Stabilizing Tube Feedback Amplifiers Without Math: Part 1

By Merlin Blencowe. This article first appeared in AudioXpress, May 2026.

When global negative feedback is wrapped around a tube amplifier it promises measurable improvements; less distortion, wider bandwidth, lower output impedance etc. When applied thoughtfully it makes a good amplifier better. But these improvements come at the cost of stability; the more feedback we apply, the greater the chance our amplifier will become an oscillator. Judging by the many internet forums I frequent it seems many DIYers apply feedback with no more precautions than crossed fingers. Feedback theory is well covered in tube textbooks and articles[1,2,3] but admittedly it can be difficult to interpret and to apply to new situations. The notation used in older books can be opaque, the techniques were optimised for slide rules, and they focus a lot on Nyquist diagrams which have never been very intuitive. Frustratingly, they often fall short of giving examples that include a transformer inside the loop.

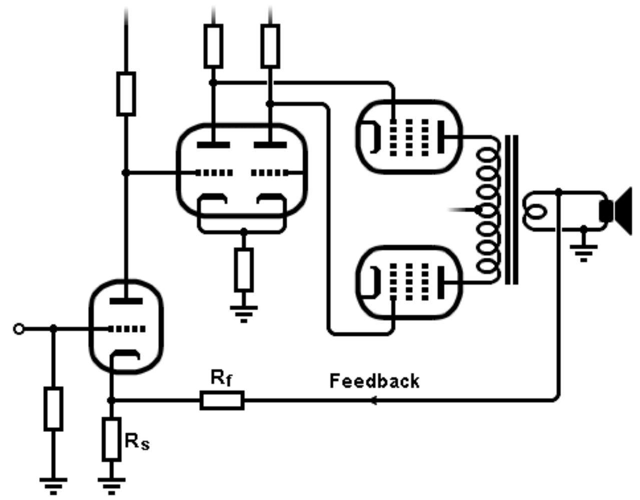

Fig. 1: Typical non-inverting feedback architecture.

In this article I hope to cover the essentials of feedback design for tube power amplifiers in a digestible and practical way using amplitude and phase plots. Bode plots will also be used when it is helpful, i.e. where amplitude curves are approximated using straight lines of specific slope (phase curves do not conform so nicely to straight line approximations so I will show only true phase responses). I will also show how the stability of a tube amplifier can be assessed by practical bench methods, and how stability can be ensured no matter how much feedback you might want to use –and what is sacrificed in the process. Nevertheless, I expect readers to be familiar with basic principles of amplifier design already, or this article would never end.

There are many ways to implement feedback but in this article I will focus on the familiar situation where it is taken from the output transformer secondary, i.e. across the loudspeaker, and is applied to an earlier stage though a simple resistor divider, something like fig. 1. This is a particularly troublesome arrangement because the output transformer is inside the loop, and transformers add phase shift in a very inconvenient way at high frequencies. Fortunately, one advantage of a tube amplifier is that it can be tested open-loop, before adding any feedback, so we can get reliable information about its gain and phase on the bench.

The Universal Feedback Equation

Readers ought to be familiar with the universal feedback equation, and for the purposes of this article I choose to write it:

$$A_c = \frac{A_o}{1 + A_o B} \tag{1.1}$$

A_o is the open-loop gain; the gain before any feedback is applied;
A_c is the closed-loop gain; the gain after feedback has been applied;
B is feedback fraction; the fraction of the output signal that we feed back.

The denominator $1+A_oB$ is called the *feedback factor* since it is the factor by which the overall gain of the system is reduced; it is the amount of feedback being used. It is normal to express the feedback factor in decibels and to talk about an amplifier as using so-many dBs of feedback.

The term A_oB is called the *loop gain*, since it is the total gain experienced by the feedback signal as it passes around the loop. It is the gain we measure by breaking the feedback loop, injecting a signal into the appropriate side of the break, and observing the resulting signal appearing on the other side of the break. The loop gain is equal to the amount of feedback being applied, minus one. For example, if we apply 20dB of total feedback, i.e. $1+A_oB = 10$, then the loop gain is 9, or 19.1dB, and we would adjust the feedback divider to produce this much loop gain. It doesn't matter if the open-loop gain of the amplifier is 100 or 100000, if we apply 20dB of feedback the loop gain is always 9, or 19.1dB (though for large amounts of feedback we can often ignore the 'minus one' difference, since it is small).

[1] Cooper, G. F. (1950). Audio Feedback Design, *Radio Electronics*, October, pp49-50.
[2] Crowhurst, N. (1956). Stabilizing Feedback Amplifiers, *Radio Electronics*, December, pp34-7.
[3] Keroes, H. I. (1959). Stabilizing Feedback Amplifiers, *Radio Electronics*, January, pp45-7.

Stability problems arise at the extreme ends of the amplifier bandwidth –especially the high end– because there are many hidden filters in the circuit, all of which introduce phase shift. At sufficiently high (or very low) frequencies the total phase shift in a multi-stage amplifier is sure to exceed 180°, at which point negative feedback becomes positive feedback. If the loop gain equals unity when this happens, then $A_oB = -1$ and the universal feedback equation blows up to infinity: the circuit will oscillate. Even if outright oscillation does not occur, marginal stability often reveals itself as a resonant peak in the frequency response, and poor transient behaviour, observable as ringing on square-waves. For the best audio we want a flat, well-behaved frequency response, even outside the audible range.

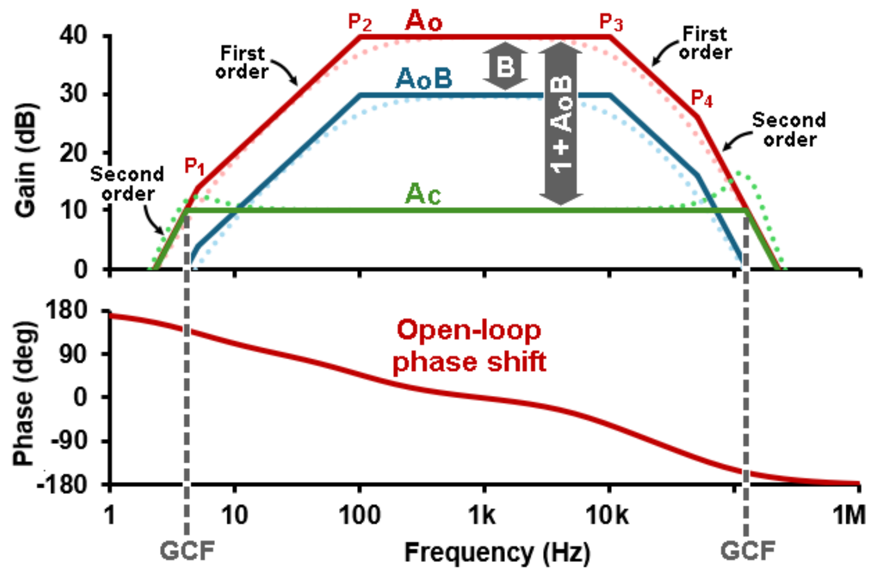

Fig. 2: The universal feedback equation in the form of amplitude and phase plots.

A better grasp of the feedback equation can be acquired by looking at gain and phase plots. Fig. 2 shows the response of a fictitious amplifier, with Bode approximation in bold, and the actual response dotted. Starting with the red line (A_o) we see the amplifier has a mid-band gain of 40dB. It also happens to have four cut-off frequencies or 'poles', P_1 to P_4, created by various unavoidable time constants within the imaginary circuit. Each pole contributes 45° phase shift and tends to 90° at infinitely low or high frequencies, leading at the low end, lagging at the high end. Each pole also adds 20dB/decade attenuation. In this case the response therefore ultimately falls at 40dB/decade (second order) at the frequency extremes, and phase shift approaches ±180°.

If 30dB of feedback is applied then the gain of this amplifier is reduced to 10dB and its bandwidth becomes much wider, shown by the green line (A_c). The vertical difference between red and green is therefore the feedback factor $1+A_oB$. It is now obvious that we really only have 30dB of feedback in the mid band; it is less for frequencies outside of P_2 and P_3.

The point where A_c intersects with A_o is called the *gain crossover frequency* (GCF) and defines the closed-loop bandwidth. This amplifier has two GCFs, one for the high end and one for the low end, whereas a DC amplifier would have no lower limit. Beyond the GCF there is essentially no effective feedback so A_c eventually returns to the original A_o curve. It is at the GCF(s) that an amplifier might try to oscillate, so this area gets all our attention when it comes to assessing stability.

The blue line shows the loop gain (A_oB). Assuming feedback is applied with a simple resistor divider with no time constants of its own, the A_oB curve is identical to the A_o curve except for being shifted vertically down by the feedback fraction B. Its phase response is also identical (or shifted bodily down by 180° depending on our frame of reference, as we will see later). Notice that the point where A_oB drops to exactly 0dB is also the GCF. We can therefore assess stability by looking at the GCF point(s) on the A_o curve, or on the A_oB curve; both can give us the same information. However, measuring the latter is both easier and more precise in practice. Once any amplifier has been set up so we can measure its loop gain, we just look for the point where it drops to 0dB and we know we are looking at the GCF, by definition.

The green line shows the closed-loop response of the amplifier, after feedback has been applied. Looking at the true curve (dotted) we find the gain is actually 'peaking' at each end of the bandwidth, exactly at the GCFs. This is happening because the loop phase shift is getting rather close to ±180° at these two critical frequencies. In this case peaking at the low end is not as bad as at the high end, owing to the positions of the poles and the phase shift each one contributes. If there were more poles then phase shift would grow more extreme and these peaks would grow higher and sharper. Eventually one of them might shoot up to infinity and we arrive at oscillation, which we don't want. There are some useful guidelines to adhere to: looking at the A_oB curve, if the GCF lies on a section that is falling at a first-order rate then the system will always be stable, because in a linear system the phase will never exceed 180° on a first-order section of slope. If the GCF lies on a second-order section of slope the system may or may not be stable, depending on how close higher-order poles happen to be. If the GCF lies on a third-order (or more) section of slope the system will certainly oscillate. In this example both GCFs lie on second-order sections of slope but the phase shift has not yet reached 180°, so the amplifier will not oscillate. The peaks in the closed-loop amplitude response will, however, produce ringing on square waves.

Measuring Loop Gain and Phase

The best way to assess the stability of a real circuit is to measure its true loop gain and phase. Valve feedback amplifiers are invariably built in a non-inverting configuration with feedback returned to the cathode of the input valve, or sometimes to the

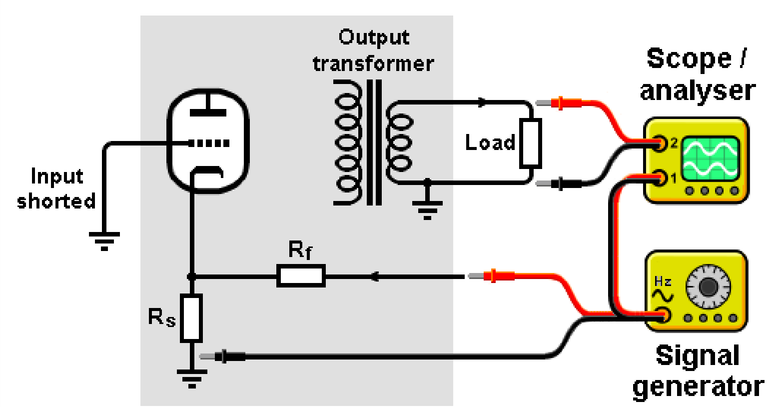

Fig. 3: Measuring loop gain by breaking the feedback path and injecting a signal into the feedback divider.

'spare' grid of a long-tailed pair phase inverter. In the non-inverting configuration the feedback signal is subtracted from the input signal, so the output (speaker) signal must be *in phase* with the input for feedback to be negative. But when we come to measure *loop gain* we will find it is always *inverting* for feedback to be negative. In other words, loop phase shift is ideally 180° in the mid band and becomes problematic as it approaches 0° or 360°.

To measure loop gain it is important that we break the loop at a point that does not significantly alter the way the amplifier inside the loop functions, or we will get misleading answers. The most practical way to do this is to make the break where the output transformer connects to the feedback divider, illustrated in fig. 3. The test signal is first attenuated by the divider that sets the feedback fraction B, then it passes through the rest of the amplifier with gain A_o, giving the total loop gain A_oB (the main amplifier input should be shorted to ground during measurement).

A modern audio analyser should be able to plot gain and phase automatically. If you don't have one then you can use a signal generator and a dual-channel oscilloscope to monitor simultaneously the test signal and the signal retuning from the amplifier; most modern digital 'scopes provide a function to read out the phase difference. You will mainly be looking for the point where loop gain drops to unity, and/or the phase shift reaches zero degrees, or close to it. An analog 'scope is unlikely to have a direct readout of phase, in which case it is possible to measure it the old-fashioned way using Lissajous figures. However, this is so laborious and prone to mistakes that, in this century, it is honestly worth investing in alternative equipment.

Stability Margin

Fig. 4 shows the loop gain and phase we might measure in another fictitious amplifier, concentrating on the high-frequency end. In this case the loop has a maximum gain of 30dB because that is the amount of feedback being applied (actually 30.3dB if we account for the 'minus one' mentioned earlier), at 100Hz at least. The circuit has, for whatever reason, three poles labelled P_1 to P_3. The total phase lag therefore tends towards 3×90° = 270° which is obviously more than the dangerous figure of 180°. However, loop gain is inverting, so we are actually looking for the point where the phase shift drops *from* 180° *to* 0°.

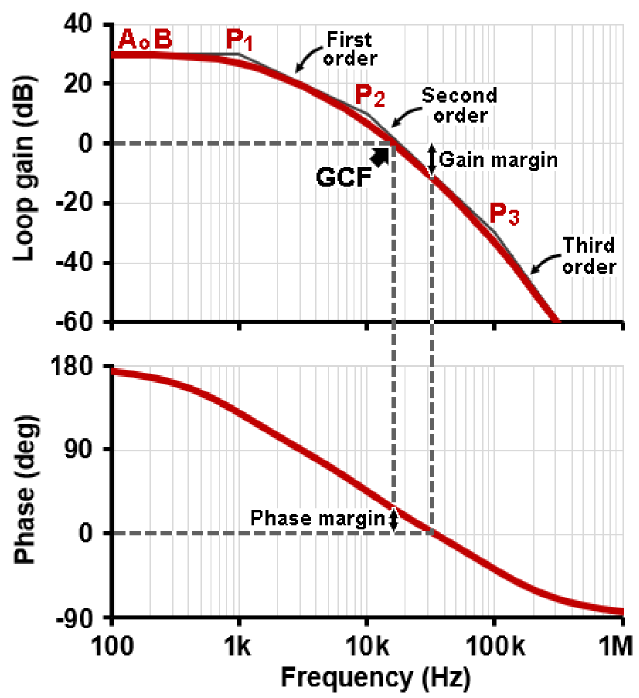

Fig. 4: Loop gain and phase of a fictitious amplifier, illustrating the gain and phase margin.

Will this amplifier oscillate if we close the loop? The GCF is the point where the loop gain falls to 0dB. In this contrived example we can see that it lies on a second-order section of slope, so immediately we know we're somewhere between 'cannot oscillate' and 'will oscillate'. How helpful. Nevertheless, following the dashed line we see that at the GCF the phase shift has only dropped to 30°. The difference between this and 0° is of course 30° and is called the *phase margin*. The phase margin is how much more phase shift would be needed at the GCF to reach the dangerous figure of 0°.

Now look at the point where the phase shift really does hit 0°; here the loop gain is –11dB. The difference between this and 0dB is of course −11dB and is called the *gain margin*. This is how much more loop gain would be needed to hit 0dB where the phase shift is 0°. Together the phase margin and gain margin can be called the stability margin of the amplifier. Exactly the same logic can be applied to the low end of the amplifier's bandwidth, not shown here, except the phase would be leading rather than lagging, i.e. we would be looking for the point where the phase hits 360° rather than 0°. In any case, phase is relative so it doesn't really matter how you choose to plot it as long as you keep track of what you are looking for. In this example the phase shift has not reached 0dB at the GCF, so this amplifier is stable at high frequencies.

Fig. 5 shows the response before and after the loop is closed. The open-loop gain A_o happens to be 40dB, so with 30dB of feedback the closed-loop gain A_c has been reduced to 10dB, but we have some peaking at the GCF. Peaking is an indicator of stability margin; the smaller the margin the worse the peaking. The math of high-order systems is complex but, as a rule of thumb, peaking normally disappears when the phase margin is better than about 60°.

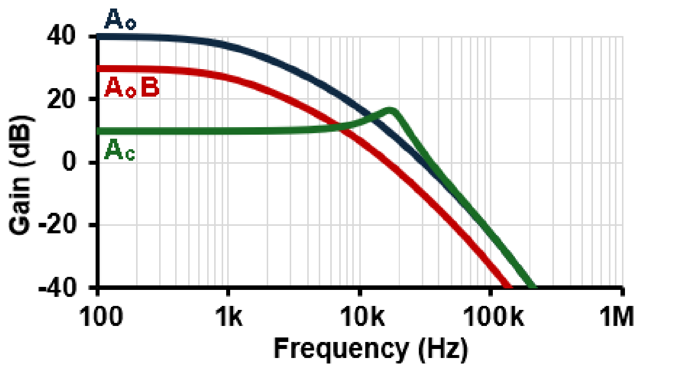

Fig. 5: Open and closed loop response of the fictitious amplifier. Note peaking at the GCF, which will cause ringing on square waves.

Stabilising an amplifier involves altering the loop gain and phase curves until a comfortable stability margin is achieved. How much margin is needed? There is no absolute answer but a phase margin of at least 30° and a gain margin better than −6dB is widely accepted as a good target to aim for, just before the loop is closed.[4,5] Once the loop is closed we may get peaking/ringing but this can finally be tuned-out with a speed-up capacitor as explained in Part 2. The stability margin of the completed design will then be even greater than the figures obtained when the loop was first closed.

Another point to note is that if we apply 100% feedback, i.e. the amplifier's output is connected directly to its input, then the A_o and A_oB curves become one and the same. Putting it another way, if the *open-loop* response drops below 0dB before the phase shift reaches ±180° then *any* amount of feedback could be applied and the system would not oscillate. This is called *unconditionally stable* or *unity-gain stable*. Tube amplifiers are rarely built for unity gain though.

The Problem of the Output Transformer

Wrapping feedback around an output transformer presents a particular problem, owing to its leakage inductance and distributed capacitance, mainly in the primary winding. These stray impedances are difficult to measure and they can vary somewhat with operating conditions, so transformer manufacturers are understandably reluctant to publish such figures. However, since we cannot alter these hidden components (unless you are in the business of winding your own transformers) we don't necessarily need to know their values, we only need to understand why they create difficulties when adding feedback.

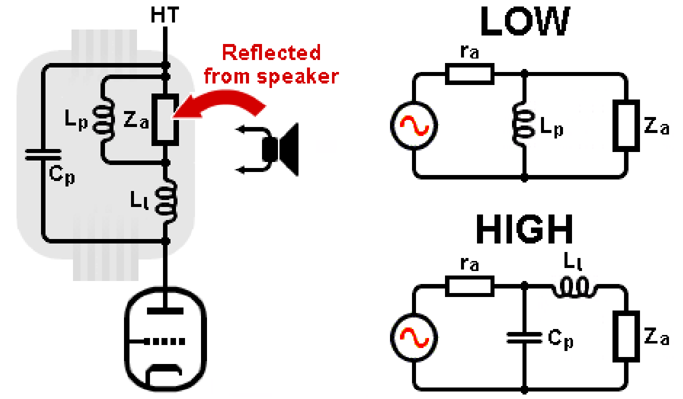

Fig. 6: Simple model of an output transformer, further simplified for low and high frequencies.
r_a = Internal plate resistance of the power tube;
Z_a = Primary reflected load impedance;
L_p = Primary inductance
C_p = Primary winding capacitance;
L_l = Primary leakage inductance.

Fig. 6 shows a simple equivalent circuit of a transformer, further broken down into equivalent circuits for low and high frequencies. Fig. 7 shows the gain and phase response we can expect from a typical tube output stage using such a transformer. At high frequencies the primary leakage inductance L_l, and lumped primary winding capacitance C_p, form a second-order RLC filter. A transformer therefore always introduces *at least two* high-frequency poles (for fuller treatment see Cherry and Hooper[6] or van der Veen[7]). Since the rest of the amplifier also contributes high-frequency poles you can probably anticipate the difficulty this creates if we want to apply feedback around the whole lot.

The larger the load resistance Z_a, or the smaller the internal resistance of the power tube/s r_a, the greater the damping of this RLC circuit. Pentodes have very high internal resistance which often results in under-damping, producing a small resonant peak in the open-loop amplitude response, which then falls off at a second-order rate. This is another way of saying P_1 and P_2 occur at the same frequency (mathematicians would say they are a complex conjugate pair). The phase will snap rapidly to −90° at resonance, and approach −180° equally rapidly.

[4] Hammond, P. H. (1958). *Feedback Theory and its Applications*, English Universities Press Ltd., p109.
[5] Cherry, E. M. & Hooper, D. E. (1968). *Amplifying Devices and Low-Pass Amplifier Design*, John Wiley and Sons Inc., p506.
[6] Ibid, p343.
[7] Van der Veen, M. (2010). *High-End Valve Amplifiers* 2, Wilco Amersfoort.

By contrast, power triodes have much lower internal resistance which helps to damp the circuit and may cause the two transformer poles to 'split', as illustrated in fig. 8 (the same pole-splitting effect will occur if an amplifier is tested with no load). The response therefore starts to fall off at a first-order rate at P_1, then transitions to a second-order rate at P_2. This results in a less rapid phase change, which tends to make triode amplifiers less difficult to stabilise if feedback is applied. The response of an ultra-linear output stage will be somewhere between the pentode and triode extremes.

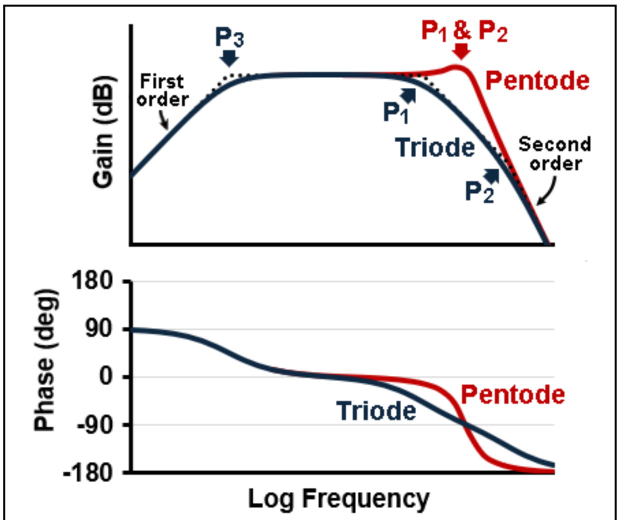

Fig. 7: Typical response of a valve output stage. The transformer introduces two high-frequency poles P_1 and P_2, but only one low-frequency pole P_3.

At low frequencies the situation is much simpler because tiny stray impedances are no longer important, only the primary inductance matters. This creates a simple first-order RL filter with a pole at:

$$f_{P3} = \frac{Z_a \parallel r_a}{2\pi L_p} \qquad (1.2)$$

Phase shift will be +45° at the pole, heading towards +90°. In many designs it is the transformer that creates the first low-frequency pole of the amplifier, often around 20Hz. Many manufacturers publish the primary inductance, at least for single-ended transformers, so this pole can be calculated easily. In a push-pull transformer the inductance will vary more with applied voltage, making it difficult to quote a single useful number. Just another reason to measure true loop gain rather than rely on calculation, then.

The Problem of Miller Capacitance

In textbooks you may find examples where the response of a multi-stage circuit is worked out on paper to assess its stability margin.[8] These examples are sometimes misleading because they often assume the response of each stage considered in isolation can be summed to find the total. In other words, they imply that we can work out the output resistance of each tube stage, and the Miller capacitance of the following stage, and use these figures to find every high-frequency pole in the circuit. But this does not work!

For example, fig. 8 shows a three-stage circuit where every stage has a gain of ×29, an output resistance of 10kΩ, and 2pF of anode-grid capacitance. The naïve approach suggests each stage therefore has (29+1)×2pF = 60pF of Miller capacitance. Each stage sees 10kΩ source resistance, so each stage must have a pole at 265kHz. Adding the three together should therefore give an overall third-order response above this frequency, right? Wrong. The actual response is shown in blue; the naïve approach is not even close. The problem is that each local feedback network acts as a reactive load upon its neighbours, changing their behaviour. By sketching in some Bode slopes (dashed) we can probably guess at a couple of poles, but there must be more at very high frequencies since the phase keeps on diving. Nor would it be correct to say any particular stage is responsible for any particular pole, they all interact. Trying to alter one, e.g. by adding more capacitance to C_{ga}, will unavoidably move the others too. However, it is nice to discover that the attenuation and phase change are not as rapid as anticipated, since this works in our favour if we want to wrap feedback around the circuit.

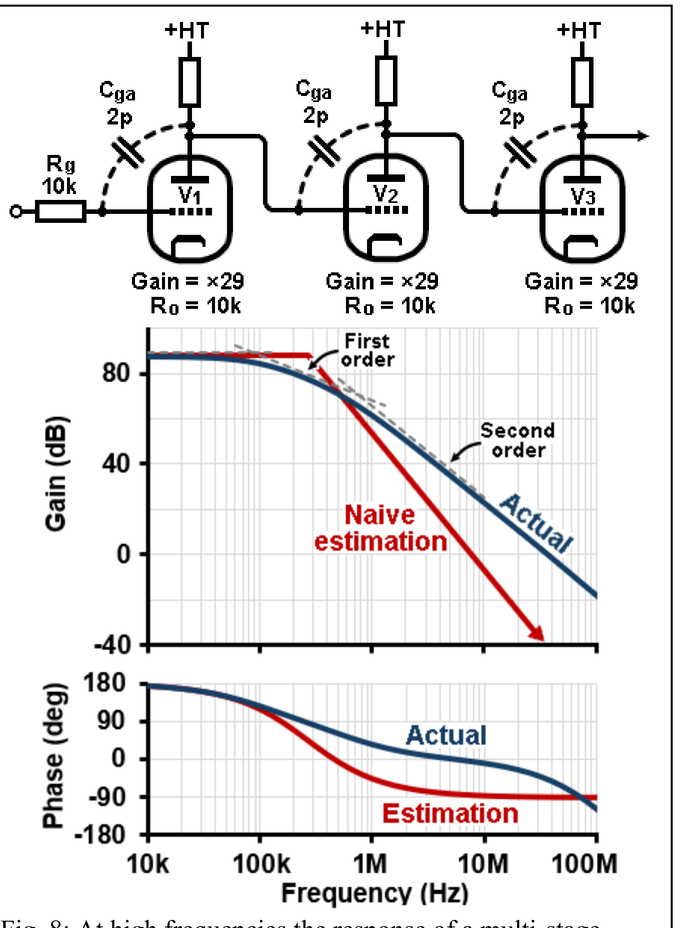

Fig. 8: At high frequencies the response of a multi-stage amplifier does not behave as the simple sum of the responses each stage would have in isolation.

[8] Crowhurst, N. H. & Cooper, G. F. (1957). *High-Fidelity Circuit Design*, Gernsback Library Inc., p35.

I present this example to illustrate how difficult (even futile) it is to try to predict the amplitude and phase response of an amplifier on paper, at high frequencies. Even Cherry and Hooper, masters of feedback theory, concede that "the complete analysis of a multistage amplifier in which the input impedance of one device forms part of the load for the preceding device is a formidable undertaking."[9] Today we can use computer simulation of course, but when a transformer is included –for which we probably don't have a good model– even simulation has its limits. There really is no substitute for measuring loop gain in real life, especially at the high end.

A corollary of this applies when global feedback is returned to the amplifier input stage. Referring to fig. 9, this stage will have an initial pole created between its output resistance and the load capacitance of the following stage, C_1. But if there is a resistance R_g in series with the grid then it will combine with the tube's own Miller capacitance to create a hidden pole and zero, which are pulled to lower frequencies as R_g is made larger. The response of the stage will vary as shown in the figure and, usually, the phase lead introduced by this hidden 'step network' will tend to improve the overall stability margin of the amplifier –more in Part 2.

If there happens to be a volume potentiometer at the input, this pole-zero pair will vary with pot rotation as the source resistance of the pot varies. Consequently, after stabilising the amplifier with the input shorted (e.g. pot turned down to zero) it is important to check that it remains stable at every other pot position too –look for square wave overshoot while rotating the pot. A grid-stopper helps to swamp such changes in source resistance, keeping it more constant regardless of pot position. In other words, an input grid stopper can be a valuable stabilising component even though it looks like it is outside the global feedback loop!

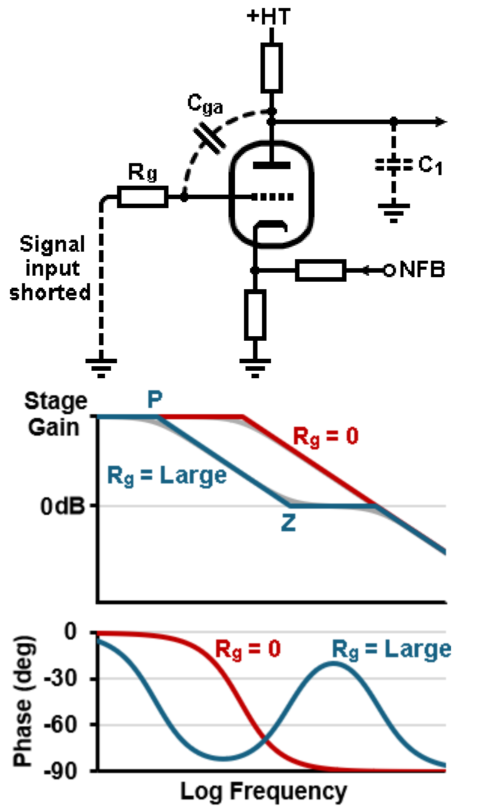

Fig. 9: C_{ga} creates a hidden pole-zero pair in the stage where feedback is applied, which depend on the resistance in series with the grid.

The Lesser-Problem of Low Frequencies

The low-frequency end of things is much easier to handle because we really can sum the responses of each individual stage to find the total, i.e. low-frequency poles don't interact, they can be moved independently. We are in full control of (nearly) all low-frequency poles and zeros since they are created by coupling capacitors, and to a lesser extent cathode bypass capacitors and power supply decoupling capacitors. The only one outside our command is the pole created by the output transformer, or as Edwin puts it "it is very much easier to increase the size of the coupling capacitances elsewhere in the amplifier than it is to increase the inductance of the transformer".[10]

If you do feel like sketching the low frequency response of an amplifier, the following approach might be used. Fig. 10 shows an example circuit and I assume the reader is familiar with basic amplifier design so should know these formulas already. A cathode bypass capacitor creates a zero at:

$$f_Z = \frac{1}{2\pi C_k R_k} \tag{1.3}$$

And a pole at:

$$f_P = \frac{1}{2\pi C_k r_k} \tag{1.4}$$

Where r_k is the internal cathode impedance of the tube. For the first stage this gives us P_1 = 4.8Hz and Z_1 = 1.6Hz. For the second stage we have P_2 = 23.9Hz and Z_2 = 8Hz.

A coupling capacitor creates a pole at:

$$f_P = \frac{1}{2\pi C_g (R_o + R_g)} \tag{1.5}$$

Which in this case gives P_3 = 6.9Hz. The output transformer creates a pole given by equation (1.2) so we have P_4 = 15.9Hz. The mid-band gain of the circuit is the gain of each stage multiplied together, including the turns ratio of the transformer which we will suppose is 0.04, giving a total gain of 20×50×0.04 = 40, which is 32dB, from input to loudspeaker.

[9]Ibid, p342.
[10] Edwin, G. (1962). Fundamentals of Feedback Design, *Wireless World*, January, pp27-28.

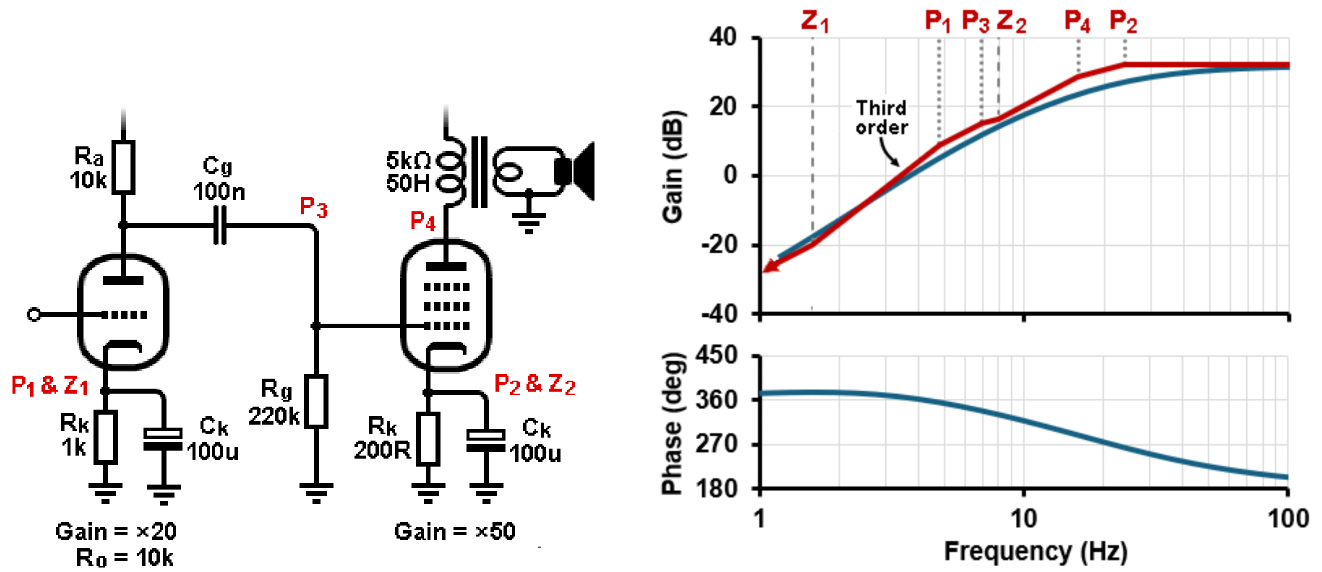

Fig. 10: Estimating low-frequency response is fairly straightforward, and most poles and zeros are fully under our control.

Marking of all the poles and zeros on a graph we can now sketch a Bode plot, starting at 100Hz say, and proceeding from right to left. Any time we encounter a pole the line should bend down by an additional 20dB/decade; any time we encounter a zero the line should bend less steep by 20dB/decade (if a pole and zero lie on top of one another they therefore cancel out). The result is shown in the figure in red, together with the true response in blue. This time the agreement is good. Phase response is not so easy to sketch, so only the true response is shown here. But neither is it really necessary to sketch it, since we can see from the Bode plot that we encounter a third-order slope at P_1, telling us that below this point would certainly be a dangerous place for the GFC to lie. The true phase plot confirms it; phase lag reaches 360° here. It is nearly always coupling capacitors that have the greatest impact on the LF stability margin, so it is desirable to minimise the number of them inside any feedback loop. In this example we could afford to make C_g larger, pushing P_3 lower, but we are free to play will all the capacitor values.

Something we have not accounted for is the power supply. At ultra-low frequencies where power supply decoupling capacitors no longer provide good bypassing, gain will try to rise, but the effect this has on the stability margin is nearly always negligible. Fig. 11 shows an example which is deliberately exaggerated. The power supply capacitor C_p and dropping resistor R_p create a pole at:

$$f_P = \frac{1}{2\pi C_p R_p} \qquad (1.6)$$

And a zero at:

$$f_Z = \frac{1}{2\pi C_p (R_p \| R_a)} \qquad (1.7)$$

Even with the unlikely component values used here, the 'gain boost' at low frequencies is less than 4dB and the phase shift less than 12°. Moreover, the phase is *lagging*, which therefore marginally improves the *leading* LF loop phase of the overall amplifier. Calculation is complicated when power supply filters are shared between multiple gain stages, but assuming the capacitors are a few tens of microfarads (why wouldn't they be?) they are still unlikely to have much influence on the stability margin. With triode stages the effect of power supply filters is smaller still.

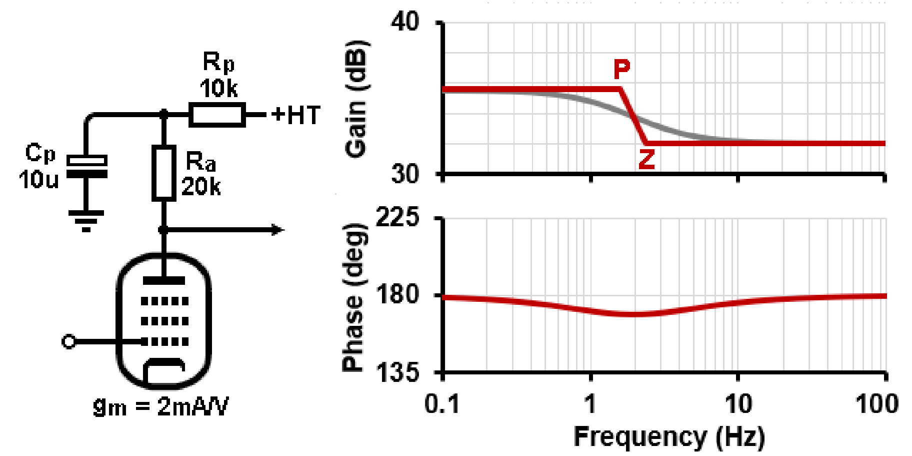

Fig. 11: Power supply decoupling capacitors have minimal effect on the stability margin, and it is in a favourable direction.

Closing Remarks

By working with gain and phase plots it is possible to build an intuitive understanding of how loop gain, bandwidth, and stability margin relate to one another. But so far I have only covered diagnosis of the patient, rather than treatment. We have not yet addressed the practical question every builder eventually faces: how to alter an amplifier that does not have enough stability margin for the amount of feedback desired. That will be the subject of Part 2 where I will cover the most common methods used to stabilise tube amplifiers, together with a real-world example that is taken from entirely unstable to firmly obedient. If you have a head for mathematics and want more quantitative information on feedback and stability, most of the references cited are available for free at: https://www.worldradiohistory.com.