

Notes on the Cathodyne Phase-Splitter

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Performance of the split-load circuit has often been questioned by designers and constructors. The author shows why most of the doubts are unfounded.

THE SPLIT-PHASE CIRCUIT of Fig. 1 is quite commonly used to drive a push-pull output stage, and has several interesting features. It is stated in various texts that the impedance seen looking into the plate circuit is different from that looking into the cathode circuit, yet for equal loads, the output voltages and frequency response are identical. It would appear that the cathode circuit, having the lower apparent source impedance, would have the better high-frequency response when both outputs are shunted by equal (and inevitable) circuit capacitances. In fact, some people think that this type of circuit is inherently unbalanced at the higher frequencies, whereas in actuality it is just as well balanced at the high frequencies as it is at the low frequencies, provided the two output load impedances are at all times equal.

The impedance looking into plate terminal A_p , (exclusive of Z_L), is

$$Z_{OP} = r_p + (1 + \mu)Z_K \quad (1)$$

where r_p is the plate resistance of the tube and μ is its amplification factor. The impedance looking into cathode terminal A_K , (exclusive of Z_K) is

$$Z_{OK} = \frac{r_p + Z_L}{1 + \mu} \quad (2)$$

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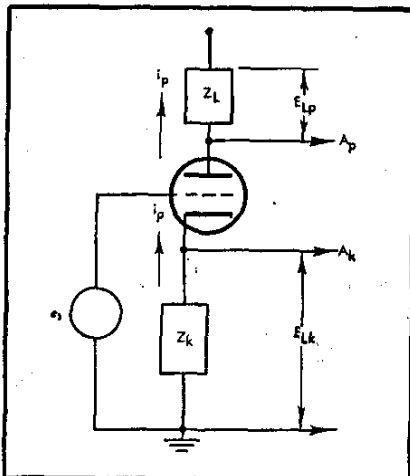


Fig. 1. Basic phase-splitter circuit.

It is evident that Z_{OK} , for reasonable values of Z_L , is much less than r_p , and Z_{OP} is greater than r_p by the amount of $(1 + \mu)Z_K$. In spite of this, the regulation of both portions of the circuit are the same, so long as $Z_K = Z_L$. For example, if Z_K and Z_L each are represented by a resistance paralleled by a capacitance, their high-frequency responses are the same, (as stated above), in spite of the fact that one is driven by a high-impedance source and the other by a low-impedance source.

The apparent paradox can be explained if we formulate the gain expression for either terminal A_p or A_K . Let the applied grid signal voltage be e_g , and let it develop output voltages E_{LK} across Z_K and E_{LP} across Z_L , where at first we do not assume $Z_K = Z_L$. The net voltage between the grid and cathode is e_p , where

$$e_p = e_g - E_{LK} \quad (3)$$

The equivalent voltage in the plate circuit is μe_p , and it causes a current i_p to flow through Z_K , r_p of the tube, and Z_L . Hence, by a simple application of Ohm's law, we have

$$i_p = \mu e_p / [r_p + Z_K + Z_L] \quad (4)$$

Current i_p , in flowing through Z_K , sets up the voltage

$$E_{LK} = i_p Z_K \quad (5)$$

and in flowing through Z_L , sets up the voltage

$$E_{LP} = i_p Z_L \quad (6)$$

If we substitute Eq. (5) in Eq. (3), and then Eq. (4) for the value of i_p , we obtain, after some algebraic manipulation:

$$e_p = e_g \left[\frac{r_p + Z_K + Z_L}{r_p + (1 + \mu)Z_K + Z_L} \right] \quad (7)$$

Eq. (7) can now be substituted in Eq. (4) to obtain:

$$i_p = \mu e_g / [r_p + (1 + \mu)Z_K + Z_L] \quad (8)$$

which, when substituted in Eqs. (5) and (6), yields finally

$$E_{LK} = \mu e_g Z_K / [r_p + (1 + \mu)Z_K + Z_L] \quad (9)$$

and

$$E_{LP} = \mu e_g Z_L / [r_p + (1 + \mu)Z_K + Z_L] \quad (10)$$

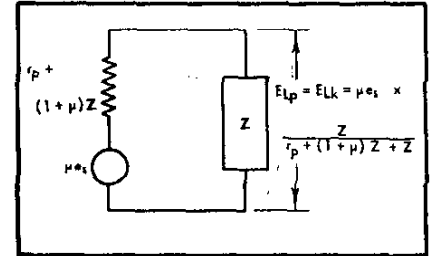


Fig. 2. Equivalent circuit for plate or cathode output.

If we now impose the phase-splitter condition that $Z_K = Z_L = Z$, we obtain:

$$E_{LK} = E_{LP} = \mu e_g Z / [r_p + (1 + \mu)Z + Z] \quad (11)$$

Equivalent Circuit

Equation (11) can be represented by the equivalent circuit shown in Fig. 2. The output voltage across Z is the fraction of the apparent generated voltage μe_g that Z is of the total circuit impedance. The latter consists of an apparent source impedance $[r_p + (1 + \mu)Z]$ that is greater than r_p by the amount $(1 + \mu)Z$. Note, then, that from this viewpoint, we have the same apparent generated voltage μe_g acting for either output, and the same apparent source impedance $[r_p + (1 + \mu)Z]$, rather than unequal source impedance as given by Eqs. (1) and (2).

We seem to be in greater mess than ever. However, let us go back to Eq. (9) and divide numerator and denominator of the right side by $(\mu + 1)$. We obtain:

$$E_{LK} = \frac{\mu}{\mu + 1} e_g \left[\frac{Z_K}{\left(\frac{r_p + Z_L}{\mu - 1} \right) + Z_K} \right] \quad (12)$$

The equivalent circuit is shown in Fig. 3. It holds whether Z_K is equal to Z_L , as in the case of the phase-splitter, or Z_K is not equal to Z_L . Note that the apparent source impedance appears here lower than r_p (for reasonable value of Z_L) owing to the factor $(\mu + 1)$ appearing in the denominator. But the apparent generated voltage is now also lower; it is also reduced by the factor $(\mu + 1)$.

In the special case where $Z_K = Z_L$, the circuit of Fig. 3 becomes exactly equivalent to that of Fig. 2, so that we can say in this case that the internal source im-

pedance for the plate output terminal Z_p of Fig. 1 is also as low provided we also accept a lower apparent generated voltage. Or, we can say that the apparent source impedance for this terminal is higher, namely $[r_p + (1 + \mu)Z]$, provided we also specify the higher apparent generated voltage. When $Z_L = Z_K$, the individual impedances lose their separate identities, as do also Eqs. (9) and (10), whereupon we can regard either output terminal as having a higher or lower source impedance, provided we also adjust the apparent generated voltages to correspondingly higher or lower values. It is only when we permit Z_K and Z_L to be unequal that we must use Eqs. (8) and (9) separately rather than use Eq. (11) for both output voltages.

We see, therefore, that the paradox is resolved if we take into account not only the change in source impedance but also the change in source-generated voltage. One can compensate for the other, but only in the case where $Z_L = Z_K$. Otherwise, a variation in either impedance causes the opposite effect upon the output voltage of the other terminal.

Practical Case

For example, suppose Z_L , the plate-load resistor, is increased but Z_K , the cathode resistor, is maintained constant. From Eq. (9) we see that E_{LK} decreases because only the denominator of the right expression increases. At the same time, Eq. (10) shows that E_{LP} increases because the numerator of its right expression increases faster than the denominator. The opposite effects are of course obtained if Z_L decreases.

On the other hand, if Z_K increases, E_{LP} decreases and E_{LK} increases by the same line of argument; and vice versa for a decrease in Z_K . We can summarize the effects by stating that a variation in either impedance causes a similar effect on its output voltage and an opposite effect on the output voltage of the other terminal; for equality of impedance, the two output voltages are equal.

The action described above concerning the difference in source impedance and the compensating difference in generated voltage can be more specifically explained in terms of what may be called a little network theorem. In (A) of Fig. 4 we see the basic circuit. The generated voltage is E_G ; the source impedance is

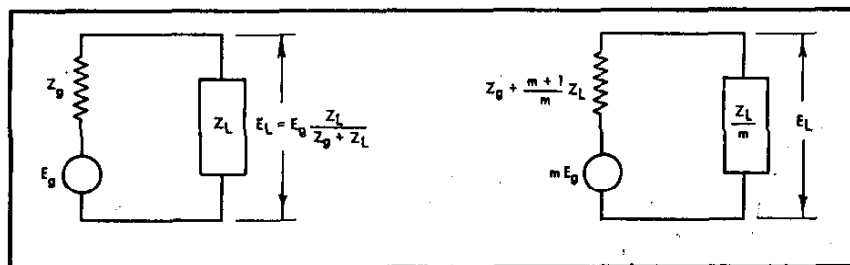


Fig. 4. Basic circuit and an equivalent circuit.

Z_G ; and the load impedance is Z_L . The output voltage is

$$E_L = E_G \frac{Z_L}{Z_G + Z_L} \quad (13)$$

just as in the case of Eqs. (9) and (10).

If we multiply E_G by some factor m , divide Z_L by the same factor m , increase

Z_G an amount $\left(\frac{m-1}{m}\right)Z_L$, and substitute

these altered values in Eq. (13), we obtain the same value E_L as before. Actually, the circuit impedance is unchanged;

$Z_G + \left(\frac{m-1}{m}\right)Z_L + \frac{Z_L}{m}$ equals $Z_G + Z_L$,

so that a physical interpretation of this alteration is that in (A) of Fig. 4 we have tapped down on Z_L at a point $1/m$ th its impedance and take off the voltage E_L at this tap. The rest of Z_L is then lumped with Z_G to represent an increased source or generator impedance. However, at the tap the voltage would be $1/m$ th its previous value; to compensate we make our generated voltage m times as great and thereby obtain the same value of E_L as before.

In the case of the phase-splitter circuit, $Z_K = Z_L = Z$, and in this case a term involving Z that is associated with the apparent source resistance can be shifted so as to become part of the load resistance or vice versa without changing the algebraic form of the gain expression. Under these conditions, we can shift terms so as to make them appear as part of the source resistance or as part of the load resistance, providing we compensate the apparent generated voltage correspondingly. The result is the same; the two output voltages are the same whether we regard the internal impedances of the two output terminals as equal or unequal. It is only when Z_K is not equal to Z_L that such shifting of terms is inadmissible, and it is then that the voltage regulations of the two terminals are different.

Before concluding this article, the writer cannot help but mention a result he obtained many years ago.¹ In using a vacuum-tube stage to feed a coaxial cable of, say, 75 ohms characteristic impedance, one can feed it from the plate circuit by first paralleling the r_p of the tube by a suitable low plate-load resistor so as to make the apparent source impedance 75 ohms to match the cable, or

one can feed it from the cathode circuit by first paralleling the lower apparent source resistance of $r_p/(1 + \mu)$ with a suitably higher load resistance to get the same apparent source resistance of 75 ohms.

The interesting thing is that when the characteristic impedance of the cable is matched from either point, the gain of the stage is the same. The advantages of feeding the cable from the cathode (cathode-follower stage) are that the benefits of inverse feedback are obtained (although amplitude distortion is not so serious a factor in a video amplifier), and more importantly from a practical viewpoint, the cable sheath can be placed at ground potential.

In the case of the phase-splitter circuit just analyzed, we can conclude that so long as the two load impedances are maintained equal at all frequencies of interest, no concern need be felt about the differences in apparent source impedance. The writer employed this type of phase-splitter back in the days when cathode-ray oscilloscopes had a bandwidth of 3 mc at most, and could maintain a flat response with such a circuit out to 8 mc.² The only difficulty encountered was that the screen signal current flowed through the cathode resistor but not through the plate load resistor, so that the latter had to be increased in value to obtain equal voltage outputs, but that is another story.

The only important disadvantage of the phase-splitter is that its maximum output voltage from either load resistor is only half that from a single-ended stage of the same total load impedance, so that some difficulty might be obtained in attempting to drive some power-tube grids. Ordinarily, however, the maximum output voltage is sufficient to drive the grids of even the larger power tubes used in high-fidelity audio amplifiers.

Some readers may wonder why the two terms involving Z in the denominator of Eq. (11) were not combined. They were left separate to show the same form as that of Eq. (13), namely, $Z_L/(Z_G + Z_L)$. Ordinarily, we would write Eq. (11) as $E_{LK} = E_{LP} = \mu e_p Z / [r_p + (2 + \mu)Z]$ (14) which is the form you will find in the handbooks and other texts. E

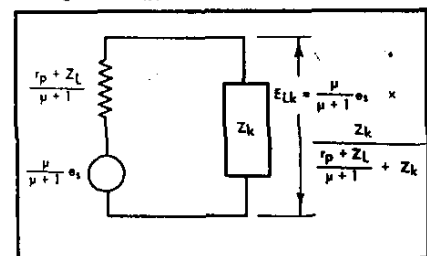


Fig. 3. Equivalent circuit for cathode output.

¹ A. Preisman, "Some notes on video amplifier design," *RCA Rev.*, April, 1938.

² See "A wide-range video amplifier for a cathode ray oscilloscope," A. Preisman, *RCA Rev.*, April, 1939.