

# 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.

## 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

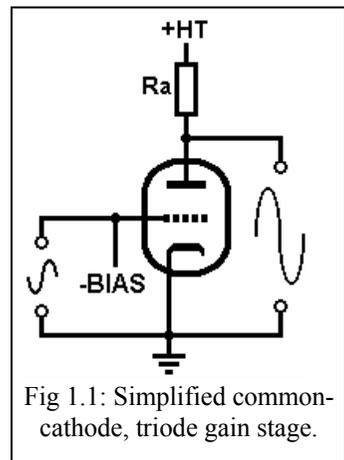


Fig 1.1: Simplified common-cathode, triode gain stage.

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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.

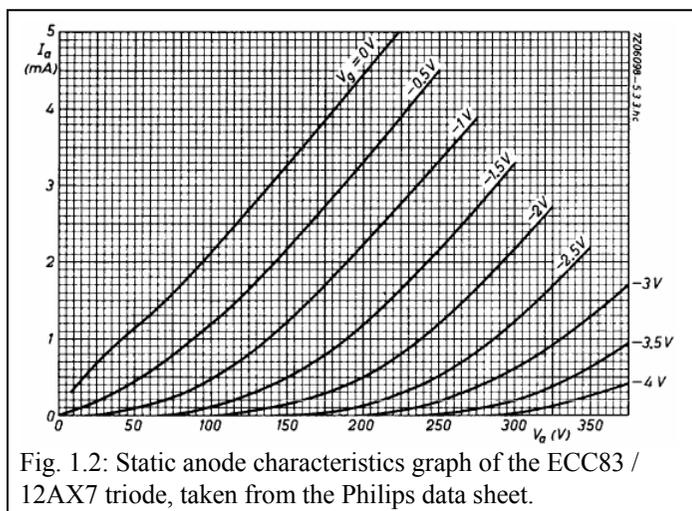


Fig. 1.2: Static anode characteristics graph of the ECC83 / 12AX7 triode, taken from the Philips data sheet.

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

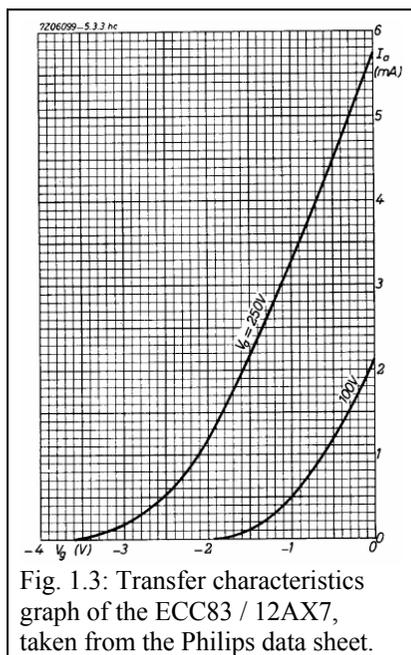


Fig. 1.3: Transfer characteristics graph of the ECC83 / 12AX7, taken from the Philips data sheet.

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.

The static anode characteristics in fig. 1.2 show anode current,  $I_a$ , on the y-axis: this is simply the current flowing from anode to cathode at any time. The x-axis shows anode voltage,  $V_a$ : this is the voltage dropped across the valve between anode and cathode. The set of curved lines shows the grid-to-cathode voltage,  $V_{gk}$ , 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

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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**),  $R_a$ , and forms the load. For small-signal valves the value of the anode resistor is usually in the region of  $100k\Omega$ , 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  $100k\Omega$  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  $I_a = 0mA$ ,  $V_a = 300V$  (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  $I_a = 3mA$ ,  $V_a = 0V$  (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.

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\* 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.

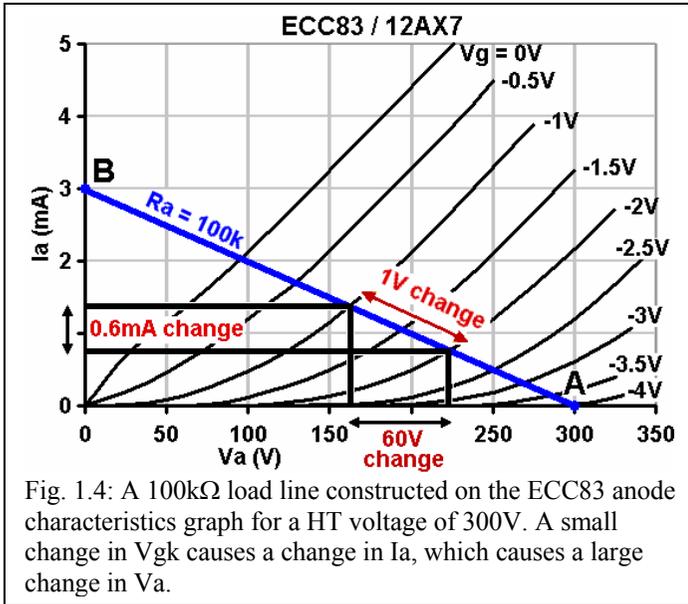


Fig. 1.4: A 100kΩ load line constructed on the ECC83 anode characteristics graph for a HT voltage of 300V. A small change in  $V_{gk}$  causes a change in  $I_a$ , which causes a large change in  $V_a$ .

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  $V_a$  and  $I_a$  will be for any given value of  $V_{gk}$ . 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 *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

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**biasing** the valve. For example; we could fix the grid voltage at  $-1\text{V}$ , 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

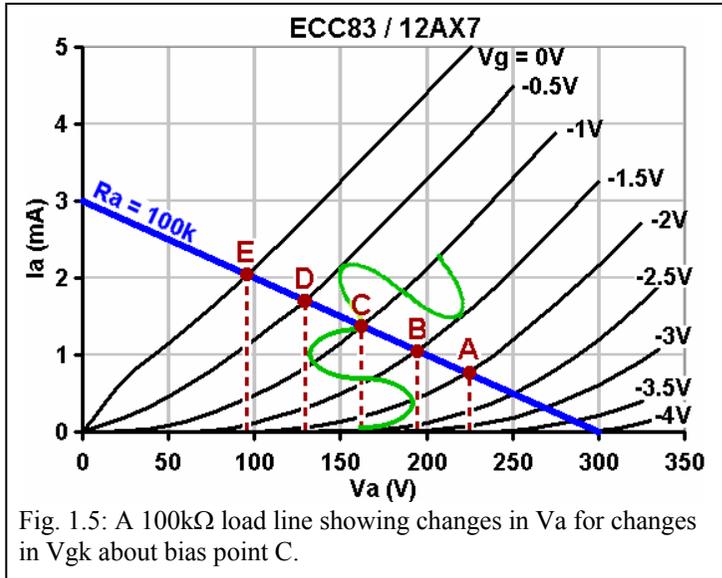


Fig. 1.5: A  $100\text{k}\Omega$  load line showing changes in  $V_a$  for changes in  $V_{gk}$  about bias point C.

that our quiescent bias voltage of  $-1\text{V}$  causes a quiescent anode current of  $1.4\text{mA}$ , and a quiescent anode voltage of  $164\text{V}$ .

We will now input a  $1\text{V}$  peak-to-peak sine wave ( $1\text{V}_{p-p}$ ). At first the grid voltage swings more positive to D ( $-0.5\text{V}$ ) and causes  $V_a$  to fall to  $130\text{V}$ . Then the grid swings more negative to B ( $-1.5\text{V}$ ) causing  $V_a$  to rise to  $195\text{V}$ . You can perhaps imagine the input signal streaming down the  $-1\text{V}$  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  $180^\circ$  out of phase with respect to the input; the stage is said to be an **inverting** gain stage.

Furthermore, our  $1\text{V}_{p-p}$  input signal has produced a  $195 - 130 = 65\text{V}_{p-p}$  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:

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

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

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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  $2 \times 65 = 130V$  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 2<sup>nd</sup> harmonic distortion is given by<sup>1</sup>:

$$H2\% = \frac{CD - CA}{2(CD + CA)} \times 100 \dots \dots \dots II$$

$$H2\% = \frac{71 - 59}{2 \times (71 + 59)} \times 100$$

$$= 4.6\% \text{ 2}^{\text{nd}} \text{ harmonic distortion.}$$

Using the same method on our earlier example of a 1Vp-p input signal yields a value of 3.1% 2<sup>nd</sup> 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

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

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than triodes and have more ‘S-shaped’ transfer characteristic. This tends to generate mainly 3<sup>rd</sup>, 5<sup>th</sup> 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  $-2.5\text{V}$  and  $V_a$  rises to 250V. But, when we try and make the grid more positive we find there are no more grid curves shown beyond  $V_{gk} = 0\text{V}$ . 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**<sup>\*</sup>, 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.

This effect does not happen instantly. Forward grid current actually begins to flow around  $V_{gk} = -1\text{V}$ , though this varies between

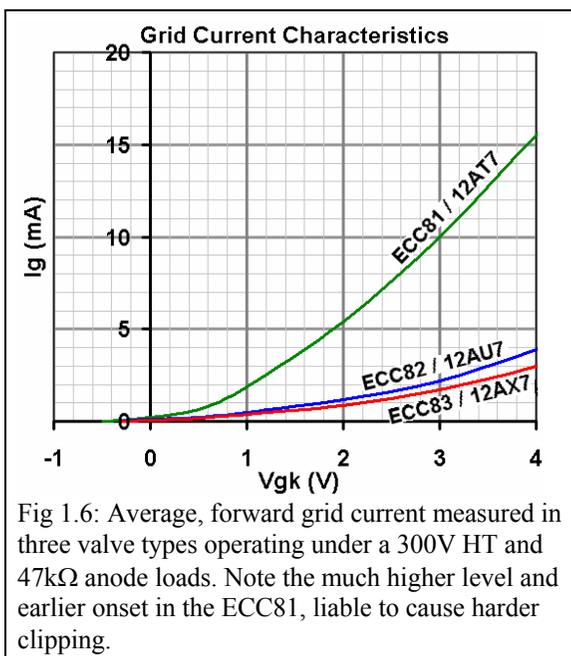


Fig 1.6: Average, forward grid current measured in three valve types operating under a 300V HT and 47k $\Omega$  anode loads. Note the much higher level and earlier onset in the ECC81, liable to cause harder clipping.

<sup>\*</sup> 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.

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valve types. The data sheet may even quote this as ' $V_{gk(max)}$ ', indicating the point at which grid current will begin to exceed  $0.3\mu A$ , 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 2<sup>nd</sup> and 4<sup>th</sup> order harmonics, while the a 'flat top' to the clipped wave indicates the introduction of some 3<sup>rd</sup> and 5<sup>th</sup> 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  $100k\Omega$  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.

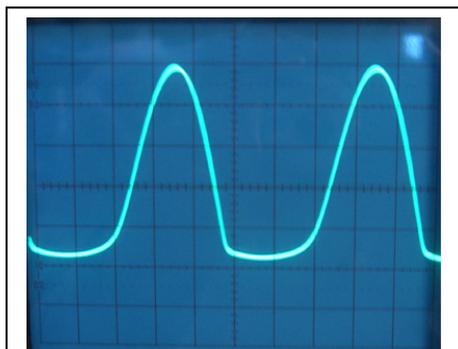


Fig. 1.7: Soft clipping due to grid-current limiting in an ECC83.

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.

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#### 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  $V_{gk} = 0V$  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  $V_{gk} = -2.5V$  (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.

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  $V_{gk} = -4V$ , 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.

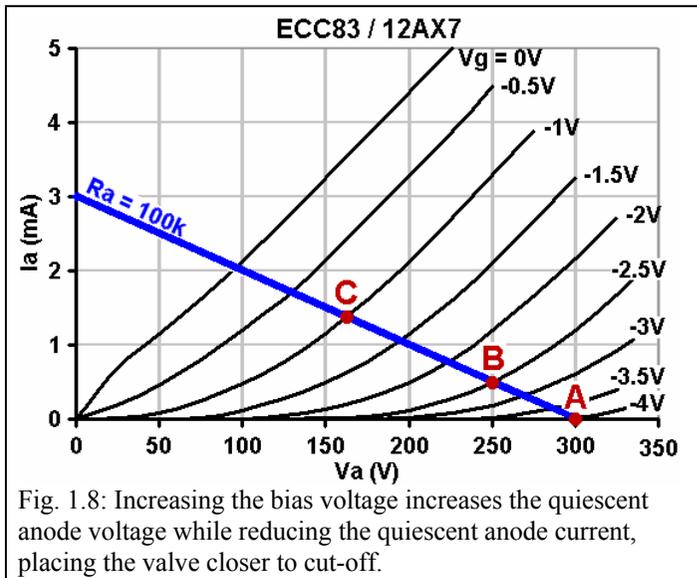


Fig. 1.8: Increasing the bias voltage increases the quiescent anode voltage while reducing the quiescent anode current, placing the valve closer to cut-off.

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 2<sup>nd</sup> and 3<sup>rd</sup> harmonics are introduced, with many higher harmonics diminishing smoothly, as indicated by the wave's flat top.

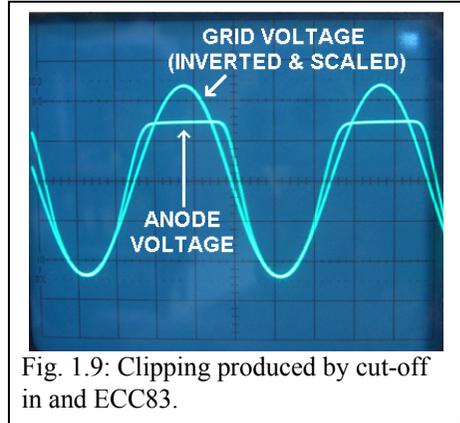


Fig. 1.9: Clipping produced by cut-off in and ECC83.

### **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 2<sup>nd</sup>), 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.

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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’

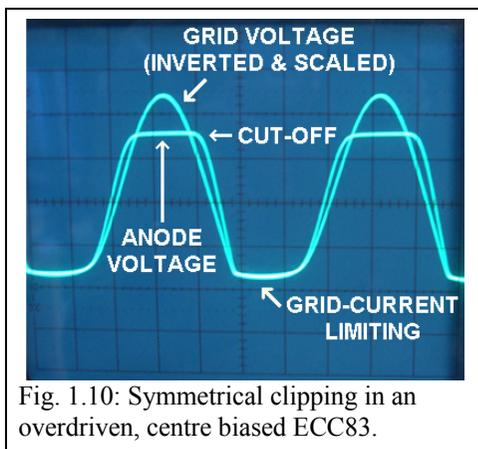


Fig. 1.10: Symmetrical clipping in an overdriven, centre biased ECC83.

\* Older texts may refer to this as ‘anode bend rectification’.

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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 **Pa<sub>(max)</sub>** or **Wa<sub>(max)</sub>**. 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:

$$P / V = I$$

$$1 / 500 = 2.0\text{mA}$$

$$1 / 400 = 2.5\text{mA}$$

$$1 / 300 = 3.3\text{mA}$$

$$1 / 200 = 5.0\text{mA}$$

### The common cathode, triode gain stage

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!

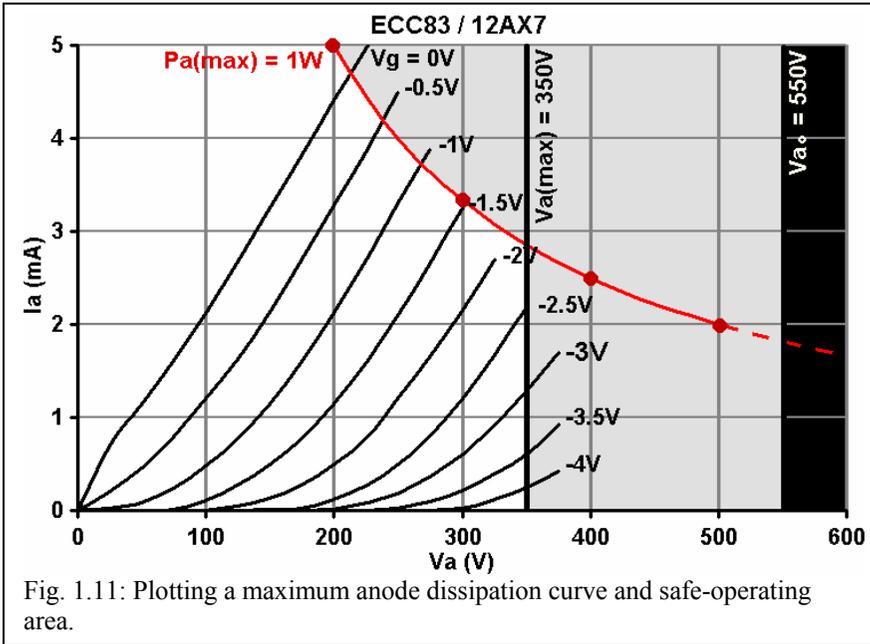


Fig. 1.11: Plotting a maximum anode dissipation curve and safe-operating area.

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  $V_{a_0}$ , 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  $V_{a(\max)}$ , and for the ECC83 this is 350V. Therefore our bias point must MUST NOT lie to the right of  $V_a = 350V$ , 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  $V_{a_0}$  at any time.

## Designing Valve Preamps for Guitar and Bass

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

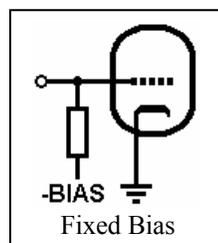
We have now fully defined the area in which it is safe to bias the valve, and in this instance the load line doesn't even fall outside any of the boundaries. Power output stages often work much closer to these boundaries, so more care must be taken when choosing operating conditions.

### **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

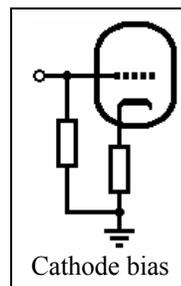


could be derived from the power supply or from a battery. Deriving a negative voltage from the power supply is fairly straightforward, though it would need to be very low noise indeed if we were to use it for biasing preamp valves, which adds complexity. However, this method is often used on power valves to help maximise output power.

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.



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

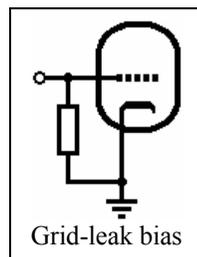
## The common cathode, triode gain stage

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,  $V_{gk}$ .

### **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:



- 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  $10M\Omega$  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  $100k\Omega$  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: 220k $\Omega$ , 100k $\Omega$  and 47k $\Omega$ , 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.

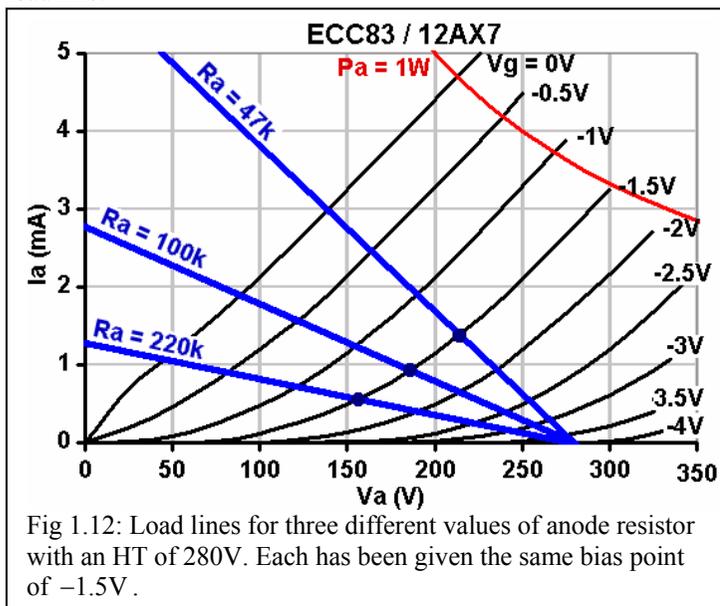


Fig 1.12: Load lines for three different values of anode resistor with an HT of 280V. Each has been given the same bias point of -1.5V.

It can be seen that the larger the load resistance, the larger the possible output voltage swing. With the 47k $\Omega$  load the maximum unclipped output signal is roughly 125Vp-p, while for the same bias voltage the 220k $\Omega$  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 47k $\Omega$ , 100k $\Omega$  and 220k $\Omega$  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 470k $\Omega$ .

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 47k $\Omega$  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 10k $\Omega$ , 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:

### The common cathode, triode gain stage

47k $\Omega$ : 10% 2<sup>nd</sup> harmonic

100k $\Omega$ : 7.2% ‘

220k $\Omega$ : 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 47k $\Omega$ . A high-gain design would tend to use a larger resistance, more than 100k $\Omega$  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 100k $\Omega$ .

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 100k $\Omega$  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:

$$R = V / I$$

$$= 1.5 / 0.0009$$

$$= 1667\Omega$$

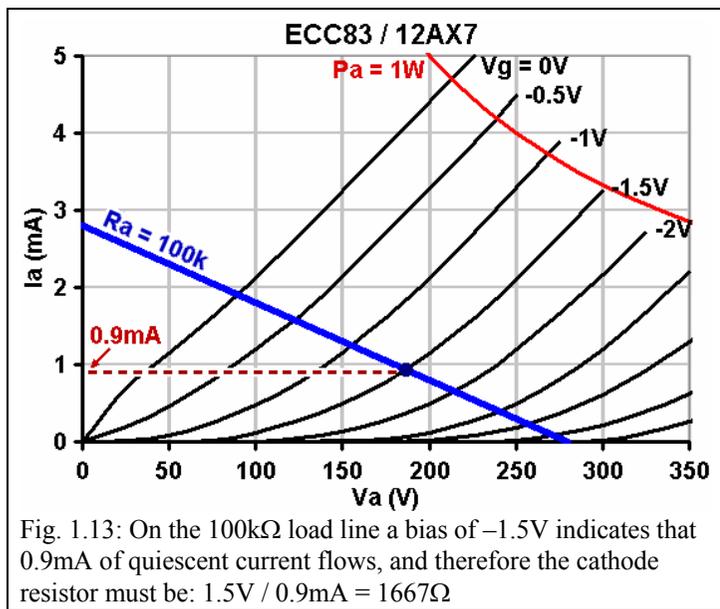


Fig. 1.13: On the  $100k\Omega$  load line a bias of  $-1.5V$  indicates that  $0.9mA$  of quiescent current flows, and therefore the cathode resistor must be:  $1.5V / 0.9mA = 1667\Omega$

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  $1.5k\Omega$ . 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:

$$\begin{aligned} V &= IR \\ &= 0.0015 \times 1500 \\ &= 2.25V \end{aligned}$$

This point can now be plotted on the characteristics graph in fig. 1.14 at  $I_a = 1.5mA$ ,  $V_{gk} = 2.25V$  (point A).

Now let us suppose that  $0.5mA$  flowed through it:

$$\begin{aligned} V &= 0.0005 \times 1500 \\ &= 0.75V \end{aligned}$$

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 of cathode resistor.

### The common cathode, triode gain stage

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

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 runaway 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

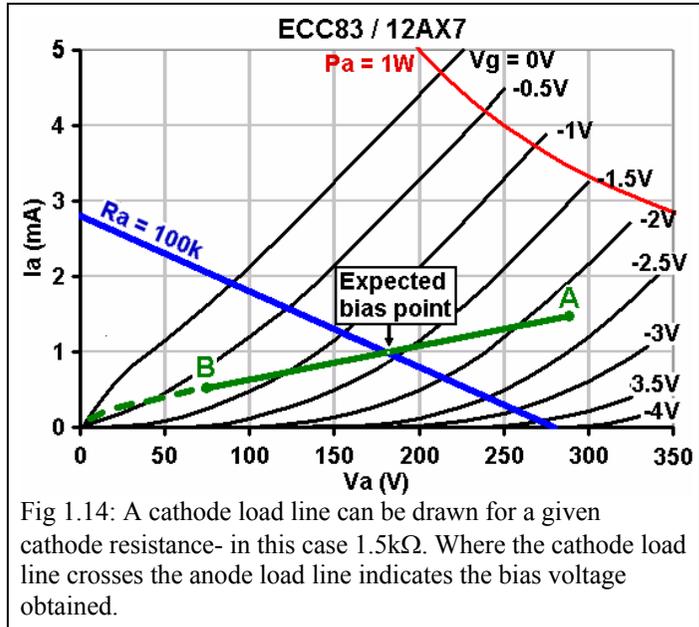


Fig 1.14: A cathode load line can be drawn for a given cathode resistance- in this case  $1.5\text{k}\Omega$ . Where the cathode load line crosses the anode load line indicates the bias voltage obtained.

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

## Designing Valve Preamps for Guitar and Bass

given as  $2M\Omega$ , although some quote  $22M\Omega$  provided the anode current does not exceed  $5mA$ . Almost all circuits use  $1M\Omega$  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  $1M\Omega$  since higher resistances increase noise while offering little extra benefit

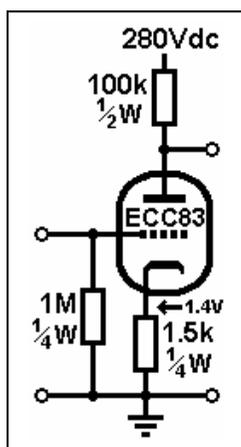
Power valves will usually have two maximum values listed: one for fixed bias and one for cathode bias, which will be somewhat higher. The reason for this is that with cathode bias, anode current cannot increase to such a great extent due to the self-regulating effect of the cathode resistor, inhibiting thermal runaway. 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:

$$\begin{aligned} P &= I^2 R \\ &= 1^2 \times 100 \\ &= 100mW \end{aligned}$$

To ensure a reliable working life it is customary to use a component having at least twice the power handling capacity, so a  $\frac{1}{4}W$  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  $\frac{1}{2}W$  resistor to be safe.



The cathode resistor has only a small voltage across it, and will dissipate only:  
 $1^2 \times 1.5 = 1.5mW$ , and even if the valve could somehow pass the maximum current of  $2.8mA$  the power would never exceed  $11.8mW$ , so a  $\frac{1}{4}W$  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  $\frac{1}{4}W$  would require an impossible  $\sqrt{0.25W \times 1M\Omega} = 500V_{rms}$  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  $\frac{1}{4}W$  resistor should be sufficient.

### The common cathode, triode gain stage

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<sup>1</sup>, causing them to generate their own 3<sup>rd</sup> 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<sub>m</sub>):**

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<sub>m</sub> 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 I<sub>a</sub> for a small change in Vgk. In this case changing Vgk from -1V to -2V (a net change of 1V) causes a change in I<sub>a</sub> of 1.7 - 0.3 = 1.4mA. The g<sub>m</sub> is therefore:

$$g_m = \frac{\Delta I_a}{\Delta V_{gk}} = \frac{1.4}{1} \\ = 1.4\text{mA/V or } 1.4\text{mS}^*$$

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<sup>1</sup> 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 ( $\mu$ ):**

When a load is placed in series with the valve a small change in  $V_{gk}$  causes a much larger change in  $V_a$ , 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  $\mu$  (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  $V_a$  for a change in  $V_{gk}$ . In this case a change in  $V_{gk}$  from  $-1V$  to  $-2V$  (a net change of  $1V$ ) results in a change in  $V_a$  of  $240 - 140 = 100$ . This gives a  $\mu$  of:

$$\mu = \mu = \frac{\Delta V_a}{\Delta V_{gk}} = \frac{100}{1} = 100$$

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  $V_a$  would cause a change in  $I_a$ . 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

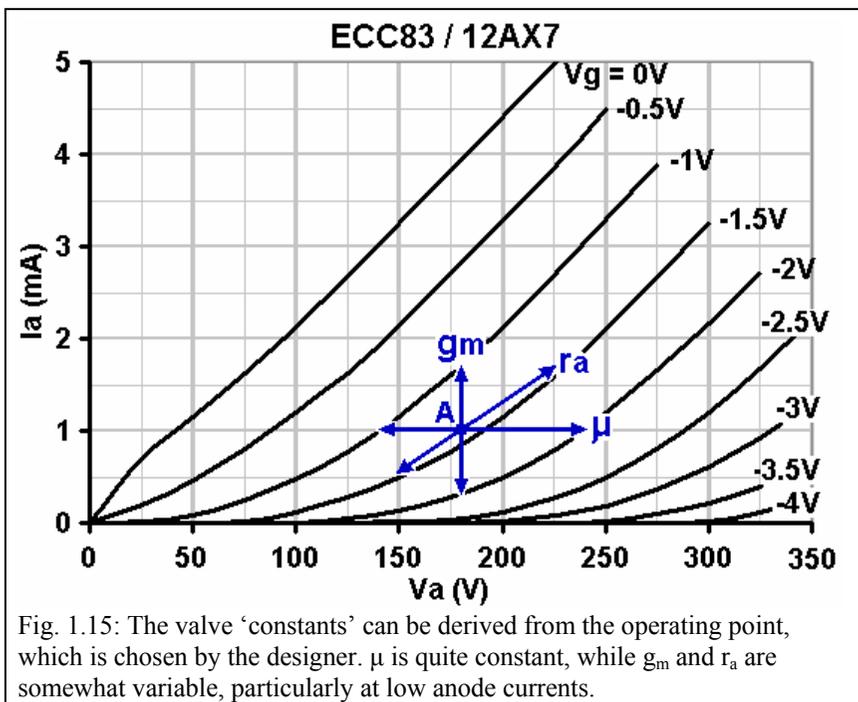


Fig. 1.15: The valve 'constants' can be derived from the operating point, which is chosen by the designer.  $\mu$  is quite constant, while  $g_m$  and  $r_a$  are somewhat variable, particularly at low anode currents.

The common cathode, triode gain stage

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

$$r_a = \frac{\Delta V_a}{\Delta I_a} = \frac{75}{1.05} = 71k\Omega$$

The data sheet will provide typical values of  $r_a$ ,  $g_m$  and  $\mu$ , 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  $\mu$  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<sup>2</sup>:

$$\mu = r_a \cdot g_m$$

so that only  $g_m$  and  $\mu$  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  $\mu$ ,  $g_m$  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  $\mu$ . This amplifier forces a signal current to flow down the series combination of  $R_a$ ,  $r_a$  and any other impedance which might be present, which together form a potential divider. If the signal is taken across  $R_a$  then the gain of the potential divider,  $\beta$ , is simply  $R_a / (R_a + r_a)$  and so the voltage gain of a simple resistance loaded, common cathode, triode gain stage is<sup>3</sup>:

$$A = \frac{-\mu R_a}{R_a + r_a} \dots\dots\dots III$$

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<sup>2</sup> 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.

<sup>3</sup> Scroggie, M.G.,(1958). *Foundations of Wireless* (7<sup>th</sup> 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  $R_k$ , we might expect the gain of the circuit in fig 1.16 to be:

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

$$= 58.$$

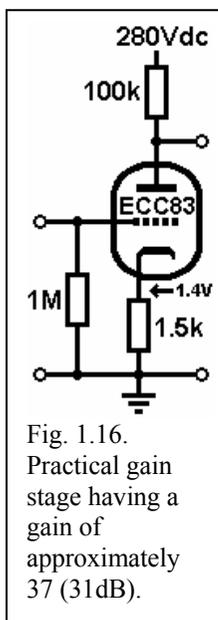


Fig. 1.16.  
Practical gain stage having a gain of approximately 37 (31dB).

However, we can only ignore  $R_k$  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  $R_k$ , 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  $V_{gk}$  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  $V_{gk}$ , 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 amplifies 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<sup>4</sup>:

$$A = \frac{\mu R_a}{R_a + r_a + R_k(\mu + 1)} \dots \dots \dots IV$$

It was shown earlier that in this case  $\mu$  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

<sup>4</sup> Langford-Smith, F. (1957) *Radio Designer’s Handbook*, p485. Iliffe and Sons Ltd., London.

### The common cathode, triode gain stage

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 ‘smooths 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 sound. Adding the capacitor removes this feedback effect, result in somewhat harder clipping and a more aggressive overdrive tone.

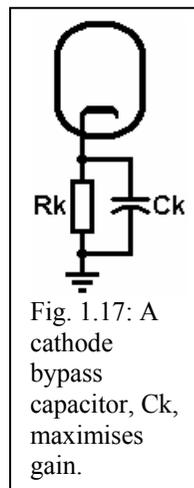


Fig. 1.17: A cathode bypass capacitor,  $C_k$ , maximises gain.

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<sup>5</sup>:

$$A = \frac{\mu R_a}{R_a + r_a} \cdot \sqrt{\frac{1 + (2\pi f R_k C_k)^2}{\left[1 + \frac{R_k(\mu + 1)}{R_a + r_a}\right]^2 + (2\pi f R_k C_k)^2}} \dots\dots\dots V$$

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<sup>6</sup>:

$$f_{(\text{half boost})} = \frac{1}{2\pi R_k C_k} \cdot \sqrt{1 + \frac{R_k(\mu + 1)}{2(R_a + r_a) + \frac{1}{2} R_k(\mu + 1)}} \dots\dots\dots VI$$

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

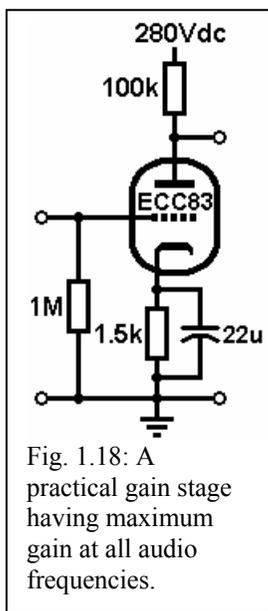


Fig. 1.18: A practical gain stage having maximum gain at all audio frequencies.

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

(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:

$$C_k = \frac{1}{2\pi \times 1.5k \times 5} = 21\mu F$$

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

<sup>5</sup> Langford-Smith, F. (1957) *Radio Designer's Handbook*, p484. Iliffe and Sons ltd., London.

<sup>6</sup> Blencowe, M & James, D. I., (2008). ‘Choosing Cathode Bypass Capacitors’. *Audio Xpress*, (August), pp. 19-20.

### The common cathode, triode gain stage

designs use a 25V component. However, it is an unfortunate property of electrolytic capacitors that they have a limited working life, relatively high ESR<sup>\*</sup>, 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<sup>7</sup>. 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 around 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

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

<sup>7</sup> Self, D. (2006). *Audio Power Amplifier Design Handbook* (4<sup>th</sup> ed.), p175, Newnes, Oxford.

## Designing Valve Preamps for Guitar and Bass

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  $C_k$ . 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  $C_k$ , which can be found using formula VI.

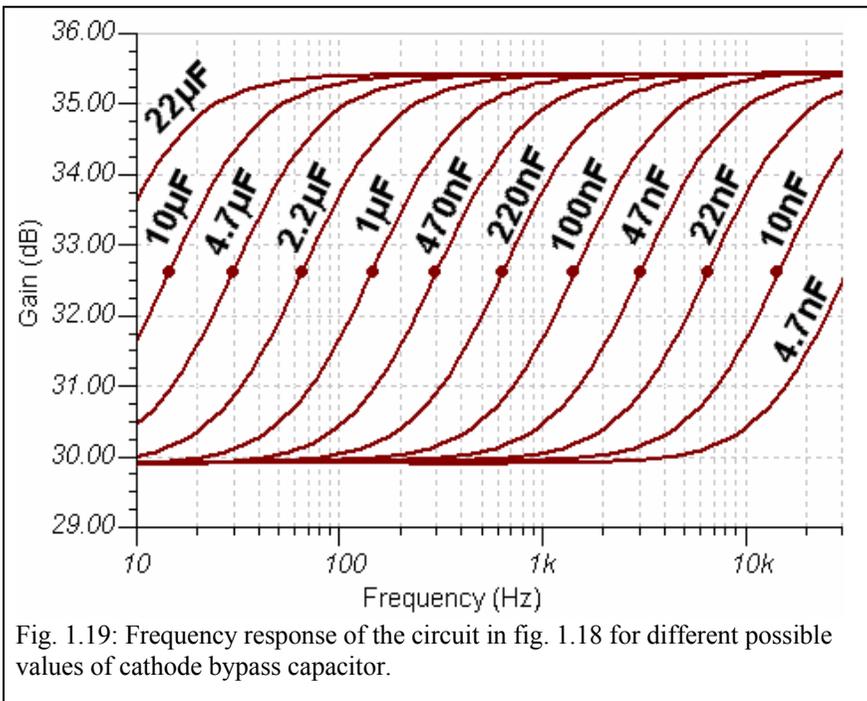


Fig. 1.19: Frequency response of the circuit in fig. 1.18 for different possible values of cathode bypass capacitor.

### **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  $C_k$ .

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:

The common cathode, triode gain stage

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

$$= 38$$

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

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

$$= 47$$

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

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

$$= 80\text{Hz}$$

An obvious modification to the circuit in fig. 1.20a 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.

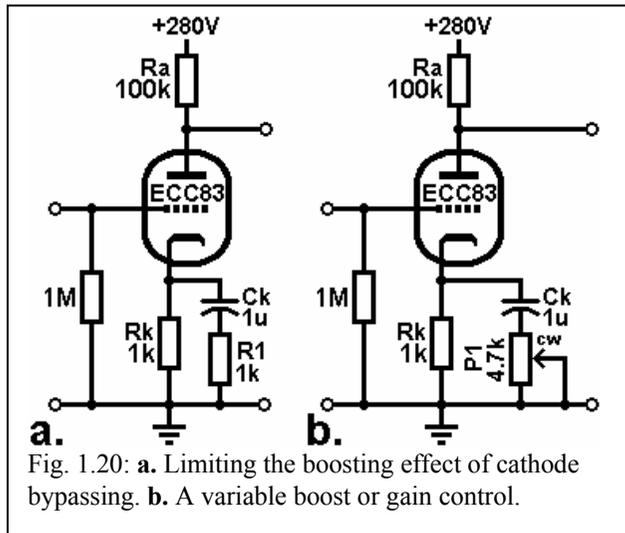
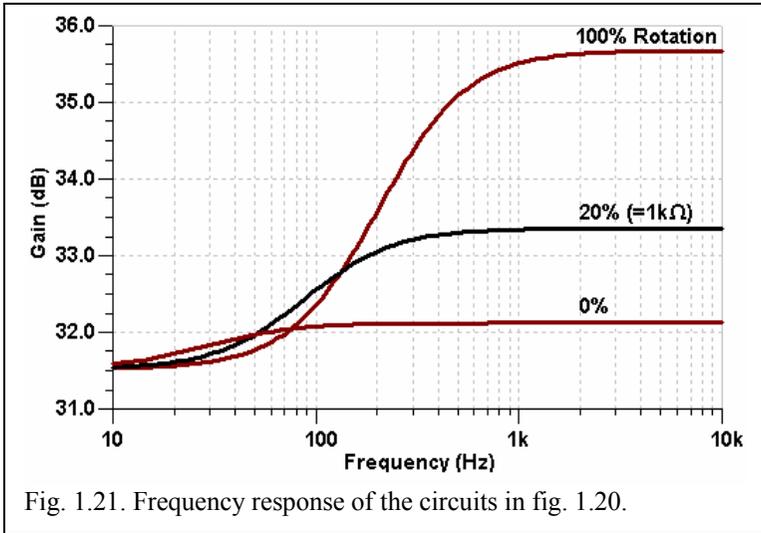


Fig. 1.20: **a.** Limiting the boosting effect of cathode bypassing. **b.** A variable boost or gain control.

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.



Because the full gain of the circuit in 1.20a 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:

$$f_{(\text{half boost})} \approx \frac{1}{2\pi \times 1000 \times 100 \times 10^{-9}}$$

$$= 1.6\text{kHz}$$

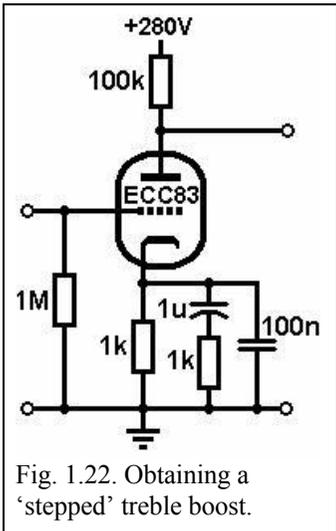
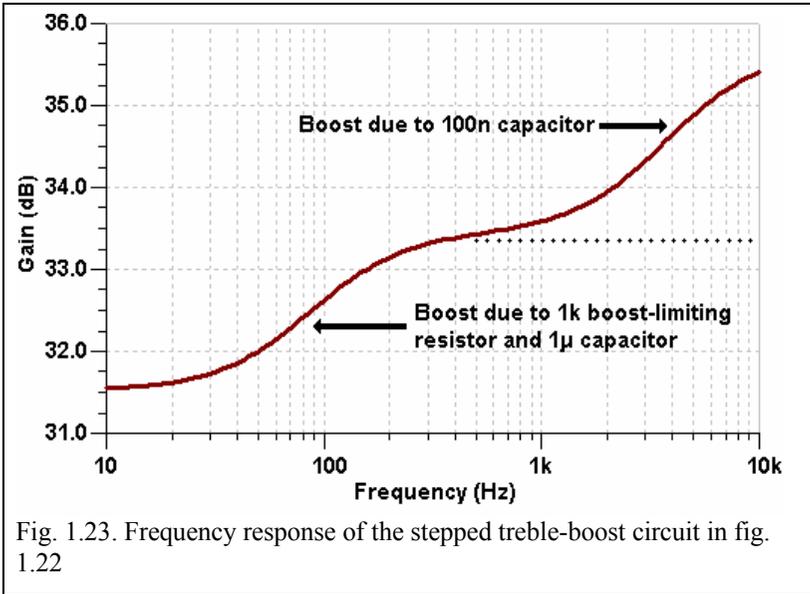


Fig 1.23 illustrates how this has modified the previous circuit's response. At very high frequencies the full gain of the stage is now realised, and the stage has a 'stepped' treble boost. For comparison, the dotted line shows the response of the previous circuit.

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.

### The common cathode, triode gain stage



#### **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**.

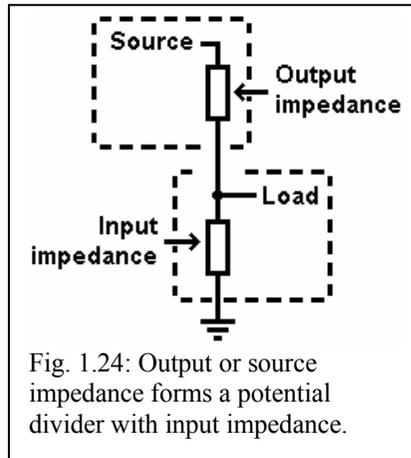


Fig. 1.24: Output or source impedance forms a potential divider with input impedance.

#### **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 ( $R_g$  in fig. 1.25), which may be any desired value provided it does not exceed the published maximum. Usually  $1M\Omega$  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

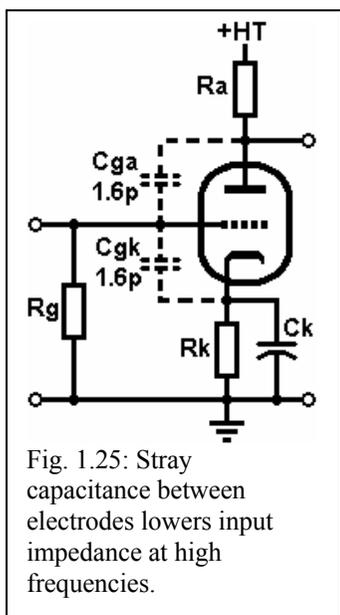


Fig. 1.25: Stray capacitance between electrodes lowers input impedance at high frequencies.

total input capacitance, but it is not so simple. The published values for the ECC83 are indicated in the circuit in fig. 1.25.

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

$$C_{in} = C_{gk} + C_{ga} \cdot A \dots\dots\dots VII$$

Where:

$C_{gk}$  = grid-to-cathode capacitance\*

$C_{ga}$  = 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  $C_{in}$  will be:

$$C_{in} = 1.6 + (1.6 \times 58) = 94.4pF$$

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  $C_{ga}$ , 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 in giving a smoother treble sound.

\* Some datasheets may not provide  $C_{gk}$  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  $R_a$  and  $r_a$ , so the output impedance is:

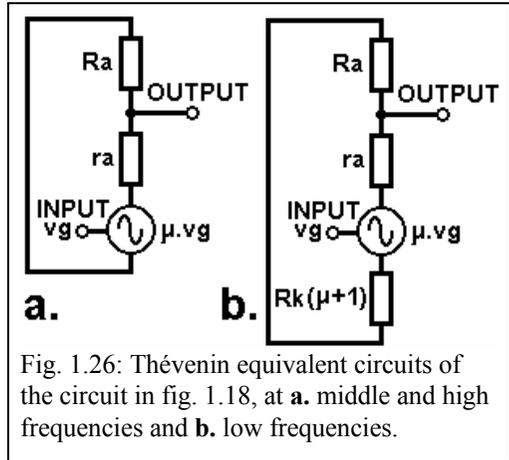


Fig. 1.26: Thévenin equivalent circuits of the circuit in fig. 1.18, at **a.** middle and high frequencies and **b.** low frequencies.

$$Z_{out (R_k \text{ bypassed})} = R_a \parallel r_a \dots \dots \dots \text{VIII}$$

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

$$\begin{aligned} Z_{out} &= 100k \parallel 70k \\ Z_{out} &= \frac{100k \times 70k}{100k + 70k} \\ &= 41k\Omega \end{aligned}$$

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  $R_k$  is now in series with  $r_a$ . Because  $R_k$  is below the cathode its value appears multiplied by  $\mu+1$ . Therefore the output impedance when the cathode is unbypassed becomes:

$$Z_{out (R_k \text{ unbypassed})} = R_a \parallel (r_a + R_k(\mu+1)) \dots \dots \dots \text{IX}$$

## Designing Valve Preamps for Guitar and Bass

If  $R_k$  is  $1.5k\Omega$  and  $\mu$  is 100 then:

$$Z_{out (R_k \text{ unbypassed})} = 100k \parallel (70k + 1.5k \times (100 + 1))$$

$$Z_{out (R_k \text{ unbypassed})} = 100k \parallel 251.5k$$

$$Z_{out (R_k \text{ unbypassed})} = \frac{100k \times 251.5k}{100k + 251.5k}$$

$$= 72k\Omega$$

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 than  $100\mu A$ ). 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

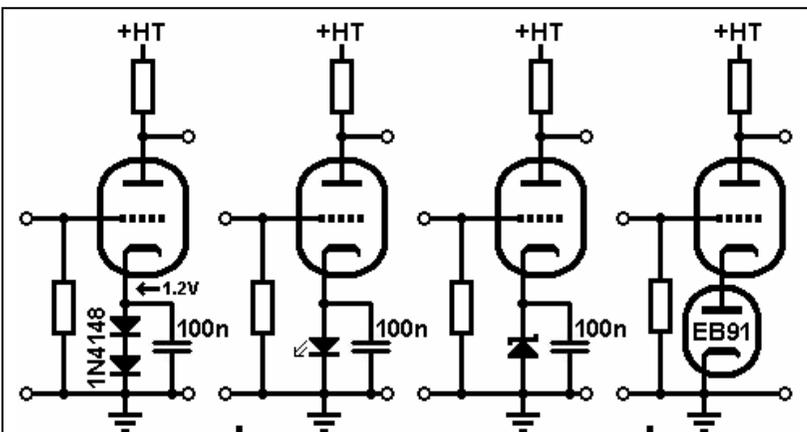


Fig. 1.27: Cathode biasing using: **a.** silicon signal diodes, **b.** an LED, **c.** a zener diode, **d.** a valve diode.

### The common cathode, triode gain stage

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.

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. 1.27d 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.

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.

**Summary of formulae:**

I; Voltage gain in decibels:

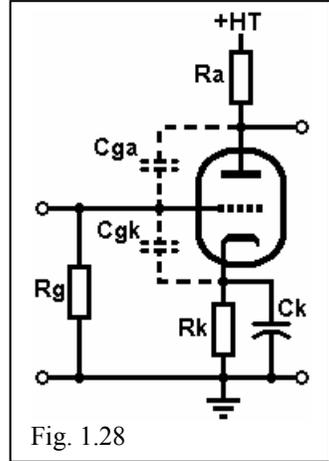
$$A_{(dB)} = 20 \log \frac{V_{out}}{V_{in}}$$

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

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

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

$$A = \frac{-\mu R_a}{R_a + r_a + R_k(\mu + 1)}$$



V; Voltage gain at any frequency:

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

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

$$f_{(half\ boost)} = \frac{1}{2\pi R_k C_k} \cdot \sqrt{1 + \frac{R_k(\mu + 1)}{2(R_a + r_a) + \frac{1}{2} R_k(\mu + 1)}}$$

Solving for Ck gives:

$$C_k = \frac{1}{2\pi f R_k} \cdot \sqrt{1 + \frac{R_k(\mu + 1)}{2(R_a + r_a) + \frac{1}{2} R_k(\mu + 1)}}$$

Where:

f = the desired half-boost frequency.

If  $R_a + r_a > R_k(\mu + 1)$  then to a reasonable approximation:

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

Input impedance (before grid current):

$$Z_{in} = R_g$$

The common cathode, triode gain stage

VII; Total input capacitance:

$$C_{in} = C_{gk} + C_{ga.A}$$

VIII; Anode output impedance (cathode resistor bypassed):

$$Z_{out} = R_a \parallel r_a = \frac{R_a \cdot r_a}{R_a + r_a}$$

IX; Anode output impedance (cathode resistor unbypassed):

$$Z_{out} = R_a \parallel r_a + R_k(\mu + 1) = \frac{R_a [r_a + R_k(\mu + 1)]}{R_a + r_a + R_k(\mu + 1)}$$

Where in all cases all notations are as in fig. 1.28 and;

$r_a$  = internal anode resistance of valve at operating point.

$\mu$  = amplification factor of valve.

All resistances in ohms. All capacitances in farads.