# Handbook of Sound Reproduction

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## Chapter 13. Voltage Amplifiers and Phase Splitters

**A discussion of tube and circuit characteristics and their effect on frequency response and distortion in the low-level stages of audio amplifiers. Circuit configurations for phase splitters are shown and described.**

THE VOLTAGE GAIN of an amplifier stage is defined as the output signal voltage divided by the input signal voltage. If vacuum tubes had no internal resistance this gain would be equal to the amplification factor (referred to by the symbol $\mu$) of the tube. We have seen, however, that the open circuit output voltage of any practical generator is not applied directly to the load, but to a voltage divider made up of the internal resistance of the generator and the load resistance in series. The voltage gain of a stage of amplification must therefore always be less than the $\mu$ of the tube, in an amount determined by the ratio of the load resistance to the internal plate resistance. The greater the relative value of the load, the more closely the value of the gain can approach $\mu$.

### Equivalent Circuits

Amplifier analysis is facilitated by the use of a.c. equivalent circuits, from which characteristics such as gain, effective internal impedance, and frequency response can be readily derived. *Figure* 13—1 illustrates a typical stage of voltage amplification, and equivalent plate circuits. An a.c. generator, with an open-circuit output voltage equal to $\mu E_g$ (the input signal multiplied by the amplification factor) is substituted for the tube and the associated circuit elements that are required to make the tube work. All elements that are not in the signal circuit are eliminated; thus the B supply and the cathode resistor, which when properly bypassed present negligible resistance to the signal, do not appear in the diagram. The plate resistance of the tube, $R_p$, and the external load presented to the tube make up the total load seen by the generator. In C of *Fig.* 13—1 the effects of the coupling capacitor and of the input capacitance to the following stage are assumed negligible at signal frequencies; the single load $R_L$ represents the plate resistor and the following grid resistor in parallel, that is, $R_l R_g/(R_l + R_g)$.

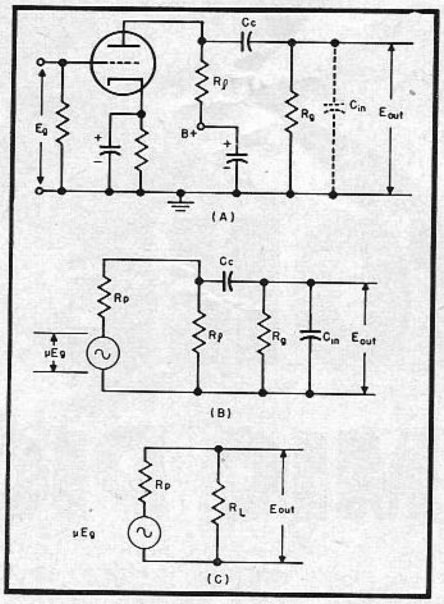

**Fig. 13—1. (A) Voltage amplifier stage. $C_{in}$ is the effective input capacitance to the following stage. (B) Equivalent a.c. circuit. (C) Simplified equivalent a.c. circuit, assuming $C_c$ shorted and $C_{in}$ negligible at signal frequencies. $R_L$ is the parallel combination of $R_l$ and $R_g$**

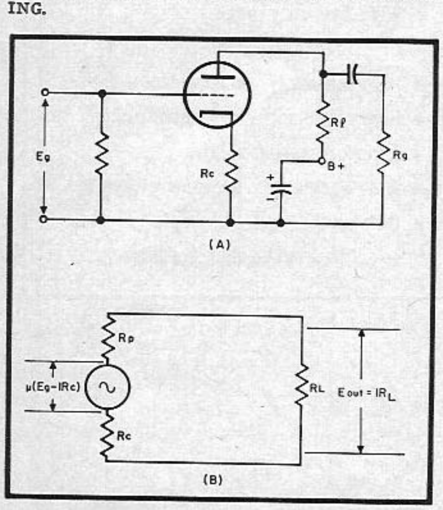

**Fig. 13—2. (A) Voltage amplifier with negative current feedback secured from unbypassed cathode resistor. (B) Equivalent a.c. circuit.**

It can be seen by inspection of C in *Fig.* 13—1 that the output voltage appearing across $R_L$ will be equal to:

$$E_{out} = \mu E_g \times \frac{R_L}{R_L + R_p}$$

The voltage gain is thus equal to:

$$\frac{E_{out}}{E_g} = \frac{\mu R_L}{R_L + R_p}$$

The same expression for gain may also be derived by applying Ohm's law, calculating the signal current from the applied voltage and total load, then calculating the IR drop across $R_L$. The current thus derived represents only the a.c. component of the plate current, and has nothing to do with the actual number of d.c. milliamperes flowing in the tube.

The Ohm's law derivation of the gain equation for a stage with an unbypassed cathode resistor $R_c$, (see *Fig.* 13—2), is shown below for the purpose of familiarizing the reader with the construction and use of equivalent circuits. In this case the original input voltage $E_g$ is reduced by $IR_c$, the signal voltage drop across the cathode resistor. It must be remembered that the d.c. component of the tube current is neglected.

$$\text{Voltage gain} = \frac{E_{out}}{E_{in}} = \frac{IR_L}{E_g}.$$

$$I = \frac{E_{generator}}{R_{total}} = \frac{\mu(E_g - IR_c)}{R_L + R_p + R_c}$$

$$\mu(E_g - IR_c) = (R_L + R_p + R_c)I$$

$$\frac{\mu E_g}{I} = R_L + R_p + R_c + \mu R_c$$

$$I = \frac{\mu E_g}{R_L + R_p + R_c(\mu + 1)}$$

$$\text{Voltage Gain (with negative current feedback)} = \frac{\mu R_L}{R_L + R_p + R_c(\mu + 1)}$$

To calculate the source impedance seen by the load, a generator of arbitrary voltage $E$ is substituted for the load impedance, and the generator representing the tube is short-circuited, indicating zero input signal. Any feedback voltages are represented by series generators with a polarity appropriately related to that of the load generator; thus a generator representing a negative current feedback voltage will have a polarity opposite to that of the load generator, (see *Fig.* 13—3) while negative voltage feedback will be represented by a generator with the same polarity as the load generator. The

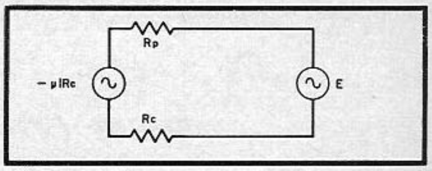

**Fig. 13—3. Equivalent a.c. circuit to stage with unbypassed cathode resistor, constructed for purpose of calculating effective internal impedance. This is the impedance seen by a generator, of arbitrary voltage E, which has been substituted for the load.**

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derivation given here, referred to in *Fig.* 13—3, is for negative current feedback; the reader may wish to check the equations for source impedance with negative voltage feedback, positive current feedback, etc.

Source Impedance $(R_{int}) = \frac{E}{I}$

$$I = \frac{E_{total}}{R_{total}} = \frac{E - \mu I R_c}{R_p + R_c}$$

$$I\,(R_p + R_c) = E - \mu I R_c$$

$$R_p + R_c = \frac{E}{I} - \mu R_c = R_{int} - \mu R_c$$

$$R_{int} = R_p + R_c + \mu R_c = R_p + R_c\,(\mu + 1)$$

For a multistage feedback loop $\mu$ becomes the gain with the final load open, $R_c$ the resistor in series with the load.

### Frequency Response

At low frequencies the coupling capacitor $C_c$ of (B) in *Fig.* 13—1, and the grid resistor $R_g$, form a voltage divider that reduces the input signal to the following stage. When the frequency becomes low enough so that the reactance of $C_c$ is numerically equal to the resistance of $R_g$ the signal voltage across $R_g$ will be reduced, by approximately 3 db, from the total signal across the voltage divider. (The value of the total load presented to the previous stage will also be increased somewhat.)

Although it might be expected that when the reactance of $C_c$ is equal to $R_g$ the signal voltage will be halved, and the reduction amount to 6 db, it must be remembered that the reactance of $C_c$ is a vector quantity.

The low-frequency amplification of a stage may be expressed as a fraction of the mid-frequency gain:

$$A_L = \frac{A_M}{\sqrt{1 + \frac{1}{(\omega R_1 C_c)^2}}}$$

where $A_L$ = Low-frequency gain (does not include phase shift)
$A_M$ = Mid-frequency gain
$\omega = 2\pi \times$ frequency concerned
$R_1 = R_g + \frac{R_l\,R_p}{R_l + R_p}$
$R_p$ = tube plate resistance; $R_l$ = plate resistor; $R_g$ = grid resistor
$C_c$ = Capacitance of coupling capacitor, in farads

*Figure* 13—4 plots the low-frequency transmission of amplifier networks with given coupling capacitors and grid resistors. It should be evident that the lower the value of the grid resistor the greater must be the value of the coupling capacitor for the same low-frequency transmission.

The low-frequency response of an amplifier may also be reduced by insufficiently large cathode and screen bypass capacitors. If the reactance of the cathode bypass at low frequencies becomes appreciable, relative to the bias resistor, discriminative negative current feedback and bass attenuation are introduced. If the capacitor bypassing the screen to ground is too small to be effective at low frequencies, signal voltage variations appear on the screen grid, of such polarity as to reduce the cathode-plate signal current and cause the signal frequencies concerned to be attenuated.

Simplified expressions[1] for calculating approximate required minimum capacitances in audio applications (not considering phase shift) appear below:

$$C_c = \frac{1.6 \times 10^6}{f_L R_g}\ \mu\text{f}$$

$$C_k = \frac{1.6 \times 10^6}{f_L R_c}\ \mu\text{f}$$

$$C_s = \frac{1.6 \times 10^6}{f_L R_s}\ \mu\text{f}$$

where $C_c$ = Coupling capacitor
$C_k$ = Cathode bypass capacitor
$C_s$ = Screen bypass capacitor
$f_L$ = Low-frequency limit in cps
$R_g$ = Grid resistor, ohms
$R_c$ = Cathode resistor, ohms
$R_s$ = Screen dropping resistor, ohms

The treble response of an amplifier stage is limited by the shunt effect of the tube capacitances, illustrated in *Fig.* 13—5. The effective input capacitance of a stage is not merely the sum of the inter-electrode tube capacitances, but is affected by the amplification of the tube:

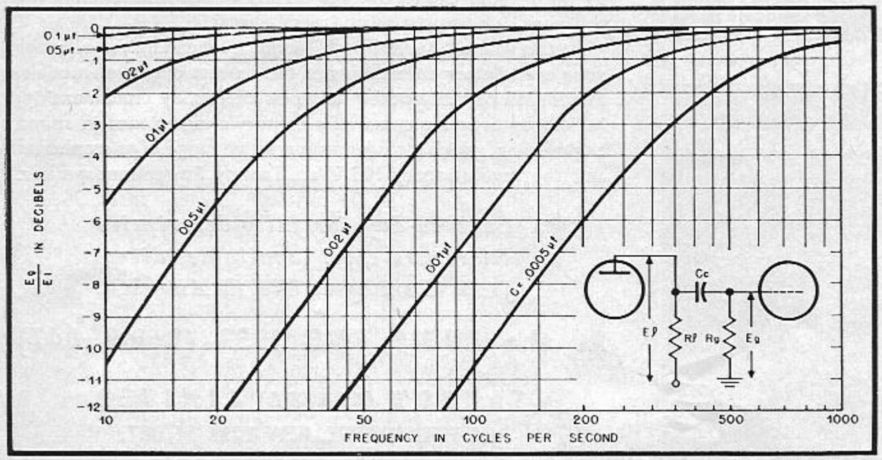

Fig. 13—4. Approximate low-frequency transmission of R-C coupling networks into grid resistor of 1.0 megohm (assuming $R_p$ and $R_1$ of first stage small). For other values of grid resistor multiply C by 1.0 meg/$R'_g$. To include effect of $R_p$ and $R_1$, use $R_g + R_1R_p/(R_1 + R_p)$ for $R_g$. *(After F. Langford Smith)*

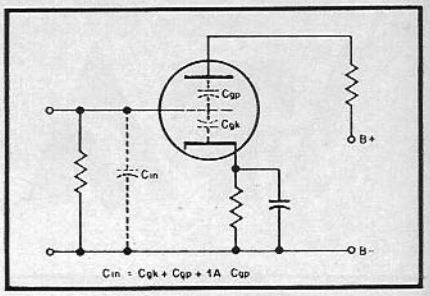

Fig. 13—5. Inter-electrode tube capacitances, whose combined effect is represented by $C_{in}$, the input capacitance of the stage.

$$C_{in} = C_{gk} + C_{gp} + |A|\,C_{gp}$$

where $C_{in}$ = Effective input capacitance of tube
$C_{gk}$ = Grid-cathode inter-electrode capacitance
$C_{gp}$ = Grid-plate inter-electrode capacitance
$|A|$ = Absolute value of voltage gain

The effective tube capacitance forms a shunt across the lower arm of the plate resistance-load voltage divider. (The lower arm is the parallel combination of $R_l$ and $R_g$. When the resistance of the combination of $R_l$ and $R_g$ is very large the shunt capacitance has an appreciable effect on the value of the total load impedance at high frequencies, but when the resistance of the combination is low the further shunt reactance changes the total load value very little. The effect of the change in load value is reduced by a low plate resistance that causes the stage to approach constant-voltage operation. Thus, high-frequency response is aided by low values of plate and grid resistors, by low tube amplification and plate resistance, and by low inter-electrode capacitance, especially that between control grid and plate. With normal R-C coupled circuits the upper frequency limit is well above the audio range.

The high-frequency amplification of a stage is equal to:

$$A_H = \frac{A}{\sqrt{1 + (\omega R_2 C_{in})^2}}$$

where $A_H$ = High-frequency gain (does not include phase shift)
$A_M$ = Mid-frequency gain
$\omega = 2\pi$ times the frequency concerned
$R_2$ = The parallel combination of $R_p$, $R_l$ and $R_g$
$= \frac{R_pR_lR_g}{R_pR_l + R_pR_g + R_lR_g}$
$R_p$ = tube plate resistance; $R_l$ = plate resistor; $R_g$ = grid resistor
$C_{in}$ = Effective input capacitance, in farads.

The insertion of a series element between stages, such as the upper arm of a potentiometer used as a volume or tone control, adds to the effective value of the source impedance seen by the input terminals of the following stage, and therefore reduces the high-frequency range for the same circuit values (counting the effect of the reduced resistance of the

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[1] Charles P. Boegli, "A note on volume controls," AUDIO ENGINEERING, p. 40, April, 1953.

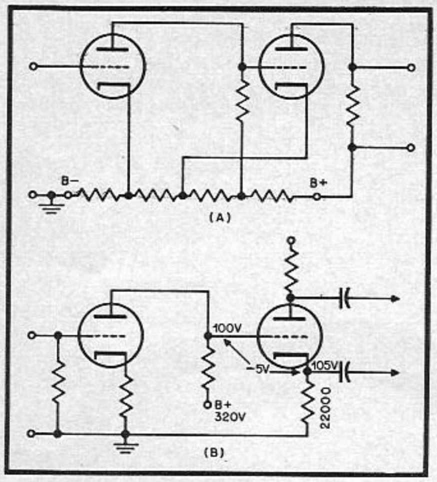


**Fig. 13—7. (A) Direct-coupled amplifier using tapped B supply. (B) Direct-coupled circuit used in Williamson amplifier. The 100 volts (relative to ground) on the grid is offset by the 105 volts between cathode and ground.**

lower arm of the volume control).[2] The restriction, however, is generally from one ultrasonic value to a lower one, and therefore not significant; controls can never be included within a feedback loop where ultrasonic response may be needed. One circuit position where the effect of the control on high frequencies may be important, however, is the input to a long cable.

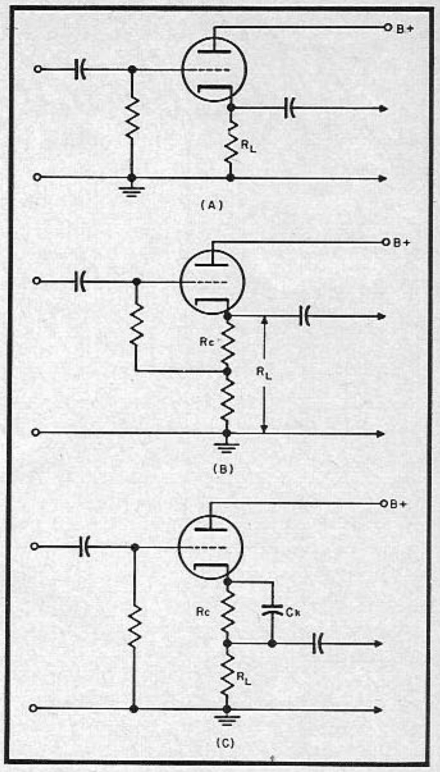


**Fig. 13—8. (A) Cathode follower voltage amplifier. All of the voltage across $R_L$ is fed back to the input, constituting 100 per cent negative voltage feedback. (B) Method of providing correct bias for cathode follower stage when $R_L$ is too large. The bias voltage is formed across $R_C$. (C) Method of providing correct bias for cathode follower stage when $R_L$ is too small. The bias voltage is formed across the series combination of $R_L$ and $R_C$. If $C_K$ is eliminated, $R_C$ will provide negative current feedback.**

From the above it may be seen that the frequency range of a resistance-capacitance coupled stage can within limits be chosen at will, and is not ordinarily dependent upon any subtle virtues or defects of the circuit components.

### Distortion in Voltage Amplifiers

Voltage amplifiers are, under normal circumstances, operated in class $A_1$. As in the case of power amplifiers the percentage of distortion is a function of the signal amplitude; the voltage amplifier stages that are most subject to distorting influences are therefore the "high-level" stages just prior to the power amplifier. The same methods that are used to decrease distortion in the power amplifier may be used here—the application of inverse feedback (both voltage and current feedback are equally beneficial) and the use of push-pull circuitry.

Negative current feedback may be inserted simply by omitting the bypass capacitor from the cathode bias resistor. Harmonic distortion and other undesirable effects are reduced, as with voltage feedback, by the gain reduction factor, while the increase of source impedance is of no consequence. If the loss of gain can be afforded, the only disadvantage of an unbypassed cathode resistor lies in the increased danger of hum picked up from heater-cathode leakage.

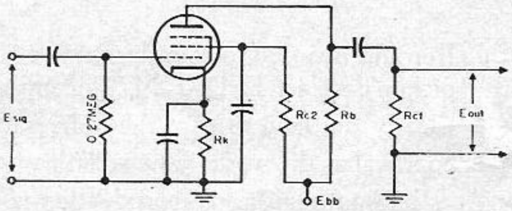
Voltage amplifiers in push-pull exercise the same discrimination against even harmonic distortion as power amplifiers. The actual cancellation does not take place, however, until the signal passes through a common load such as the output transformer.

### Bias Voltage

Bias in a voltage amplifier is obtained from a cathode resistor in the overwhelming majority of cases. Occasionally the grid resistor method is used in low-level stages such as phonograph preamplifiers. In the latter case a grid resistor of high value, of the order of 10 megohms, is used between grid and ground, while the cathode is grounded.

### The Use of Tube-Manual Data in Designing R-C Coupled Voltage Amplifiers

Although circuit equations may be used to design voltage-amplifier stages, the values of circuit elements for most of the common applications can be determined by referring to prepared charts. *Figure* 13—6 illustrates a section of such a chart, and includes a diagram of the circuit elements to which the symbols refer. The gain and distortion characteristics accompanying various combinations of circuit values, supply voltages, signal amplitude, and in some cases type of bias may be read directly. The importance of not over-driving stages located prior to the volume control (where the increased signal amplitude will not be



**RESISTANCE COUPLED AMPLIFIER DATA**

Self Bias Operation

6SJ7

| | Ebb - 100 VOLTS | | | | | | | Ebb - 250 VOLTS | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| Rb | 0 1 | | 0 27 | | | 0 47 | | 0 1 | | 0 27 | | | 0 47 | |
| Rc2 | 0 39 | | 1 2 | | | 1 8 | | 0 39 | | 1 2 | | | 2 2 | |
| Rcf | 0 27 | 0 47 | 0 27 | 0 47 | 1 0 | 0 47 | 1 0 | 0 27 | 0 47 | 0 27 | 0 47 | 1 0 | 0 47 | 1 0 |
| Rk | 1200 | 1200 | 2700 | 2700 | 2700 | 4700 | 4700 | 560 | 560 | 1200 | 1200 | 1200 | 1800 | 1800 |
| Ib | 0 645 | 0 645 | 0 259 | 0 259 | 0 259 | 0 165 | 0 165 | 1 77 | 1 77 | 0 675 | 0 675 | 0 675 | 0 402 | 0 402 |
| Ic2 | 0 18 | 0 18 | 0 068 | 0 068 | 0 068 | 0 045 | 0 045 | 0 50 | 0 50 | 0 183 | 0 183 | 0 183 | 0 102 | 0 102 |
| Ec1 | 0 99 | -0 99 | 0 882 | -0 882 | -0 882 | -0 99 | -0 99 | -1 27 | -1 27 | -1 03 | -1 03 | -1 03 | -0 908 | -0 908 |
| Ec2 | 29 8 | 29 8 | 18 5 | 18 5 | 18 5 | 19 0 | 19 0 | 55 | 55 | 30 5 | 30 5 | 30 5 | 25 5 | 25 5 |
| Eb | 35 5 | 35 5 | 30 2 | 30 2 | 30 2 | 22 5 | 22 5 | 73 | 73 | 67 8 | 67 8 | 67 8 | 61 2 | 61 2 |
| Esig | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 | 0 1 |
| Eout | 6 85 | 7 8 | 8 2 | 10 2 | 12 5 | 10 2 | 13 1 | 10 2 | 11 5 | 13 6 | 17 9 | 21 6 | 19 5 | 25 6 |
| Gain | 68 5 | 78 0 | 82 | 102 | 125 | 102 | 131 | 102 | 115 | 136 | 179 | 216 | 195 | 256 |
| % Distortion | 0 6 | 0 7 | 3 4 | 2 6 | 2 3 | 2 8 | 3 2 | 0 7 | 0 8 | 2 2 | 1 8 | 1 5 | 3 1 | 2 4 |
| Esig(1) | 0 2 | 0 2 | 0 14 | 0 14 | 0 14 | 0 13 | 0 13 | 0 5 | 0 5 | 0 25 | 0 25 | 0 25 | 0 15 | 0 15 |
| Eout | 13 15 | 14 9 | 11 1 | 13 9 | 17 2 | 12 8 | 16 6 | 47 | 54 | 33 | 41 8 | 50 | 28 | 37 |
| Gain | 65 8 | 74 5 | 79 4 | 99 5 | 123 | 98 5 | 128 | 94 | 108 | 132 | 167 5 | 200 | 187 | 247 |
| % Distortion | 3 0 | 2 9 | 5 1 | 4 3 | 3 7 | 4 6 | 5 0 | 4 2 | 5 0 | 5 2 | 4 4 | 4 7 | 4 5 | 3 7 |

Note (1) At grid current point, less than 1/8 microampere grid current

**Fig. 13—6. Sample of R-C coupled amplifier data used for selecting circuit values. (Part of a chart from the Sylvania Tube Manual)**

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[2] Technical Manual—Sylvania Radio Tubes, Sylvania Electric Products, Inc., Emporium, Penna., 1946, p. 32.

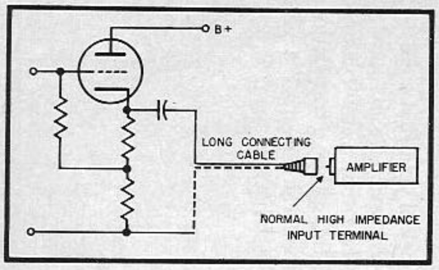

**Fig. 13—9. Use of a cathode follower for overcoming effects of long connecting cable on high-frequency response and hum level. Common value for cathode load is 10,000 ohms.**

evident in the final output) and of choosing tubes and circuits adequate to handle the signal voltages that will be fed to them, is apparent from the distortion data. Since negative feedback reduces only gain and not the output capabilities of the stage, the allowable input signal is increased by at least the feedback gain reduction factor.

### Direct-Coupled Stages

The function of the coupling capacitor is to keep d.c. from the following grid, while allowing the signal to pass. This capacitor may be eliminated, and the plate connected directly to the following grid, if the positive d.c. plate voltage applied to the grid is over-balanced, in such measure that the resultant bias voltage between grid and cathode is negative and of the proper value. (A) of *Fig.* 13—7 illustrates the principle of a direct-coupled amplifier. Although such amplifiers present stability problems much greater than occur in conventional R-C designs and need special power supplies, they are required in certain applications. The measurement of minute body potentials in medical work, for example, calls for an amplifier that responds to input voltage stimuli which may change very slowly or not at all.

Direct-coupled stages may also be incorporated in audio amplifiers for the purpose of reducing phase shift within the feedback loop. (B) of *Fig.* 13—7 illustrates the direct-coupled input stages of the Williamson circuit. The positive voltage on the second grid is more than offset by the IR drop across the large cathode load resistor, so that the final bias voltage is – 5 volts, as shown.

### The Cathode-Follower Voltage Amplifier

The cathode-follower circuit mentioned in Chapter 12, various forms of which are illustrated in *Fig.* 13—8, is most commonly used in voltage applications. The voltage gain is always less than one, indicating that the stage can serve no purpose simply as a voltage amplifier, however low the distortion. Other characteristics, however, make the cathode follower an extremely useful circuit. These characteristics include:

1. Very low input capacitance—little more than the grid-plate inter-electrode capacitance of the tube—making possible use of the cathode follower as a buffer stage between a high-impedance source and a high-capacitance load.
2. Near-constant-voltage operation associated with the 100 per cent negative voltage feedback. The range of load impedance values that can be used is much greater than in conventional circuits, making it possible to match the characteristic impedance of low-impedance lines.
3. The output has one side grounded and is in phase with the input.
4. Very low distortion, wide frequency range, and low phase shift.
5. High signal-handling capability.

When a signal is applied to the grid of the cathode follower, the instantaneous voltage across the cathode load resistor adjusts itself to a value somewhat less than the signal voltage. The voltage "gain" varies from about .5 for very low values of load impedance, such as provided by low-impedance transmission lines, to about .9 for resistances of 10,000 ohms or more.

The voltage gain of a cathode follower, ignoring reactances, is equal to:

$$\frac{\mu R_L}{R_p+R_L}\times\frac{1}{1-A\beta}$$

$$\left[\begin{array}{l}\text{where } \beta=1\\ \qquad A=\dfrac{-\mu R_L}{R_p+R_L}\end{array}\right]$$

$$=\frac{\mu R_L}{R_p+R_L(\mu+1)}$$

where $R_L$ = Cathode load resistor
$R_p$ = Tube plate resistance

The cathode follower is useful in vacuum-tube voltmeters, where it is very important that the input impedance be very high so as not to disturb the circuit being measured, and in video applications. The cathode follower is also useful in audio systems as a high-quality, noncritical matching or buffer stage. A high impedance source—for example, the output of a preamplifier tube—may feed an amplifier that is located at an appreciable physical distance. The capacitance between the shield and central conductor of the cable provides an effective shunt across high-impedance circuits and attenuates the treble frequencies. In ad-

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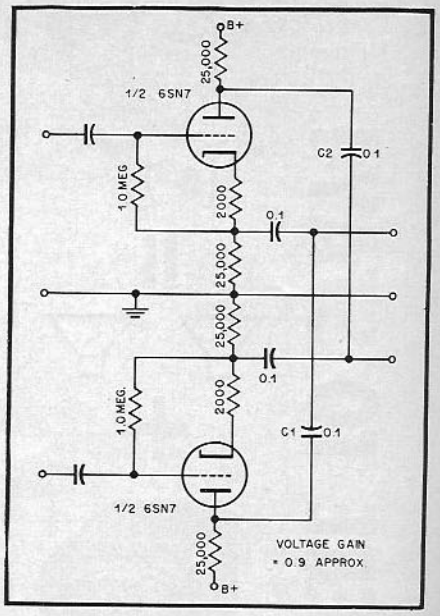

**Fig. 13—11. "Phase compressor" for correcting an imperfectly balanced push-pull signal.**

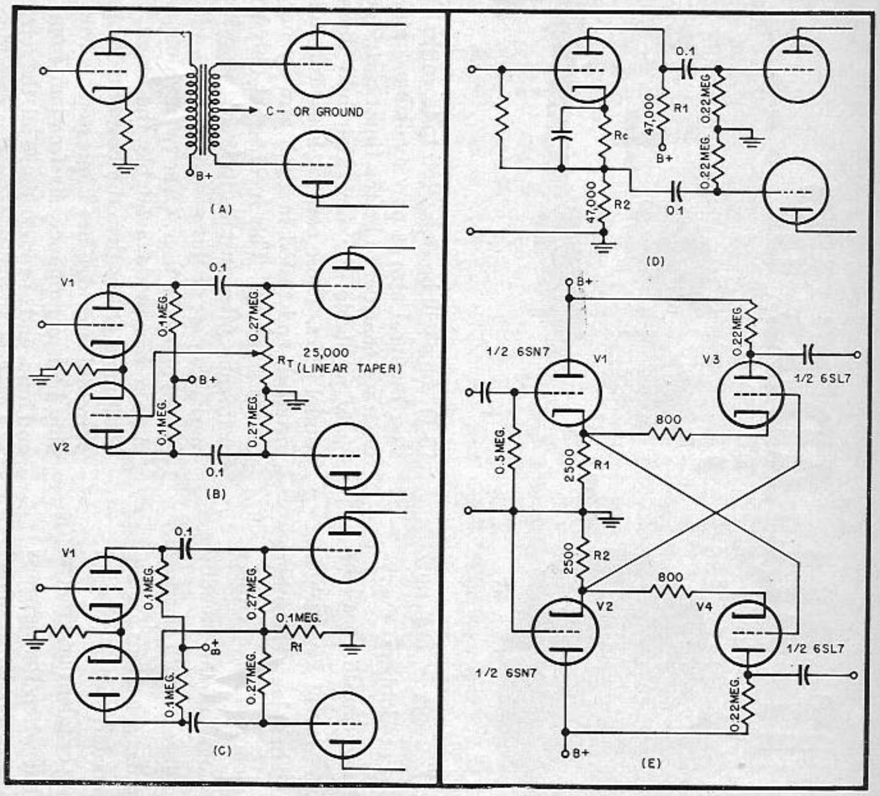

**Fig. 13—10. (A) Transformer phase splitter. (B) Tapped-voltage phase splitter, with common circuit values. (C) Floating-paraphase phase splitter. (D) Split-load or cathodyne phase splitter. $R_c$ is selected on the basis of the total load, $R_1 + R_2$. (E) Cross-coupled phase splitter.**

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*[From page 48]*

dition, the line which is a physical extension of the circuit beyond the boundaries of the chassis—is susceptible to hum pick-up. Now suppose that a cathode-follower stage is inserted between the preamplifier and the line, as in *Fig.* 13—9. The capacitance presented to the preamplifier plate is negligible, while the resistance across the line is low enough so that the shunt cable capacitance loses any significance at audio frequencies. The impedance of the parallel combination of cable capacitance and output resistance varies very little over the signal frequency range; even if a higher load value were used the feedback secures low source impedance and near-constant-voltage operation. The danger of hum pickup from the cable is likewise reduced or eliminated.

The ill effects of a long cable connecting high-impedance circuits are thereby overcome, at the price of a virtually distortionless extra stage and a slight loss of voltage. Use of the cathode follower for this application is undoubtedly cheaper and more effective (in low-current circuits) than a pair of step-down and step-up transformers connected at the preamplifier output and amplifier input respectively.

A tube used as a cathode follower requires the normal operating grid bias. If the value of the cathode load resistor is chosen to equal the correct bias resistance the problem is solved automatically. In *Fig.* 13—8, (B) and (C) illustrate methods of providing bias for cathode followers when the load resistance is either lower or higher than the correct bias resistor value.

### Phase Splitters

An amplifier stage located between the plate of a single-ended stage and the grids of a push-pull amplifier is the phase splitter. The signal channel up to the phase splitter is "unbalanced," which is to say, the signal potentials are on one line relative to ground. The phase splitter converts the signal channel to a balanced one, and ground becomes the midpoint between two lines whose instantaneous potentials are always opposite in polarity.

*Figure* 13—10 illustrates various ways of dividing the signal into two components identical in every way except that they are opposite in phase. In (A) advantage is taken of the fact that the two ends of the transformer secondary are 180 deg. out of phase with each other. Transformer coupling has the disadvantage of requiring an input transformer of critical characteristics and relatively high price, but the method is simple, stable, and suited to fixed-bias circuits.

(B) illustrates an often-used design sometimes referred to as the "tapped voltage" circuit. A fraction of the output of $V_1$ is tapped off by a voltage divider and applied to the grid of $V_2$. The output signal of a vacuum-tube is always reversed in phase from the input, so that the output of $V_2$ is 180 deg. out of phase with the output of $V_1$. (Since $V_2$ is in the circuit solely for phase reversal purposes it is properly called a *phase inverter*. This term is often used, although incorrectly, to apply to any complete phase-splitter circuit.) If the fraction of the signal voltage tapped off is equal to the reciprocal of the gain of $V_2$, the output of $V_2$ will be equal in amplitude although opposite in phase from the output of $V_1$, and the two signals may be applied to the inputs of a push-pull stage. $R_T$ may be a fixed resistor of approximately correct value, or it may be a potentiometer that allows a balance adjustment.

The operation of the circuit of (C) is somewhat similar, except that the voltage tapped off by $R_1$, the lower arm of the voltage divider, is an appreciable part of the total voltage, commonly on the order of 30 per cent. This would create an output from $V_2$ far in excess of the output of $V_1$, but the signal from $V_2$ is applied to a voltage divider with the same resistor $R_1$ as lower arm. The two signals across $R_1$ are opposite in phase and tend to cancel until push-pull balance is achieved. The circuit is called a "floating paraphase," because the top of $R_1$ has a potential which is not anchored, relative to a single signal source, in the conventional manner. The two plate resistors of $V_1$ and $V_2$ are often made equal, but correct balance usually requires that the plate resistor of $V_2$ have a value slightly greater than the resistor of $V_1$. If a variable resistance is used for the plate resistor of $V_2$ the circuit may be balanced manually, and the limited self-balancing characteristics relied upon to maintain balance over a period of time. The paraphase circuit has potentially greater stability than the circuit of (B) for the same number and quality of parts. In both circuits the use of a common, unbypassed cathode resistor for $V_1$ and $V_2$ (with a value half of that required for the single tube) is advantageous for balance. When the signals are perfectly balanced there is no feedback, but when there is any unbalance negative current feedback is applied to the stage with the larger signal, and effectively positive current feedback to the stage with the smaller one. This is a permissible condition when the inherent distortion is low and when the stage is within a negative feedback loop.

The use of a partial cathode follower for phase splitting is illustrated in (D) of *Fig.* 13—10. All of the voltage across $R_2$ is fed back to the input, and therefore the output voltage at $R_2$ is less than the input voltage. If $R_1$ is equal to $R_2$ the signal across $R_1$, which is $IR_1$, must equal the signal across $R_2$, which is $IR_2$; these two signals are balanced relative to ground, and together have an amplitude equal to about 1.8 times the input signal. The circuit is extremely stable, because with the same signal current necessarily flowing through $R_1$ and $R_2$

(and their parallel grid resistors), balance is entirely dependent upon the equality of resistor values, and independent of tube characteristics or of operating voltages. Plate resistors and the following grid resistors should be matched by actual measurement rather than by reliance on color codes.

The cathode-loaded phase splitter has very low distortion and the ability to handle large input signals. It will be seen that the output across $R_1$ is subject to negative current feedback, and that the output across $R_2$ is subject to negative voltage feedback. The main drawback of the "split-load" or "cathodyne" circuit is a matter of economy, in space and cost; other circuits furnish appreciable gain. The fact that the cathode is well above ground potential also makes the stage susceptible to hum.

The cross-coupled phase splitter,[3] illustrated in (E) of Fig. 13—10, is a self-balancing circuit in which the two voltage amplifiers $V_3$ and $V_4$ are driven by cathode followers $V_1$ and $V_2$. Note that $R_1$ and $R_2$ are connected as cathode resistors for $V_3$ and $V_4$ respectively, and as grid resistors for $V_4$ and $V_3$ respectively, so that the voltage across either is coupled to both $V_3$ and $V_4$, but in opposite phase. (Cathode-to-ground and grid-to-ground signal voltages are always opposite in sense).

A separate circuit for correcting an imperfectly balanced push-pull signal[4] appears in *Fig.* 13—11. Restoration of balance is achieved through the cross-coupling capacitors, $C_1$ and $C_2$ connected to the cathode follower outputs.

REFERENCES

L. B. Arguimbau, *Vacuum-tube Circuits,* John Wiley and Sons, Inc., New York, N. Y., 1948, Chapter 3.

E. Milton Boone, *Circuit Theory of Electron Devices,* John Wiley and Sons, Inc., New York, N. Y., 1953, Chapter 3.

Herbert J. Reich, *Theory and Application of Electron Tubes,* McGraw-Hill Book Co., Inc., New York, N. Y. 1944, Chapters 4 and 6.

F. Langford Smith (editor), *The Radiotron Designer's Handbook,* 3rd ed., Wireless Press, Sydney, Australia, (Reproduced by Radio Corporation of America) Chapters 1 and 8.

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[3] J. N. Van Scoyoc, "A cross-coupled input and phase inverter circuit," *Radio and Television News,* Nov., 1948.

[4] D. H. Parnum, "The phase compressor—a resistance-capacity output circuit complementary to the phase splitter," *Wireless World,* Jan., 1945; Correction Feb., 1945.

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