at least twice that of C2), so that C1, C2 and C3 can be rated only for the AC voltage.

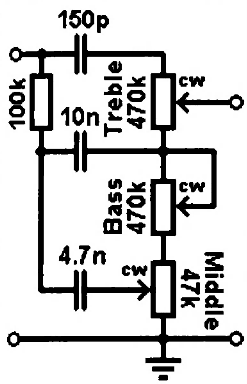


Fig. 10.35: The completed design.

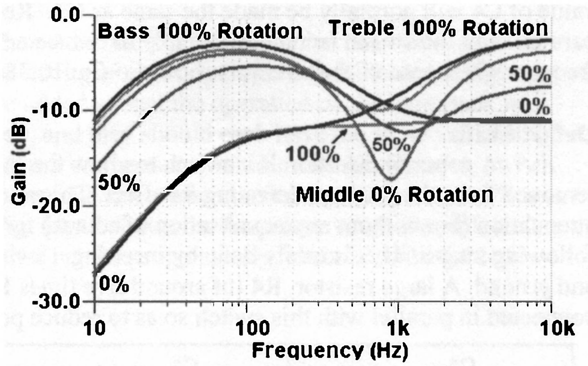


Fig. 10.36: Frequency response of the FMV tone stack in fig. 10.35.

Expanding the FMV tone stack:

Although the FMV tone stack is universally popular, it does not suit all tastes. In more advanced amplifiers, switchable values and additional controls may be provided; just a few of the more useful options are given here.

Ultra-low switch:

By switching in a capacitor, C4, between the bottom of the slope resistor and ground (fig. 10.37), treble frequencies are permanently attenuated before reaching the middle control. This makes the treble and middle controls less interactive, increases the inherent mid-scoop and shifting the middle response down in frequency somewhat (although it robs the middle control of some of its range too). This gives the impression of a low-bass boost. This feature appeared on some later Fender *Bassman* models and was labeled the 'deep' switch. The

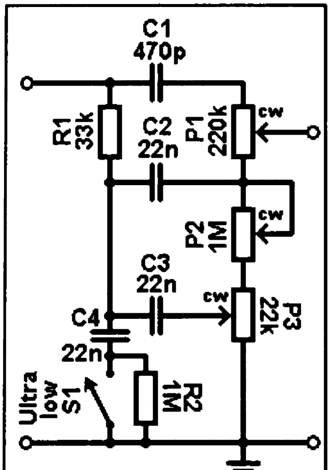


Fig. 10.37: An additional capacitor, C4, shifts the middle response down in frequency, giving the illusion of a bass boost.

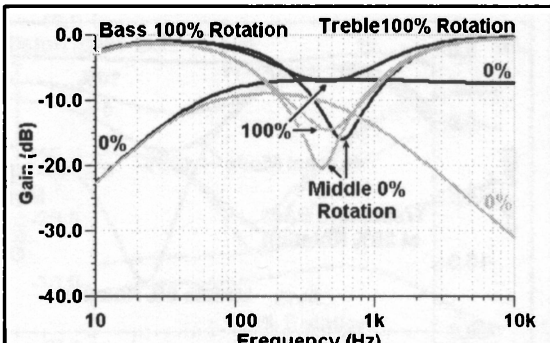


Fig. 10.38: Frequency response of the FMV tone stack in fig. 10.37. Bold traces are the 'normal' response while the faint traces show the response with the 'ultra low' switch closed.

value of C4 will normally be made the same as C3. Resistor R2, connected in parallel with the switch is included simply to reduce any popping sounds. The frequency response of the circuit is shown in fig. 10.38.

Defeat / Lift:

A common addition is a switch to allow the tone stack to be effectively removed from the preamp, defeating its effect. This allows the signal to pass without attenuation (but without any equalisation of course) for more gain / overdrive in following stages. This is easily done by inserting a switch between the bottom of P3 and ground. A large resistor, R4 (of around five times R1 or more) should be connected in parallel with this switch so as to reduce popping sounds and to retain a

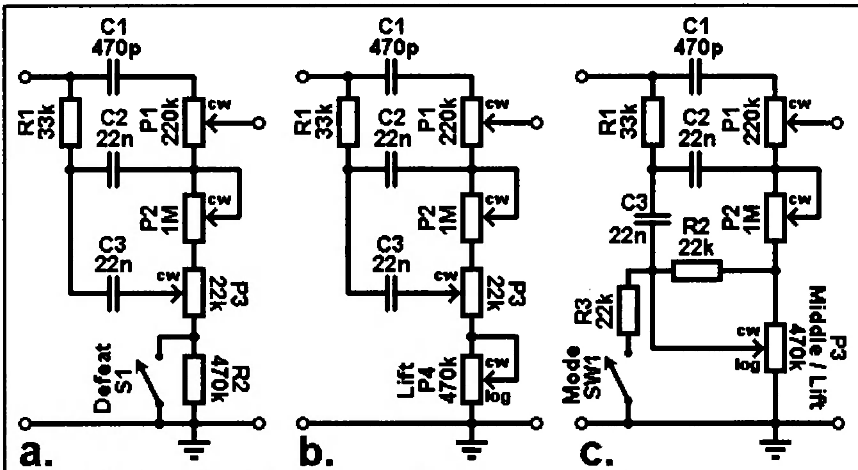


Fig. 10.39: The circuit in **a.** shows a defeat switch, which allows the effect of the tone stack to be removed, for maximum gain The arrangement in **b.** makes this variable. In the circuit in **c.** both 'lift' and middle control are available from a single potentiometer.

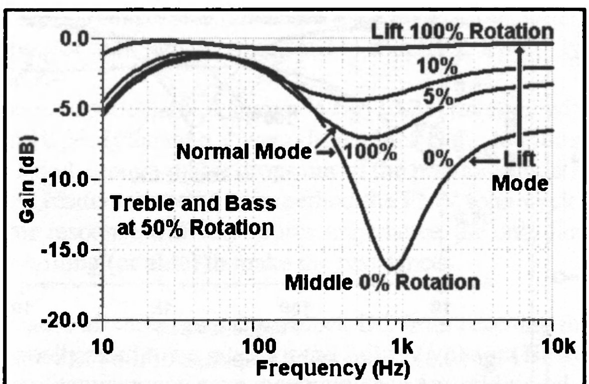


Fig. 10.40: Frequency response of the circuit in fig. 10.39*c*, illustrating the 'normal' and 'lift' modes.

proper grid-leak path for any following valve. This is shown in fig. 10.39*a*. Of course, replacing this extra resistor with a potentiometer, as in fig. 10.39*b*, allows the tone stack to be progressively 'elevated' of 'lifted', so the degree of equalisation can be adjusted to taste. At maximum setting the tone controls are again completely defeated. The same idea can be applied

to any tone stack of course.
If front-panel space is limited then the middle control, P3, can also serve as a lift control if a switch is added to select its mode of operation, as indicated in fig. 10.39*c*. Resistors R2 and R3 are added so that the operation of the tone stack is unchanged when in normal mode, and they should each have the same value as the 'normal' value of P3 when used as a conventional middle control. Opening switch SW1 allows P3 to fully elevate the tone stack when turned to maximum, so P3 acts as a lift control in this mode. Fig. 10.40 shows the frequency response plot, and illustrates how the range of control can be progressively 'shrunk', until the tone stack is completely defeated.

Shift Control:

Although the circuit is normally 'designed around' a chosen value of slope resistor, once completed the slope resistor can be used to shift the response envelope. The shift is not as uniform as for the James stack because as the resistance of R1 is increased bass and middle frequencies are attenuated more, so the range treble control becomes greater at the expense of bass / middle control. However, since the 'traditional' arrangement is fairly bass-heavy anyway, this can actually be very useful. Reducing the value of R1 has the opposite effect, and is normally avoided. Fig. 10.41 shows such a shift control, P4, added to a conventional circuit. Because it is necessary to keep DC off this potentiometer an additional coupling capacitor may be required (C4 in this case), having a value of at least twice C2.
The effect on the response envelope is shown by the frequency response plots in fig. 10.42.

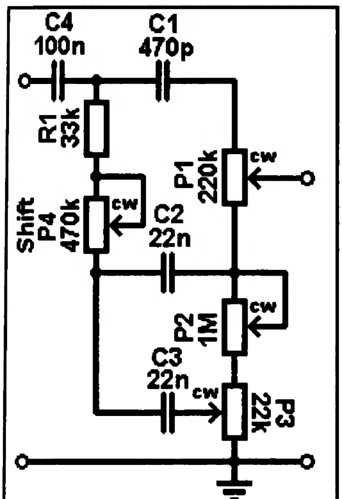


Fig. 10.41: Increasing the value of R1 shifts the response envelope down in frequency.

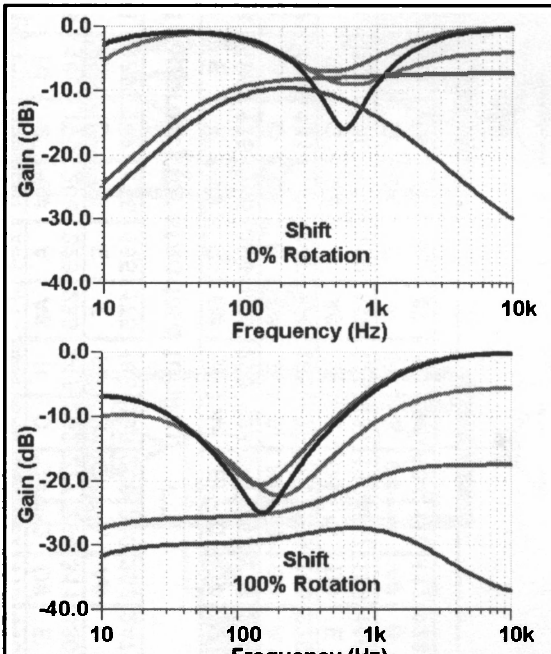


Fig. 10.42: Frequency response of the FMV tone stack with shift control as shown in fig. 10.41.

Guitar

OPEN	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20
E 330	F 349	F# 370	G 392	G# 415	A 440	A# 466	B 494	C 523	C# 554	D 587	D# 622	E 659	F 698	F# 740	G 784	G# 831	A 880	A# 932	B 988	C 1047
B 247	C 262	C# 277	D 294	D# 311	E 330	F 349	F# 370	G 392	G# 415	A 440	A# 466	B 494	C 523	C# 554	D 587	D# 622	E 659	F 698	F# 740	G 784
G 196	G# 208	A 220	A# 233	B 247	C 262	C# 277	D 294	D# 311	E 330	F 349	F# 370	G 392	G# 415	A 440	A# 466	B 494	C 523	C# 554	D 587	D# 622
D 147	D# 156	E 165	F 175	F# 185	G 196	G# 208	A 220	A# 233	B 247	C 262	C# 277	D 294	D# 311	E 330	F 349	F# 370	G 392	G# 415	A 440	A# 466
A 110	A# 117	B 123	C 131	C# 139	D 147	D# 156	E 165	F 175	F# 185	G 196	G# 208	A 220	A# 233	B 247	C 262	C# 277	D 294	D# 311	E 330	F 349
E 82	F 87	F# 92	G 98	G# 104	A 110	A# 117	B 123	C 131	C# 139	D 147	D# 156	E 165	F 175	F# 185	G 196	G# 208	A 220	A# 233	B 247	C 262

Note and frequency in Hertz (approx.) for standard tuning.

Bass

G 98	G# 104	A 110	A# 117	B 123	C 131	C# 139	D 147	D# 156	E 165	F 175	F# 185	G 196	G# 208	A 220	A# 233	B 247	C 262	C# 277	D 294	D# 311
D 73	D# 78	E 82	F 87	F# 92	G 98	G# 104	A 110	A# 117	B 123	C 131	C# 139	D 147	D# 156	E 165	F 175	F# 185	G 196	G# 208	A 220	A# 233
A 55	A# 58	B 62	C 65	C# 69	D 73	D# 78	E 82	F 87	F# 92	G 98	G# 104	A 110	A# 117	B 123	C 131	C# 139	D 147	D# 156	E 165	F 175
E 41	F 44	F# 46	G 49	G# 52	A 55	A# 58	B 62	C 65	C# 69	D 73	D# 78	E 82	F 87	F# 92	G 98	G# 104	A 110	A# 117	B 123	C 131

Chapter 11: Effects Loops

Line levels. Send, line out and loop drivers. Recovery stages and serial effects loops. Parallel effects loops.

An effects loop is simply a break in an amplifier's signal path, into which external effects may be effectively inserted (see fig. 11.1*a*). The loop is usually placed somewhere towards the end of the preamp, so that most of the basic tonal shaping and overdrive has already been generated by the preamp, before being sent to the effects unit. After the effects have been added, the signal is returned to the amplifier for final power amplification. Of course, the signal being 'sent' does not always have to be returned, it may instead feed a mixing desk, recording device, P.A. or some other slave amp. Consequently, 'line out' functions are included in this chapter.

The demand for amplifiers to include effects loops is quite strong now, though is still a matter of some small controversy. There are those who would argue that the tone of a good valve amplifier should not be 'spoilt' by the addition of tasteless solid-state effects, or that a loop which is *not* being used must be to the detriment of the 'raw' amplifier tone since the signal will surely pass through the unwanted circuitry. Certainly, some effects loops found in commercial amplifiers (mainly earlier ones from the 1970s and 1980s) are extremely poorly designed, giving noise, loss of treble or lack of drive for the power output stage, and may have led to discontent among some players (who might blame the results on the effects themselves).
And then there are the more adventurous musicians who find they cannot live without the ability to layer effects on top of the ordinary amplifier's sound.

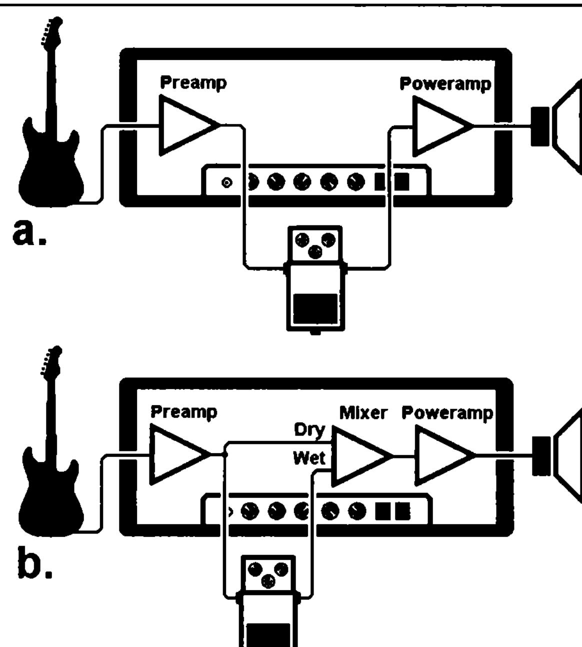


Fig. 11.1: Functional diagrams of **a.** a serial effects loop and **b.** a parallel effects loop.

Effects loops fall into one of two categories: **serial** or **parallel**. Basic functional diagrams for both are shown in fig. 11.1. In the serial loop in fig. 11.1*a*. the amplifier's signal chain is completely broken at some point, and the effect is inserted in between. Electronically this is very simple to arrange, and the very earliest of amplifiers

offering loops were of this type. However, without sufficient care over design the results may be disappointing, as the basic tone of the amplifier will be completely changed by the insertion of the effects, which operate at a much lower signal level than the valves.
Fig. 11.1*b* shows a parallel loop (or 'side chain'). This is a more advanced arrangement in which the main amplifier signal chain is not completely broken. The signal returning from the effect, known as the **wet** signal, is mixed with the unaffected or **dry** amplifier signal. This helps to preserve the basic tone of the amp and also allows effects to be added more subtly. The degree of wet-to-dry mixing is often made adjustable by some means. A parallel loop also has the small advantage that if the effects unit or interconnecting cables were to fail, the amplifier would not be completely silenced.

There is considerable room for variation as to where certain effects are placed in the signal chain. Equalisers (tone controls) tend to perform better when used *before* distortion is added (e.g., between the guitar and amplifier), so that they can manipulate the relative proportions of frequencies being distorted, rather than simply altering the frequency response of an already overdriven sound, although good results can be obtained either way. Compressors, volume/expression pedals and boosters are also normally placed before the input of the amp. Wah-wah effects are *usually* placed before distortion is added, but this is by no means universal. Delay effects such as echo, reverb, chorus and flangers tend to be best placed *after* most of the distortion has already been generated. Applying them before adding distortion will distort the delay effect too, and can result in a rather messy 'noise', if it is done carelessly, and it is these effects that are most likely to be used in an effects loop. The same considerations apply to phasers, tremolo and vibrato effects. Of course, all musicians will have their own preferred arrangement, from subtle to garish.

Obviously, the sort of amplifier most likely to benefit from an effects loop is one in which most of the distortion is generated in the preamp, which is the case for modern high-gain designs, so that the loop may be inserted after the preamp but before the power output stage, which is not significantly overdriven itself. In lower-gain amplifiers of more traditional design the output stage may be heavily overdriven, in which case there is not much to be gained by having an effects loop since the effects themselves will be distorted no matter where they are inserted. Clean amplifiers are the least likely to need an effects loop since there is little difference in using the effects before or after the preamplifier if the raw guitar sound has barely changed (and clean players are generally the sort to use very few effects, of course).
It is often said that an effects loop should be designed to be 'transparent', so that it does not impinge on the fundamental tone of the amp. This is not a very practical statement where valves are concerned, since the signal returning from the effects unit must inevitably be re-amplified to a level commensurate with the signals levels used in the amp, and it can be difficult (though not impossible) to incorporate this gain stage without affecting the 'fundamental' tone of the amp. It is better to

request that the effects loop should be a *part* of the main amplifier, 'designed in' as it were, so that the loop in a fully integrated piece of circuitry that also contributes to the raw sound of the amp. In this way the player will not be aware of any significant changes in the basic tone of the amp, whether the loop is in use or not.

Line levels:

Interestingly, the design of 'valve driven effects loops' seems to be regarded as something of a highly skilled, 'black-art', perhaps because there are no classic, old designs to copy them from. In reality they are exceedingly simple, the problem really lies in deciding what sort of devices we are trying to cater for.

Often we find the term **line level** used in audio, although its meaning has changed over the years. It refers to the nominal (which is effectively the maximum average) signal level being sent down a 'line', that is, down a cable between two devices.* The 'line level' is measured relative to a reference level, which is either **decibel volts, dBV**, or **decibels unloaded, dBu**.

Decibel volts are used for consumer electronics, into which category the guitar amplifier falls, and has a reference level of 1mW into 1kΩ, or 1Vrms (2.8Vp-p). Signals at this level are then called 0dBv.

Decibels unloaded is used for professional purposes (though not in the sense of 'professional musician') and has a reference level of 1mW into 600Ω, or 0.77Vrms (2.2Vp-p). Signals at this level are then called 0dBu.

The actual line level in question is measured relative to either of these references according to:

$$\text{Line Level (dB)} = 20.\log\frac{\text{Actual rms voltage}}{\text{Reference rms voltage}}$$

Thus, for a consumer product, a line level of 316mVrms would be described as:

$$\text{Line Level (dBV)} = 20.\log\frac{\text{Actual rms voltage}}{1\text{Vrms}} = 20.\log\frac{0.316}{1}$$

=-10dBV

And this is a very common level used in consumer hifi and music products, and is not dissimilar from the output level of a humbucker.

Effects pedals (stomp boxes) are designed to receive **instrument level** signals. Actual instrument levels are enormously variable of course, but a line level of −10dBV , or simply 1Vp-p (350mVrms), may be regarded as the most that is required. Higher levels may cause some pedals to clip. The output from such pedals will be of similar order, and since nearly all pedals operate from a 9V supply the

* The EIA defines 'line level' as any signal over 25mVrms, although this is not very revealing where guitar amps are concerned.

maximum output rarely exceeds 5Vp-p (they usually have their own output level control of course).
The 'line level' inputs and outputs on P.A, mixing desks and rack-mount equipment are usually designed to handle 0dBV level input signals. CD players generally produce an output level of +4dBV. Microphones, on the other hand, produce very low-level signals in the region of -50dBV (3mVrms), although condenser microphones using phantom power may boost this to 0dB or more. Additionally, P.A., mixers and recording equipment may only use balanced XLR and unbalanced RCA phono sockets, whereas guitar-orientated devices almost exclusively use only 6.35mm (1.4") mono jack sockets.
From the hobbyist's perspective it is not practical to cater for all of these interfacing possibilities, so most of this chapter will assume that the effects loop is intended for use with ordinary guitar effects pedals, handling signals levels of about −10dBV.

Send, line out and loop drivers:

Perhaps the most important aspect in the design of and effects loop is the output impedance of the **line out** or **send** section. If the output impedance is high then it will be easier for spurious EM noise to couple into the cable used for sending the audio signal, particularly since we are unlikely to be using balanced (microphone) cables. Fluorescent lights are particularly notorious for causing mains frequency hum. The lower the output impedance, the better this environmental noise is shunted to ground. Also, the input impedance of many pedals is depressingly low, sometimes only 47kΩ, so if the effects loop has a high source impedance then there may be an unexpected loss in level when such pedals are used, compared with high input impedance pedals. What's more, some pedals, particularly older designs using discrete transistors, have significant input (Miller) capacitance. This often causes noticeable loss of treble if the source impedance is rather high, which dulls the tone, and is probably the most common complaint heard about effects loops. As a general rule, we would like the output impedance of the 'send' section to be not greater than about 10kΩ, preferably much less. This simple requirement is often overlooked even in commercial amplifiers, and leads to considerable dissatisfaction among users.

Simple line-out arrangements:

There is a wide range of ways we might provide a low-level, relatively low output impedance signal, in a conventional valve amp. The simplest of these is a passive potential divider. This can be placed at any convenient point in the preamp, and fig. 11.2 shows it placed immediately after a gain stage. We need to consider the signal level at this point. The maximum, peak-to-peak output signal swing of a triode is around 2/3HT. Most amplifiers operate the preamp stages from a supply voltage

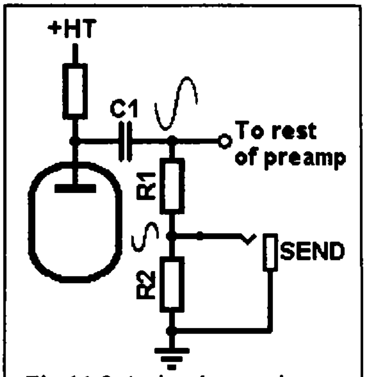


Fig 11.2 A simple, passive line-out arrangement.

in the region of 250V to 400V, so assuming the triode is fully driven (or overdriven) we should expect a signal level of around 165V to 260Vp-p, or about 60V to 90Vrms. More accurate values, for a particular circuit, could be derived from a load line of course.

A common mistake made by beginners is to simply provide a 1MΩ potentiometer for the line out, but this obviously suffers from the problem that its output impedance is very high (except at very low settings) and if turned up, the effects unit could be damaged by the very high signal voltage levels, so a dropping resistor is required.

Since we demand an output impedance of not more than 10kΩ then the lower resistor in the divider should be around 10kΩ, and this will normally be a 'send level' potentiometer. The value of the upper resistor depends on our desired maximum output line level. Although effects pedals only require about -10dBV, we cannot guarantee that the triode will always be fully driven. It would be better to provide an output level of 0dBV when the triode is fully driven, so that we have sufficient range still available if it is *not* fully driven. Taking the expected signal level produced by the triode to be 60Vrms, and remembering that 0dBV is 1Vrms, the upper resistor must be:

$$V_{out} = V_{in}.\frac{R2}{R1+R2}$$

So:

$$R1 = \frac{V_{in}R2}{V_{out}} - R2 = \frac{60 \times 10k}{1} - 10k$$

$=590k\Omega$

So we might choose a standard value of 560kΩ. However, this may be a rather heavy load to place on the triode (if it uses a fairly large anode resistor), so the next highest standard of 680kΩ might be a reasonable compromise. Even a common value of 1MΩ would probably provide sufficient output level for most purposes, without loading the triode significantly.

The output impedance (ignoring the source impedance of the triode itself) is simply:

$$Z_{out} = R1 \parallel R2 = \frac{680k \times 10k}{680k + 10k}$$

$= 9.9k\Omega$

The final circuit is shown in fig. 11.3, and since the divider is placed after the usual coupling capacitor, C1, no additional coupling capacitor is required. This type of line out is easily fitted to an existing amplifier, where a more complex, active output buffer is undesirable or difficult to install.

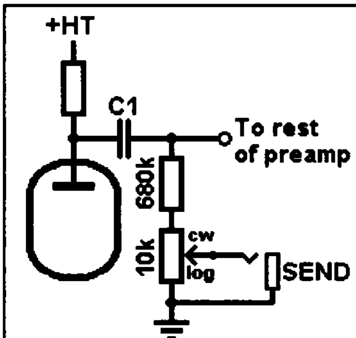


Fig. 11.3: A simple line-out arrangement with level control.

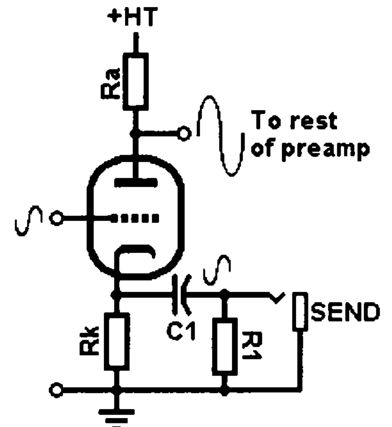


Fig. 11.4: A simple, buffered line out.

Another convenient place to take a low impedance line-out signal from, is the cathode of a conventional gain stage, provided the cathode resistor is unbypassed of course. This is shown in fig. 11.4.
The stage resembles a cathodyne phase inverter, except that the anode and cathode resistors are simply those of a normal gain stage. Since we take advantage of cathode-follower operation this has the obvious advantage that the output impedance is very low indeed, being given by:

$$Zout = \frac{Ra + r_a}{\mu + 1} \parallel Rk$$

For a typical ECC83 gain stage with $Ra = 100k\Omega$, $Rk = 1k\Omega$, $r_a = 65k\Omega$ and $\mu = 100$, this yields an output impedance of only:

$$Zout = \frac{100k + 65k}{100 + 1} \parallel 1k$$

$=620\Omega$.

What's more, the output signal swing at the cathode is slightly less than that appearing at the grid which, assuming Rk is of a conventional value between about 470Ω and 4.7kΩ say, is itself limited to the input sensitivity of the valve being used. If the output level from the anode is known (which is easily derived from a load line) then the corresponding signal level at the cathode is easily found using:

$$vk = Rk.\frac{va}{Ra}$$

Where:
vk = AC signal level at the cathode.
va = AC signal level at the anode.
Rk = the cathode resistor.
Ra = the anode resistor.

This is only a couple of volts for most preamp valves, so the signal voltage at the cathode is already at a convenient level for line out purposes and cannot become large enough to damage any solid-state effects that might be connected. R1 is included simply to allow the coupling capacitor, C1, to discharge when not in use, so that no loud 'pops' occur when plugging into the send socket. It could be replaced by a send-level pot' of course, having a value of around 4.7kΩ to 10kΩ. Assuming the input impedance of the effects unit being used is much greater than R1, then the –3dB low roll-off frequency caused by the output coupling capacitor, C1, is given by:

$$f_{-3dB} = \frac{1}{2\pi C1(Zout + R1)}$$

If we wish to pass all audio frequencies we might choose a frequency of 1Hz. If we choose R1 to be 4.7kΩ, then rearranging the above gives:

$$C1 = \frac{1}{2\pi f(Zout + R1)} = \frac{1}{2\pi \times 1 \times (0.62k + 4.7k)}$$

$= 29.9\mu F$

So we could take the nearest standard of 22μF, or some convenient larger value such as 47μF. Smaller values could be used, but they will cause the output impedance to rise at lower frequencies, which will reduce the circuit's immunity to low-frequency noise.

A further modification which could be made is to use a switched jack socket so that, when nothing is plugged into the socket, C1 is grounded and therefore acts as the cathode bypass capacitor.

Of course, this means that when the loop is not in use the gain of the valve will be higher then when the loop *is* used. This loss in gain can be useful when designing parallel effects loops since it automatically reduces the level of the dry signal somewhat, making it a little easier to mix with the low level, wet signal returning from the effects unit. Obviously the tone of the preamp will be different depending on whether the loop is in use or not, which could be seen as a shortcoming, or could be regarded as a useful 'voicing' feature.

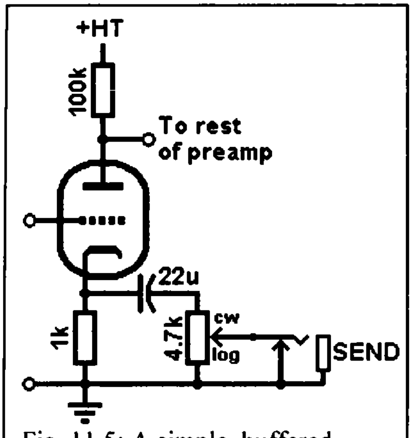


Fig. 11.5: A simple, buffered line-out in which the potentiometer acts both as an output level control, and as a gain control when the loop is not in use.

If R1 is a potentiometer as in fig. 11.5, however, then it functions as a send-level control, *or* as a gain control when the loop is not in use, as was described in chapter 1, fig. 1.20. This version of the circuit is easily added to an existing amplifier, since a pre-existing cathode bypass capacitor can function as the output coupling capacitor, provided it is large enough. With this arrangement the original tone of the amp is preserved as long as the loop is not in use and the pot' is turned to maximum gain.

A slight disadvantage to this arrangement is that when the valve is overdriven, grid current flows down the cathode and therefore contributes to the cathode signal. Hence the signal at the cathode will not appear clipped in exactly the same way as the signal found at the anode. Fortunately the small difference is unlikely to be a serious problem for most players.

To illustrate some of the many ways in which a simple line-out can be added to an existing circuit, fig. 11.6*a* shows a simple three-stage preamp of typical topology. Maximum expected rms signal voltages are shown. Fig. 11.6*b* shows five possible means of taking a low-impedance line-out at approximately 0dBV level, from the same preamp.

1. 'Send 1' is the same potential divider described earlier, giving a maximum output impedance of approximately 10kΩ and a *maximum* output level of 1Vrms. However, the addition of this divider will load down the first triode somewhat, which might cause too much tonal change for some purposes.
2. 'Send 2' applies the same divider principle as 'Send 1' except that the pre-existing 1MΩ gain control now functions as a convenient dropping resistance, thereby avoiding unnecessary loading of the first triode. Again,

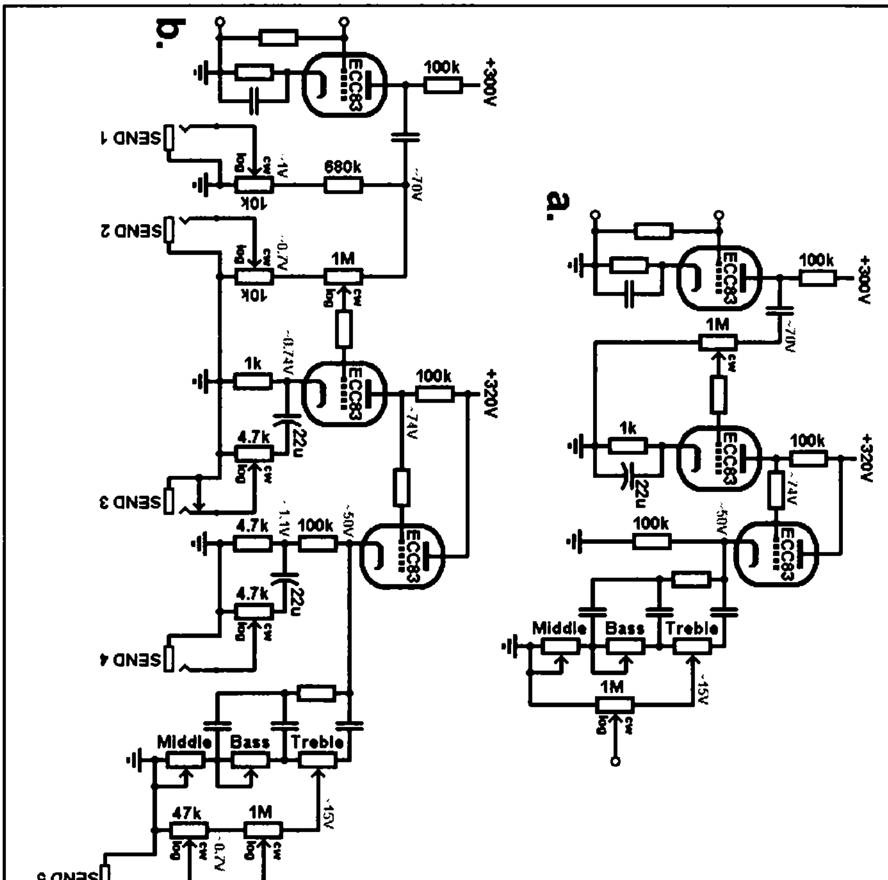


Fig. 11.6: **a.** A typical section of a preamp, and **b.** several ways in which a line-level output could be derived. AC signal voltages shown are approximate *maximum* rms values.

the maximum output impedance is roughly 10kΩ, though the output level is slightly lower at 0.7Vrms, which is still sufficient for conventional pedals. However, the addition of the 10kΩ output level potentiometer now prevents the original gain control from being turned fully to zero (see chapter 2, fig. 2.16), but most players have no need of a completely silent amp anyway!

3. 'Send 3' is the simple buffered output described earlier, and makes use of the pre-existing cathode bypass capacitor to act as the line output coupling capacitor. The switching jack socket ensures that the original tone of the amp can be retained, when nothing is plugged into the socket, by turning the potentiometer to maximum.
4. 'Send 4' takes advantage of the cathode follower's load resistance to act as a dropping resistance. A 4.7kΩ resistor is added (which is too small compared with the 100kΩ load to affect the original tone) and the output signal is tapped off the junction. However, because DC flows in the cathode load resistance an additional output coupling capacitor is required (a high-voltage capacitor is not required, of course). The maximum output impedance is only 4.7kΩ||4.7kΩ||100kΩ = 2.3kΩ while the maximum output level is roughly 0dBV. This is very similar to the effects loop driver used in the Soldano *SLO100*.
5. 'Send 5' is a variation on 'Send 2' but takes the output after the tone stack. In this position the maximum signal level is reduced, due to the heavy attenuation caused by the tone stack. Of course, this varies depending on the values used in the tone stake, the setting of the tone controls and the reference frequency. The signal levels shown in fig. 11.6 assume a frequency of 1kHz with all the tone controls in the middle position. Consequently, the output level potentiometer must be larger to achieve sufficient output level which, unfortunately, also raises the maximum output impedance to roughly 47kΩ.

Any of these line-out designs can be added, without any changes required, to most existing guitar amps assuming the HT is within the normal range of 250V to 400V. It is usually desirable to take the line-out from a point near the end of the preamp, of course, so as to capture as much of the preamp's character in the send signal as possible.

Active loop drivers:

The simple line-out arrangements discussed previously are useful for providing a line-level output to an amp, without great cost. However, if we wish to take an output signal from *anywhere* in an amp, particularly the difficult position immediately after a tone stack, without affecting the original tone, then a dedicated buffer stage is required. Also, a modern effects loop will normally require a recovery stage to re-amplify the return signal, and since this will normally use one triode this leaves another triode available (if a dual-triode valve is used), so it makes sense to use the extra triode as the send buffer.

The loop driver/buffer should offer a low output impedance and a high input impedance, so that it does not load down the existing preamp stages in any way, so a cathode follower is the obvious solution. An active stage such as a cathode follower also has the advantage that it will usually be capable of driving the required line level into relatively heavy loads. It should be pointed out that the simple circuits described earlier, although offering relatively low output impedance to keep noise low and treble response good, will not drive equally low input impedances without heavy attenuation of the signal. This is not a great concern for conventional effects pedals since they have input impedances of at least 47kΩ, but some mixing desks, P.A. amplifiers and computer sounds cards may have input impedances as low as 10kΩ. A valve with heavy local feedback could offer a low output impedance but, again, is not capable of maintaining this performance into a heavy load, so a cathode follower is the only logical option.

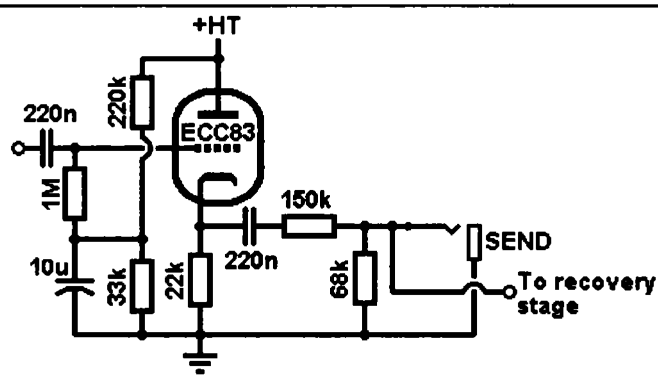


Fig. 11.7: The effects-loop driver circuit used in the Orange *Rockerverb* has a high output impedance despite using a cathode follower, and is a poor design.

The design of cathode followers was explained in chapter 5, so only a few important design requirements need to be discussed here. A common mistake made by amateurs and professionals alike, is to design a perfectly good cathode follower but to then follow it with a large dropping resistance or potentiometer to bring the output down to line level. For example, fig 11.7 shows the loop driver used in the Orange *Rockerverb*. The output impedance of the cathode follower itself is a respectable 620Ω, but it is followed by a potential divider using high-value resistors which raise the output impedance of the whole circuit to a poor $(620\Omega+150k\Omega)||68k\Omega = 47k\Omega$. The input impedance of the cathode follower is also not as high as it could have been if cathode-biasing had been used [see chapter 5, fig. 5.5]. Thus all the useful characteristics of a cathode follower have been wasted. The same mistake is made in the Mesa Boogie *Dual Rectifier*, in which the cathode-follower, loop driver is followed by a 100kΩ send-level potentiometer, which raises the output impedance to tends of kilo-ohms (unless the pot' is always at maximum), again making the cathode follower redundant.

Another common mistake is to use an output coupling capacitor which is too small, meaning that the output impedance rises considerably at low frequencies due to the capacitor's rising reactance. This is very undesirable since most noise interference is low in frequency (particularly mains frequency), so we must maintain a low output impedance down to very low frequencies so as to maintain good noise immunity. This error is made in the Soldando *SLO100* which has an output impedance of 2.2kΩ at mid' frequencies, but this more than triples at mains frequency and below, due to a small output coupling capacitor.

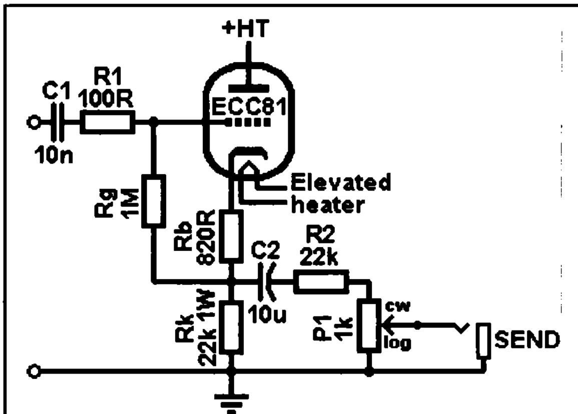


Fig. 11.8: A high-quality effects-loop driver.

If the cathode follower is to be expected to drive heavy loads, then cathode load resistor must be lowest value possible, and the quiescent current as high as is practical. This also results in the lowest output impedance, of course. If it is to be DC coupled to a preamp gain stage then there is a limit to how small we can make the cathode resistor before quiescent grid current will flow, heavily loading the previous stage. If the cathode follower is AC coupled, however, then the cathode load resistor can be made relatively small as in fig. 11.7. However, the smaller we make the cathode resistor, the greater is the chance that the cathode follower itself may be overdriven, since the maximum peak-to-peak output swing is given by formula XXXIV:

$$V_{O(max\ p\text{-}p)} = \frac{HT}{1 + r_a / Rk}$$

The best option is therefore to use a low r_a valve, such as an ECC81 / 12AT7 or ECC82 / 12AU7 which naturally require a low value of load resistor, while retaining good headroom.

For example, fig. 11.8 shows an ECC81 cathode-follower driver circuit capable of driving loads as low as 10kΩ with no discernable loss of signal level. To obtain the maximum headroom it is roughly centre biased, and the input impedance is over 9MΩ! The output impedance is roughly: R2||P1 = 22kΩ||1kΩ = 960Ω. R2 attenuates the maximum signal level to around +5dBV, and should suit most existing preamps. Unfortunately, the price which must be paid for such a high-quality driver is that it requires between 3mA and 5mA from the power supply, depending on the voltage used (up to 400V) and the output coupling capacitor, C2, must be rated for at least 2/3HT. $Vhk_{(max)}$ for the ECC81 is 90V, so the heater should be elevated by about 50V.

An ECC82 or ECC83 could also be used in this circuit, though Rb should be increased to 1.5kΩ or 1kΩ respectively, for maximum headroom.

Although the circuit in fig. 11.8 gives excellent performance –better than most commercial loop drivers– the full drive capabilities of the cathode follower are still not realised, as they are still hampered by the necessary dropping resistor and output level potentiometer. However, in many cases an extremely high input impedance is not required, particularly if the preamp is being designed from scratch. In such cases we could move the signal-attenuation components to the *input* side of the circuit, allowing the cathode follower to drive the loop directly, as shown in fig. 11.9.

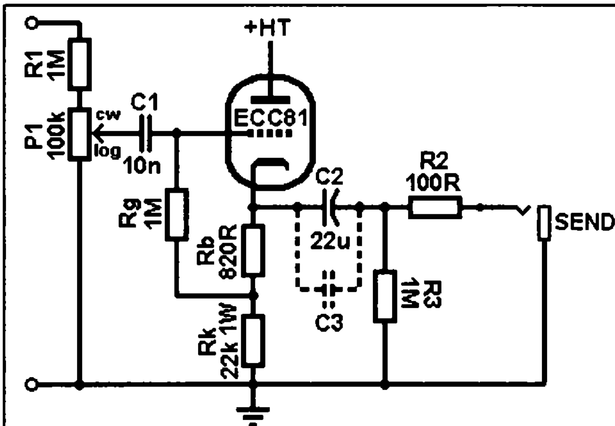


Fig. 11.9: Moving the signal-attenuation components to the input side allows the cathode follower to realise its full drive capabilities.

The attenuating resistor, R1, and send-level potentiometer, P1, are now placed at the input of the circuit. The value of these components could be chosen to suit any signal level of course. A build-out resistor, R2, isolates the follower from cable capacitances while R3 is included simply to allow the output coupling capacitor C2 to discharge, to avoid popping sounds.
The output impedance of the circuit is roughly 670Ω, and because it may be expected to drive input impedances as low as a few kilo-ohms the output coupling capacitor must be fairly large in order to maintain this output impedance at very low frequencies (again it must be rated for at least 2/3HT). This demands an electrolytic capacitor, which is unfortunate. Therefore an optional capacitor, C3, is shown dashed. The purpose of this component is to compensate for any shortcomings in the electrolytic capacitor at higher frequencies and so reduce capacitor distortion. C3 should therefore be a good poly' type; its value is not critical, and 100nF or greater would be suitable. The same could be applied to fig. 11.8 of course. For guitar purposes, however, this component is not essential.

In a studio environment it is sometimes desirable to isolate the amplifier's ground from other equipment, to eliminate any possible ground-loop hum. If this is the case then it will be necessary to use an isolating transformer. A ground-lift switch may also be included, as shown in fig. 11.10. Coupling capacitor C2 is still required to keep DC from flowing in the transformer's primary, which would otherwise cause saturation of the core. This arrangement is known as **parallel feed**[1] or simply **parafeed**.
The primary impedance of the transformer should be high enough that it does not load down the cathode follower, and 10kΩ is a common standard. The secondary impedance depends somewhat on the expected input impedance of the device we

[1]Langford-Smith, F. (1957). *Radio Designer's Handbook* (4th ed.), p 350. Illife & Sons, Ltd., London.

are driving. If the circuit is only expected to drive high-impedance loads then a secondary impedance of 10kΩ might be appropriate, though transformers with impedance ratios closer to 10kΩ:600Ω are more commonly available. The formula for turns ratio was given on page 228:

$$n = \sqrt{\frac{Z_{pri}}{Z_{sec}}} = \sqrt{\frac{10k}{0.6k}}$$

$= 4$

For a 0dBVlevel output, the input to the transformer must therefore by four times greater, or +12dBV, and this may need to be taken into account when selecting R1 and P1. As shown, the circuit requires roughly 50Vrms input for 0dBV output with P1 turned to maximum.

It is important that the load impedance seen by the secondary of the transformer is not much higher than the expected value, otherwise the increased reflected impedance to the primary is liable to cause significant bass attenuation, unless the primary inductance is extremely high (which is unlikely). Therefore, a permanent load resistor, R2, has been added to the circuit in fig. 11.10, which should ensure that the reflected impedance to the primary is about right, even when using a high input impedance device. A 50mW transformer is more than sufficient for the circuit shown. If the transformer has a centre-tapped secondary then a balanced output could be provided, for the greatest noise rejection, and the connections for an XLR socket are shown. Be aware that because of the inductive load, a grid-stopper MUST be used, to avoid any chance of oscillation, and this role is served by R1 provided it is not positioned far from the valve itself. Fortunately, an isolated line-out is beyond the requirements of most players.

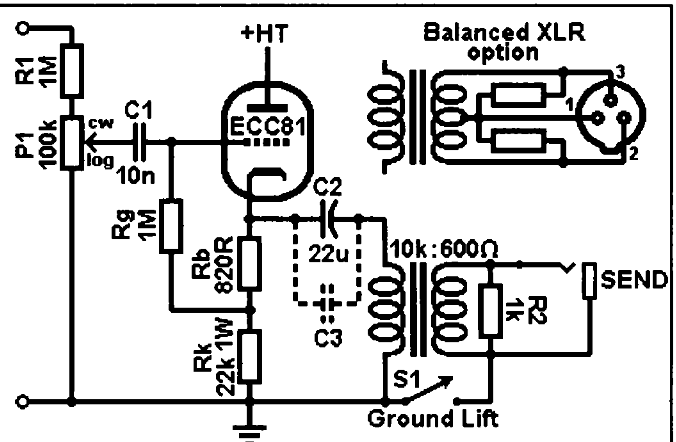


Fig. 11.10: Parafeed transformer coupling allows complete isolation of the amplifier and external equipment circuit grounds.

Recovery stages and serial effects loops:

In some cases we may only want a line-out function for the preamp, but if we are designing a complete loop then we must provide means for the return signal to be injected back into the amplifier. If we are designing a serial loop then we must ensure that when the loop is *not* in use, the amplifier signal chain is unbroken.

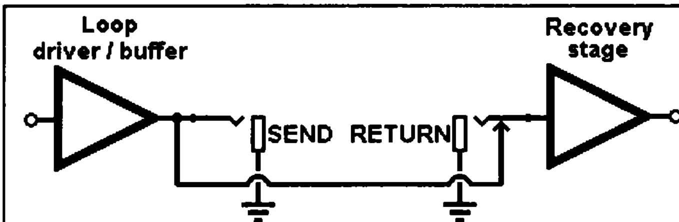


Fig. 11.11: Functional diagram of a serial effects loop.

The universal solution to this is to use a switched jack socket for the return input, as shown in fig. 11.11. If the loop is not in use, or if a cable is plugged only into the send jack, the amplifier signal chain is unbroken since the output of the loop driver is connected to the switch on the return jack socket. If a cable is plugged into the return jack, the amplifier signal chain is broken by the switch. This also allows instruments and equipment to be plugged directly into the recovery stage and power amp, bypassing the preamp altogether, if desired. This is useful if the amplifier is being used as a make-shift P.A. amp.

In a solid-state amp there may be no need for dedicated driver and recovery stages since the impedances and signal levels in the amplifier are of similar order to those used by effects pedals. This is not true for valve amps. The 'wet' signal returning from the effects pedals will normally be at a much lower level than the normal 'dry' signal present in the amp when the loop is not used, so an additional gain stage is usually required to boost or 'recover' the wet signal. Without this stage it will be difficult to drive the power amp to full output when the loop is used, and is one of the reasons why early effects loops acquired a bad reputation.

Passive serial loops:

The only difficulty with designing effects loops is in obtaining broadly the same tone from the amp whether the loop is used or not, as this is what most players expect. If, say, a reverb pedal is used in the loop, most players expect this effect to be layered 'on top' of the normal tone of amp, and not to obtain some wildly different tone. This is why adding a loop to an existing amp can be troublesome. (Of course, if the signal is heavily distorted *after* the effect has already been added then any sound could result, and there is nothing the designer can do to alter this.)

For example, fig. 11.12 shows a section of a conventional preamp, into which a passive effects loop has been inserted (i.e., it has no dedicated driver or recovery stage), indicated inside the dashed box. The first remark should be that the output impedance of the send jack is high because the pre-existing 1MΩ volume control doubles as the send level control, so the loop has almost no noise immunity and might cause treble attenuation with some pedals. The second remark is that because there is no recovery stage we probably won't be able to drive the output stage to full power if the loop is used, since the wet signal being delivered to the phase inverter will probably be only a few hundred millivolts in level, rather than several volts. (Of course, this does depend on how sensitive the power valves are). On the other hand, an advantage to this arrangement is that the original tone of the amp is almost completely unaffected (the 3.3MΩ discharge resistor should have minimal

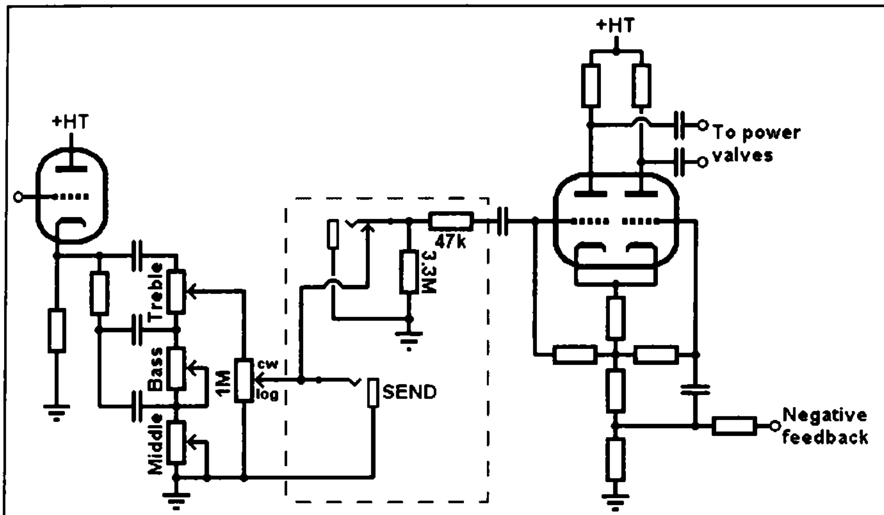


Fig. 11.12: A passive, serial loop installed in a typical preamp. This arrangement is extremely simple, but performance is primitive.

loading effect on the tone stack). It is also extremely simple to install, and some less-demanding customers might be satisfied with this.

If, as in this example, the return signal is being injected directly into a conventional long-tailed-pair phase inverter, which also has global negative feedback, the following modification may prove useful. By using a *stereo* jack socket it is possible to 'switch off' the negative feedback when something is plugged into the return jack. This raises the effective gain of the phase inverter to open-loop level, and may compensate somewhat for the lower input signal level received from the effects pedal. Of course, the removal of

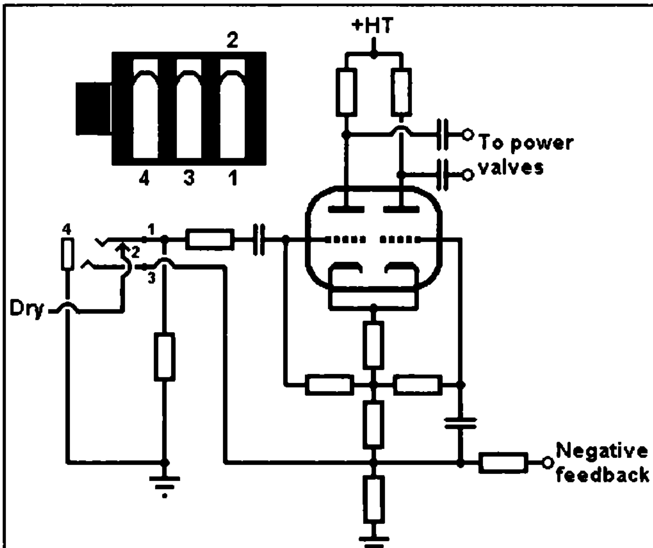


Fig. 11.13: A stereo jack can be used to 'switch off' global negative feedback when the effects loop is in use.

feedback will considerably change the tone of the power output stage, but this may be an acceptable forfeit for some users who expect a different sound anyway, when effects are used. The circuit is shown in fig. 11.13. When a conventional, mono jack plug is inserted into the jack socket, the global negative feedback is shunted to ground via the body of the plug. Connections for an ordinary insulated jack socket are shown for reference.

Active serial loops:

To see how the problems of a passive loop might be overcome, consider fig. 11.14, which shows the same section of preamp with a fully active loop installed in the same position. A dedicated ECC81 / 12AT7 cathode follower, loop driver is now included, and the second triode in the same envelope is used as a recovery stage with a gain of about 50. The HT voltages are not critical. The ECC81 is used because it has a low r_a and high g_m, making it a strong cathode follower, but it is also capable of high gain for the recovery stage.
In order to keep the tone of the amp unchanged (or nearly so) the original volume control has been replaced by a potential divider, which attenuate the dry signal to instrument level by a factor of roughly 1/50, the inverse of the gain of the recovery stage. This ensures that when the loop is *not* used,

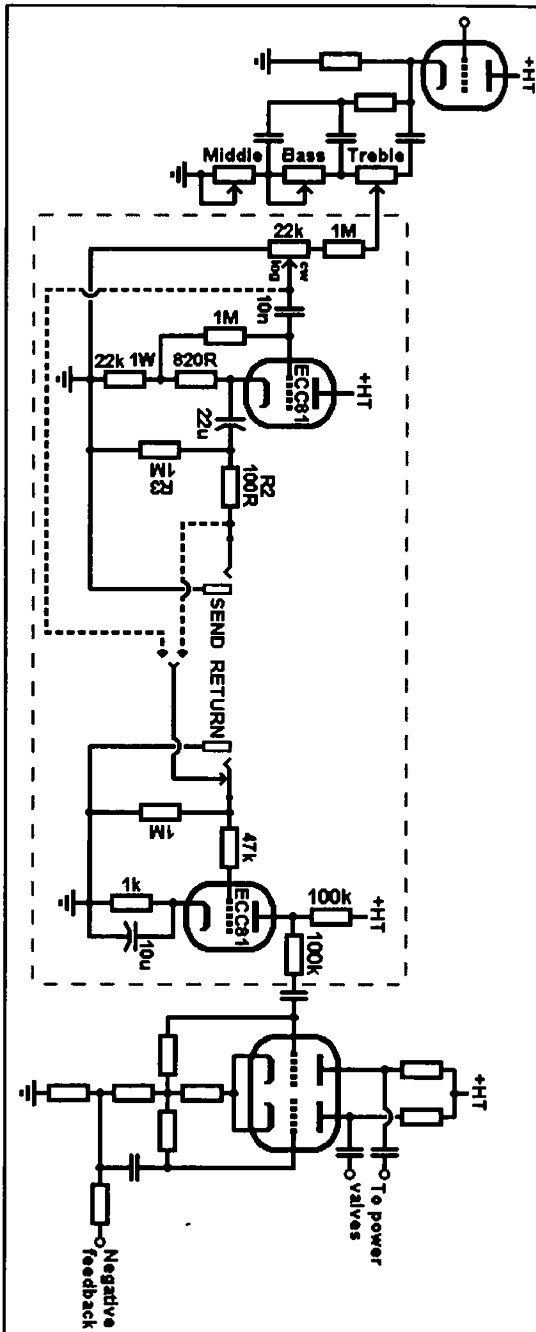


Fig. 11.14: An active serial loop installed in a typical preamp. This arrangement is intended to have minimal impact on the original tone of the amp.

the overall signal level passing through the loop is broadly similar to that when the loop *is* used. In this way neither the driver or recovery stage should ever become overdriven; they perform only clean amplification and should therefore not significantly affect the original tone of the amp (this also allows an ECC81 to be used without worry of unpleasant overdriven tones). A 100kΩ grid stopper is added to the phase inverter to simulate the source impedance of the tone stack before the loop was inserted. It causes minimal attenuation since the input impedance of the long-tailed-pair is normally several meg-ohms.
Two serial connecting paths are shown dashed. The upper path is from the output of the cathode follower, which is the conventional approach. However, in this case the cathode follower is acting solely as a clean buffer / driver stage. Since it offers no tonal advantage to the 'dry' signal, the serial path could be derived from *before* the driver stage (the lower path) which minimises the influence of the loop on the original tone, even further.

Of course, if the preamp is being designed from scratch then the loop can be 'built in' as part of the amp's ordinary tone. An example is shown in fig. 11.15. Here the cathode follower is DC coupled to the previous gain stage, which greatly reduces the number of passive components required. The output level is attenuated in the same manner as was suggested in fig. 11.6*b* (send 4), and the recovery stage is itself another gain stage. The same principle is used in the Soldano *SLO100*. Two serial connecting paths are shown dashed. The lower one is taken from the output of the send level pot', so that it does double-duty as a gain control when the loop is not in use. Alternatively, the upper path could be used, in which case the pot' only controls the line out level.
For a serial loop, the recovery stage will normally be a simple gain stage, usually using an ECC81 or ECC83, since these offer the highest gain of the common valve types. Another useful valve to consider is the ECC832 / 12DW7 currently

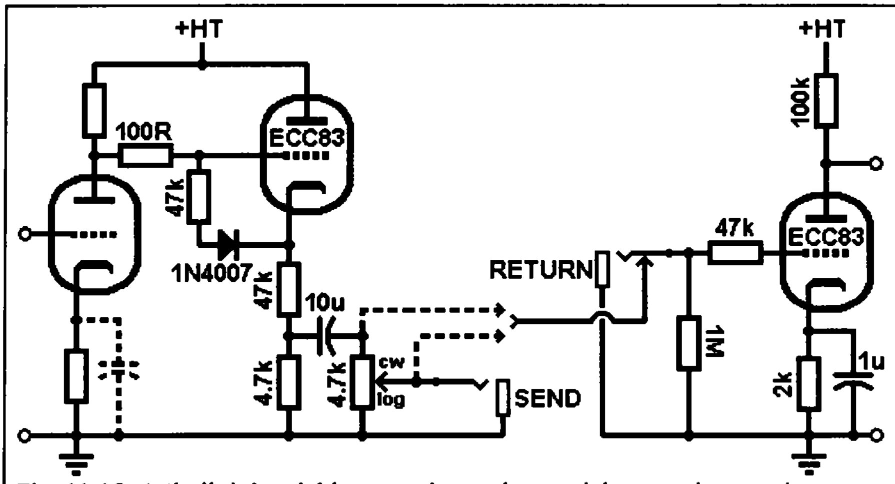


Fig. 11.15: A 'built in' serial loop need not take special precautions against sounding 'transparent'; it is *part* of the amp's fundamental tone.

manufactured by JJ Electronic. This valve contains one low-gain, high current triode similar to an ECC82 and one high-gain triode similar to the ECC83. The low gain section would be used for the loop driver and the high-gain section for the recovery stage. Because the ECC832 is pin-compatible with the ECC83 etc. it could be replaced with an ECC81/2/3 (in an emergency say) and the loop would still work, albeit sub-optimally.
Another possibility is to use a pentode-triode such as the ECF80, whose triode could be used for the cathode follower with the pentode used for recovery.

Parallel effects loops:

Serial effects loops are conceptually straight forward, but the previous section illustrates how implementing one usually requires careful design to avoid affecting an amp's existing tone, and that they are better suited to new designs where they can be 'designed in'. Parallel effects loops are a little more forgiving since the dry signal is uninterrupted. The greater difficulty is in mixing the dry and wet signals effectively, whilst retaining a simple design topology.
Designers often trouble over the phasing of a parallel loop. That is, if wet and dry signal are in-phase with one another they will tend to cancel out when mixed (unless a differential amplifier is used for the mixing, in which case the opposite is true). Unfortunately there is no industry standard as to the phasing of effects pedals; we cannot know if the wet signal will have been inverted by the effects pedals or not. Fortunately, this is not as great a problem as might be expected, since most effects change the waveform so much that relative phasing becomes irrelevant. Only if a very mild effect such a volume / swell pedal or modest equalisation is used, on a relatively clean guitar tone, are we likely to encounter significant phase cancellation problems. This is not a particularly common situation since such mild effects are normally placed at the input of the amp anyway, although a phase-switchable loop is suggested at the end of this chapter.

Passive parallel loops:

Adding a passive, parallel loop to an existing amp may not be easy, unless the existing design has some convenient place to inject the wet signal without an additional recovery stage. Fig. 11.16 shows what is probably the simplest arrangement, which should work with most existing preamps.
The 'send' signal is taken from the cathode load of the pre-existing cathode follower as in fig. 11.6 'send 4', although it could be derived from anywhere in the preamp, of course. The return signal is mixed with the dry signal, purely resistively. The pre-existing volume control is labelled 'dry' since it controls the dry signal level. The 'wet' signal is delivered to its own level pot', which is connected in the opposite manner to most potentiometers (input to wiper). This is so that the setting of this pot' does not alter the loading on the pre-existing volume control. R1 ensures that if the 'dry' control is turned fully down, the wet signal is not shunted to ground too. R2 is included so that if the 'wet' control is turned fully down, the effects pedal's output is not directly short-circuited to ground, which might otherwise overload its internal circuitry.

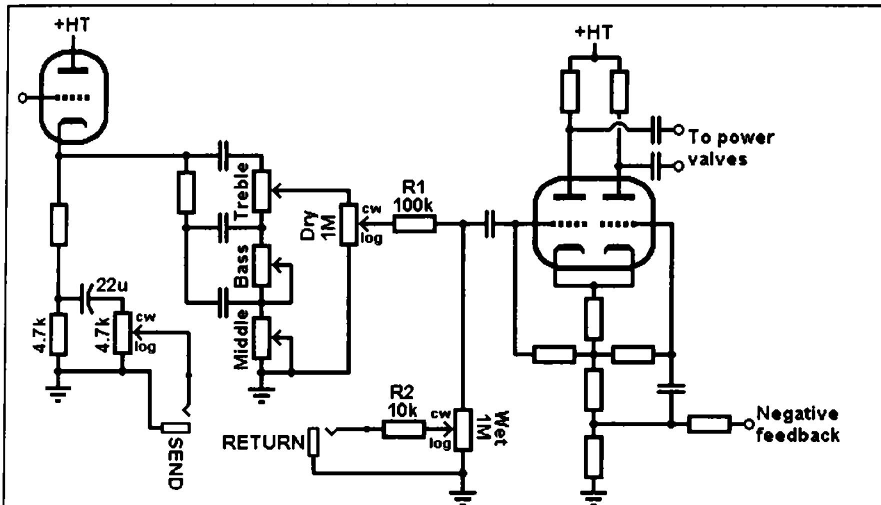


Fig. 11.16: A passive, parallel loop installed in a typical preamp. This arrangement is extremely simple, but performance is primitive.

Of course, this arrangement is primitive because the wet and dry signals are not at comparable levels, so obtaining a strong 'wet' mix while *also* driving the output stage to full power may be difficult. The 'wet' control potentiometer must also have a high resistance of 1MΩ to avoid loading down the pre-existing volume control too much, which is unfortunate, because a very high resistance in series with the low-level wet signal results in a poor signal-to-noise ratio. Nevertheless, this circuit may be sufficient for some users, and might be regarded as 'better than nothing'.

Alternatively we might choose to take advantage of the long-tailed pair, and use it to actively mix the wet and dry signals. If global negative feedback is present, it may again be eliminated by the use of a switching jack. This is illustrated in fig. 11.17, from which it should

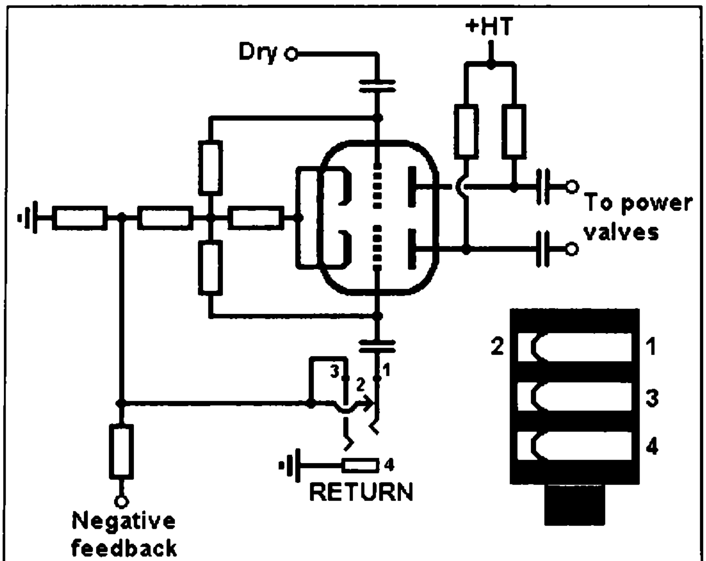


Fig. 11.17: The long-tailed pair as a mixer in which global negative feedback is removed while the effects loop is in use.

be clear that when a mono jack plug is inserted the negative feedback is grounded via the body of the plug while the pre-existing grid decoupling capacitor doubles as the coupling capacitor for the wet signal.

Active parallel loops:

An active, parallel effects loop is something of a 'holy grail' for effects loops. In spite of the mysticism, they are actually relatively easy to design, since they offer so many possibilities for novel and inventive design now that we have active stages to play with.

We must derive wet and dry signals from somewhere in the preamp and then make arrangements for them to be mixed at roughly equal amplitude, so that one does not 'swamp' the other, and we will probably want a means of blending the wet/dry levels too. This either requires boosting the wet signal to match the dry signal before mixing, which can complicate the mixing mechanism, or we can attenuate the dry signal to match the wet signal before mixing, which worsens the signal-to-noise ratio and is more likely to affect the raw tone of the amp.

If we are willing to accept *some* change in tone through the inclusion of a loop, a little imagination will devise any number of send, return and mixing topologies, just a few examples are described here. Generally we would like to dedicate no more than two triodes to the loop. We could either use one as a dedicated

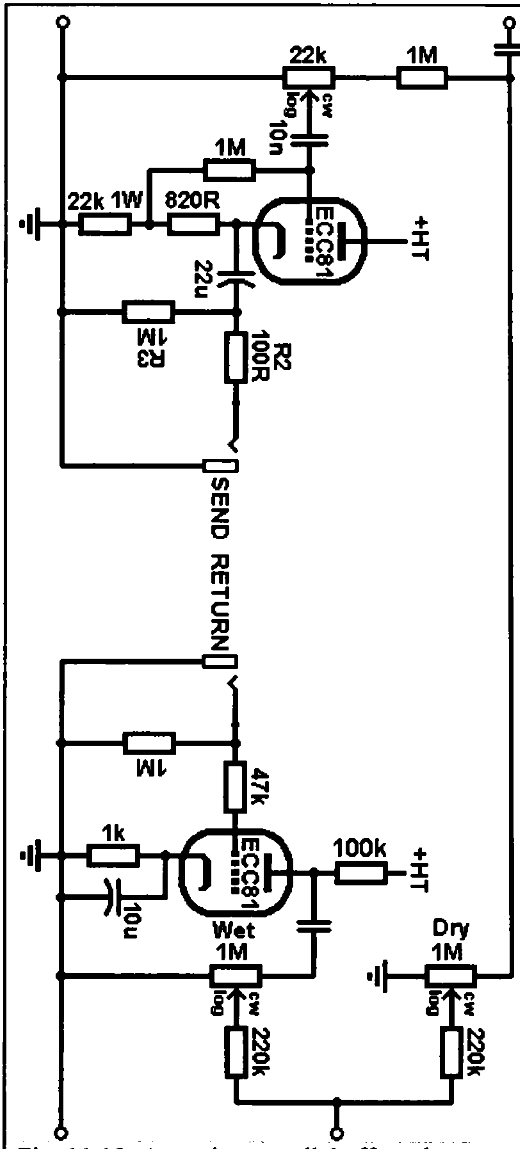


Fig. 11.18: An active, parallel effects loop employing passive mixing.

send buffer and one as a recovery stage or mixer, or we could derive the send signal passively using one of the methods shown in fig. 11.6, and dedicate both triodes to a more elaborate recovery / mixing arrangement. For example, fig. 11.18 shows a simple arrangement, which is probably the easiest way to add an active parallel effects loop to an existing amp. Comparing it with fig. 11.15 it can be seen that it comprises the same send and recovery stage, except that the dry signal is now mixed with the output of the recovery stage. Wet and dry level controls are provided and each signal is summed via the 220kΩ mixing resistors (each is required to prevent the other signal being shunted to ground if the opposite level control is fully turned down). Larger mixing resistors could be used to reduce interaction between controls, but may cause treble attenuation when fed to a gain stage having sufficient input capacitance. Of course, 'bright' capacitors could be added to overcome this [see chapter 2]. In this circuit, the 'dry' signal path is relatively unadulterated by the addition of the loop, except for the unfortunate interaction of the passive mixing. Alternatively, the two signals could be mixed via each input of a long-tailed-pair phase inverter, if no global negative feedback is used, which is quite successful.

Fig. 11.19 shows a different loop, suitable for a new preamp design. The send signal is derived from an existing gain stage [fig. 11.6, send 3]. The wet and dry signals are mixed using a common anode mixer [see chapter 4, fig. 4.9] which is arranged for a gain of around 60 for the 'wet' input, but only about 26 for the higher amplitude 'dry' input. A dry-level control is shown, which also acts a gain control when the loop is not in use. A wet-level control could also be added in the same fashion, if desired, although the same function is served by the output-level control on the effects pedals.

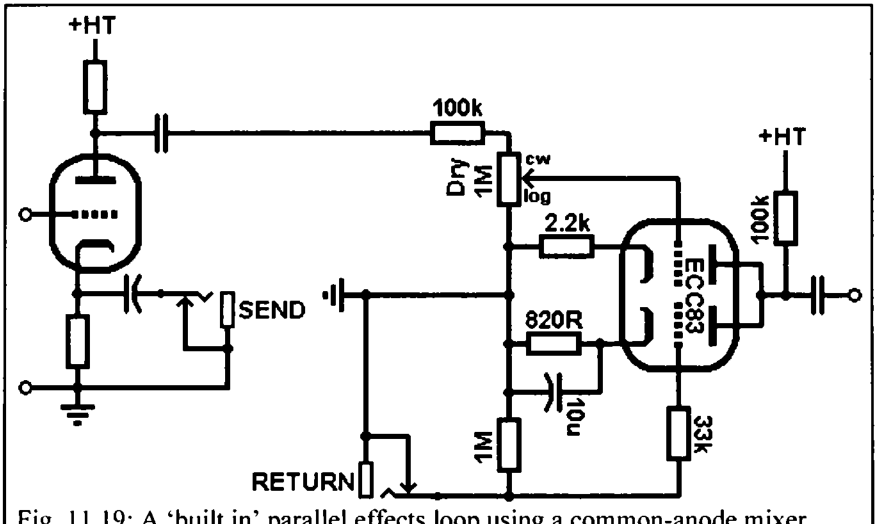


Fig. 11.19: A 'built in' parallel effects loop using a common-anode mixer.

Another useful mixer circuit is the virtual-earth mixer described in chapter 9. Fig. 11.20 shows a loop which incorporates both a dedicated, cathode-follower send buffer and an active, victual-earth mixer / recovery stage. The inputs to the mixer are weighted to suit the wet and dry amplitudes, while the gain / mix pot' allows the input resistance of each input to be varied. The gain of the mixer is therefore controllable from about 0.9 to 8 for the dry signal, and 0.9 to 28 for the wet signal. This topology is ideal for mixing signals of different level, roughly equally, without significant cross-talk between channels, and allows the dry signal to be handled at full level which helps preserve the signal-to-noise ratio. The resistor labelled 'load' represents the following load resistance, which forms the grid-leak for the virtual-earth mixer. An optional switch, S1, allows the dry signal path to be broken for either serial or parallel loop operation. It should be pointed out, however, that if the switch is open and the loop is not in use, both signal paths are broken and the amp will be silent unless the switch is closed again. This may confuse some users, and it may be necessary to use a double-pole switch, the other pole being used to operate a warning LED to indicate when the loop is in serial mode, or similar. C1, C2 and C4 must be rated for the full HT voltage, while C3 need only be rated for 63V or more.

This circuit may be inserted in series with an existing preamp with fairly minimal impact on the existing tone, except that additional gain is available if the gain / mix pot' is turned up. An ECC81 or ECC832 would also be ideal for this topology.

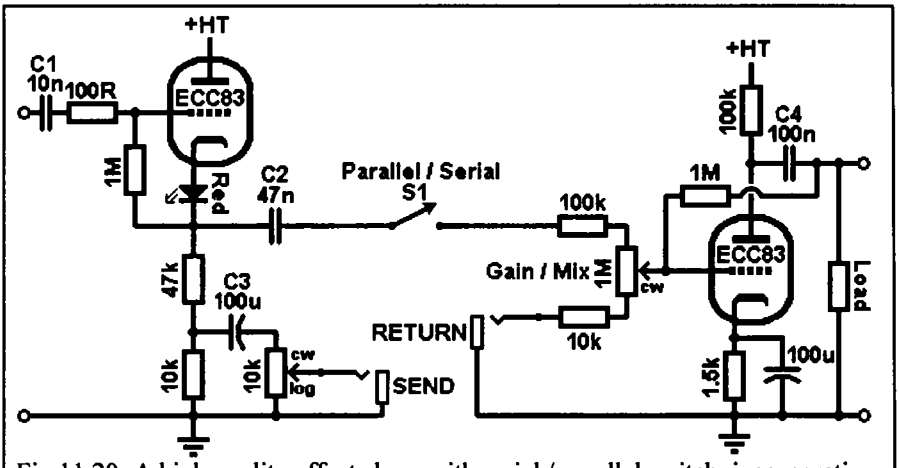


Fig 11.20: A high quality effects loop with serial / parallel switch, incorporating a variable gain, virtual-earth mixer.

If the problem of phase cancellation is thought to be likely, then fig. 11.21 shows a modified version of the loop in fig. 11.20. The first triode is now arranged as a cathodyne phase inverter [see chapter 6]. A switch, S1, selects between either the cathode or anode output, so that the phase of the dry signal may be ideally matched to the wet signal. The 'dry' input resistor has been increased to 330kΩ to avoid heavily loading the cathodyne. This limits the gain of the virtual-earth mixer to about 0.7 to 2 for the dry signal, which is still quite sufficient, given the low level of the wet signal. The 1MΩ resistors shown dashed may be included to eliminate any popping sounds when throwing S1.

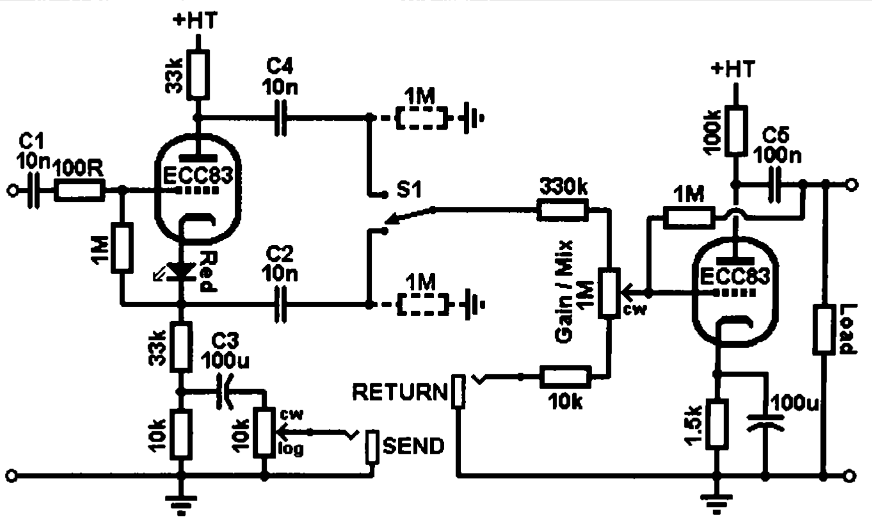


Fig 11.21: A high quality effects loop with phase switch and mix control.

If front-panel space is restricted, then a heavily simplified phase-variable loop is shown in fig. 11.22. The recovery stage is again a simple gain stage and the wet and dry signals are mixed passively.

The two opposing phases of dry signal are applied to each end of the phase / mix pot', so that any degree of phase shift from 0° to 180° is available. However, when the pot' is in its central position the two applied phases cancel out and we obtain zero dry signal, effectively giving serial-loop operation. Rotating the pot in either direction from this point gradually increases the level of dry signal almost to its full, un-attenuated level. In this way, phase, blending / mixing and parallel / serial operation are all accomplished with a single potentiometer! This circuit nicely illustrates the wide range of functionality that is possible with quite simple valve circuits, and a little imagination.

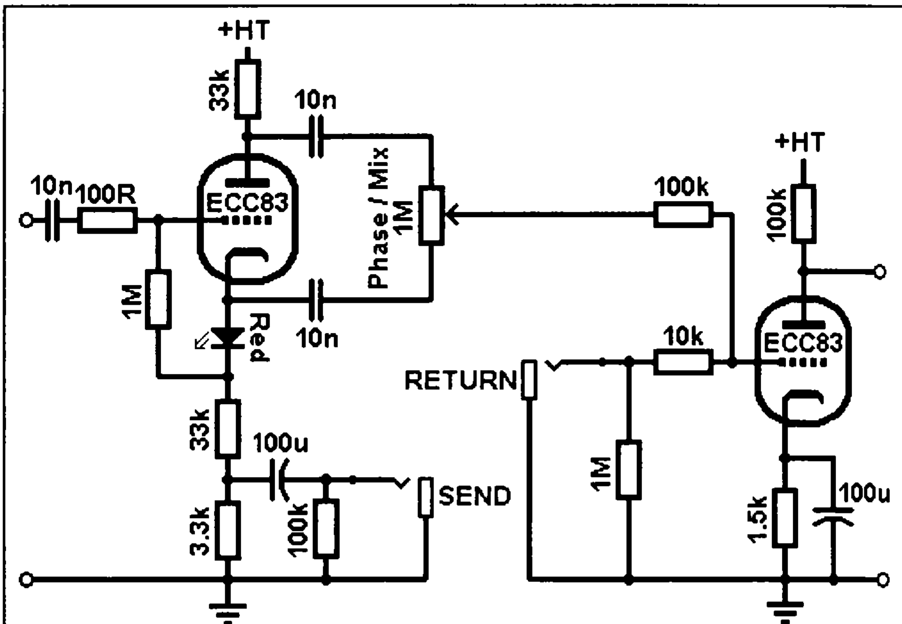


Fig 11.22: A deceptively simple effects loop with excellent functionality.

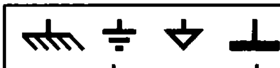
Chapter 12: Notes on Construction

Grounding. The power supply. The heater supply. A note on lead dress.

This book is mainly concerned with the design of preamp circuits rather than their ultimate construction, which is up to the readers own requirements and abilities. However, there are a few aspects of amplifier building which apply to all cases, and are worth clarifying.

Grounding:

Ground refers to the common 'reference voltage' that all the parts of a circuit share. For all the circuits in this book, which also applies to most existing guitar amplifiers, ground is zero volts, or earth, and is normally represented by one of the schematic symbols in fig. 12.1. All four are more-or-less interchangeable, though that in *a.* is normally reserved for connections to the chassis or Earth.



a. b. c. d.

Fig. 12.1: Ground / earth symbols.

Many of these symbols may appear in a single schematic, but in reality they are all ultimately connected together. However, when physically building a circuit it is important to adopt a suitable **ground scheme**, particularly in the preamp. We must not simply connect all the ground wires to each other randomly, even though it might appear that one bit of ground wire is much the same as another: they are not. A good ground scheme will achieve two things:

- Minimise the resistance in the ground circuit and avoid ground loops, making it harder for external EM fields to induce a noise voltage in the ground.
- Remove the possibility for heavy power-amp and power-supply currents from flowing in the ground circuit of other parts of the amp, particularly the preamp. This is crucial, since current flowing around the power transformer / rectifier / reservoir capacitor occurs in heavy pulses, many times greater than the average supply current, which will develop an ugly buzzing noise in the preamp if it is allowed to flow into the preamp ground circuit.

The Earth bond:

Most guitar amps are built in a metal chassis. Even if it is enclosed in a wooden box, it is usually possible for the user to touch the metal somehow, via fixing screws or when replacing valves etc. **For the appliance to be safe, it must be completely impossible for the metal chassis to become live. This is achieved by physically connecting the chassis to planet Earth via the mains earth wire.** Once the chassis is earthed, if any live wire were to touch the chassis it would be immediately shorted to earth, and so the chassis cannot shock the user, whether a fuse blows or not.

Where the mains cable enters the chassis, usually via an IEC inlet, a heavy-gauge wire should be soldered to the mains earth connection, and then connected to the chassis with a solder tag, as shown in fig. 12.2. The chassis area should be cleaned with emery paper before hand to ensure a good electrical connection. The wire should be short and should have the same colour scheme as the local mains supply. The earth wire is green-and-yellow striped in Europe and green in the US.

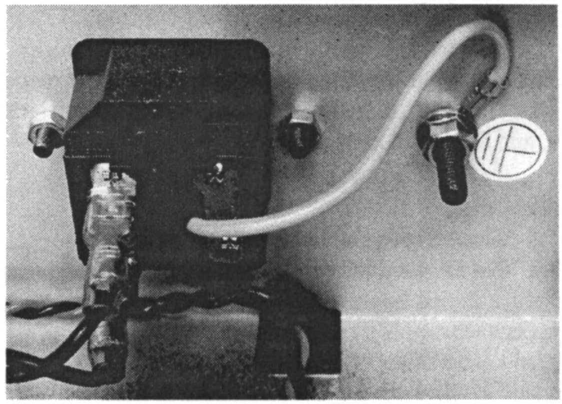


Fig. 12.2: The Earth bond should be made to a dedicated screw, close to the mains inlet.

Where this wire is bolted to chassis is known as the **Earth bond**, and it should be a dedicated screw/bolt, *not* a screw which is used to fix some other piece of hardware which might become loose over time. A nyloc nut should be used, or else a shake-proof or star washer should be used, with two ordinary nuts, well tightened. This wire is the most important connection in the amplifier and is legally required, and must be completely sound.

The Earth bond is for safety only; it plays no part in circuit operation and may be regarded as just another part of the chassis. Although the terms 'earth' and 'ground' are often used interchangeably, the *audio* circuit does not necessarily have to be connected to planet Earth. The entire amplifier circuit *could* be built 'floating' inside the metal chassis, with no connection to the chassis at all. However, in reality the circuit *will* be connected to chassis at some point, since this ensures that the amplifiers working voltages are properly defined with respect to zero volts, otherwise it might be possible for the user to receive a shock if he were to touch the guitar strings and earth at the same time while an amplifier fault had occurred.

This ground-to-chassis connection is normally made at the input of the amplifier circuit. If not, then a capacitor should be connected between the input-ground and the chassis, immediately at the amp's input, as shown in fig. 12.3. This ensures a low-impedance ground connection at RF without creating a ground loop at audio frequencies. A value of 10nF to 100nF is suitable, and may be a ceramic type. This is true whatever ground scheme is used.

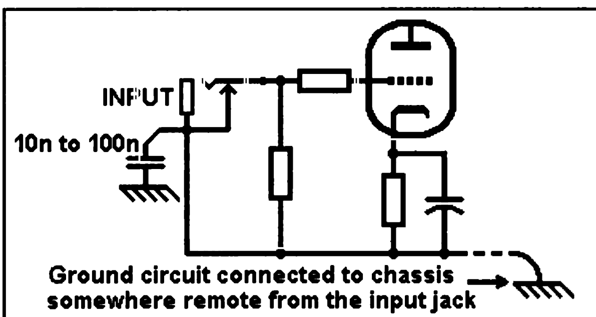


Fig. 12.3: If the circuit ground-to-chassis connection is not made at the input of the amp, then a capacitor should be connected from input-ground to chassis, as shown.

Ground loops:

When grounding stages, it is important to avoid ground loops. That is, a complete loop of wire (the chassis may also form part of the loop) into which a current could be induced by an external EM field.
For example, fig. 12.4*a* illustrates a case where one end of the shield on a shielded cable has been soldered to the tab on a potentiometer, which is also bent over and soldered to the body of the potentiometer (which is in contact with the chassis). The other end of the shield is connected to circuit ground, which has inadvertently been connected to the chassis somewhere else; a ground loop now exists, as indicated. If there is sufficient resistance in this loop then any induced currents will develop a voltage directly in series with the audio signal, creating hum, noise and even radio interference.
It is not advisable to simply bend over the tabs on potentiometers and solder them to

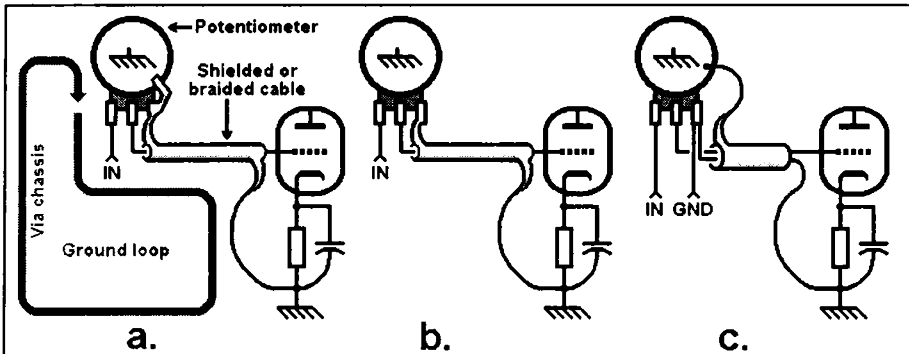


Fig. 12.4: **a.** a ground loop caused by too many chassis connection in a single circuit will lead to hum. **b.** breaking the ground loop; there is now only one chassis connection. **c.** improved layout with a completely separate shield.

the back, as this can easily lead to inadvertent ground loops, and the connection between potentiometer and chassis can become loose or corroded over time. Fig. 12.4*b* shows the corrected circuit. Better still is to avoid using the shield as part of the circuit ground. Instead, the 'hot' signal and circuit ground each have their own wire, and both are then encased in a shield or braid as in *c*. This braid should be connected to the chassis or ground at *one end only*. It therefore acts simply as a metal tube –an extension of the chassis– and is not part of the electronic circuit itself. Twisting the ground and signal wires together also aids in hum rejection.
In America, un-insulated jack sockets seem to be prevalent, which immediately force the builder to make a chassis connection where he may not want one. Therefore, insulated jack sockets are recommended, and have been more-or-less standard in British amplifiers for decades.

Ground loops can also be accidentally created if two pieces of equipment are connected together (such as an amplifier and slave amp, say) if each appliance has its own connection from circuit-ground to chassis. Some appliances will offer a **ground lift** switch that allows this connection to be broken, defeating the ground loop. This is illustrated in fig. 12.5; note that the Earth bond between chassis and earth must *never* be broken. Such switches are usually included in low-level preamps and mixers, rather than guitar amps. The easiest way to avoid ground loops is to adopt a proper, logical ground scheme.

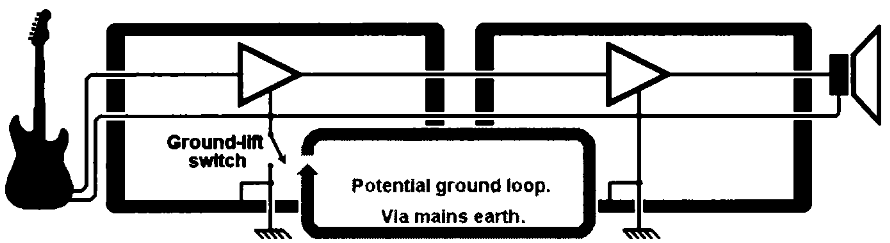


Fig. 12.12.5: A ground loop caused by connecting two pieces of equipment together via interconnecting cables. The potential loop is broken by the ground-lift switch in one of the appliances. The Earth bond must *never* be broken, however.

Bus grounding:

A popular method of grounding is the bus ground. This requires a single, heavy-gauge wire or **bus wire** being run through the chassis, to which all the ground connections are made. The path of the bus wire should follow the natural path of the circuit from preamp to power amp (if present) to rectifier/power supply, and all the ground connections should be made progressively along it, e.g., a power-amp ground should not be connected amongst the preamp ground connections. An example of this is shown in fig. 12.6. Note that that the various power-supply smoothing capacitors should be close to the stages that they feed.

The bus wire should be connected to the chassis at one point only, usually at the

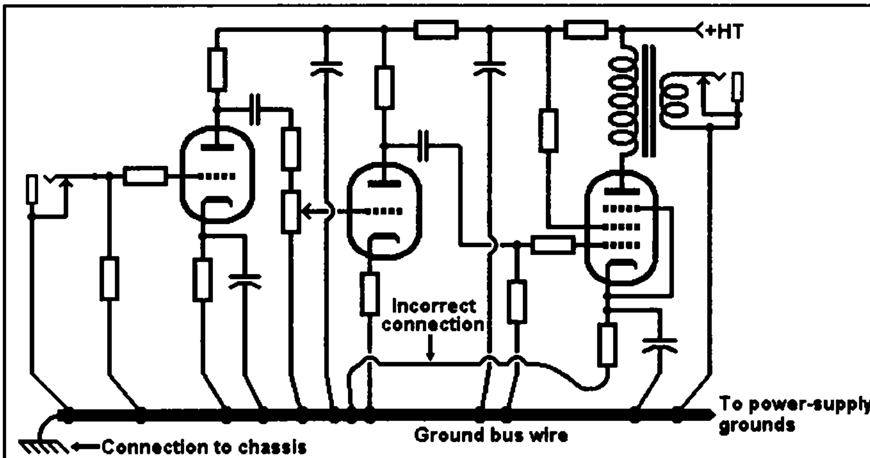


Fig. 12.6: A simple circuit extract, employing a bus ground. This method does not *always* guarantee noise-free operation, but it is methodical and does lend itself to hand-wired designs.

input of the amplifier. Tinned-copper wire can be bought on the roll, but a piece of stripped 24A or 32A solid-core mains wire is a cheap alternative for making the bus. This method of grounding does not completely guarantee noise-free operation because each of the wire runs has a resistance, albeit a small one, across which a noise voltage could be developed. Nevertheless, it is simple (difficult to get wrong!) and well suited to hand-wired designs, and once all the components are soldered to it, it can form quite a rigid, floating structure.
Some amplifiers adopt a sort of pseudo bus ground by using the chassis as the bus ('ground plane' is perhaps a better term), but this is never advisable in anything but the simplest, low-gain circuit.

Star grounding:

The simplest arrangement is the single star ground, in which all of the circuit ground connections in the amplifier are brought to a single point, where they are connected to the chassis. Fig. 12.7 shows this in essence. All of these ground wires should be as short as possible. This method is not *always* guaranteed to give low-noise performance, though it is normally suitable for very small, low gain amplifiers. It quickly becomes impractical in a large, complex circuit.

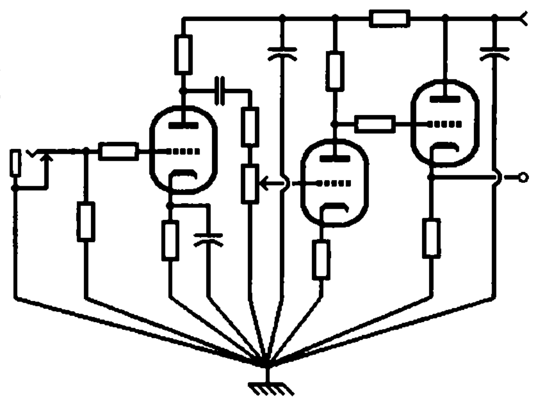


Fig. 12.7: A simple circuit extract, employing a single star ground. This type of construction is normally limited to very simple circuits.

The only 'proper' way to ground an audio amplifier is by the multi-star ground scheme. With this method, each power-supply smoothing capacitor defines a start point. All of the stages which are supplied by a particular capacitor have their ground and power supply connections made *directly to that capacitor*, and these star points are then daisy-chained together in sequence, from preamp to power amp to rectifier / power supply. One of the stars will then be connected to the chassis, usually the one nearest the amplifier's input. Alternatively, all the stars may be connected to single point –a remote star– which is bonded to the chassis, often close to the main reservoir capacitor.

This method minimises the resistance in the ground and power supply wires immediately in the vicinity of each gain stage. This is where low impedances matter the most since *signal* currents circulate within each 'loop' and not along the daisy chain, which is part of the power supply's ground loop. The proper arrangement is shown in fig. 12.8. Note that any components inbetween two cascaded stages

'belong' to the second stage, since they are part of its grid-leak circuit, and should be grounded at its star point, (fig. 12.9 ⑥ shows an incorrect example of this.)
If two stages are run from the same capacitor then the two should share the same star, but if this is very inconvenient then the capacitor may be dedicated to one or other of the gain stages without too much trouble.
Fig. 12.9 shows some of the possible mistakes which might be made, some worse than others:

1. The power supply take-off point for other parts of the amplifier is not taken directly from the smoothing capacitor, so currents drawn by other stages may produce noise voltages on the power supply here.
2. The second triode is not taking its current immediately from the smoothing capacitor; its anode signal currents may develop a noise voltage which feeds back into the first triode (this can lead to motorboating).
3. The smoothing capacitor is too far from the stages it feeds, making it easier for noise voltages to develop between the capacitor and the triodes.
4. The ground connection to other parts of the amp is not taken from a star, so any ground currents induced by other stages may affect the first triode here.
5. The ground-wire connecting the first triode to its power supply capacitor is too long, making it easier for noise voltages to develop between the triode and capacitor.
6. The second triode's grid leak has been grounded at the first triode, making it easier for noise voltages to develop between grid and cathode of the second triode (which it will then amplify).

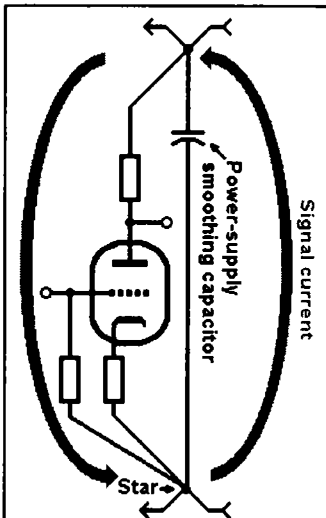


Fig. 12.8: Signal currents flow around the loop formed by a power-supply capacitor and the immediate stage/s it feeds. Ideally, connections should be made directly to the capacitor.

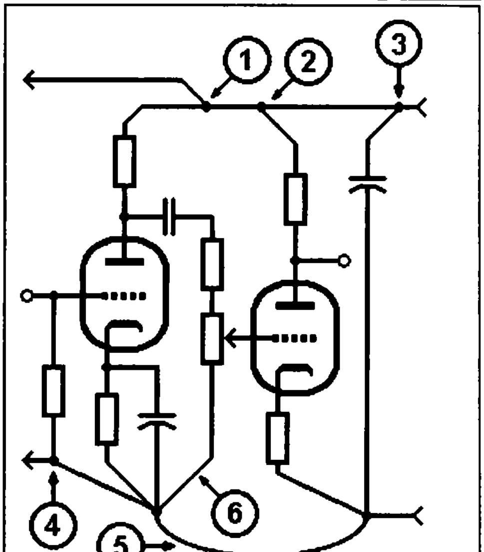


Fig. 12.9: A ground scheme with many errors which might lead to unnecessary noise.

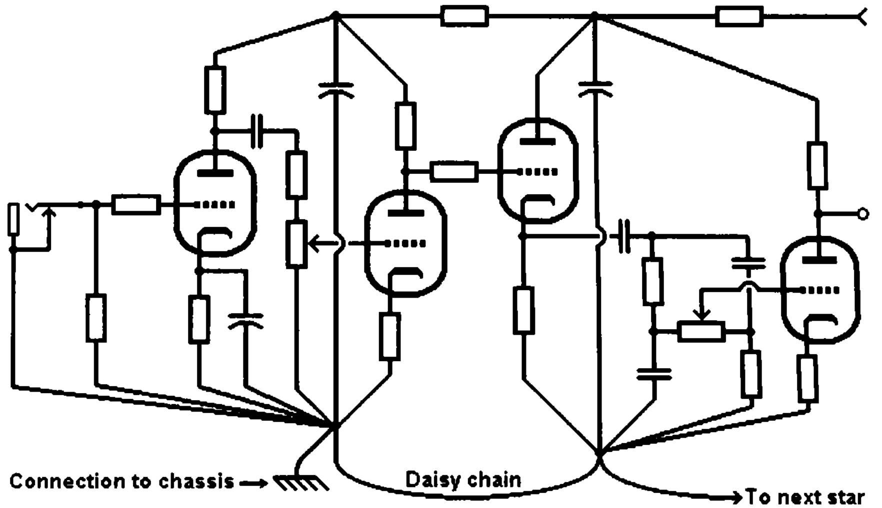


Fig. 12.10: A mutli-star ground scheme. Each power supply capacitor defines a star point, all of which are then daisy-chained together in sequence. Alternatively, all the stars may be connected to single point –a remote star- which is bonded to the chassis, often close to the main reservoir capacitor.

Fig. 12.10 shows an ideal multi-star ground scheme. It must be noted, though, that this is the optimum circumstance. Sometimes a rather less-desirable ground scheme will be used out of necessity, and indeed, many commercial amps use quite atrocious ground scheme and 'get away with it'. Perhaps the worst habit is to place all of the power supply smoothing capacitor in one place and then run long wires to the relevant parts of the amplifier. This inevitably leads to unnecessary hum, although in a low to medium gain design it may just be bearable, or we may just 'get lucky'. In a high-gain design the importance of star grounding is much greater, and bad practice will usually make a high-gain amp unusable.

Vintage amplifiers often used so-called 'can capacitors' in the power supply. This is a single, metal can, containing several capacitors sharing a single negative terminal. Such components were convenient in the days when electrolytics were bulky or expensive, but it is not ideal with respect to grounding, and can-caps should be avoided in modern circuits as a result.

The power supply:

This book has assumed throughout that a suitable power supply is available. It is not the intention of the author to discuss the operation or design of power supplies, save for the following comments:

The various stages of a preamp are normally supplied via a chain of RC filters, each comprising a series dropping resistor and **smoothing** or decoupling capacitor, from a high voltage source. A typical circuit extract is shown in fig. 12.11. Each of these filters serves to attenuate any AC noise which might be on the power supply, reducing it to pure DC. Of course, the earliest stages in the amplifier are the most

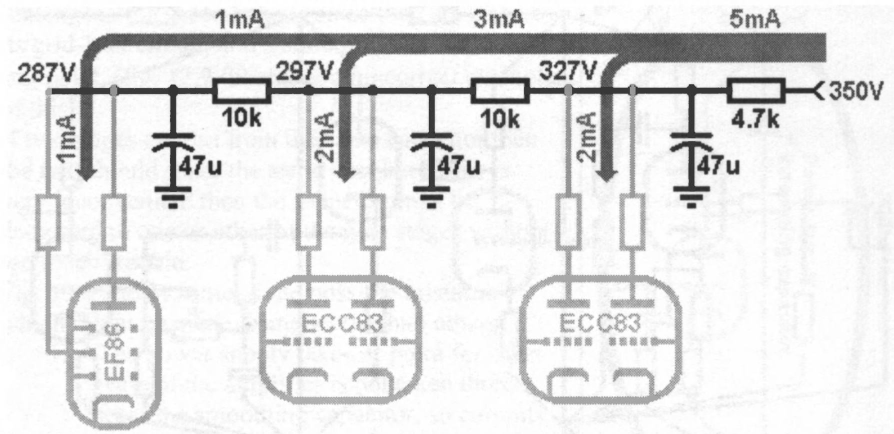


Fig. 12.11: A typical power supply arrangement. The dropping resistors are chosen to give the desired voltage drop to each HT node, as well as serving as RC smoothing filters.

sensitive to power supply noise, so these will usually be placed at the end of this chain, where the cumulative smoothing effects are maximum.

When setting up these HT filters, the series dropping resistors may be chosen first. Because each gain stage draws current from the power supply, through the dropping resistors, the HT voltage will fall progressively towards the last filter stage.
Since the quiescent current of each gain stage is known (or can be closely estimated) the voltage at each HT node is easily determined according to Ohm's law. The power rating of these resistors is just as easily calculated, and in a preamp will rarely exceed ½W per dropping resistor.
As a rule of thumb, a typical ECC83 / 12AX7 triode will demand about 1mA. Usually, using the same value resistors throughout is quite sufficient, since the actual HT voltage is not especially critical. The filtering *is*, however. Ideally we would like each RC filter to have a –3dB roll-off frequency of 0Hz (DC). Of course, this is not possible in reality, but a value of 1Hz may be taken as a maximum. The minimum suitable value for a smoothing capacitor is therefore:

$$C = \frac{1}{2\pi R}$$

A typical dropping resistor might be 4.7kΩ, requiring a smoothing capacitance of:

$$C = \frac{1}{2\pi \times 4.7k} = 34\mu F$$

We might accept 33µF, though a common standard for high-voltage capacitors is 47µF. Vintage amplifiers often use rather under-sized capacitors since sufficiently large values were expensive, bulky or not readily available, and many old amps suffer from dubious low-frequency stability. This trend should not be copied in a new design; modern capacitors are cheap, physically small and always improving.

The voltage rating of the capacitors should exceed the maximum possible supply voltage, taking into account any rise which may occur due to warm-up time at switch-on (when the power supply is not fully loaded) and mains voltage variations. Voltage ratings higher than 450V are not readily available, so it is advantageous to design amplifiers operating at voltage less than 400Vdc. However, if we do find that we need a higher voltage rating, then capacitors may be connected in series to increase the voltage rating of the whole. Equal value resistors should be placed in parallel with each capacitor to ensure the HT is divided equally between them, as in fig. 12.12. The resistors should be large enough that they do not waste too much HT current, but should pass at least five times the expected leakage current of the capacitor, which is typically in the region of 0.05mA. Values of around 220k to 470k ½W are typical. They can also act as a 'bleeder' network that will allow the power supply capacitors to discharge when the amp is switched off. The total capacitance achieved will be less than one capacitor alone, since capacitors sum in the opposite manner to resistors. To counter this, additional capacitors could be connected in parallel to increase the total, as in fig. 12.12*b*.

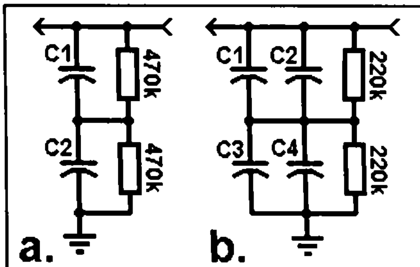


Fig. 12.12: Using capacitors in series to increase the voltage handling of the whole, though the total capacitance is reduced. For **a.**:

$$C_{total} = \frac{C1.C2}{C1 + C2}$$

For **b.**:

$$C_{total} = \frac{(C1 + C2)(C3 + C4)}{C1 + C2 + C3 + C4}$$

This can be particularly useful for reservoir capacitance, since large, good quality 250V capacitors are readily available and very cheap, and with two in series a working voltage of 500V is obtained.

It is common for every two triodes in a preamp to be supplied by one smoothing capacitor. Usually, when one of the triodes draws more current, the other draws less (if they are cascaded) so the change in the total current supplied from the capacitor is minimal. If more than two valves are supplied from the same capacitor, however, the variations in current can become great and, unless the capacitor is extremely large, the HT voltage will be forced to fluctuate at low frequencies. This low-frequency signal will be fed back to the earliest stage, resulting in low-frequency ringing or, ultimately, motorboating. A 'boomy' or 'loose' bass tone is often a symptom of insufficient power-supply decoupling, and as a rule no more than two triodes should ever be fed from the same smoothing capacitor.

A common error made by amateurs is to add a new valve to an existing amplifier circuit, but to supply it from one of the pre-existing smoothing capacitors, and this frequently leads to low-frequency instability. A new RC filter should be installed to feed the new valve.

Similarly, some users may attempt to reduce the bass response of an amp by deliberately using a small-value smoothing capacitor. This must never be done; coupling capacitors may be reduced instead.
Small-signal pentodes should ideally be supplied from their own, dedicated smoothing capacitor, as they are much more sensitive to power-supply noise.

As a further note, electrolytic capacitors have a finite life-span of a few years before they begin to dry up. Their life-span also reduces dramatically if they are allowed to become warm during operation, so electrolytics should never be placed too close to any hot valves or resistors. As the electrolyte evaporates, ESR increases, reducing the ability to decouple the HT, and bass tone will begin to suffer. Old capacitors must never be used in a new amp, and replacing aging capacitors with new ones nearly always 'tightens up' the bass, giving a faster attack to bass notes. The conscientious player will have *all* electrolytic capacitors at least every ten years (including any electrolytic cathode bypass and bias-supply capacitors).

The heater supply:

The heater supply is one of the four most likely sources of noise in a valve amp, together with power supply noise, improper grounding and resistor hiss. Most of the heater noise is due to the EM field around the heater wires coupling into the signal circuitry, while some is due to direct leakage between heater and cathode. Both of these can be minimised by good design.

A useful but little-used addition to a heater circuit is to connect a large poly' type capacitor across the heater supply (connecting it directly between the heater pins of the input valve is ideal). This helps to filter out high-frequency noise which is coupled into the heater circuit from the HT rectifier, if both heater and HT windings are on the same transformer. A value of 100nF to 470nF is suitable for this. Although it will not alter the overall hum-levels, it serves to reduce any harsh-sounding buzz from the heater supply, which is more annoying than the low-frequency hum.

Heater lead dress:

Most guitar amps use AC heaters (usually 6.3Vrms). The heater current in an amplifier is much greater than any of the other currents in the amp, so the EM field around heater wiring is very strong. Because an EM field is proportional to current flowing in a conductor, and *not* to voltage, it is desirable to operate preamp valves such as the ECC83 / 12AX7 from a 12.6V supply since this requires only half the current and therefore generates only half the EM field strength; a 6dB noise improvement over 6.3V operation. Power valves can also be operated from a 12.6V supply by connecting them in series, of course.

The heater supply is normally a single chain, daisy-chaining from one valve socket to the next and supplying all the heaters in parallel. The current flowing in the *supply* end of the chain is therefore the sum of all the heater currents, while at the other end of the chain it only has to supply one valve. Power valves require the most current and are the least sensitive to heater hum so should be at the beginning of the chain

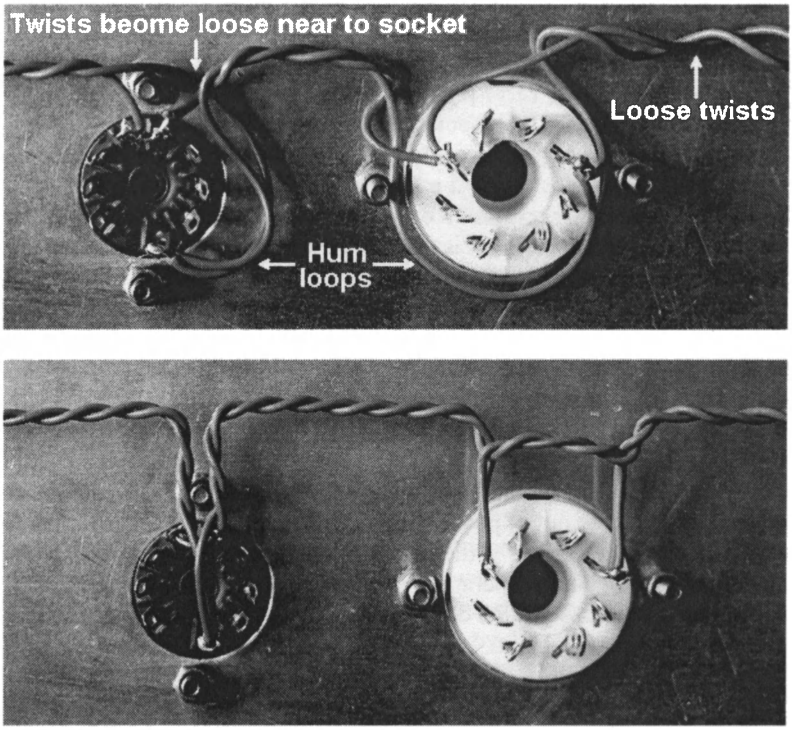


Fig. 12.13: **Upper**: Bad heater wiring with common mistakes, leading to excess heater hum. **Lower**: Good heater wiring keeps the wires away from the socket where possible, and keeps the twisting tight, right up to the socket.

(heavier gauge wire may be necessary in this region). The heater chain should then progress through the amplifier towards the input, and the input valve will be the last in the chain, where current is least.

AC heater wires must always be *tightly* twisted as this helps to suppress their EM fields, since each twist is out-of-phase with its neighbours. Note, **a loose twist is useless, only a tight twist will do!** This is easily done by anchoring the wires at one end and holding the other in the chuck of a drill, keeping reasonable tension on the wires while twisting. Stretching it gently before releasing will discourage the wire from twirling back on itself. It is better to use stranded wire for this, as solid-core wire usually develops metal fatigue when twisted and will often fail immediately, or if not, at some point in the future.

The heater wires should be kept as far from any signal wires as possible, and pushed into the corner of the chassis if possible. Orientating the valve sockets so that the heater wiring can approach from the edge of the chassis will help with this. If signal

wires must cross heater wiring, they should do so at right-angles, to reduce the possibility of EM coupling between the two.
As the heater wires approach a valve socket, *the twisting must be kept tight right up to the socket*, since the other valve pins are in close proximity. Allowing the twists to become loose near the socket will spoil a lot of hard work!
It is also very important not to create a loop of heater wiring around a valve socket[*], since any wires or valve pins inside the loop will be subject to strong EM interference. The heater wiring should only ever approach from one side of the socket and if it must cross it (which is usually the case for preamp valves), it should go directly across the socket and straight back. An example of good heater wiring is shown in the lower image in fig. 12.13; this takes care and patience, and will usually be obscured by other wiring once the amp is complete, so it is worth spending the time on getting it right at the start. The upper image shows some typical mistakes, hum loops being the most common and problematic.

Hum balance:

AC heater wiring should be kept electrically balanced. This ensures that hum induced by one wire out of phase with hum induced by the other, so the two will cancel out to a large extent. If the (heater) transformer has a centre tap it should be grounded, to give balanced operation, as shown in fig. 12.14*a*. It does not normally matter where this ground connection is made; it will normally be connected to the chassis at the nearest point.

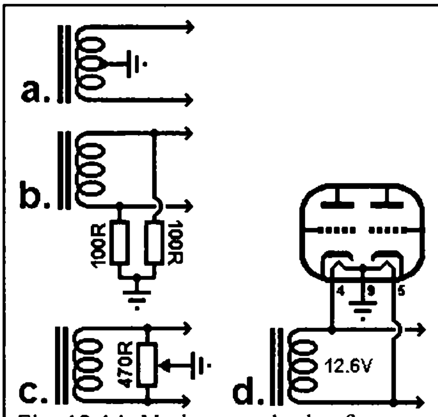


Fig. 12.14: Various methods of ensuring a balanced heater supply:
a. Transformer centre tap.
b. Artificial centre tap.
c. 'Humdinger' potentiometer.
d. Centre tap on a valve heater.

If there is no centre tap then an 'artificial centre tap' can be created by connecting a resistor from each heater leg to ground, as shown in fig. 12.14*b*. The resistors should have a low resistance so the reference to ground is as close to zero as possible. Values of 100Ω ½W, or 220Ω ¼W are usual. They will of course cause a small amount of extra current draw from the transformer (32mA when using 100R resistors at 6.3V) so bare this in mind. If preamp valves such as the ECC83 / 12AX7 are operated at 12.6Vrms then pin-9, which is a tapping between the heater of each triode, may be used to create a centre tap, if the transformer does not have one. This is shown in fig. 12.14*d*. Note, *only one* valve should have its pin-9 connected to ground to achieve this.

[*] Most commercial amplifiers do make this mistake, and correcting it will always reduce the hum levels.

Another traditional method to perfectly balance the heater-hum signals is to connect a potentiometer across the heater legs, with the wiper connected to ground, as shown in *c*. Again, a low value of 220Ω to 500Ω is ideal. This **humdinger** allows the minimum hum to be dialled in precisely, and is very effective.

Heater elevation:

Occasionally a circuit will demand the heater supply to be elevated, so as not to exceed the $Vhk_{(max)}$ ratings of the valves. This is done simply by 'adding' a DC voltage to the heater supply. The heaters still operate at 6.3V (or whatever), but this 'floats' on top of a DC reference voltage.

This method is also used to reduce audible heater hum by raising the heater voltage above the cathode voltage and 'switching off' the stray current between filament and cathode. This works provided the DC reference voltage is sufficient to raise the negative AC peaks above the cathode voltage of the valves. Reference voltages used are typically between 5V and 100V, and combining elevation with a humdinger can reduce heater noise to very low levels indeed.

A convenient way to apply the DC reference is by connecting the centre tap (any of those given in fig. 12.14) to the cathode of a power valve, assuming the power valve is cathode biased of course. The bias voltage of most power valves is usually more than 8V, so the heater supply will be 'jacked up' by this amount. This is shown in fig. 12.15*a*. No current flows in the centre tap, so there are no implications for the bias of the power valve.

Alternatively, we may take the DC reference from a potential divider from some convenient point on the HT –it does not matter where– as shown in fig. 12.15*b*. Typical voltage references are around 20V to 90V in this case, placing the heater supply well above the potential of most cathodes in the amp. The potential divider should have a fairly high resistance so as not to waste HT current, although R2 should not be excessively large or the maximum heater-to-cathode resistance may be exceeded, and it is advisable not to make it

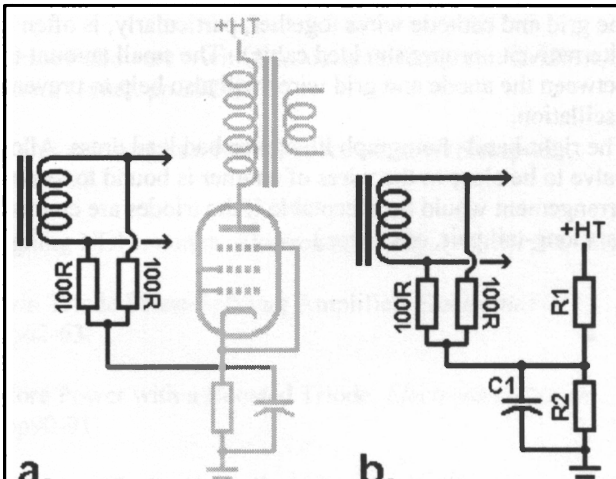


Fig. 12.15: Elevating the heater supply by a fixed DC potential can reduce heater hum to a considerable extent.

greater than 100kΩ. An arbitrarily large-value capacitor, C1, may also be added to ensure the DC reference is noise free, and anything over 10uF should be ample.

A note on lead dress:

Most of the common preamp valves contain two or more triodes. When making connections between the valve socket and a PCB or turret board, wires 'belonging' to one triode should be kept away from wires belonging to any other triode. This is essential to prevent feedback via parasitic capacitance from one to the other, which may otherwise lead to oscillation in higher-gain designs.

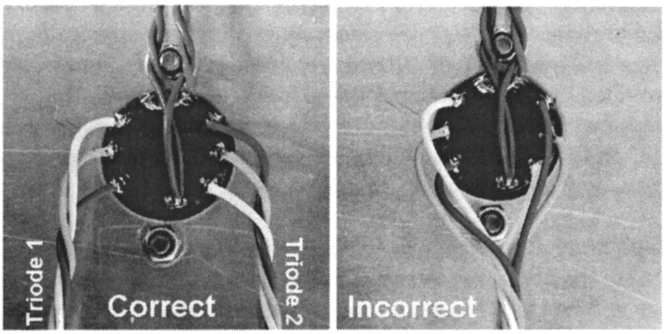


Fig. 12.16: **Left:** The wires leading to different triodes must be kept separate. **Right:** Grouping all the wires together will lead to parasitic ringing or oscillation.

Twisting the wires of each individual triode together is recommended, as shown in the left-hand photograph of fig. 12.16. This will make it visually clear which valve each bundle of wires is serving and creates a strong physical link, preventing individual wires from moving out of place over time. The low-impedance cathode wire will also help to shield the others from external interference (twisting the grid and cathode wires together, particularly, is often a cheap and convenient alternative to proper shielded cable). The small amount of additional capacitance between the anode and grid wires will also help to prevent high-frequency oscillation.

The right-hand photograph illustrates bad lead dress. Allowing the wires of one valve to be close to the wires of another is bound to lead to problems. (This arrangement would be acceptable if the triodes are connected directly in parallel, or as a long-tail pair, of course.)

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NB: At the time of writing, many of these references are available for free as .pdf files from **http://www.thermionic.info/** and **http://www.pmillett.com/**

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