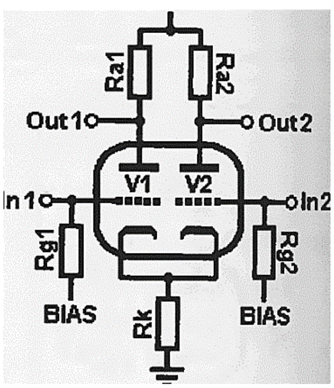
Chapter 8: The Long-Tailed-Pair Phase Inverter

The long-tailed pair is the most popular phase inverter circuit found in guitar amps. It is also known as a differential amplifier, cathode-coupled inverter or Schmitt inverter after Otto Schmitt described its operation in 1938 although, the conventional form of the circuit can be traced as far back as 1934. The cathodyne phase inverter described in chapter 7 is well suited to driving relatively sensitive power tubes to high levels of overdrive but if we wish to drive larger tubes to similar levels of distortion then we require large signal amplitudes. This could be obtained from a cathodyne operating under a high B+ voltage but this is not always available. The long-tailed-pair offers a greater output swing for a given B+ but requires two tubes which will usually be in the same envelope. It also has more gain, although it should be pointed out that in combination with an additional gain stage, a cathodyne offers more total gain with the same two tubes. It also seems to be less sensitive to triode types producing a good tone for many different triodes. The 12AY7 is a popular choice for low to moderate gain designs and has a particularly rich tone. The ECC82/12AU7 is suitable for very clean or bass amplifiers.

Basic operation of the long-tailed pair:

A simplified diagram of a long-tailed pair is shown in fig. 8.1; it can be seen that the circuit has two inputs plus ground and two outputs plus ground, making it a six-pole, entirely balanced or push-pull circuit. The plate current of each tube flows into the shared cathode resistor Rk which forms the tail of the circuit.
If we input a positive signal to V1, it's plate current increases. This current flows through Rk causing the voltage across it to increase, which is effectively the same as the grid voltage of V2 decreasing, so the plate current through V2 decreases. V2 operates the opposite of V1 with antiphase signals appearing at each output; V1 acts a common gain stage and a cathode follower.



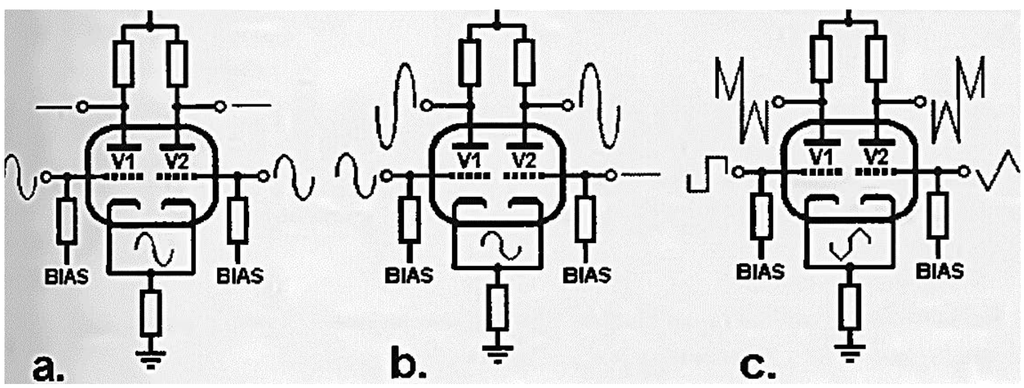
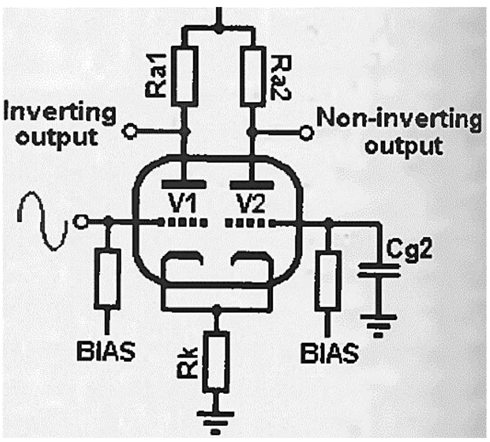


Fig. 8.1: Simplified diagram of the long-tailed pair.
Fig. 8.2: The long-tailed pair only amplifies the difference between its inputs. In a. there is no difference between inputs so no output signal is produced. b. the circuit as a simple phase inverter. c. the circuit as a differential mixer and phase inverter.

Schmitt, O. H. (J 938). Cathode Phase Inversion, Journal of Scientific Instruments 15 (March), pp 100-101.
Matthews, B. H. C. (1934). A Special Purpose Amplifier. The Journal of Physiology, 81, 28P-29P.

The output from its cathode is passed to the cathode of V2, which amplifies in a non-inverting fashion. The output from V1 is out-of-phase with the input of V1, while the output of V2 is in phase with the input of

V1, as a phase inverter should. The same arguments follow if we had input the signal to V2. In a perfectly balanced circuit the output signal amplitudes will be equal, which is only possible if each tube sees one half of the total input signal; the signal appearing at the cathode must be one half the amplitude of the input signal.
However, suppose we input the same positive signal to both tubes simultaneously. Both tubes attempt to increase their plate current. This flows in Rk, increasing the voltage across it; each cathode attempts to follow the corresponding grid. Under ideal circumstances the cathode voltage rises by the same amount as the voltage on each grid, so the total change in the grid-to-cathode voltage of each tube is zero. Therefore, there is no signal for the tubes to amplify and therefore no output. From this we discover that the long-tailed-pair will only amplify the difference in voltage between its inputs, which is why it is also known as a differential amplifier. When we input the same signal to both grids the difference between the two is zero, so nothing is amplified. This property is known as Common-Mode Ripple Rejection or CMRR. It is one of the most desirable features of balanced amplifiers, since it discriminates between the wanted signal(s) and common-mode hum and noise; noise which is the same or common to both inputs.
We could also input two completely different signals. The circuit would mix the two by amplifying the difference between them and provide two out-of-phase output signals. Hence the long-tailed pair can be used as a phase inverter and as a phase-inverter/mixer, as shown in fig. 8.2.



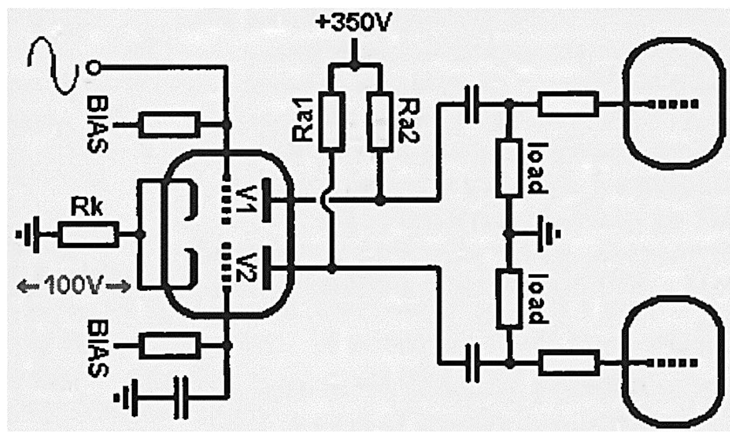


Fig. 8.3: Simplified diagram of the long-tailed pair as a phase inverter.
Fig. 8.4: A simplified circuit arrangement with 100V allowed across Rk.

Fundamental parameters of the long-tailed pair:

In most guitar amps the stage is used simply as a phase inverter in which the second grid is grounded as far as AC is concerned by a decoupling or bypass capacitor Cg2 and the input signal is applied to the first grid. This is shown simplified in fig. 8.3 and by referring to fig. 8.2b we see that the output from V1 is inverted, while that from V2 is non-inverting. If the second grid were not bypassed then the input signal may find its way onto the second grid and we would have no difference between the inputs with no output at all or very unbalanced output signals, depending on the circuit arrangement used.

Gain:

Because V1 also act as a cathode follower with close to unity gain, not all of the input signal is passed to V2, so the gain of each output may not be identical and the stage will be unbalanced. It can be shown that for identical triodes, if Ra1 = Ra2 = Ra, the gain to the non-inverting output is:

$$A2 = \frac{\mu Ra}{(Ra+ra)\times\left(2+\frac{Ra+ra}{Rk(\mu+1)}\right)} \qquad \mathbf{55}$$

But the gain to the inverting output will be greater than this by a factor of:

$$\frac{A1}{A2} = 1 + \frac{Ra+ra}{Rk(\mu+1)} \qquad \mathbf{56}$$

From which we see that if the plate resistors are equal, the overall balance depends on the value of Rk and

perfect balance is achieved when Rk is infinite. If Rk is not infinite then balance is improved by using high-μ tubes. We could also use mis-matched plate resistors to bring about equal gain to each output which will be described later.
Alternatively, we could measure the output signal between both outputs, resulting in the differential gain. Assuming perfect balance this equals:

$$A = \frac{\mu Ra}{Ra + ra}$$

Which is the same as formula 3 for a common-cathode gain stage. The gain of either output of a long-tailed pair is roughly half that which would be obtained from the same, single triode, if used as a common gain stage with the same plate resistor. If a cathodyne were coupled to the same, common gain stage then that gain would be measured at both outputs, which demonstrates why it actually offers more gain from the same two tubes.

Valley, G. E. & Wallman, H. (eds.) (1948). Vacuum Tube Amplifiers (1st ed.), p447. McGraw-Hill, New York.

Input capacitance:

The total input capacitance is given by:

$$Cin \approx Cgk(1 - Ak) + Cga \times Aa \qquad \textbf{57}$$

Where:
Ak = the gain from grid to cathode, being approximately 0.5
Aa = gain to the anode, which will be approximately $Aa \approx \frac{\mu Ra}{2(Ra+ra)}$
And since $Cgk(1 - Ak)$ is very small, this may be approximated to: $Cin \approx Cga \times Aa$

Input impedance:

In the circuit of fig. 8.3, the input impedance is equal to the grid-leak resistor. In other circuit arrangements, the input impedance can be made much greater.

Output impedance:

If the circuit is perfectly balanced then the output impedance is the same as that of a common gain stage, given by formula 8:

$$Zout = Ra \| ra = \frac{Ra \times ra}{Ra + ra}$$

However, if the stage is unbalanced then the output impedance at the highest gain, will increase Zout. However, the change is not much, reaching a maximum of:

$$Zout = Ra \| ra + R$$

Where:

$$R = \frac{Ra + ra}{1 + \frac{Ra + ra}{Rk(\mu + 1)}}$$

Jones, M. (1999). Valve Amplifiers (2nd ed.), p 116. Newnes. Oxford.

If Rk and μ are large then R ≈ Ra + ra and the output impedance may be simplified to:

$$Zout \approx Ra \| 2ra + Ra = \frac{Ra}{2} \qquad \textbf{58}$$

Input sensitivity:

Half the input signal appears at the cathode of a long-tailed pair so, the input sensitivity is halved when compared to the same triode, used in a conventional gain stage with a fully bypassed cathode. For example, a typical ECC83/12AX7 gain stage has in input sensitivity of about 4Vp-p when center-biased. But when connected as a long-tailed pair this is halved to about 8Vp-p, making the long tailed pair more difficult to overdrive (i.e. it has more headroom); this is more sensitive than a cathodyne.

Designing a long-tailed pair:

For guitar purposes, the design of the long-tailed pair is quite straight forward, since precise balance is not required. In fact, the small amount of imbalance which is inherent in most designs encourages even-harmonic distortion in both the phase inverter and the power output stage which produces greater texture to the tone; this is in contrast to the technical advice published in some guitar magazines that state that perfect balance is essential. Normally, it is in the nature of push-pull circuits to cancel out even-harmonic distortion, since asymmetrical non-linearity in one device is matched by the same non-linearity in the opposing device but with reversed phase so that the distortion products are equal but opposite and cancel out when they are mixed together in the output transformer. Distortion in a perfectly balance long-tailed pair using triodes is dominated by third-order intermodulation distortion, with harmonic distortion appearing at a lower level. Depending on the amount of imbalance however, odd and even harmonic distortion can be increased relative to the intermodulation products. This develops the characteristic growly sound of the long-tailed pair. However, too much imbalance can result in the opposite effect, making the tone too flat.

An easy way to design a long-tailed pair is to begin by deciding how much voltage can be spared for the tail resistor, Rk. A greater voltage implies a larger value for Rk, which improves the inherent balance of the circuit but also reduces the available output swing. Obviously, the greater the B+ then the more voltage drop can be tolerated across Rk. A clean amplifier can spare the most voltage of about 20-50% of B+, since a large output swing is not necessary and it is more likely to benefit from a well balanced phase inverter, whereas a moderate to high-gain amp might allow about 10-30% of B+.

Abuelma'atti, M. (2005). Large Signal Analysis of Differential Triode Tube Amp, Audio Xpress (online). Available from: http://www.audioxpress.com/magsdirx/ax/addenda/media/muhammad.pdf

For example, if the B+ is 350V we might allow 100V for Rk, leaving 250V of effective B+ for the tubes, as shown in fig. 8.4. Since we know the output swing of a triode is typically about 2/3 B+, we can predict about 170Vp-p output. This is certainly enough to overdrive almost any power tube.

We can now select a suitable load and draw a load line using this effective B+ of 250V. For example, if we were to use an ECC83/12AX7 we might use a typical load of 100KΩ. The load line drawn in fig. 8.5 suggests about 170Vp-p maximum output swing however, it is worth considering the additional load placed on the circuit by the power tubes' grid-leak resistors (labeled load in fig. 8.4).

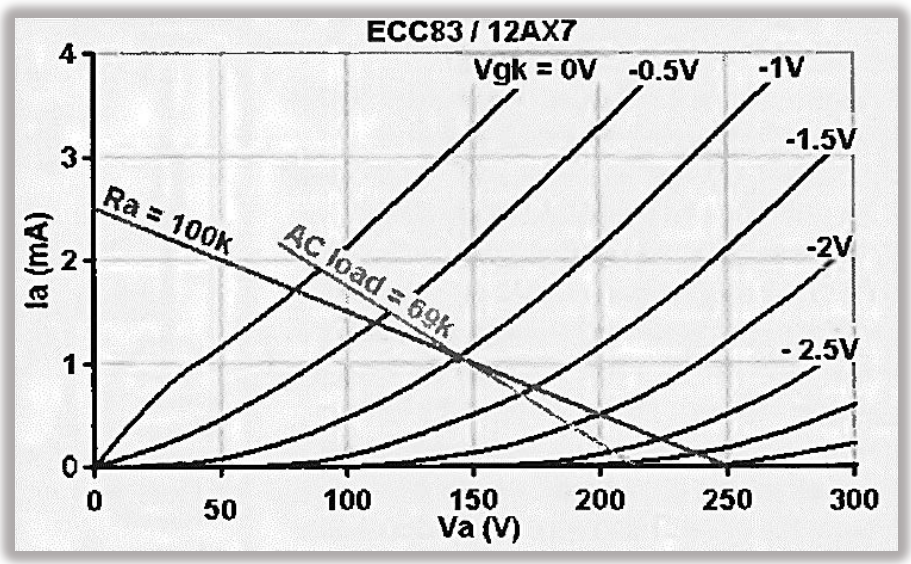
Normally we are free to ignore such following loads but the maximum value of grid-leak for a power tube is much less than for a preamp tube. If the grid-leaks were 220KΩ each, which would be typical if we were driving EL34s or 6L6GCs, then the AC load on each triode is 100K ∥ 220K = 69KΩ. Choosing a bias point of Vgk = -1V the AC load line is shown in fig. 8.5, which shows that the actual output swing is reduced to about 140Vp-p. Fortunately, this is enough for most purposes.

Cold biasing will encourage cut-off clipping in the phase inverter but at the expense of output swing. The result is that the output tubes may be driven less with more phase-inverter overdrive being heard, shifting the tonal balance in favor of preamp distortion rather than power amp distortion. Warm biasing has the opposite effect; even though grid-current clipping will happen sooner in the phase inverter, this produces a descending output swing which will not be amplified by the corresponding power tube since it is in cut-off, unless the output stage is Class A.

Having selected a bias point, the quiescent plate current can be found. In this case, fig.8.5 indicates about 1.1mA. However, the current flowing in Rk is the sum of the currents in both triodes, resulting in twice this value or 2.2mA. Since we initially allowed 100V across Rk and we now know the current though it to be 2.2mA, applying Ohm's law calculates its value as: R = V/I = 100/0.0022 = 45.5KΩ. So a typical value of 47KΩ can be selected. Fig. 8.6 shows the circuit partially completed. Since the required bias is Vgk = -1V and the cathode voltage is 100V, we require a grid voltage of 99V…all that remains is to bias the circuit.

DC-coupled long-tailed pairs:

Because the required grid voltage of a long-tailed pair is often relatively high, it is a suitable candidate for DC-coupling. This is a common arrangement in hi-fi amps, popularized by the Mullard 5-10 and 5-20 designs but is comparatively rare in guitar amps, the Sound City LB series being notable exceptions.



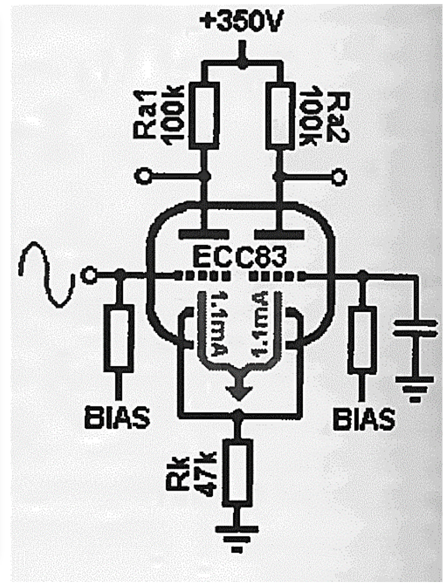
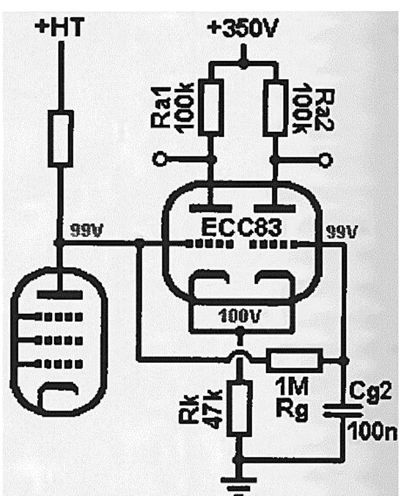
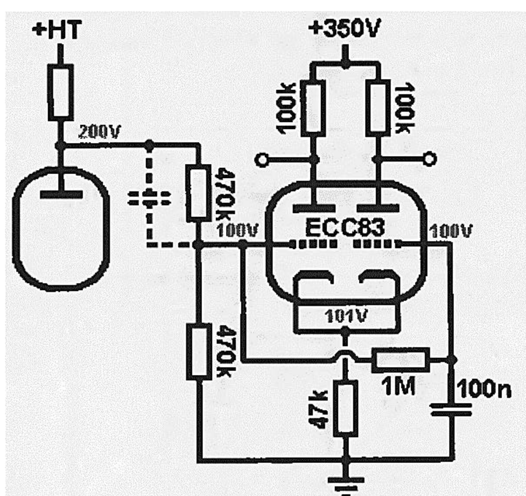


Fig. 8.5: DC, AC and cathode load lines drawn for an effective B+ of 250V with a 100KΩ plate resistor and 220KΩ following load resistance.
Fig. 8.6: A partially completed design showing plate current paths.

In this example the desired grid voltage is 99V. This might be available directly from the plate of the previous stage, especially if it is a pentode; it is taken directly from a cathode follower in the Sound City LB amps. This voltage must be applied to both grids to ensure the same bias; a pull-up resistor Rg, is connected between the grids so, the second grid is pulled up to the same DC voltage as the first. Since the second grid is to be bypassed, this resistor will form an AC load on the previous stage, so its value should be large but should not exceed the maximum allowable value of grid-leak.





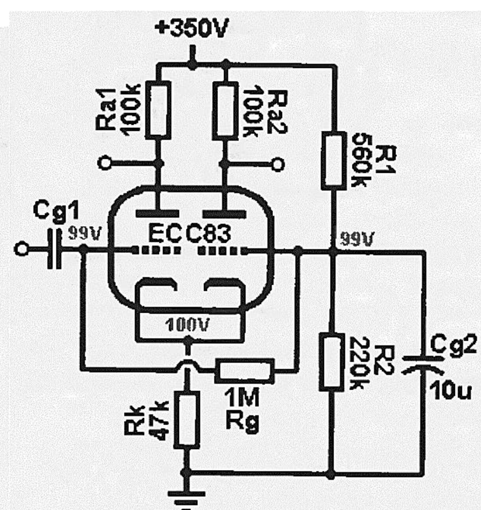


Fig. 8.7: A direct-coupled, long-tailed pair, typical of hi-fi designs.
Fig. 8.8: A level-shifted, DC-coupled long-tailed pair.
Fig. 8.9: A fixed-biased long-tailed pair.

Assuming we wish the second grid to be fully bypassed at all frequencies down to 1Hz, then we may find the value of Cg2:

$$C = \frac{1}{2\pi fR} = \frac{1}{2\pi \times 1 \times 1M} = 159nfd$$

A value of 150nfd might be chosen, although a value of 100nfd will decouple down to 1.6Hz and is the usual choice. This is shown in fig. 8.7, the grid-stopper and arc-protection diode are omitted for clarity. At power-on, Cg2 may be exposed to full B+ voltage so should have a voltage rating greater than the B+.
It was noted earlier that unless Rk is virtually infinite, the gain to the inverting output will be greater than the gain to the non-inverting output by a factor of:

$$\frac{A1}{A2} = 1 + \frac{Ra+ra}{Rk(\mu+1)}$$

In this case the difference in gain is:

$$\frac{A1}{A2} = 1 + \frac{100K+65K}{47K\times(100+1)} = 1.03$$

The inverting triode has 3% more gain than its partner, which is not a great difference. However, if we were concerned about this imbalance we might correct it by reducing Ra1 to: 100K/1.03 = 97KΩ. Perfect AC balance is obtained when:

$$\frac{Ra2}{Ra1} = 1 + \frac{Ra2+ra}{Rk(\mu+1)} \qquad \textbf{59}$$

In practice this modification is not required for guitar purposes.
Level shifting is also a convenient way to couple the long-tailed pair, particularly if the required grid voltage is rather low. Fig. 8.8 shows an example in which the previous stage has a plate voltage of 200V. This is level-shifted down to 100V, which is sufficiently close to the design value of 99V. If a treble boost cap is used then an additional grid-stopper is advisable. We should replace Ra with the value of Ra ∥ R1, where R1 is the following loading resistance; however, in practice the error is negligible.
DC-coupled, long-tailed pairs are rarely found in a guitar amp and this is usually because it is preceded by a tone stack or volume control which requires AC coupling to keep DC off the pots which would otherwise cause loud scratching noises when adjusted.

Clare, J. D. (1947). The Twin Triode Phase-Splitting Amplifier. Electronics Engineering, (February), pp62-63.

AC-coupled long-tailed pairs:

Designing an AC-coupled, long-tailed-pair follows essentially the same process as described earlier except, the input is isolated by a coupling capacitor so, we must apply the required grid voltage by another method.

Fixed bias:

The required grid voltage could be applied to the first (input) grid but since the second grid is already bypassed, it is more convenient to apply the bias at this point, as illustrated in fig. 8.9; note It's similarity to the fixed-biased cathodyne in fig. 7.5b. R1 and R2 form a potential divider which provides 99V to the second grid, while Rg pulls up the first grid to the same voltage. The bypass capacitor Cg2 must be large since, the effective resistance supplying the second grid has been reduced to R1 ∥ R2 ∥ Rg or:

$$R = \frac{1}{\frac{1}{R1}+\frac{1}{R2}+\frac{1}{Rg}} = \frac{1}{\frac{1}{560K}+\frac{1}{220K}+\frac{1}{1000K}} = 136\Omega$$

Cg2 decouples the grid at audio frequencies and also filters out any B+ noise. If we wish to filter out frequencies as low as 1Hz then:

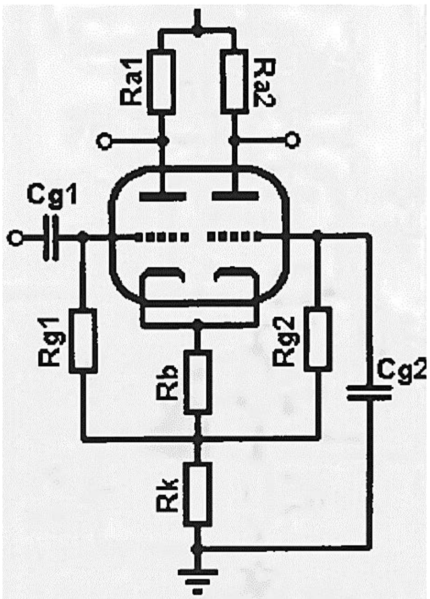
$$C = \frac{1}{2\pi fR} = \frac{1}{2\pi \times 1 \times 136K} = 1.2\mu fd$$

So a large value of 10µfd will provide good filtering and need only have a voltage rating somewhat greater than the grid voltage, 160V. The input coupling capacitor is chosen in the usual way, in combination with the input impedance which is equal to Rg.
This form of fixed-biased, long-tailed pair is rarely seen in audio amplifiers but is included here for completeness. Its lack of popularity may be because the cathode-biased version offers a higher input impedance with the same number of components and does not require a large electrolytic, grid bypass

capacitor.

Cathode bias:

This is the most common form of long-tailed pair. The bias is applied in the same way as the cathode-biased cathode follower and cathodyne circuits. The necessary grid voltage is tapped off the tail resistance by splitting it into two parts and connecting the grids to the junction through grid-leak resistors, as shown in fig. 8.10. We could go to the trouble of ensuring that Rk + Rb equal the total tail resistance that was initially specified but in practice the bias resistor Rb, is small enough that there is no need and can be stacked on top of Rk.



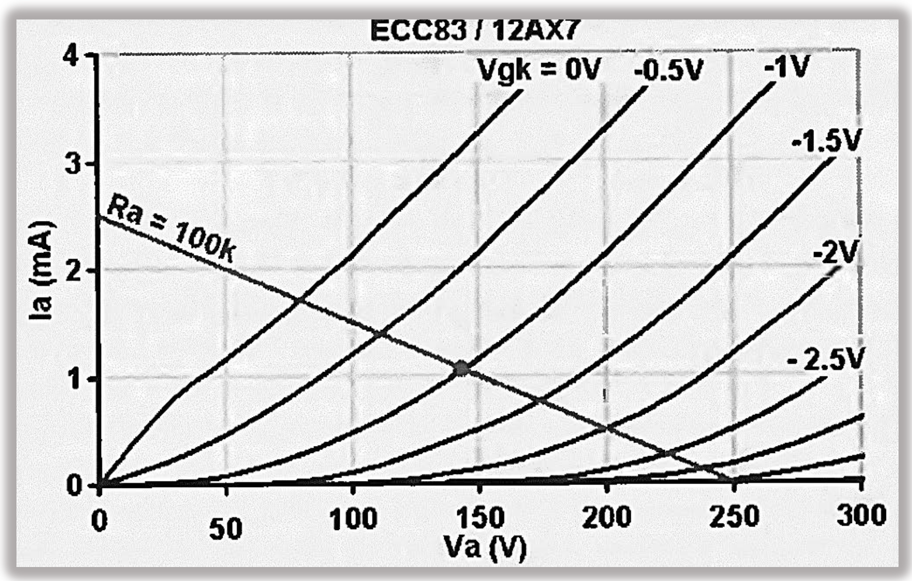


Fig. 8.10: The cathode-biased long-tailed pair is the most common phase-inverter arrangement found in guitar amplifiers.
Fig. 8.11: A 100KΩ load line reproduced from fig. 8.5. With a bias point of - 1V, the plate current is 1.1mA.

The load line for the earlier design is reproduced in fig. 8.11. Using the same bias point of Vgk = -1V; the plate current appears to be about 1.1mA. However, the current from both triodes flows in the tail resistance so, this must be doubled to 2.2mA. The necessary bias resistor is then: R = V/I = 1 / 0.0022 = 455Ω , so use 470Ω.
Half the input signal also appears at the cathode with most of this also appearing at the junction of Rb and Rk (the bottom of the grid-leak resistors), so Rgl and Rg2 are both bootstrapped.
Assuming proper balance and Rg1 = Rg2 = Rg, the input impedance to each grid becomes:

$$Zin = \frac{Rg}{1-\frac{Rk}{2(Rk+Rb)}} \qquad \mathbf{60}$$

Usually Rk/(Rk+Rb) ≈ 1, so this may be simplified to: Zin ≈ 2Rg
So using 1MΩ grid-leaks provides an input impedance close to 2MΩ; 470KΩ resistors could be used for less noise with an input impedance of 940KΩ. The input coupling capacitor Cgl may be chosen in the usual manner, using the calculated input impedance. Often, it is large, with frequency shaping being performed by an earlier inter-stage coupling network or tone stack.
The bypass capacitor for the second grid, Cg2, is chosen according to:

$$Cg2 = \frac{1}{2\pi f Zin}$$

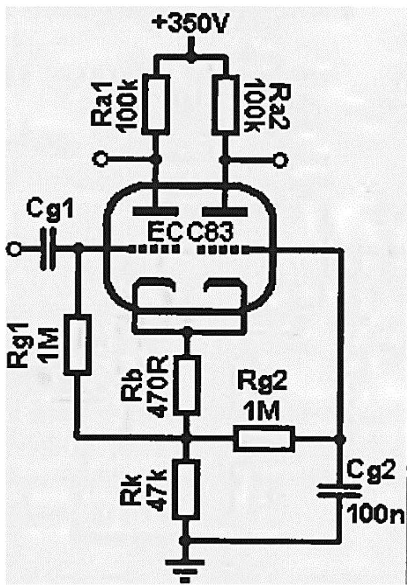
In this case, using 1MΩ grid stoppers provides an input impedance of:

$$Zin = \frac{Rg}{1-\frac{Rk}{2(Rk+Rb)}} = \frac{1000K}{1-\frac{47K}{2\times(47K+0.47K)}} = 1.98M\Omega$$

Normally we expect the second grid to be bypassed well below audible frequencies, down to 1Hz:

$$Cg2 = \frac{1}{2\pi f Zin} = \frac{1}{2\pi\times 1\times 1980K} = 80nfd$$

So the nearest common value of 100nfd is suitable. The completed circuit is shown in fig. 8. 12. This is very similar to the phase inverter used in the Vox AC30, except that it was arranged as a mixer/inverter for mixing the normal and tremolo channels, as shown in fig. 8.13.
Note that when feeding a signal into only one input, if the volume control of the opposite input is not turned down, then the opposite grid will not be properly decoupled and the balance may suffer. In practice however, the imbalance is not much more than +/-1 dB because the potential divider formed by the opposite grid-leak and corresponding volume control attenuates the AC signal appearing on the opposite grid.



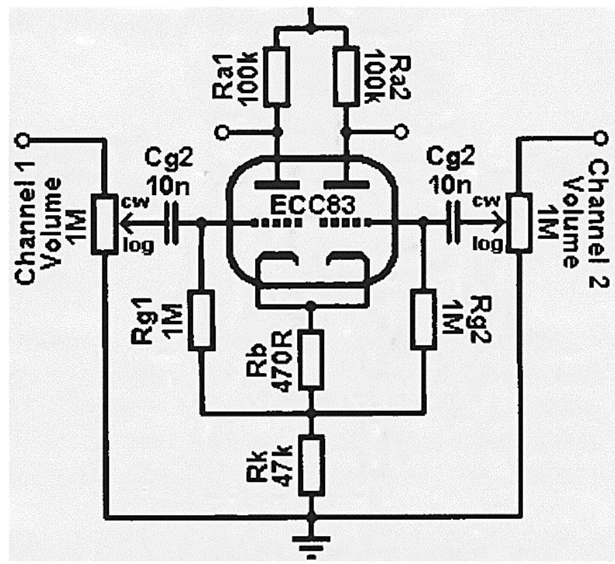


Fig. 8.12: A completed cathode biased, long-tailed pair phase-inverter.
Fig. 8.13: The cathode-biased long-tailed pair as a mixer/inverter, similar to that used in the Vox AC30 and Marshall 1974 18W.

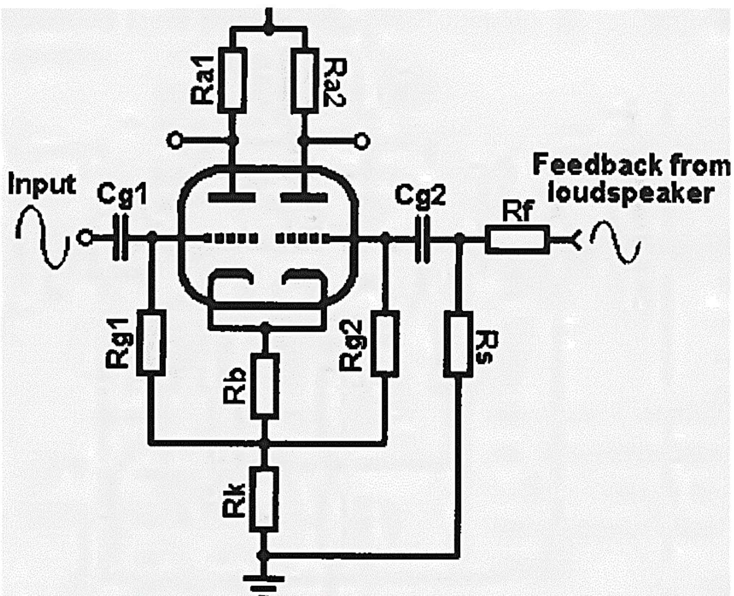
Global negative feedback and the long-tailed pair:
The previous sections have shown how to design the essential parts of a long-tailed pair. In guitar amplifiers it is common to find global negative feedback (NFB) applied to it so the subject deserves special attention. This feedback is usually taken from the secondary of the output transformer and helps to linearize the power output stage, increasing its bandwidth and flatten the frequency response. It also reduces the effective input sensitivity of the stage to which feedback is applied, increasing the headroom and making the amplifier more difficult to overdrive, which is suited for cleaner styles of playing. This is dealt with in greater detail in chapter 9. Here we describe simply how to apply this feedback and it effects on the phase inverter.
The feedback signal is normally applied to the second grid of the circuit. The feedback must be negative…it should oppose the input signal. If it was positive feedback to the input signal, it would make the amplifier oscillate uncontrollably, resulting in a wild, full-volume howl. Since the long-tailed pair is a differential amplifier, the feedback signal must be in phase with the input signal at the first grid. In this way the difference in voltage between the two grids will be reduced and so the apparent gain of the phase inverter will be reduced. In an extreme situation, if the feedback signal could be made equal to the input signal then there would be no difference between the two grids and we would obtain no output signal. In reality this is impossible since no output would also mean no feedback signal but it illustrates the phase relationship between the two grids.
The simplest arrangement is to take the feedback from the secondary of the output transformer/loudspeaker and apply it to the second grid via a potential divider Rf and Rs (fig. 8.14), which determines the amount of feedback applied. Although this is an acceptable circuit, anyone familiar with existing guitar amp designs will notice that this does not look like a typical circuit arrangement.
By moving Rs so that it forms part of the tail resistance, as shown in fig. 8.15, the circuit performance can be improved slightly. Since the input and feedback signals are assumed to be substantially similar except in

amplitude, the signal appearing at the cathode will also have the same shape. Since the feedback signal is also being applied to part of the tail resistance we now have similar signals both at the top and the bottom of the tail; Rb and Rk have been bootstrapped. The effective tail resistance is increased from Rb + Rk + Rs to a new value of:

$Rt = (Rk + Rb)\frac{Vin}{Vin - Vfb} + Rs$ or:

$Rt = (Rk + Rb) \times 10^{dB/20} + Rs$ where: dB = the amount of feedback in decibel.

In theory, increasing the effective tail resistance helps improve the balance of the phase inverter. The feedback signal supplied to the tail resistor will also help in maintaining balance when the plates are unequally loaded (e.g. when the power tubes are being overdriven) but the circuits used in guitar amps are rarely optimized for this. In fact, this form of phase inverter often produces worse balance than that in fig. 8.14.



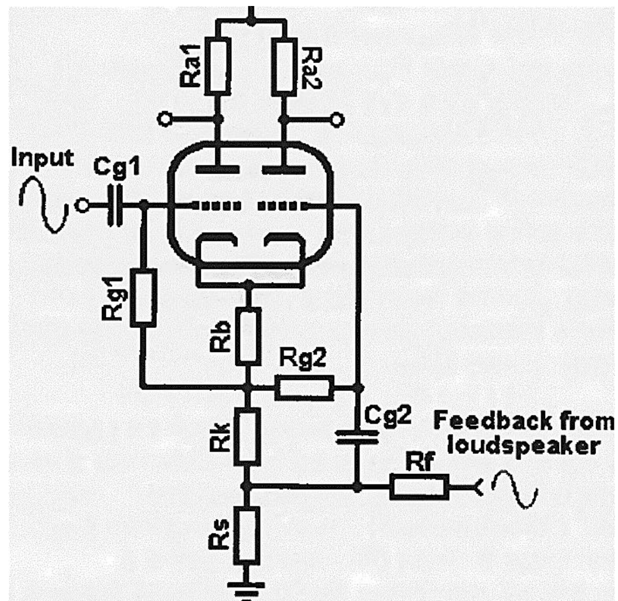


Fig. 8.14: Applying negative feedback to the long-tailed pair from a simple potential divider formed by a feedback resistor Rf and shunt resistor Rs.
Fig. 8.15: Applying negative feedback to the second grid and tail, bootstraps the tail resistance and is the most common form of long-tailed pair found in guitar amplifiers.

This has become the most common form of long-tailed pair used in guitar amplifiers and was first adopted by Fender in 1958 for the new versions of Bassman and Twin Amps and later popularized in classic Marshall amplifiers; it has become an industrial standard despite its design flaws. These are flaws in the conventional hi-fi sense. For musical instruments, expectations are relaxed and the circuit still produces good tonal results. In practice, about 4 - 10dB of global negative feedback is used; the gain of the phase inverter is reduced by 4 - 10dB or 30 – 60% of its full level before feedback is added. The feedback signal should be between 0.4 - 0.7 times the input signal. This is set by the potential divider formed from Rf and Rs, which means we must know the amplitude of the signal appearing at the loudspeaker, for a given input signal to the long-tailed pair; this can be measured directly or it may be calculated.

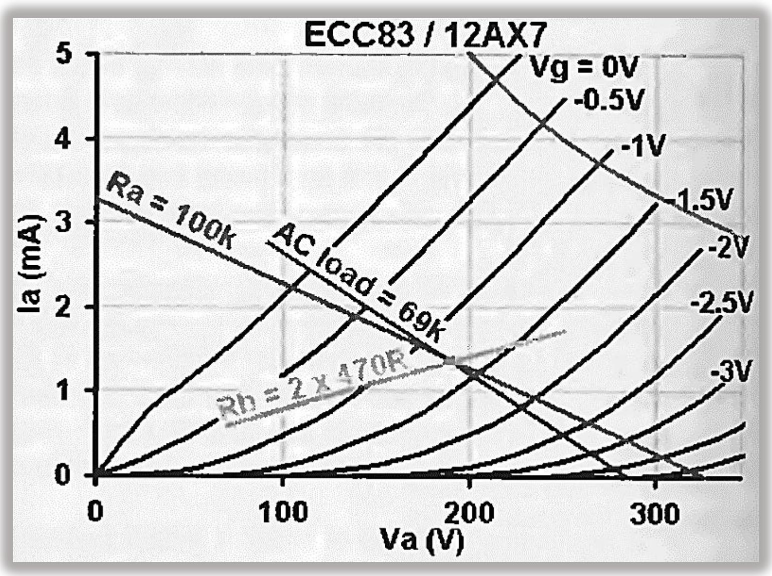
The Fender 5F6-A Twin Amp:

In the Fender 5F8-A Twin Amp, only 27V was allowed for the tail resistor. The B+ was 355V, providing an effective B+ of 328V and 100KΩ loads were chosen for use with an ECC83; fig. 8.16 shows the load lines. A 470Ω bias resistor was used but since twice the current flows in this resistor, a cathode load line of 2 × 470 = 940Ω is drawn. The expected bias point shows a quiescent current of l.35mA. Since twice this or 2.7mA flows in the tail resistor, a value of 27 / 0.0027 = 10KΩ was used. This is relatively small, so the imbalance of the circuit is more significant. The gm at the bias point is 1.6mA/V and ra is about 62.5KΩ, so the difference in gain between the two triodes is:

$$\frac{A1}{A2} = 1 + \frac{Ra+ra}{Rk(\mu+1)} = 1 + \frac{100K+62.5K}{10K\times(100+1)} = 1.16$$

In other words, the gain of the first triode is 16% greater than that of the second. This could be corrected by reducing Ra1 to: 100K / 1.16 = 86.2KΩ. The nearest standard is 82KΩ, which is what Fender used. The grid-leak resistors are set at a value of 1MΩ and Cg2 is made the usual value of 0.1µfd.

All that remains is to apply the negative feedback and this requires knowing the gain of the amplifier from the phase inverter to speaker. This could be measured directly or if we know or guestimate the maximum output-power before clipping, it may be calculated. The Fender 5F8-A Twin Amp used 4 × 5881 power tubes and quoted an output power of 80W. The 5881s were biased at -50V, so require $50/\sqrt{2}$ = 35Vrms of drive signal for full output. The gain to one output of the phase inverter is half that indicated by the AC load line or about 26. For maximum, unclipped output power, the input signal to the phase inverter will be: 35 / 26 = 1.35Vrms.



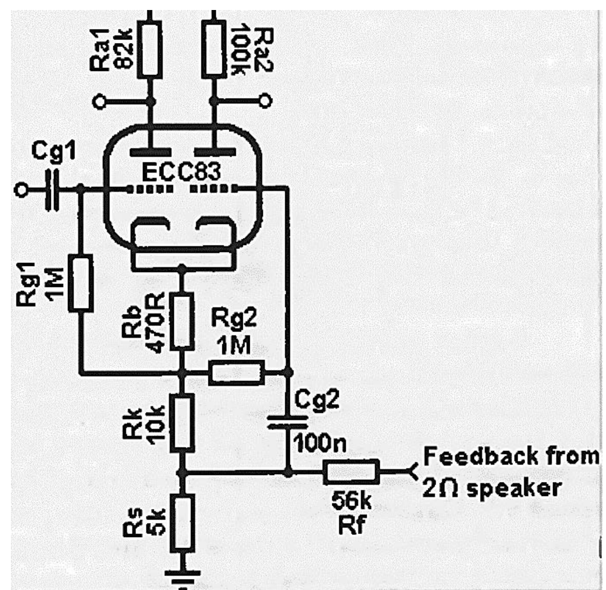


Fig. 8.16: Plate and cathode load lines drawn for an effective B+ of 328V.
Fig. 8.17: The basic, long-tailed pair circuit, first used in the Fender 5F8-A Twin Amp.

If this develops 80W into a 2Ω speaker (the original amp used two 4Ω speakers in parallel), then the voltage at the speaker must be:

P = V^2/R, $V = \sqrt{PR} = \sqrt{160} = 12.6Vrms$

We now have sufficient information to choose the feedback components. Fender applied 12.5dB of feedback. This is a relatively large amount but unsurprising since it was designed for pedal-steel players who required a very clean tone. The ratio of input signal to feedback signal is given by:

$$\frac{Vin}{Vfb} = 1 - 1\frac{1}{10^{dB/20}}$$

Where:
Vin = the input signal amplitude.
Vfb = the feedback signal amplitude.
dB = the amount of feedback in decibels.
So for 12.5dB of feedback:

$$\frac{Vin}{Vfb} = 1 - \frac{1}{10^{12.5/20}} = 0.76$$

In other words, for every 1V of input signal, 0.76V of feedback signal is returned to the second grid. It has already been calculated that for full power, 1.35Vrms of input signal is required. The feedback signal at this

level will be: 1.35 × 0.76 = 1.03V. This is applied from the speaker via a potential divider formed by the feedback resistor and shunt resistors Rf and Rs respectively. Since the voltage at the speaker is 12.6V and the required feedback signal is 1.03V, the gain of this divider must be: β = 1.03 / 12.6 = 0.082.
The absolute value of these resistors is not important, except that they must be large with respect to the speaker impedance. Rs is usually made relatively small with respect to Rk, so that it does not significantly alter the previously calculated tail resistance. However, Fender selected a rather large value of 5KΩ for Rs. The value of Rf must be:

$$Rf = \frac{Rs - Rs \times \beta}{\beta} = \frac{5K - (5K \times 0.082)}{0.082} = 56K\Omega.$$

Note that the value of the feedback resistor depends on the power output and speaker tap from which feedback is taken; we cannot simply shoehorn this circuit into a different amplifier without altering Rf or Rs to suit the new conditions. If in doubt, do not use any feedback at all, it is not essential. The completed circuit is shown in fig. 8.17. The high values of Rf and Rs were reduced in later designs, which also reduces noise and improves the stability of the feedback loop since it lowers the influence of stray capacitances.
The most significant change in the operation of the phase inverter is a decrease in input sensitivity due to the addition of feedback; the feedback signal effectively cancels out part of the input signal. By adding 12.5dB of feedback, the input sensitivity is decreased by the same degree or 4.2 times. The load line in fig. 8.16 indicated an initial input sensitivity of around 3.5Vp-p but remembering that for a long-tailed pair the headroom is twice that indicated, this results in 7Vp-p. But the addition of feedback has reduced this to: 7 × 4.2 = 29.4Vp-p, so the phase inverter is now much more difficult to overdrive. This additional headroom is one of the main reasons for using negative feedback. However, by placing a potential divider before the phase inverter without feedback can produce the same amount of headroom without any instability. Since good fidelity is rarely a requirement for most musicians, global negative feedback in guitar amplifiers is less useful today than it was in the 1950s.
Another interesting point to note about this original circuit is the use of the 82KΩ and 100KΩ plate resistors. Because a relatively low tail resistance was used, this mis-match did serve to improve the balance of the circuit before feedback was added. However, if the tail resistance is increased to slightly over 16KΩ, the difference in gain between the two triodes becomes small enough that better balance is actually obtained with identical 100KΩ plate resistors. Virtually all versions of this circuit produced today descend from this original design so, the 82KΩ:100KΩ combination is still seen in designs even when much larger tail resistances are used, verifying the fact that most modern manufacturers copy parts of existing designs, rather than design from scratch. Historically, this appears to have begun with the CBS-Fender designs and has perpetuated ever since.

The presence control:

It may be noted that the circuit in fig. 8.17 is missing a presence control. This simple control effectively removes upper middle and treble frequencies from the feedback signal by shunting them to ground through a capacitor. By removing the feedback at these frequencies the gain of the phase inverter returns to its normal level, providing an active boost in the upper middle and treble frequencies. The earliest designs replaced Rs by a pot, with the shunt capacitor connected to the wiper, as shown in fig. 8.18a. However, since DC flows in this resistor, when operated the control develops a loud scratching sound, so the modified version in fig. 8.18b should be used, allowing C1 to block DC from the pot.
P3, is normally larger than Rs so that it does not alter the calculated amount of feedback; this is not critical, since this resistance will reduce feedback slightly, which is much safer than increasing it.
For example, if Rs is 100Ω then P1 might be 4.7KΩ. If the presence control is turned fully CCW (0Ω), then the +3dB transition frequency between the flat response and the boosted level is:

$$f = \frac{1}{2\pi \times C1(Rf \| Rs)} = \frac{1}{2\pi C1\left(\frac{Rf \times Rs}{Rf + Rs}\right)}$$

In the case of the Fender 5F8-A example, Rf = 56KΩ, Rs = 5KΩ and a value of 100nfd was used for C1, providing a boost frequency of:

$$f = \frac{1}{2\pi \times 100 \times 10^{-9} \times \left(\frac{56K \times 5K}{56K + 5K}\right)} = 347Hz$$

This covers almost all the middle range frequencies and above. Other manufacturers design for frequencies in the range of 350 - 600Hz. Interestingly, some guitarists find the presence control invaluable and go to great lengths to add one to amplifiers which don't already have one, while other players find the presence control to be a nuisance. This may be due to some amplifiers using so little feedback that the effective boost has little audible effect. This formula breaks down if less than 6dB of feedback is used, although it will still be sufficiently accurate for most practical purposes.

Variable feedback:

The effects of negative feedback are detailed in chapter 9. Because these effects are not to everyone's tastes, a useful option is to make the feedback variable between its full level and zero. For users who find the presence control of limited use, it may be converted into a feedback control by increasing the value of the presence capacitor Cl (fig. 8.18) so, that it shunts virtually all the feedback signal to ground; this is shown in fig. 8.19a.

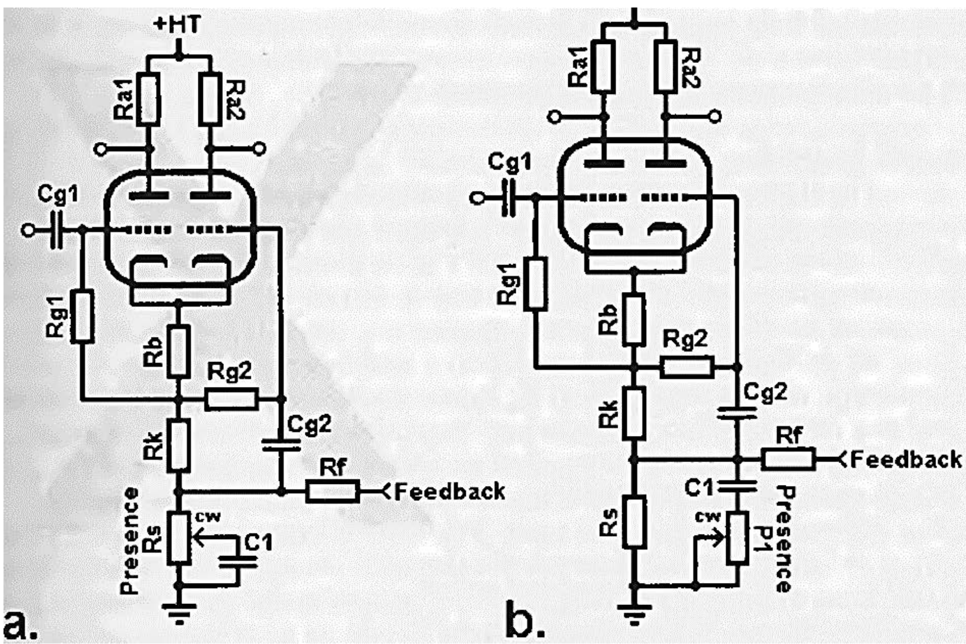
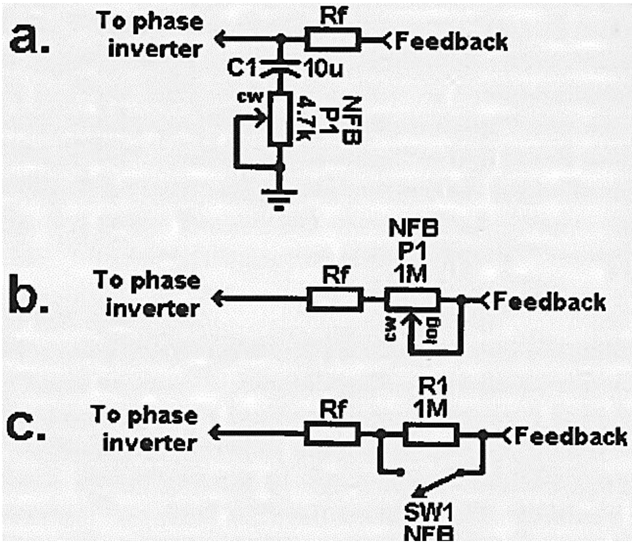


Fig. 8.18: Long-tailed pairs with presence control. The early model design in a. produces an obnoxious scratching sound due to DC on the pot so, the circuit in b. is preferred for new designs.

Alternatively, variable feedback may be implemented by connecting a pot, wired as a variable resistor, in series with the feedback resistor, as shown in fig. 8.19b; this effectively increases the feedback resistance. However, due to the relatively small amount of feedback used in guitar amplifiers, often there is little audible difference in the tone over much of the control's range. For many users, the ability to switch between full feedback and virtually no feedback is sufficient and this is shown in fig. 9.19c. Note that R1 is included so that the feedback loop is never broken, reducing popping sounds. Its value is made sufficiently large, reducing feedback to almost nothing.



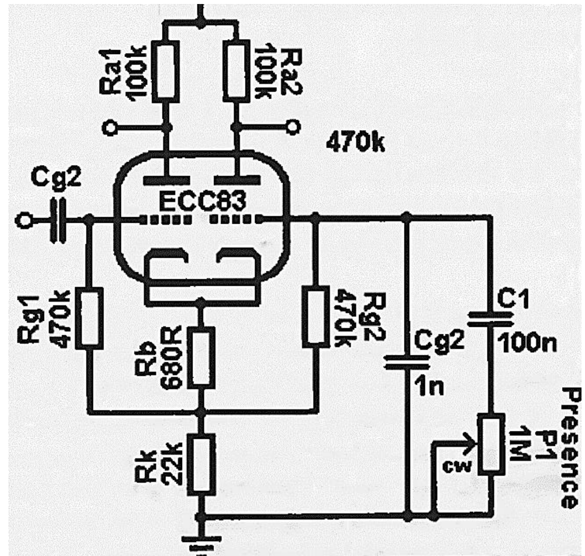


Fig. 8.19: Methods of implementing variable feedback, a: modifying an existing presence control by increasing C1. b: adding a variable resistance in series with Rf. c: switching between two levels of feedback.
Fig. 8.20: A presence control for an amplifier without global negative feedback.

Presence control without global negative feedback:

Global negative feedback may not suit all playing styles and can also increase instability (see chapter 9). For builders who are only interested in adding a presence control, the use of feedback can seem like an unfortunate necessity. Thankfully, the long-tailed pair phase inverter allows the traditional, feedback presence control to be emulated. By manipulating the size of the second grid's bypass capacitor we can alter the gain of the circuit to some extent, enough that it can be turned into a presence control as shown in fig. 8.20. The grid-bypass capacitor Cg2, is made quite small so that only middle and treble frequencies are bypassed, allowing these frequencies to achieve full gain. A larger capacitor Cl, is connected in parallel so that all frequencies can be bypassed, while the pot in series, allows the degree of bypass at low frequencies to be controlled. Effectively, this is a bass-cut control with the same tonal effect as a conventional presence control; the frequency response is given in fig. 8.21.

In this case the tail resistance is greater than 16KΩ so identical 100KΩ plate resistors have been used, although some players may prefer the tonal effect caused by using the 82KΩ/100KΩ imbalance. In any case, the tail resistance and plate load resistances are not specific to the presence control. The maximum degree of bass cut is determined by the potential divider formed by Rg2 and Pl. Rg2 is bootstrapped to an effective value of:

$$Zin = \frac{Rg}{1-\frac{Rk}{2(Rk+Rb)}} = \frac{470K}{1-\frac{22K}{2\times(22K+0.68K)}} = 913K\Omega$$

P1 will be 1MΩ, therefore the maximum degree of bass cut is:

$$\beta \approx 1-\frac{P1}{P1+Zin} = 1-\frac{1000K}{1000K+913K} = 0.47\ or -6.4dB.$$

This illustrates that using smaller grid-leak resistors, increases the maximum degree of cut. We may also note that the degree of cut does not depend upon the type of tubes used but only on the tail and grid-leak resistances.

This control could easily be applied to any existing design where only one input of a long-tailed pair is used and no global negative feedback is applied.

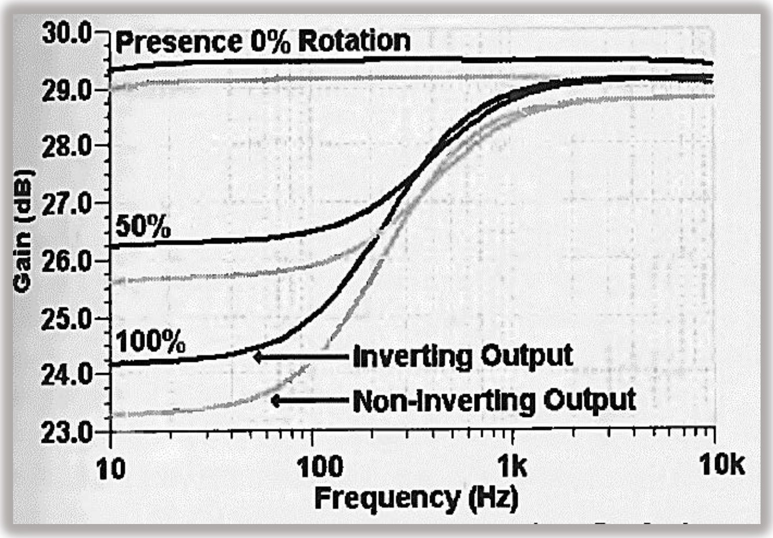


Fig. 8.21: Frequency response of the circuit in fig. 8.20.

Scale control:

The gain and output swing of the long-tailed pair can be controlled by varying it's bias. It might be thought that the tail resistance would have more effect on the latter but in practice this does not seem to be the case; varying Rk has very little effect on the characteristics and tone of the circuit, provided it is not smaller than 10KΩ. By varying Rb, operation can be controlled from its normal mode of moderate gain and high output swing, to less gain and much less output swing; this will serve as a pseudo-power-scaling effect. By increasing the bias the output from the phase inverter can be reduced to a few volts, while the phase inverter itself can be overdriven more, reducing the volume of the amp while retaining the same amount of perceived overdrive levels. Three possible circuits for achieving this are shown in fig. 8.22.

The circuit in fig. 8.21a is perhaps the most obvious and easy to install in an existing circuit, in which the bias resistor is replaced with a large variable resistance in the region of 10K - 47KΩ. This allows the bias to be varied from zero: high gain and output swing, to a high value: low gain and output swing. The rest of the phase inverter is conventionally designed.

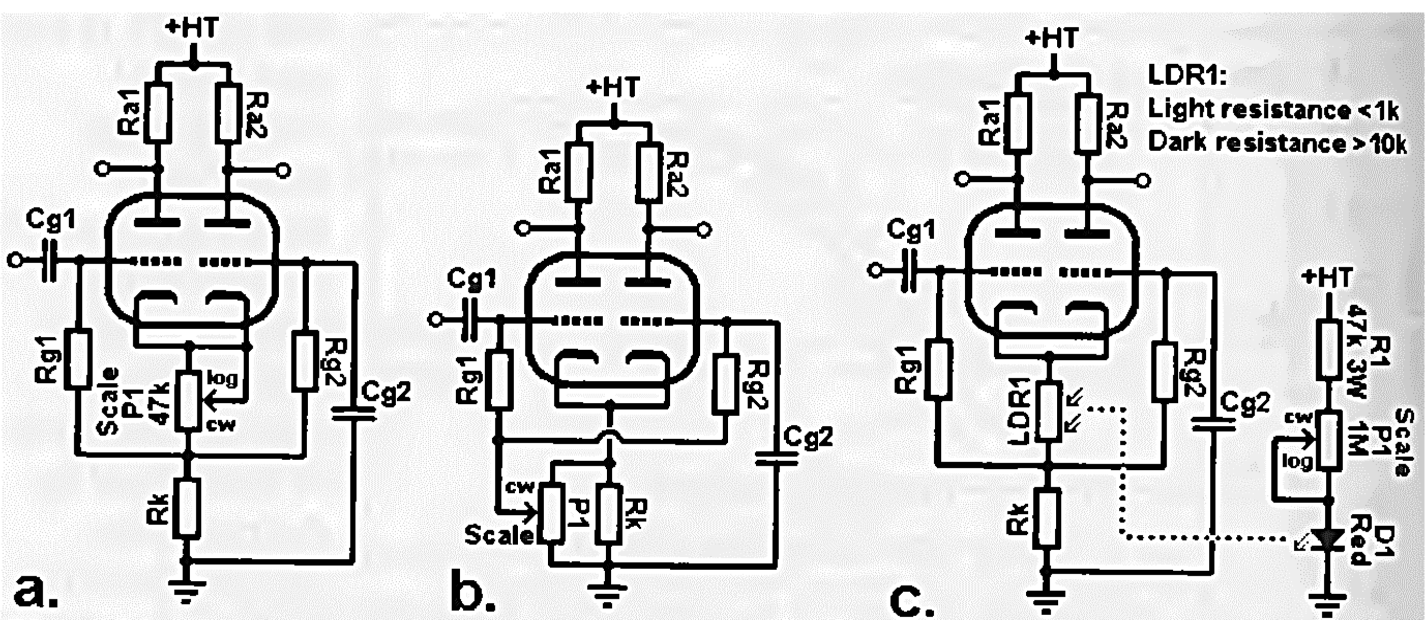


Fig. 8.22: Three possible scaling circuits, a. and b. are simple but will produce a scratchy sound when operating, while c. varies the bias remotely through an opto-isolator.

The circuit in b. removes the bias resistance altogether and a high-value pot is connected in parallel with Rk, the combined resistance of the two, equaling the original tail resistance. The grid bias voltage is now tapped off the tail resistance. This method has the advantage that the total tail resistance no longer varies with the bias setting, which is preferable if global negative feedback is applied. This is the same circuit as the limiting control used in the Carlsbro 60TC. A high value pot, 100K - 470KΩ should be used so that most of the tail current flows only in Rk, with little power being dissipated in Pl.

Both the previous circuits have DC on a pot which will cause scratching sounds when operated, although many users may be willing to suffer this, considering how useful this control is. For a new design though, it may be preferable to vary the bias remotely. One such example is shown in fig. 8.2lc. This is essentially the same as the circuit in a., except that the bias resistance is replaced with a light-dependent resistor (LDR) whose resistance is varied by an LED. This filters out the DC changes in the circuit as Pl is rotated, reducing spurious noises. R1 determines the maximum current and brightness for the LED and this can be reduced to almost nothing with P1. Most LEDs will be sufficiently bright at around 5 - 8mA current. As shown, the LED is powered from the B+ but any convenient DC supply will do, if R1 is of proper value. The LED and LDR should be mounted facing one another and sealed from external light with heat-shrink tubing. Alternatively, an opto-isolator can be used. In typical circuits, any of these methods will allow the output swing to be varied from normal to a couple of volts peak-to-peak, heavily overdriven.

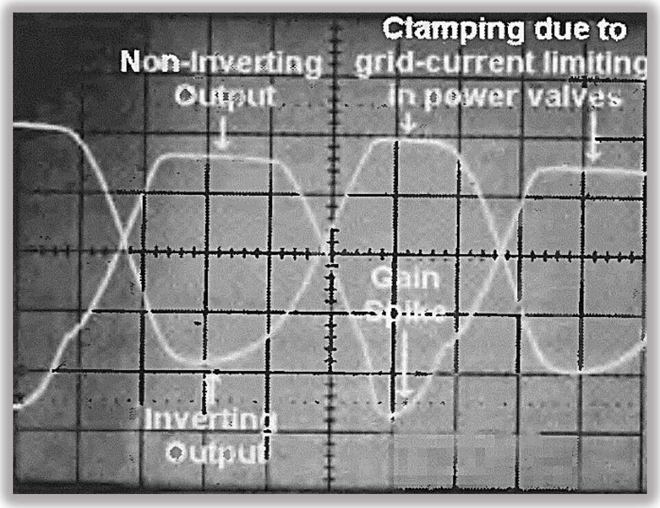
Overload characteristics of the long-tailed pair:

The overload characteristics of phase inverters is of particular importance in guitar amps because they are consistently overdriven. It was shown in the previous chapter how some quite unexpected waveforms can be produced by a cathodyne phase inverter, which are rarely desirable and given it a bad reputation in the community. It is true that the long-tailed pair is slightly more predictable in this regard but is not immune. The large resistance in the cathode circuit of the long-tailed pair makes it prone to the frequency-doubling effect (see chapter 7, fig. 7.12) and swirl.

Fortunately, this is only an issue if the tail resistance is very large, more than 100KΩ or if it is replaced with a constant-current sink, which is rarely the case in a guitar amp. However, it may surprise some folks to learn that the long-tailed pair does suffer from the gain-spike effect first described in chapter 7 (fig.7.11), though for different reasons. This spike occurs at the non-inverting output when the inverting output voltage is clamped, a result of overdriven power tubes. When the inverting-output signal is large enough, it will begin to overdrive the corresponding power tube. As the power tube begins to draw grid current it will clamp the plate voltage of the inverting plate, so that it cannot rise any further even though the input signal to the long-tailed pair is still moving negatively. Since the plate voltage remains steady, it appears grounded for AC conditions; the input triode now acts as a pure cathode follower. Normally only half of the input signal appears at the cathode and is passed to the second triode but now that the first triode is acting as a cathode follower with near unity gain, almost all of the input signal appears at the cathode. This is passed to the second triode and amplified, causing its output signal to suddenly increase, creating the gain spike. This is seen in the oscillogram in fig. 8.23 which was produced from the circuit in fig. 8.12, while overdriving a pair of EL84s without grid- stoppers.

If the output stage is class AB, then this negative gain spike is not always a concern because the corresponding power tube is virtually cut-off when the negative spike occurs. However, if the power tubes are poorly matched or if the power output stage is class A, then this spike will be amplified by the corresponding power tube and will introduce high-order intermodulation distortion products which are unpleasant.

As with the cathodyne, this effect may be overcome or eliminated by using larger grid stoppers on the power tubes. Since the power tubes are typically pentodes or beam tetrodes, they’ll have very low input capacitance so, large grid-stoppers can be used without affecting the treble response and values of 100KΩ will virtually eliminate the gain-spike effect with high certainty. An example of this application, occurs in the Marshall Studio 15. Larger grid-stoppers greatly reduce the chance of blocking distortion and can also be used to assist in stabilizing global negative feedback (see chapter 9, fig. 9.11). For most classes of amplification, larger grid stoppers of about 10KΩ are recommended to help maintain consistency if the power tubes do not age equally.



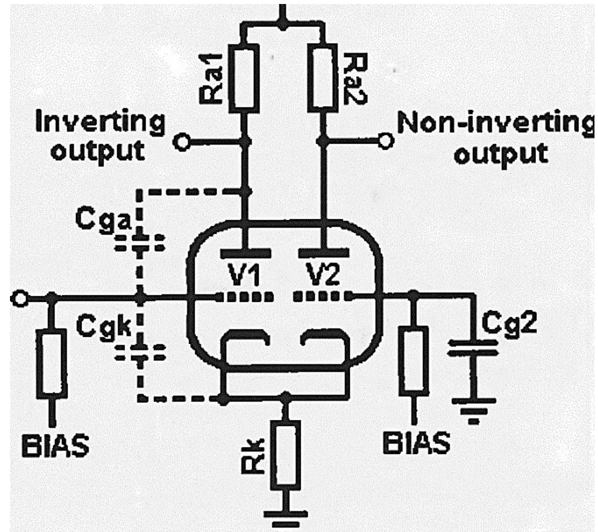


Fig. 8.23: Oscillogram showing the gain spike effect when the power tubes are overdriven, 1KHz, 5V/div.
Fig. 8.24

Summary of formulae:

55: Gain to non-inverting output assuming a symmetrical circuit where Ra1 = Ra2 = Ra:

$$A2 = \frac{\mu Ra}{(Ra+ra)\times\left(2+\frac{Ra+ra}{Rk(\mu+1)}\right)}$$

56: Difference in gain between outputs:

$$A = \frac{A1}{A2} = 1 + \frac{Ra+ra}{Rk(\mu+1)}$$

Differential gain assuming perfect balance:

$$A = \frac{\mu Ra}{Ra+ra}$$

57: Total input capacitance: $Cin \approx Cgk(1 - Ak) + Cga \times Aa$ where:

Ak = the gain from grid to cathode, being approximately 0.5.

Aa = gain to the anode, which will be approximately: $Aa \approx \frac{\mu Ra}{2(Ra+ra)}$

But since Cgk (1 – Ak) is very small, this may be approximated as: $Cin \approx Cga \times Aa$

58: Output impedance assuming unequal loading:

$$Zout \approx Ra \| 2ra + Ra = \frac{Ra}{2}$$

Output impedance for equal loading:

$$Zout = Ra \| ra = \frac{Ra \times ra}{Ra+ra}$$

59: Perfect AC balance is obtained when:

$$\frac{Ra2}{Ra1} = 1 + \frac{Ra2+ra}{Rk(\mu+1)}$$

60: Input impedance of the circuit in fig. 8.25 where Rg1= Rg2 = Rg:

$Zin = \frac{Rg}{1-\frac{Rk}{2(Rk+Rb)}}$ usually Rk/(Rk+Rb) ≈ 1, so this may be simplified to: $Zin \approx 2Rg$

Where in all cases notations are as in figs. 8.24 and 8.25.

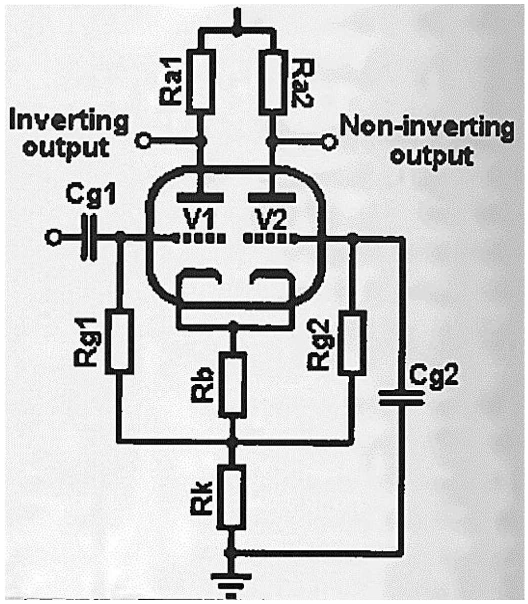

Ra1
Ra2
Inverting output
Non-inverting output
Cg1
V1
V2
Rg1
Rb
Rg2
Cg2
Rk

Fig. 8.25