Push-pull Input Circuits

Part 2.—Cathode-follower Phase-splitter

By W. T. COCKING, M.I.E.E.

ONE of the most widely used phase-splitters has the form of a cathode-follower, but with a coupling resistance in the anode as well as in the cathode circuit. It is by no means new and it preceded the cathode-follower as such, probably in time, and certainly in popularity. The earliest reference[1] to it which the writer has been able to trace is October 1935.

In basic form the circuit is the same as that of Fig. 7 (Part I), but with the input voltage applied between grid and earth instead of between grid and cathode. It is shown in Fig. 9 in its commonest form. It has the very desirable feature that the input and both output voltages all have one common earth terminal, so that it can readily be used after a circuit which has one of its output terminals earthy. The input voltage is E_{AB} and the outputs are E_{32} and E_{12}.

If the bias resistor R_b in Fig. 9 (a) is considered as short-circuited to alternating currents it is obvious that the input voltage E_{AB} must be equal to the sum of the grid-cathode voltage e_{gc} and the cathode output voltage E_{32}. With resistive circuit elements it is also obvious that all these voltages are in the same phase. Therefore, E_{32} must always be less than the input voltage by the amount of the grid-cathode voltage. The "amplification" $A_c = E_{32}/E_{AB}$ is thus always less than unity.

With the unearthed input circuit of Fig. 7 an amplification of 10-20 times is possible, but when one input terminal is earthed the amplification drops to less than unity. This is the price which must be paid for the convenience of the earthy input circuit. There is, however, also a considerable gain in linearity, through the negative feedback provided by R_c.

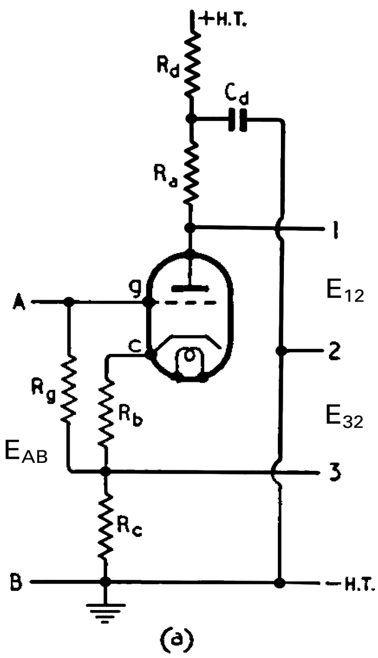
It is obvious that the circuit of Fig. 9 suffers from the same defect as that of Fig. 7 at low frequencies, which is that the output at the anode tends to increase relative to the output at the cathode because of the rising impedance of C_d, the decoupling capacitor. By analogy with Fig. 7 one would expect to obtain equality of the outputs at other frequencies when $R_a = R_c$ and this is actually the relation usually adopted in practice.

However, strictly speaking, this does not equalize the outputs, for the anode current of the valve is not the only current through R_a and R_c. There is a current through the grid leak R_g which flows through R_c and increases the cathode output. At high frequencies there are also currents through the grid-cathode and grid-anode capacitances C_{gc} and C_{ga}. These currents are not in phase with the anode current and they have the effect of making the anode and cathode output voltages unequal in amplitude and of giving them a phase difference which is not equal to the ideal 180°.

Over the middle range of frequencies, where the capacitances all have a negligible effect, the equivalent circuit has the form of Fig. 9 (b). The usual expression for the amplification is given by Eqn. (1) in Appendix II and it shows the cathode and anode outputs as being equal when $R_a = R_c$. It is accurate only when the frequency is such that the capacitances exercise a negligible effect and when R_g is infinitely large. This last condition is approached very closely in practice if R_g is returned to a potential divider across the H.T. supply instead of to the cathode circuit. This is shown in Fig. 10 and in using Eqn. (1) for this circuit we write $R_b = 0$, since there is no point in providing a bias resistor when the bias is otherwise obtained.

Although it is the better from this point of view, the circuit of Fig. 10 is not often used. It demands more parts than the other and the conditions for correct grid bias are rather more critical.

With cathode bias (Fig. 9) and when $R_a = R_c$, as is usual in practice, Eqn. (2) gives the unbalance in the two outputs; that is, the value of this equation is the fraction by which the cathode exceeds the anode output. It is at once obvious that a pentode is likely to be better than a triode, for the numerator will be



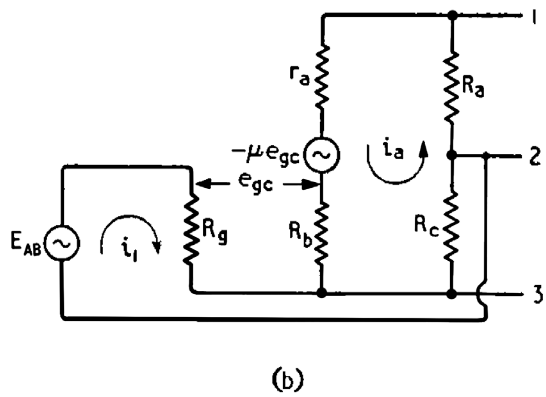


Fig. 9. The usual circuit of a cathode-follower type phase-splitter is shown at (a) and the equivalent circuit for low and medium frequencies at (b).

smaller owing to the higher A.C. resistance of the valve, and the denominator may well be somewhat larger.

However, the pentode is inconvenient in this circuit because of the screen supply which *must* be decoupled to cathode if the valve is not to become effectively a triode. This introduces further possibilities of error at extremes of frequency. A triode is, therefore, almost invariably used.

It is usual to make R_a and R_c equal and about equal to r_a, while R_b is rarely more than one-tenth of R_a. Under these conditions the unbalance is of the order of

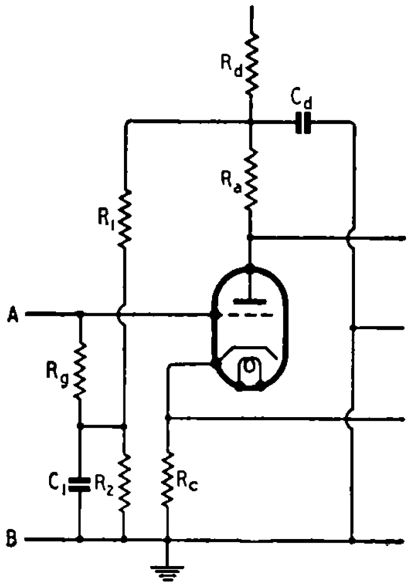


Fig. 10. A modifierd form of bias circuit is shown here.

$5/g_mR_g$. Now R_g can be as high as 2MΩ in most cases and g_m will rarely be less than 2 mA/V. Under these conditions the unbalance will be 0.125 per cent. In no practical case is the unbalance from this cause likely greatly to exceed this figure, so that it can nearly always be ignored. It is likely to reach practical importance only when R_g is below about 100kΩ.

At high frequencies the equivalent circuit has the form shown in Fig. 11, ignoring the anode-cathode capacitance of the valve. If the current i_1 through R_g can also be ignored, and it has been indicated above that this usually is permissible, the unbalance is given by Eqn. (4) of Appendix II. The expression is in two parts one with and one without the operator j attached to it. The part without j indicates a difference of amplitude between the anode and cathode outputs, the voltages so compared being correctly in opposite phase. Such an error can be corrected by a subsequent balance adjustment except in so far as its frequency-dependent term is concerned.

The part prefixed by j indicates the fractional amplitude of a component of one output in phase quadrature with the main output. It cannot readily be corrected in any subsequent circuit.

In a typical practical case we may well have g_m = 2mA/V, $R_a = R_c = r_a$ = 20k Ω, and R_b = 2kΩ. If C_C = 100 pF and it is unlikely to be higher, and C_{gc} = 5 pF, the phase unbalance at 10 kc/s is 0.02 per cent. The inphase unbalance is some 0.2 per cent.

These figures are so small that they are without much practical significance. In spite of the fact that the cathode-follower phase-splitter is inherently unbalanced, the magnitude of the unbalance is so small that for all ordinary purposes in A.F. amplifiers it is quite negligible. Practically speaking, it is necessary only to make R_a and R_c equal, and also the shunt capacitances C_a and C_c and to keep the grid leak of as high a value as possible. The capacitances C_a and C_c are usually composed mainly of the input capacitances of the two halves of the following push-pull amplifier, and so normally tend to be approximately equal. It is usually unnecessary to equalize them artificially.

As already mentioned the low-frequency unbalance is the same as with an earlier circuit and is given by Eqns. (7) and (8) Part I. It is almost entirely a phase unbalance and can be made negligible by the use of a large enough value for the decoupling capacitance C_d. Under normal conditions it should have a minimum value of 8 μF.

The input impedance of the stage is high. It is defined as the ratio of the input voltage E_{AB} to the total current flowing from the input voltage source into the input lead. Referring to Fig. II, $Z_{in} = E_{AB}/(i_1 + i_2 + i_3)$. At low and medium frequencies i_2 and i_3 are negligibly small, and it has already been shown that $i_1 = e_{gc}/R_g$; therefore, $Z_{in} = R_g E_{AB}/e_{gc} = R_g E_{AB}/(E_{AB} - E_{32}) = R_g/(1 - A_c)$. By inserting typical values into Eqn. (I) it is found that A_c usually lies

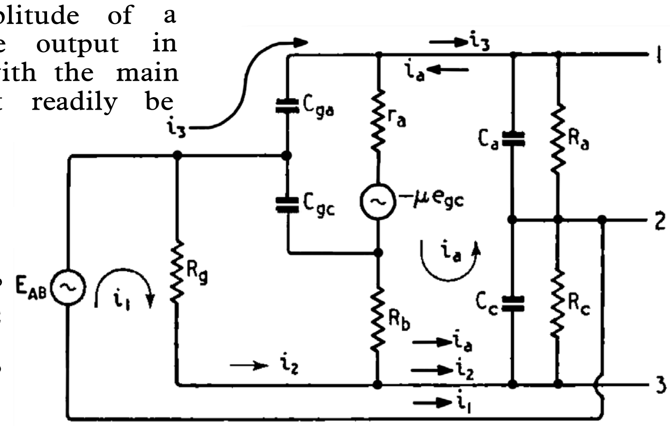


Fig. II. The circuit equivalent at high frequencies to that of Fig. 9 (a) is given here. The anode-cathode capacitance of the valve can usually be ignored without serious error.

between 0.85 and 0.95. If it is 0.9, Z_{in} = 10 R_g, and with the usual 2 MΩ for R_g, the input impedance becomes 20 MΩ

A similar effect occurs at high frequencies with C_{gc} as long as E_{AB} and E_{32} are nearly in phase. As A_c approaches unity. the cathode-earth voltage approaches the grid-earth voltage in value, and the difference between them, which is the grid-cathode voltage, is small, so that the current is small and the effective input capacitance from the element becomes very small and tends to zero.

The effect of the grid-anode capacitance is increased, however. In the limiting case when A_a = 1, if E_{AB} and E_{21} are in phase, the voltage acting to drive the current i_3 through C_{ga} is $E_{AB} + E_{21}$ = $2E_{AB}$, and then the effective input capacitance is $2C_{ga}$.

With normal values of components and over the audio frequency range it is sufficiently accurate for most ordinary purposes to take the input impedance as comprising a resistance 10 R_g shunted by a capacitance $2C_{ga}$. The inequalities of, and phase errors between, the two outputs are negligible, and the amplification A_c [given by Eqn. (I)] is of the order of 0.9.

No mention has so far been made of the output impedance of the stage. That at the cathode tends

Push-Pull Input Circuits—
towards that of a cathode follower whereas that at the anode conforms to the output impedance of a stage with negative current feedback. The cathode output impedance is much lower than r_a whereas the anode output impedance is much higher than r_a in normal applications of the circuit, however, these facts are without much practical significance.

In choosing circuit values it is generally satisfactory to make R_a and R_c about one to two times the working value of r_a and to make R_g as high as possible without making it so high that reverse grid current in the valve, or surface leakages on components, become troublesome. Because the input resistance is about 10 times R_g, the value of the input coupling capacitance can be about one-tenth of that appropriate to the value of R_g alone.

The decoupling resistor R_d should be as high as possible consistent with obtaining the requisite output from the stage, and C_d should be large, say 8-16 μF. The bias resistor R_b must be chosen to suit the valve and its operating conditions, but is usually 1-2 kΩ.

Turning now to the output available, this is of the same order as that given by the same valve working as a normal amplifier with a coupling resistor of $2R_a$, the same decoupling resistor R_d and bias resistor R_b, and the same H.T. supply voltage, but the linearity will be better because of the negative feedback provided by R_c. The output referred to above is the total output, $E_{32} + E_{21}$.

The exact conditions can readily be calculated by the usual graphical method. The D.C. load line for a resistance $R_a + R_b + R_c + R_c$ is drawn from the H.T. supply voltage on the anode-volts/anode-current valve curves and the desired operating point is selected ; the mean anode current I_a and anode-cathode voltage V_{ac} are then known. The A.C. load line for $R_a + R_b + R_c$ is then drawn through the point.

It is convenient to tabulate the anode currents and grid cathode voltages corresponding to the intersections of the line with the valve curves and to convert them to changes of current and voltage about the mean values by deducting these mean values from them. The cathode earth voltage is then the product of the current changes and $R_b + R_c$ while the cathode output voltage is the product with R_c. The sum of the grid-cathode voltage changes and the cathode-earth voltage changes give the grid-earth voltage changes, — the input. The dynamic characteristic is the plot of cathode output against input voltages and is the actual working characteristic taking feedback into account.

This procedure, while easy, takes some little time to carry out and it is helpful, therefore, to have a quick means of roughly estimating the output. With a triode the anode-cathode voltage cannot usually be swung below 25 + $V_c/6$ volts (where V_c is the voltage across C_d) without driving the valve into grid current.

The maximum anode-cathode voltage is usually about the same amount less than the mean voltage across C_d ; i.e., $V_c - 25 - V_c/6$ volts. The total swing is thus $\frac{2}{3}V_c - 50$, and the peak outputs at anode and cathode are each $(\frac{2}{3}V_c - 50)/4$. This is a very rough figure, but is useful for an initial estimation of the possibilities. If $V_c = 200$ V, for instance, an output at anode and cathode of the order of 20 V peak each can be expected. With 300V the output will be about 37.5V peak.

The mean anode-cathode voltage is about $V_c/2$ and the mean anode current about $V_c/2\,(2R_a + R_b)$. The mean cathode-earth voltage is about $\frac{V_c}{2}$.

$$\frac{R_a + R_b}{2R_a + R_b} \approx \frac{V_c}{4}.$$

This is important, for with many valves there is a maximum permissible heater cathode voltage and it is usually desirable to earth the heater. In the case of the EF37 valve, for instance, the rating is 100V. There is also for this valve a maximum figure of 20 kΩ quoted by the makers for the resistance between heater and cathode, so that $R_a + R_b$ must not exceed 20 kΩ.

With such a valve therefore, V_c is limited to about 400V and the outputs to about 55V peak, and the mean anode current will be of the order of 5 mA. This is within the maximum rating of 6 mA.

As an example of the determination of operating conditions and to illustrate the degree of accuracy of this rough method, the dynamic characteristic will now be deduced by the accurate method given earlier. We shall take an EF37 valve strapped as a triode. Since Rb + Re ≤ 20 kΩ, we shall take $R_a = R_c =$

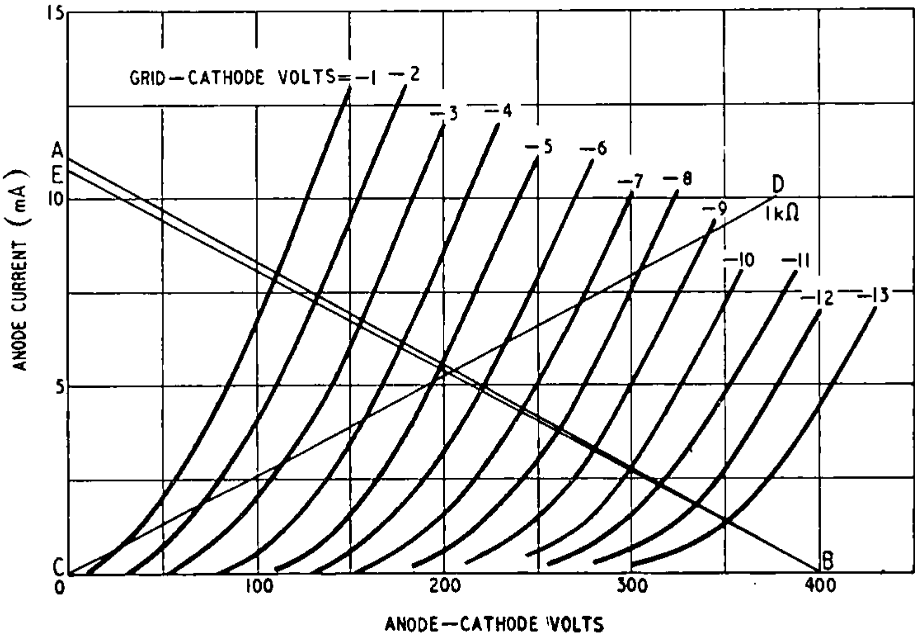


Fig. 12 Characteristics of the EF37 valve as a triode with load lines AB (36kΩ), and EB (37kΩ) and a bias line CD (1kΩ).

18 kΩ, since this is the nearest preferred value in the 5 per cent and 10 per cent tolerance ranges. The valve data places a limit of

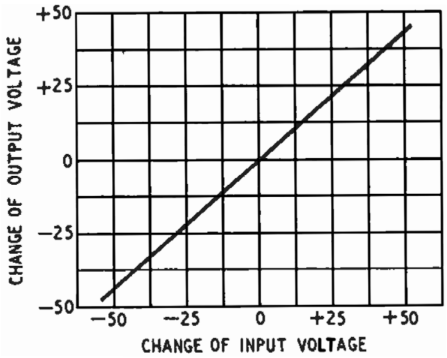


Fig. 13. Dynamic characteristics of the phase-splitter.

3 MΩ on R_g, and we can with confidence settle this at once at the standard value of 2.2 MΩ. We shall take V_c as 400 V.

The valve curves are shown in Fig. 12. As we do not know R_b at this stage we cannot draw the final load line, and we start off by drawing AB for 36 kΩ. It is obvious that the bias should be about − 6 V. With an input of 5 V peak, the grid-cathode voltage would swing from − 1 V to − 11 V and grid current should just be avoided. The anode-cathode voltage would swing from 105 V to 308 V with a mean value of 220 V. The outputs would be − 115 V and +88 V so that there is considerable distortion. The cathode-earth voltage would be (400 − 220)/2 = 90 V and the anode current 5 mA, so that the valve would operate within its rating.

A slightly lower bias would be better, but it cannot be much lower without the rating of the valve being exceeded. It is convenient to use a standard resistor for R_b, so let us try 1 kΩ. We draw the bias resistor line[2] by joining the intersections of the current ordinates with the grid- volts curves corresponding to the product of the current and the resistance. This is the line CD in Fig. 12. The new load line is now for 37 kΩ and is BE, and the no-signal operating point is the intersection of CD and BE at a current of 5.3 mA. The heater-cathode voltage is 5.3 × 19 = 100 V. It is just on the rating of the valve and it would be desirable to reduce it somewhat by reducing the H.T. voltage. The grid bias is 5.3 V.

The next step is to tabulate the grid voltages and the corresponding anode currents as in columns 1 and 2 of the table. Then prepare columns 3 and 4 for the changes of voltage and current, by deducting the no-signal values from columns 1 and 2, and produce column 5 by multiplying the figues of column 4 by the total cathode resistance $R_b + R_c = 19\,k\Omega$; this gives the change of cathode voltage. The sum of columns 3 and 5, in 6, gives the change of input voltage. Finally, column 7 is prepared by multiplying the figures of column 4 by the resistance R_c across which the output voltage is developed, in this case by 18 kΩ. The output at the anode is the same but with the signs reversed.

The relation between input and output voltages is shown by the curve of Fig. 13 and it will be seen that this is a straight line within the limits of accuracy imposed by rather small-scale graphical calculations. The maximum input is set by the onset of grid current, and is at a grid-cathode voltage of − 1, corresponding to a grid-earth potential of +51.8 V. the corresponding output being + 45 V. The amplification is 45/51.8 = 0.87 times.

The output of 45 V peak is somewhat below the figure of 55 V estimated earlier, but the agreement is reasonable since the method of estimation is a very rough one intended only for giving a preliminary indication of the output. In this case the output is limited by grid current and this indicates that a somewhat higher value of bias resistor would be better. There are, however, signs in Fig. 13 that the curve is starting to bend beyond − 50 V input and but little increase in bias resistance would be practicable.

A stage such as this will just feed a pair of push-pull PX4 valves directly but in view of the high value of H.T. supply needed there is nothing to spare for decoupling. Fortunately in this case decoupling is usually unnecessary.

The heater-cathode voltage with no signal is some 100V, the maker's maximum rating. On full output it rises to 145 V peak. It is not clear from the published figures whether this is permissible or not. Since a large heater-cathode voltage is normally used only with a superimposed signal it has probably been taken into account.

It will be seen from this that the conditions are rather tight when the phase-splitter is called on to feed a triode output stage directly and because of this the writer usually prefers to use an intermediate push-pull stage with an amplification of the order of 10 times. The phase splitter is then called on to provide an output of 4.5 volts or so only, and the valve can very easily be operated well within its limits.

In conclusion, it must be pointed out that condition of R_a and R_c

TABLE

1 Grid-cathode volts	2 Anode current (mA)	3 Change of grid-cathode volts	4 Change of Anode current (mA)	5 Change of cathode volts	6 Change of input voltage	7 Change of output voltage
−1	7.8	+4.3	+2.5	+47.5	+51.8	+45
−2	7.2	+3.3	+1.9	+36.1	+39.4	+34.2
−3	6.6	+2.3	+1.3	+24.7	+27.0	+23.4
−4	5.95	+1.3	+0.65	+12.5	+13.45	+11.65
−5	5.45	+0.3	+0.15	+2.85	+3.15	+2.7
−5.3	5.3	0	0	0	0	0
−6	4.8	−0.7	−0.5	−9.5	−10.2	−9.0
−7	4.3	−1.7	−1.0	−19.0	−20.7	−18.0
−8	3.75	−2.7	−1.55	−29.5	−32.2	−27.9
−9	3.2	−3.7	−2.1	−40.0	−43.7	−37.8
−10	2.65	−4.7	−2.65	−50.0	−55.1	−47.7
−11	2.3	−5.7	−3.0	−57.0	−62.7	−54.0
−12	1.8	−6.7	−3.2	−61.0	−67.7	−57.5

Push-Pull Input Circuits—
being equal, which has been assumed in all the foregoing, is one which must be closely observed in practice. It is usually desirable to employ resistors matched within about 2 per cent, but the usual tolerance of ± 20 per cent on actual value is quite satisfactory.

APPENDIX II

Referring to the circuit of Fig. 9(*b*), under the conditions that the reactance of C_d is negligibly small, that $R_g = \infty$ and that $R_a = R_c$,

$$A_c = \frac{E_{32}}{E_{AB}} = A_a = -\frac{E_{12}}{E_{AB}} = \frac{g_m R_c}{1 + \frac{R_b + 2R_c}{r_a} + g_m(R_b + R_c)} \quad (1)$$

where $g_m = \mu / r_a$ = mutual conductance.

When R_g is not infinite and $R_a = R_c$, the unbalance is

$$U = \frac{E_{32} - E_{12}}{E_{12}} = \frac{i_1}{i_a} \approx \frac{1 + \frac{R_b + 2R_c}{r_a} + g_m R_b}{g_m R_a} \quad .. \ (2)$$

provided that $i_1 \ll i_a$

At high frequencies the circuit has the form of Fig. 11. Assuming that C_{ac} and R_g have a negligible effect, the unbalance is:

$$U \approx \frac{i_1 + i_3}{i_a} \approx \frac{(r_a + R_b + 2Z_c)(1 + Z_{gc}/Z_{ga}) + 2\mu Z_c Z_{gc}/Z_{ga}}{\mu Z_{gc}} \quad .. \quad .. \ (3)$$

When $Z_{gc} = Z_{ga} = 1/j\omega C_{gc}$ and $Z_a = Z_c = R_a/(1 + j\omega C_c R_c)$ this becomes:

$$U \approx \frac{2\omega C_{gc}}{g_m}\left[1 + \frac{R_b}{r_a} + \left(g_m + \frac{2}{r_a}\right)\frac{R_c}{1 + \omega^2 C_c^2 R_c^2}\right] + j\,\frac{2\omega C_{gc}}{g_m} \bullet \frac{\omega C_c R_c}{1 + \omega^2 C_c^2 R_c^2}\left(g_m R_c + \frac{2R_c}{r_a}\right) \quad .. \quad .. \ (4)$$

References

1 "Resistance Coupling for Push-Pull amplification," by Walther Richter, Electronics, October 1935, Vol. 8, p. 382.

2 "Self-Bias and the Valve Load Diagram," by W. T. Cocking, Wireless Engineer, December 1934, Vol. 11, p. 655.

This article shows that provided the anode and cathode load impedances (*including R_k and R_a and the grid impedances they feed*) are equal the voltage outputs will be equal even at high audio frequencies despite the cathode output impedance being very low due to 100% negative feedback and the anode r_a being very high due to the large un-decoupled cathode resistor.

This work was overlooked by Morgan Jones, *Wireless World* Jan. 1996 (*although he did credit other prior art*) when he advised the cathode follower or 'concertina' phase splitter with equal R_a and R_k (R_c) requires a series resistor in the cathode output to provide balance. This is not the case and hardly, a fresh look at valve power, as the article was titled.

Achieving a perfect balance at the phase-splitter and or push-pull output stage may not be desirable becasue the even harmonics will get cancelled leaving relatively higher level odd harmonics.

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