

Choosing Cathode Bypass Capacitors

With little practical information in print, it was necessary to derive this simple solution to the problem of triode cathode-bypassing.

When designing a simple cathode-biased triode, voltage-gain stage of the type shown in **Fig. 1**, the designer is usually faced with choosing a cathode bypass capacitor in order to maximize gain. If the cathode resistor is left unbypassed, negative current feedback will limit the available gain of the stage to some minimum value. Usually this is a simple matter of using an arbitrarily large value of capacitance to ensure proper cathode decoupling well below audible frequencies.

Occasionally though, some of you may wish to use a small value of capacitance in order to obtain a treble boost action, because the cathode will be fully bypassed at high frequencies and effectively unbypassed at low frequencies. This might be called “partial bypassing,” and results in the stage operating as a first-order shelving filter. Possible applications include phono stages or pre-amplifiers for musical instruments and microphones, where a bass cut is often

desirable.

Calculating the maximum possible (fully bypassed) gain is easy, and is given by¹:

$$A_{\max} = \frac{\mu \cdot R_a}{R_a + r_a} \quad (1)$$

Likewise, the minimum possible (unbypassed) gain is just as easily found using²:

$$A_{\min} = \frac{\mu \cdot R_a}{R_a + r_a + R_k(\mu + 1)} \quad (2)$$

where, in both cases:

μ = the amplification factor of the valve.

R_a = the anode load resistor.

r_a = the internal anode resistance (plate resistance).

R_k = the cathode bias resistor.

(If the stage is heavily loaded by a following stage, you should substitute R_a for $R_a || R_l$, where R_l is the following loading resistance, which is usually the grid leak resistor.)

LOOKING FOR ANSWERS

But at what point is the transition made from one extreme to the other? And how should you choose the cathode bypass capacitor to give the desired result?

We consulted numerous sources, but none gave satisfactory answers to these questions. The older texts all assume that an audible bass-cut would be undesirable and give this subject almost no treatment at all, while some contemporary texts give incorrect treatment³. *The Radiotron Designer's Handbook* (4th ed.) does give a formula for gain at any frequency, on page 484:

$$\left| \frac{A'}{A_{\max}} \right| = \sqrt{\frac{1 + (\omega R_k C_k)^2}{\left[1 + \frac{R_k(\mu + 1)}{R_a + r_a} \right]^2 + (\omega R_k C_k)^2}}$$

where, all symbols as above, and:

A' = gain at the frequency of interest

$\omega = 2\pi f$

But this is hardly a convenient formula, and even if you solved it for C_k , you must pick an arbitrary level of gain “from the air,” and work from there. Of course, SPICE simulation could solve this problem without trouble, but that could be just as time consuming as solving the above formula. Therefore, we thought it would be more useful to have a universal and simple equation that would allow quick “back of an envelope” choice of the cathode bypass capacitor, or quick analysis of an existing circuit.

A SIMPLE SOLUTION

Normally, filters are defined according to their $\pm 3\text{dB}$ pole/zero frequency. However, this convention is less useful for this type of shelving filter because the difference between $A_{\max}$ and $A_{\min}$ may be much less than 6dB, in which case the pole would be lower in frequency than the zero! In fact, if the difference between $A_{\max}$ and $A_{\min}$ is less than 3dB, then there will be no conventional pole/zero at all! Instead, the best way to define such a filter, which is applicable to all circumstances, is by the point at which gain is halfway between minimum and maximum, which hereafter we call the “half boost point”⁴.

The problem of defining this point is most easily solved using a Nyquist plot of the denominator of equation 2 (**Fig. 2**), first by drawing a vector equal to the instantaneous current through R_a and r_a , and in series with this, current

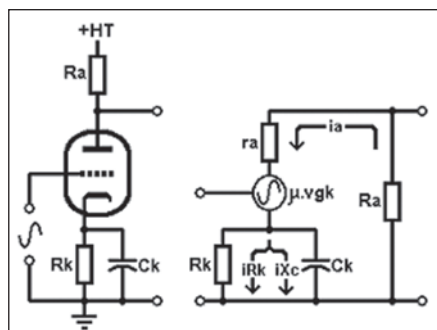


FIGURE 1: Left, a simple triode amplifier stage and its Thévenin equivalent circuit, right.

through $R_k(\mu + 1)$. At zero frequency, no current flows in C_k , and anode current (i_a) is equal to and in phase with current through R_k (iR_k). At maximum gain, C_k is effectively a short circuit, and current through it (iX_c) is also in phase with i_a . At intermediate gain levels, i_a is the vector sum of iR_k and iX_c , the vectors of which describe a semicircular arc, as shown.

At the half boost point, you can determine the total phase shift (θ) in i_a by drawing an arc of radius $R_a + r_a + \frac{1}{2}R_k(\mu + 1)$ as shown. Where this vector is tangential with the arc $\frac{1}{2}R_k(\mu + 1)$, then AB and BC indicate the relative magnitudes of iR_k and iX_c , respectively, while a/b yields the product of R_kC_k in radians per second, so that:

$$a/b = \omega R_k C_k$$

It is then a simple matter of finding a/b , by treating the problem as two geometrical problems as in **Fig. 3**, and solving them simultaneously:

To aid algebraic manipulation, put $(R_a + r_a) = A$ and $\frac{1}{2}R_k(\mu + 1) = B$.

From elementary geometry:

$$a^2 = b(2B - b) \quad (I)$$

and,

$$a^2 = (B - b)(2A + B + b) \quad (II)$$

Equating I and II,

$$2Bb = 2AB + B^2 - 2Ab$$

and collecting terms in b ,

$$b = \frac{B(2A + B)}{2(A + B)} \quad (III)$$

so,

$$b^2 = \frac{B^2(2A + B)^2}{4(A + B)^2} \quad (IV)$$

Substituting III into I and simplifying,

$$\begin{aligned} a^2 &= \frac{4A^2B^2 + 6AB^3 + 2AB^3 + 3B^4}{4(A + B)^2} \\ &= \frac{4B^2(A^2 + 2AB) + 3B^4}{4(A + B)^2} \quad (V) \end{aligned}$$

Putting V over IV and simplifying yields

$$a^2/b^2 = \frac{4A^2 + 8AB + 3B^2}{(2A + B)^2} = \frac{(2A + B) + 2B}{(2A + B)} = 1 + \frac{2B}{(2A + B)}$$

so,

$$a/b = \sqrt{1 + \frac{2B}{(2A + B)}} \quad (VI)$$

the limits of which are 1 and $\sqrt{3}$. Finally, putting $A = R_a + r_a$, and $B = \frac{1}{2}R_k(\mu + 1)$ and $a/b = \omega R_k C_k$,

$$\omega R_k C_k = \sqrt{1 + \frac{R_k(\mu + 1)}{2(R_a + r_a + \frac{1}{2}R_k(\mu + 1))}} \quad (VII)$$

and solving for $f_{(\text{half boost})}$;

$$f_{(\text{halfboost})} = \frac{\sqrt{1 + \frac{R_k(\mu + 1)}{2(R_a + r_a + \frac{1}{2}R_k(\mu + 1))}}}{2\pi R_k C_k} \quad (VIII)$$

Where all resistances are in Ohms, and all capacitances in Farads.

Figure 4 shows the idealized frequency response of the circuit given in **Fig. 1** using an ECC83 (12AX7), constructed by drawing a first-order (6dB/octave) slope through the half boost point, found with the above equation and converting to decibels. The heavy line shows the response produced by SPICE simulation, showing excellent correlation. In fact, the accuracy of equations 1, 2, and VIII is limited only by the accuracy of the value used for r_a , but this will usually be swamped by the tolerance of the capacitor and triode used. We hope you find this a useful and practical formula to accompany equations 1 and 2 given at the very beginning of this article, when designing triode amplifiers. **aX**

REFERENCES

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