CHAPTER 11

DIRECT-COUPLED AMPLIFIERS

INTRODUCTION

A direct-coupled amplifier is one in which the input, output, and interstage couplings are conductive, i.e., direct connections or resistive networks. No "blocking" condensers or coupling transformers are employed. Thus, d-c signals may be amplified, whereas the bandpass characteristics of other types of amplifiers drop to zero gain at zero frequency.

Usually zero frequency is the most important part of the pass band, and it has the highest gain, except in certain servoamplifiers where, for damping purposes, the d-c is attenuated in comparison with low audio frequencies. Where rapid fluctuations are to be followed, the pass band must extend correspondingly high in the frequency spectrum, but only in a very special case would a direct-coupled amplifier be expected to pass frequencies much above the audio spectrum.

11·1. Applications of Direct-coupled Amplifiers.—Below are listed several fields of application, with a brief statement of some special features of each. A direct-coupled ampl fier is not the only type applicable in each of these cases, since a sequence of modulation, a-c amplification, and demodulation may sometimes be used.

Vacuum-tube Voltmeter.—The purpose of the direct-coupled amplifier in a vacuum-tube voltmeter is to convert, with a linearity and constancy of gain of from 5 to 0.1 per cent, a d-c voltage derived from a high-impedance source into a current in a low resistance. Precision of linearity and gain may be achieved by the use of negative feedback. Except for battery-operated types, the instrument ought to operate independently of supply-voltage variations. The input impedance should be low enough so that the grid current of an ordinary low-power tube will cause negligible additional loading.

Electrometer.—This is a class of vacuum-tube voltmeter in which special precautions are taken to keep the input current very low. Special tubes are used, and low electrode voltages are employed in order to minimize ionization, which would cause grid current to lower the input impedance. Batteries are the most common source of power.

Resistance Measurement.—This application of direct-coupled amplifiers includes ohmmeters for the measurement of high resistance, bolometers, resistance-strain gauges, and photocell amplifiers. Usually a vacuum-tube voltmeter is used in some sort of a bridge circuit.

Mirror Oscillograph Drivers.—A direct-coupled amplifier is needed when, as is often the case, a record of the d-c component of the signal is desired. A higher frequency limit to the pass band and, in general, more current output are required than for a vacuum-tube voltmeter.

Cathode-ray-tube Deflection.—A direct-coupled amplifier is required for some types of measurement with oscilloscopes, such as the measurement of tube characteristics. For electrostatic deflection a large linear voltage swing, in push-pull form but with very little power output, is needed. Magnetic deflection requires large current and power, and the inductance of the coils makes the bandpass problem rather difficult.

Relays and Solenoids.—It may be desired to operate a magnetic device when a d-c voltage, current, or a resistance reach some predetermined value. Linearity is not important here, but sensitivity and stability are.

D-c Servoamplifier.—In this case, a motor must be driven in a direction corresponding to an "error signal," which comprises the amplifier input signal and which the resulting motion of the motor will tend to reduce or reverse. The input signal may be the difference between two variable d-c voltages or between a variable voltage and a fixed potential. The amplifier may furnish all the power to the motor but often only controls its motion by a differential field control or some other device. "Proportional control" is usually, although not always, desired; it implies a rough linearity between error input and motor-control output in the neighborhood of zero error. For large errors the output is expected to be limited in the proper sense. Since the servo loop itself comprises a complete negative-feedback system, no resistive negative feedback is normally employed in the amplifier except, perhaps, to reduce the d-c gain in favor of a-c gain to achieve better damping. D-c stability and sensitivity are the important factors in the amplifier.

Voltage and Current Regulators.—A voltage-regulator amplifier is similar to a servoamplifier, in that an error voltage, the difference between a standard and the regulated output or a fraction thereof, is amplified and used to control the output in order to minimize the error. In a current regulator the output current is passed through a constant resistance, and the resulting voltage is compared with the standard voltage to give the error.

Impedance Changing.—When a variable d-c voltage supplied by a high-impedance source is required to operate a low-impedance load, a d-c impedance changer of unit voltage gain is needed. This, depending on the requirements, may be a simple cathode follower or a higher-gain direct-coupled amplifier with negative feedback.

Phase-splitting; Subtracting; Level Changing; Summing.—These and similar operations that are sometimes performed in d-c computing circuits can be performed by certain types of direct-coupled amplifiers, usually with large negative feedback.

Integration and Differentiation.—High-gain direct-coupled amplifiers are used in the "amplified time constant" methods of voltage integration and differentiation. In the first case, a series resistance is connected to the input terminals, and a condenser gives over-all negative feedback. In the second case, the resistance and condenser functions are reversed. Thus, the differentiator is not direct-coupled to its input terminal and is independent of the input level, but interstage and output coupling are direct.

11·2. Problems Peculiar to Direct-coupled Amplifiers.—The fact that conductive coupling must be used throughout results in several special problems of design and stability. One obvious design problem arises because plate potentials are considerably higher than grid potentials. It is therefore necessary either to have the cathode of a following stage at this higher potential from ground or to use a device to lower this potential the correct amount. Since batteries are usually impractical here, a negative supply voltage and a resistance divider are generally required, with resulting decrease in gain.

The "zero adjustment" problem is the most difficult one resulting from conductive coupling. If the effects of noise and pickup are neglected, a nonconductively coupled amplifier gives zero output for zero input regardless of circuit parameters such as resistor values, supply voltages, and tube characteristics. This is not true of a direct-coupled amplifier; the d-c output at a given d-c input (e.g., zero) depends on all these things. A certain output current or voltage is usually required for a specified input. In order to meet this requirement some kind of "zero" adjustment is needed to compensate for variation of the above parameters (1) between amplifiers of the same design, due to tolerances of like components, and (2) because of effects of aging, temperature variation, vibration, etc. For rational circuit design, the expected extreme limits of voltage, resistor, and tube characteristics must be known, as well as their effects on the circuit. The zero adjustment should provide sufficient latitude to compensate for these extremes, and the incidence of these extremes, together with the appropriate zero adjustment, should not put any part of the circuit out of its good operating range.

Voltage-supply variation will depend on the kind of regulation provided. Its effect on zero shift and some methods of minimizing or canceling this effect will be discussed in connection with specific circuits.

Resistor tolerances, temperature characteristics, and stabilities are covered in another volume.[1] The use of wire-wound or other high-stability resistors will be recommended for certain critical positions in precision amplifiers. The effects of production tolerances, heater-voltage variation, and aging on the static characteristics of vacuum tubes will also be examined.

SPECIAL ASPECTS AND EFFECTS OF VACUUM-TUBE PROPERTIES

11·3. Variability of Vacuum-tube Characteristics.—The Joint Army-Navy Specification, JAN-1A, for Electron Tubes[2] is a standard of what limits of variability may be expected in tubes of any given type. Pertinent data therefrom, for common tube types, are presented in the *Components Handbook.* Maximum and minimum plate current are specified for given plate voltage and grid bias or self-biasing cathode resistor. The limits of transconductance and amplification factor are also given for the same test conditions.

In direct-coupled amplifiers, the operating plate current and voltage are usually established by design, and grid bias is then adjusted to compensate for tube tolerance. Therefore, it is desirable to interpret the limits in terms of grid bias. For the conditions of the JAN test, the grid-bias spread may be computed by dividing the plate-current spread by average transconductance. If, in the test, bias were obtained from a cathode resistor, the change in bias between upper and lower limits would have to be added to the above figure. This value is obtained by multiplying the biasing resistance by the spread of cathode current. The three examples following illustrate the computation:

1. *Fixed bias, pentode:*
   Tube:
   6SJ7 Sharp-cutoff pentode.
   Test conditions:
   $E_b = 250$ volts, $E_{c2} = 100$ volts, $E_{c1} = -3$ volts.
   Plate current:
   Minimum $I_p = 2.0$ ma, maximum $I_p = 4.0$ ma, $\Delta I_p = 2.0$ ma.
   Transconductance:
   Minimum $S_m = 1325$ μmhos, maximum $S_m = 1975$ μmhos,
   Average $S_m = 1650$ μmhos.
   Computed grid bias tolerance:
   $$\Delta E_c = \frac{\Delta I_p}{S_m} = \frac{2.0}{1.65} = 1.21 \text{ volts.}$$

2. *Cathode bias, triode:*
   Tube:
   6J6 (one section) miniature double triode.
   Test conditions:
   $E_b = 100$ volts, $R_k = 50$ ohms, grid grounded.
   Plate current:
   Minimum $I_p = 5.5$ ma, maximum $I_p = 12.5$ ma,
   $\Delta I_p = 7.0$ ma.
   Transconductance:
   Minimum $S_m = 4000$ μmhos, maximum $S_m = 7300$ μmhos,
   Average $S_m = 5650$ μmhos.
   Actual test grid bias ($-R_kI_b$):
   Minimum $E_c = -0.275$ volts, maximum $E_c = -0.625$ volts.
   Computed grid-bias tolerance:
   $$\Delta E_c = \frac{7.0}{5.65} + 0.35 = 1.6 \text{ volts.}$$

3. *Cathode bias, tetrode or pentode:*
   Tube:
   6AJ5 low-voltage miniature pentode.
   Test conditions:
   $E_b = E_{c2} = 28$ volts, $R_k = 200$ ohms, grid grounded.
   Plate current:
   Minimum $I_p = 1.8$ ma, maximum $I_p = 4.0$ ma, $\Delta I_p = 2.2$ ma.
   Screen current:
   Minimum $I_{c2} = 0.1$ ma, maximum $I_{c2} = 2.0$ ma.
   Transconductance:
   Minimum $S_m = 2000$ μmhos, maximum $S_m = 3500$ μmhos,
   Average $S_m = 2750$ μmhos.
   Cathode current (sum of $I_p + I_{c2}$):
   Minimum $I_k = 1.9$ ma, maximum $I_k = 6.0$ ma.
   Actual test grid bias ($-R_kI_k$):
   Minimum $E_c = -0.38$ volts, maximum $E_c = -1.2$ volts
   Computed grid-bias tolerance:
   $$\Delta E_c = \frac{2.2}{2.75} + 0.82 = 1.62 \text{ volts.}$$

The tolerances of the first stage of a direct-coupled amplifier have by far the greatest effect on the zero adjustment. This stage is most often a voltage amplifier, operating with plate voltages and currents considerably lower than were employed in the JAN tests. The question therefore arises as to the validity of the tolerance computed as above.

In the case of a triode, if the assumption is made that the differences between samples of a tube type lie only in plate resistance and in amplification factor and that the characteristics are linear and obey the formula

$$I_p = \frac{E_b + \mu E_c}{r_p},$$

it follows that variations of μ and/or r_p result in the following variation in grid bias required at a given plate voltage and current:

$$\Delta E_c = \frac{I_p}{\mu} \Delta r_p - \frac{E_c}{\mu} \Delta \mu.$$

The same sort of formula applies to a pentode if μ and r_p refer to screen voltage. Actually, about the only generalization that can be made is that the variation between tubes measured at the grid is less for small currents and grid biases.

11·4. Vacuum-tube Characteristics at Low Currents.—In voltage amplification, which is the function of all but the output stage in most d-c amplifiers and of the output stage as well in some, it is generally desirable to employ each vacuum tube in such a way as to obtain the maximum gain, with little regard for the resulting impedance levels. This involves operation at rather low currents, for which the ordinary published tube data are unsatisfactory for use in a critical analysis. This section is largely the result of special measurements of vacuum-tube characteristics in the low-current region, from which certain generalizations are made regarding voltage amplifiers.

In hot-cathode vacuum tubes the best simple formulation of the relationship between anode current and electrode potentials is generally considered to be the three-halves power law. According to this law, the plate current in a diode is proportional to $e_{pk}^{3/2}$, and in a triode the total anode current (including the grid current if the grid bias is so small that this current is appreciable) is proportional to $[e_{gk} + (e_{pk}/\mu)]^{3/2}$. The basis for the derivation of this law is the assumption of a copious supply of electrons at the cathode surface, all of which are emitted with no excess energy so that the potential gradient at the cathode surface can be zero. For large currents, where e_{pk} or $[e_{gk} + (e_{pk}/\mu)]$ is large compared with the initial electron velocities (expressed in electron-volts), the three-halves power law is valid. For low currents an entirely different relationship obtains.

The portion of the total number of electrons emitted from a hot surface having normally directed velocities greater than e electron-volts is $\epsilon^{-\frac{Te}{11{,}600}}$, where T is the temperature in degrees Kelvin. Thus, if this expression represents the plate current of a diode, there must be a point in the tube e volts below the cathode potential that the rest of the electrons have insufficient energy to attain. This minimum potential usually results from a space charge of excess electrons; but if the plate potential is itself negative, the minimum potential in the tube may actually be at the plate. Below the plate potential where this becomes true, the plate current is an exponential function of plate voltage. (Contact potential also is important, but it can be considered to be due to a constant voltage added to the plate supply.) The exponent is such that, for oxide-coated cathodes operating at from 1000° to 1100° K, an increment of plate voltage of about 0.21 volt produces a 10-fold increase of plate current. If the $\log_{10}$ of current is plotted against plate voltage, the result should be a straight line with a slope of 1/0.21.

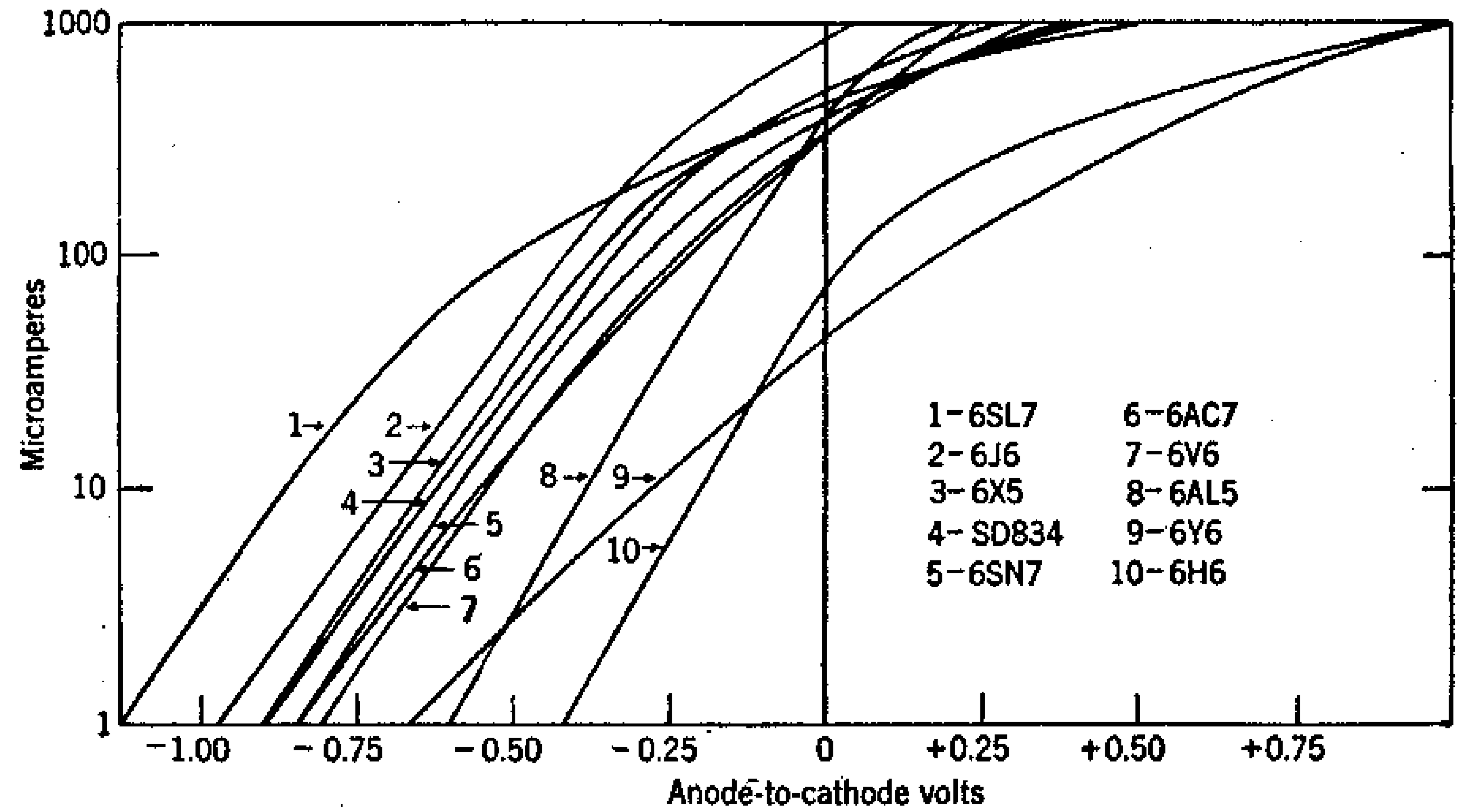

FIG. 11·1.—Low-current characteristics of several diodes and diode-connected tubes, with logarithmic current scale.

The preceding theory is fairly descriptive of all tubes with unipotential cathodes, at low currents. Figure 11·1 shows the logarithmic characteristics of samples of several tube types, either diodes or connected as such (grids connected to plate). The curves are straight for currents below 50 to 500 μa, depending on cathode area and spacing, and most of them have about the same slope: approximately 1/0.25. (These curves are not necessarily typical of the various tube types. Considerable shift in the voltage scale is encountered from tube to tube. For example, for several 6SL7's, the voltage at 1 μa ranged from −0.7 to −1.3 volts, although the slopes and shapes of the curves were fairly consistent.)

In a triode operated as such, where potential differences exist between plate and grid, the relationship of anode current to grid-cathode potential seems still to be logarithmic, up to the same current level as it is when the plate-to-grid potential is zero. The slope of the curve is decreased by the plate-grid potential, probably because of distortion of the unipotential surfaces around the grid. This distortion and the decrease of slope of the curve are roughly proportional to grid bias and therefore to $1/\mu$ times the plate potential. Thus, the curve at a given plate potential is, in general, steeper for a high-μ than for a low-μ triode.

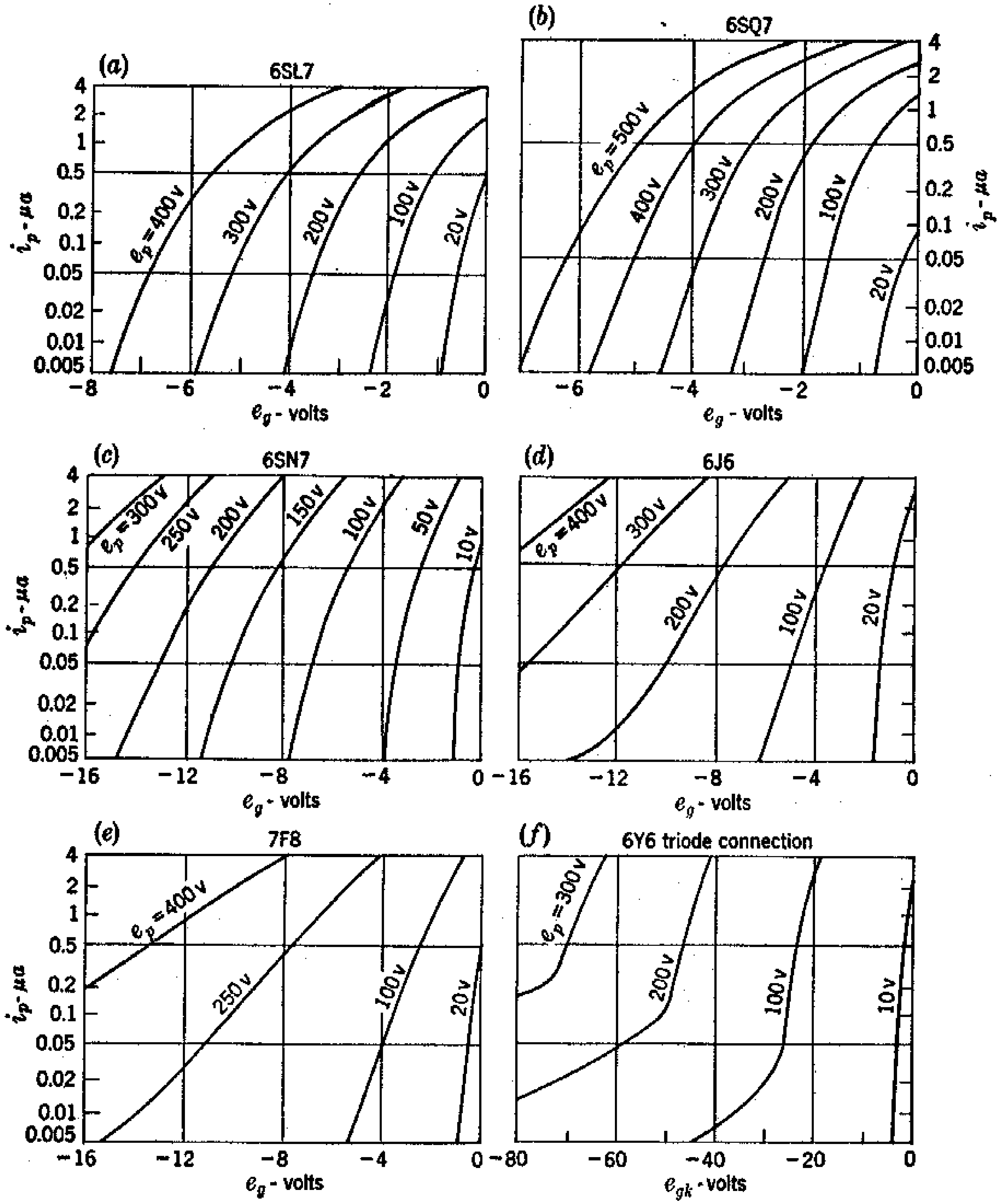

FIG. 11·2.—Logarithmic low-current characteristics of certain triodes.

The curves of Fig. 11·2 are the conventional "mutual" or "transfer" characteristics of certain triodes, but with plate current in a logarithmic scale. The lower portions of the curves are fairly straight and have slopes that approach that of the diode-connected tube for low plate potentials but that decrease from this value as plate voltage is increased. (An aberration is evident in the case of the triode-connected 6Y6—the loss of control at low currents, because of leakage around the grid.) As explained above, this decrease of slope with respect to plate voltage is less noticeable for the higher-μ triodes.

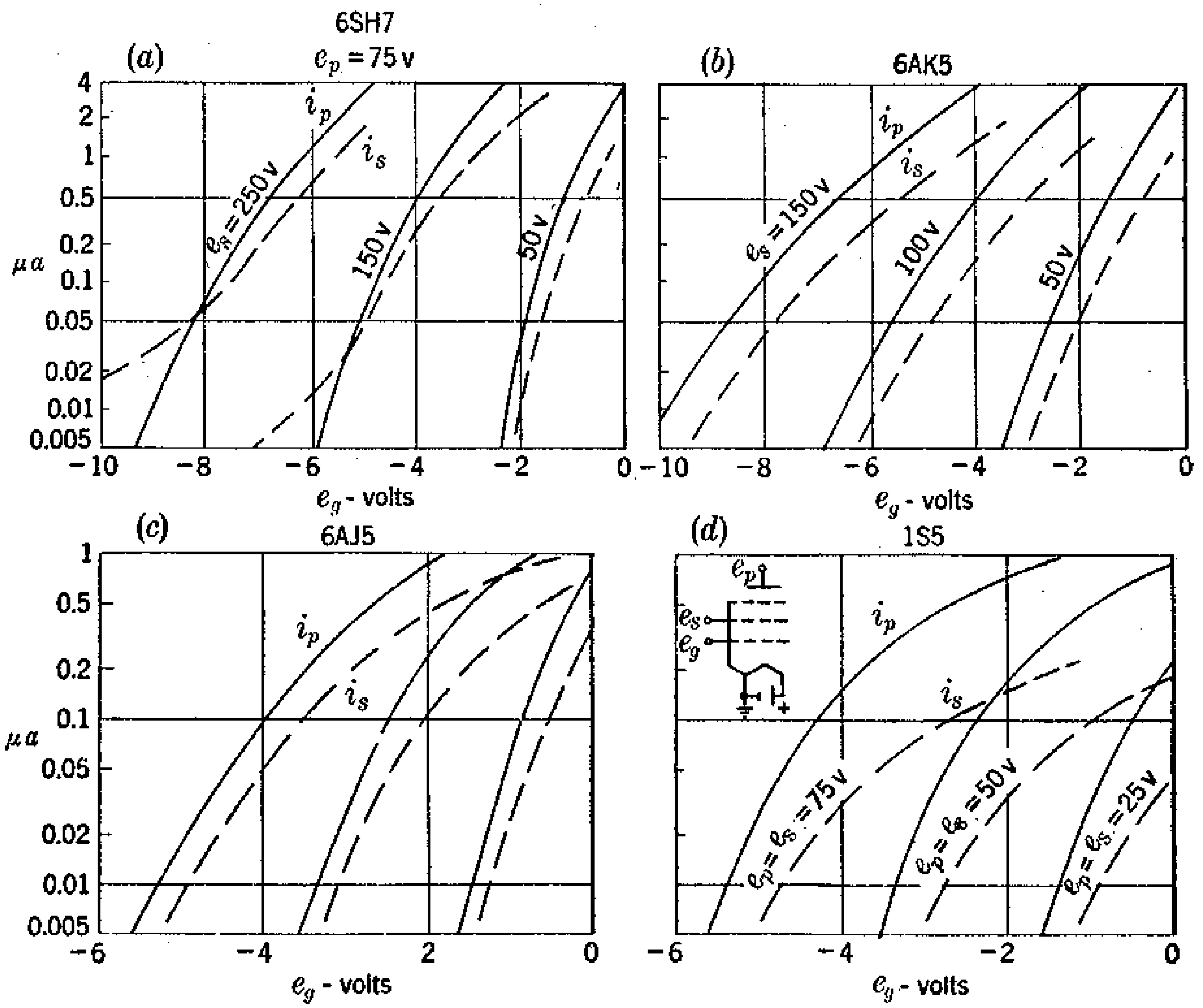

FIG. 11·3.—Logarithmic low-current characteristics of certain pentodes.

If E_0 is the reciprocal of the slope of any curve in Fig. 11·2, i.e., the plate current increases 10-fold for each E_0 increment of e_{gk}, the mutual conductance at this point is

$$g_m = \frac{2.3 i_p}{E_0}$$

and is therefore proportional to the plate current while the curve is straight. It is evident that for a given plate voltage and plate current, the g_m in the low-current region is greater for high-μ than for low-μ triodes, regardless of rating. Also, for a given triode at a given plate current, g_m is greater at lower plate voltages.

From the curves and from the formulas developed in a later section, it will be evident that the maximum voltage gain is obtainable in a d-c amplifier if the tube operates at as low a plate voltage as possible and at a plate current corresponding to the top of the straight part of the curve as in Fig. 11·2.

Similar pentode characteristics are given in Fig. 11·3, from which it is evident that the maximum gain obtains at as low a screen-grid voltage as is permissible without resulting in the flow of control-grid current.

11·5. Grid Current.—A knowledge of the grid-current characteristics of vacuum tubes is essential to an understanding of certain important limitations of direct-coupled amplifiers.

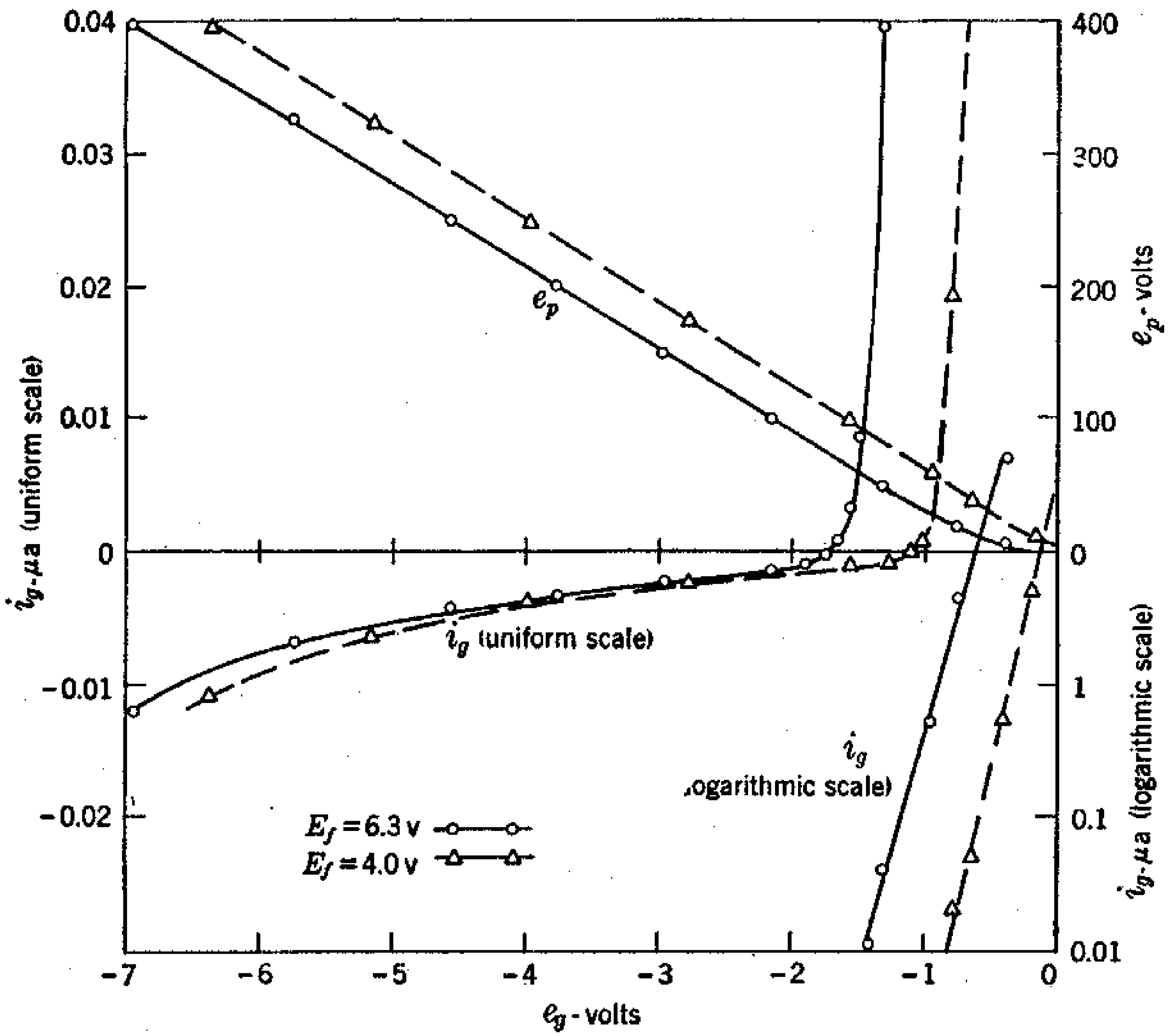

FIG. 11·4.—Negative and positive grid current in a triode operated with constant-plate current (6SL7; i_p = 0.1 ma).

Negative Grid Current.—When the grid is negative with respect to the cathode, a current, which is the result of positive ions that are attracted to the grid, flows from the grid terminal. The ionization of the residual gas in the tube depends on voltage gradients, and therefore the negative grid current increases with higher plate voltage as well as with greater negative grid voltage. Since the positive-ion current is the result of residual gas, its magnitude varies widely from tube to tube. For ordinary receiving tubes under usual operating-voltage conditions it may be anything from 0.00001 to 0.1 μa. The JAN specifications usually limit the grid current to 1 or 2 μa, but generally this value may be considered as being very conservative. The JAN specifications for the 6AK5 grid current is 0.1 and for the 6SU7, 0.02 μa. Thus the latter tube is a good choice where a very high input impedance is needed, although the 6SL7 and many other tube types, when operated at reasonably low plate voltage, may generally be relied on for grid currents as low as this. The effect of plate potential on positive-ion current (negative-grid current) is illustrated for one example of a high-μ triode in Fig. 11·4. (Variation of this part of the curve, from tube to tube, is considerable; the illustration is presented only to show the trend.) Plate current was held constant, so that as the grid voltage was lowered, the plate voltage rose μ times as much. Negative grid current varies about as the square of the plate potential. Reduced cathode temperature has very little effect on this relationship; a given plate potential occurs at a reduced grid bias (see Sec. 11·5), but a given plate potential produces a certain positive-ion current almost independently of cathode temperature.

"Electrometer tubes" are specially designed and constructed not only to have very little gas to be ionized but also in such a way that they are operative at very low electrode potentials in order to minimize the possibility of ionization. In this way and with special insulation, these tubes operate with grid current of the order of 10^{-16} amp. When the positive-ion current is less than about 0.0001 μa, a photoelectric grid current may be large enough to be of importance. Roberts,[1] operating a 606 pentode at plate and screen potentials of 20 volts, observed that a decrease of grid current from 10^{-10} to 10^{-12} amp resulted from shielding the tube from light. A reduced cathode temperature can help in this respect by decreasing the illumination from the cathode.

Another source of negative grid current exists in some tubes, namely, grid emission. This emission may occur if the grid is contaminated and especially if the tube power dissipation is high, even though the grid dissipation itself is negligible. It rarely amounts to more than 1 or 2 μa if the plate dissipation is not above the rating, but even this small amount may be a consideration where a high resistance is employed in the input circuit to a power stage.

Positive Grid Current.—Positive grid current, caused by electrons flowing to the grid from the cathode, is much greater and much less erratic than the negative grid current. The grid receives electrons as if it were the plate of a diode. Because of the initial velocities of the thermoelectrons, many of them have sufficient energy to arrive at the grid even if it is at a considerable negative potential with respect to the cathode. Thus, this positive grid current becomes as great as the negative (positive-ion) grid current, making the net current zero at some negative grid potential, usually about -1.0 to -1.8 volt for oxide-coated unipotential cathodes at temperatures of 1000° to 1100° K. This is the "floating-grid" potential, which the grid assumes if disconnected. It is largely independent of plate voltage but tends to be slightly more negative at the lower values.

Because of the distribution of electron initial velocities, the grid current increases exponentially with respect to grid potential, increasing itself by a factor of 10 for every 0.2- or 0.3-volt increase of potential. (The fact that the grid potential moves as the logarithm of the grid current is used in certain computing devices.) This exponential type of increase ends at from 10 to 100 μa; thereafter the nature of the increase depends on the tube type and is also considerably affected by plate voltage—the lower the plate voltage the more the grid current.

The logarithmic portion of the grid-current curve is shown in Fig. 11·4. This part of the curve shows much less variation from tube to tube than the negative portion. A reduction of cathode temperature reduces the grid current at a given bias, but it also increases the plate potential at a given plate current. The result is that if the grid current is not to exceed some predetermined magnitude, for a certain plate current the minimum allowable plate potential is not affected by cathode temperature.

Minimum allowable plate potential is of great importance in d-c amplifier design. It depends primarily on the maximum allowable positive grid current, which depends on input resistance, and on the permissible voltage error resulting from their product. For example, a grid current of 0.02 μa in a series input resistance of 1 megohm causes the actual grid voltage to be 20 mv lower than the input potential. In the case of Fig. 11·4, with a plate current of 0.1 ma, this would occur at e_p = about 50 volts, regardless of cathode temperature. In this same example, a reduction of plate current by a factor of 10 (to 0.01 ma) would lower the plate potential by only 15 or 20 volts.

The higher the μ of a triode the higher will be the minimum allowable plate potential (assuming a given allowable grid current and a given plate current), the latter being roughly proportional to the former.

In a pentode, the plate can operate at much lower potentials than in a triode, provided that the screen is at a sufficiently high potential to permit ample negative bias on the control grid. The only requirement on the pentode-plate potential is that it be high enough to keep the screen from taking all of the current; most low-power pentodes will operate satisfactorily as voltage amplifiers with plate potentials in the neighborhood of 10 volts.

11·6. The Effect of Heater-voltage Variation.—Variation of the cathode temperature of a vacuum tube in which the current is limited by electrode potentials and space charge rather than by cathode emission has the effect of varying the average initial electron velocity and therefore also the electrode voltages required to obtain a given electron flow.[1] When the plate current is very small compared with cathode emission, as is often the case in the first stage of a d-c amplifier, this effect is fairly

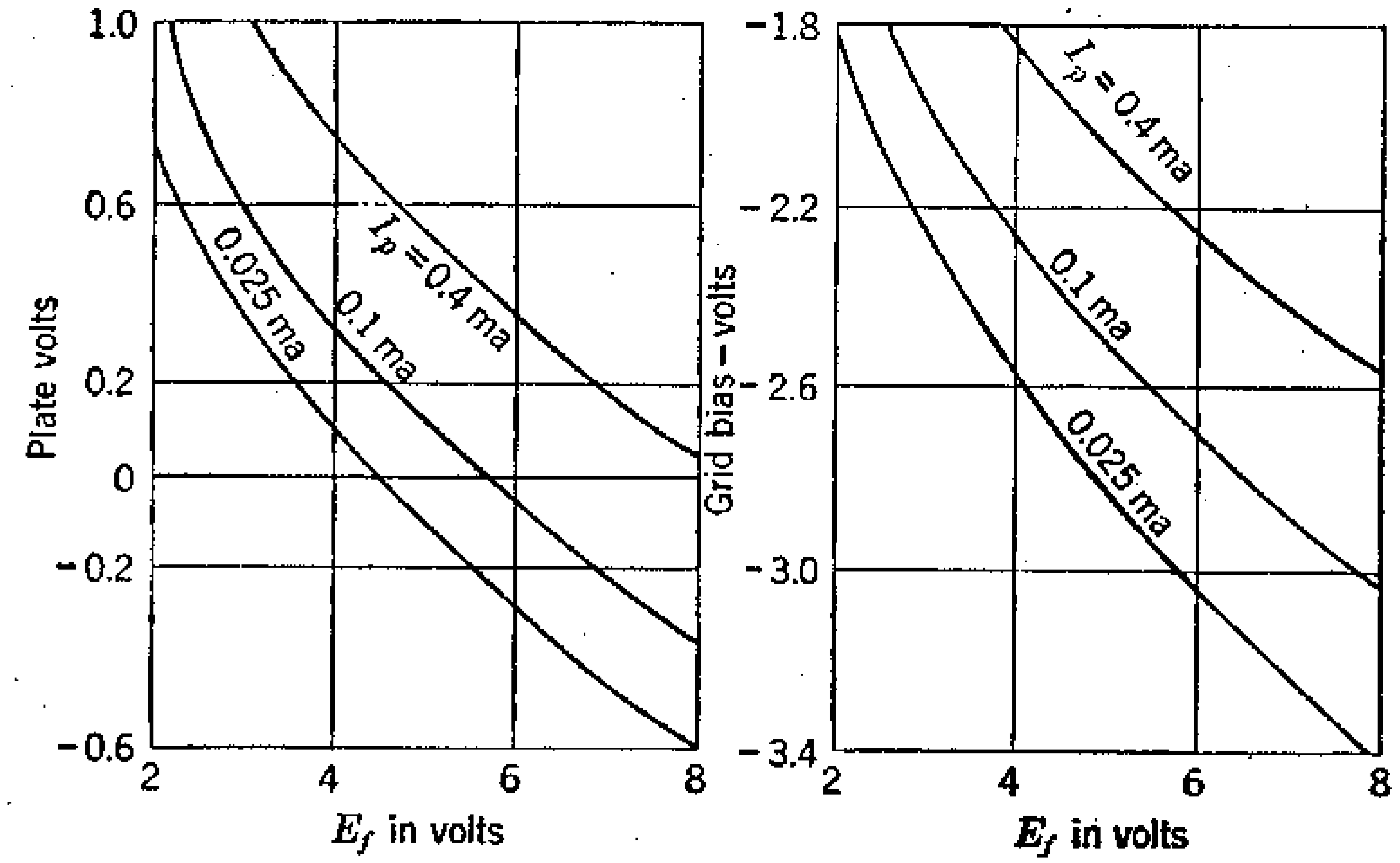

Fig. 11·5 Fig. 11·6

Fig. 11·5.—Diode heater-voltage characteristics (6SL7 as diode).
Fig. 11·6.—Triode heater-voltage characteristics (6SL7, E_p = 150 volts).

independent of plate current. It may be expressed in terms of the amount by which the cathode voltage must be changed relative to the other electrode voltages in order to hold the current constant as the cathode temperature is varied.

For oxide-coated unipotential cathodes, this amount is approximately 0.2 volt for a 20 per cent change of heater voltage about the normal value, whether the tube is a diode, a triode, or a multigrid tube. Figure 11·5 illustrates the relationship for a diode. (A 6SL7 triode with grid connected to plate was used as a diode in this test, for comparison with Fig. 11·6. An ordinary diode gives the same result, however.) It is seen that the slope $\partial E_p/\partial E_k$ at a given value of heater voltage is not greatly affected by current at the low-current values employed.

Figure 11·6 gives the same sort of information for the same tube used as a triode. Here the plate-to-cathode voltage was fixed; and as the cathode temperature varied, the current was held constant by adjustment of the grid potential. To have been completely comparable with the diode characteristics, the plate-to-grid potential should have been held constant; however, the results differ only by $1/\mu$. It is seen that the slope $\partial E_c/\partial E_h$ is about the same as was the slope $\partial E_p/\partial E_h$ for the diode and that in this case, too, it remains fairly constant with changes in plate current.

If no compensation is employed, a direct-coupled amplifier will suffer a drift in the zero with any variation in the heater voltage. As will be shown later, it is not possible to cancel this drift with inverse feedback. As measured at the input, the drift caused by the first stage is simply $(\partial E_c/\partial E_h)\Delta E_h$; i.e., about 200 mv for a 20 per cent heater-voltage change (assuming an oxide-coated cathode and low plate current). Drift in subsequent stages is less important, depending on the preceding gain.

Various practical methods of counteracting the effect of change in heater voltage will be detailed in later sections. The most important of these rely, in some manner, on the canceling effect of an auxiliary tube that has the same heater supply and therefore undergoes the same sort of drift as the amplifying tube. Usually, equivalent tube types are used for the compensation, but some methods employ a diode to compensate a triode or pentode. The use of double tubes, such as the 6SL7, is preferable, since the cathode characteristics of a pair of triodes in the same envelope, made on the same day by the same manufacturer, are likely to be more alike than the characteristics of a random pair. To check this, the characteristics of 18 6SL7 double triodes (36 triodes) were measured. For each triode was recorded the change in grid bias that was needed to hold constant a plate current of 0.2 ma at a plate voltage of 150 volts while the heater voltage was changed from 10 per cent below normal to 10 per cent above normal. The tubes were of various ages and makes. The average value of ΔE_h for the 20 per cent change was 210 mv, and 90 per cent of the values lay between 191 and 210 mv. Pairing the triodes at random without regard to envelopes, the average difference in ΔE_h between pairs[1] would be 14 mv, whereas the average difference between the two triodes in each envelope was only 8 mv.

A set of mutual characteristics for a 6SL7 is given in Fig. 11·7 for two different heater voltages. For a given current, $\partial E_c/\partial E_h$ is not affected by the value of E_p; nor is $\partial E_c/\partial E_h$ appreciably affected by the value of I_p for currents less than 1 ma, such as are employed in most voltage amplifiers. For higher currents, no longer negligible compared with cathode emission, $\partial E_c/\partial E_h$ changes somewhat more with change in current and is also subject to much more variation from tube to tube.

Table 11·1 gives the average characteristic (in terms of ΔE_c required by a 20 per cent change of E_h about normal) and dispersion of the characteristic for rather limited samples of certain tube types. Both high and low values of current were applied, and the results show that for

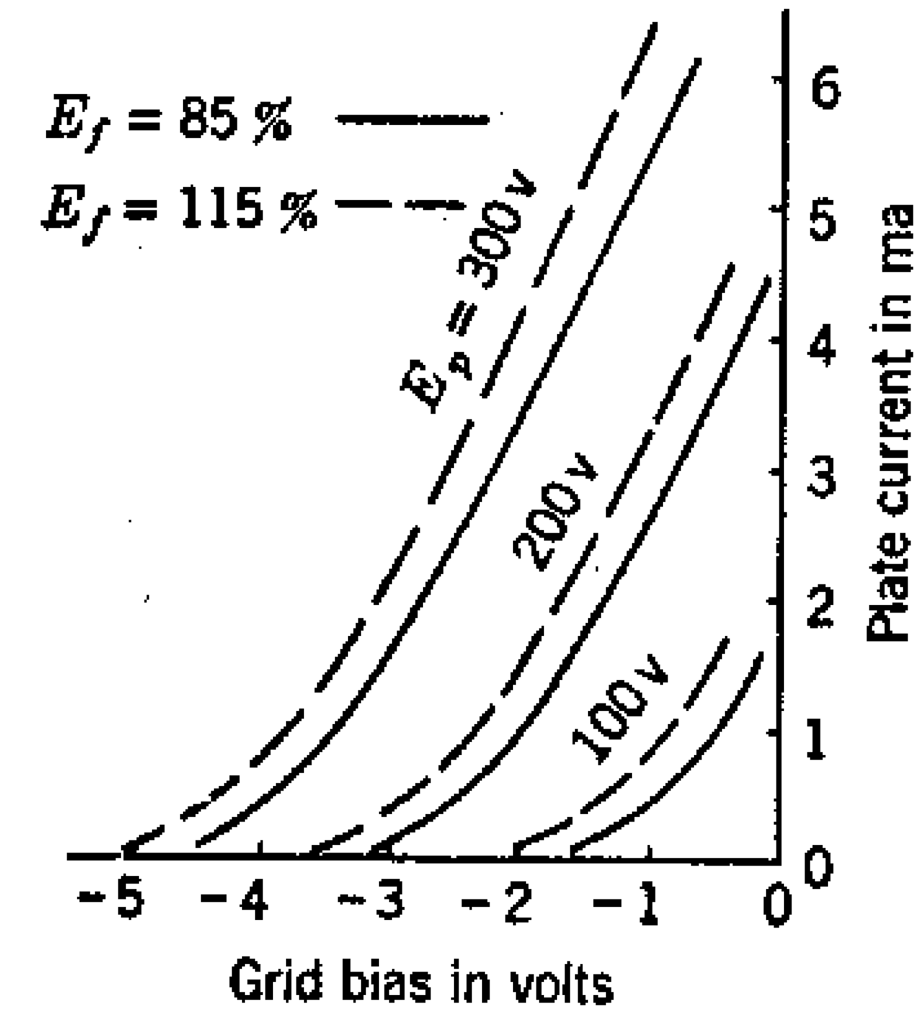

FIG. 11·7.—Effect of heater voltage on 6SL7 characteristics.

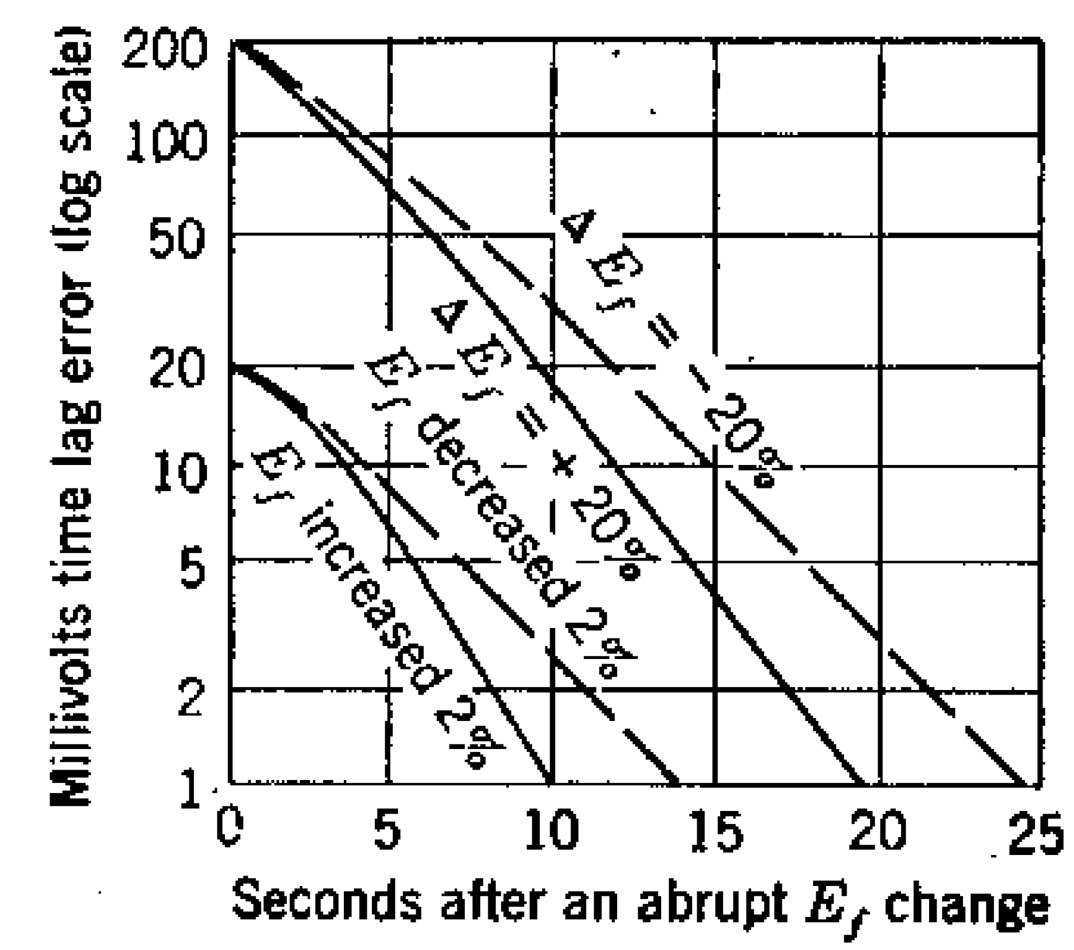

FIG. 11·8.—Heater-cathode time lag of a 6SL7.

high currents the effect of heater-voltage variation is greater and much more erratic. Pentodes were connected as triodes for these tests; this type of operation is legitimate, because it is found that the ratio between plate and screen currents is not affected by this connection.

TABLE 11·1.—EFFECT OF ELECTRODE VOLTAGES AND CURRENTS ON THE AVERAGE DRIFT RESULTING FROM HEATER-VOLTAGE VARIATION AND ON THE DEVIATION FROM THIS AVERAGE, AND THE GRID BIAS AND DEVIATION THEREFROM

Tube type	No. of tubes in sample	Plate (and screen) potential, volts	Plate (and screen) current, ma	Average grid bias (E_h = 63v), volts	Probable deviation of grid bias, volts	Average bias shift required for ΔE_h = 20 %, mv	Probable deviation of bias shift, mv
6SN7	10(20 sec.)	60	2.0	−1.65	±0.15	210	± 15
		240	10.9	−6.70	±0.40	425	± 90
6SJ7	10	60	2.5	−1.60	±0.10	215	± 15
		240	10.0	−7.20	±0.35	400	±110
6AC7	10	60	2.0	−1.30	±0.25	210	± 15
		240	8.0	−5.10	±0.60	290	± 40
6AG7	10	60	10.0	−3.20	±0.20	220	± 20
		240	40.0	−7.90	±0.25	325	± 50

[1] In a gaussian distribution of values, the average difference between pairs of values is $2\sigma/\sqrt{\pi}$, where σ is the standard deviation of the distribution.

There is a certain amount of thermal lag between a change of heater voltage and the resulting change of emission. Its magnitude, for a sudden increase or decrease of heater voltage on a 6SL7, is shown in Fig. 11·8 in terms of the lag in bias for a given current. This lag is of particular interest where the method of compensation involves the direct application of heater voltage as a circuit parameter. Where the compensation employs another tube as mentioned above, it is of interest only in so far as the lag varies from tube to tube.

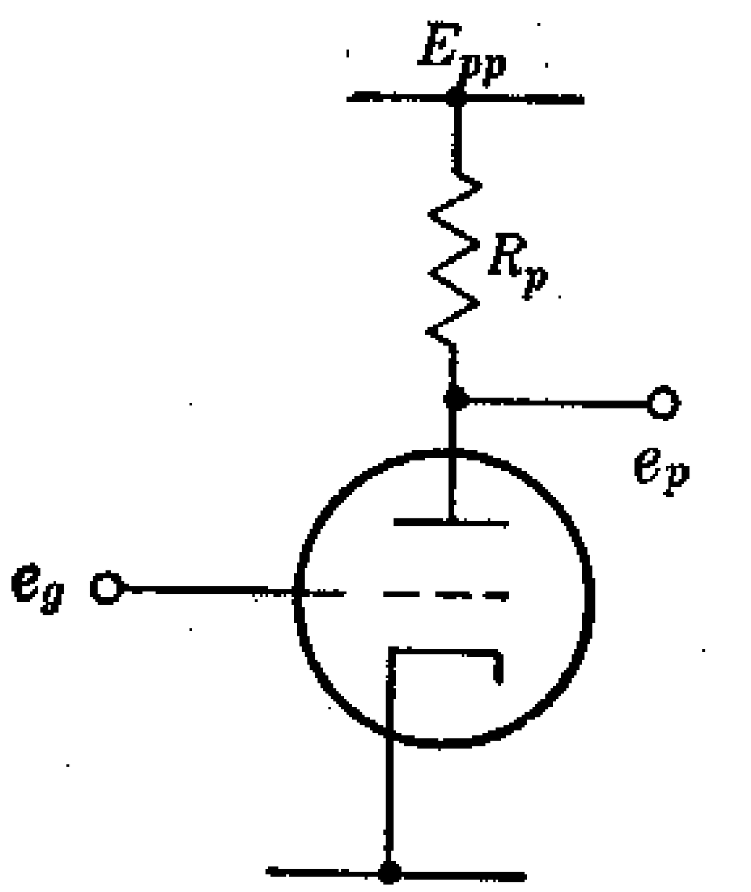

FIG. 11·9.—Simple triode amplifier.

DESIGN PRINCIPLES

11·7. Single-ended Triode Amplifiers.—The amplifier of Fig. 11·9, with fixed cathode potential, may be analyzed by drawing a load line on the graph of plate characteristics in the conventional manner (see Fig. 11·10). The plate current and plate-to-cathode voltage may then be read directly from the figure for any value of grid voltage.

In so far as the characteristics may be considered to be straight and parallel and to start from zero plate current at zero plate and grid voltage,

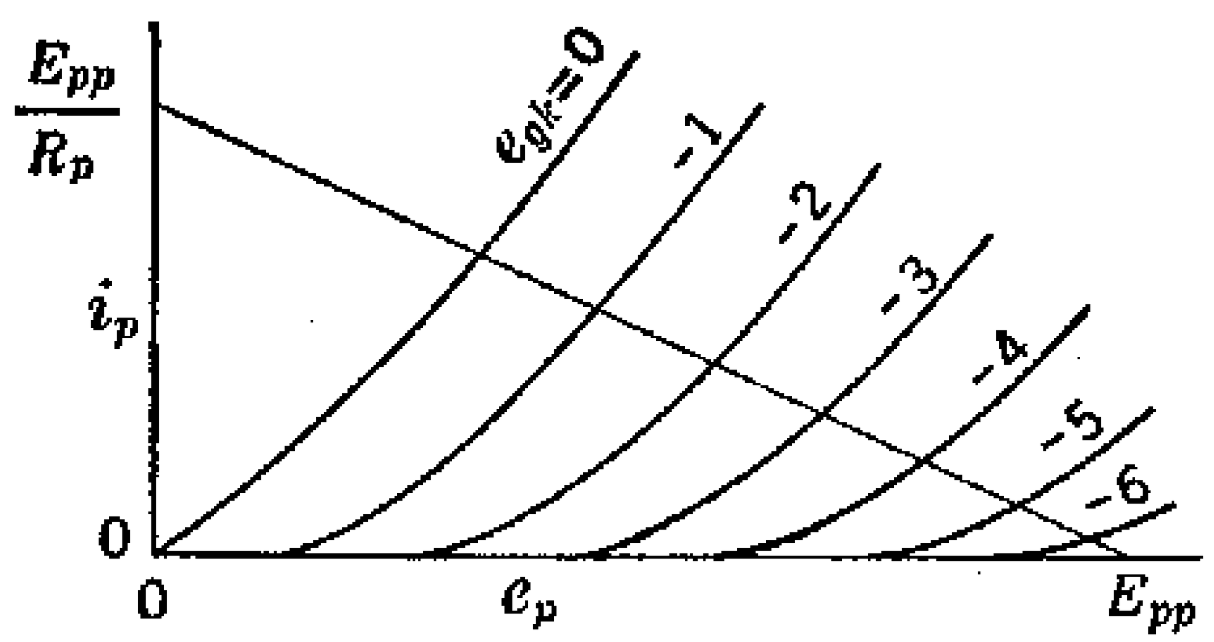

FIG. 11·10.—Plate characteristics with load line.

the operation of the amplifier may be analyzed mathematically from the formula for plate current:

$$i_p = \frac{e_{pk} + \mu e_{gk}}{r_p} \qquad (1)$$

where μ is $\left.\frac{\partial e_{pk}}{\partial e_{gk}}\right|_{i_p}$ and r_p is $\left.\frac{\partial e_{pk}}{\partial i_p}\right|_{e_{gk}}$. Substituting $E_{pp} - R_p i_p$ for e_{pk}, the plate current is found to be

$$i_p = \frac{E_{pp} + \mu e_{gk}}{r_p + R_p}, \qquad (2)$$

and the plate voltage is

$$e_p = E_{pp} - R_p i_p = \frac{\frac{r_p}{R_p} E_{pp} - \mu e_{gk}}{1 + \frac{r_p}{R_p}}. \qquad (3)$$

From Eq. (2) the current gain (considering R_p as the load) is

$$\frac{\partial i_p}{\partial e_{gk}} = \frac{\mu}{r_p + R_p}, \qquad (4)$$

and from Eq. (3) the voltage gain is

$$\frac{\partial e_p}{\partial e_{gk}} = \frac{-\mu}{1 + \frac{r_p}{R_p}}. \qquad (5)$$

Thus the current gain approaches g_m if R_p is small compared with r_p; and if R_p is large compared with r_p, the voltage gain approaches μ.

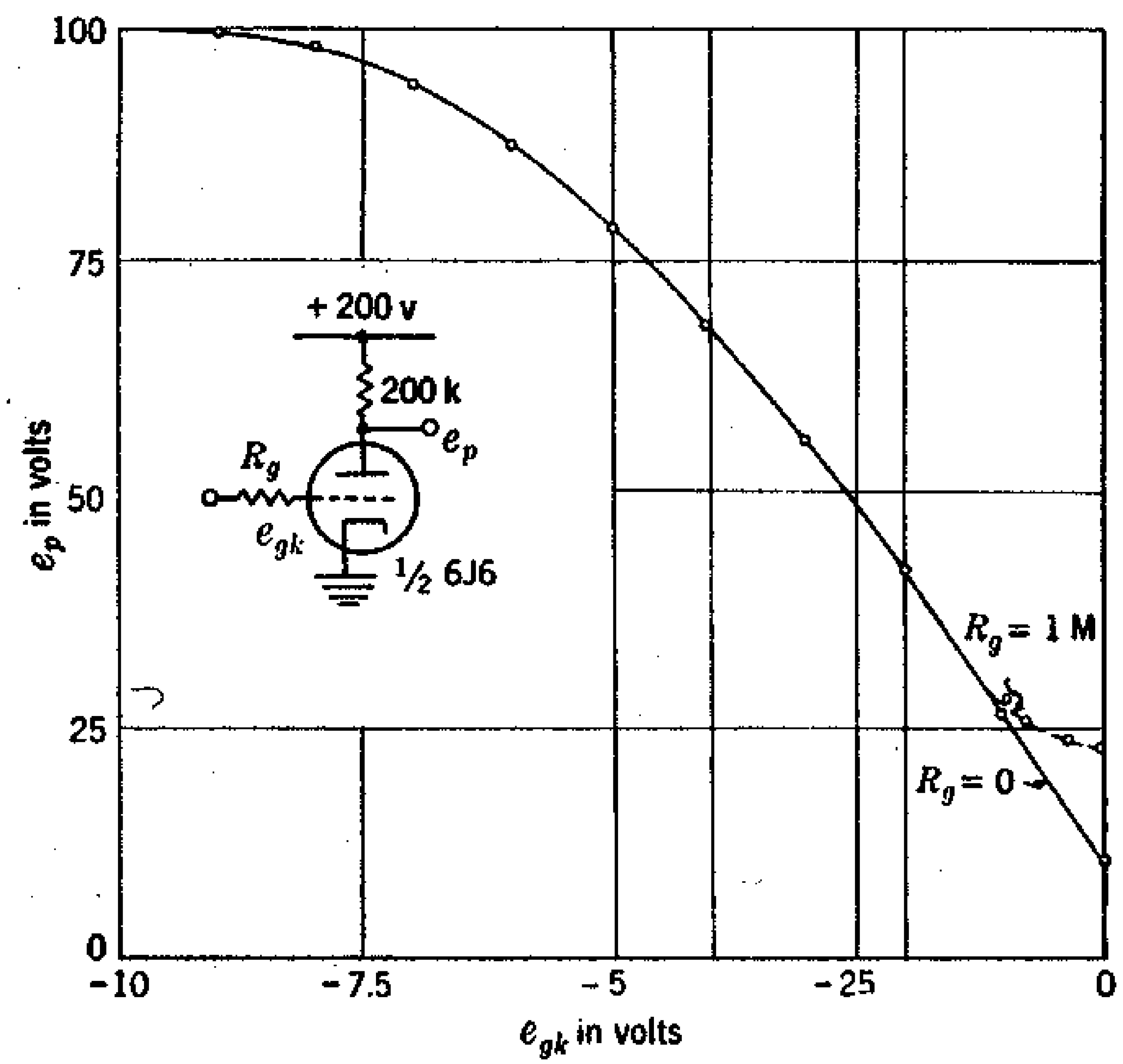

FIG. 11·11.—Typical triode amplifier characteristics.

This relation would indicate that a very high resistance should be employed for R_p if a voltage amplifier is to realize the maximum gain. Actually, r_p is approximately inversely proportional to plate current for small values of i_p, and μ drops slightly as plate current is decreased. Therefore, for a given E_{pp}, the gain ceases to increase as R_p is increased beyond a finite multiple of r_p. For instance, in a 6J6 triode, by measuring the increment of plate voltage obtained between $e_g = -2.5$ and -5

volts, it can be shown that if $E_{pp} = 200$ volts, increasing R_p above about 75,000 ohms gives no increase in the gain. On the other hand, an increase of E_{pp} gives an appreciable increase of gain.

Except where tube rating is exceeded, the factors limiting the range of voltage or current output are the plate-supply voltage (or zero-load current) and the point where grid current becomes excessive as determined by the input impedance. The latter, except for a low input impedance, generally occurs at a grid voltage of from -1 to -2 volts.

The linearity of a simple triode amplifier, considered over the entire output range, is rather poor because the plate resistance increases and the amplification factor decreases toward the low-current end of the load line. Figure 11·11 shows a typical triode amplifier characteristic and is derived by drawing a suitable load line on a family of i_p vs. e_p curves for a type 6J6 triode. The dashed curve results from grid current flowing in the input resistor.

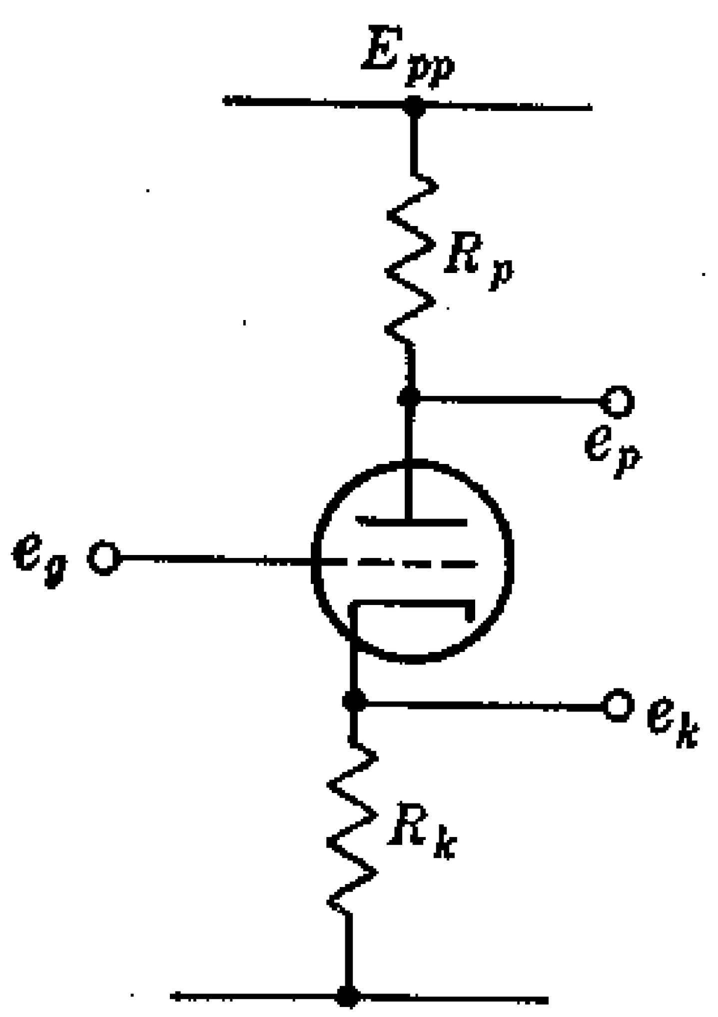

FIG. 11·12.—Triode with cathode and plate resistors.

However, linearity over a given range can be greatly improved by raising the plate-supply voltage to a value considerably above the upper limit required to obtain the desired output voltage. For many triodes, μ is constant with respect to plate voltage if plate current can be maintained constant. Thus, for good linearity as well as for maximum gain, the ideal voltage amplifier would have infinite R_p and infinite E_{pp}. Some constant-current device, such as will be described in connection with pentode amplifiers, may be employed in the plate circuit to approach this ideal. The linearity of output under conditions of constant plate current is illustrated for a 6SL7 in Fig. 11·20.

Triode Amplifier with Cathode Resistor.—When a resistance is inserted in series with the cathode, as in Fig. 11·12, the load line, if it is derived from $R_p + R_k$ instead of R_p, still describes the performance of the amplifier. But since e_{gk} is no longer the same as e_g but is equal to $e_g - i_pR_k$ (assuming no external cathode load), it is not possible to obtain an operating point directly from the load line. For a given e_g, one estimates i_p, computes the corresponding i_pR_k and e_{gk}, refers to the diagram (Fig. 11·10) to see how far the resulting i_p is from the estimate, and repeats the process with some intermediate estimate of i_p. When R_k is large (comparable, for instance, with r_p or greater) and when e_g is positive and sufficiently large so that e_{gk} will be small in comparison, the first estimate of i_p is assumed to be equal to or a little larger than e_g/R_k. This usually proves to be an accurate assumption. When an operating point on the load line has been determined, its interpretation is as follows: The difference between e_{pk} and E_{pp} is apportioned between R_p and R_k in proportion with their relative values.

When R_k is small, it may be convenient to redraw the plate characteristics to include the effect of this resistor and to consider the new graph as if an equivalent tube (having a different set of characteristics from the original) were being used. The procedure is illustrated in Fig. 11·13, where the plate characteristics of a 7F8 triode are modified by a 1000-ohm cathode resistor. The points are plotted by assuming values of e_g and e_{gk} and computing i_p from $i_p = (e_g - e_{gk})/R_k$. The resulting curves (shown solid) are the plate characteristics of the equivalent tube, upon which load lines may be drawn in the usual manner with no further consideration of R_k.

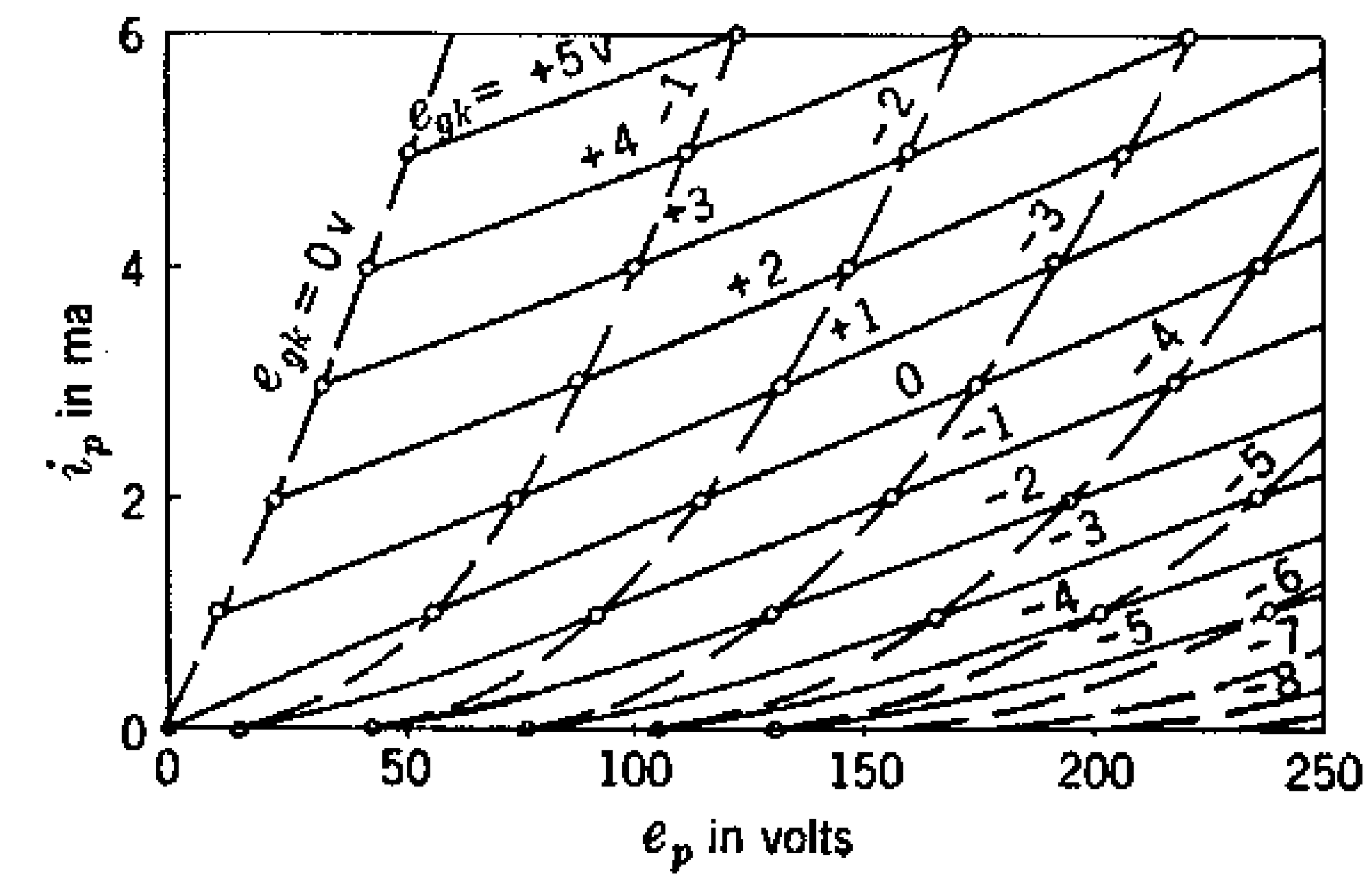

FIG. 11·13.—Modification of triode plate characteristics by a cathode resistor (7F8 triode, $R_k = 1000$ ohms).

It is helpful to analyze the circuit of Fig. 11·12 on the basis of the simple approximation of Eq. (1). Substituting $E_{pp} - R_pi_p - R_ki_p$ for e_{pk} and $e_g - R_ki_p$ for e_{gk}, the plate current is found to be

$$i_p = \frac{E_{pp} + \mu e_g}{r_p + R_p + (\mu + 1)R_k}. \tag{6}$$

A rather useful concept is obtained by comparing Eqs. (6) and (1). It is seen that when a triode has resistors in series with its cathode or plate or both, the triode with its resistors may be considered equivalent to a simple triode having the same μ but having a plate resistance equal to $r_p + R_p + (\mu + 1)R_k$. Figure 11·14 illustrates this equivalence. Figure 11·13 bears this out for the case of the cathode resistor: μ is unchanged in going from the dashed to the solid curves, but r_p is increased by $(\mu + 1)R_k$.

From Eq. (6) it follows that the plate and cathode voltages are (assuming negligible grid current)

$$e_p = \frac{\dfrac{r_p + (\mu + 1)R_k}{R_p} E_{pp} - \mu e_g}{1 + \dfrac{r_p + (\mu + 1)R_k}{R_p}}, \tag{7}$$

and

$$e_k = \frac{E_{pp} + \mu e_g}{\mu + 1 + \dfrac{r_p + R_p}{R_k}}. \tag{8}$$

Equation (7) could have been derived from Eq. (3) by considering R_k as part of an equivalent triode and consequently adding $(\mu + 1)R_k$ to r_p, as suggested in the foregoing paragraph.

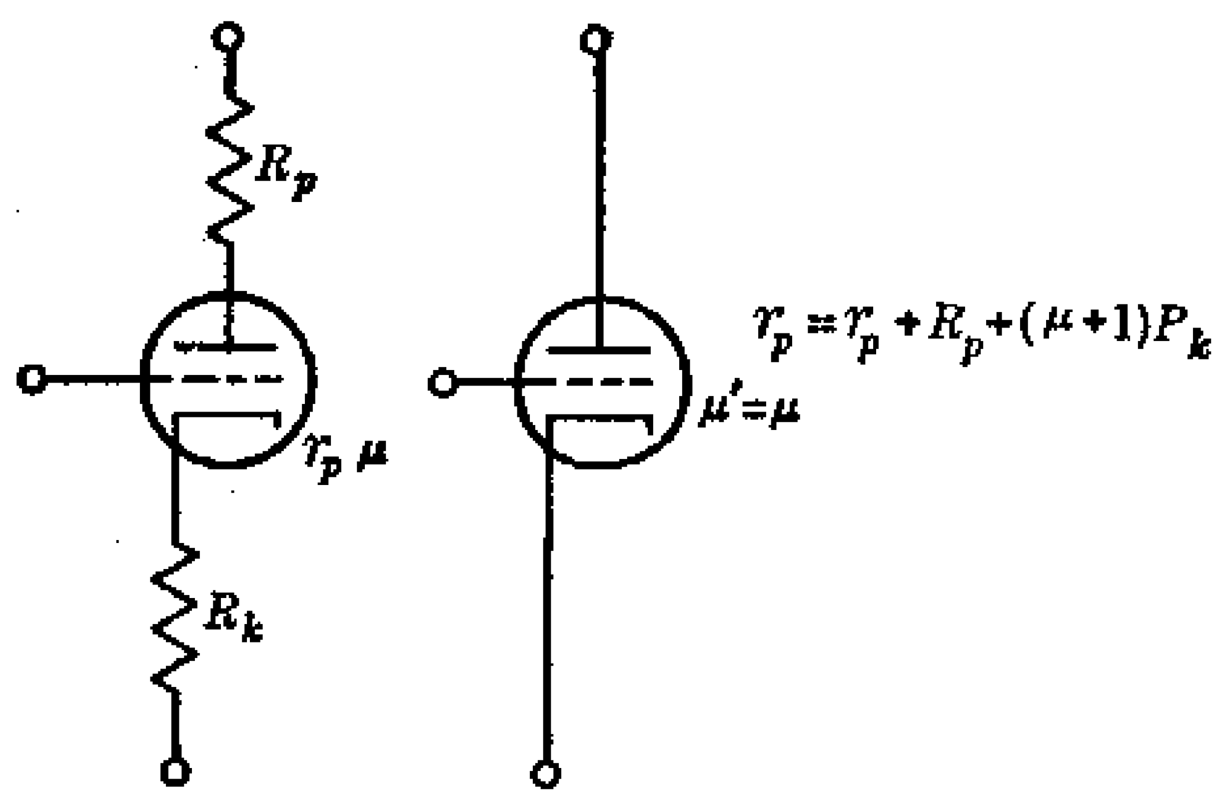

FIG. 11·14.—Triode equivalent of triode with plate and/or cathode resistor.

Equation (7) shows a gain from grid to plate of

$$\mathcal{G} = \frac{-\mu}{1 + \dfrac{r_p + (\mu + 1)R_k}{R_p}}. \tag{9}$$

The cathode resistor affords a negative feedback that, in addition to reducing the gain, reduces the variation in gain caused by variations of r_p and μ. This tends to keep the gain constant when tubes are changed and also to make the output-vs.-input curve more linear. The effect of variations of r_p and μ on the gain may be found by partial differentiation of Eq. (9). The resulting fractional change of gain is

$$\frac{\Delta\mathcal{G}}{\mathcal{G}} = \frac{r_p + R_p + R_k}{r_p + R_p + (\mu + 1)R_k}\,\frac{\Delta\mu}{\mu} - \frac{r_p}{r_p + R_p + (\mu + 1)R_k}\,\frac{\Delta r_p}{r_p} \tag{10}$$

Thus if R_k is such that $(\mu + 1)R_k$ is n times $(r_p + R_p)$ the gain is reduced by a factor of $n + 1$ [see Eq. (9)]. The fractional change in gain caused by a given variation of r_p is then reduced by the same factor as that caused by a variation of μ if μ is large compared with n.

If μ is large and μR_k is large compared with $r_p + R_p$, the gain is approximately R_p/R_k. This gain and the deviation from it are shown by Fig. 11·15*a*, which is one form of graphical representation of Eq. (9). Figure 11·15*b* is a different form for use where R_k is low.

The variational plate output impedance of a simple triode amplifier (Fig. 11·9) is equivalent to r_p and R_p in parallel. By applying the equivalent-triode theory developed above, it is found that the plate

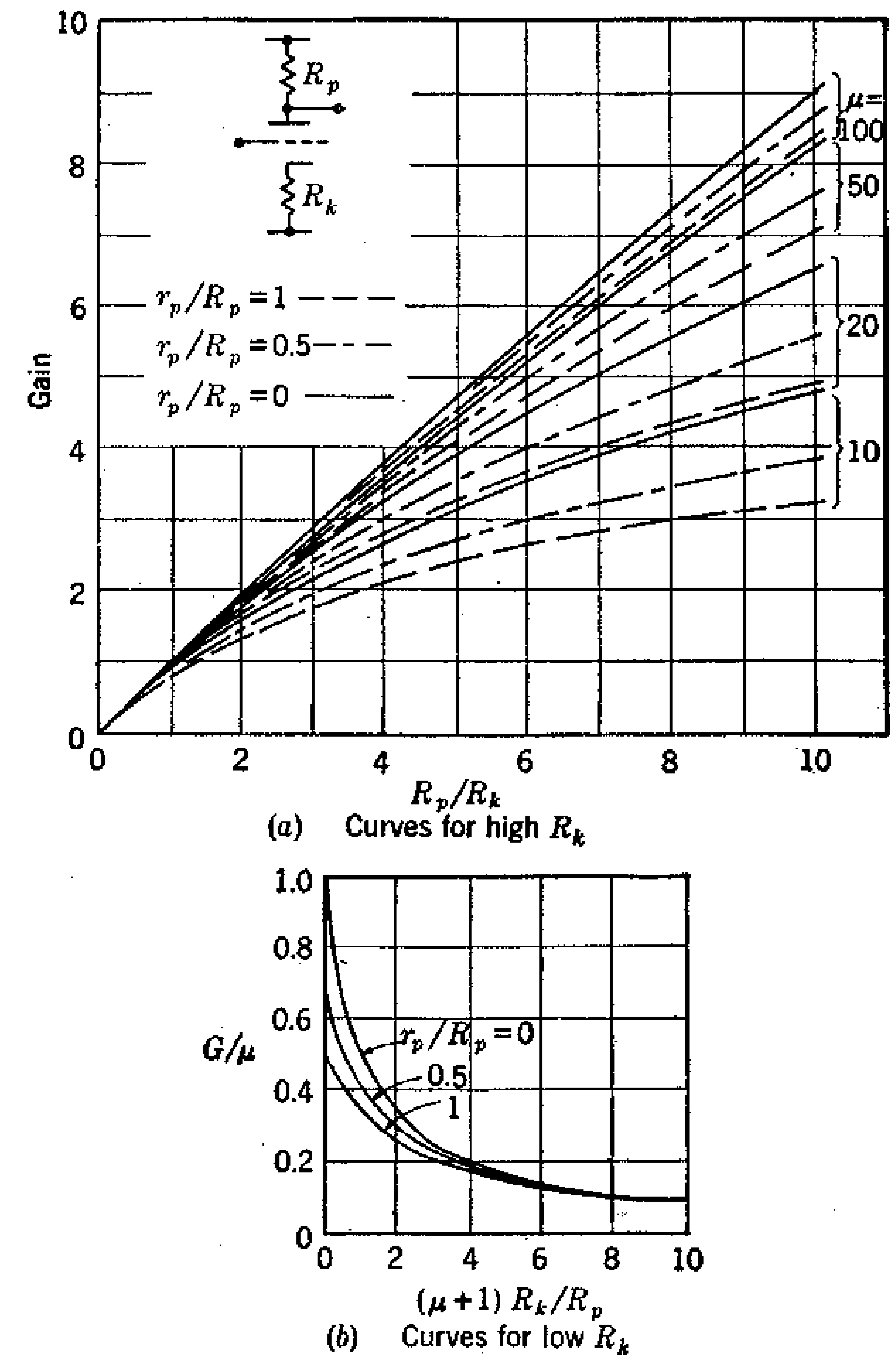

FIG. 11·15.—Gain of triode amplifier with cathode resistor.

output impedance for a triode amplifier with cathode resistor (Fig. 11·12) is

$$Z_p = \frac{r_p + (\mu + 1)R_k}{R_p + r_p + (\mu + 1)R_k} R_p. \tag{11}$$

To find the cathode output impedance, consider the change in the current flowing into the cathode terminal in Fig. 11·9 if the cathode is ungrounded and e_k is varied by external means. This current is $-i_p$, which from Eq. (2) is

$$-i_p = \frac{E_{pp} - e_k + \mu(e_g - e_k)}{r_p + R_p};$$

therefore, if e_k is raised by an increment Δe_k without changing E_{pp} or e_g,

$$\frac{\Delta(-i_p)}{\Delta e_k} = \frac{+1}{r_p + R_p}, \tag{12}$$

which is thus the admittance of the cathode terminal. If this impedance is put in parallel with R_k, the cathode output impedance of a triode with cathode resistor (Fig. 11·12) is

$$Z_k = \frac{r_p + R_p}{\mu + 1 + \dfrac{r_p + R_p}{R_k}}. \tag{13}$$

A similar procedure may also be employed in the derivation of Eq. (11).

D-c Cathode Follower.—If the cathode is to be used as the output terminal, the circuit of Fig. 11·12 is a "cathode follower," although the usual practice is to omit R_p. In the latter case Eqs. (6), (8), and (13) become

$$i_p = \frac{E_{pp} + \mu e_g}{r_p + (\mu + 1)R_k}, \tag{14}$$

$$e_k = \frac{E_{pp} + \mu e_g}{\mu + 1 + \dfrac{r_p}{R_k}}. \tag{15}$$

$$Z_k = \frac{r_p}{\mu + 1 + \dfrac{r_p}{R_k}}. \tag{16}$$

Generally μ and R_k are large enough so that for practical purposes the following approximations can be used:

$$i_p \approx \frac{E_{pp}}{\mu R_k} + \frac{1}{\left(1 + \dfrac{r_p + R_k}{\mu R_k}\right)} \frac{e_g}{R_k}, \tag{17}$$

$$e_k \approx \frac{E_{pp}}{\mu} + \left(1 - \frac{r_p + R_k}{\mu R_k}\right) e_g, \tag{18}$$

$$Z_k \approx \frac{1}{g_m}. \tag{19}$$

The cathode follower is not a voltage amplifier but is a simple and excellent impedance changer or power amplifier, in which a high degree of stability and linearity is afforded by the inherent negative feedback. The variability of gain may be obtained directly from Eq. (10); and since the same current flows in both R_p and R_k, the fractional gain variation is the same measured across either resistor. Dropping R_p and rearranging Eq. (10) for clarity,

$$\frac{\Delta \mathcal{G}}{\mathcal{G}} = \frac{1 + \dfrac{R_k}{r_p}}{1 + (\mu + 1)\dfrac{R_k}{r_p}} \frac{\Delta\mu}{\mu} - \frac{1}{1 + (\mu + 1)\dfrac{R_k}{r_p}} \frac{\Delta r_p}{r_p}. \tag{20}$$

For example, if μ is 20 and R_k is twice the average value of r_p, and if μ increases by 20 per cent and r_p decreases by 50 per cent of its average value as the working range is traversed, the percentage change of gain is

$$\frac{\Delta \mathcal{G}}{\mathcal{G}} = \frac{1 + 2}{1 + 21 \times 2} \times 0.2 - \frac{1}{1 + 21 \times 2} \times (-0.5)$$
$$= 0.014 + 0.012 = 2.6 \quad \text{per cent},$$

which indicates a 2.6 per cent change in slope of the output-vs.-input curve, or a deviation from linearity of less than ± 0.2 per cent.[1] Equation (15) shows that a given variation of plate-supply voltage results in a variation in the output of a cathode follower of only $1/(\mu + 1 + r_p/R_k)$ times this change of voltage.

The output-vs.-input curve of a cathode follower may be obtained from a load line with a slope of $-1/R_k$ drawn on the plate characteristics (Fig. 11·10). (E_{pp} is, of course, measured from the fixed terminal of R_k in case this is not at ground potential.) The output voltage is simply $E_{pp} - e_{pk}$; and for any given output, the input is this value plus e_{gk} at this point on the load line. The load line reveals immediately, without the necessity of plotting the output curve, the useful range of the cathode follower (as limited by grid current and by low plate current) and the tube power dissipation.

The linearity of a cathode follower may be greatly increased for special applications where precision is paramount by the use of a "constant-current" device in place of R_k. As has been mentioned, the μ of many triodes at constant current is independent of plate voltage. Therefore, good linearity is obtainable if the load line can be made horizontal. The devices of Fig. 11·16 afford good approximations to constant current. The pentode circuit of Fig. 11·16*a* relies on the fact that the pentode plate current is fairly independent of plate voltage. Current is determined by the self-biasing resistor R and by the screen potential. The latter should be as low as possible, because if the pentode plate falls below the

[1] *Cf.* Fig. 11·52, Eq. (102).

level of the screen, the screen current will increase. This effect, in turn, will decrease the plate current by increasing the biasing current in R_k and by lowering the screen voltage by extra loading of the bleeder. If a suitable negative voltage source is available, it may be convenient to connect R_k and the grid to this source and to connect the screen to ground.

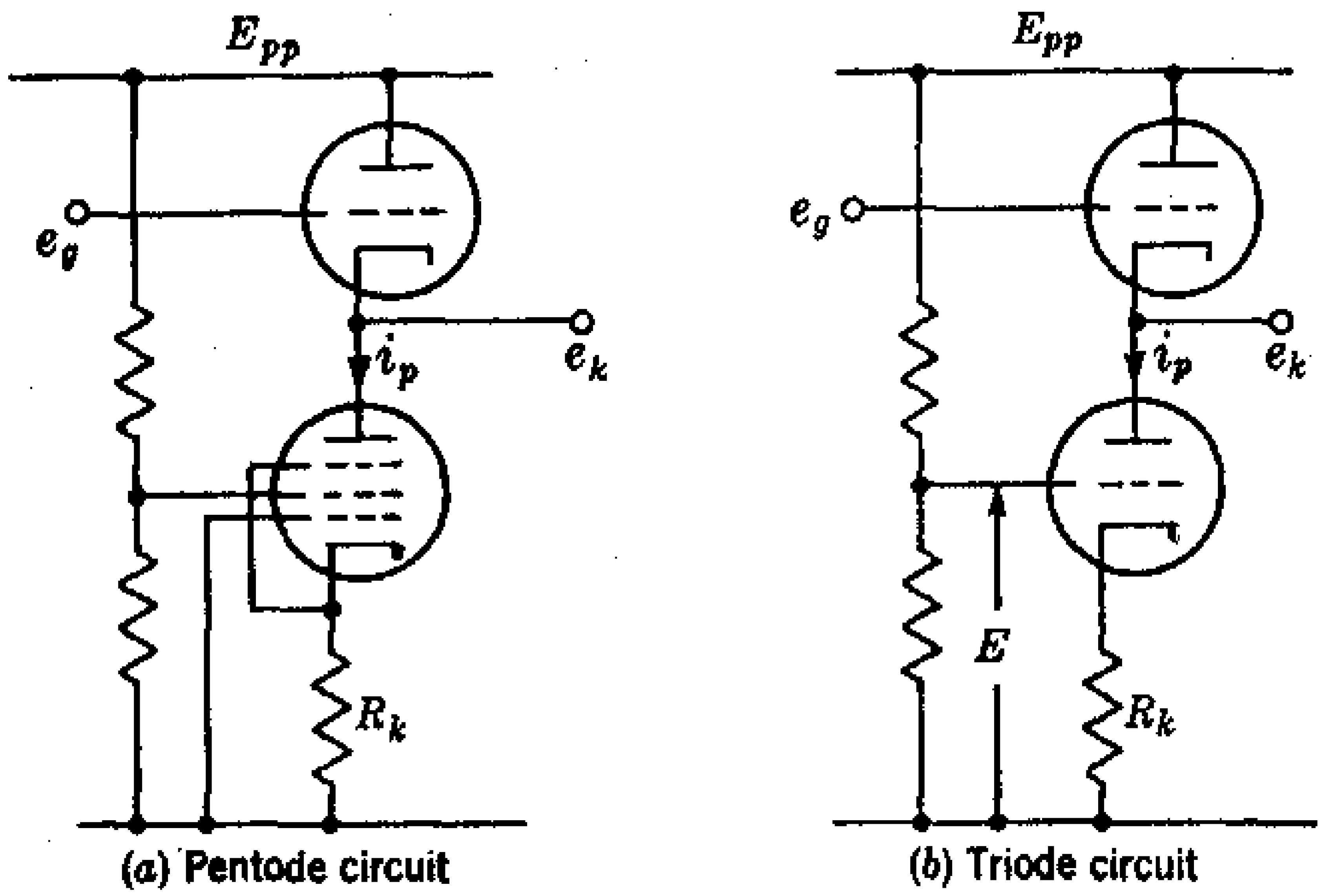

FIG. 11·16.—Constant-current devices replacing R_k of a cathode follower.

The triode circuit of Fig. 11·16b (if it is assumed that e_k remains sufficiently above E so that there is negligible grid current) may be analyzed by application of Eq. (6).

$$i_p = \frac{e_k + \mu E}{r_p + (\mu + 1)R_k}.$$

Thus the equivalent resistance is

$$\frac{\partial e_k}{\partial i_p} = r_p + (\mu + 1)R_k,$$

and it behaves within the limit of useful range as if its nether end were connected to a potential μE below the reference level. This triode constant-current device has much greater stability with respect to tube drift than has the pentode because a given drift of grid-to-cathode voltage causes less percentage change of the drop across R_k. Figure 11·20b shows the improvement in linearity of a cathode follower that may be achieved by using such a constant-current device.

If a constant-current device is employed, the cathode-follower output impedance [Eq. (16)] is simply the reciprocal of the transconductance.

11·8. Single-ended Pentode Amplifiers.—If both the cathode and screen grid of a pentode (or tetrode) are at fixed potentials, with no series resistances, the operation of the amplifier may be determined by means of the load line, as for a triode (see Fig. 11·17). The plate characteristics employed must apply to the existing screen potential.

A complete mathematical description of the approximate operating characteristics, such as was based on Eq. (1) for the triode, is not feasible here. However, since μ and r_p are derivatives at an operating point, the gain, which is also a derivative, is correctly expressed by Eq. (5).

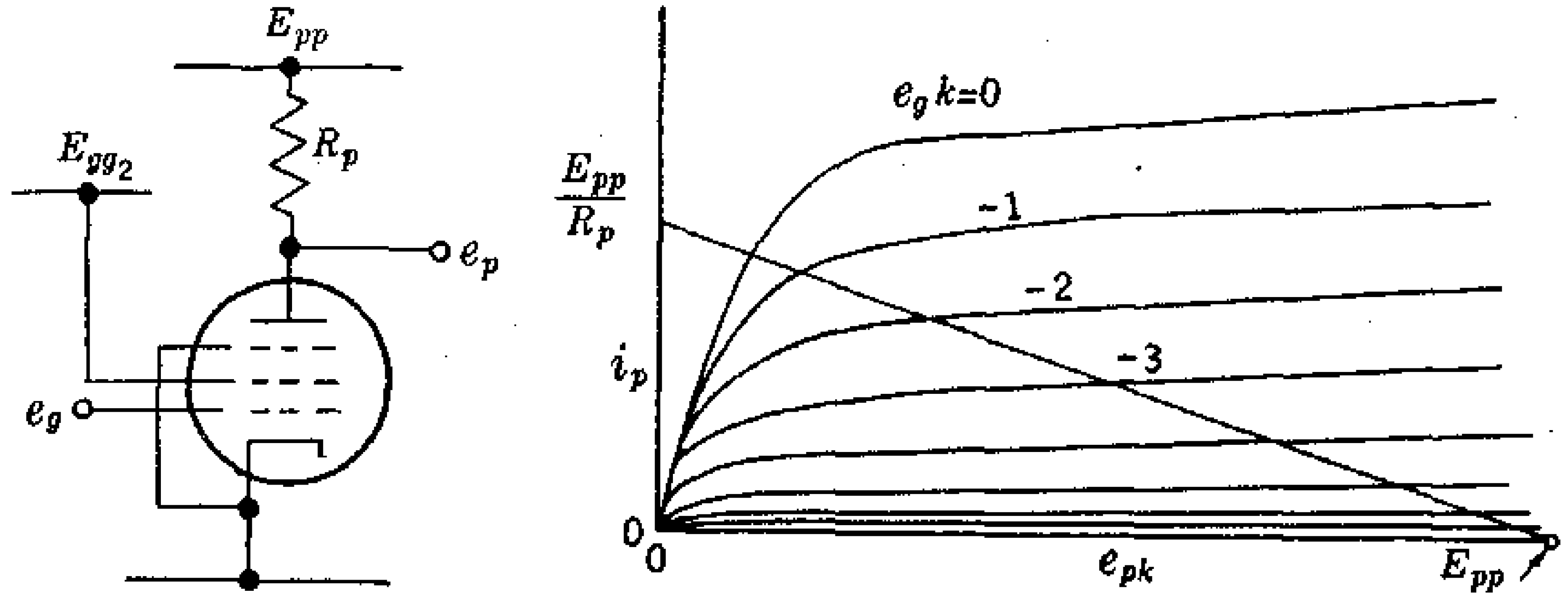

FIG. 11·17.—Pentode amplifier and load line on plate characteristics.

As μ is in practice large and rather indeterminate, the gain is better expressed for computation in the form

$$\frac{\partial e_p}{\partial e_g} = -\frac{g_m R_p}{1 + \frac{R_p}{r_p}}. \tag{21}$$

Since, in general, r_p is very great compared with R_p, a good approximation is

$$\frac{\partial e_p}{\partial e_g} \approx -g_m R_p. \tag{22}$$

It would seem that the voltage gain is proportional to the plate-load resistance except for very high values of R_p. Actually, in the case of any ordinary "sharp cutoff" pentode operating as a voltage amplifier, the transconductance is very nearly proportional to the plate current. Thus for a given plate-supply voltage and a certain operating range of plate voltage, the gain is independent of R_p over a very wide range of values. On the other hand, it is roughly proportional to $E_{pp} - e_p$.

If E_{pp}, together with R_p, can be raised without limit, a very high gain may be attained. The same effect may be achieved by the use of some kind of constant-current device in place of the load resistor. Such a device is shown in Fig. 11·18. This is the same as the constant-current element shown in Fig. 11·16b except that here a battery or its equivalent is needed. The operation of the *triode* (if there is assumed to be sufficient plate-to-cathode voltage to keep its grid from drawing current) may be determined by substitution of the "equivalent triode" with R_k included. This substitution changes the plate resistance to

$$r_p + (\mu + 1)R_k,$$

and in applying Eq. (1), $e_{pk} = E_{pp} - e_p$ and $e_{gk} = E$ (Fig. 11·14),

$$i_p = \frac{E_{pp} - e_p + \mu E}{r_p + (\mu + 1)R_k}. \tag{23}$$

Thus, except for an upper working limit somewhat below E_{pp}, this triode arrangement behaves like a resistor of value $r_p + (\mu + 1)R_k$ (or roughly, μR_k), whose upper terminal is attached to a potential μE above E_{pp}. This device is not often practical because it requires a battery or floating power supply.

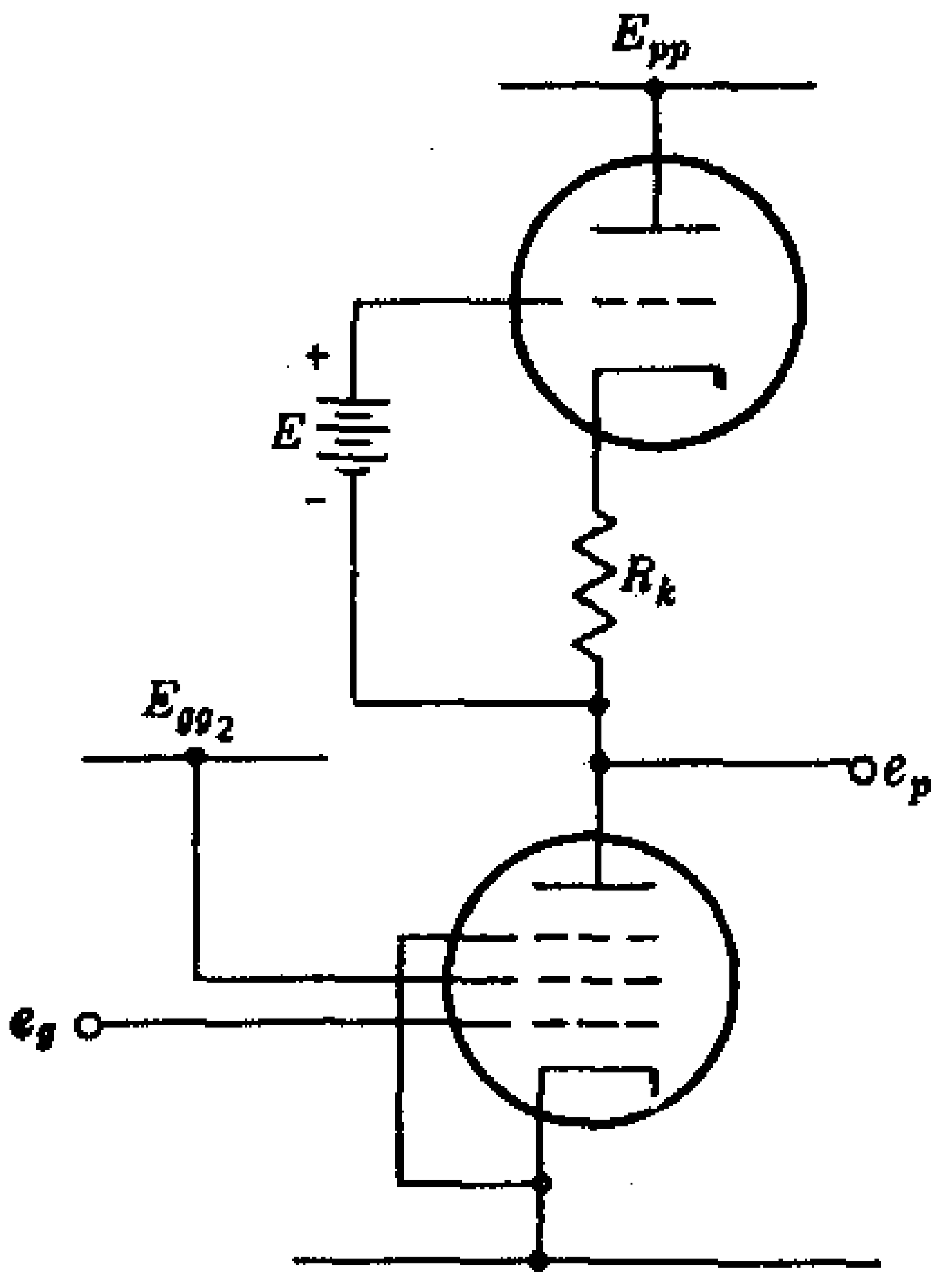

FIG. 11·18.—Constant-current device as a pentode load resistor.

In general, with a given plate-supply voltage and without the use of a special circuit such as the foregoing, the maximum voltage gain is obtained when the screen grid is operated at the lowest potential permissible from the standpoint of control-grid current. The g_m of ordinary sharp cutoff pentodes, at any given plate current, increases slightly with reduced screen potential.

If the screen-to-cathode voltage is subject to variation, the analysis of the operation of an amplifier is more involved than if it is fixed. For rough calculation, advantage may be taken of the facts that over the usual range of operation, both plate and screen current are fairly independent of plate voltage, and screen current is a substantially constant ratio of plate current (between $\frac{1}{4}$ and $\frac{1}{2}$). An equation similar to Eq. (1) is sometimes useful, although it is not so accurate or complete a description of the characteristics. The plate current is the same sort of function of screen voltage as it was of plate voltage in the case of the triode. Therefore, in applying Eq. (1) to a pentode, $\partial e_{g_2}/\partial i_p$ is used instead of $\partial e_p/\partial i_p$, and $\partial e_{g_2}/\partial e_g$ instead of $\partial e_p/\partial e_g$.

$$i_p = \frac{e_{g_2} + \dfrac{\partial e_{g_2}}{\partial e_g} e_{gk}}{\dfrac{\partial e_{g_2}}{\partial i_p}}. \tag{24}$$

Thus, if the screen grid has a series resistor, R_{g_2} (e.g., from a bleeder whose output resistance is appreciable), and if the assumption is made that i_{g_2} is a constant fraction n of i_p, then

$$i_p = \frac{E_{gg_2} - R_{g_2} n i_p + \dfrac{\partial e_{g_2}}{\partial e_g} e_g}{\dfrac{\partial e_{g_2}}{\partial i_p}},$$

from which

$$i_p = \frac{E_{gg_2} + \dfrac{\partial e_{g_2}}{\partial e_g} e_g}{\dfrac{\partial e_{g_2}}{\partial i_p} + nR_{g_2}}, \tag{25}$$

and

$$\begin{aligned} e_p &= E_{pp} - i_p R_p \\ &= E_{pp} - \frac{R_p E_{gg_2} + R_p \dfrac{\partial e_{g_2}}{\partial e_g} e_g}{\dfrac{\partial e_{g_2}}{\partial i_p} + nR_{g_2}}. \end{aligned} \tag{26}$$

The gain is

$$\frac{\partial e_p}{\partial e_g} = - \frac{R_p \dfrac{\partial e_{g_2}}{\partial e_g}}{\dfrac{\partial e_{g_2}}{\partial i_p} + nR_{g_2}} = - \frac{R_p g_m}{1 + nR_{g_2} \dfrac{\partial i_p}{\partial e_{g_2}}}. \tag{27}$$

If the pentode amplifier is further complicated by a cathode resistor, as in Fig. 11·19, the current flowing in this resistor must be considered to include screen current. But if the assumption is made that $i_{g_2} = ni_p$, where n is a constant, then i_k is $(n + 1)i_p$. Equation (24) yields

$$i_p = \frac{E_{gg_2} - nR_{g_2} i_p - (n + 1)i_p R_k + \dfrac{\partial e_{g_2}}{\partial e_g}[e_g - (n + 1)i_p R_k]}{\dfrac{\partial e_{g_2}}{\partial i_p}}$$

from which

$$i_p = \frac{E_{gg_2} + \dfrac{\partial e_{g_2}}{\partial e_g} e_g}{\dfrac{\partial e_{g_2}}{\partial i_p} + nR_{g_2} + \left(\dfrac{\partial e_{g_2}}{\partial e_g} + 1\right)(n + 1)R_k}. \tag{28}$$

The voltage gain is

$$\frac{\partial e_p}{\partial e_g} = - \frac{R_p g_m}{1 + nR_{g_2} \dfrac{\partial i_p}{\partial e_{g_2}} + (1 + n)R_k \left(g_m + \dfrac{\partial i_p}{\partial e_{g_2}}\right)}. \tag{29}$$

Pentode Cathode Follower.—In a-c applications a cathode follower can make capital of the high gain afforded by a pentode, for the purposes of obtaining good linearity and low attenuation, by the use of a condenser to maintain constant screen-to-cathode potential. In d-c circuits, however, where (unless a battery or floating power supply is used to set the screen at a fixed potential with respect to the cathode) the screen-to-cathode potential varies along with the plate-to-cathode potential, the pentode operates as a triode. In computing the cathode-follower performance, the triode-connected plate characteristics apply. There are cases, however, where some advantage may be taken of the extra electrode. For example, if a stable output at high power level is desired, and if a regulated, but low-power, voltage source is available, the screen can be tied to this regulated source, and the plate, which furnishes most of the power, can be connected to an unregulated source. The output will then hardly be affected by plate-voltage fluctuation.

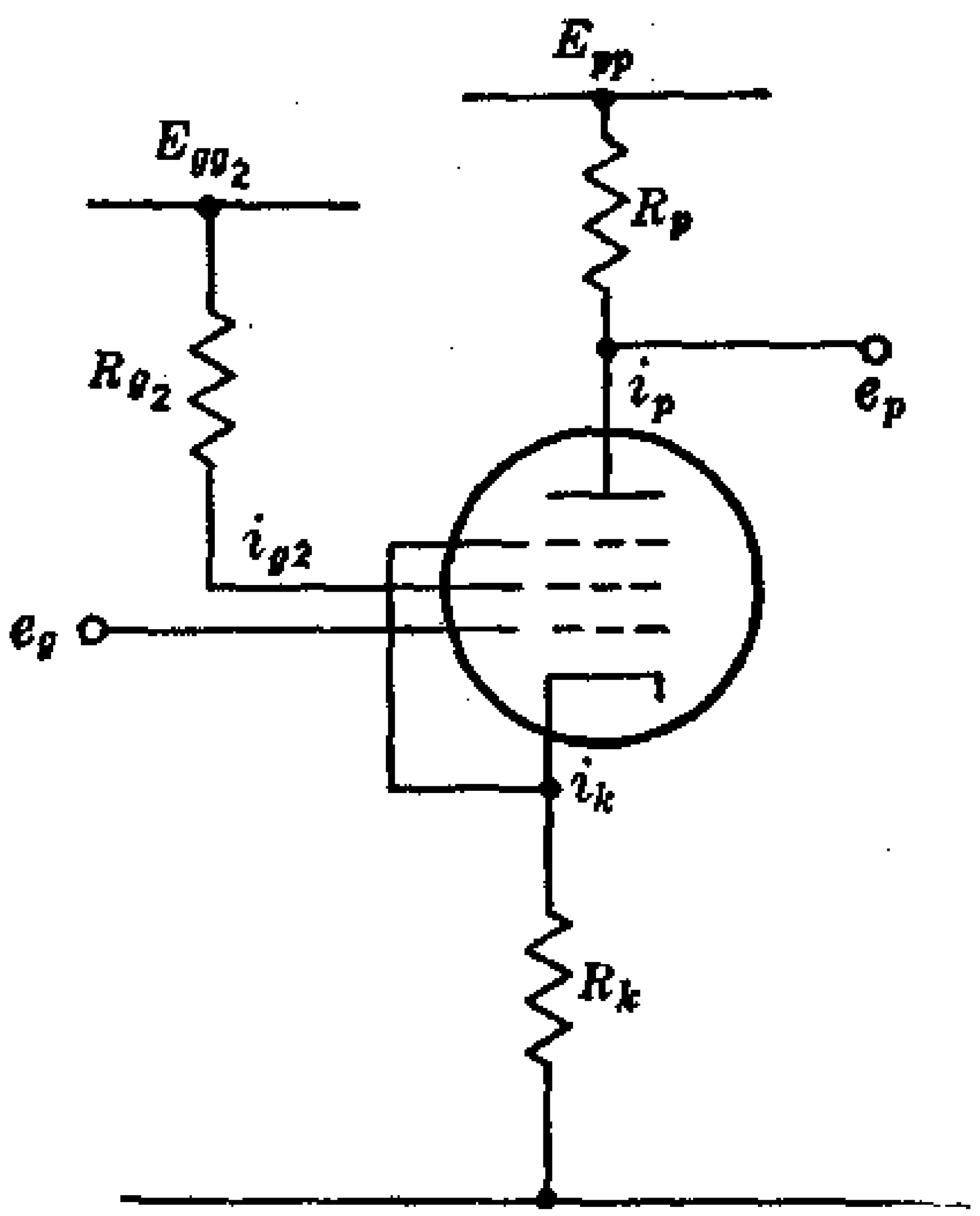

FIG. 11·19.—Pentode amplifier with cathode and screen resistors.

11·9. Cascode and Other Series Amplifiers. *Constant-current Plate Load.*—It is evident that the gain of an amplifier is a maximum when both R_p and E_{pp} are infinite and i_p is finite. In this case, also, the output signal is most linear, since the gain, being simply μ, is independent of r_p. For most triodes μ, at any given plate current, tends to be constant over a wide range of plate voltage. The effect of infinite or very great E_{pp} and R_p may be achieved by the use of a "constant-current" device between the plate and the $B+$ voltage supply.

One such device is a diode with such a low filament voltage that its current is limited by cathode emission and is thus independent of its plate-to-cathode potential. A diode with an indirectly heated, oxide-coated cathode does not have really constant current at low temperature because the emission depends partly on plate voltage,[1] but the partial effect is of some value. Figure 11·20*a* shows the increase in both gain and linearity of a triode amplifier using a diode instead of a fixed plate resistor. A diode having a pure tungsten cathode would be more effective but less convenient. In either instance, the device is subject to considerable drift if the filament supply voltage varies. Figure 11·20*b* shows analogous data for a cathode follower using the constant-current triode circuit shown in Fig. 11·16*b*.

A high-vacuum *photoelectric* cell has a current independent of voltage at any given value of illumination and thus may be used as a constant-current device. The current is at most a few microamperes, and the illumination must be constant; however, the scheme may be of value in certain applications.

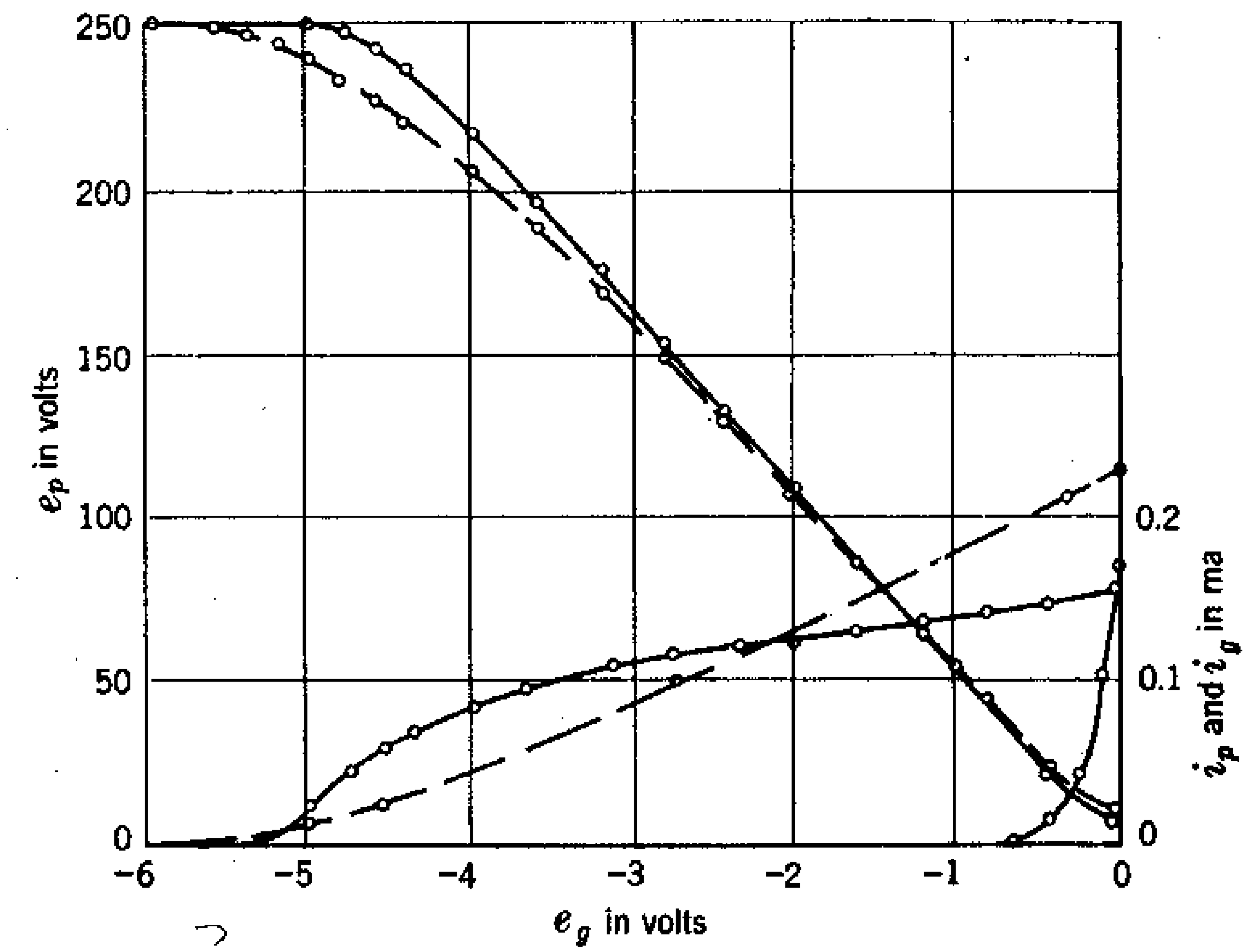

FIG. 11·20*a*.—Output characteristics of a triode amplifier with and without a "constant-current" diode as plate load (6SL7 triode, E_{pp} = 250 volts). Solid curves: plate load is a 6H6 diode, E_f = 2.4 volts; dashed curves: plate load is a 1.1-megohm resistor.

The constant-current arrangement of Fig. 11·16*b*, as used in a cathode-follower circuit, may be adapted as an amplifier plate load as shown in Fig. 11·21. In this adaptation, since the cathode end rather than the plate end of the device varies in potential, a battery or floating supply is needed to determine the grid potential, although no appreciable current is required from this source. A small constant-voltage glow tube, with its current furnished by a resistor from the positive supply, may be used as the voltage standard in place of the battery, but this current is variable and therefore must be small compared with plate current.

The device of Fig. 11·21 is a cathode follower in which the current is approximately E/R. The greater E and μ_2 are the more nearly constant will be the current. From Eq. (6), the current is

$$i_p = \frac{E_{pp} - e_p + \mu_2 E}{r_{p2} + (\mu_2 + 1)R}, \qquad (30)$$

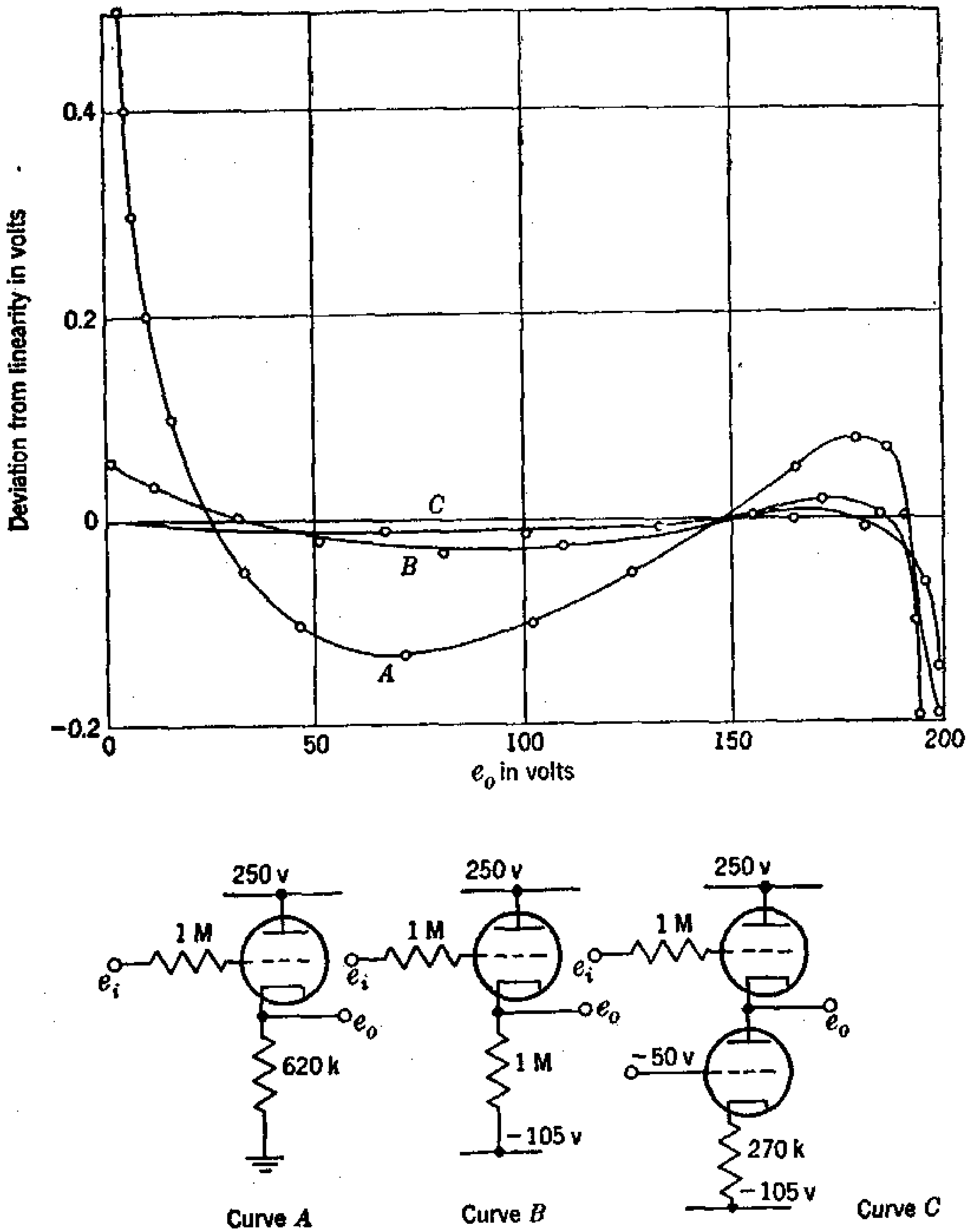

FIG. 11·20b.—Cathode-follower linearity.

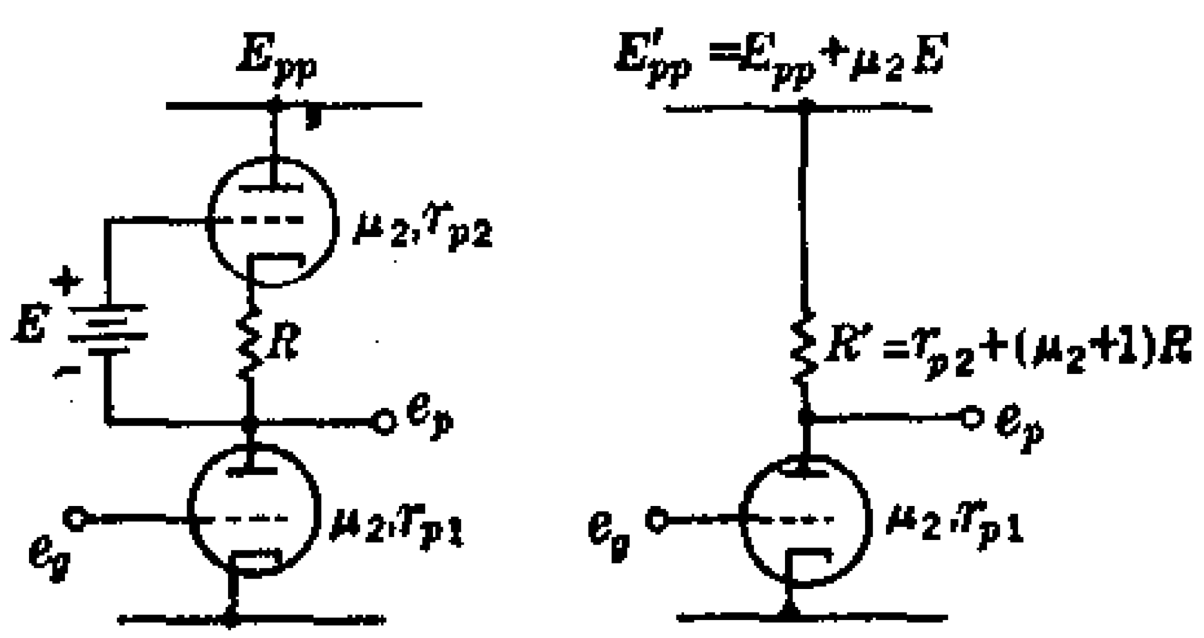

FIG. 11·21.—Triode amplifier with "constant-current" plate load. (*a*) Complete circuit; (*b*) equivalent circuit.

and its rate of change with respect to amplifier plate voltage is

$$\frac{di_p}{de_p} = -\frac{1}{r_{p2} + (\mu_2 + 1)R}. \tag{31}$$

The device becomes inoperative before e_p reaches E_{pp}, but it is seen from Eq. (30) that over its operating range, the current behaves as if it would be zero when $e_p = E_{pp} + \mu E$. Thus, except for the limited operating range, the device is equivalent to a plain resistance of value

$$R' = r_{p2} + (\mu_2 + 1)R \tag{32}$$

returned to a $B+$ voltage of

$$E'_{pp} = E_{pp} + \mu_2 E,$$

as shown in Fig. 11·21*b*.

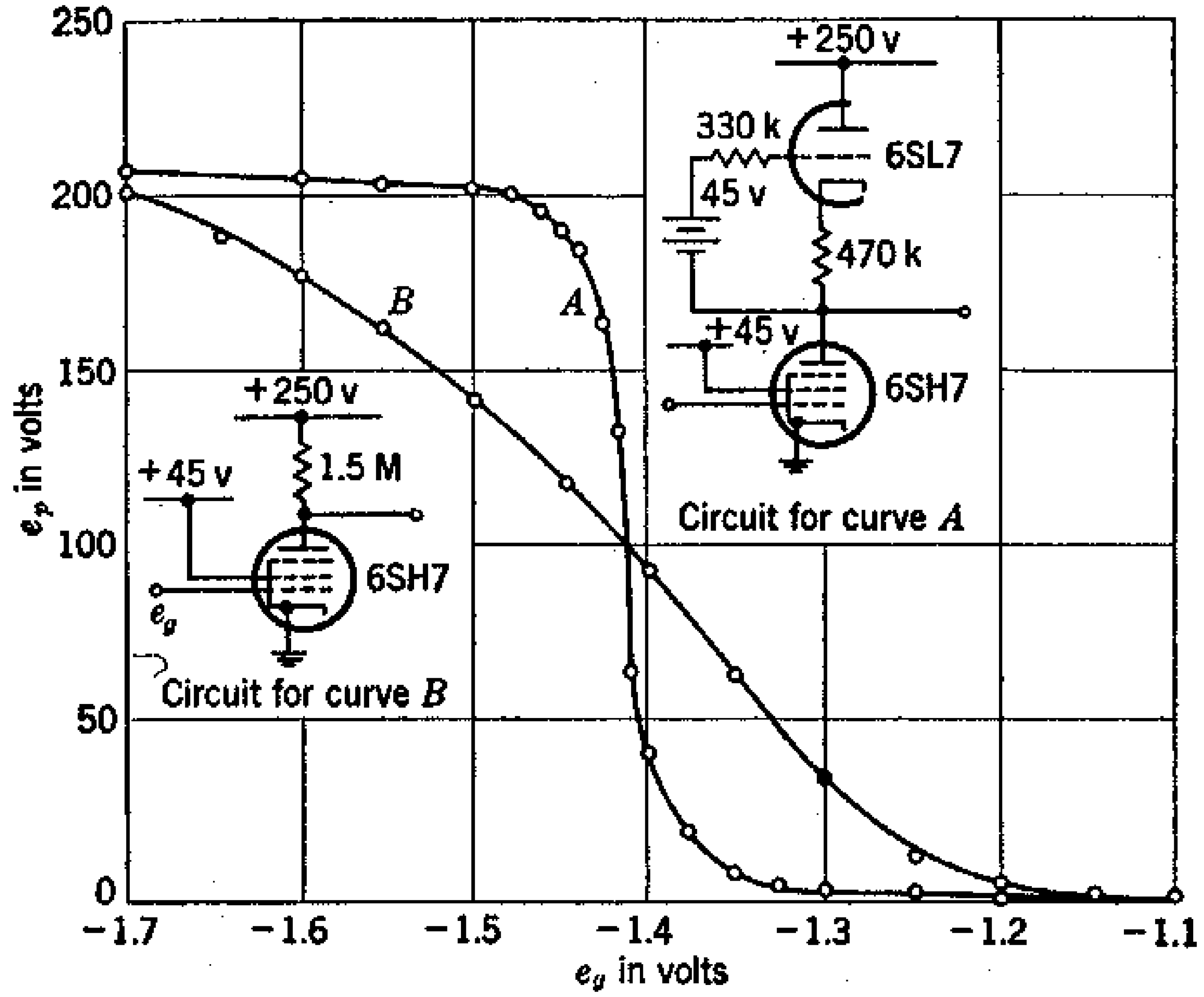

FIG. 11·22.—Output characteristics of a pentode amplifier with and without a constant-current plate load.

The constant-current circuit of Fig. 11·21 is especially effective in increasing the gain of a pentode because the gain with constant current is μ, which, for a pentode at low current, is very high. Figure 11·22 shows the output curve of a 6SH7 pentode with a 6SL7 constant-current plate load, where $E = 45$ volts. The gain is about 7000, whereas with a simple resistor and the same $B+$ voltage it is only 600.

The output impedance of a pentode amplifier with a constant-current load is very high; and if it is followed by a resistance divider, the gain may be seriously decreased. In effect, the divider resistance parallels the equivalent resistance of the constant-current circuit. If the voltage level must be lowered before being applied to the grid of the next stage, a cathode follower is advisable for the purpose.

If the battery is omitted in Fig. 11·22, the circuit does not, in any sense, comprise a constant-current device. It behaves like a simple resistor, of value $r_{p2} + (\mu_2 + 1)R$, attached to the E_{pp} source. Therefore, no increase of gain obtains. If the two triodes are similar, however, and if the amplifier triode also has a cathode resistor of value R, the arrangement is useful in that the output voltage is more linear than that of a simple amplifier and the effect of cathode temperature variation is canceled (see Sec. 11·11).

The pentode constant-current circuit of Fig. 11·16*a* may be adapted as a plate load for an amplifier by the use of a battery or its equivalent for the screen supply. This type of supply has the disadvantage that appreciable current is drawn therefrom. A simple divider for the screen supply attached between the cathode and E_{pp} is ineffective, because in this case i_p will tend toward zero as e_p approaches E_{pp}, as is the case with a plain resistance plate load.

Cascode Amplifier.—The preceding constant-current circuits have the aim of making the gain equal to μ by eliminating the $1/g_mR_p$ term from the gain equation

$$\mathcal{G} = \frac{-1}{\dfrac{1}{\mu} + \dfrac{1}{g_mR_p}}. \tag{33}$$

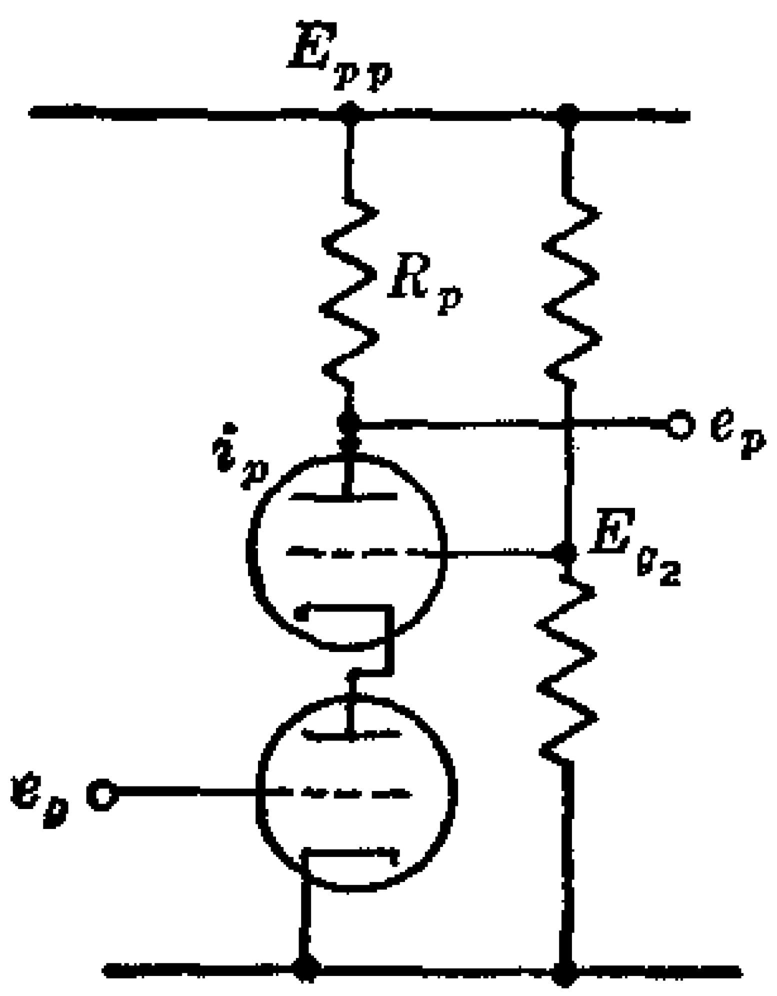

FIG. 11·23.—Cascode amplifier.

Thus they are most effective for a pentode amplifier where g_mR_p normally is much less than μ. In a triode amplifier, on the other hand, μ is much smaller, and the first term in the denominator of Eq. (33) is usually the more important. It is the result of variation of the plate voltage, which has very little effect in a pentode because of the isolating effect of the screen on the plate. A cascode amplifier tends to hold the triode plate potential fixed but still permits its current to flow in a load resistor. Thus the behavior is similar to that of a pentode, with the advantage that no screen current is required.

A circuit arrangement is shown in Fig. 11·23. The upper tube has a fixed grid, so that the excursions of its cathode are limited. If e_{p1} were really held constant, the current gain of the lower triode would be g_m and the voltage gain from e_g to e_p would be $-g_mR_p$, as in a pentode.

Actually the lower triode has as its plate load the cathode input impedance of the upper tube, which is, from Eq. (12), $(r_p + R_p)/(\mu + 1)$. The current gain is thus

$$\frac{\Delta i_p}{\Delta e_g} = \frac{\mu}{r_p + \dfrac{r_p + R_p}{\mu + 1}}, \tag{34}$$

and the voltage gain is

$$\mathcal{G} = -\frac{\mu R_p}{r_p + \dfrac{r_p + R_p}{\mu + 1}} = -\frac{1}{\dfrac{1}{g_mR_p} + \dfrac{r_p + R_p}{R_p}\,\dfrac{1}{\mu(\mu + 1)}}. \tag{35}$$

Thus, if R_p is large in comparison with r_p, μ^2 takes the place of the μ in Eq. (33).

The value of E_{g2} should be sufficiently high to keep the input grid from drawing appreciable grid current, and the operating range of e_p should be far enough above E_{g2} so that the current in the upper grid will not become comparable with i_p.

11·10. Differential Amplifiers. *Cathode as Input Terminal to an Amplifier.*—The cathode or grid of a tube or both may be employed as input terminals. From Eq. (2), the plate current in Fig. 11·24 is

$$i_p = \frac{E_{pp} + \mu e_g - (\mu + 1)e_k}{r_p + R_p}. \tag{36}$$

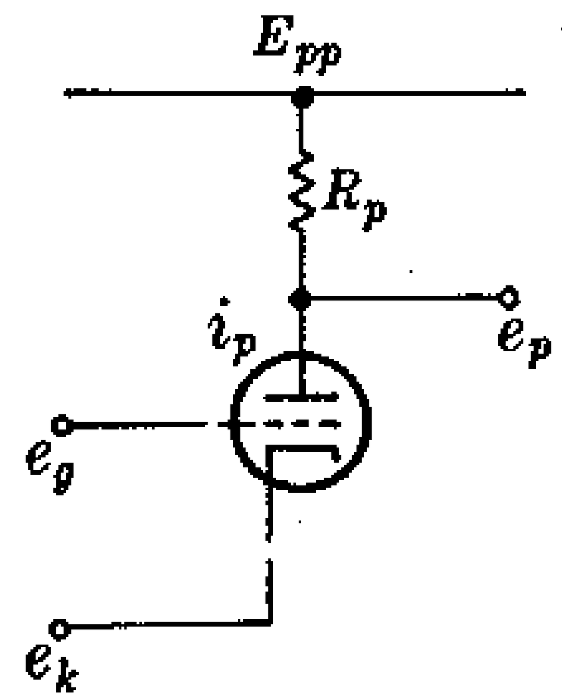

FIG. 11·24.—Triode with two input terminals.

The gain of the amplifier with respect to the cathode input terminal is $-(\mu + 1)/\mu$ times that with respect to the grid input terminal. If both input terminals are used, the output voltage is much more sensitive to changes in their difference than to changes in their common level. If their difference is held constant, e_p will change only $R_p/(r_p + R_p)$ as much as any common change of e_k and e_g, whereas e_p responds by μ times this much to any change of their difference.

The input impedance to the cathode is low and is equal to

$$Z_k = \frac{r_p + R_p}{\mu + 1}. \tag{37}$$

Any resistance in the cathode circuit R_k' has the same effect as adding $(\mu + 1)R_k'$ to the plate resistance. Thus, Eq. (36) becomes

$$i_p = \frac{E_{pp} + \mu e_g - (\mu + 1)e_k'}{r_p + R_p + (\mu + 1)R_k'}, \tag{38}$$

where e_k' is the cathode input voltage outside of R_k'. The output voltage is still μ times as responsive to input voltage difference as to mean input voltage level.

In spite of the low impedance of the cathode, it is often convenient to employ it rather than the grid as the input terminal, because the amplifier used in this way gives an output voltage of the same sense as that of the input voltage.

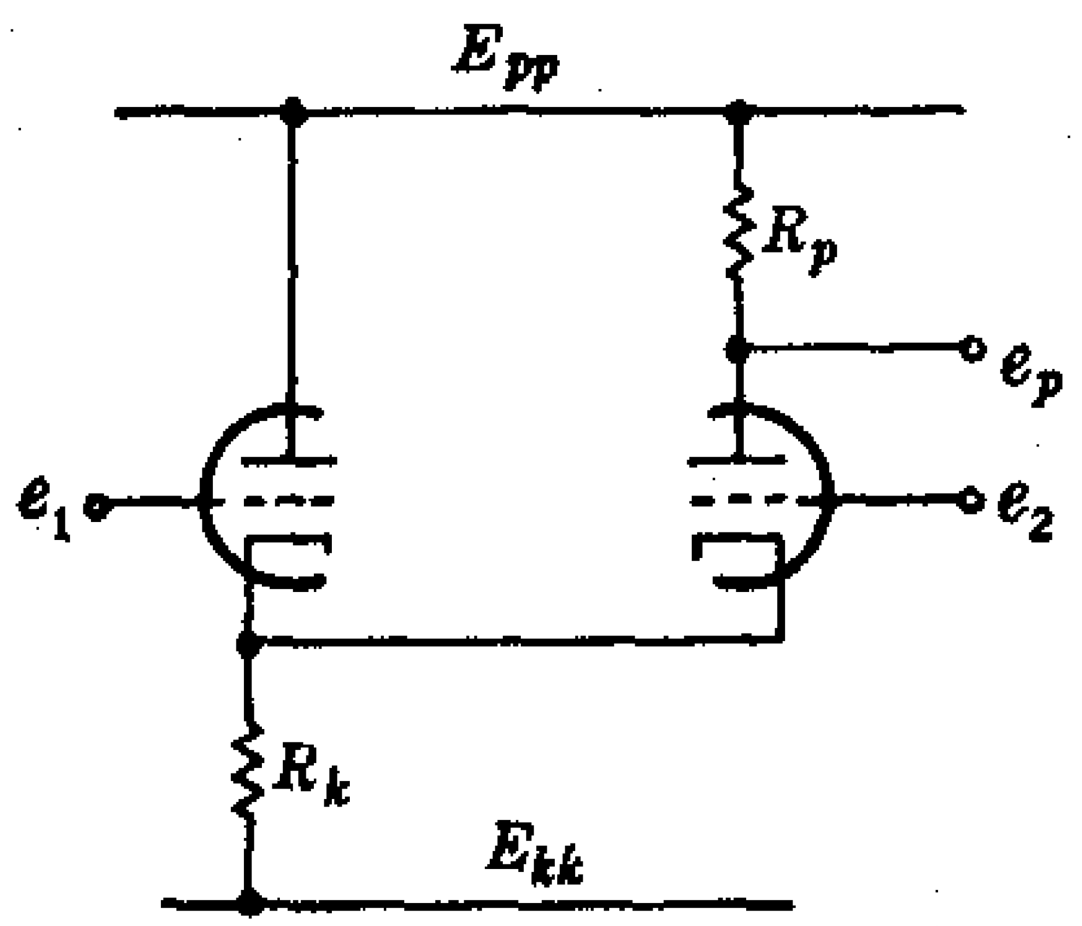

FIG. 11·25.—Cathode follower as an input means.

Cathode Input Voltage Applied through a Cathode Follower.—A cathode follower affords a convenient means of obtaining the necessary low impedance for the cathode input terminal. In addition to high input impedance and amplification without inversion, the arrangement illustrated in Fig. 11·25 provides a balancing of the effect of heater-voltage variation and a convenient high-impedance terminal (e_2) for zero adjustment or feedback or both.

The input voltage is attenuated according to the voltage gain of the cathode follower, which is $\mu/(\mu + 1 + r_p/R_k)$. Also, the output impedance of the cathode follower, $r_p/(\mu + 1 + r_p/R_k)$, acts as a resistance in series with the cathode of the amplifier. Thus, the gain from e_1 to e_p is

$$\mathcal{G}_1 = \left(\frac{\mu_1}{\mu_1 + 1 + \dfrac{r_{p1}}{R_k}}\right)\left(\frac{(\mu_2 + 1)R_p}{r_{p2} + R_p + \dfrac{(\mu_2 + 1)r_{p1}}{\mu_1 + 1 + \dfrac{r_{p1}}{R_k}}}\right), \tag{39}$$

in which the subscript numbers 1 and 2 refer to the cathode follower and amplifier triode, respectively. If these are the same, the gain is

$$\mathcal{G}_1 = \frac{\mu}{1 + \dfrac{2r_p}{R_p} + \dfrac{r_p(r_p + R_p)}{R_pR_k(\mu + 1)}}. \tag{40}$$

Cathode Follower Used as a Cathode Return.—The circuit of Fig. 11·25 may even be used when voltage inversion is acceptable; in this case the input voltage may be applied to the amplifier grid. Possible reasons for using the circuit in this way are (1) The input voltage may be at some level far from ground, and a bleeder of sufficiently low resistance for the cathode return may be undesirable; (2) the cathode follower grid may be used for negative or positive feedback; (3) the cathode follower may be used for a zero adjustment for the input level; (4) the effect of heater-voltage variation on zero shift is largely canceled; and (5) the output characteristic may be more nearly linear because of a mutual cancellation of the variation of plate resistance. The gain from e_2 to e_p is computed from Eq. (9), where the cathode resistor is the cathode-follower output impedance.

$$\mathcal{G}_2 = \frac{-\mu_2 R_p}{r_{p2} + R_p + \dfrac{(\mu_2 + 1)r_{p1}}{\mu_1 + 1 + \dfrac{r_{p1}}{R_k}}}. \tag{41}$$

If the tubes have the same characteristics,

$$\mathcal{G}_2 = -\frac{\mu + \dfrac{\mu}{\mu + 1}\dfrac{r_p}{R_k}}{1 + \dfrac{2r_p}{R_p} + \dfrac{r_p(r_p + R_p)}{R_pR_k(\mu + 1)}}. \tag{42}$$

If the μ's are equal and R_k is large, the last term in the denominator of Eq. (41) is approximately r_{p1}. The sum of the two plate currents, which is equal to the current in R_k, is substantially constant, so that a decrease in the amplifier current causes an almost equal increase in the plate current of the cathode follower. If the two plate currents are of the same order of magnitude, the variations of plate resistance caused by variation of plate current will be about equal and opposite. The denominator of Eq. (41) will thus be fairly constant over a considerable output range, and an increase in linearity will result.

Differential Input.—The circuit of Fig. 11·25 is more suitable than that of Fig. 11·24 for the amplification of the difference between two voltages, because both input terminals have a high impedance, the cathodes mutually offset the drift due to heater-voltage variation, and the output voltage may be less affected by the "common-mode variation" (common change of level) of the two input voltages.

This independence of the common mode is a feature of importance when the amplifier is to be used as a comparator,[1] in which the output voltage ideally should be a function only of the difference between the input voltages and not of their average. The degree of rejection of the common mode may be expressed by the ratio of the gain of the amplifier with respect to the average of the input voltages to that with respect to their difference. If the two input potentials are e_1 and e_2, an output voltage variation may be expressed as

$$\Delta e_0 = \mathcal{G}_1\,\Delta e_1 + \mathcal{G}_2\,\Delta e_2 \tag{43}$$

or

$$\Delta e_0 = \frac{\mathcal{G}_1 - \mathcal{G}_2}{2}(\Delta e_1 - \Delta e_2) + (\mathcal{G}_1 + \mathcal{G}_2)\frac{\Delta e_1 + \Delta e_2}{2}. \tag{44}$$

The second coefficient in Eq. (44), $\mathcal{G}_1 + \mathcal{G}_2$, should be zero or very small in comparison with the first coefficient $(\mathcal{G}_1 - \mathcal{G}_2)/2$.

In the case of Fig. 11·24 the proportional response to the common mode, expressed as the ratio of $(\mathcal{G}_1 + \mathcal{G}_2)$ to $(\mathcal{G}_1 - \mathcal{G}_2)/2$ is approximately $1/\mu$. For Fig. 11·25, where the triodes have been assumed to be identical, comparison of Eqs. (40) and (43) shows the ratio to be approximately

$$\frac{\mathcal{G}_1 + \mathcal{G}_2}{\frac{\mathcal{G}_1 - \mathcal{G}_2}{2}} \approx \frac{-r_p}{(\mu + 1)R_k}. \tag{45}$$

With given voltage levels, this ratio cannot be reduced indefinitely by increasing R_k because the resulting decrease of current causes r_p to increase

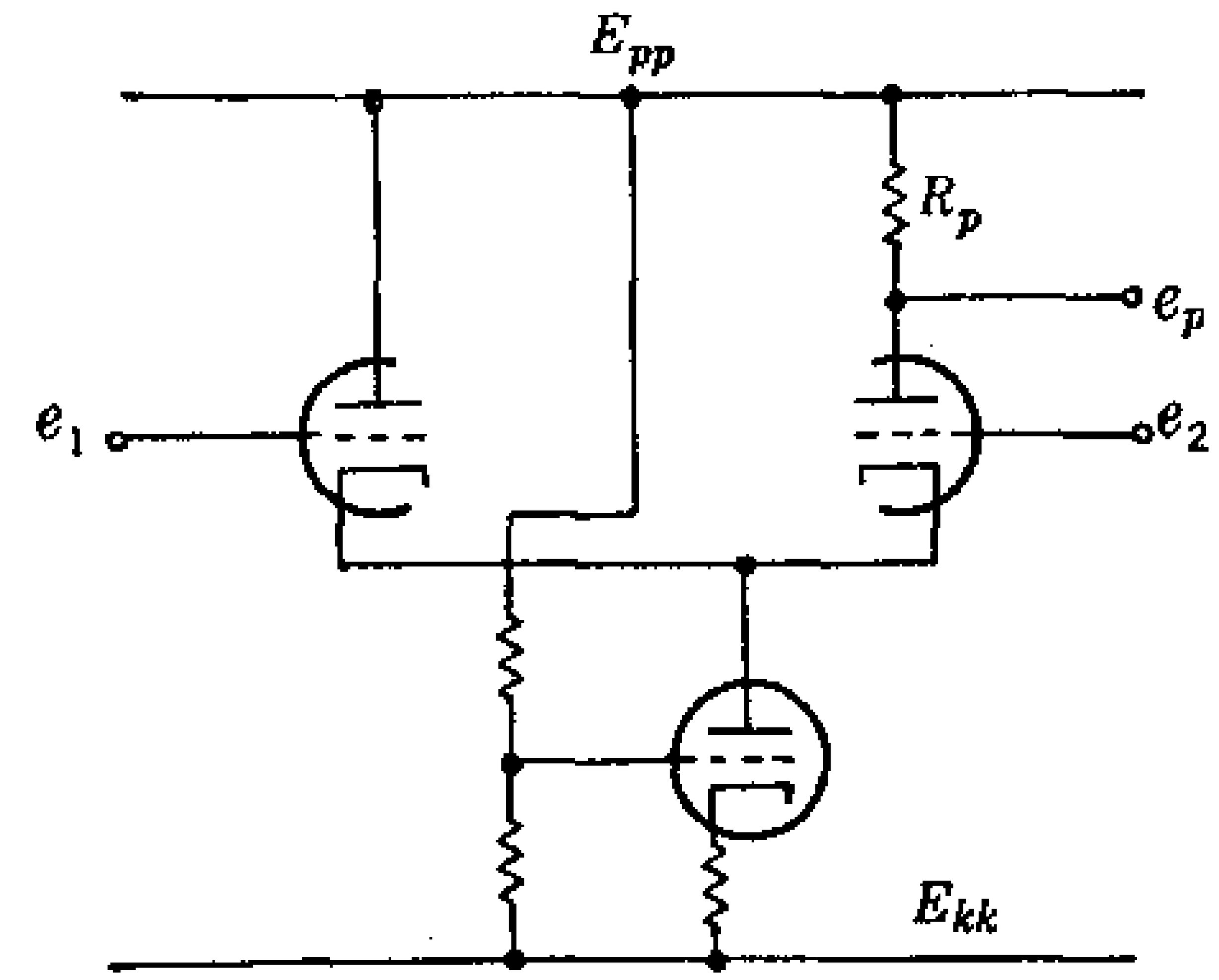

FIG. 11·26.—Use of constant-current device to minimize common-mode effect.

as well. A large negative return potential for B_k is advisable. Figure 11·26 illustrates the use of a constant-current device to simulate a very large R_k without reduction of current.

Even with an effectively infinite R_k, however, the common-mode rejection depends on equality between the μ's of the two triodes. The assumption of $R_k = \infty$ in Eqs. (39) and (41) shows that if the tubes in Fig. 11·22 are dissimilar, even though the constant-current device is perfect, the proportional common-mode variation is

$$\frac{\mathcal{G}_1 + \mathcal{G}_2}{\frac{\mathcal{G}_1 - \mathcal{G}_2}{2}} = \frac{\mu_1 - \mu_2}{\mu_1\mu_2 + \frac{\mu_1 + \mu_2}{2}}, \tag{46}$$

$$\approx \frac{\mu_1 - \mu_2}{\mu_1\mu_2}, \tag{47}$$

or approximately $1/\mu$ times the fractional difference of the μ's. Thus, high-μ tubes are indicated, and double triodes are preferable because they are generally better matched. For 6SL7's, for example, Eq. (47) rarely exceeds 0.1 per cent. The 6SU7 double triode is a specially designed and selected 6SL7 in which the quantity expressed by Eq. (47) is held to a minimum.[1]

From Eqs. (11) and (16), the output impedance of the circuit of Fig. 11·25 is

$$Z_p = \frac{R_p r_p\left(1 + \frac{\mu + 1}{\mu + 1 + \frac{r_p}{R_k}}\right)}{R_p + r_p\left(1 + \frac{\mu + 1}{\mu + 1 + \frac{r_p}{R_k}}\right)}. \tag{48}$$

When R_k is large, this is approximately

$$Z_p \approx \frac{2R_p r_p}{R_p + 2r_p}, \tag{49}$$

or R_p in parallel with twice r_p.

The foregoing equations in this section relate mainly to variational quantities. The equations for the d-c levels are cumbersome and are of

[1] Care should be exercised not to read into Eq. (45) more than it actually says. For instance, although pentodes have very large μ's, Eq. (45) does not apply to their use, since in its derivation it is assumed that $R_k \gg r_{p1}$; but in view of the large plate resistance characteristic of pentodes it is unlikely that this condition can easily be made to obtain. Thus, when two 6SJ7 pentodes ($\mu > 1000$) are employed in a circuit like that of Fig. 11·22, the common-mode rejection is found to be no better than when two 6SL7 triodes are employed.

Again, if R_k is made large by the use of a large ohmic resistor, the plate currents of the two tubes may be so small that their individual μ's may differ by more than the amount measured under normal operating conditions. Nor can Eq. (45) be expected to apply if, when two tubes of differing μ are employed, some other means is used to balance the amplifier to give zero output for a certain pair of grid potentials.

The above is possibly the explanation of the fact that when highly sensitive amplifiers are used to detect cross talk between the two tubes being tested, a variation of the common-mode bias, for balanced tubes, passes through zones where a minimum, a maximum, or a point of inflection is obtained at the output. In these regions the output is almost independent of small variations of common-mode signals, whereas between these regions the output varies at an appreciable rate. The length of these zones, in which the output is independent—within the noise level of the amplifier—of the common-mode signal, varies with the type of tube. For example, it is found that the region may extend over several hundred millivolts of common-mode signal for type 6SL7 tubes. But for type 6SN7 tubes the region may be as great as 1 or 2 volts. Another series of experiments indicated that this region was between 500 and 600 mv for pentodes (type 1620), when the criterion was the observation of a signal corresponding to a 10 μv difference between the two grid potentials. Editor.

less value in analysis and design. The problem in this regard would be, typically, to determine $e_1 - e_2$ for a certain value of e_p and for a given level of the input voltages. The sum of the two tube currents will be known in the case of the circuit of Fig. 11·26, and in Fig. 11·25 it is

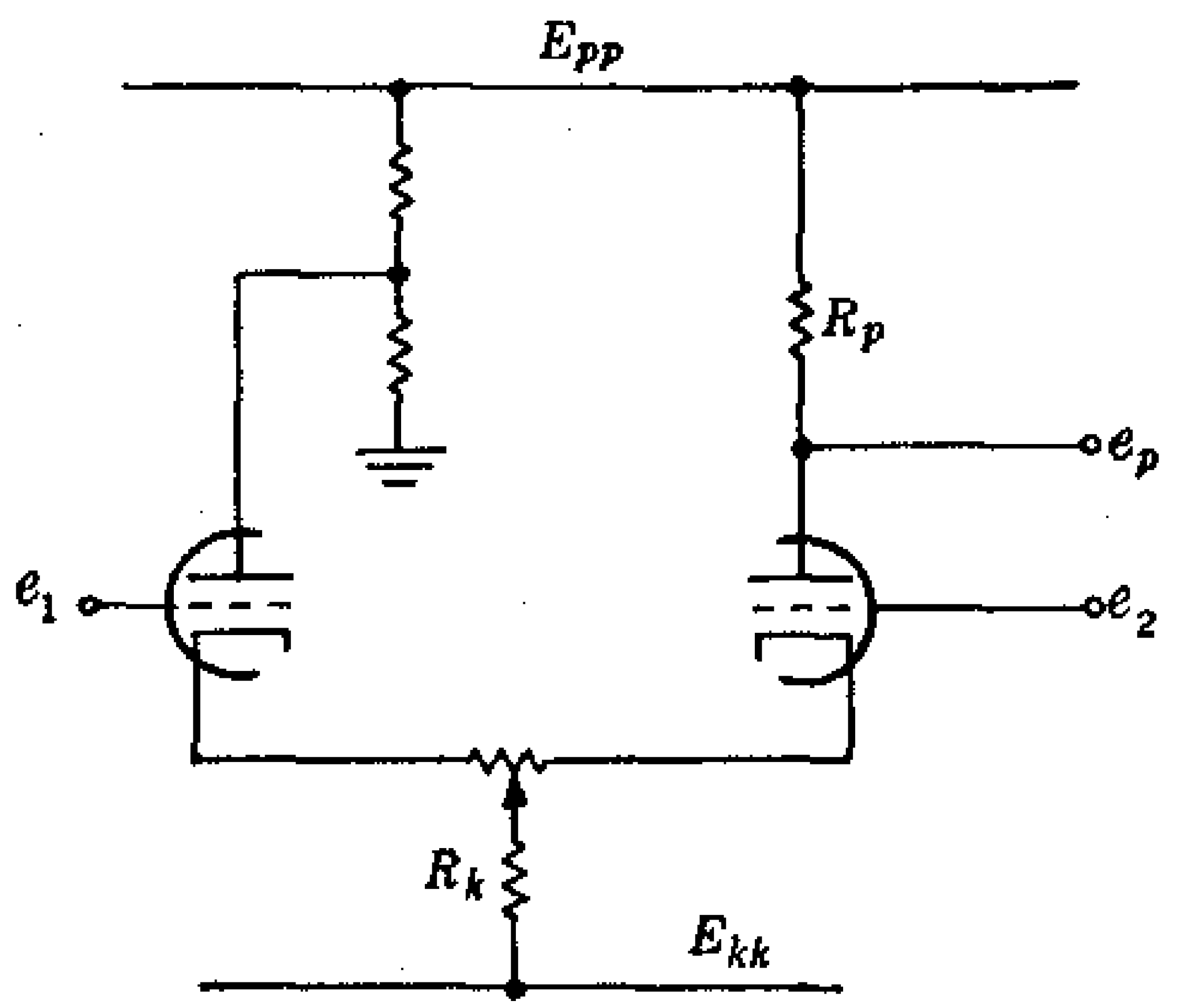

FIG. 11·27.—Method of zero adjustment.

usually approximately e_1/R_k. The current in the amplifier tube is $(E_{pp} - e_p)/R_p$, and that in the cathode follower is the difference between this and e_1/R_k. Thus the approximate e_{pk} and i_p are known for each tube, and the e_{gk}'s may be found by consulting the plate characteristic curves. Since the e_k's are the same, $e_1 - e_2$ is the difference between the e_{gk}'s.

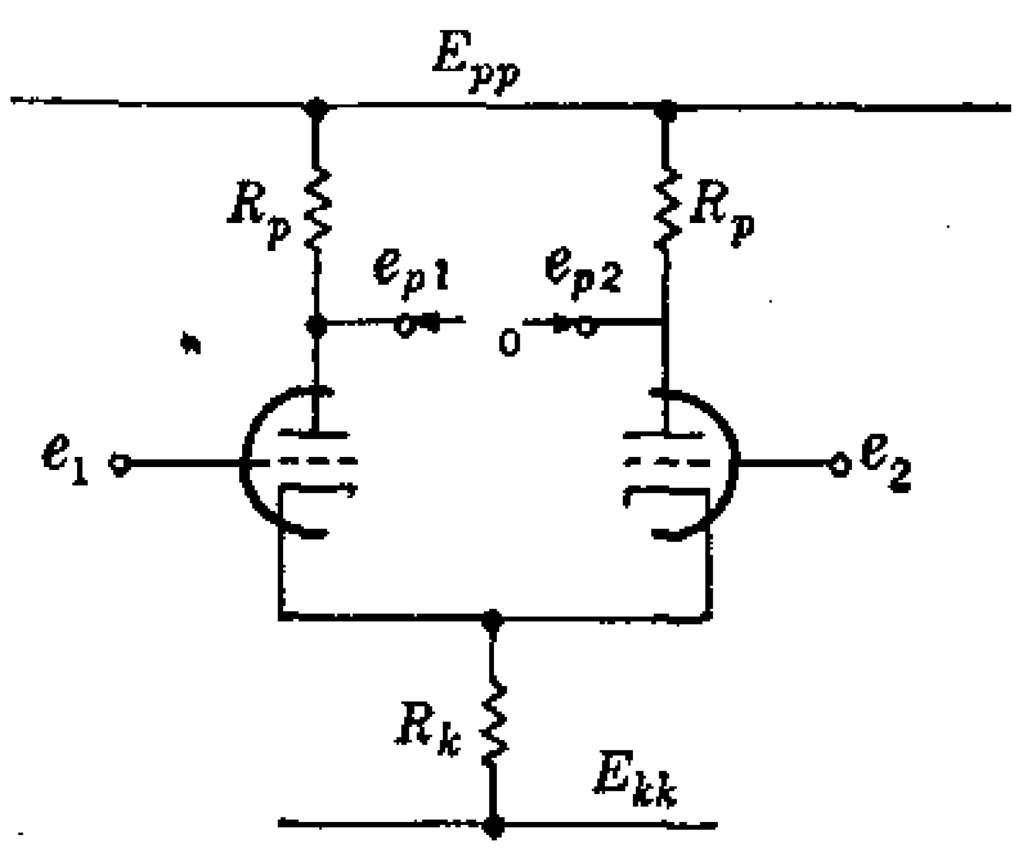

FIG. 11·28.—Symmetrical differential amplifier.

The cathode-follower input voltage e_1 generally will be somewhat lower than e_2 because of the higher plate voltage. The difference may be reduced by employing a lower-potential source for the cathode-follower plate; in fact, a potentiometer to this plate may comprise a zero adjustment. More commonly, however, the zero adjustment, if required, is a potentiometer between the cathodes as shown in Fig. 11·27. This decreases the gain by effectively adding $(\mu + 1)$ times half the resistance of the potentiometer to the r_p of each triode. (The resulting negative feedback may or may not be of value, depending on the application.) It also increases the common-mode susceptibility [Eq. (45)]. The value of this resistance depends on tube tolerances, but it may be held to a minimum by proper choice of the cathode-follower plate potential.

Differential Output.—A differential amplifier with two output terminals, such as the type illustrated by the symmetrical arrangement of Fig. 11·28, finds frequent application. It might be used as the first stage of a multistage amplifier; as a differential voltage output stage (actuating the plates of an oscilloscope, for example); or as a current or power output device, where the load (an ammeter, a relay, a motor field, a pair of magnetic oscilloscope deflecting coils, etc.), appears either between the two plates or, in two parts, as the R_p's.

To find the voltage gain from e_1 to e_{p2}, which is the same as that from e_2 to e_{p1} (assuming equal tubes), Eq. (39) may be applied. Because r_{p1} is increased by R_p, the resulting gain is

$$\mathcal{G}_{12} = \mathcal{G}_{21} = \frac{\mu R_p}{(R_p + r_p)\left[2 + \frac{R_p + r_p}{(\mu + 1)R_k}\right]}. \tag{50}$$

Similarly, from Eq. (41), the gain from e_1 to e_{p1} or from e_2 to e_{p2} is

$$\mathcal{G}_{11} = \mathcal{G}_{22} = \frac{-\mu R_p}{(R_p + r_p)\left[2 + \frac{R_p + r_p}{(\mu + 1)R_k}\right]}\left[1 + \frac{R_p + r_p}{(\mu + 1)R_k}\right] \tag{51}$$

The gain from either input terminal to the two output terminals across which the voltage e_0 appears is the difference between Eqs. (50) and (51).

$$\mathcal{G} = \frac{\mu R_p}{R_p + r_p} \tag{52}$$

This is identical with the gain of a simple triode amplifier and is not affected by R_k (except in so far as R_k determines the current and therefore affects r_p).

The differential output voltage is unaffected by common-mode variation of the input voltage (provided the μ's and r_p's are equal). Thus, for example, the input voltage might be applied to only one grid, the other remaining fixed, and the output voltage between the plates would be the same as if the input voltage were applied in "push-pull," one half to each grid. In the latter case, however, the output voltage would be truly "push-pull," whereas with one grid fixed the plate of the triode with the moving grid would move $1 + \frac{R_p + r_p}{(\mu + 1)R_k}$ times as far as the other plate. Usually this difference is very slight; but if R_k is small and it is important to have balanced output voltage, this may be accomplished by using slightly unequal R_p's.

The differential output voltage will be affected by the input voltage level if there are inequalities between the triode characteristics. This effect may be appreciable where the amplifier is being used to detect the difference between two potentials, each of which is subject to considerable variation. When the triode characteristics are different, the gain from e_1 to e_0 may be found from Eqs. (39) and (41) in a manner similar to that by which Eq. (52) was obtained. It is

$$\mathcal{G}_1 = \frac{\mu_1 R_p \left[2 + \frac{r_{p2} + R_p}{(\mu_2 + 1)R_k}\right]}{\frac{\mu_1 + 1}{\mu_2 + 1}(r_{p2} + R_p) + (r_{p1} + R_p) + \frac{(r_{p1} + R_p)(r_{p2} + R_p)}{(\mu_2 + 1)R_k}}, \quad (53)$$

where the subscripts refer to the respective triodes. The gain $\mathcal{G}_2$, from e_2 to e_0, is the same as $\mathcal{G}_1$ with opposite sign and subscripts. If the differences between the μ's and r_p's are small, the proportional response to the common-mode input voltage is approximately

$$\frac{\mathcal{G}_1 + \mathcal{G}_2}{\frac{(\mathcal{G}_1 - \mathcal{G}_2)}{2}} \approx \frac{1 + \frac{(r_p + R_p)}{2R_k}}{\mu + 1}\,\frac{\mu_1 - \mu_2}{\mu} + \frac{r_p}{2(\mu + 1)R_k}\,\frac{r_{p2} - r_{p1}}{r_p} \quad (54)$$

Thus, if a given percentage of dissimilarity between either the μ's or the r_p's is assumed, the appearance of the common-mode input voltage in the differential output voltage is lessened by the use of higher-μ triodes and a higher R_k.

In some applications it is desirable that not only the differential output voltage but also the level of the two output voltages be unaffected by the input voltage level. The variation of the output voltage level is approximately $-R_p/2R_k$ times that of the input voltage; this variation may be eliminated by the use of a constant-current circuit in place of R_k. This method also minimizes the effect on the differential output voltage [Eq. (54)], thereby eliminating the effect of difference between the r_p's.

As has been pointed out, unless R_k is very small, the output voltage of a differential amplifier is very nearly "push-pull" even if the input voltage is applied to only one grid. This effect comes about by virtue of the cathode coupling; the cathodes undergo excursions about half as great as those of the moving grid, thus furnishing an input voltage to the cathode of the tube with fixed grid and simulating a true push-pull input. An important result is improved linearity over a large range of output voltages. This result is indicated by Eq. (53), since the sum of r_{p1} and r_{p2} tends to remain more nearly constant than either of them alone. Figure 11·29 illustrates the degree of linearity effected, as well as the balanced output voltage obtained with a single input voltage (*cf.* Fig. 11·11).

Equation (52) indicates the rate of change of differential output voltage with respect to differential input voltage. By symmetry, the output voltage should be zero at zero input voltage. Therefore, from Eq. (52), the voltage equation is

$$e_{p2} - e_{p1} = \frac{\mu R_p}{R_p + r_p}(e_1 - e_2). \quad (55)$$

It is evident that the differential output voltage is independent of plate-supply voltage, whereas in the case of the single-triode amplifier [Eq. (3)]

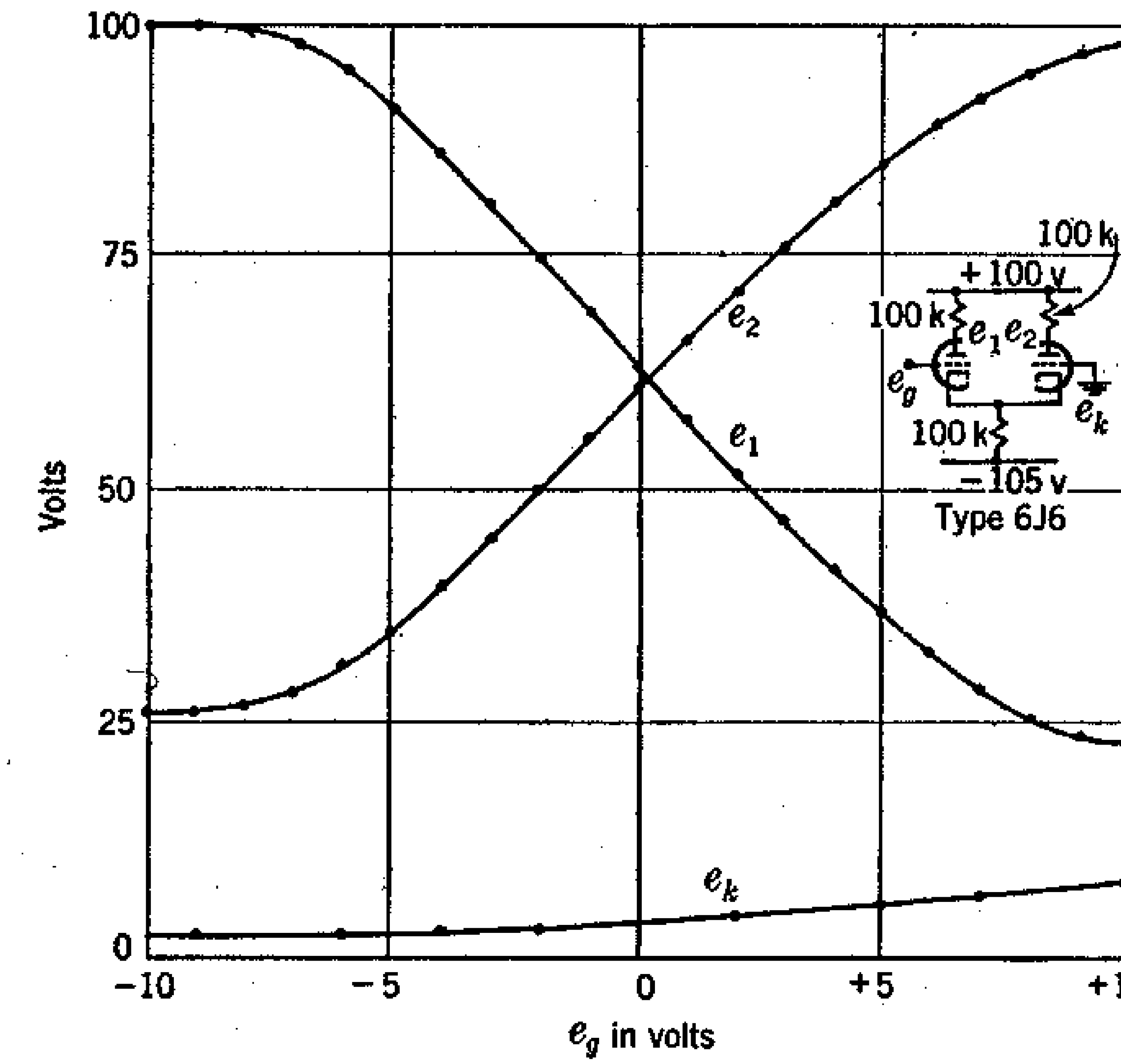

FIG. 11·29.—Example of differential amplifier and characteristic

any variation of E_{pp} appears in the output voltage multiplied by

$$\frac{r_p}{(R_p + r_p)}.$$

Inequalities between the two triodes will result in Eq. (55) being in error by a constant amount. Often this error is immaterial, as the zero adjustment can be located elsewhere in the circuit. If zero output voltage at zero input voltage difference is required, a potentiometer between the cathodes, as in Fig. 11·27, usually is satisfactory, although it decreases the gain to some extent. It is sometimes preferable to employ a potentiometer in the plate circuit to increase one R_p and decrease the other. In either case, the unbalance required by the zero adjustment tends to spoil the rejection of the common mode and the push-pull nature of the output voltage; if these are important, a pair of triodes like the 6SU7, which requires the minimum of zero adjustment, should be used.

Addition of a separate resistance in series with each cathode decreases the gain as given by Eq. (52) by the addition to the denominator of $\mu + 1$ times this resistance. This resistance may be used as a gain adjustment where desired. (The increase of linearity accompanying the decrease of gain is an advantage over simple attenuation.) To avoid the use of a double potentiometer, the π-connected equivalent circuit as shown in Fig. 11·30 may be used instead of the T-connected cathode circuit. The gain potentiometer is generally of low resistance as compared with the cathode-return resistance, and each of these is about twice the value that R_k would have in the T-circuit. There is a minimum of interference between the gain and zero adjustments as shown.

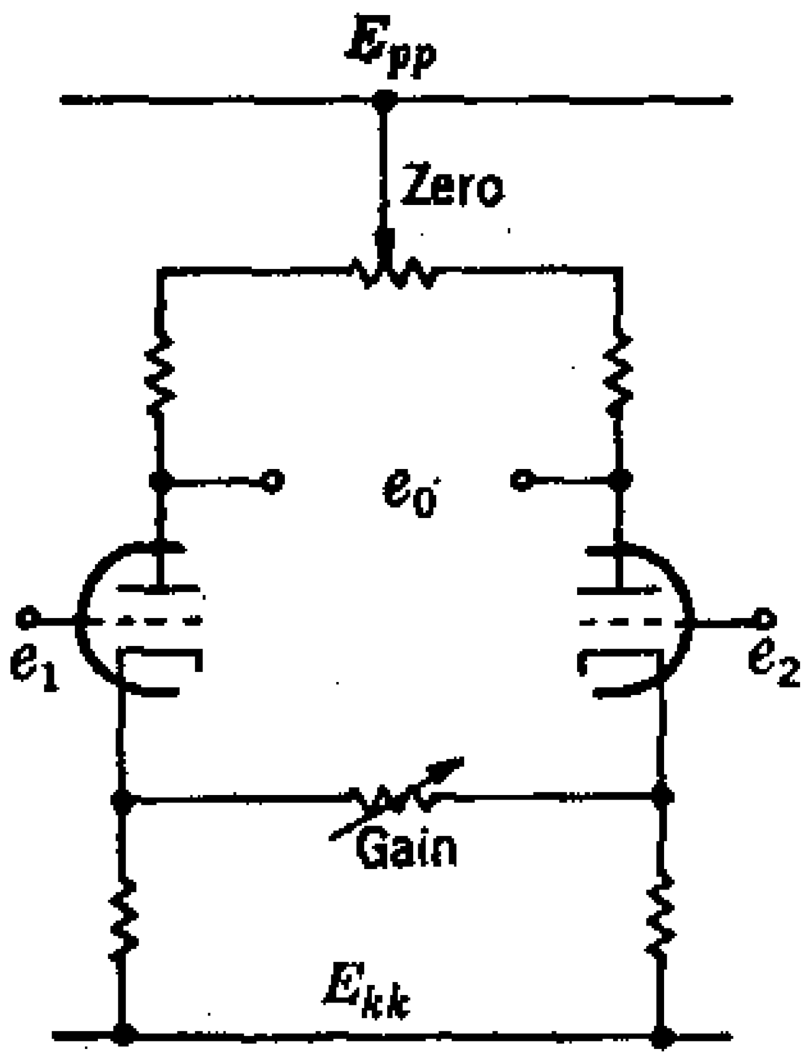

FIG. 11·30.—Differential amplifier with gain and zero adjustments.

Pentode Differential Amplifier.—Equation (55) may be written in the form

$$e_{p2} - e_{p1} = \frac{g_m R_p}{1 + \dfrac{R_p}{r_p}} (e_1 - e_2). \tag{56}$$

If pentodes are employed, the plate resistance may be assumed infinite, and

$$e_{p2} - e_{p1} = g_m R_p (e_1 - e_2). \tag{57}$$

Thus a somewhat higher voltage gain is obtainable with pentodes than with triodes. The special problem in the case of the pentode differential amplifier concerns the screen potential and current.

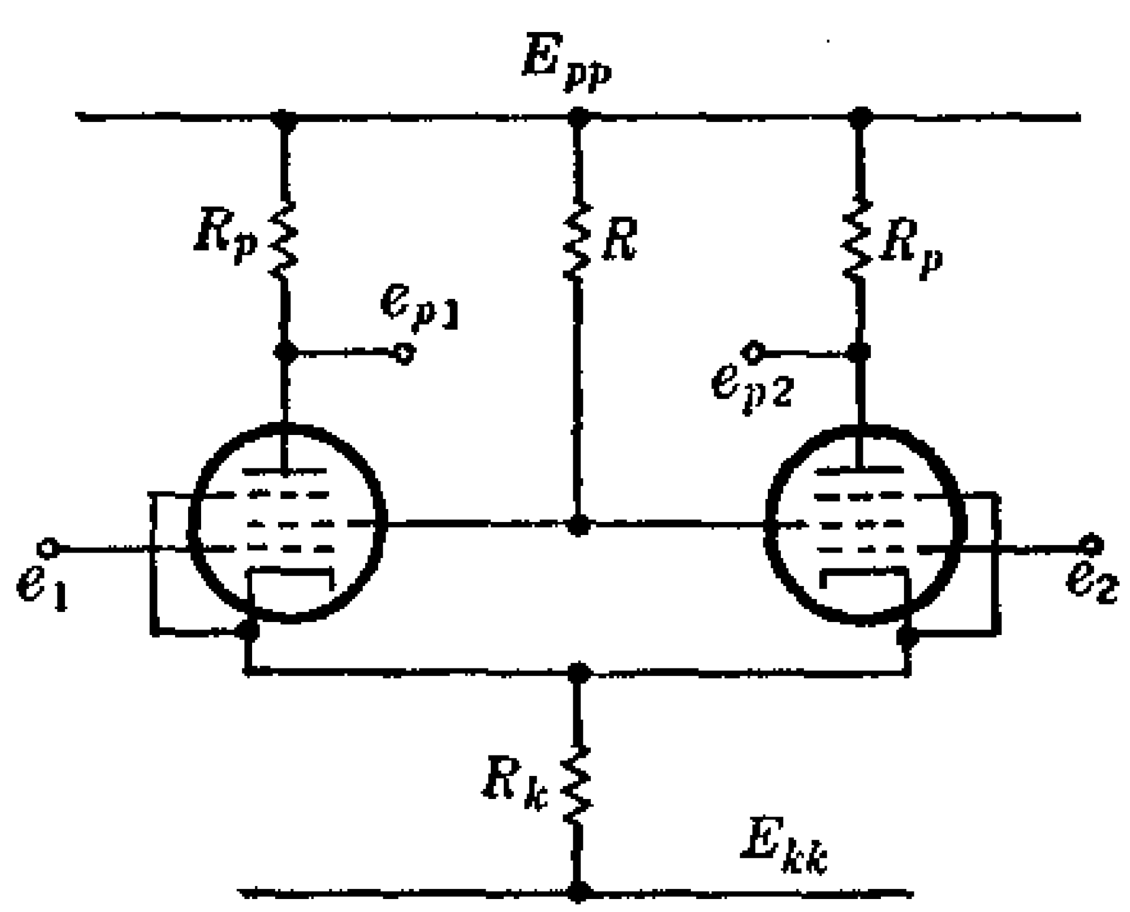

FIG. 11·31.—Pentode differential amplifier.

If a battery or floating supply can be used to furnish cathode-screen potential, good common-mode rejection may be realized. Also, since the screen current does not flow in the common cathode resistor, the output voltage is balanced as in the triode amplifier.

In general, however, the screen grids are supplied from a fixed source or by means of a common resistor as in Fig. 11·31. The current in R_k is then the sum of the plate and screen currents. If the total screen current varies, the push-pull aspect of the output voltage will suffer. This effect is likely to be noticeable if either plate potential falls much below the screen potential. Therefore, R_s should be chosen carefully, with this in mind.

11·11. Output Circuits. *Requirements.*—The type of output signal required of a direct-coupled amplifier varies greatly. Therefore, each application must be considered separately in choosing the appropriate output stage. Loads may be classified according to any of the following characteristics; the impedance, voltage, and current range required; whether or not both positive and negative voltage or current are expected; the degree of linearity desired; and the number of terminals needed. An example of a single-terminal load is an amplifier in which a voltage is to be fed back in a d-c computing circuit, wherein the amplifier is a voltage-equating device. A two-terminal load might be the deflecting plates of an oscilloscope, a voltmeter or ammeter, a relay, a motor field, or a magnetic oscillograph. Some loads, such as magnetic oscilloscopes, differential relays, magnetic amplifiers, and servo-motor fields, have three or four terminals for differential excitation.

If there are no preceding stages in the amplifier, the design must take account of the input voltage level and provide an adjustment for variations of components in accordance with the principles of preceding sections. Generally, however, there will be one or more stages of voltage amplification ahead of the output stage, and it will be assumed that the zero adjustment has been taken care of there. The actual output voltage from the preceding stage will be considerably above ground and usually must be dropped down to fit the input voltage level of the output state by a divider that also attenuates. Therefore, a high input voltage level is usually preferable.

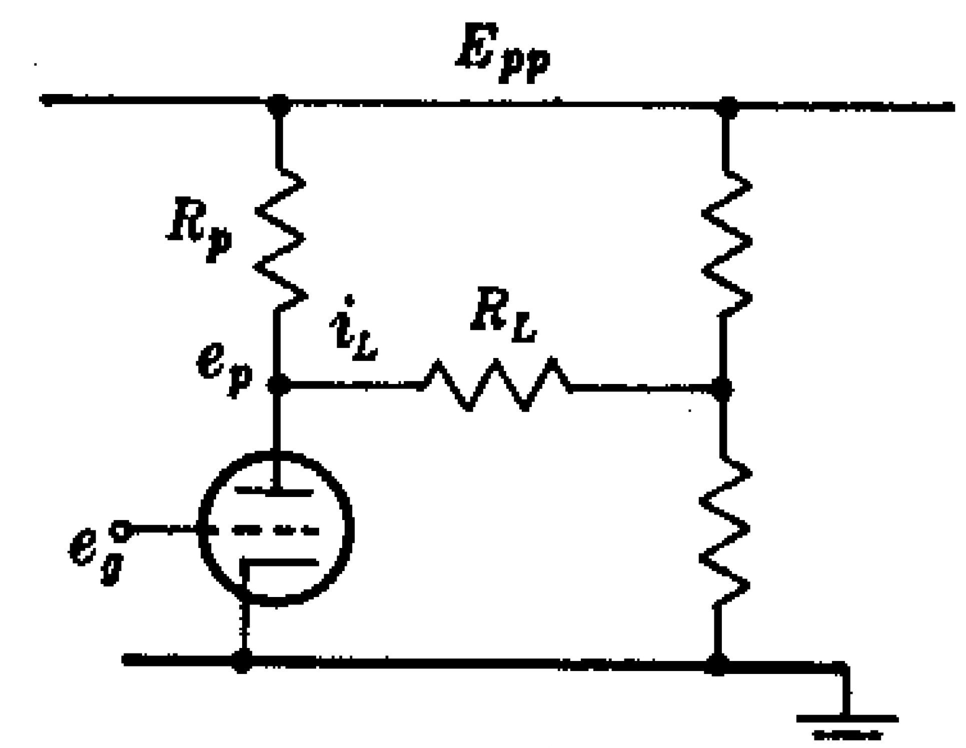

FIG. 11·32.—Plate as output terminal.

Simple Amplifier with Plate as Output Terminal.—If the load is of very high resistance or if the required voltage swing is small, the load may be connected between some fixed point and the plate of the output tube, as in Fig. 11·32. In computing the performance of the amplifier with the load, Thévenin's theorem may be applied, using the unloaded characteristics and the output impedance. Another method involves considering R_L in parallel with R_p as the net plate resistor. In this case a fictitious E_{pp}, which is equal to the value of voltage assumed by e_p with the tube removed, should be employed to determine the output characteristic. From Eq. (5), with R_p paralleled with R_L, the voltage gain is

$$\mathcal{G} = \frac{\mu}{1 + \frac{r_p}{R_p} + \frac{r_p}{R_L}}. \tag{58}$$

For a low-resistance load, such as a meter or a relay, the voltage gain is of much less interest than the rate of change of load current with input voltage, which is called the "current gain." This quantity is equal to the voltage gain divided by R_L.

$$\frac{\Delta i_L}{\Delta e_g} = -\frac{g_m}{1 + \frac{R_L}{r_p} + \frac{R_L}{R_p}}. \tag{59}$$

Often the load resistance may be neglected, in which case

$$\frac{\Delta i_L}{\Delta e_g} \approx -g_m. \tag{60}$$

If the output voltage is to be "single-ended" with a large working range encompassing ground potential, a negative voltage supply will

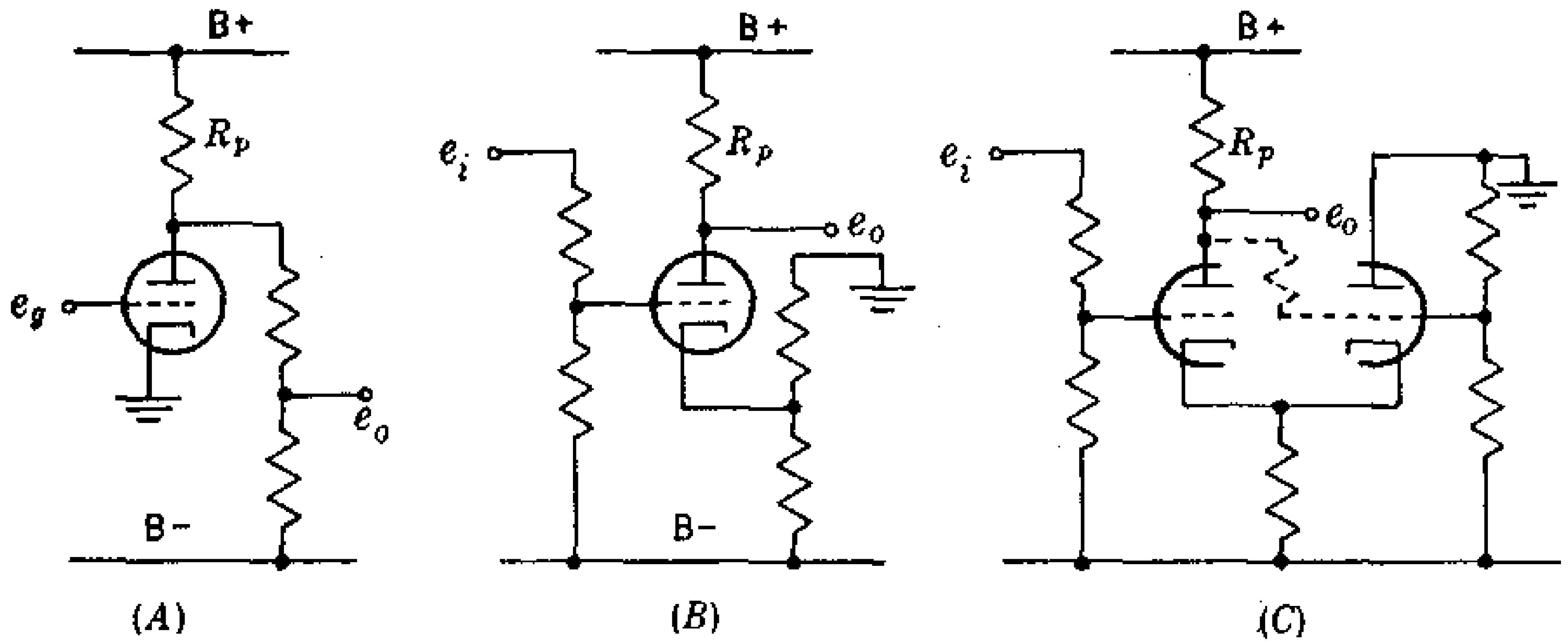

FIG. 11·33.—Voltage output embracing ground potential.

be needed. Two general methods of obtaining this voltage are shown in Fig. 11·33. In (*a*) the output voltage is obtained from a divider, and in (*b*) the input voltage and the cathode supply are taken from dividers. If it is assumed that the tube permits a certain minimum plate voltage without having appreciable grid current and that a certain $B-$ voltage is available, it is found that the output voltage is somewhat less attenuated in (*a*) than is the input voltage in (*b*); also, in (*b*) the impedance of the cathode tie point decreases the gain. On the other hand, the output impedance in (*a*) is much higher than in (*b*), and the upper limit of the output range is lower. For these reasons it is sometimes advisable to employ the depressed cathode method, with a cathode-follower tie point as in Fig. 11·33*c*. In addition to providing a low-impedance cathode return, this method permits the application of positive feedback (dashed resistor), which is beneficial if used in conjunction with over-all negative feedback (see Sec. 11·13). In any case a high-resistance plate resistor should be employed because the greater the plate current the higher the minimum allowable plate voltages. A pentode tube is much superior to a high-μ triode in point of low limiting plate voltage and is therefore generally preferable in this application.

Cathode-follower Output Stage.—If the output voltage requirement is the same as in the foregoing paragraph but the current load is appreciable, a cathode follower may be added to any of the circuits of Fig. 11·33. Its cathode resistor should be returned to the $B-$ voltage to permit the output voltage to reach zero without much nonlinearity. Its input voltage will have to drop somewhat below zero, but this amount is minimized by the use of a high-μ tube and by a plate supply that is no higher than needed in view of the maximum output voltage required.

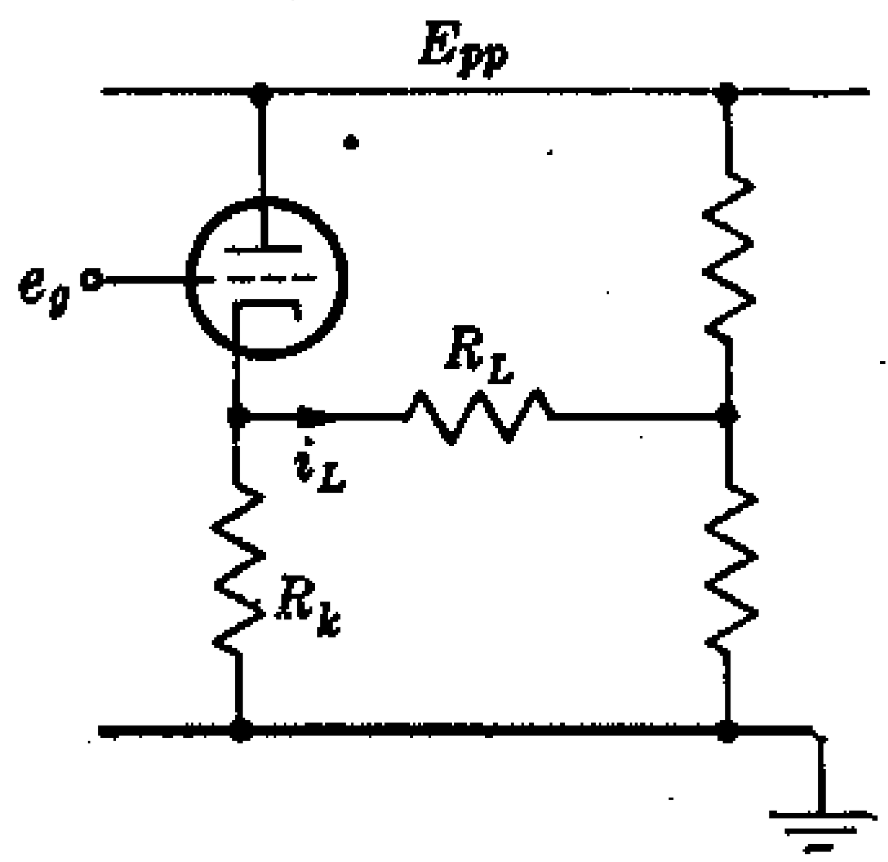

FIG. 11·34.—Cathode-follower output stage.

If one terminal of the load is or can be at some fixed positive potential, the arrangement might be as shown in Fig. 11·34. By Thévenin's theorem, from Eqs. (15) and (16), the voltage "gain" with the load is

$$\mathcal{G} = \frac{\mu}{\mu + 1 + \frac{r_p}{R_k} + \frac{r_p}{R_L}} \tag{61}$$

(R_L includes the resistance of the load tie point). The current gain is this expression divided by R_L.

$$\frac{\Delta i_L}{\Delta e_g} = \frac{g_m}{1 + (\mu + 1)\frac{R_L}{r_p} + \frac{R_L}{R_k}}, \tag{62}$$

which, if R_L is negligible, approaches

$$\frac{\Delta i_L}{\Delta e_g} \approx g_m, \tag{63}$$

just as in the case of Fig. 11·32. In fact, if R_L is negligible, the only difference between Figs. 11·32 and 11·34 is the nature of the saturation, and even here the difference is one of sense only. In Fig. 11·34 the maximum reverse current is the current obtained with the tube removed and depends on the magnitude of R_k; the maximum forward load current depends only on tube capability and plate voltage. A similar, but reverse, situation obtains for Fig. 11·32. Where R_L is appreciable, the current equations (59) and (62) differ by μR_L in the denominator. In this case, the gain of Fig. 11·34 is lower, but its inherent negative feedback gives it greater linearity.

Differential Cathode Follower.—The resistance of the return point for the load resistor may decrease the current gain considerably if a bleeder is used as in Fig. 11·34. It may be advisable to employ a second cathode follower as a tie point as in Fig. 11·35. Similarly to Eq. (62), the current gain is found to be

$$\frac{\Delta i_L}{\Delta e_g} = \frac{g_m}{2 + (\mu + 1)\dfrac{R_L}{r_p} + \dfrac{R_L}{R_k}}, \tag{64}$$

or, for very small load resistance,

$$\frac{\Delta i_L}{\Delta e_g} \approx \frac{g_m}{2}. \tag{65}$$

Saturation in either direction occurs when one of the tubes cuts off. The load current is then simply the current in R_k. In the forward direction there is still a slight current increase beyond this point, and Eq. (62) then applies with R_L increased by R_k.

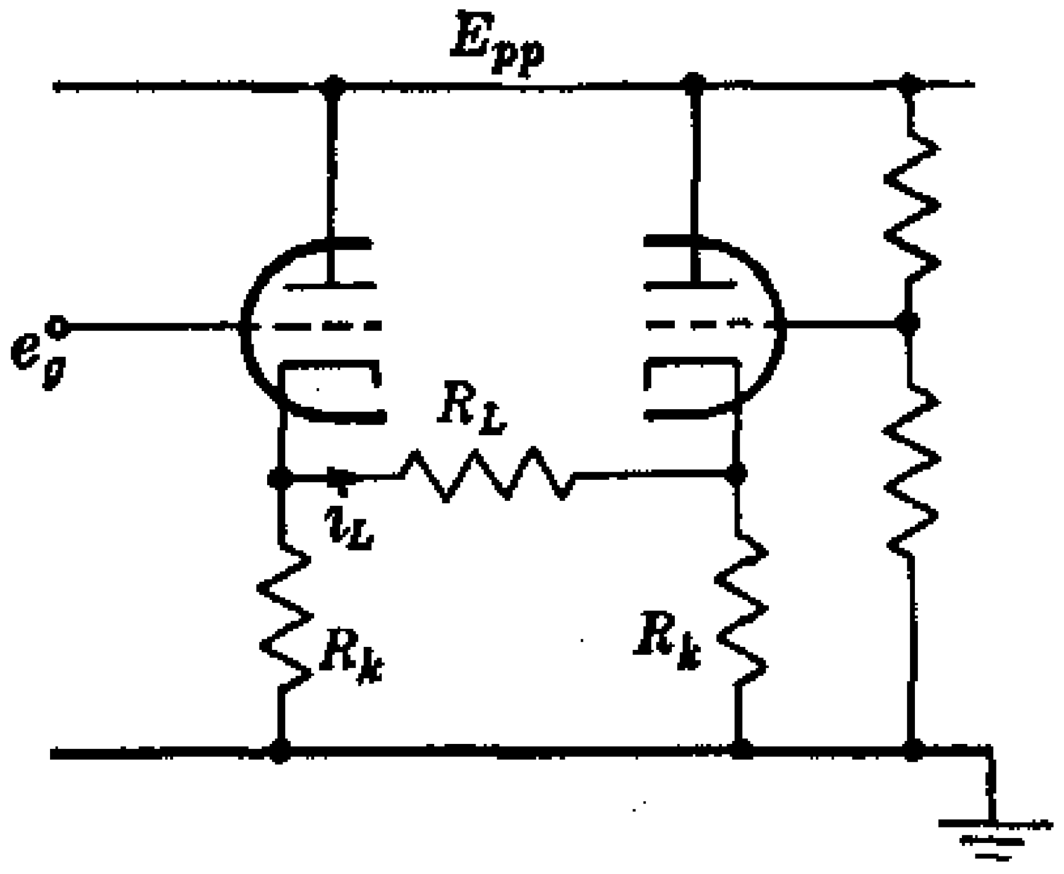

FIG. 11·35.—Differential cathode follower.

This circuit is convenient for driving ammeters and magnetic oscillographs, which have negligible resistance and which need the protection against overloads afforded by the saturation. Good linearity is afforded by mutual cancellation of plate resistance variation. Extra linearity may be obtained at the expense of gain, simply by the addition of resistance in series with the load.

The input terminals may conveniently be at both grids, instead of at only one as shown, from the two plates of a differential amplifier.

Differential Amplifier.—The case of a high-impedance push-pull load, such as oscilloscope plates, is covered by the differential amplifier analysis of Sec. 11·10. The voltage range available depends on the $B+$ voltage supplied, and the push-pull nature of the output voltage, if the input voltage is single-ended, depends on the $B-$ voltage being large in comparison with the input voltage.

To describe the performance of the differential amplifier with a load, as in Fig. 11·36, it is perhaps easiest to begin with the assumption of zero load resistance. In this case, since the cathodes are tied together and the plate voltages are equal, the difference between the plate currents is

$$i_{p1} - i_{p2} = g_m(e_1 - e_2). \tag{66}$$

The currents in the two R_p's are $i_{p1} + i_L$ and $i_{p2} - i_L$ respectively. Since the R_p's are equal and receive the same voltage, these currents are equal, and Eq. (66) becomes (for $R_L = 0$)

$$i_L = \frac{g_m}{2}(e_2 - e_1). \tag{67}$$

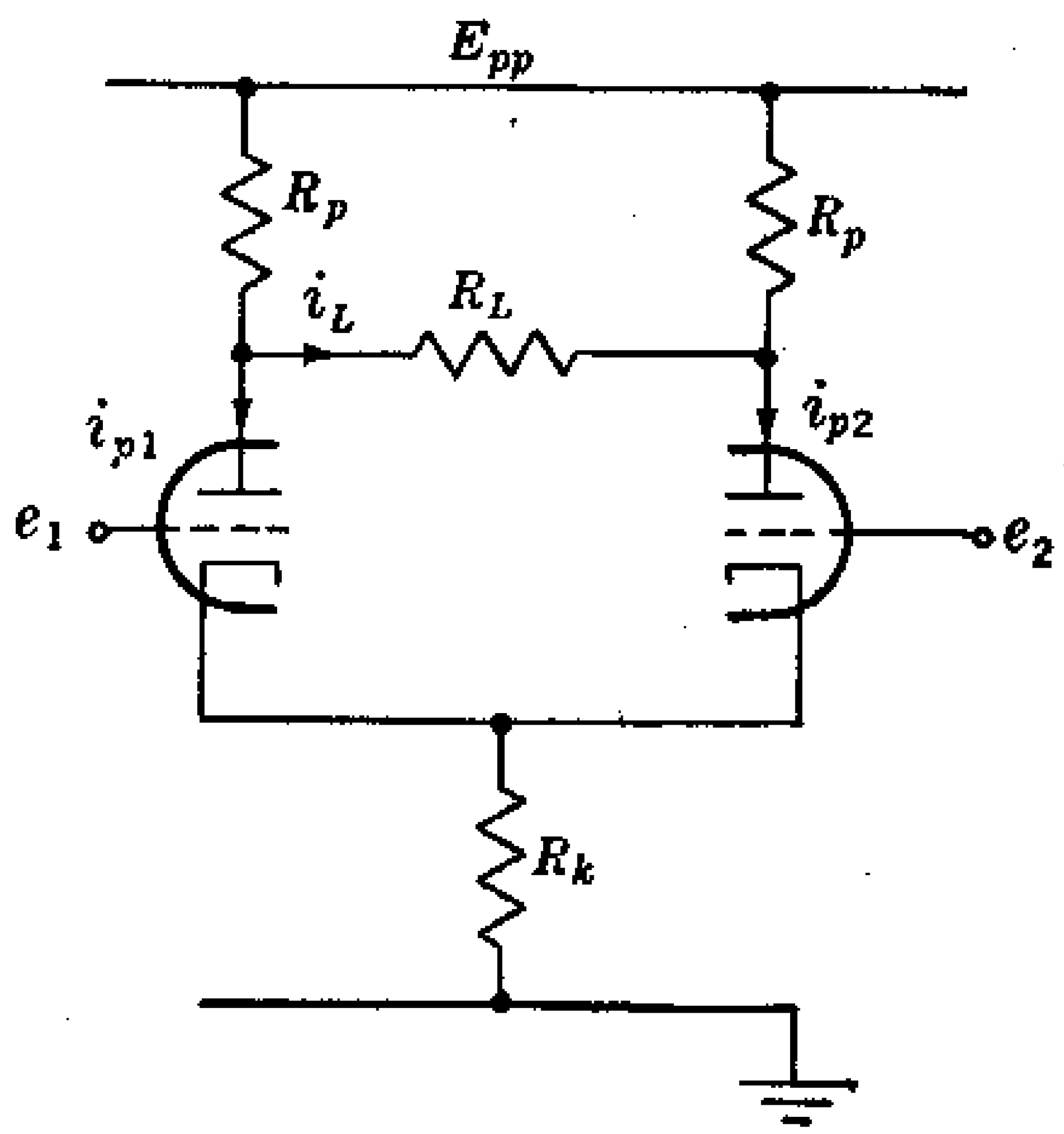

FIG. 11·36.—Differential amplifier with load.

The output impedance of the amplifier is the ratio of open-circuit voltage to short-circuit current and thus is found, from Eqs. (67) and (55), to be

$$Z_0 = \frac{2r_pR_p}{R_p + r_p}, \tag{68}$$

or twice the output impedance of a simple amplifier. Application of Thévenin's theorem to Eqs. (55) and (68) gives the voltage gain with load as

$$\mathcal{G} = \frac{-\mu}{1 + \dfrac{r_p}{R_p} + \dfrac{2r_p}{R_L}}, \tag{69}$$

from which the current gain is found to be

$$\frac{\Delta i_L}{\Delta(e_1 - e_2)} = -\frac{g_m}{2 + \dfrac{R_L}{r_p} + \dfrac{R_L}{R_p}}. \tag{70}$$

It is evident that when, as in an ammeter, the load resistance is negligible, the gain of a differential amplifier is identical with that of a differential cathode follower. The linearity of the two types is the same, and the linearity of the differential amplifier may be improved at the expense of gain by insertion of resistance between the cathodes, as in Fig. 11·30. A possible reason for preferring the cathode follower would be the extra damping afforded a meter by the low output impedance. For loads with considerable resistance, like a motor field, a relay, or a solenoid, the differential amplifier is preferable when maximum gain is desired. Also, when a quick current response in spite of load inductance is required, it is preferable to use the plate circuit. In this case, the current through the load tends to follow the grid voltage without being influenced by the voltage across the load as much as it would be if the load were in the cathode circuit.

If a pentode differential amplifier, where μ and r_p are almost infinite, is used as an output circuit, Eq. (70) reduces to

$$\frac{\Delta i_L}{\Delta(e_1 - e_2)} = -\frac{g_m}{2 + \dfrac{R_L}{R_p}}. \tag{71}$$

Twice as much current gain can be realized if the load is divided into two parts, as in the case of a differential relay or a magnetic oscilloscope, where each of the two parts is simply one of the R_p's and the output is the difference between these currents. For a pair of triodes, this output current is [adding R_p to r_p in Eq. (61)]

$$i_1 - i_2 = \frac{\mu}{r_p + R_p}(e_1 - e_2). \tag{72}$$

In the case of a pair of pentodes, the load resistance does not decrease the gain, and Eq. (66) applies directly.

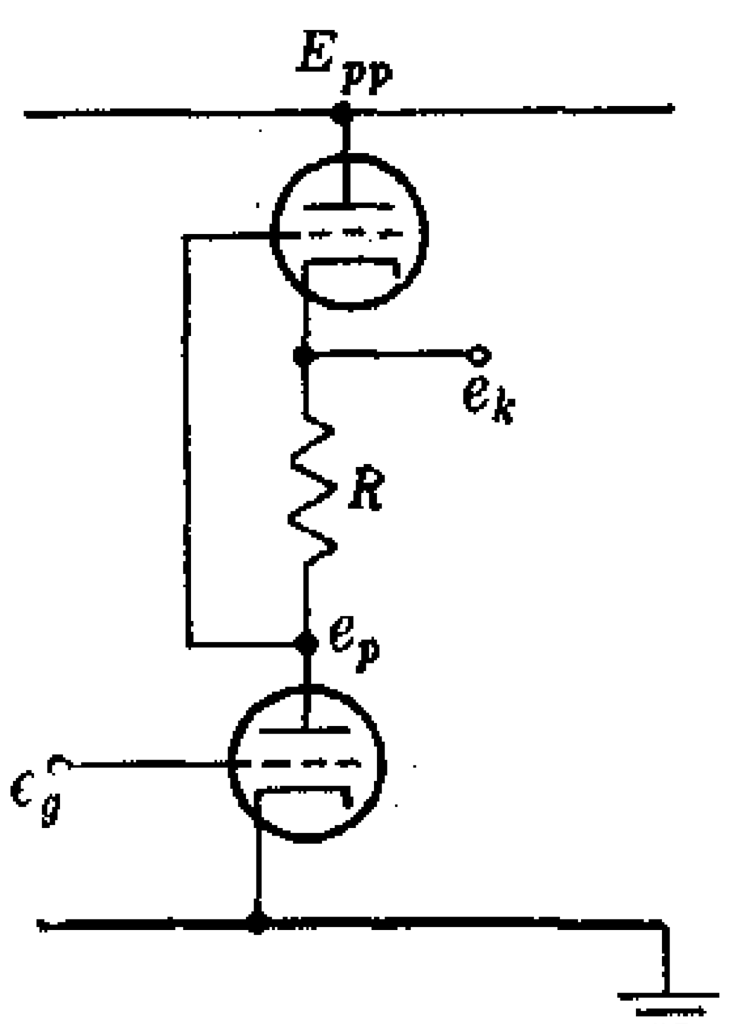

FIG. 11·37.—Series amplifier.

Two-tube Series Arrangement.—The circuit of Fig. 11·37, where e_k is the output voltage, is sometimes useful as a power amplifier. A similar circuit, having another resistor equal to R in series with the lower cathode and with the output terminal at e_p, affords cancellation of heater-voltage variation and is described in Sec. 11·12. The circuit between E_{pp} and e_p, comprising the plate load for the lower tube, resembles the constant-current circuit of Fig. 11·18, but without the battery it is not actually a constant-current device. It is merely the equivalent of a simple resistance, of value $r_p + (\mu + 1)R$, returned to E_{pp}. Thus in Fig. 11·37 the voltage gain to the e_p terminal is

$$\frac{\Delta e_p}{\Delta e_g} = \frac{-\mu[r_p + (\mu + 1)R]}{2r_p + (\mu + 1)R}. \tag{73}$$

Since all the current flows through R, the cathode is at a point at a distance R from the e_p end of the total equivalent resistance $r_p + (\mu + 1)R$, and therefore e_k moves $\dfrac{r_p + (\mu + 1)R - R}{r_p + (\mu + 1)R}$ times as much as e_p. Thus, the gain to the cathode (with no load) is

$$\frac{\Delta e_k}{\Delta e_g} = \frac{-\mu(r_p + \mu R)}{2r_p + (\mu + 1)R}. \tag{74}$$

To find the gain with a load, it is easiest first to determine the current gain in the case of a zero-resistance load. If R_L is zero in Fig. 11·38, so that e_k is fixed at E, a change Δe_g will produce a plate current increment Δi_{p1} of $\mu\,\Delta e_g/(r_p + R)$. This increment, in turn, lowers the upper grid by a voltage increment R times this, so that the upper tube current changes by an amount Δi_{p2}, equal to $-\mu^2 R\,\Delta e_g/r_p(r_p + R)$. Thus, if $R_L = 0$, the net current gain is

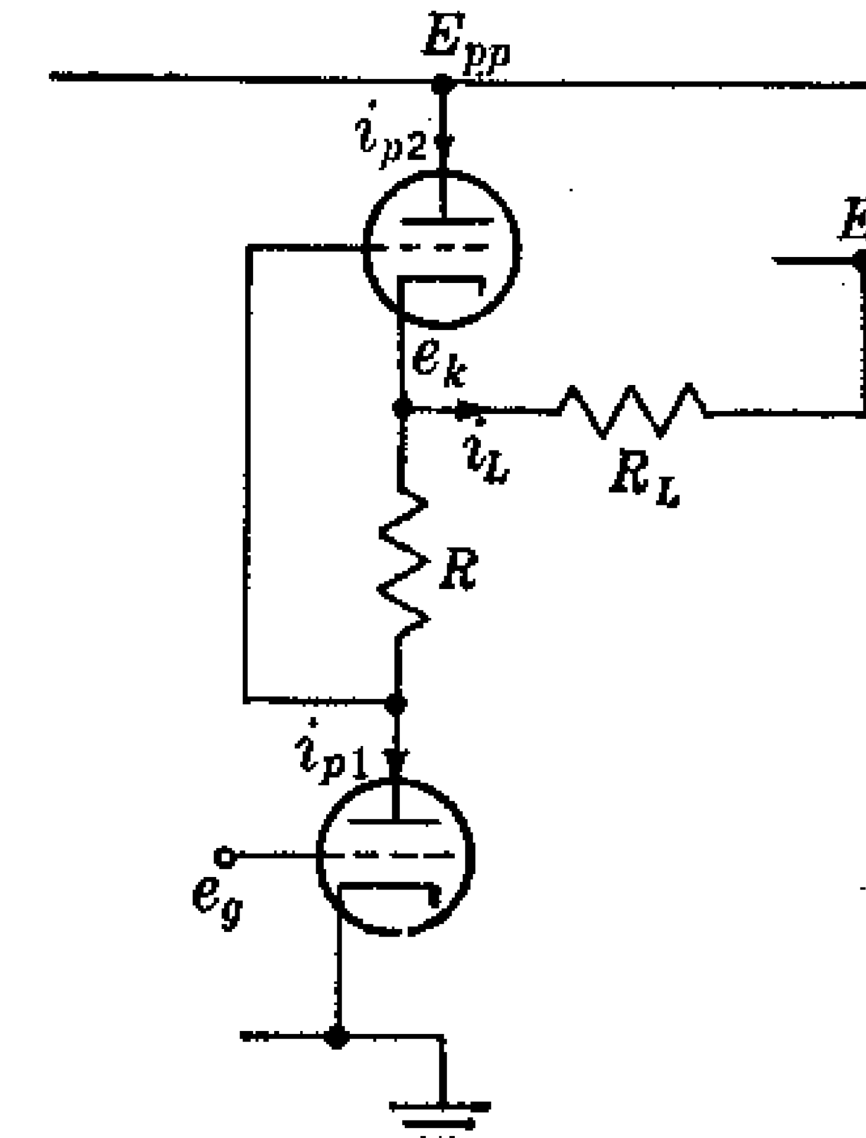

FIG. 11·38.—Series amplifier with load.

$$\frac{\Delta i_L}{\Delta e_g} = -\frac{\mu}{r_p + R} - \frac{\mu^2 R}{r_p(r_p + R)} = -g_m\frac{r_p + \mu R}{r_p + R}. \tag{75}$$

The current gain with a load of negligible resistance, where a suitable intermediate voltage source exists for a load tie point, is considerably greater than that for a simple amplifier or a differential amplifier or cathode follower.

The output impedance is the ratio of open-circuit voltage gain to short-circuit current gain.

$$Z_0 = r_p\frac{r_p + R}{2r_p + (\mu + 1)R}. \tag{76}$$

From Eq. (74), by means of Thévenin's theorem, the voltage gain with any load resistance is found to be

$$\mathcal{G} = \frac{-\mu(r_p + \mu R)}{2r_p + (\mu + 1)R + (r_p + R)\dfrac{r_p}{R_L}}. \tag{77}$$

The current gain with any R_L is

$$\frac{\Delta i_L}{\Delta e_g} = -g_m\frac{r_p + \mu R}{r_p + R\left[1 + (\mu + 1)\dfrac{R_L}{r_p}\right] + 2R_L}. \tag{78}$$

It is apparent from Eq. (78) that as long as R_L is considerably smaller than r_p, R may be chosen so that the current gain is several times g_m. The maximum range of output current in both positive and negative directions is achieved with $E = E_{pp}/2$ and $R = 1/g_m$, but the maximum gain occurs with an R of several times $1/g_m$.

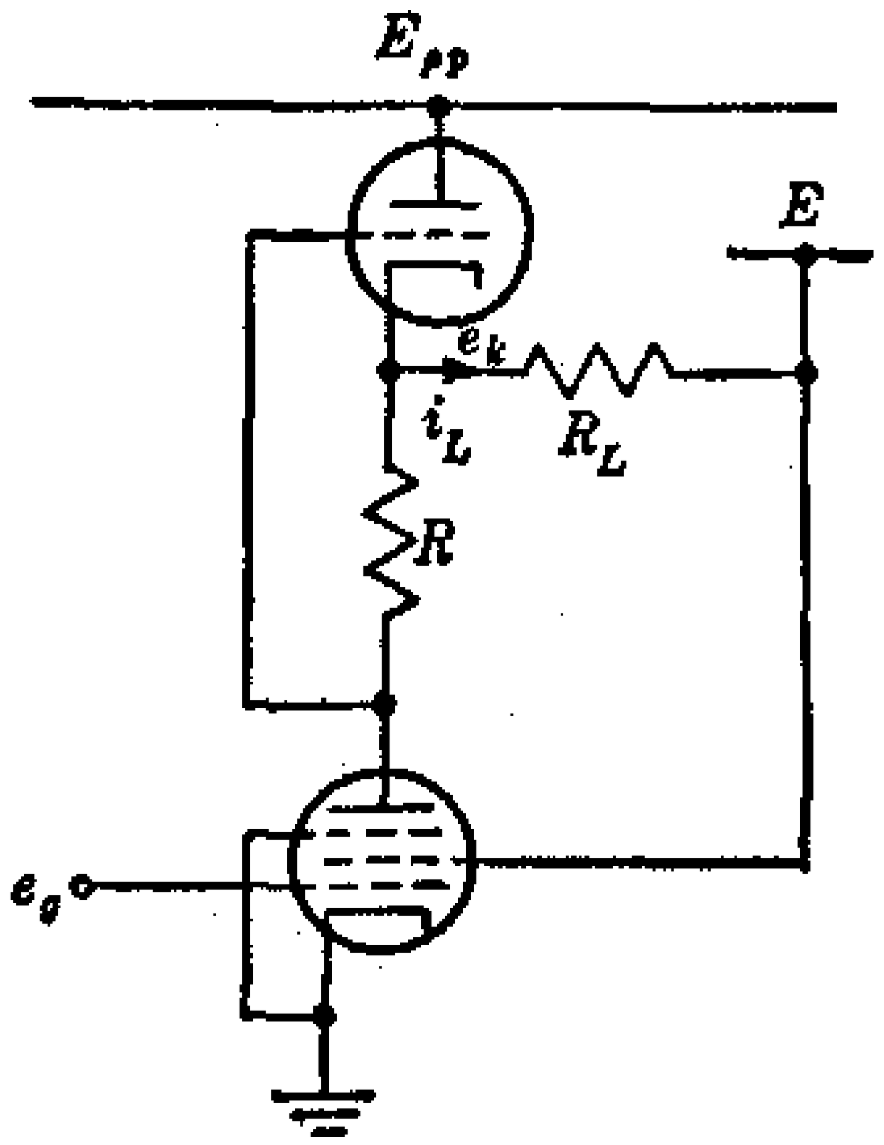

FIG. 11·39.—Series amplifier using pentode.

If a pentode is used in the lower position, as in Fig. 11·39, the formulas are simpler. In this case, Eq. (74), the voltage gain with $R_L = \infty$, becomes

$$\frac{\Delta e_k}{\Delta e_g} = -g_m(r_p + \mu R), \qquad (79)$$

where g_m refers to the pentode and μ and r_p to the triode. The output impedance is simply

$$Z_0 = r_p, \qquad (80)$$

because the pentode current is independent of its plate voltage, so that if e_k is moved by external means, the triode bias will remain constant and its current will vary according to plate resistance. From Eqs. (79) and (80) the voltage gain with load is found to be

$$\mathcal{G} = -g_m \frac{r_p + \mu R}{1 + \frac{r_p}{R_L}}, \qquad (81)$$

and the current gain is

$$\frac{\Delta i_L}{\Delta e_g} = -g_m \frac{r_p + \mu R}{r_p + R_L}. \qquad (82)$$

A practical example of this output circuit is given in Fig. 11·42. Both tubes are pentodes, but the upper tube behaves like a triode because its plate and screen both are fixed.

A comparison of this circuit with the differential amplifier shows that for tubes of the same capabilities, the former has at least four times the gain and twice the maximum output current in both directions as the latter. On the other hand, this circuit requires a low-impedance intermediate voltage source, and, for a given available $B+$ voltage, the input voltage level must be considerably lower than that for a differential amplifier.

11·12. Cancellation of Effect of Heater-voltage Variation.—The fundamental effect of heater-voltage variation was explained in Sec. 11·6: A definite change of heater voltage is the equivalent of a definite change of the cathode potential relative to the other electrode potentials. For oxide-coated cathodes, a 10 per cent increase of heater voltage is the same as a cathode-potential decrease of about 100 mv, although this figure is subject to some variation from tube to tube and is larger and more erratic if the cathode current is very great. Thus the effect can be canceled either by an equal displacement in the opposite direction of the cathode potential or by approximately the same grid potential displacement in the same direction. If such a cancellation is not applied, the output of the amplifier will shift by its gain times this equivalent cathode-potential decrease. If more than one stage is involved, each will contribute to the heater effect by an amount that depends on the gain of the following stages, but, in general, the gain of the first stage will be great enough so that the heater effect in following stages is negligible in comparison.

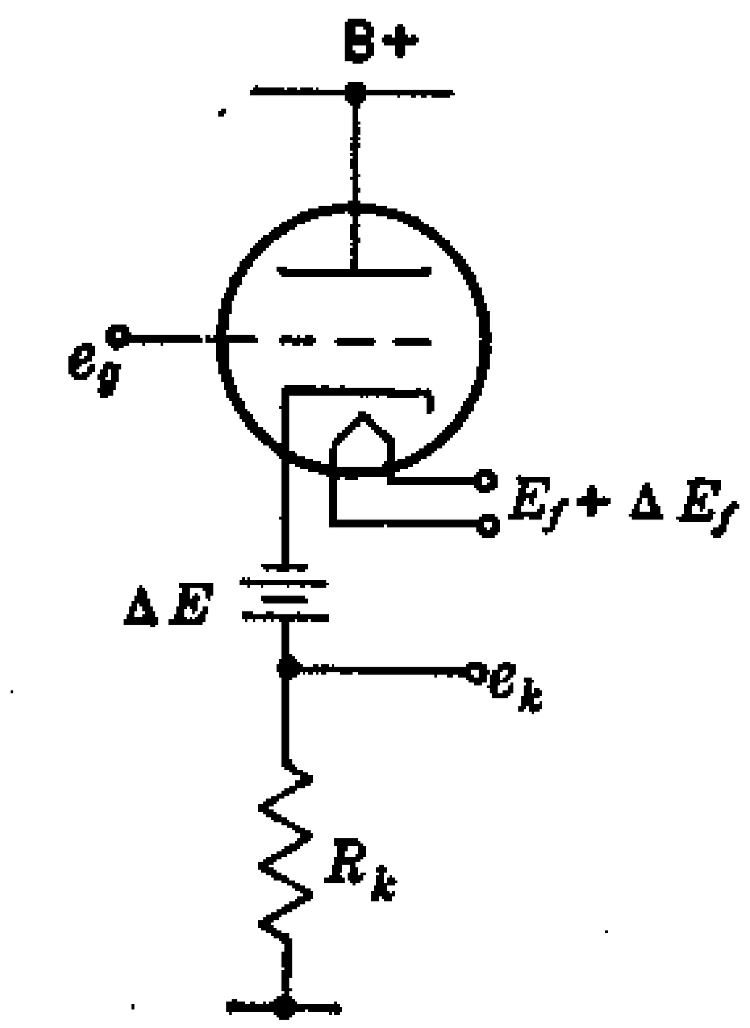

FIG. 11·40.—Heater-voltage variation effect on a cathode follower.

Negative feedback cannot reduce the amount of adjustment required at the input to cancel this effect. This fact is obvious in the case of a cathode follower. The effect of an increase of E_f is the same as the insertion of a low-voltage battery of zero resistance in series with the cathode, as shown in Fig. 11·40. If this were done, with no change in e_g, e_k would change by an amount almost equal to the battery voltage ΔE. To cancel this change, e_g must be lowered by just ΔE.

In a majority of applications, some method of canceling the effect of variations of heater voltage is necessary or advisable. (Of course, if the heater supply is well regulated, or if the resulting error is within allowable limits in a specific application, no cancellation will be needed.)

Diode Cancellation.—A comparison of Figs. 11·5 and 11·6 shows that the slopes of these curves for the diode are substantially the same as those for the triode. (At extremely low currents the triode curves become a little steeper, because incomplete grid control over the electrons is more noticeable with increasing temperature.) Thus the variation in a diode, whose heater is connected to the same source as that of the triode, may be used to offset the variation in the amplifying tubes.

One method of diode cancellation is shown by Fig. 11·41. The negative B voltage is large compared with the expected fluctuation at the cathode, so that the current in R may be considered constant (R is required to be large compared with the diode's variational resistance). It is desired that, with no change in e_g, i_p (and thus e_p) will remain constant. To permit this, a certain increase of E_f must be compensated for by a certain rise Δe_k at the cathode of the amplifier. But if the diode has the same cathode characteristic, this same increase Δe_k will occur at the diode, since the diode current is constant. If the two cathodes have different heater-voltage characteristics, the compensation will suffer accordingly. But generally the error will be less than one-tenth of that with no correction; e.g., the error due to a 10 per cent change of E_f, measured at e_g, will be less than 10 mv.

The error due to the fact that R is finite is not noticeable in comparison with that due to tube differences unless the voltage across R is less than about 10 volts. It can be shown that this error is only $r_d/(R + r_d)$ of the error with no compensation, where r_d is the diode variational resistance.

The value of R must be such that its current will be greater than the maximum in the amplifying tube; otherwise the diode will cut off, and the amplifier will cease to function as such. The cathode return of the amplifier is through the variational diode resistance r_d (in parallel with R, which should be large enough to be negligible in comparison). A loss of gain [Eq. (9)] results unless μr_d [or $(\partial e_{g_1}/\partial e_g) r_d$ if a pentode is used] is very small compared with R_p. It must be remembered, in selecting R, that r_d is a function of the diode current and may be determined from the slope of the diode current-voltage curve. At very low currents (e.g., below 0.2 or 0.3 ma, for a 6AL5) r_d becomes inversely proportional to the current, with a multiplier that is not a function of the diode type or even of the number of diodes in parallel (since increasing the number would decrease the current and increase the variational resistance of each). Thus, for all diodes with unipotential oxide-coated cathodes, the variational resistance at very low current approximates

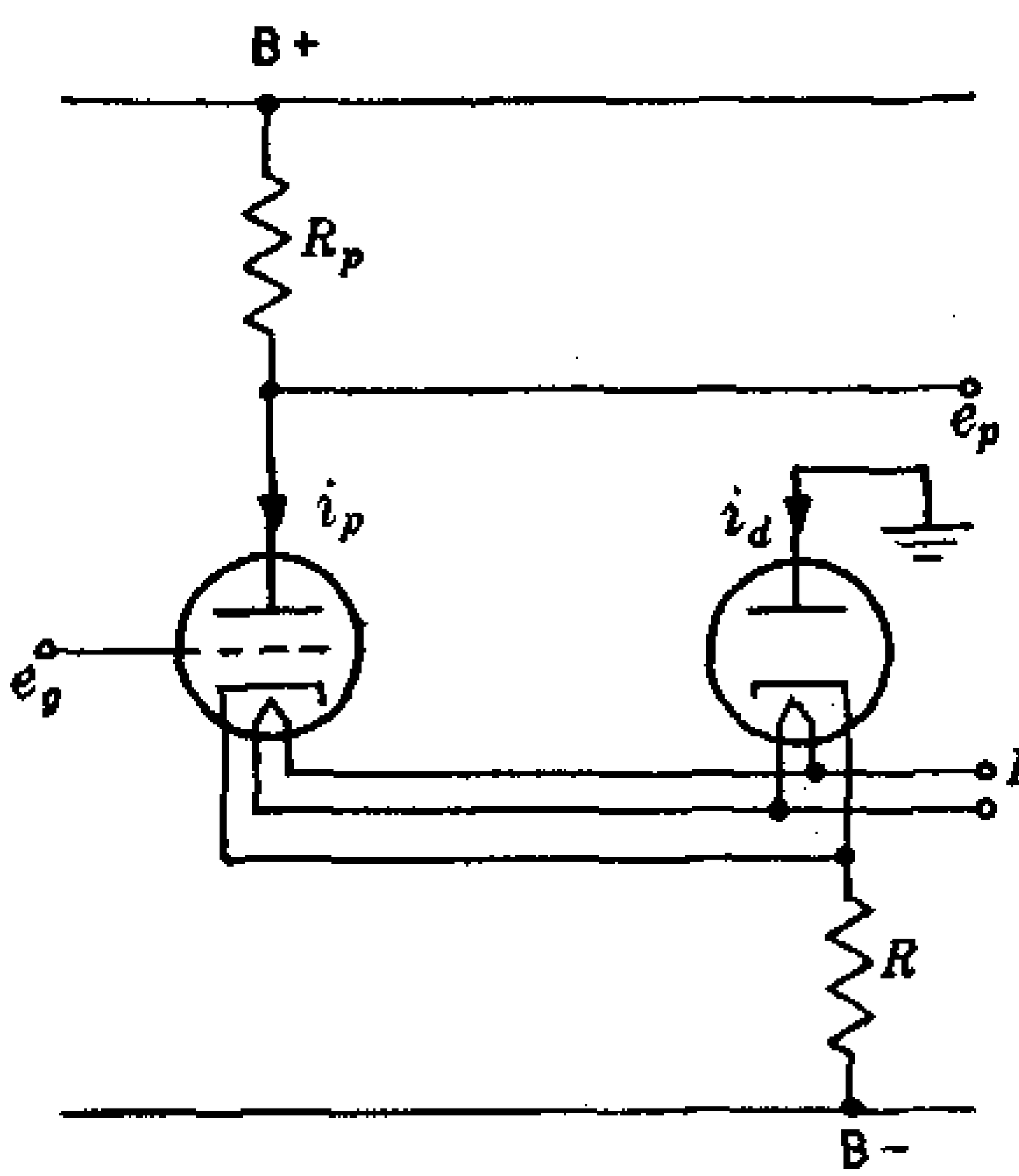

FIG. 11·41.—Balancing E_f variator by means of a diode.

$$r_d \approx \frac{0.09}{i_d} \quad \text{kilohms} \tag{83}$$

if i_d is in milliamperes.[1] This variation of the diode impedance might conceivably be used to linearize a portion of the amplifier output curve by varying the gain in a direction opposite to its natural curvature.

[1] This equation derives from the fact that the current at small values is exponential with respect to voltage: $i_d \approx i_0 e^{11,600 e_{pk}/T}$, where T is the absolute temperature of the cathode and is 1000° to 1100°K in this instance.

Tubes with diodes using the same cathode as the triode or pentode are especially useful in the circuit shown in Fig. 11·41, since the heater-voltage characteristics are bound to be nearly identical. For example, a sample of six 6SQ7 double-diode triodes was tested in the circuit of

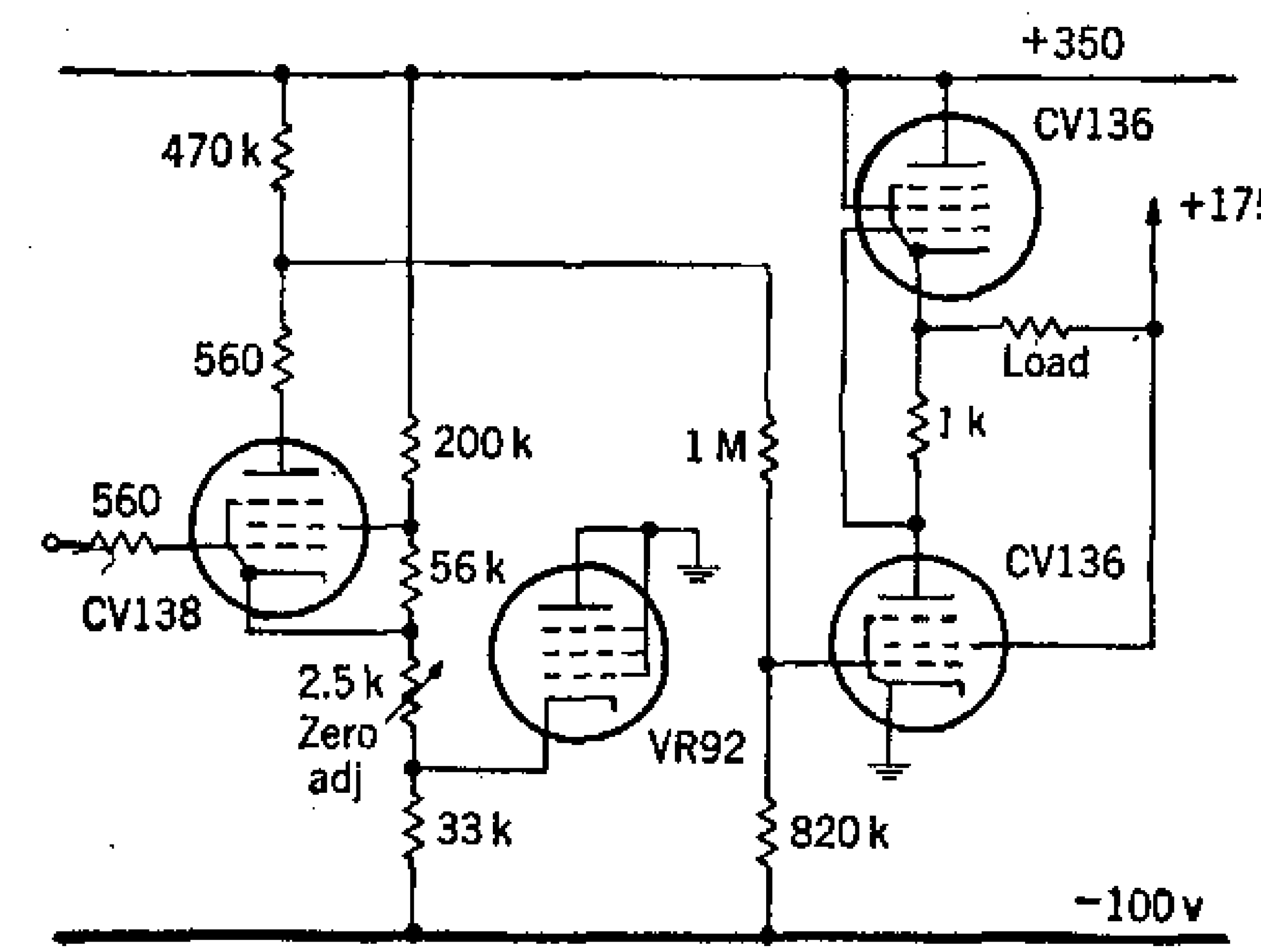

FIG. 11·42.—D-c servo amplifier.

Fig. 11·41, with the diode plates in parallel ($e_p = 150$ volts; $i_p = 0.1$ ma; $i_d = 0.08$ ma). The adjustments required at the grid by a 20 per cent heater-voltage change were −10, −5, 0, +2, +3, and +6 mv, whereas the adjustments required with the cathode grounded ranged from 195 to 212 mw. The drift due to aging of the heater or cathode may also be fairly well canceled in this way.

Figure 11·42 is the circuit diagram of a direct-coupled velocity servo amplifier,[1,2] which furnishes field excitation for a d-c servo motor. The input stage of this amplifier uses diode cancellation of heater-voltage variation. The zero-adjustment resistance and screen-grid bleeder do not impair this function because the currents here are substantially constant. The diode current is two or three times the cathode current or the pentode.

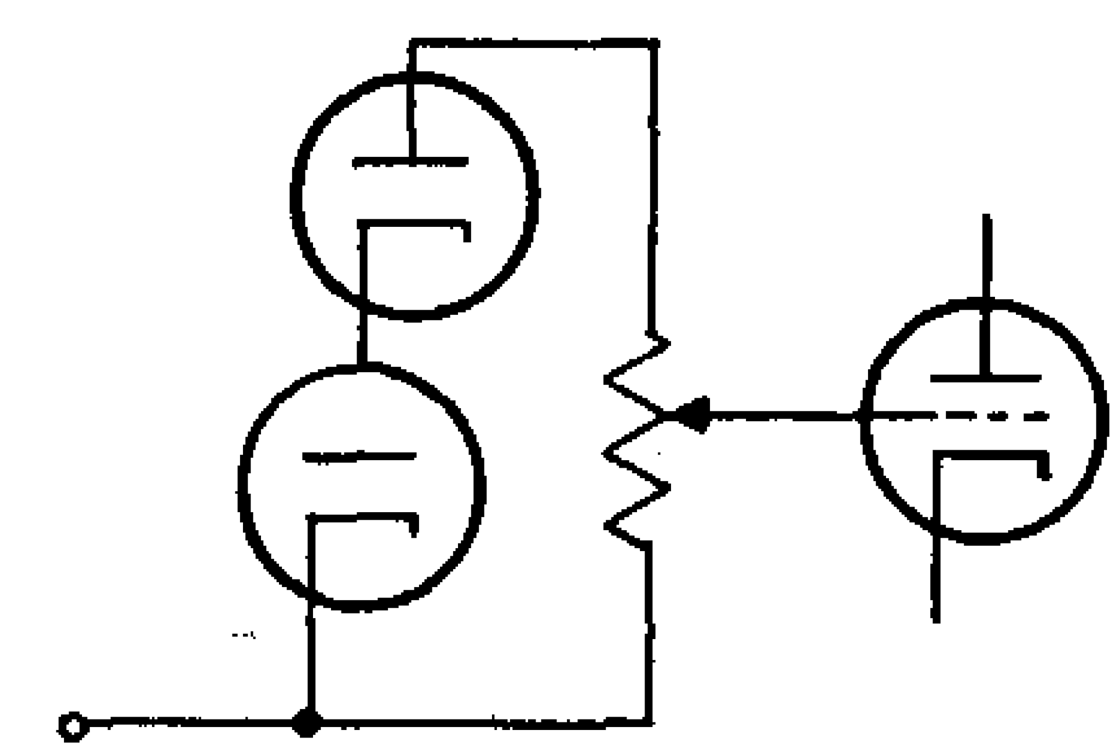
FIG. 11·43.—Diode heater-voltage cancellation at amplifier grid.

A method that requires two diodes but needs no negative supply and permits manual adjustment for differences in cathode characteristics is illustrated by Fig. 11·43. The circulating current due to initial electron velocity (Edison effect) in the diodes inserts a negative voltage in series with the grid. This voltage increases with increasing heater voltage, and its slope factor may be adjusted with the potentiometer. The resistance employed may be anywhere from 10,000 ohms to 1 megohm. A single diode is not potent enough because of the load imposed by the resistance.

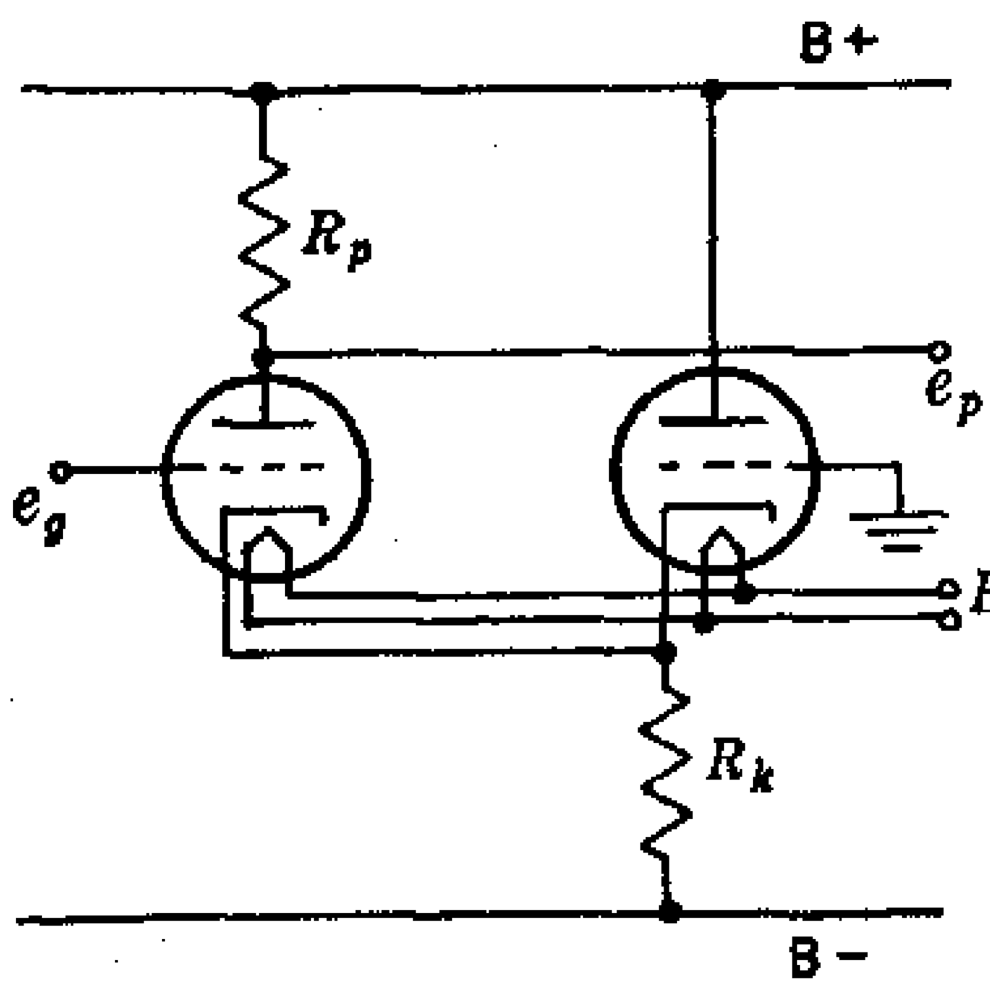

FIG. 11·44.—Balancing E_p variation with a cathode follower.

Cathode Follower and Differential Amplifier Cancellation.—In Fig. 11·44 the diode of Fig. 11·41 is replaced by a cathode follower. The same reasoning that explained the action in the case of the diode circuit is equally appropriate here, as are the remarks concerning the assignment of the value of R_k. For the cancellation to be as effective as possible, the $B-$ voltage must be large enough so that R_k may be very large compared with the reciprocal transconductance of the cathode follower. In the comparison with the diode circuit, this latter quantity corresponds to the diode variational resistance.

The grid of the cathode follower provides a convenient high-impedance point for zero adjustment for the amplifier. This and other features of the circuit have been discussed under differential amplifiers. If the load resistor is transferred to the right-hand triode in Fig. 11·44, to obtain an output of the same sense as the input, or if the circuit is made into a symmetrical differential amplifier, the cancellation of heater-voltage variation still obtains in the same way. Double triodes, such as the 6SL7 or 6SU7, are convenient for this type of application and give reasonable assurance of similar cathode characteristics. (The 6J6 should be very appropriate, since the same cathode is used for both triodes; but it seems subject to considerable drift because of its type of construction.)

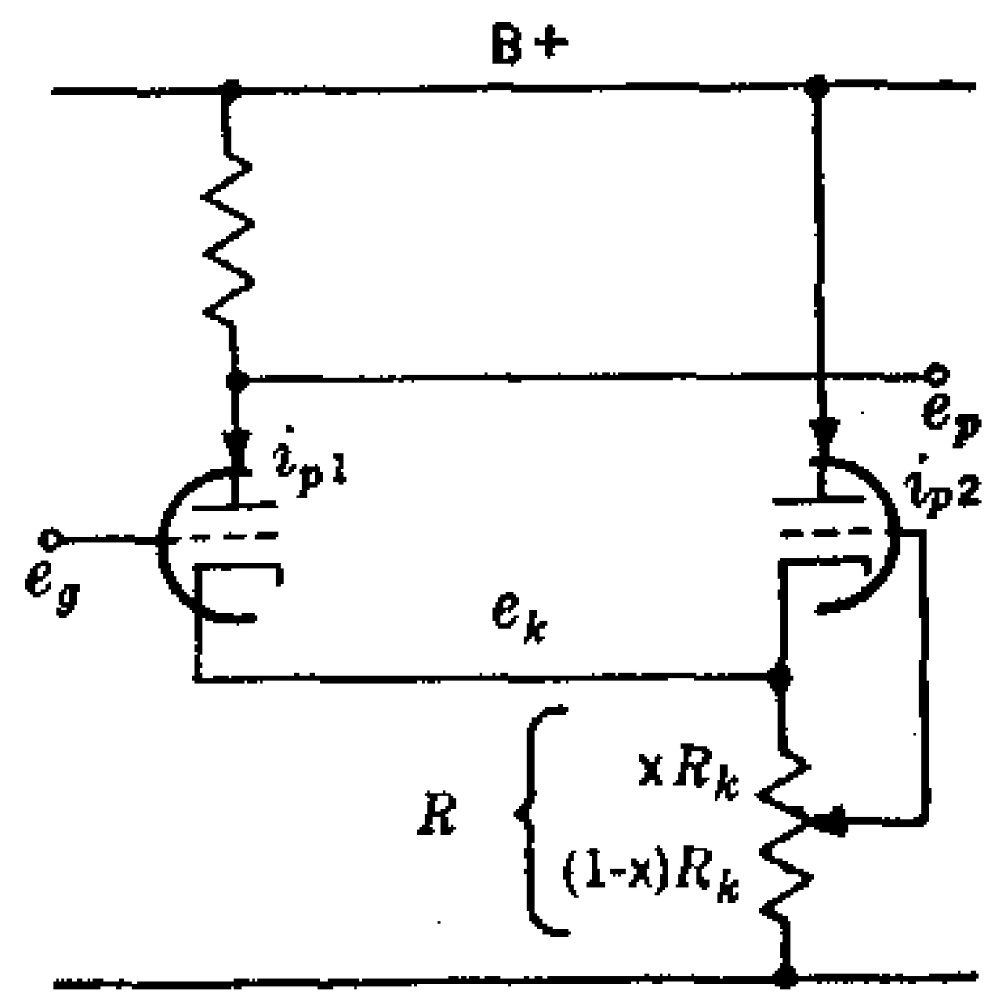

FIG. 11·45.—Balancing E_f variation with a self-biased cathode follower.

The circuit of Fig. 11·45, which employs a self-biasing cathode follower as a cathode return for the amplifier tube, permits adjustment for the inequality that may exist between the cathode characteristics of the two tubes. If i_{p1} is to be held constant in spite of any given change of heater voltage, e_k must be changed by some amount ΔE_1. Since i_{p1} is constant, the resulting change of current in R_k all occurs in the cathode-follower tube.

$$\Delta i_{p2} = \frac{\Delta E_1}{R_k}. \tag{84}$$

If x is the portion of R_k between the cathode and the grid tap, the change of grid bias is

$$\Delta e_{gk} = -x\,\Delta E_1 \tag{85}$$

But this tube behaves as if a small voltage $-\Delta E_2$ (which is approximately equal to $-\Delta E_1$) had been inserted in series with its cathode (as in Fig. 11·40). Thus the effective grid-to-cathode voltage has changed by the amount $\Delta E_2 - x\,\Delta E_1$, and effective plate-cathode voltage has changed by $\Delta E_2 - \Delta E_1$. So from Eq. (1),

$$\Delta i_{p2} = \frac{\Delta E_2 - \Delta E_1 + \mu_2(\Delta E_2 - x\,\Delta E_1)}{r_{p2}} = \frac{(\mu_2 + 1)\,\Delta E_2 - (x\mu_2 + 1)\,\Delta E_1}{r_{p2}}. \tag{86}$$

Combining (86) and (84),

$$R_k = \frac{r_{p2}}{\mu_2\left(\frac{\Delta E_2}{\Delta E_1} - x\right) + \frac{\Delta E_2}{\Delta E_1} - 1}. \tag{87}$$

Or, approximately,

$$\left(\frac{\Delta E_2}{\Delta E_1} - x\right) R_k \approx \frac{r_{p2}}{\mu_2}. \tag{88}$$

If the two cathodes have exactly equal characteristics, so that $\Delta E_1 = \Delta E_2$, the portion of R_k below the grid tap is simply

$$(1 - x)R_k = \frac{r_{p2}}{\mu_2}, \tag{89}$$

where r_{p2} is the plate resistance of the cathode-follower tube, and if there is a resistor inserted in series with this plate, it is included in r_{p2}.

The amplifier tube cathode is, in effect, connected to a voltage $E_{pp}/(\mu + 1)$ through a resistance $r_{p2}/(\mu + 1)$, regardless of the value of the portion of R_k above the grid tap (although this value does determine the cathode-follower plate current, which in turn affects r_{p2}). Thus the amplifier tube will be just about cut off at zero grid voltage, and it will have a gain [Eq. (9)] of $\mu R_p/(R_p + 2r_p)$. The value of R_k is not critical but should be between two and eight times r_p/μ. The triodes should be aged for a while and cycled several times through the extremes of heater voltage before the adjustment is made.

Cancellation by a Series Triode.—In the circuit of Fig. 11·46, the plate resistor for the amplifier (lower) tube comprises a circuit resembling the constant-current device of Fig. 11·18 except that the voltage E is omitted. Thus this is not a constant-current device but is simply the equivalent of a resistor whose upper end is attached to E_{pp} and whose value is $r_p + (\mu + 1)R$ [Eq. (23)]. Thus from Eq. (7), if the two triodes are similar,

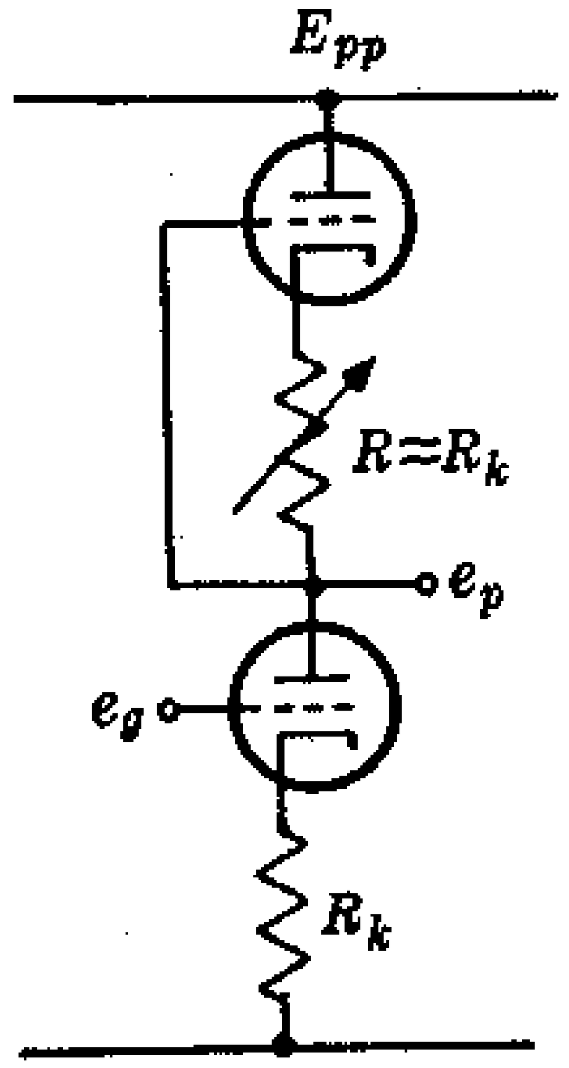

FIG. 11·46.—Cancellation of E_f variation with a series triode.

$$e_p = \frac{\dfrac{r_p + (\mu + 1)R_k}{r_p + (\mu + 1)R} E_{pp} - \mu e_g}{1 + \dfrac{r_p + (\mu + 1)R_k}{r_p + (\mu + 1)R}}. \tag{90}$$

If the two cathodes respond equally to a change of heater voltage, R should equal R_k for cancellation of the effect.[1] With e_g fixed, a given increase of E_h causes a certain small increase of current; but since this increase is the same in both tubes, the grid biases change identically. Therefore, the two plate-to-cathode voltages suffer no change, and e_p remains constant. If $R_k = R$, Eq. (90) becomes

$$e_p = \frac{E_{pp}}{2} - \frac{\mu e_g}{2}. \tag{91}$$

Thus, when e_g is zero, the output voltage is half the plate-supply potential (as is already apparent by symmetry), and the gain is $\mu/2$. The output voltage is linear with respect to e_g because r_p does not appear in Eq. (91). Of course, the two r_p's were assumed equal, but this assumption is not far in error, as the currents in the two tubes are equal.

The output impedance, from Eq. (11), is (if $R_k = R$)

$$Z_p = \frac{r_p + (\mu + 1)R_k}{2}. \tag{92}$$

If the two cathodes have different heater-voltage characteristics, either R or R_k may be adjusted until cancellation is obtained. This adjustment will not usually change the gain from $\mu/2$ by more than 5 per cent.

Cancellation by Means of a D-c Potential Proportional to Heater Voltage. If the d-c load on the power supply is fairly constant, its unregulated output will vary in proportion with the a-c line voltage and therefore with E_h. Thus this output may be employed in some way to offset the effect on the cathode of the variations of E_h. For example, in Fig. 11·47, if E_h increases, e_k will rise a small amount, which, if the adjustment is right, will keep the plate current constant with no change of e_g. If the cathode is on the bleeder at a point 1 volt from ground, then regardless of the total voltage on the bleeder, e_k will rise 0.1 volt with a 10 per cent rise in the unregulated source. This rise is just about the required amount. Another resistor, inserted in series with the cathode for degeneration or zero adjustment, will not affect the cancellation.

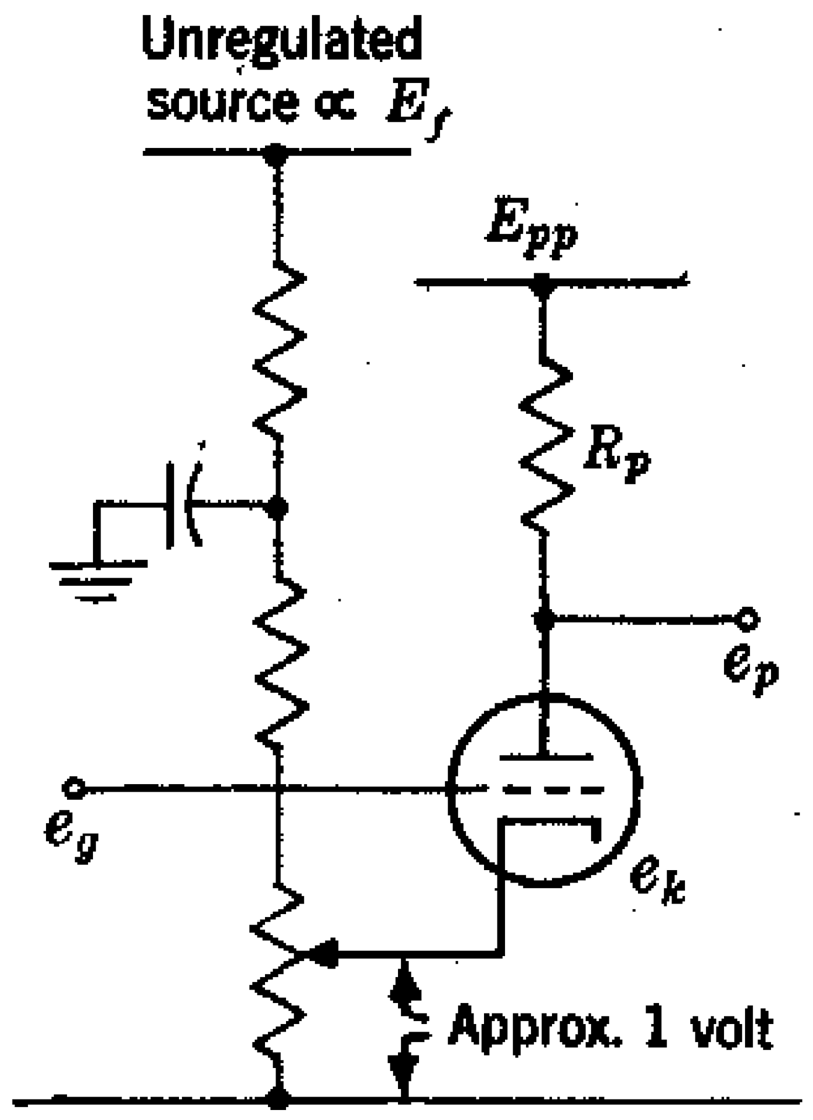

FIG. 11·47.—Variation of e_k with e_p.

If the source is not well filtered, filtering may be done at the cathode. But a condenser of a given capacity is more effective if placed as shown, about midway on the bleeder.

Rapid line fluctuations will produce errors whose nature and duration is revealed by Fig. 11·8. Another disadvantage of this method is that the correction is linear with respect to E_h whereas the effect itself is curved (Fig. 11·6); therefore, the cancellation is of limited range.

Cancellation by Special Connections of Multigrid Tubes.—The circuit of Fig. 11·48, wherein the second grid of a tetrode or pentode is used as the control grid for the plate current, is actually very much like the circuit of Fig. 11·41. Here the first grid instead of a separate diode plate is grounded (or otherwise fixed) and acts as a diode plate to furnish a large part of the cathode current. If the cathode temperature is increased, its potential will rise; and since i_g is practically constant (e_p is assumed to be held constant, and the variation of e_k is small compared with the drop across R_k), its rise will be just enough to neutralize the effect on the other electrodes in the tube. The advantage over the diode-cancellation method of Fig. 11·41 is that the same cathode area is used in the cancellation as in the amplification. The drift that is caused by change of average initial electron velocity with time is also canceled in the same way as that due to temperature change.

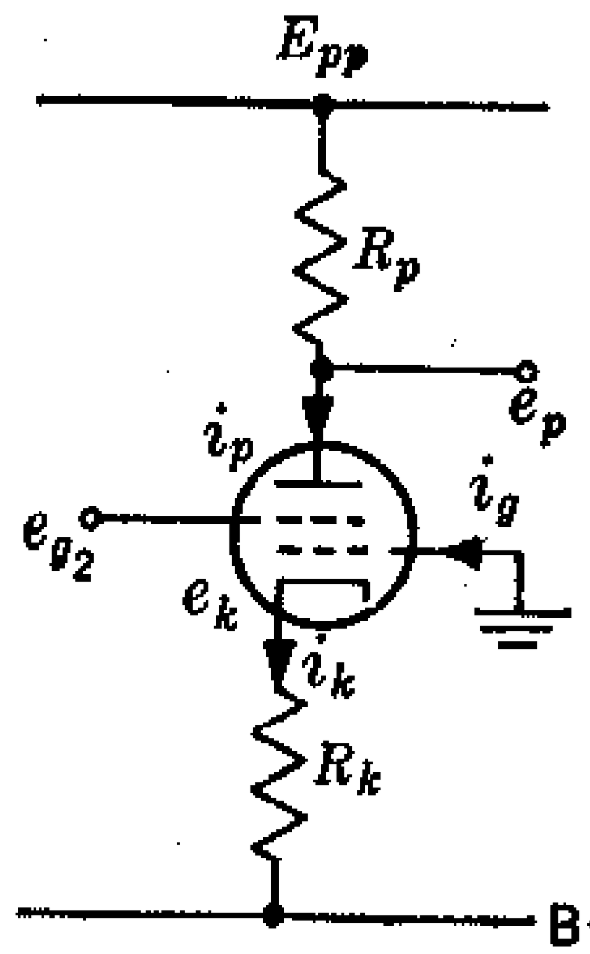

FIG. 11·48.—First grid as diode plate.

Ordinary pentodes are generally unsatisfactory in this circuit because the screen grid has so much control over the plate current that the latter is completely cut off (except of extremely high plate voltage) when the screen-grid potential is lowered far enough so that its own current is negligible. The suppressor grid, on the other hand, has very little control. In a tetrode like the 6V6, however, the effect of the screen grid is between these two extremes. Figure 11·49 shows the plate characteristics of a 6V6 in the circuit of Fig. 11·48, with a negative supply of 45 volts, and R_k = 90,000 ohms, so that i_k is just about $\frac{1}{2}$ ma. The curves are similar to those of an ordinary triode with a μ of about 40,

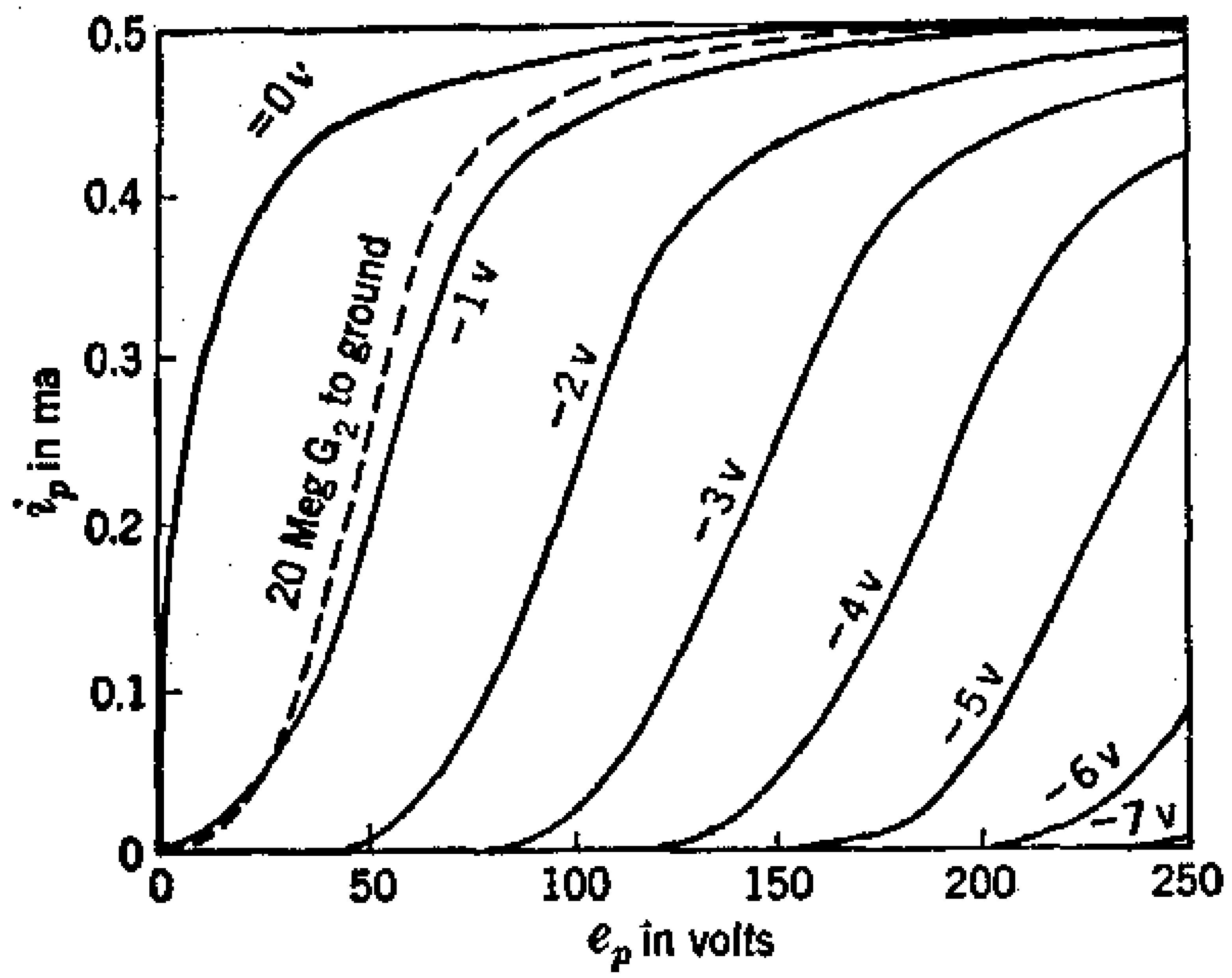

FIG. 11·49.—Plate characteristics of a 6V6 triode.

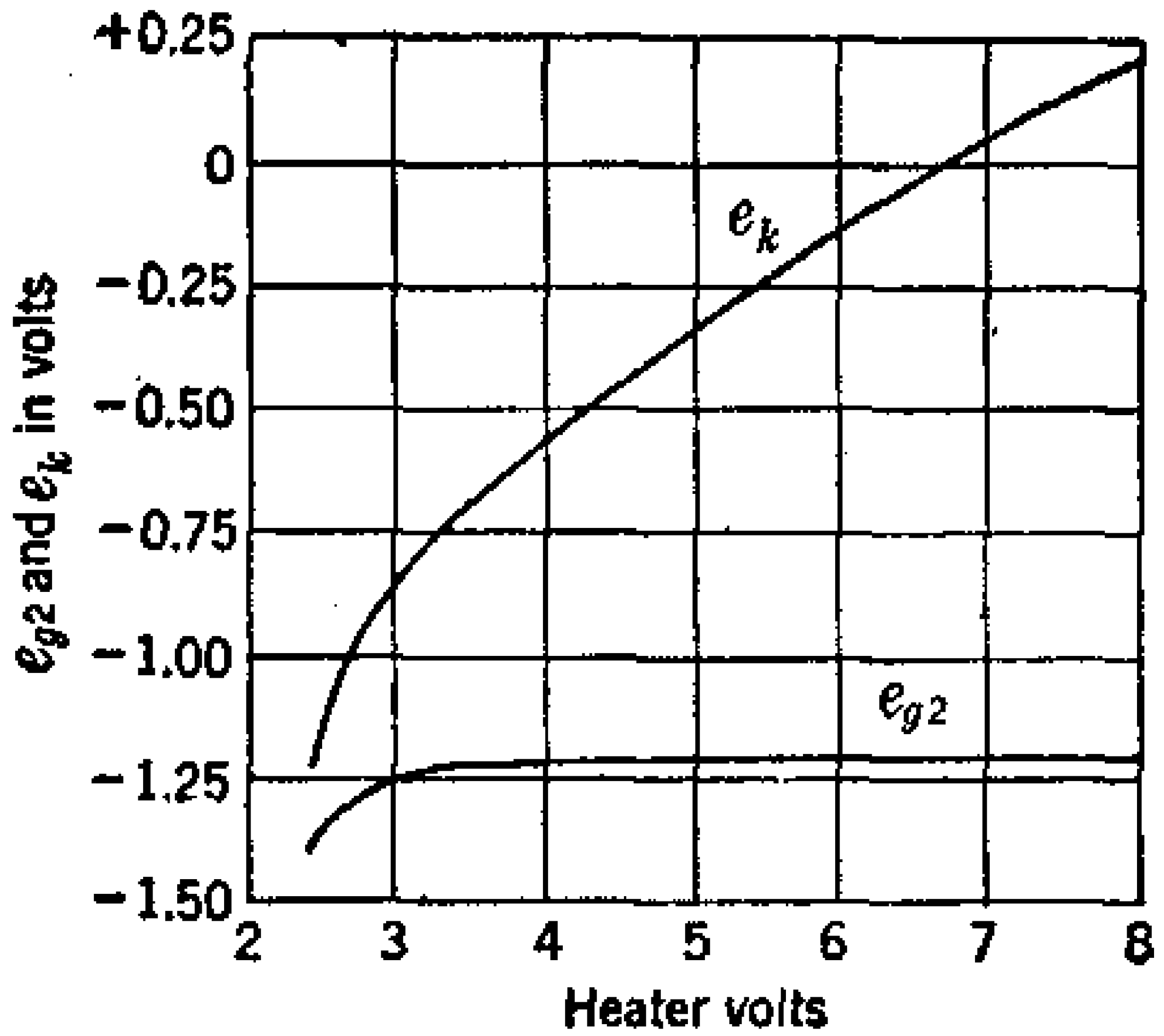

FIG. 11·50.—Effect of heater voltage on e_{g2} and e_k.

except that the current has an upper limit of $\frac{1}{2}$ ma. Figure 11·50 shows the effect of heater voltage on e_k and e_{g2}, the latter being adjusted so as to maintain constant plate current. When i_p and i_g are about the same, e_{g2} is fairly constant over a large temperature range.

The 6AS6 miniature pentode has a specially designed suppressor grid that has unusually effective control over the plate current. This tube may be used in a manner similar to the one in the circuit of Fig. 11·48, but with the suppressor as the control grid and the screen grid at some positive potential. This method of operation has the advantage of producing plate characteristics more like those of a pentode, but the cancellation of heater-voltage effect seems to be less effective over a wide operating range than in the case of the tetrode. A pentagrid converter tube might also be applicable in the same way as the 6AS6. More research is indicated in this field, with a possible objective of developing a suitable tube for the purpose.

11·13. The Use of Feedback in D-c Amplifiers. *Conductive Negative Feedback.*—Conductive, or direct-coupled, negative feedback can reduce the dependence of gain upon output d-c voltage or current (i.e., nonlinearity or amplitude distortion) and tube characteristics. It cannot reduce the effect of zero drift or displacement of a given amplifier (as referred to the input terminals) no matter whether it is due to heater-voltage variation, tube aging or replacement, or microphonics. This fact was illustrated in Sec. 11·12 for the case of a cathode follower, in which the zero drift referred to the input terminals was found to equal the tube drift in terms of the shift of grid-to-cathode potential required to maintain constant current. In some instances, as in the arrangement of Fig. 11·51, negative feedback can actually increase the effect of drift as seen at the input terminals. In Fig. 11·51, if the amplifier drifts so that a given change Δe_g is required to hold e_0 constant, the change required at the input terminals is $\Delta e_1 = \Delta e_g(R_1 + R_2)/R_2$.

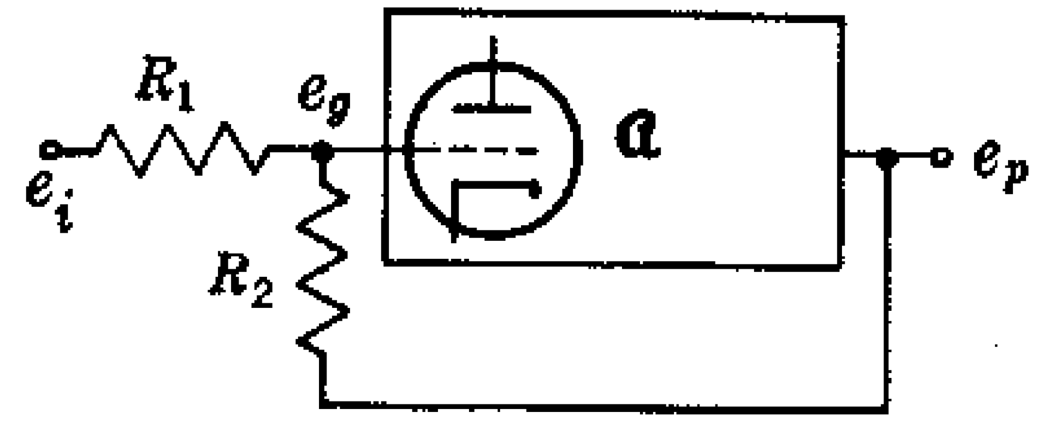

FIG. 11·51.—Direct feedback using resistance adding.

However, if a given over-all gain is required, negative feedback can reduce the effect on zero displacement of changes in certain parts of an amplifier by permitting the use of more amplification ahead of these parts. This is often of great value, since a power-output stage is much more susceptible to shifts caused by variations of load, supply voltage, heater voltage (because of greater tube currents), etc., than is a voltage amplifier with very low power output.

The common form of the feedback equation is

$$\mathcal{G} = \frac{\mathfrak{A}}{1 - \mathfrak{A}\beta}. \tag{93}$$

where $\mathfrak{A}$ is the gain without feedback, $\mathcal{G}$ is the gain with feedback, and β is the fraction of output voltage that is added to the input voltage. This equation is based on the assumption of simple addition involving no attenuation of the input voltage and no disturbance of the amplifier parameters by the feedback action. In Fig. 11·51 the resistance addition of input voltage to a fraction of the output voltage attenuates the input voltage by the factor $R_2/(R_1 + R_2)$, and Eq. (93) therefore becomes[1]

$$\mathcal{G} = \frac{R_2}{R_1 + R_2} \frac{\mathcal{A}}{1 - \frac{\mathcal{A}R_1}{R_1 + R_2}}. \tag{94}$$

For negative feedback the sense of the amplifier must be such that $\mathcal{A}$ is negative.

Perhaps the most common form of negative feedback is through the cathode. Some of the aspects of this method, as applied to single-stage voltage amplifiers, were covered in Sec. 11·7. Equation (93) cannot be applied directly because the operation of the amplifier itself is somewhat affected by this type of feedback. From Eq. (9) it is seen that the insertion of a cathode resistor R_k gives the triode amplifier a gain of

$$\mathcal{G} = -\frac{\mu}{1 + \frac{r_p}{R_p} + (\mu + 1)\frac{R_k}{R_p}}, \tag{95}$$

whereas the gain without this degeneration is simply

$$\mathcal{A} = -\frac{\mu}{1 + \frac{r_p}{R_p}}. \tag{96}$$

Combination of these equations to obtain a form like Eq. (93) results in

$$\mathcal{G} = \frac{\mathcal{A}}{1 - \mathcal{A}\frac{\mu + 1}{\mu}\frac{R_k}{R_p}}. \tag{97}$$

Thus the feedback factor is, in effect,

$$\beta = \frac{\mu + 1}{\mu}\frac{R_k}{R_p}, \tag{98}$$

which is rather different from the common form in that it is a function of a characteristic of the triode, although a given percentage change of μ will cause only $1/(\mu + 1)$ times as much percentage change in β.

If β is constant in Eq. (93), differentiation shows that a change $\Delta\mathcal{A}$ of the gain of the amplifier will produce a change with the feedback of

$$\Delta\mathcal{G} = \frac{\Delta\mathcal{A}}{(1 - \mathcal{A}\beta)^2} \tag{99}$$

$$= \frac{\mathcal{G}^2}{\mathcal{A}^2}\Delta\mathcal{A}, \tag{100}$$

or

$$\frac{\Delta\mathcal{G}}{\mathcal{G}} = \frac{\mathcal{G}}{\mathcal{A}}\frac{\Delta\mathcal{A}}{\mathcal{A}}. \tag{101}$$

This equation indicates that the introduction of negative feedback reduces fractional deviations from constancy of gain in the same ratio as it reduces the gain. Of course, if cathode feedback is employed so that Eq. (98) obtains, Eq. (101) is not complete for any change of μ but is complete for a change or r_p. This agrees with the conclusions drawn from Eq. (10) in Sec. 11·7.

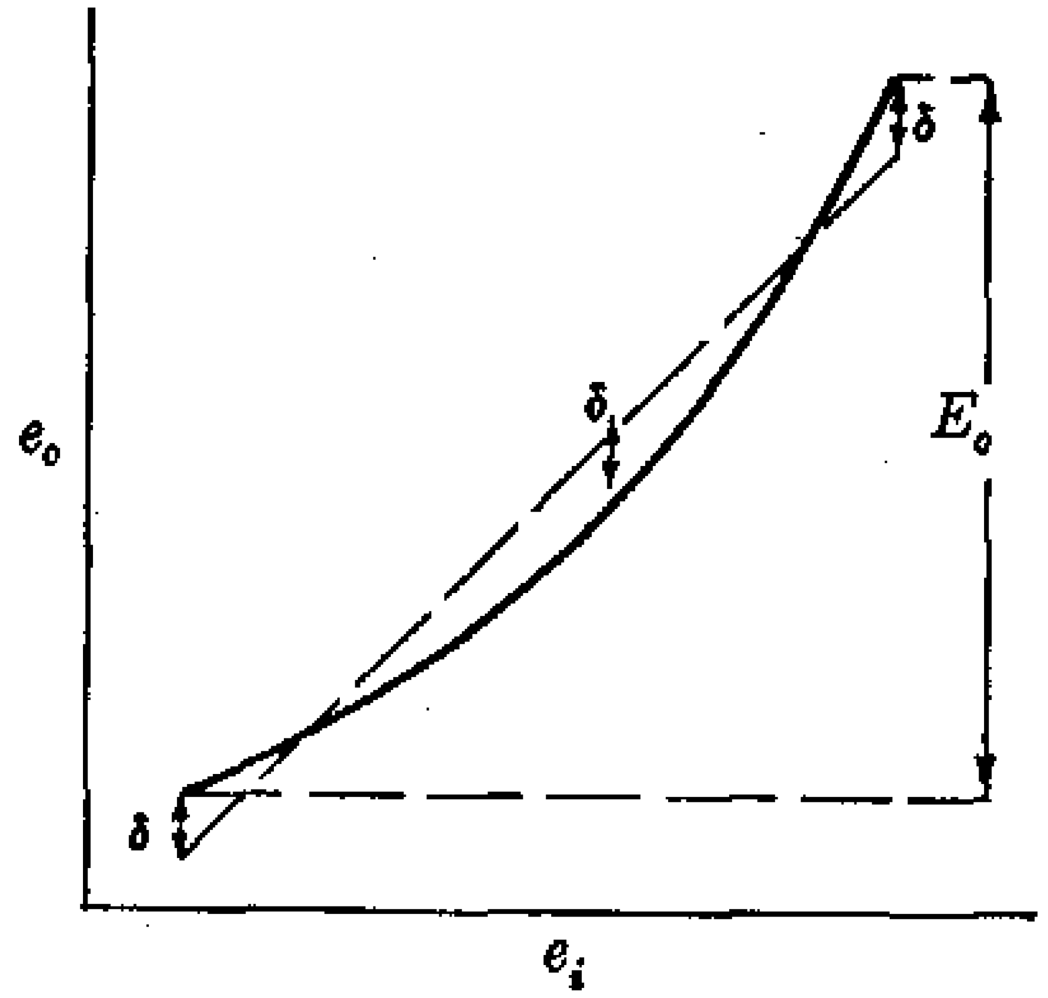

FIG. 11·52.—Relation of nonlinearity to change of gain.

Although negative feedback tends to stabilize the gain with respect both to changes of circuit parameters at any given value of output voltage and to changes of the output voltage itself, these two factors may be considered separately to some extent. One or the other of them may be of greater importance in a given application, and this may affect the nature of the feedback employed.

Variation of gain with respect to output voltage results in nonlinearity. The common way of expressing nonlinearity is in terms of maximum deviation of the curve of output voltage plotted against input voltage from the best linear approximation thereto, given as a percentage of the total working range of output voltage. This quantity may be related to the change of gain over the length of the curve if the nature of this change is known. For example, if the gain, which is the slope of the curve, changes uniformly with respect to input, the curve is part of a parabola, as shown in Fig. 11·52. The dashed line is the best linear approximation, and δ is the maximum deviation. Thus the nonlinearity is expressed as δ/E_0, where E_0 is the range of output voltage.

[1] In the resistance addition of e_0 and e_i to obtain e_g,

$$\Delta e_g = \frac{R_2\,\Delta e_i + R_1\,\Delta e_0}{R_1 + R_2}. \tag{i}$$

But the amplifier gain is

$$\mathcal{A} = \frac{\Delta e_0}{\Delta e_g}, \tag{ii}$$

and over-all gain with the feedback is

$$\mathcal{G} = \frac{\Delta e_0}{\Delta e_i}. \tag{iii}$$

Combination of Eqs. (i), (ii), and (iii) gives Eq. (94).

A geometrical calculation reveals that the nonlinearity is

$$\frac{\delta}{E_0} = \frac{1}{16}\frac{\Delta \mathcal{G}}{\mathcal{G}_0}, \tag{102}$$

where $\mathcal{G}_0$ is the slope of the dashed line and $\Delta\mathcal{G}$ is the difference between the slopes at the extremes of the curve. In most cases where the curvature is greater near one end, the fractional nonlinearity is an even smaller fraction of the fractional maximum change of gain. In any event, Eqs. (102) and (101) show that the fractional nonlinearity is decreased by negative feedback in the same ratio as is the gain.

If there is more than one stage in an amplifier, the question may arise as to whether the negative feedback should be over all or over the individual stages. The total amplification is

$$\mathcal{A} = \mathcal{A}_1\mathcal{A}_2\mathcal{A}_3, \cdots, \tag{103}$$

where $\mathcal{A}_1$, $\mathcal{A}_2$, etc., are the amplifications of the stages. If each has its own deviation in gain, which is assumed to be small, then differentiation shows that the fractional deviation of $\mathcal{A}$ is simply the algebraic sum of the individual fractional deviations:

$$\frac{\Delta\mathcal{A}}{\mathcal{A}} = \frac{\Delta\mathcal{A}_1}{\mathcal{A}_1} + \frac{\Delta\mathcal{A}_2}{\mathcal{A}_2} + \frac{\Delta\mathcal{A}_3}{\mathcal{A}_3} \cdots. \tag{104}$$

If over-all feedback, which reduces the gain to $\mathcal{G}$, is employed, Eq. (101) gives the deviation of gain as

$$\frac{\Delta\mathcal{G}}{\mathcal{G}} = \frac{\mathcal{G}}{\mathcal{A}}\left(\frac{\Delta\mathcal{A}_1}{\mathcal{A}_1} + \frac{\Delta\mathcal{A}_2}{\mathcal{A}_2} + \frac{\Delta\mathcal{A}_3}{\mathcal{A}_3} + \cdots +\right). \tag{105}$$

On the other hand, if feedback is on an individual basis so that each stage has its gain reduced from $\mathcal{A}_1$ to $\mathcal{G}_1$, etc., then the linearity of each stage is improved according to Eq. (100):

$$\frac{\Delta\mathcal{G}_1}{\mathcal{G}_1} = \frac{\mathcal{G}_1}{\mathcal{A}_1}\frac{\Delta\mathcal{A}_1}{\mathcal{A}_1}, \text{ etc.}$$

Thus, as in Eq. (104), the over-all deviation in this case is

$$\frac{\Delta\mathcal{G}}{\mathcal{G}} = \frac{\mathcal{G}_1}{\mathcal{A}_1}\frac{\Delta\mathcal{A}_1}{\mathcal{A}_1} + \frac{\mathcal{G}_2}{\mathcal{A}_2}\frac{\Delta\mathcal{A}_2}{\mathcal{A}_2} + \frac{\mathcal{G}_3}{\mathcal{A}_3}\frac{\Delta\mathcal{A}_3}{\mathcal{A}_3} + \cdots +. \tag{106}$$

If the over-all gain is to be the same in both cases, i.e., if $\mathcal{G} = \mathcal{G}_1\mathcal{G}_2\mathcal{G}_3, \cdots$, then Eq. (106) yields a larger result than Eq. (105), since $\mathcal{G}/\mathcal{A}$ will be much smaller than any of the individual ratios $\mathcal{G}_1/\mathcal{A}_1$, etc. Thus the over-all degeneration is by far the more effective in linearizing and stabilizing the gain.

In many cases, it is more convenient to apply the degeneration to individual stages, especially in view of the simplicity of the cathode-resistor method or the fact that proper phasing of the feedback may not otherwise be possible. The question then arises as to what the apportionment of degeneration should be among the stages. This will depend to some extent on whether linearity or stability of gain with respect to circuit elements is desired. If linearity is of greater importance, most of the degeneration should be placed where there is the greatest range of voltage (or current) to be encountered because the fractional nonlinearity tends to be greatest there. But if it is desirable to have constancy of gain with respect to tube parameters, etc., each stage should be stabilized regardless of the magnitude of its output voltage swing.

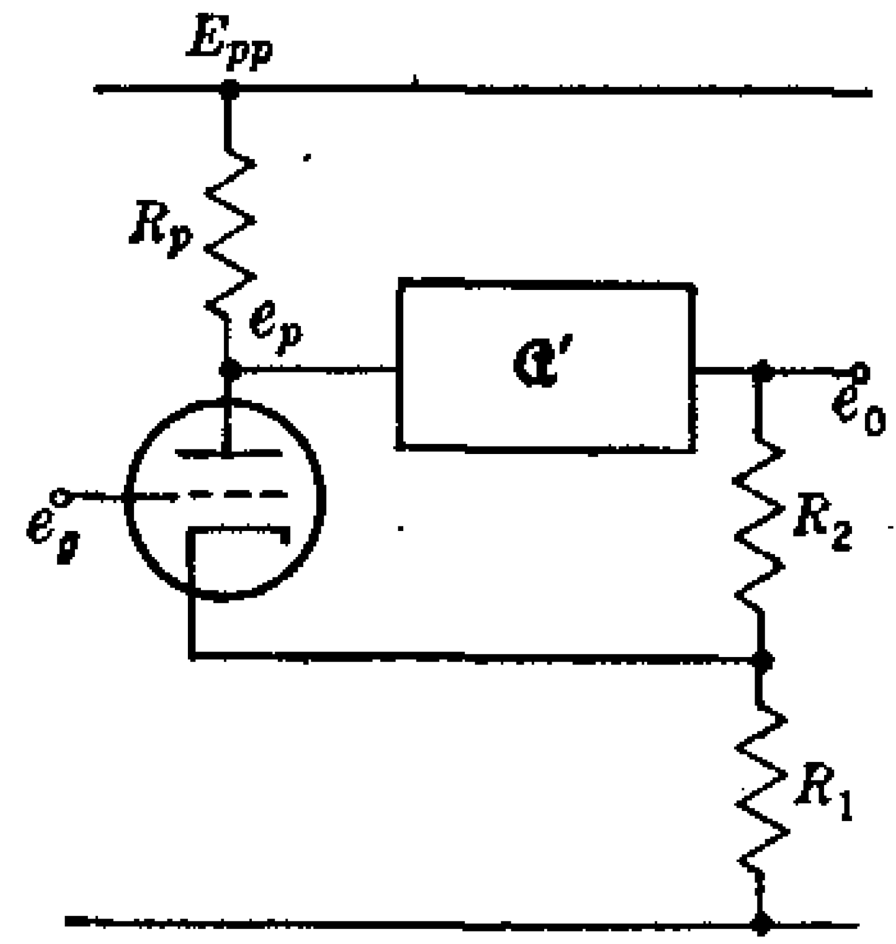

FIG. 11·53.—Over-all cathode feedback.

Over-all negative feedback to the cathode of the first stage may be applied as shown in Fig. 11·53, where the rest of the amplifier beyond the first stage is generalized and assigned the gain $\mathcal{A}'$. From Eq. (13) the portion of the output fed back to the cathode is found to be

$$k = -\frac{R_1}{R_1 + R_2 + (\mu + 1)\dfrac{R_1R_2}{r_p + R_p}}$$

This attenuating network also has the effect of inserting in series with the cathode of the first tube a resistance of value

$$R_k = \frac{R_1R_2}{R_1 + R_2}.$$

As explained in Sec. 11·7 this has the effect of increasing the plate resistance of the triode by $(\mu + 1)R_k$. Thus,

$$\Delta e_p = \frac{(\mu + 1)k\,\Delta e_0 - \mu\,\Delta e_g}{1 + \dfrac{r_p + (\mu + 1)R_k}{R_p}}. \tag{107}$$

But $\Delta e_p = \Delta e_0/\mathcal{A}'$, and the over-all gain may therefore be found as

$$\mathcal{G} = \frac{\Delta e_0}{\Delta e_g} = \frac{\mu}{(\mu + 1)k - \dfrac{1}{\mathcal{A}'}\left[1 + \dfrac{r_p + (\mu + 1)R_k}{R_p}\right]}. \tag{108}$$

But the gain of the first stage alone (including the effect of R_k) is

$$\mathfrak{A}'' = -\frac{\mu}{1 + \dfrac{r_p + (\mu+1)R_k}{R_p}}. \tag{109}$$

Substituting this in Eq. (105),

$$\mathfrak{G} = \frac{\mu}{k(\mu+1) + \dfrac{\mu}{\mathfrak{A}'\mathfrak{A}''}}. \tag{110}$$

The total amplification without the feedback (but with R_k) is $\mathfrak{A} = \mathfrak{A}'\mathfrak{A}''$, so that the gain with feedback, in the form of Eq. (93), is

$$\mathfrak{G} = \frac{\mathfrak{A}}{1 + \dfrac{\mu+1}{\mu}k\mathfrak{A}}, \tag{111}$$

which is very much like the single-tube case of Eq. (97). The feedback factor here is

$$\beta = -\frac{\mu+1}{\mu}k \tag{112}$$

The negative sign shows that for negative feedback, $\mathfrak{A}$ must be positive, i.e., $\mathfrak{A}'$ must be negative.

The possible variation of β can cause variation of gain with this type of feedback, no matter how large $\mathfrak{A}$ is. Differentiation of Eq. (111) shows that if μ and $\mathfrak{A}$ are both subject to small changes, the resulting change of gain is

$$\frac{\Delta\mathfrak{G}}{\mathfrak{G}} = \frac{\mathfrak{G}}{\mathfrak{A}}\frac{\Delta\mathfrak{A}}{\mathfrak{A}} + \frac{k\mathfrak{G}}{\mu}\frac{\Delta\mu}{\mu}. \tag{113}$$

If $\mathfrak{A}$ is very large compared with $\mathfrak{G}$, Eqs. (111) and (113) become

$$\mathfrak{G} \approx \frac{\mu}{\mu+1}\frac{1}{k} \tag{114}$$

and

$$\frac{\Delta\mathfrak{G}}{\mathfrak{G}} \approx \frac{1}{\mu+1}\frac{\Delta\mu}{\mu}. \tag{115}$$

Thus the limiting factor in the degree of linearization and gain stabilization achieved is expressed by Eq. (115). If a considerable range is covered by the input voltage, the plate-to-cathode voltage will vary, since the cathode voltage follows the input voltage. Because of the concommitant variation in μ, nonlinearity may result, but it can be kept to a minimum by keeping the plate current in the tube practically constant, i.e., by having $\mathfrak{A}'$ large enough so that the variation of e_p will be small compared with the drop across R_p. Obviously a high-μ tube is preferable in the input position. As an example of the effect of tube replacement, μ for 6SL7's at a plate current of 0.2 ma varies approximately from 61 to 73 for different tubes. Therefore, the maximum change of gain to be expected from Eq. (115) would be 0.26 per cent.

If a pentode with its screen at a fixed potential is employed in the input stage, the preceding analyses and formulas still apply, except that μ must be replaced by $\partial e_{sk}/\partial e_{gk}$ and r_p does not appear. The screen may be kept at a constant potential above the cathode by means of an auxiliary network, as in Fig. 11·54 where $R_4/R_3 = R_2/R_1$. Thus the expression on the right-hand side of Eq. (115) is very nearly equal to zero. If the variation in pentode plate current is small, and if the screen potential never swings much above the plate potential, the screen current will be reasonably constant, so that R_3 and R_1 may be rather large. A simpler arrangement than that of Fig. 11·54 puts the screen on a bleeder from cathode to $B+$. The operation is then somewhere between the two conditions described.

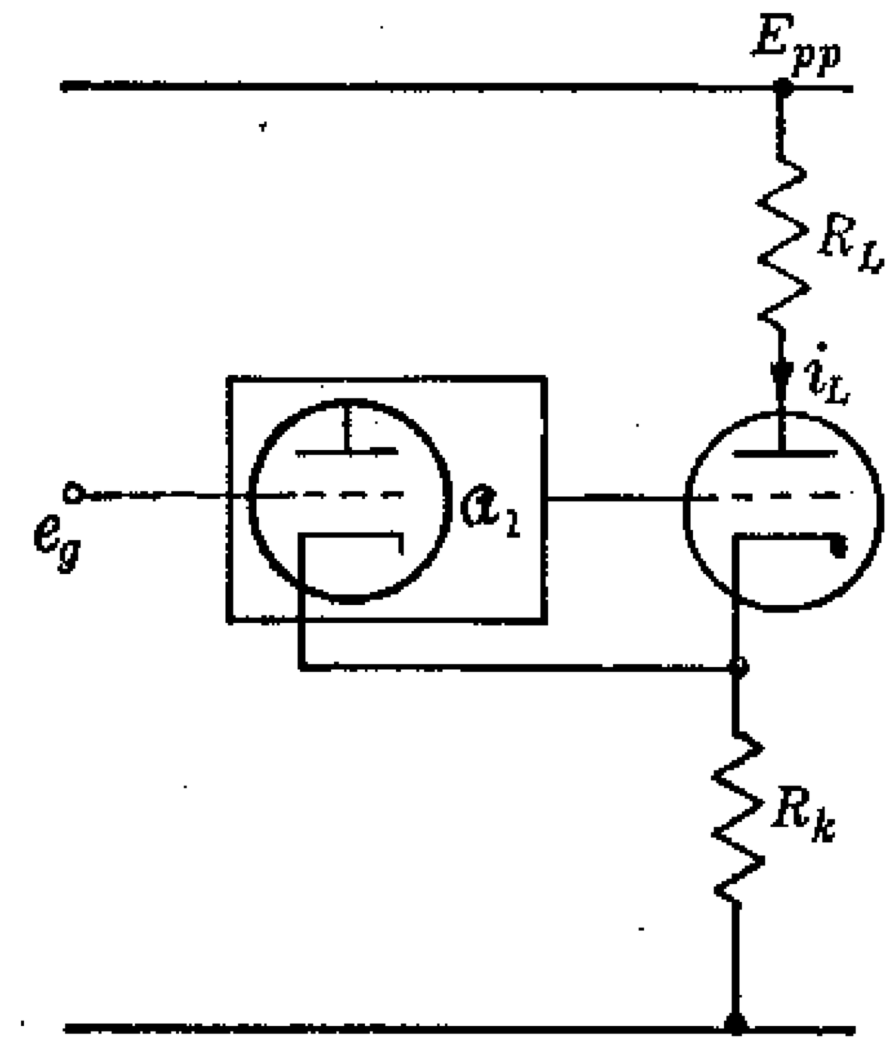

FIG. 11·54.—Cathode and screen feedback.

In all of the previous discussion, the loading effect of the feedback network on the amplifier has been assumed to be included in $\mathfrak{A}$. If the amplifier also delivers power to an external load, which is not constant, $\mathfrak{A}$ will be caused to vary, and $\mathfrak{G}$ will vary in accordance with Eq. (113).

The feedback arrangements of Figs. 11·53 and 11·54 as well as those of Fig. 11·51 are *voltage* negative feedback; they operate to stabilize the output voltage with respect to changes of load; i.e., the effective output impedance is decreased. In fact it may readily be demonstrated that the output impedance of the amplifier is decreased in the ratio $\mathfrak{G}/\mathfrak{A}$.

This is not the case with the simple cathode-resistor system of Fig. 11·12 where the voltage output terminal is at the plate. Equation (11) shows that the output impedance is actually increased. This type of feedback, which operates to stabilize the current in the load resistor by feeding back a voltage developed by this current in a fixed resistor, is termed *current* feedback. Figure 11·55 shows a more elaborate example of current feedback. The output is the current in R_L. As this current also flows in R_k, the fedback voltage is proportional to it. (The current of the input tube also flows in R_k, but this is usually much smaller and is also nearly constant.) Figure 11·55 would be the same as Fig. 11·53 if the output voltage were considered to be the voltage at the cathode of the output tube (and $k = 1$ in Fig. 11·53). Thus the output current may be derived from these considerations by simply dividing the output voltage by R_k. The current in R_L is linearized with respect to e_g and stabilized with respect to changes of R_L, E_{pp}, etc., in a manner identical with that of the output voltage of Fig. 11·53.

The cathode feedback arrangement of Fig. 11·53 leaves the input circuit with no compensation for variation of heater voltage. In the resistance-adding feedback of Fig. 11·51, compensation may be provided at the input stage, but this type of feedback is often less desirable than the other because of the lower input impedance and because of the need for high-resistance precision resistors. Compensation may be provided

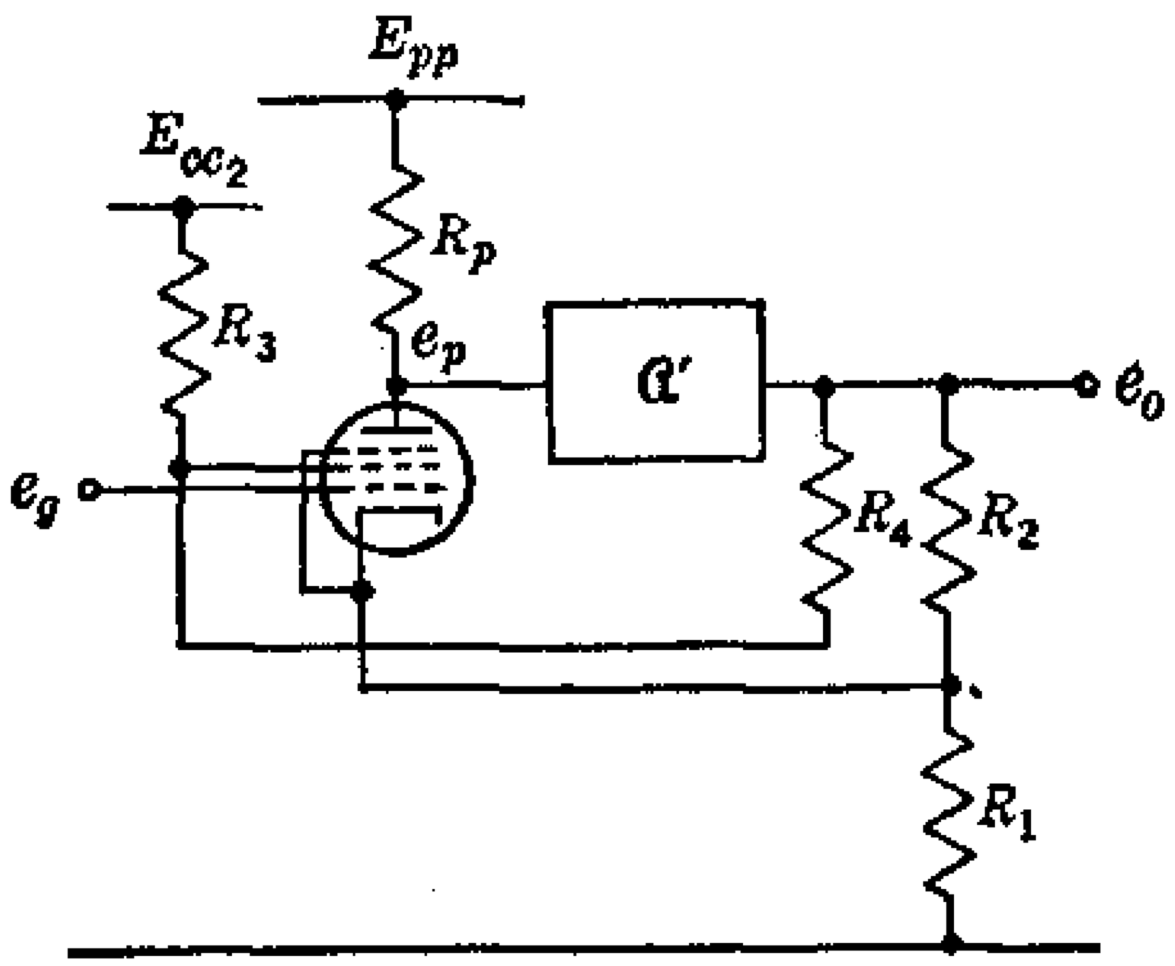

FIG. 11·55.—Example of current feedback.

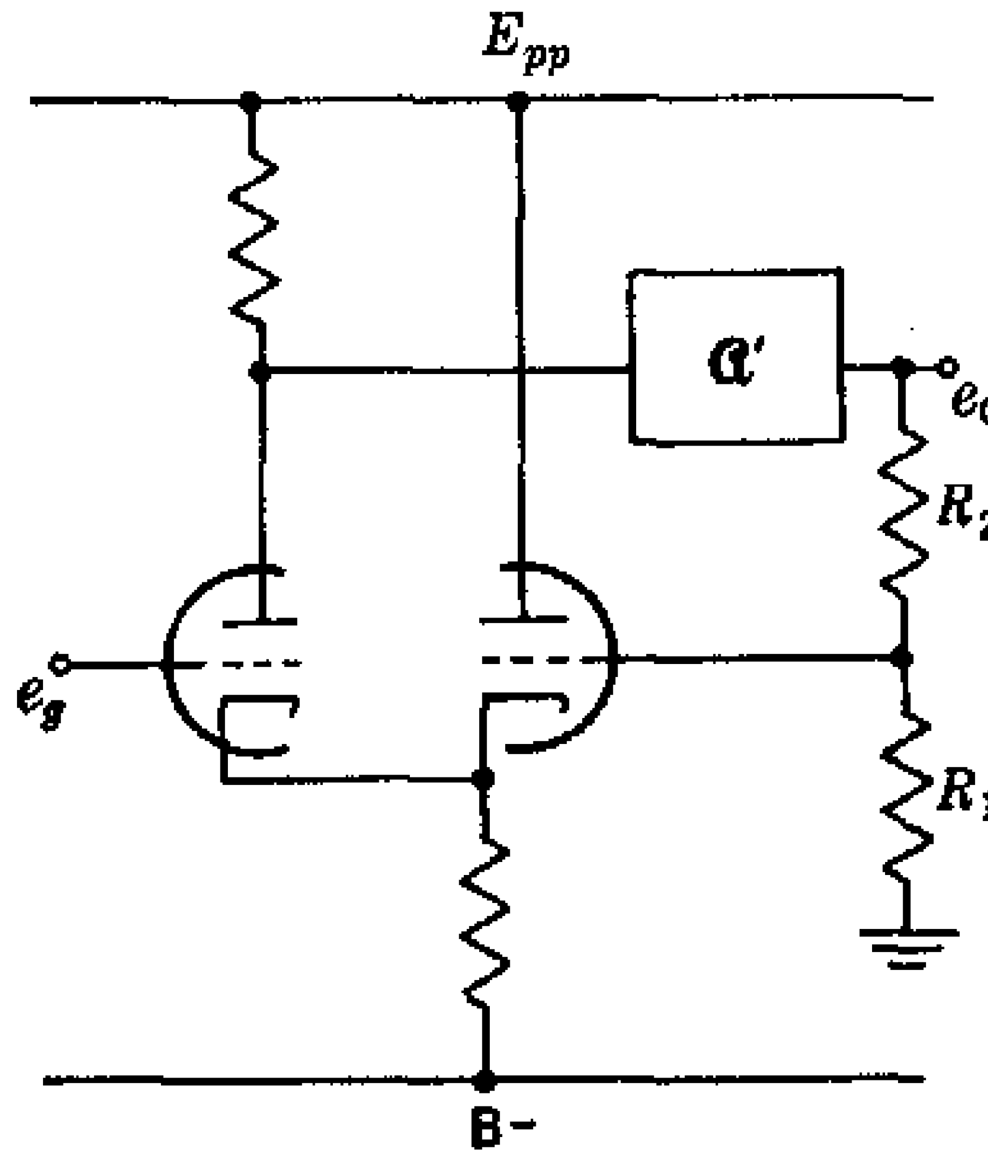

FIG. 11·56.—Cathode-follower feedback.

in the case of cathode feedback, by feeding back to the plate of a diode arranged as in Sec. 11·12. A more common method employs a cathode follower as in Fig. 11·56, which has the advantage that R_1 and R_2 do not reduce the gain of the input stage and therefore may be large enough to avoid serious loading of the output stage. The effective resistance in series with the cathode of the first triode is, from Eq. (16),

$$\frac{r_p}{\mu + 1 + \dfrac{r_p}{R_k}}$$

where r_p and μ refer to the cathode follower. If the triodes are similar, this effective resistance has the effect on the gain of the triode of slightly more than doubling its plate resistance. From Eq. (15) it is apparent that the effective feedback factor of Eq. (112) should be multiplied by $\mu/(\mu + 1 + r_p/R_k)$ so that if the μ's are the same,

$$\beta = -\frac{\mu + 1}{\mu + 1 + \dfrac{r_p}{R_k}} k, \tag{116}$$

where $k = R_1/(R_1 + R_2)$. The two μ's, however, are subject to small changes independently of each other. If μ_1 and μ_2 refer respectively to the amplifier and cathode follower and r_p to the latter, Eq. (115) becomes

$$\frac{\Delta \mathcal{G}}{\mathcal{G}} \approx \frac{1}{\mu_1 + 1} \frac{\Delta\mu_1}{\mu_1} - \frac{1 + \dfrac{r_p}{R_k}}{\mu_2 + 1 + \dfrac{r_p}{R_k}} \frac{\Delta\mu_2}{\mu_2}. \tag{117}$$

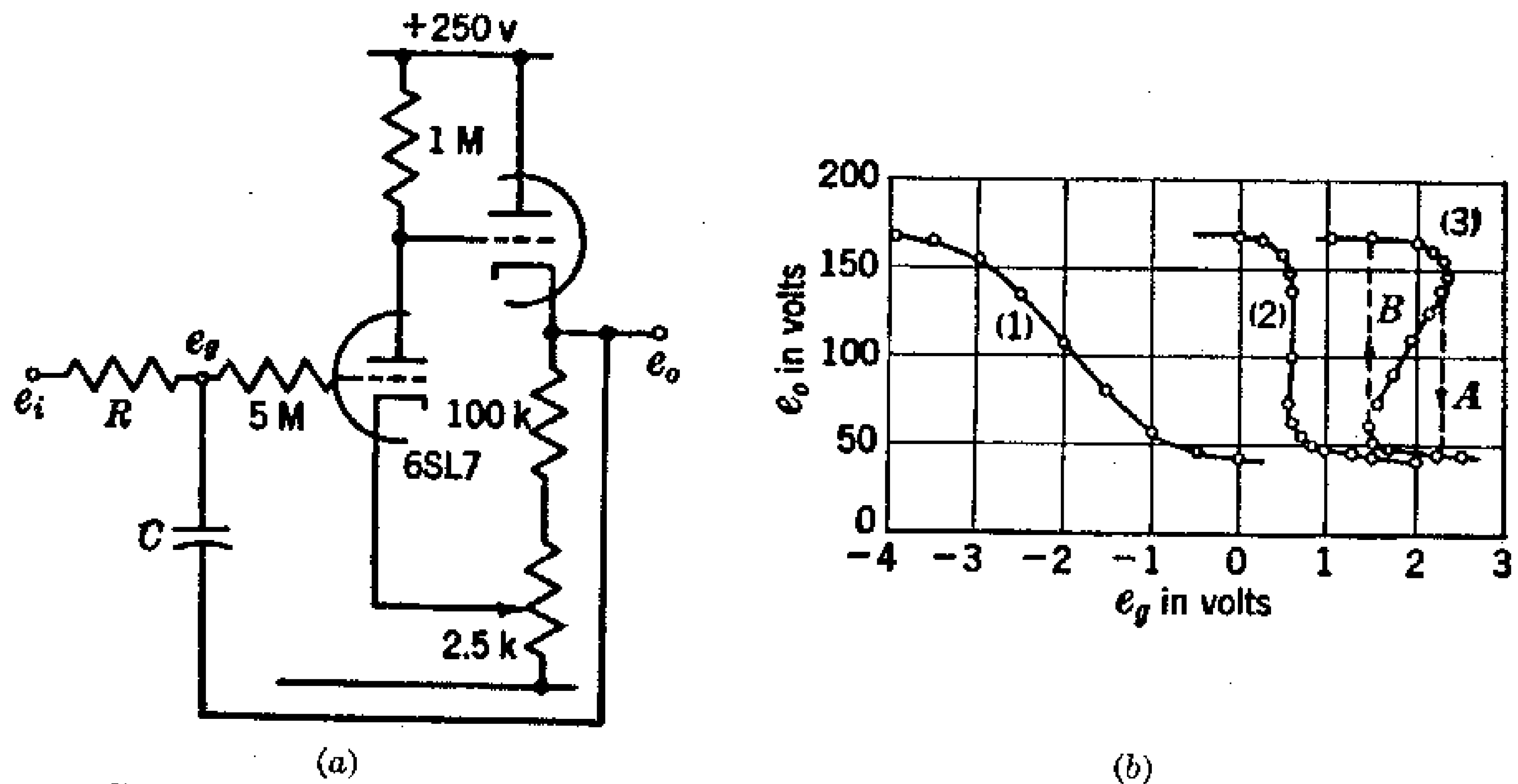

FIG. 11·57a.—Linear sweep or integrating circuit employing positive feedback.
FIG. 11·57b.—Amplifier characteristics of Fig. 11·57a with different feedback adjustments.

Much the same conditions obtain if the second triode in Fig. 11·56 is used as the amplifier so as to obtain the opposite sense of output voltage or if the two triodes are employed as a differential amplifier. The peculiarities of these arrangements are discussed in Sec. 11·10.

Positive Feedback.—Positive feedback has the effect of increasing the gain. Also, because Eq. (101) applies as well to positive as to negative feedback, it decreases the linearity and stability of gain. In some applications, such as when the output-voltage range is small and when it is feasible to readjust the feedback when tubes are replaced, this limitation is not serious. Amplifiers used in voltage regulators, linear sweep circuits, and servomechanisms may sometimes employ positive feedback to advantage.

In a linear sweep circuit, or voltage integrator, employing a direct-coupled amplifier and a resistor and condenser in a "fed-back time-constant" arrangement, perfect operation requires infinite gain. This may be obtained by adjusting the feedback factor β so that it is the reciprocal of the amplification, thereby causing the denominator in Eq. (93) to disappear. The circuit will still operate if β is so great that the denominator is actually negative—there will simply be an error of the same nature, but of opposite sign, as if the gain were finite and positive. Figure 11·57 gives an example of a simple linear sweep circuit or voltage integrator employing a high-gain amplifier with over-all capacitive negative feedback through C and R, such that, in so far as e_g is constant, the rate of change of output voltage is proportional to the input voltage. The amplifier between e_g and e_0 is a single-stage amplifier with a cathode-follower output stage and positive feedback to the cathode, as in Fig. 11·53. The curves are input vs. output voltage characteristics for the amplifier. The upper and lower limits are caused by grid current in the cathode follower and in the amplifier triode respectively (the grid resistor being for the purpose of demonstrating the latter). Curve 1 represents the characteristics with no positive feedback and shows a maximum gain of about 50. For Curve 2 the feedback was adjusted so as to afford infinite gain over a limited range. In the case of Curve 3 even more feedback was used, resulting in negative gain over a portion of the output voltage range. This reflex part is unstable; without R and C the output voltage would jump along dashed line A or B (respectively) if e_g were being raised or lowered, but with R and C in operation the whole curve is actually traced out.

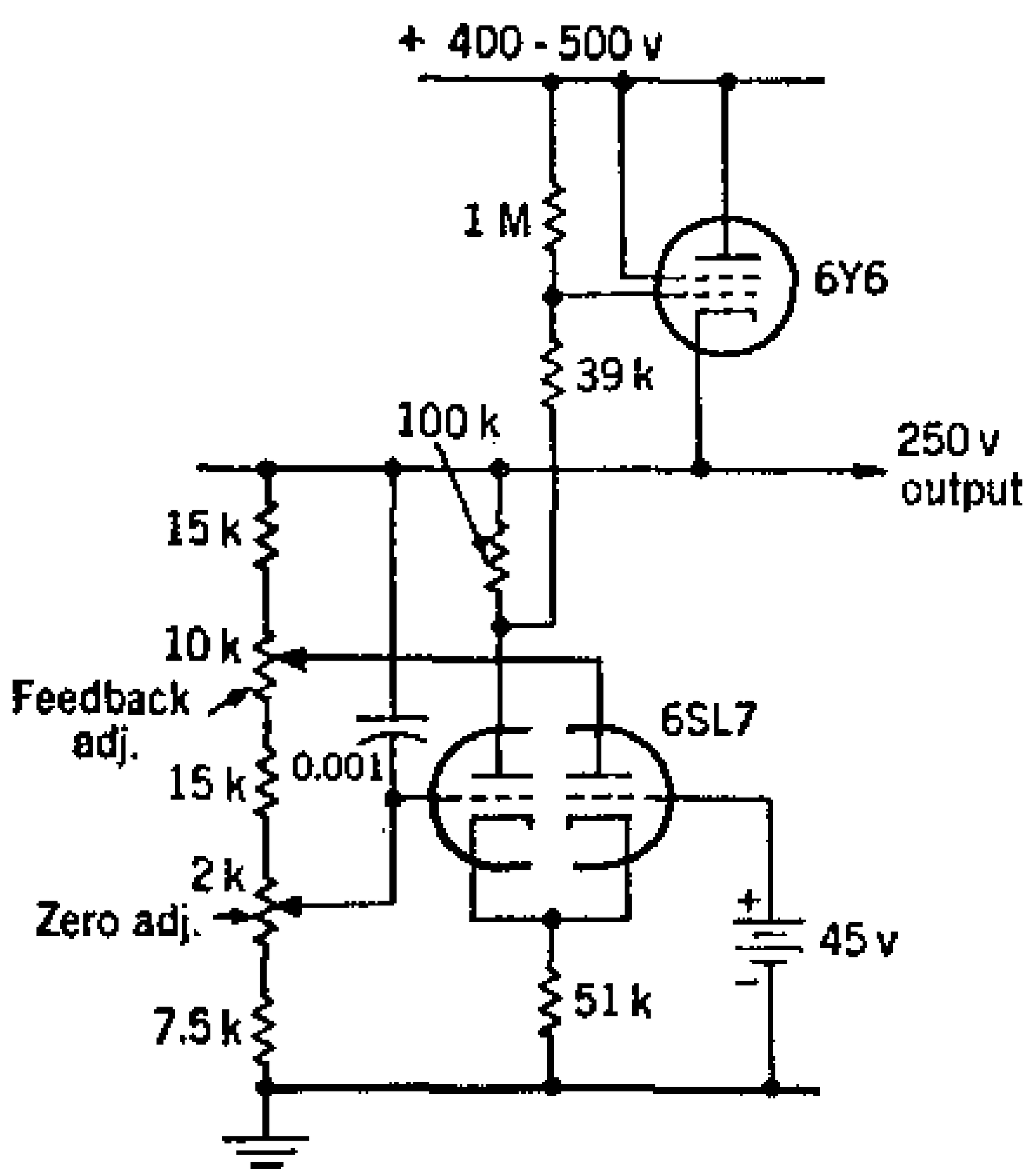

FIG. 11·58.—Voltage regulator employing positive feedback.

For perfect operation, an electronic voltage regulator, requires an amplifier with infinite gain, which may be approached or achieved conveniently with the aid of positive feedback. The regulator circuit also includes over-all negative feedback, which results in stability even though the positive feedback may be slightly more than is needed for infinite gain. The regulator of Fig. 11·58 may be adjusted to give an almost constant output over a limited range of load-current or supply-voltage variation. Positive feedback is provided by the resistive coupling between the cathode-follower plate and the amplifier grid. If the amplifier grid voltage is lowered, the decrease in plate current results in an almost equal increase in plate current in the cathode follower, which acts on the bleeder resistance to add to the input change. The cathode follower also provides heater-voltage compensation and loading of the reference source. The small condenser increases the negative feedback at high frequency, thus preventing oscillation. This circuit is given not as an example of a good voltage regulator but only as an illustration of the use of positive feedback as a simple means of increasing amplification.

In a servoamplifier the input signal comprises the error, and the gain should therefore be as high as possible. Because linearity is not of importance, it is permissible to employ positive feedback in many applications. Usually the upper limit of gain is determined by consideration of stability of the servomechanism, but there are instances[1] where infinite gain may be employed to advantage, and the feedback may be even greater, as in Fig. 11·57b, Curve 3, without causing oscillation, although an error of the reverse sense results.

If the feedback is adjusted for infinite gain [$\beta = 1/\mathcal{A}$ in Eq. (93)] and thereafter the amplification without feedback changes by a certain percentage (e.g., because of tube replacement), the gain with feedback will become some finite positive or negative quantity whose magnitude will be proportional to the original amplification. If $\beta = 1/\mathcal{A}$ in Eq. (93) and if $\mathcal{A} + \Delta\mathcal{A}$ is substituted for $\mathcal{A}$, the gain with feedback is found to be

$$\mathcal{G} = -\mathcal{A}\,\frac{1 + \Delta\mathcal{A}/\mathcal{A}}{\Delta\mathcal{A}/\mathcal{A}}. \tag{118}$$

Thus, for example, if β in Fig. 11·57 were adjusted to give Curve 2 and then the tube were replaced by one with an amplification that was 10 per cent higher, the curve would assume a shape slightly more like that of Curve 3, with a slope of $-50(1 + 0.10)/0.10 = -550$.

When positive feedback is to be applied to a multistage amplifier, the question arises as to whether to apply it over all or to the individual stages. In the case of negative feedback it was determined that from the standpoint of constancy of gain, it is better when arranged overall. The same reasoning gives the opposite answer for positive feedback. When $\mathcal{G}/\mathcal{A}$, $\mathcal{G}_1/\mathcal{A}_1$, etc., are *greater* than one, the expression for $\Delta\mathcal{G}/\mathcal{G}$ in Eq. (106) is smaller than that in Eq. (105). Thus, it is a general rule that positive feedback, if employed, should be applied only over single stages rather than over the whole amplifier. Of course, it should be applied to the stages that are subject to the least percentage variation of gain.

It follows, then, that a multistage amplifier with over-all negative feedback may be further stabilized as to gain by the use of positive feedback over individual stages. Figure 11·59 gives a computed illustration

[1] For example, in velocity servomechanisms employing RC feedback for stability.

of this principle in the case of an amplifier of two stages, each with a normal amplification of 20 and each being subject to the same variation. For each curve, the gain was computed from Eq. (93); the feedback factors were assumed to remain constant at values giving an over-all gain of five at the normal value of $\mathcal{A}$. The individual positive feedback is seen to give a great improvement in constancy of gain. The improvement is appreciable even when the positive feedback is applied to only one of the stages.

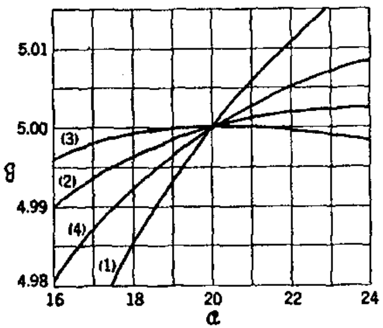

FIG. 11·59.—Variation with respect to $\mathcal{A}$ of the over-all gain $\mathcal{G}$ of two stages each having amplification $\mathcal{A}$ with fixed over-all negative feedback adjusted so that $\mathcal{G} = 5$ when $\mathcal{A} = 20$. Curve (1), no positive feedback; Curve (2), fixed individual positive feedback adjusted to give each stage a gain of 100 when $\mathcal{A} = 20$; Curve (3), fixed individual positive feedback adjusted to give each stage infinite gain when $\mathcal{A} = 20$; Curve (4), fixed positive feedback over only one of the stages adjusted to give it infinite gain when $\mathcal{A} = 20$.

Figure 11·60 is the circuit of a two-stage voltage amplifier employing SD-834 subminiature triodes ($\mu \approx 15$), with over-all negative feedback as in Fig. 11·51. The input voltage is applied to the cathode of the first amplifier tube by way of a cathode follower, permitting the necessary voltage inversion in two stages. Individual positive feedback is provided for each stage: in the first stage by coupling from the cathode-follower plate to the amplifier grid and in the second stage by coupling from the output stage (cathode follower) to the amplifier cathode, as in Fig. 11·57a. In the linearity tests whose results are presented in Fig. 11·61, the nonlinearity of the amplifier was exaggerated by operation up to an output voltage of 100 volts, where the second-stage amplifier tube is near cutoff. Figure 11·61 shows an improvement in the linearity of more than 10-fold obtained by the positive feedback, with the same over-all gain. Substituting a weaker set of tubes, without readjustment of feedback, lowered the gain from 10 to 7.95 when the positive feedback was not employed and from 10 to 9.78 when it was employed. A stronger set of tubes had the effect of changing the gain from 10 to 10.30 without and to 9.91 with positive feedback. In the latter case, the positive feedback was more than enough for infinite gain in each stage, but the

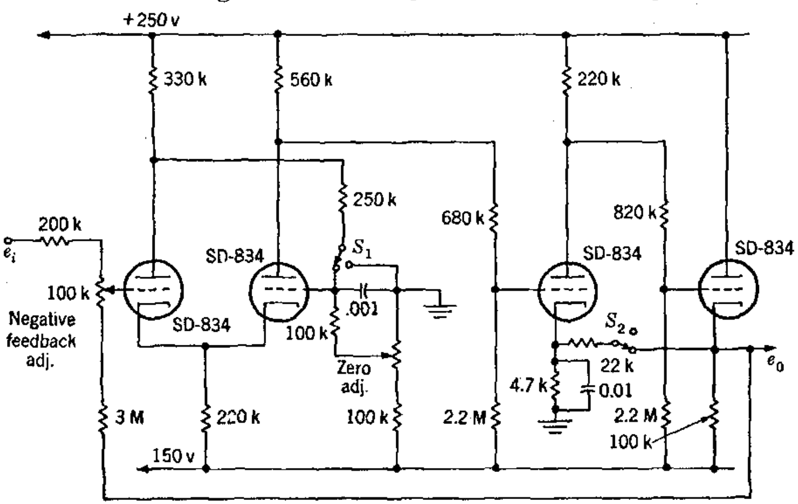

FIG. 11·60.—Two-stage voltage amplifier with gain stabilized at 10 by over-all negative feedback. Positive feedback operative for each stage with switches S_1 and S_2 in positions shown.

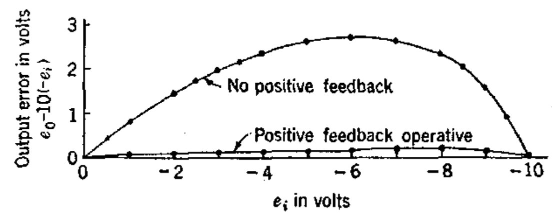

FIG. 11·61.—Deviation from linearity of output voltage, with and without positive feedback for each curve; negative feedback was adjusted so $e_o = 0$ at $e_i = 0$ and $e_o = 100$ at $e_i = -10$.

condensers prevented oscillation by reducing the effect at high frequencies without impairing the negative feedback.

No improvement can be obtained by the preceding method in the gain stability of a single-stage amplifier. Two β's between the same output and input points simply add algebraically.

EXAMPLES OF SPECIAL-PURPOSE AMPLIFIERS

In the next three sections are circuit diagrams and qualitative explanations of a few examples of direct-coupled amplifiers. The circuits are selected with the primary view of illustrating the principles discussed in previous sections, and they are not necessarily the best that have been developed for the purpose.

11·14. Current-output Amplifiers. *Meters.*—The single-stage differential amplifier of Fig. 11·62, which actuates a $\frac{1}{2}$-ma meter, illustrates the independent gain and zero adjustments shown in Fig. 11·30. The differential action affords a linearity commensurate with that of most portable meters. For a grid input voltage of 1 volt at full scale (as shown), very little resistance is needed between the cathodes; if the minimum full-scale input voltage can be 2 or 5 volts, the higher resistance allowed between the cathodes will appreciably improve the linearity. Saturation limits the overload on the meter to about 100 per cent.

A more precise voltmeter, suitable for actuating a recording meter, is given in Fig. 11·63. Current feedback is employed, the current in

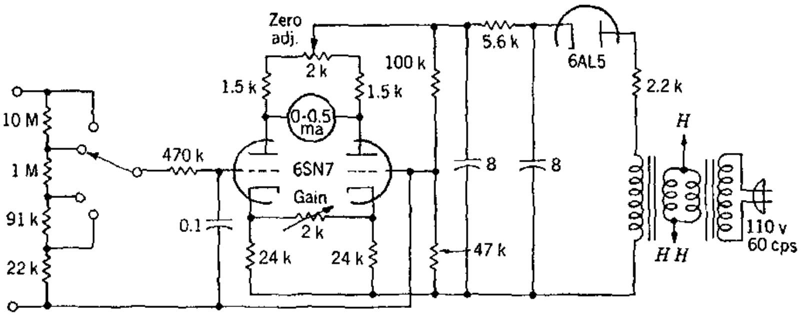

FIG. 11·62.—Simple vacuum-tube voltmeter.

the meter being measured by a stable resistance and a voltage proportional to it fed back to the auxiliary grid of the input differential amplifier.

The second stage is a pentode amplifier with its cathode returned to a cathode follower. This arrangement permits differential input so that the full gain of the first stage is realized. It also allows input voltages

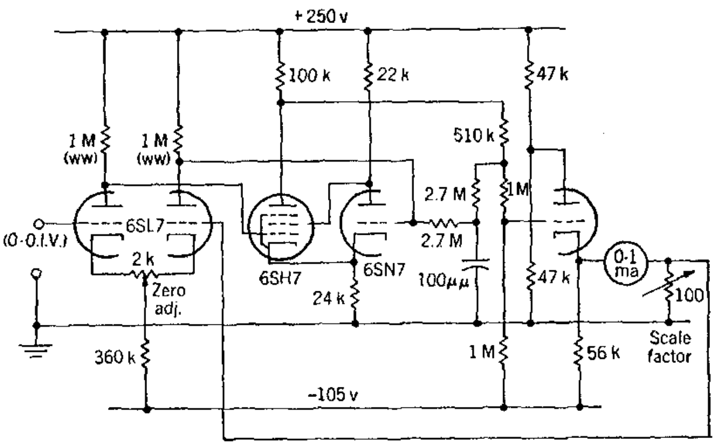

FIG. 11·63.—Voltmeter using high-gain amplifier.

at the plate level of the first stage; therefore, no dropping divider is needed at this point where the drift of such a divider would be felt much more than after the second stage. Another feature permitted by the cathode follower in the second stage is the local positive feedback (by means of the 5.4-megohm resistor), which increases the linearity with over-all negative feedback (Sec. 11·13). Connecting the pentode screen and triode plate together prevents the common dropping resistor from causing a reduction in gain; in addition, the reduction in screen voltage produces higher gain in the pentode. The output cathode follower is designed to limit the current in the 1-ma meter at about 2 ma in either direction; the two limits result from plate-current cutoff and from grid current in the input resistance.

The circuit may be used for full-scale input voltage as high as 10 volts, the limit depending on the value of the feedback resistor. The lower limit of full-scale input voltage is determined by the drift of the input tube. This limit is about 100 mv, and at this scale factor the drift may be noticeable over a period of hours. In order to minimize this drift,

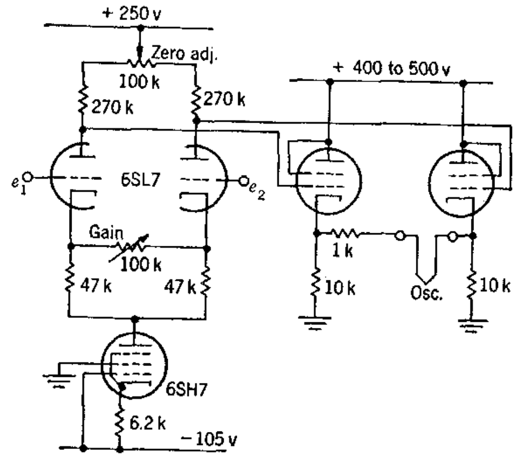

FIG. 11·64.—Two-stage oscillograph driver.

the input tube should be aged several hundred hours with the heater on. Selection of this tube is also recommended, as some tubes are subject to rapid erratic changes equivalent to input "noise" of 2 or 3 mv or more. This effect may be observed on the meter if the current-feedback resistor is reduced to 5 or 10 ohms.

Magnetic Oscillograph Driver.—The first stage of the driving circuit of Fig. 11·64 illustrates the use of a constant-current circuit in a differential amplifier to eliminate the input common-mode variation from the output voltage. The input voltage may be either differential or single-ended (with the unused input terminal fixed at the appropriate level) and may be at any level from -50 to $+100$ volts. Voltages beyond these limits may be reduced by the use of dividers. Thus, several driving circuits for measuring voltages at various levels may use the same power supply.

The gain may be varied from 6 to 25 ma per volt; large input voltages may be reduced by an input attenuator, but the amplifier gain should be set at minimum in such cases. The maximum output current is about ±18 ma; it can be increased by the use of smaller cathode-follower cathode-return resistances and larger output tubes. The negative feedback is applied individually in this amplifier, because over-all feedback is impractical where the input level is so variable. Linearity is satisfactory for the purpose.

Servoamplifiers.—The servoamplifier in Fig. 11·65 is designed to drive the differently wound fields of a small instrument servo motor.

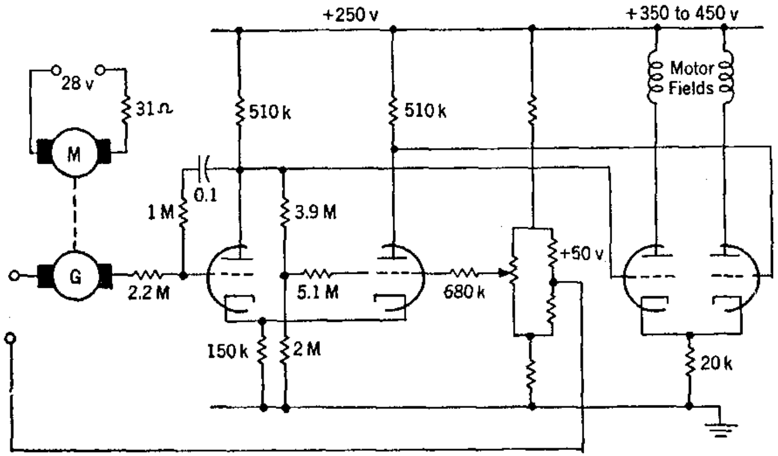

FIG. 11·65.—Velocity servo amplifier driving differentially wound motor fields.

The arrangement shown comprises a velocity servomechanism, wherein the motor drives a tachometer generator whose output voltage is subtracted from the input speed-control voltage. The difference voltage is the "error signal" to be amplified. No negative feedback is used in the amplifier itself, as this would detract from the linearizing effect of the over-all negative feedback, which includes the tachometer. On the other hand, local positive feedback is beneficial (Sec. 11·13), and this is provided in the first stage by the resistive coupling from the first plate to the second grid. This is designed for approximately infinite gain. The capacitive negative feedback, in combination with the resistances in series with the condenser, provides damping of oscillations of the servomechanism.

The input stage of the servoamplifier, which was shown in Fig. 11·42, employs a diode to balance the effect of heater-voltage variation on the cathode of the pentode. The zero-adjustment resistance and screen-grid voltage divider do not impair this function, as the currents in the various parts of the divider are substantially constant. The diode current is two or three times the pentode current. A slight loss of gain results from the zero adjustment and diode variational resistance, but these are very small compared with the plate-load resistance. The small auxiliary resistors in series with grid and plate are designed to dampen possible high-frequency oscillation. The operation of the output stage has been explained in Sec. 11·9.

Current Regulator.—Where large currents are to be held constant, the constant-current circuits of Sec. 11·8 are inadequate. The power control and the voltage amplification must be done separately, (and the current must be fed back and compared with a standard). The

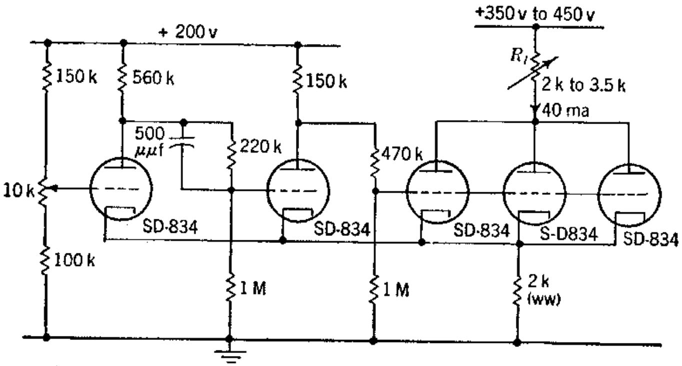

FIG. 11·66.—Current-regulating d-c amplifier.

function of the circuit of Fig. 11·66 employing an unregulated voltage source is to maintain a constant current in a load resistor whose resistance is subject to large variation. The load current flows in a constant resistance in the cathode circuit of the power stage, and the resulting voltage is fed back to the cathode of the input stage, whose grid receives the standard voltage. The cathode of the second voltage stage is connected to the same point, partly for convenience and partly because the slight cathode coupling between the first and second stages comprises positive feedback, increasing the gain. The condenser inhibits oscillation by decreasing phase retardation at high frequencies. If the 200-volt source changes, the output current changes in the same proportion. No compensation is provided for heater-voltage variation; a 10 per cent change amounts to an input drift of about 100 mv, which causes an output variation of 0.13 per cent.

11·15. Voltage-output Amplifiers. *Electrostatic Oscilloscope Driver.*—An electrostatic oscilloscope requires a large deflecting voltage, applied in a push-pull manner. The two-stage driver in Fig. 11·67 employs a pair of pentodes to get a large voltage swing without using an excessive supply voltage. The pentodes are arranged in a differential amplifier with single-ended input and with considerable linearizing degeneration from resistance between the cathodes. The input stage illustrates the method of cancellation, with a series triode, of drift due to heater-voltage variation, as explained in Sec. 11·2. The cathode resistor may be adjusted to give the best cancellation (after the tube has been aged for several hours) by observing the output change upon variation of the heater voltage. Thereafter, the zero adjustment is made at the plate of the upper triode. The input grid resistor is designed to protect the grid from excessive current due to high input voltage.

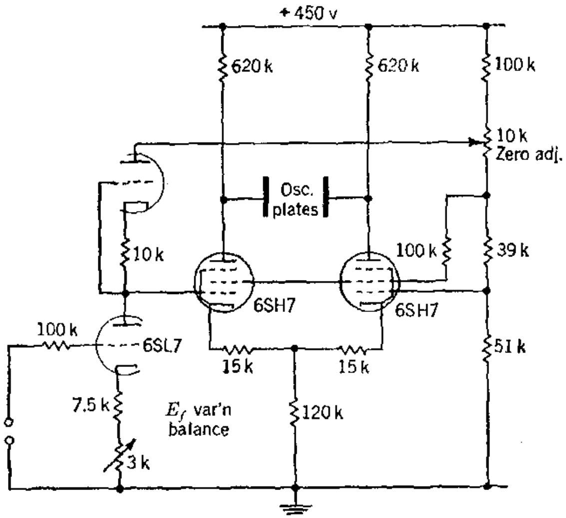

FIG. 11·67.—Two-stage voltage amplifier for oscilloscope deflection.

Comparison Amplifier.—In d-c circuits and especially in computing devices, the need sometimes arises for a high-gain amplifier with two input terminals and a single output terminal, in which the output voltage depends only on the difference and not on the mean level of the input voltages. The general application is in the electronic solving of an algebraic equation. The two input voltages are the two equal parts of the equation, and the output voltage is the "unknown," which actuates other circuit elements involved in the equation in such a way as to force the two input voltages into equality. Thus there is an over-all negative feedback in the application of the amplifier, and linearization by means of negative feedback within the amplifier is superfluous and detrimental. On the other hand, local positive feedback, if designed toward obtaining infinite gain, can improve the performance.

In the amplifier of Fig. 11·68 freedom from the effect of the common-mode variation is obtained by the use of a differential input stage with a constant-current triode as the common-cathode return. As the input voltage level is changed from 0 to +80 volts, the input voltage difference needs to be changed by less than 50 mv, for the majority of 6SU7 tubes, in order to hold the output voltage constant.

In addition to permitting the use of both output terminals of the first stage, the differential arrangement of the second stage provides a linearizing effect and allows the maximum output voltage range, as in Fig. 11·33*c*. It also provides a simple means for local positive feedback,

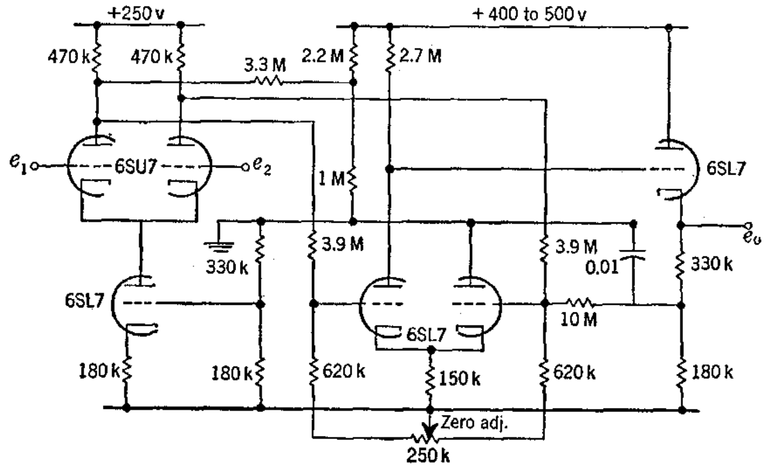

FIG. 11·68.—Two-stage amplifier with differential input and single-terminal cathode-follower output.

which is applied by resistive coupling from the cathode-follower output terminal to the cathode-follower grid in the second stage. The condenser in this network prevents oscillation by decreasing the positive feedback at high frequencies.

The output voltage range is from somewhat below zero to considerably above 250 volts. Without the positive feedback the gain of the amplifier is about 500, with very good linearity. With the positive feedback operative the output voltages traverse at least 250 volts while the input difference changes not more than 50 mv.

The resistance coupling between the high-voltage supply and the plate of the first tube is used to cancel the effect on the output of variations in this voltage. The +250 volts and −150-volt supplies are regulated. Heater-voltage variations are, of course, balanced by the differential arrangement.

The comparison amplifier may be used as a driver, similar to a cathode follower but having much greater precision, by connecting the output terminal to one of the input terminals (e_2 in Fig. 11·68). It may also be used as a linear, adjustable-gain amplifier, by feeding back any desired fraction of the output voltage through a resistance attenuator.

Inverting Amplifier.—For inverting a signal voltage (i.e., obtaining the negative of it, not its reciprocal), the simple degenerative triode circuit of Fig. 11·8 may be employed with plate and cathode resistors nearly equal. Besides having only medium precision, such an arrangement is limited in application because the output voltage contains a large

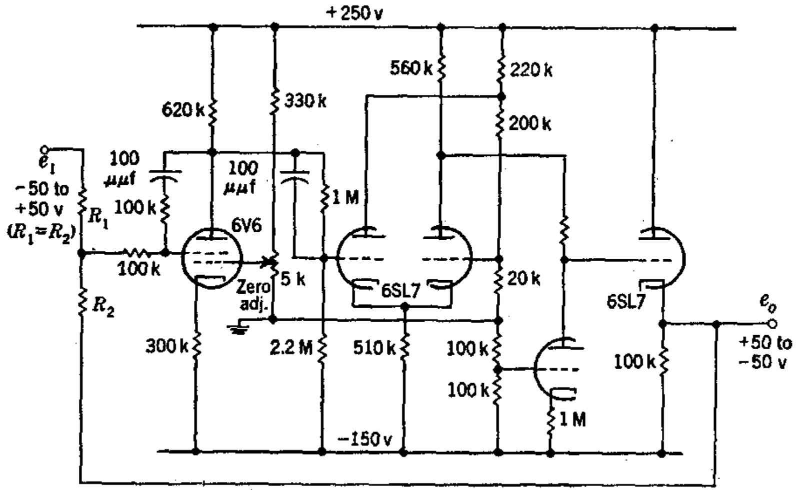

FIG. 11·69.—Voltage inverter employing a two-stage amplifier with cathode-follower output.

additive constant. This constant voltage is directly affected by any supply-voltage variation, and both it and the gain are affected by output load.

More precise voltage inversion, without these disadvantages, is accomplished by a high-gain amplifier whose input voltage is the average of the voltage to be inverted and the amplifier output voltage. The amplifier is sensed to invert, so the averaging process is done by negative feedback. If the gain is high enough so that the amplifier input voltage is negligible, and if the averaging resistors are equal, the output voltage is the negative of the input voltage.

In Fig. 11·69 a two-stage amplifier is made to invert by the use of cathode-follower input to the second stage. Positive feedback is applied in the second stage by means of resistive coupling between the first plate and the second grid. The potential is dropped to the grid of the output cathode follower, without attenuation, by the use of a triode constant-current circuit. This not only prevents attenuation but also permits a greater range of output voltages than would be possible with an ordinary divider. The input stage illustrates the use of a first grid as a diode plate to eliminate drift due to cathode change. For voltage inversion without change in scale factor R_1 and R_2 are equal. Amplification or attenuation may be obtained by other ratios between these resistances. The condenser-resistor feedback at the input stage prevents oscillation.

Voltage Level Changer.—Figure 11·70 illustrates a use of the push-pull quality of the output of a differential amplifier with a single input terminal. It is a simple device for providing a voltage that remains at a fixed potential difference above a variable input voltage. A possible

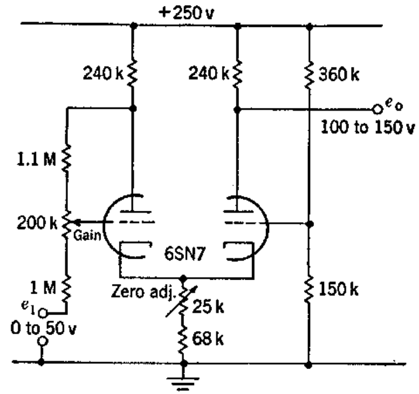

FIG. 11·70.—Differential amplifier arranged as a level shifter.

application might be to hold the screen of a pentode at a certain voltage with respect to the cathode regardless of movement of the latter. The resistance negative feedback from the first plate to the first grid is adjusted so that the gain from the input terminal to this plate is -1. Since the output is push-pull, the gain to the other plate is $+1$, but with a shift in voltage level.

11·16. A Galvanometer-photoelectric Tube Feedback Amplifier.—The minimum input scale factor of direct-coupled amplifiers is limited by drift of the characteristics of the input tube. The minimum significant input-voltage variation, with ordinary vacuum tubes, is seldom much less than 1 mv, even under short-time laboratory conditions.

The usual method of amplifying or measuring d-c voltages, too small for a direct-coupled amplifier, is to convert the direct current to alternating current by a modulator that is not subject to drift and to amplify the resultant voltage with an a-c amplifier. Depending on the output requirements, the amplifier output voltage may or may not be demodulated. Such modulation systems are outside the scope of this chapter.

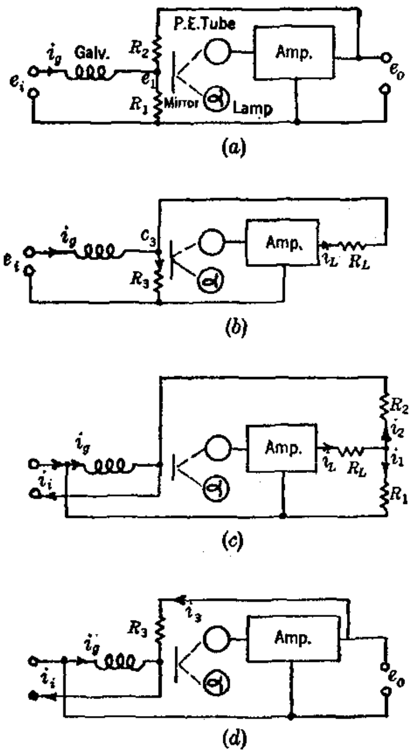

FIG. 11·71.—Feedback arrangements with galvanometer phototube input. (a) Voltage input, voltage output; (b) voltage input, current output; (c) current input, current output; (d) current input, voltage output.

An alternative to the a-c conversion method employs a device that is inserted ahead of a direct-coupled amplifier and is capable of increasing the amplitude of the input signal without drift. Linearity in the device itself is not needed if it is possible to arrange over-all negative feedback.

Such a scheme is afforded by the combination of a mirror galvanometer with a photoelectric tube. The general arrangement is shown in Fig. 11·71. A current in the galvanometer in one direction turns the mirror and alters the light received by the phototube in such a way as to raise the amplifier output voltage, and a current in the galvanometer in the other direction lowers the amplifier output voltage in a similar manner. A very small current is required to turn the mirror far enough to obtain full amplifier output voltage in either sense.

The negative feedback through a potential divider, illustrated in Fig. 11·71*a*, acts to reduce the galvanometer current by making e_1 almost equal to e_i. If equality is attained and i_g is zero, the output voltage becomes

$$e_0 = \frac{R_1 + R_2}{R_1} e_i. \tag{119}$$

If i_g is not negligible, the output voltage is

$$e_0 = \frac{R_1 + R_2}{R_1} (e_i - Ri_g), \tag{120}$$

where R is the parallel combination of R_1 and R_2, plus the galvanometer resistance, plus any series input resistance between e_i and the galvanometer. For good linearity Ri_g should be small in comparison with e_i. Ideally, therefore, the galvanometer should have no mechanical restoring torque and no friction.

The series feedback of Fig. 11·71*b*, where the output current is measured by being passed through a fixed resistance R_3, gives an output current of

$$i_L = \frac{e_i}{R_3} \tag{121}$$

if i_g is negligible.

Figure 11·71*c* is an arrangement for current amplification. If i_g is negligible, so that $i_i = i_2$, the output current is

$$i_L = \frac{R_1 + R_2}{R_1} i_i. \tag{122}$$

Also, the input voltage is zero, which is a requirement of an ideal current-measuring device.

The output voltage of the arrangement of Fig. 11·71*d* is

$$e_0 = R_3 i_i \tag{123}$$

if the galvanometer current is assumed to be zero. If the requirements are such that this resistance is too large for practical application, e_0 may be attenuated before being fed back as a current through R_3.

In the circuits of Figs. 11·71*b* and 11·71*c*, which have current feedback, the output current is not affected by the load resistance, the output voltage adjusting itself to compensate for any variations thereof.

By the use of a suspension galvanometer with a minimum restoring torque, the galvanometer current required for full output may be made a very small fraction of a microampere. In Figs. 11·71*a* and 11·71*b* the galvanometer and feedback resistances can be low [e.g., approximately 1000 ohms for R in Eq. (120)]; therefore if the input circuit resistance is also low, voltages well below 1 mv may be amplified with good linearity and without interference from drift. Drift of the amplifier itself is of negligible importance because it is offset by a very small shift of the mirror position.

The device actually comprises a servomechanism and is therefore subject to mechanical oscillation under some conditions. This oscillation may be obviated by negative feedback through a condenser, in addition to the resistive negative feedback and at a somewhat higher amplification.

The rapidity of response does not depend on mechanical restoring torque. The latter can (and should) be zero; the negative feedback will always act to hold the galvanometer rigidly in its proper position. Any change of input voltage, which results in a galvanometer current and a turn of the mirror, is immediately counteracted by feedback of the proper amount to cause the mirror to assume its appropriate new position. Any small error in this position results in a large restorative galvanometer current. Since the total movement is small and the electrical restoring torque is large, the rate of response of the device is much higher than that of a galvanometer as such. The frequency response is limited by the inertia, the movement required, and the condenser feedback needed to prevent oscillation.

The circuit of Fig. 11·72 has been used successfully as a microammeter.[1] The feedback is as illustrated in Fig. 11·71*c*, where R_1 and R_2

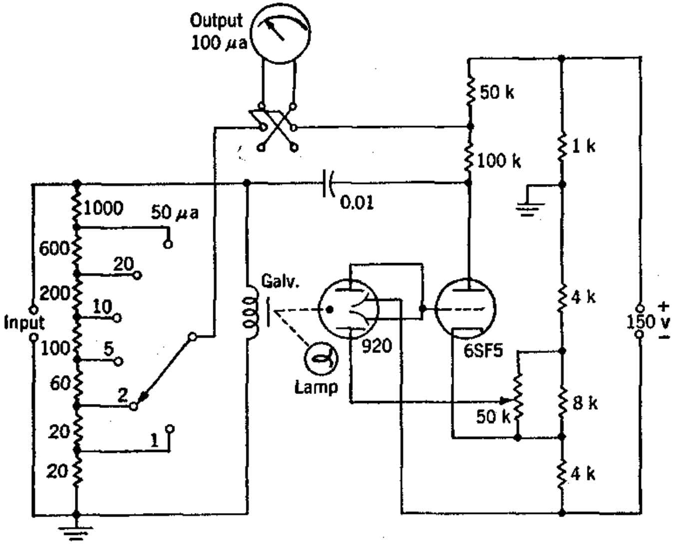

FIG. 11·72.—Microammeter based on Fig. 11·71c.

are in the form of an ayrton shunt to give several scale factors. The condenser feedback to prevent hunting is derived at a point of higher gain than the resistive feedback, by employing a tap on the plate-load resistor for the latter. No heater-voltage compensation is needed, as this effect is not noticeable; most of the gain of the system is ahead of the amplifier tube.

A differential phototube is employed, in a manner that affords very high output voltage to the amplifier grid for a small shift of the light beam. At the low voltages employed, the rate of change of anode-to-cathode voltage with respect to illumination, for a given current in each part of the phototube, is high. With the two parts connected in series as shown, the two currents are equal (if output current to the amplifier grid is neglected), with the result that an increase of light to one cathode and a decrease of light to the other gives the same output voltage that would be obtained if an infinite load resistor were used for each section. A change of total light does not change the voltage level as it would if load resistors were used.

The electrical isolation between the amplifier input terminals and the galvanometer permits the former to be at a lower potential than the latter, so that the output signal need not have its potential lowered for the feedback.

The effective input resistance is much lower than that of the galvanometer alone, because the feedback tends to keep the voltage across the galvanometer at zero regardless of input current.

11·17. D-c Amplifier Analysis.—This section deals with the analysis of the performance of a d-c amplifier from inspection of its circuit dia-

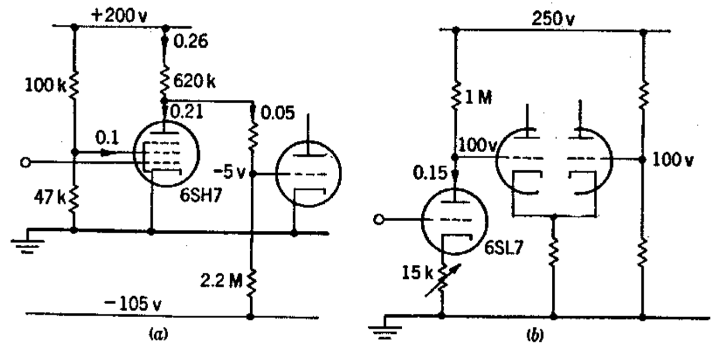

FIG. 11·73.—Computation of tube-operating conditions, based on an estimated grid potential for the following tube.

gram. The primary object is to determine the approximate voltages and currents at every part of the circuit in order to ascertain if each tube is being used in a satisfactory manner. Other objects of analysis may be the determination of linearity, effect of variation of supply and filament voltages, effects of the incidence of extremes of tube and resistor tolerances, etc.

Because of the rather large possible discrepancies between nominal and actual resistance values, only a fairly rough computation is justified in most instances. Simple application of Ohm's law, Thévenin's theorem, and the principles of the foregoing sections as applied to individual stages are sufficient to describe adequately the d-c aspects of any multistage amplifier.

An estimate of the grid-cathode potential in a tube is a good starting point for determining conditions in the preceding stage. From a knowledge of the tube type and the plate load the grid bias can be estimated within 2 or 3 volts. (For a low-μ power tube a consultation of its characteristic curves may be advisable.) The computation proceeds backward in the circuit from this point. Some specific examples will illustrate the method; no attempt will be made to formulate a general set of rules.

In Fig. 11·73*a* the bias for the following tube has been estimated as -5 volts. The potential across the lower part of the divider is thus 100 volts; therefore (if there is no grid current) the pentode plate potential must be 1/2.2 of 100 above -5, or $+40$ volts, and the divider current is about 0.05 ma. The current in the plate resistor is $\frac{160}{620} = 0.26$ ma, and the plate current is $0.26 - 0.05 = 0.21$ ma. Screen current will be about one-fourth to one-half of this, for example, 0.1 ma. The resistance of the screen bleeder as viewed from the screen is 32,000 ohms, so its regulation due to screen current is only 3 volts. Screen potential is about 60 volts. The grid potential of the pentode is found, from characteristic curves, to be about -2.0 volts (subject to some variation from tube to tube). This bias is ample to ensure against positive grid current.

From the characteristic curves, the transconductance, at $i_p = 0.21$ ma and $e_s = 60$ volts, is found to be about 0.8 ma/volt. The value of μ_s, or $-\partial e_s/\partial e_g$, is 35. The effective load resistance is the parallel resistance of 620,000 ohms and 3.2 megohms, or 520,000 ohms. The gain, from Eq. (27), is therefore (assuming $n = 0.5$)

$$\mathcal{G} = \frac{-(520)(0.8)}{1 + \frac{(0.5)(32)(0.8)}{35}} = \frac{-420}{1 + 0.37} = -300.$$

(The second term in the denominator is due to screen-bleeder resistance; if this were reduced by a factor of 4 the gain would be 380, an increase that might be worth the extra current drain.) The output divider to the following grid attenuates the gain to about 200.

In Fig. 11·73*b* the output to the differential amplifier will evidently be in the neighborhood of $+100$ volts. Since there is no divider, the plate current is 150 volts/1000k = 0.15 ma. From Fig. 11·73*a*, the grid bias appears to be about -1.6 volts. If the cathode resistor is for the purpose of zeroing the input potential for an output of 100 volts, its value will be $1.6/0.15 = 10.7$ kilohms. Plate resistance and amplification factor at $i_p = 0.15$ ma, $e_p = 100$ volts, are 130,000 ohms and 65 (approximately); therefore the gain is [Eq. (9)]

$$\mathcal{G} = -\frac{65}{1 + \frac{130 + (66)(10.7)}{1000}} = -\frac{65}{1 + 0.83} = -35.$$

In both (*a*) and (*b*) of Fig. 11·73, an increase of 10 per cent in heater voltage will require a lowering of input potential of about 0.1 volt if the output is to be held constant; otherwise the output will change by this amount times the gain.

Since the output variation required for full-range operation of the following tube is only a few volts at most, the preceding computations should suffice. If the required output range of the tube in question is great, the computation must be repeated for each limit. This usually is necessary only at the output stage.

The output of the amplifier of Fig. 11·69 is supposed to range from -50 to $+50$ volts with respect to ground. At the lower limit, the cathode follower has a plate-cathode voltage of 300, a plate current of 1 ma, and a grid bias, therefore, of -4 volts; its grid potential will then be -54 volts. The current in the constant-current circuit comprising the lower part of the divider feeding the cathode follower is about 75 volts/1 megohn = 0.075 ma; the amplifier plate therefore is at a level 150 volts above the cathode-follower grid, or $+96$ volts. Current in the plate resistor is $(250 - 96)/560 = 0.275$ ma, so that the plate current is this quantity minus 0.075 or 0.2 ma. The grid of this triode is in the neighborhood of $+10$ volts, and the cathode will be slightly above the grid; therefore the current in the common cathode resistor is 0.32 ma, and 0.12 ma is left in the first triode of the differential amplifier. At the upper output limit, the amplifier plate is at $+199$ volts, plate current is 0.09 ma, and plate current in the other triode is 0.32 minus this, or 0.23 ma. In this condition the latter triode still has ample plate-cathode voltage (about 90 volts) to ensure that the grid current is negligible in view of the input resistance.

The average plate resistance of each of the differential amplifier triodes is about 120,000 ohms. (One increases and the other decreases as the output range is traversed.) The first triode is a kind of cathode follower driving the cathode of the second. The plate-supply divider adds 110,000 ohms to the r_p of this cathode follower with a μ of 65, and Eq. (40) gives a gain from the first grid to the second plate of 40, without the positive feedback. Thus, for 1 volt of input, the second plate moves 40 volts. The first plate moves $\frac{110}{560}$ of this, or 8 volts in the opposite direction, since a certain change of current in the 560,000-ohm plate resistor implies an equal and opposite change in the other. This 8 volts of plate movement causes a 0.8-volt displacement at the second grid, so that for the assumed 1 volt of total differential input only 0.2 volt is needed at the first grid; the positive feedback increases the gain by a factor of 5.

In determining the d-c levels in a multistage amplifier the logical direction for the computation to proceed is generally from the output toward the input. An important exception is in the case of differential amplifiers, where the double-ended signal is transmitted as such from stage to stage.

For example, in the servoamplifier of Fig. 11·65, the level of the grids of the output stage cannot be estimated as in the foregoing examples; this level is determined by conditions in the first stage. The input grids are at about +50 volts, and the cathodes ride a volt or two about this, so the sum of the plate currents is 0.35 ma. At balance, then, the plates will be at 250 − (510)(0.17) = 160 volts, approximately. About 8 or 9 ma will flow in the output-tube cathode resistor, and this is divided between the two motor fields according to the amount of unbalance.

In Fig. 11·64 the first step is to compute the current in the constant-current pentode. If the i_p and i_s vs. e_g curves of the pentode are available, a simple method is to draw a line from the origin with a slope of $-1/R_k$, as shown in Fig. 11·74. The point at which this line intersects the curve of total cathode current, for the screen-cathode potential in

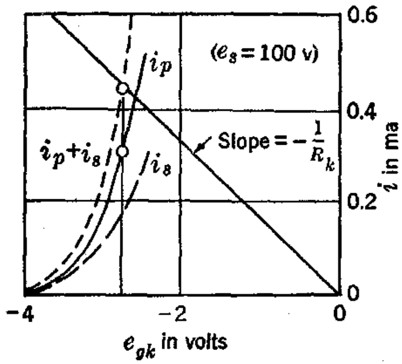

FIG. 11·74.—Determination of constant-current pentode current.

question (about 100 volts in this instance), will indicate the bias. In the example this is −2.8 volts, and the plate current is therefore 0.3 ma. This current flows in the two plates of the differential amplifier, regardless of the input level, so that the average of the two plate potentials is 250 − (320)(0.15) = 200. The cathodes of the power tubes will ride at about 20 volts above their grids; therefore, at balance, the current in each of these tubes will be $\frac{220}{10}$ = 22 ma. This is also the approximate current available for the load at saturation, i.e., when either tube is cut off.

When part of an amplifier is in differential form but the output is not, the computation may have to converge from both ends. The circuits of Figs. 11·63 and 11·68 are examples of this situation. In the latter, the grid of the output cathode follower is required to travel from about −8 volts to nearly +250 volts. This range gives the limits of plate potential (from ground) and current in the preceding triode; but to find its cathode level and the limits of current in the other triode of this differential pair, a fresh start must be made at the input stage. Here, the constant-current circuit holds the total cathode current at 0.3 ma. At balance, each plate current will be 0.15 ma. With no plate current each plate would be at about 210 volts because of the dividers to the following stage input. The effective plate-load resistance is 420,000 ohms, including the dividers, so the plates will be at

$$210 - (420)(0.15) = 145 \text{ volts.}$$

Thus, the grids of the second stage are at −103 volts, and its total plate current is about 0.33 ma.

Computation of gain may be made either by application of the gain formulas (with care to evaluate the variational tube parameters in the vicinity of the existing conditions) or by recomputation of the input for two assumed values of output. The same alternatives exist in the determination of the effects of positive and negative feedback. Linearity may be estimated by means of a gain calculation at each extreme of output and application of Eq. (102) or an appropriate variation thereof.

For determination of the effect of possible extremes of resistor values and vacuum-tube characteristics, at least two computations of the conditions must be made, after a qualitative inspection to decide which extreme of each component will result in a displacement of a given sense.

In ascertaining the shift resulting from changes of d-c supply potentials, the appropriate derivatives of the amplifier formulas [e.g., Eqs. (3), (7), (15), and (26)] may be applied. Great care must be used to consider the effect on every electrode of each tube, e.g., the screen grids of pentodes. Perhaps a surer method is a recalculation of the amplifier input for a certain output, at each of two extremes of the supply voltage; the output shift would be this value times the gain.