FUNDAMENTALS
OF
RADIO

BY

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WITH THE COLLABORATION OF

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quencies and at the higher audio frequencies. In addition, the plate and screen circuits of low-level stages can be isolated by the use of series impedances (resistances or inductances) and shunt condensers, as shown in Fig. 75*b*. Such arrangements are called filters, and serve to make the voltage actually transferred to the plate and screen circuits of a low-level stage less than the voltage developed across the common impedance. In order to be effective, the series impedance of such a filter must be

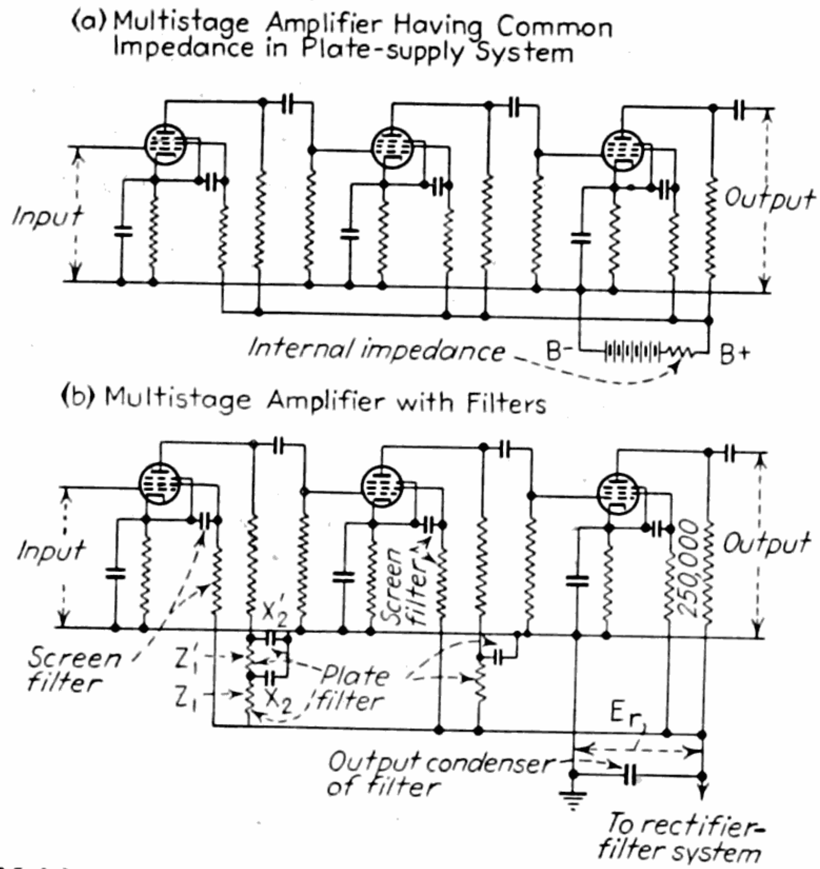

FIG. 75.—Multistage amplifier showing how the internal impedance of a common power-supply system can transfer energy from the final stage to earlier stages, together with same amplifier provided with filters for reducing this energy transfer.

much higher than the reactance of the shunt condenser at all frequencies for which the filter is to be effective. The high series impedance then prevents the voltage developed across the common impedance from sending appreciable current toward the input stage, while the low shunt impedance short-circuits substantially all of whatever current does flow. The result is that the voltage actually introduced into the plate and screen circuits of the input tube is very much less than the voltage developed across the common impedance. The reduction is given approximately by the equations[1]

[1] These equations assume that X_2 is small compared with Z_1 and the plate and coupling impedances of the input tube, and that Z_1 is much larger than the common impedance. Under these conditions a voltage E_r across the common impedance causes a current E_r/Z_1 to flow through Z_1 (Fig. 75*b*). The greater part of this current

$$\text{Reduction with one-stage filter} = \frac{X_2}{Z_1} \tag{69a}$$

$$\text{Reduction with two-stage filter} = \frac{X_2X_2'}{Z_1Z_1'} \tag{69b}$$

where Z_1 and Z_1' are the series impedances of the filter sections, and X_2 and X_2' are the reactances of the shunting condensers.

The effectiveness of filters and shunting condensers becomes less the lower the frequency, and trouble is practically always experienced in avoiding regeneration in high-gain amplifiers that have a good low-frequency response. This makes it desirable to design audio-frequency amplifiers so that the low-frequency response is no better than actually required. Even then it is often necessary to use a separate power-supply system for different parts of the amplifier.

The principles involved in designing a filter system are illustrated by the following example.

Example.—In the amplifier of Fig. 75*b* each stage gives a voltage gain of 60 in the middle-frequency range, and the filtering need not be effective below 40 cycles. The internal impedance of the plate-supply system is the reactance formed by an 8-μf shunting condenser. Assign values to the elements in the plate filter Fig. 75*b*, so that the voltage acting in the plate circuit of the first tube will not exceed 10 per cent of the voltage which this tube applies to the grid of the second tube.

On the assumption that the signal at the grid of the second tube is 1 mv, the voltage developed across the coupling resistance in the output of the final tube is

$$60 \times 60 \times 0.001 = 3.6 \text{ volts.}$$

The current through the coupling resistance is then $3.6/250{,}000 = 14.4\ \mu$a, and this flowing through 8 μf will at 40 cycles produce a voltage

$$14.4 \times 10^{-6}/(2\pi \times 40 \times 8 \times 10^{-6}) = 0.0072 \text{ volt.}$$

The filter for the first stage is then required to reduce this to 0.0001 volt, or by a factor of $1/72$. A variety of condenser-resistance combinations will do this satisfactorily. Thus substitution in Eq. (69*b*) shows that if $\frac{1}{2}$ μf shunting condensers are used, the series resistances for a two-stage filter must each be at least

$$\sqrt{72}/(2\pi \times 40 \times 0.5 \times 10^{-6}) = 68{,}000 \text{ ohms.}$$

It will be noted that little or no filter is required for the middle stage. This is because the voltage that this tube delivers to the grid at the final tube is 0.060 volt, which is quite large compared with the 0.0072 volt developed across the common impedance.

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flows through X_2 and so produces a voltage E_rX_2/Z_1 across X_2. In the case of a one-stage filter, this is the voltage actually inserted in the plate circuit of the input tube, and is less than the voltage E_r by the factor X_2/Z_1. In the case of a two-stage filter, there is a further reduction of X_2'/Z_1', giving a total reduction of X_2X_2'/Z_1Z_1', as in Eq. (69*b*).

The screen-grid filter can ordinarily be taken care of by the voltage-dropping resistance in the screen circuit, provided the by-pass condenser from screen to cathode is of reasonable size. Thus in Fig. 75*b*, if a 1-megohm series resistance is required to drop the supply voltage to the value needed for the screen, and the screen-cathode condenser is 1 μf, then the reduction at 40 cycles as calculated from Eq. (69*a*) is 251, which is obviously adequate. However, with very high-gain amplifiers it is sometimes necessary to provide a two-stage filter in the screen circuit by dividing the voltage-dropping resistance into two parts, and using two by-pass condensers.

Audio-frequency amplifiers are also troubled by regeneration arising from electrostatic and magnetic couplings. Electrostatic coupling between unshielded tubes and leads, particularly grid leads of different stages, often causes trouble. This coupling usually results in a high audio-frequency oscillation, and the remedy consists in providing adequate shielding. Magnetic coupling between stages can arise when there is more than one coupling transformer employed. The usual remedy consists in a change in relative orientation of the transformers involved, or in increased spacing.

Regeneration in Radio-frequency Amplifiers.—At radio-frequencies a common cause of regeneration is electrostatic and magnetic coupling between tuned circuits of different stages. This cause of regeneration is ordinarily controlled by enclosing coils in copper or aluminum shielding cans, and by shielding the sections of the tuning condenser from each other. It is also necessary to employ screen-grid or pentode tubes, or to use neutralized triodes, in order to prevent energy transfer directly through the tube. Care must likewise be used in placing the wiring, particularly the grid and plate leads, so that these introduce no capacitive coupling between stages.

Impedances common to two or more stages are particularly troublesome causes of regeneration in radio-frequency amplifiers. This is because at the high frequencies involved a wire only a few inches long will often have sufficient reactance to provide an effective means of transferring energy. Unavoidable common impedances, such as those from common sources of electrode voltages, can be reduced to a low value by shunting with by-pass condensers, and in addition filters may be placed in the leads running to the individual amplifier stages. Indiscriminate use of the chassis as a return circuit often gives trouble, since the chassis, when used in this way, provides an impedance common to all the circuits.

48. Feedback Amplifiers.—In the feedback amplifier a certain amount of regeneration is deliberately introduced in such a way as to reduce the amplifier gain. By properly carrying out this operation it is possible to reduce the non-linear distortion and noise generated in the amplifier, to make the amplification substantially independent of electrode voltages and tube constants, and to reduce greatly the phase and frequency distortion.

The operation of a feedback amplifier can be understood by reference to the schematic diagram of Fig. 76. Here A represents an amplifier which has a gain A when used as an ordinary amplifier. Regeneration is introduced by superimposing on the amplifier input a fraction β of the output voltage E so that the actual input consists of a signal e_s plus the feedback voltage βE. The effective gain of the amplifier is then[1]

$$\left.\begin{matrix}\text{Gain, taking into}\\ \text{account feedback}\end{matrix}\right\} = \frac{A}{1 - A\beta} \tag{70}$$

In this equation the assumption as to signs is such that when the feedback voltage opposes the signal voltage, β is negative.

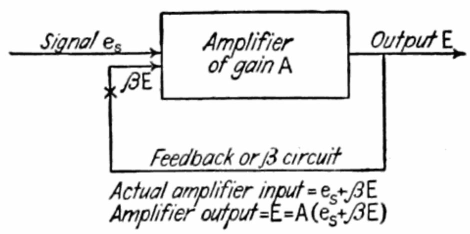

FIG. 76.—Schematic diagram of feedback amplifier.

The quantity $A\beta$ can be termed the feedback factor, and represents the amplitude of the voltage superimposed upon e_s compared with the actual voltage applied to the input terminals. Thus if $A\beta = 50$, then for each millivolt existing between the input terminals, the feedback voltage will be 50 mv. If the phase is such as to give negative feedback, a signal of 51 mv will then be required to produce 1 mv at the amplifier input terminals.

Examination of Eq. (70) shows that the amplification is reduced by the presence of negative feedback. Furthermore, when $A\beta \gg 1$, Eq. (70) reduces to

$$\left.\begin{matrix}\text{Amplification with}\\ \text{large feedback}\end{matrix}\right\} = \frac{-1}{\beta} \tag{71}$$

Expressed in words, Eq. (71) states that, when the feedback factor $A\beta$ is large, the effective amplification depends only upon the fraction β of the output voltage that is superimposed upon the amplifier input, and is *substantially independent of the gain actually produced by the amplifier itself.*

[1] Equation (70) is derived as follows: If E is the output voltage, then the feedback voltage is βE, and the actual input potential is $(e_s + \beta E)$. This input amplified A times must equal E, *i.e.*,

$$(e_s + \beta E)A = E$$

Equation (72) then follows by solving for E/e_s, which is the actual gain.

including the active part of the interelectrode capacity[1] as well as the resistance effects due to electron current flow, be represented by a resistance R in series with a capacitive reactance X (*i.e.*, $Z = R + jX$). In terms of this representation one has for complete space charge[2]

$$R = R_p \frac{12}{\theta^4} [2(1 - \cos\theta) - \theta \sin\theta] \tag{69}$$

$$X = -\left\{\frac{1}{\omega C} + R_p \frac{12}{\theta^4} [\theta(1 + \cos\theta) - 2 \sin\theta]\right\} \tag{70}$$

where R_p = dynamic plate-cathode resistance of diode at low frequencies.
θ = transit-time angle, radians ($\theta = \omega$ times transit time, sec.).
$\omega = 2\pi$ times frequency.
C = active part of the plate-cathode capacity of the tube measured with the cathode cold.

At frequencies where the transit-time effects are small, Eqs. (69) and (70), after suitable transformations, become

$$R = R_p \left[1 - \frac{\theta^2}{15}\right] \tag{71}$$

$$X = -\frac{3\theta}{10} R_p \left[1 - \frac{\theta^2}{25.2}\right] \tag{72}$$

It will be noted that θ is the transit time in radians, on the basis that the time represented by one cycle is 2π radians. A curve showing the variation of R and X with

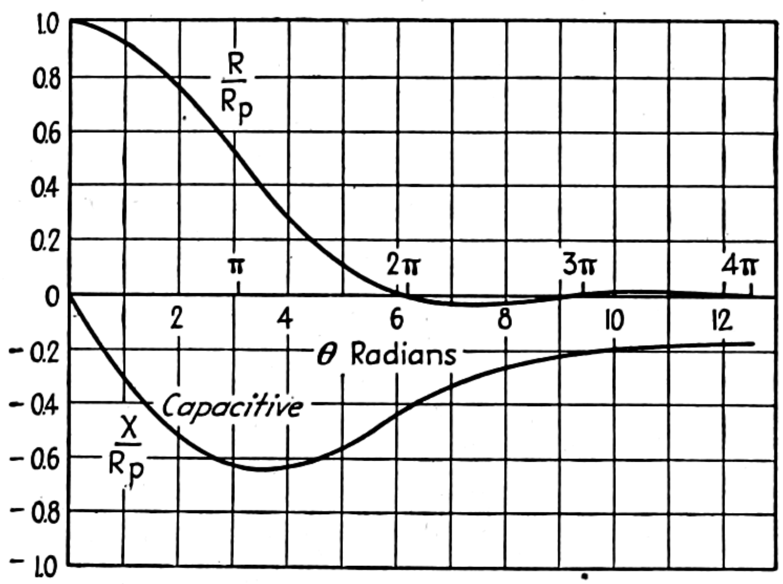

FIG. 28.—Resistance and reactance components of diode plate-cathode impedance as a function of transit angle θ (series representation).

transit angle is given in Fig. 28. It is seen that as the frequency is increased the resistance R first drops increasingly from its low-frequency value R_p until it becomes zero when the transit time is a full cycle, and with still greater transit times then

[1] The active part C of the interelectrode capacity is the capacity between parts of the tube that are effective in causing a current flow. The capacity between leads, or inactive parts of the plate or cathode structure, is not included, and represents an additional capacity that shunts the "active" impedance $R + jX$.

[2] See Llewellyn, *op. cit.*, p. 48; W. E. Benham, Electron Transit Time, *Wireless Eng.*, Vol. 16, p. 598, December, 1939. Graphical means of analyzing transit-time effects in diodes are described by R. W. Sloane and E. G. James, Transit-time Effects in Diodes in Pictorial Form, *Jour. I.E.E.*, Vol. 79, p. 291, 1936; also, Wireless Section, *I.E.E.*, Vol. 11, p. 247, September, 1936; and Rudolf Kompfner, Transit-time Phenomena in Electronic Tubes, *Wireless Eng.*, Vol. 19, p 3, January, 1942.

ELECTRONIC AND RADIO ENGINEERING

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$$g_m = \frac{3.5\sqrt{E_g + \frac{E_p}{\mu}}}{\left[d_c^{4/3} + \frac{(d_c + d_p)^{4/3}}{\mu}\right]^{3/2}} \text{ microamperes per volt per unit area} \tag{66}$$

where E_g = grid voltage.
E_p = plate voltage.
μ = amplification factor.
d_c = cathode-grid distance.
d_p = grid-plate distance.

and the electrode structure is that of Fig. 22. The transconductance is seen to depend upon the equivalent voltage $(E_g + E_p/\mu)$ and the electrode dimensions. Unlike the amplification factor, the transconductance does depend upon the cathode-grid distance. The transconductance of a plane electrode triode may also be expressed in terms of the plate current:

$$g_m = \frac{2.64 \times 10^{-4} i^{1/3}}{\left[d_c^{4/3} + \frac{(d_c + d_p)^{4/3}}{\mu}\right]^{3/2}} \text{ amperes per volt per unit area} \tag{67}$$

where i is the current density in amperes per unit area and the other symbols have the same significance as in Eq. (66).[1]

The transconductance of a *cylindrical electrode triode* with the configuration of Fig. 23 is

$$g_m = \frac{2.2\left(E_g + \frac{E_p}{\mu}\right)^{1/2}}{\left[(s_g\beta_{cg}^2)^{2/3} + \frac{1}{\mu}(s_p\beta_{cp}^2)^{2/3}\right]^{3/2}} \text{ microamperes per volt per unit length of structure} \tag{68}$$

where E_g = grid voltage.
E_p = plate voltage.
μ = amplification factor.
s_g = radius of grid-wire circle.
s_p = plate radius.
β_{cg}^2 = value of β^2 from Table 3 for argument s_g/s_c.
β_{cp}^2 = value of β^2 from Table 3 for argument s_p/s_c.
s_c = cathode radius.

9. Ultra-high-frequency Effects and Tubes.[2]—In ordinary applications of vacuum tubes the transit time of the electrons in the tube is short compared with the period of the applied voltages. As frequency is increased this becomes less and less true until finally the electron will require an appreciable fraction of a cycle to pass from one electrode to another. When this happens the behavior of the tube changes markedly.

Ultra-high-frequency Behavior of Diodes.—Even the simple diode tube experiences a difference in its behavior at ultra-high frequencies. The dynamic plate resistance of the diode drops, and at certain frequencies even becomes negative, and there are also modifications produced in the plate-cathode capacity.

When transit-time effects are of importance it is convenient for the purpose of analysis to consider that the impedance between the plate and cathode of the tube,

[1] J. H. Fremlin, Calculation of Triode Constants, *Elec. Comm.*, Vol. 18, p. 39, July, 1939.

[2] A comprehensive discussion and summary of transit-time effects is given in the book by F. B. Llewellyn, "Electron Inertia Effects," Cambridge (London), 1941. Much of this material is also given by F. B. Llewellyn, Operation of Ultra-high-frequency Tubes, *Bell System Tech. Jour.*, Vol. 15, p. 575, October, 1936.

oscillates about the zero value, being slightly negative at times. The reactive component of the impedance is always capacitive, but varies greatly. For very large transit times the tube impedance approaches that offered by the "cold" interelectrode capacity C, and the resistance component becomes very small.

In the preceding, the effect of the active part of the interelectrode capacity has been represented by a reactance in series with the equivalent plate resistance R. At frequencies where the transit-time effects are small, it is, however, more convenient to think of the tube plate-cathode impedance as representing a parallel combination of a capacity in parallel with an equivalent plate resistance. This equivalent resistance at low frequencies R is given by Eq. (69) or (71), while the equivalent low-frequency shunting capacity is $0.6C$. This low-frequency representation holds up to transit times of about $\theta = 1$. It will be noted that the presence of the electron stream in the interelectrode space causes the "hot" capacity of the tube to be only a fraction of the "cold" capacity C.

Input Resistance of Control Grids at Ultra-high Frequencies.[1]—At very high frequencies the transit time contributes a conductance component to the grid input admittance. A current will be induced in an electrode by the approach or passage of an electron even if the electron does not hit the electrode.[2] In the case of a triode or a tube with a screen grid, the currents induced in the control grid by approaching and receding electrons are different because the differences in velocities and the finite transit time impart an in-phase component to the induced current. The result of this action is that the conductance increases with frequency according to the formula[3]

$$G_g = Kg_mf^2T^2 \tag{73}$$

where G_g = conductive component of grid input admittance.
g_m = transconductance of tube.
f = frequency.
T = transit time to a reference point near the grid plane.
K = a constant for any particular tube, a function of the cathode-grid and grid-plate transit times.

Examination of the preceding formula shows that if all the dimensions of a tube are reduced by a factor M, then the input conductance will be reduced by a factor M^2. If all the voltages are increased by a factor N, the input conductance will be reduced by a factor $N^{\frac{1}{2}}$.

An idea of the magnitude of the transit-time effect upon grid input conductance can be gained from the fact that the input resistance of a 57 tube is of the order of megohms at 5 megacycles, while at 30 megacycles it has dropped to 20,000 ohms and at 100 megacycles it has dropped to about 1,500 ohms. The acorn pentode 954, which makes use of a very small cathode-grid spacing, has an input resistance that is of the order of 10 times as great as that of ordinary triodes, being about 20,000 ohms at 100 megacycles. The theoretically predicted variation of conductance with the square of the frequency is borne out by experimental observations.

Ultra-high-frequency Behavior of Triodes.—The behavior of triodes at ultra-high frequencies has been worked out to include both the effects of transit time and space charge.[4] Analysis shows that it is possible to set up a circuit that is the equivalent of

[1] The discussion here assumes that the electrode on the side of the control grid away from the cathode (screen or plate, as the case may be) is at ground potential to very high frequencies.

[2] B. J. Thompson, Review of Ultra-high Frequency Vacuum Tube Problems, *R.C.A. Rev.*, Vol. 3, p. 146, October, 1938.

[3] W. R. Ferris, Input Resistance of Vacuum Tubes as Ultra-high Frequency Amplifiers, *Proc. I.R.E.*, Vol. 24, p. 82, January, 1936; D. O. North, An Analysis of the Effects of Space Charge on Grid Impedance, *Proc. I.R.E.*, Vol. 24, p. 108, January, 1936.

[4] F. B. Llewellyn, Operation of Ultra-high Frequency Tubes, *Bell System Tech. Jour.*, Vol. 14, p. 112, October, 1935.

the vacuum tube and its transit-time effects, in terms of parameters that are independent of frequency and that are not complex. This is done by using a delta of impedances with the individual arms represented by the proper series or parallel combination of elements as shown in Fig. 29.[1] In this circuit the need for complex tube constants is avoided by the proper arrangements of circuit elements. The actual values applicable to the various parts of the circuit of Fig. 29 depend on the transit times, the dynamic plate resistance at low frequencies, and the capacities between the active parts of the various electrodes. The relations involved are too complicated and specialized to be given here, but are to be found in the references cited. The equivalent circuit of Fig. 29 is valid up to frequencies of 100 mc or more with ordinary tubes.

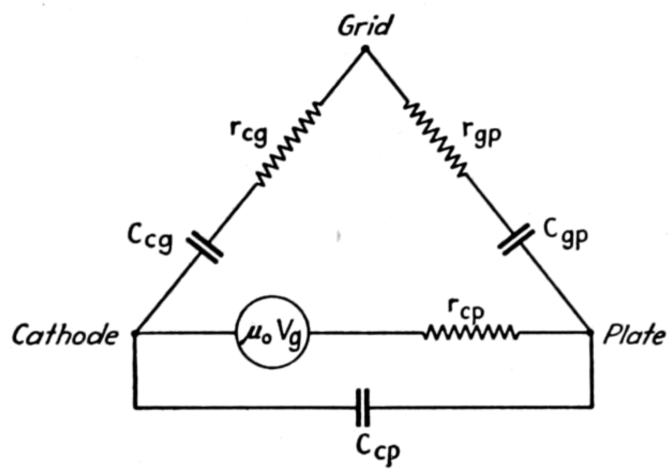


FIG. 29.—Equivalent delta of a triode at ultra-high frequencies.

The circuit of Fig. 29 is valid only for the active part of the tube, *i.e.*, the portion of the tube elements that intercept electrons. The effect of the lead impedances is not included, an important limitation to keep in mind, since the effects of the lead impedances may be as great as the effects of transit time of the electrons. Neither are the capacities between inactive parts of the electrodes and leads included. These capacities act in shunt with the equivalent circuit of Fig. 29, between the appropriate electrodes.

As far as the plate circuit of a triode is concerned, the effect of transit time can be taken into account by assuming that the ordinary equivalent plate circuit consisting of a generator voltage μe_s acting in series with the plate resistance R_p still holds (see Par. 3, Sec. 5), but that the value of amplification factor and plate resistance (and hence transconductance) is now modified in magnitude and a phase angle is associated with them. In particular, the transconductance tends to decrease in magnitude and lag by an increasingly large amount as the transit time increases,[2] while the amplification factor likewise tends to decrease and have an increasingly large phase angle.

Ultra-high-frequency Negative Grid Tubes.—Triode, beam, and other similar tubes intended for negative grid operation at very high frequencies must be especially designed for such service. The desired characteristics are those outlined in Par. 11, Sec. 6, for oscillator tubes, namely: (1) close electrode spacing to minimize transit time and give a high transconductance in proportion to electrode capacities; (2) short leads of relatively large diameter to minimize lead inductance and power losses in the leads (achieved in many cases by the use of multiple leads from the same electrode); (3) electrode and lead arrangements that facilitate operation with resonant transmission lines, the ideal being when the leads and electrodes of the tube represent extensions of the transmission line tank circuits.[3]

The tube should also be designed so that there can be no stray electrons that can shoot out of the open ends of the tube structure, and take a circuitous route in

[1] F. B. Llewellyn, Equivalent Networks of Negative-grid Vacuum Tubes at Ultra-high Frequencies, *Bell System Tech. Jour.*, Vol. 15, p. 575, October, 1936.

[2] Thus see F. B. Llewellyn, Phase Angle of Vacuum-tube Transconductance at Very High Frequencies, *Proc. I.R.E.*, Vol. 22, p. 947, August, 1934; Thompson, *loc. cit.*

[3] A tube in which this idea is carried to the limit is described by I. E. Mouromtseff and H. V. Noble, A New Type of Ultra-short-wave Oscillator, *Proc. I.R.E.*, Vol. 20, p. 1328, August, 1932.

traveling from the cathode to the plate.[1] Such electrons are likely to bombard the glass envelope and thereby release secondary electrons. These indirect paths, either with or without involving secondary electrons, result in some electrons arriving at the anode with excessively long transit times, with undesirable consequences. In order to avoid difficulties of this sort, it is desirable to close the ends of the tube structure so that it will not be possible for even a few electrons to stray along indirect routes.

When high voltages are involved, as is the case with transmitter tubes, the structural arrangement must be such as to provide an adequate amount of insulation. This is particularly important as the frequency increases, because the dielectric loss in the glass is proportional to the square of the frequency, and can reach sufficiently large values at extremely large frequencies to start a cumulative heating that results in puncture of the glass envelope.

It is possible to obtain satisfactory ultra-high-frequency operation of receiving tubes up to higher frequencies than can be achieved with power tubes. This is because the close spacings that favor ultra-high-frequency operation inevitably carry with them the necessity of making the tube dimension small, and this limits the power-dissipating ability of the tube. The amount of power that can be developed in the output of an ultra-high-frequency tube is accordingly less as the frequency limit is pushed to higher values. Ingenious arrangements have been devised for increasing the ability to dissipate heat, as, for example, the use of cooling fins attached to the plate, and in some cases also to the grid, and the use of water or forced-air cooling, even in tubes of moderate power rating. Such arrangements at best, however, merely improve an otherwise unfavorable situation, but in no case do they eliminate it.

A variety of ultra-high-frequency tubes are available. These include receiving tubes of the triode, pentode, and beam types, and also triode, beam, and pentode power tubes. Most of these are air-cooled, but a few of the larger types designed for such applications as television transmitters are arranged for water or forced-air cooling. In a few instances tubes are arranged in duplex (two tubes in a single envelope) for push-pull operation.[2]

10. Special Considerations Involved in Power Tubes.[3]—Tubes intended for use as power amplifiers must be capable of dissipating plate and grid losses in proportion to the desired power. Likewise, the electron emission from the cathode and the amount of voltage that can be applied to the anode with safety must be in proportion to the power output that the tube is to develop.

[1] W. G. Wagener, The Developmental Problems and Operating Characteristics of Two New Ultra-high-frequency Triodes, *Proc. I.R.E.*, Vol. 26, p. 401, April, 1938.

[2] Tubes especially designed for ultra-high-frequency use represent the accumulation of a great deal of specialized technical experience, much of which is of interest only to the designer of tubes. Because of this, and the fact that the subject is so new that the techniques are being steadily improved and modified, detailed descriptions of the various ultra-high-frequency tubes now available on the open market are not included here. The reader wishing to obtain a background in the subject is referred to the following articles: B. J. Thompson and G. M. Rose, Jr., Vacuum Tubes of Small Dimensions for Use at Extremely High Frequencies, *Proc. I.R.E.*, Vol. 21, p. 1707, December, 1933; M. J. Kelly and A. L. Samuel, Vacuum Tubes as High Frequency Oscillators, *Elec. Eng.*, Vol. 53, p. 1504, November, 1934; A. L. Samuel and N. E. Sowers, A Power Amplifier for Ultra-high Frequencies, *Proc. I.R.E.*, Vol. 24, p. 1464, November, 1936; A. L. Samuel, A Negative Grid Triode Oscillator and Amplifier for Ultra-high Frequencies, *Proc. I.R.E.*, Vol. 25, p. 1243, October, 1937; A. K. Wing, A Push Pull Ultra-high-frequency Beam Tetrode, *R.C.A. Rev.*, Vol. 4, p. 62, July, 1939; A Transmitter for Frequency-modulated Broadcast Service Using a New Ultra-high-frequency Tetrode, *R.C.A. Rev.*, Vol. 5, p. 327, January, 1941; A. K. Wing, Jr., and J. E. Young, A New Ultra-high-frequency Tetrode and Its Use in a 1-kilowatt Television Sound Transmitter, *Proc. I.R.E.*, Vol. 29, p. 5, January, 1941; K. C. DeWalt, Three New Ultra-high-frequency Triodes, *Proc. I.R.E.*, Vol. 29, p. 475, September, 1941; Cecil B. Haller, The Design and Development of Three New Ultra-high-frequency Transmitting Tubes, *Proc. I.R.E.*, Vol. 30, p. 20, January, 1942.

[3] An outstandingly fine discussion on the construction of large tubes is given by J. Bell, J. W. Davies, and B. S. Gossling, High-power Valves: Construction, Testing, and Operation, *Jour. I.E.E.*, Vol. 83, p. 176, 1938; also, Wireless Section, *I.E.E.*, Vol. 13, p. 177, September, 1938.

Small power tubes are often merely large receiving tubes constructed according to the standard receiving-tube techniques, involving oxide-coated cathodes, "getters" to obtain and maintain vacuum, etc. Such tubes are not capable of standing much abuse, however, and the maximum power that can be developed with this type of construction is limited.

Larger tubes, up to dissipations of several hundred watts, are normally still inclosed in a glass envelope but are constructed along lines entirely different from those of receiving tubes. The plates are commonly of molybdenum or tantalum, carbon sometimes being used. Grids are usually molybdenum, with tungsten and tantalum as alternates. Cathodes are sometimes oxide-coated, but more frequently are thoriated tungsten carburized for long life. The exhaust procedure is so thorough that a high degree of vacuum can be maintained without a getter. The various electrodes inside the tube are heated to high temperatures during exhaust, and the glass envelope is simultaneously baked at a temperature just below the softening point, and this process is continued until all the occluded gases have been removed. It is possible in this way to obtain a degree of vacuum such that a satisfactory life can be obtained with thoriated tungsten filaments even at high anode voltages, particularly if tantalum is used for the plate.

In tubes where the anode dissipation is of the order of kilowatts, the anodes of the tube are in the form of copper cylinders, which are part of the envelope and are cooled by means of circulating water or by being soldered to a radiating structure with many fins, which are cooled by means of an air blast. Analysis and experience indicate that water and forced-air cooling are about equally effective when used with vacuum tubes having external anodes.[1] Decision as to which type of cooling is preferable depends upon the individual factors involved in the particular case. Thus air-cooled tubes behave better in operation than water-cooled tubes with respect to "flashback." With forced-air cooling the anodes tend to run hotter, but the glass ends run cooler than with water cooling, which has some advantages and some disadvantages. With forced-air cooling the anode capacity to ground is increased by the large cooling structure, something to be avoided at very high frequencies.

The most effective way to employ water cooling is to arrange the water jacket so that the water passes over the anode in a very thin stream of high velocity. In this way any steam bubbles that may be formed on the anode tend to be scraped away by the velocity of the cooling water. The tendency for such bubbles to form limits the amount of heat that can be dissipated by a water-cooled anode, since, if the tendency of the bubble to stick to the copper is not overcome by the flow of cooling water, then that particular spot on the anode ceases to be cooled by direct contact with the water, and may readily become hot enough to cause failure of the tube.

In tubes with external anodes that are water-cooled, there is a tendency for the grid to produce a focusing effect on the electrons, which causes hot spots to be produced on the anode at the points where the electrons concentrate.[2] This is much more troublesome with water cooling than with forced-air cooling. With water cooling these hot spots tend to cause steam bubbles to form on the surface of the anode.

The cathodes of tubes employing water or forced-air cooling are always tungsten. This is because such high-power tubes are operated at very high anode voltages, and it is not practical to exhaust copper-anode tubes so thoroughly as glass-envelope tubes, where the anodes can be brought to incandescent heat during the exhaust process.

[1] Further discussion of both air and water cooling is given by I. E. Mouromtseff, Water and Forced-air Cooling of Vacuum Tubes, *Proc. I.R.E.*, Vol. 30, p. 190, April, 1942. See also E. M. Ostlund, Air Cooling Applied to External-anode Tubes, *Electronics*, Vol. 13, p. 36, June, 1940.

[2] I. E. Mouromtseff, The Influence of Grid Focusing Effect on Plate Dissipation Limit of a Vacuum Tube, *Communications*, Vol. 18, p. 9, December, 1938.

Thoriated tungsten filaments accordingly do not behave satisfactorily, and pure tungsten must be employed.[1]

The thoriated-tungsten and oxide-coated cathodes should be operated at the rated voltage to obtain maximum life. With tungsten filaments, however, decreasing the filament voltage will increase the useful life at the expense of reduced electron emission (and hence lowered peak anode current). Varying the voltage applied to a tungsten filament by 5 per cent from normal will double or halve the life as the case may be, while a 10 per cent variation affects the life by a factor of 4. A study of the optimum diameter of a tungsten filament, if the costs of power and tube replacement are taken into account, indicates that the optimum diameter is one that would correspond to a life of the order of 10,000 hours or more.[2]

In water-cooled or forced-air cooled tubes operating at high anode voltages (such as 15 to 20 kv), trouble is frequently encountered from arc-backs or "flash arcs" in the tube.[3] These are ordinary high-current arcs that suddenly form between cathode and plate for no apparent reason, and short-circuit the plate supply. After such an arc has been broken by opening of the circuit breakers, and the anode voltage has been reapplied, the tube may continue to operate for many hours or days without any further trouble, while in other cases another flash arc may occur almost immediately. A tube that has a tendency to arc-back will generally improve in this respect if it is put through a conditioning process[4] that involves applying an alternating potential of about 25,000 volts between plate and filament of the tube with the filament unlighted, with suitable relays and current-limiting devices in the circuit to prevent damage from arcs. This results in the production of many flash arcs in the tube, which gradually clean up the gas that is causing the trouble, and bring the tube to a satisfactory operating condition.

Power tubes are commonly operated so that the grid goes positive during a part of the cycle. This results in grid current, and causes power dissipation at the grid of the tube, which is sometimes the limiting factor in tube operation.[5] As a consequence, the grids of power tubes often operate at relatively high temperatures, and such materials as molybdenum, tungsten, or tantalum are accordingly generally used. The fraction of the primary electrons intercepted by the grid depends upon the grid potential relative to the anode potential, and upon the grid structure. In ordinary triodes with equal grid and plate voltages the effective grid area, insofar as intercepting the flow of primary electrons is concerned, is between 120 and 180 per cent of the actual grid area.[6] The grid heating that takes place is determined by the number of primary electrons intercepted by the grid, and by the grid voltage. The actual d-c grid current as measured by a meter may differ from the number of primary electrons received by the grid as a result of secondary emission causing the grid to lose secondary electrons at the same time that it receives primary electrons. The amount of current thus lost through secondary emission will be affected by the electrode potentials, by the grid temperature, and by the character of the grid surface.

[1] It is possible by keeping a water-cooled tube continuously on a vacuum pump, to maintain a degree of vacuum about one decimal point greater than feasible when the tube is sealed off. Under these conditions thoriated-tungsten filaments are entirely practical even at very high anode voltages. Such continuously evacuated tubes have been developed by various individuals and organizations for very high-power service, but as yet have not had any extensive continuing commercial use.

[2] Thus see J. J. Vormer, Filament Design for High Power Transmitting Valves, *Proc. I.R.E.*, Vol. 26, p. 1399, November, 1938.

[3] See Bell, Davies, and Gossling, *loc. cit.;* B. S. Gossling, The Flash-arc in High-power Valves, *Jour. I.E.E.*, Vol. 71, p. 460, 1932; also, Wireless Section, *I.E.E.*, Vol. 7, p. 192, September, 1932.

[4] Further details are given in Vacuum Tube Reconditioning, *Electronics*, Vol. 14, p. 84, January, 1941.

[5] I. E. Mouromtseff and H. N. Kozanowski, Grid Temperature as a Limiting Factor in Vacuum Tube Operation, *Proc. I.R.E.*, Vol. 24, p. 447, March, 1936.

[6] Karl Spangenberg, Current Division in Plane-electrode Triodes, *Proc. I.R.E.*, Vol. 28, p. 226, May, 1940.

In the case of thoriated tungsten and oxide-coated cathodes the secondary emissions may, under some conditions, become quite large as a result of cathode material that has been deposited upon the grid.

11. Receiving Tubes.—Receiving tubes can be divided into two main classes: the glass-envelope tubes, and the so-called "metal" tubes. The glass-envelope tubes, as the name implies, have a glass envelope, and the electrodes are mounted on leads running through either a stem or a glass cup, to which the bulb is sealed after the electrodes are mounted and adjusted. The metal-envelope tube employs a metal shell in place of the glass bulb, with all the leads, except perhaps a control-grid lead, brought through a glass button or glass beads in the base of the tube.

The cathodes of receiving tubes, either filaments or heaters, are practically always of the oxide-coated type. The grid and plate electrodes and side rods are usually constructed from nickel, and a "getter" is used to obtain the desired degree of vacuum. The exhaust, and also many of the other assembly operations, are performed on automatic machines.

The actual technical details of receiving tubes represent primarily a production problem, and accordingly are beyond the scope of this work. The literature available on the subject is relatively limited, and always lags behind the current practice in industry.[1]

12. "Getters."[2]—A high vacuum is obtained in receiving tubes by means of a "getter," which is volatilized inside the tube for the purpose of removing residual gas by either chemical or mechanical action. Magnesium is widely used as a getter, but other materials, such as barium, barium berylliate,[3] zirconium, phosphorus, etc., can also be employed, as well as various mixtures. A getter should be initially inert, but should be of such character that it can be made highly active by some process such as a chemical change, or vaporization of material inert at room temperature.

Some getters also act as "keepers" in that they not only remove what gas is present in the tube at the time of flashing but also combine with any gas that may subsequently be liberated within the tube.

Getters cannot be used in tubes where the anode dissipation is large. This is because the large amount of heat liberated in such tubes would vaporize the getter and destroy the vacuum.

Certain metals have the property of absorbing gases when raised to a high temperature. For example, tantalum will absorb gases when very hot, and this is one of the things that make tantalum a desirable metal to use in power tubes for plates and grids.

Zirconium is also a useful metal in this regard.[4] Zirconium at 1400° will absorb copious quantities of such gases as oxygen, nitrogen, carbon monoxide, and carbon dioxide. The optimum temperature for taking up hydrogen is, however, approximately 350°C, and at higher temperatures hydrogen begins to be given off by the metal. As a result of these properties it is possible to maintain an extremely high

[1] Useful articles on receiving tubes are M. Benjamin, C. W. Cosgrove, and G. W. Warren, Modern Receiving Valves: Design and Manufacture, *I.E.E. Wireless Proc.*, Vol. 12, p. 65, June, 1937 E. R. Wagner, Raw Materials in Vacuum Tube Manufacture, *Electronics*, Vol. 7, p. 104, April, 1934; Processes in Vacuum Tube Manufacture, *Electronics*, Vol. 7, p. 213, July, 1934; G. E. Moore, Improved Repeater Tubes, *Bell Lab. Rec.*, Vol. 18, p. 219, March, 1940; Newell R. Smith and Allen H. Schooley, Development and Production of the New Miniature Battery Tubes, *R.C.A. Rev.*, Vol. 4, p. 496, April, 1940; S. R. Mullard, The Development of the Receiving Valve, *Jour. I.E.E.*, Vol. 76, p. 10, 1935, also Wireless Section, *I.E.E.*, Vol. 10, p. 1, March, 1935.

[2] A good discussion of getters is given by E. A. Lederer and D. H. Wamsley, "Batalum," a Barium Getter for Metal Tubes, *R.C.A. Rev.*, Vol. 2, p. 117, July, 1937.

[3] See E. A. Lederer, Recent Advances in Barium Getter Technique, *R.C.A. Rev.*, Vol. 4, p. 310, January, 1940.

[4] Further information is given by J. D. Fast, Zirconium and Its Compounds with a High Melting Point, *Phillips Tech. Jour.*, World's Fair Issue, 1939.

vacuum, provided that there is zirconium in the tube at these two temperatures. This can be accomplished either by employing two zirconium filaments with different filament currents or by using a single piece of zirconium in which different parts of the metal are raised to different temperatures, so that the temperature range extends from 300 to 1400°C. This expedient is frequently used in experimental tubes to clean up gases that may be evolved during use, and can also be employed in special tubes that are particularly difficult to evacuate.

13. Effect of Gas upon Tube Characteristics—Maximum Allowable Resistance in Grid Circuits.—Very small traces of gas in vacuum tubes affect the characteristics adversely in a number of ways as a result of the positive ions produced in the tube by collision between the gas molecules and the electrons flowing to the anode. The positive ions travel in the opposite direction from the electrons, and normally end their existence by falling into the cathode or the negative control grid. Electrons that bombard the cathode tend to destroy the emission of thoriated-tungsten and oxide-coated cathodes. Positive ions collected by the negative grid result in grid current, which limits the resistance that may be inserted in series with a negative grid, and which also introduces noise.

Positive-ion currents to the grid limit the d-c resistance that may safely be placed in series with the control grid and the cathode, because the voltage drop that such a grid current produces across the resistance has a polarity that makes the grid less negative than would otherwise be the case. Thus if the tube begins to liberate gas with resulting positive-ion grid current, the grid becomes less negative, thereby increasing the space current. This increases the number of positive ions produced, and will cause additional grid current, and still greater reduction in the negative grid potential. If the resistance in the grid circuit is high enough, this process can become cumulative, and in some types of tubes can easily result in the destruction of the tube as a result of excessive plate current caused from loss of grid bias. The maximum resistance that it is permissible to place in series with the grid electrode depends upon the tube characteristics and the method of obtaining bias. It is of the order of several megohms in small tubes used for voltage amplification at audio and radio frequencies. With small power tubes, such as the output tubes of radio receivers and public-address systems, the allowable grid resistance is much less, particularly if a fixed bias is employed. The use of self-bias permits an increased resistance to be employed in the grid-cathode circuit, since self-bias provides an automatic protection against excessive increase in plate current.

Positive-ion current to a control grid has a "shot" effect associated with it, since the current flow is composed of individually charged particles, and is not a uniform fluid. The noise that is produced by this current flowing through an impedance Z_g between grid and cathode can be expressed by the formula

$$R_n = 19.3\ I_g|Z_g|^2 \tag{74}$$

where $|Z_g|$ is the absolute value of the external circuit impedance between grid and cathode, I_g is the grid current in amperes, and R_n the resistance that, when connected between grid and cathode, would produce the same thermal agitation voltage as is actually produced by the grid current flowing through the impedance Z_g.

In addition to the noise effect represented by Eq. (74), the presence of gas introduces an additional noise factor through the fact that positive ions not captured by the grid travel toward the cathode and produce irregularities in the space charge around the cathode. Little is known about this effect, however, beyond the fact that it is probably small under ordinary circumstances compared with the effect represented by Eq. (74).

14. Miscellaneous Types of Tubes. *Space-charge-grid Tubes.*—In the space-charge-grid tube there is an auxiliary grid, called the space-charge grid, located between the cathode and the control grid, and operated at a low positive potential. This grid increases the number of electrons drawn out of the space charge near the cathode. Although some of these electrons are immediately attracted to the space charge grid, many pass through its meshes into the region between the space-charge grid and the control grid. Here they are slowed down by the retarding field and form a space charge or virtual cathode, as shown in Fig. 30. The virtual cathode serves as an actual cathode as far as the remainder of the tube is concerned. The characteristics that result are similar to those for conventional tubes, as is apparent from Fig. 31.

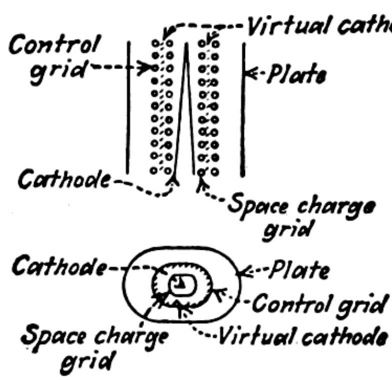


FIG. 30.—Details of a space-charge grid tube.

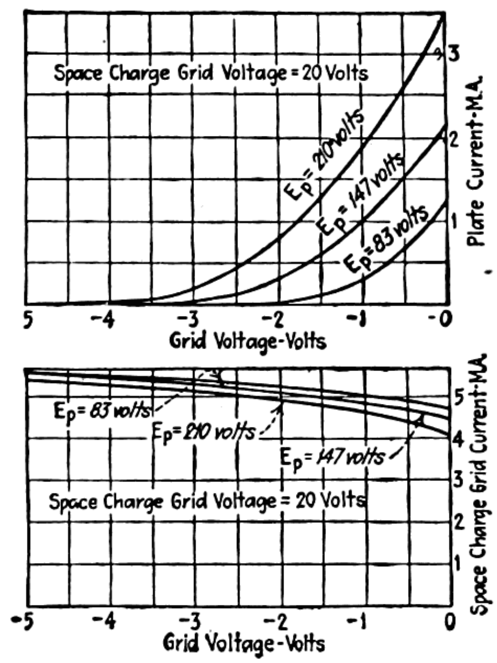


FIG. 31.—Characteristic curves of space-charge grid triode.

The space-charge-grid arrangement provides a virtual cathode with a large area located very close to the control grid. This gives a high transconductance in proportion to the plate potential. At the same time the characteristic curves of the tube tend to have excessive curvature when considered over an appreciable range of grid voltage. Thus space-charge-grid tubes have small power-handling capacity. Furthermore, the space-charge grid draws a very heavy current, usually more than half of the total space current.

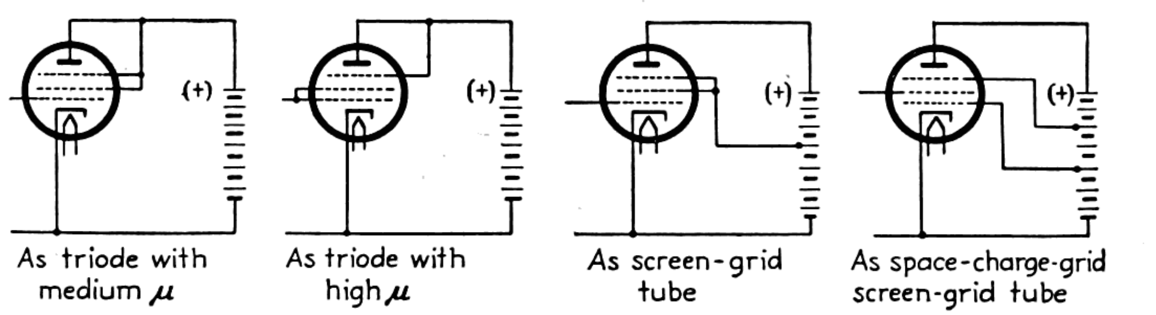


FIG. 32.—Pentode tube arranged in various ways.

Special Connections for Conventional Tubes.—It is possible to operate conventional tubes to give special characteristics, either by employing special connections or by the proper combination of electrode voltages. Thus a pentode tube can be connected to operate as a triode with medium or high amplification factor, as a screen-grid tube, or as a space-charge-grid tube, as shown in Fig. 32.

Another rearrangement of a conventional tube is to interchange the functions of the grid and plate by making the grid the anode electrode, and by using the plate as in the plate circuit is negligibly small.[1] One then has

$$\frac{\partial g_m}{\partial E_g} = \frac{4\Delta I}{E^2} \tag{102}$$

$$\frac{\partial^2 g_m}{\partial E_g{}^2} = \frac{24 I_3}{E^3} \tag{103}$$

where ΔI is the change in d-c plate current, I_3 the crest value of the third-harmonic component of the plate current, and E is the crest value of the signal voltage applied to the grid.

The Application of Exact Equivalent Circuit to Simple Cases.—The first-, second-, and third-order components of the equivalent voltage in the exact equivalent circuit for the case of signals consisting of a sine wave and of two superimposed sine waves are summarized in Table 5. An examination of this table shows that the first-order effects account for the undistorted part of the amplified output. The second-order effects give rise to second harmonics, to sum and difference frequencies, and to a direct-current or "rectified" component, all of which are proportional to the square of the signal. The third-order effect gives rise to third harmonics and odd-order combination frequencies. There is also a third-order component having the same wave shape as the applied signal (*i.e.*, as the first-order output) but an amplitude proportional to the cube of the signal, so that when this combines with the first-order part of the output, the result is a lack of proportionality between the amplitude of the input signal and the amplitude of the output. Finally, when the applied signal consists of more than a single frequency, the third-order action causes the part of the output that is at one frequency to depend upon the amplitude of the other components present, giving rise to cross-modulation.

Fourth, sixth, and other higher even-order components give effects similar in character to those produced by the second-order action, namely, even-order harmonics, even-order combination frequencies, and a rectified d-c current. Fifth, seventh, and other higher odd-order components give effects similar to those produced by the third-order action, *i.e.*, odd harmonics, odd-order combination frequencies, lack of proportionality between input and output, and cross-modulation.

Cross-modulation (Cross Talk).—Cross-modulation is said to be present in a circuit when the conditions are such that when two signals of different frequencies are applied, the amplitude of the output of one frequency depends upon the amplitude of the other frequency that is present. In the case where one of the signals is a modulated wave, cross-modulation results in the amplitude of the other wave depending on the envelope amplitude of the modulated wave. This causes the modulation of the first wave to be transferred to the second wave. Cross-modulation occurs in amplifiers as a result of third-order action.

The amount of cross-modulation can be expressed in terms of a *cross-talk* or *cross-modulation* factor that is defined as the percentage modulation that a modulated wave produces on an unmodulated superimposed wave, divided by the percentage modulation of the modulated wave. In the case of pentode and other tubes where the load resistance is small compared with the plate resistance, the cross-talk factor is[2]

$$\left.\begin{matrix}\text{Cross-talk factor} \\ \text{in pentodes}\end{matrix}\right\} = \frac{E_2{}^2}{2g_m}\frac{\partial^2 g_m}{\partial E_g{}^2} \tag{104}$$

where E_2 is the crest amplitude of the carrier applied to the grid, whose modulation is

[1] An alternative procedure for measuring the third-order coefficient is given by E. W. Herold, Simple Methods for Checking R. F. Distortion or Cross-modulation of Pentode Amplifiers, *Electronics*, Vol. 13, p. 82, April, 1940.

[2] Stuart Ballantine and H. A. Snow, Reduction of Distortion and Cross Modulation in Radio Receivers by Means of Variable-mu Tetrodes, *Proc. I.R.E.*, Vol. 18, p. 2102, December, 1930.

being transferred to the other carrier frequency. In triode tubes the cross-talk factor depends upon the load impedance to the first- and second-order currents, as well as upon the square of the carrier amplitude and the second derivative of the transconductance.

Distortion of Modulation Envelopes.—It was noted above that third-order action introduces a nonlinear relation between output and input voltage of an amplifier. When the input voltage is a modulated wave, the third-order action therefore causes the output voltage to fail to follow exactly the amplitude variations of the input voltage. This results in a distortion of the modulation envelope, and also makes the degree of modulation of the output voltage differ from that of the applied signal. When the modulation of the applied signal is sinusoidal, and the degree of modulation m is not too great, the second-harmonic distortion introduced in the modulation envelope by third-order action in the case of pentode tubes is

$$\left.\begin{array}{l}\text{Second-harmonic}\\ \quad\text{component of modulation}\end{array}\right\} = \frac{m}{2 + \dfrac{4g_m}{E_0^2 \dfrac{\partial^2 g_m}{\partial E_g^2}}} \tag{105}$$

where E_0 is the carrier amplitude of the applied signal, and the remaining notation is as above. It will be noted that when the third-order action is large, the distortion approaches $m/2$. Practically, it is found that when modulation distortion is enough to be of significance, cross-modulation is excessive. Cross talk rather than modulation distortion accordingly limits the amount of third-order action permissible over the operating range of an amplifier under ordinary conditions.

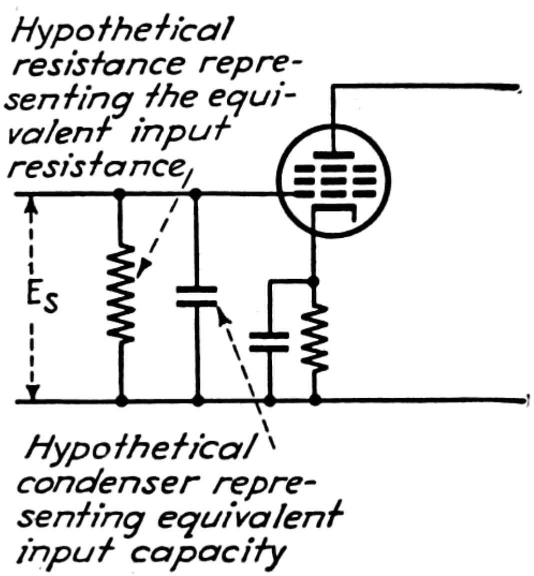


FIG. 89.—Input impedance of a vacuum-tube represented as a resistance shunted by a capacity.

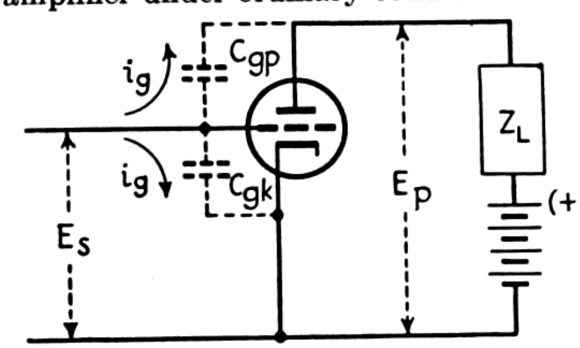


FIG. 90.—Diagrams illustrating how the interelectrode capacities permit grid current to flow when alternating voltage is applied to a negative grid.

25. Input Admittance and Output Impedance of Vacuum-tube Amplifiers. *Input Admittance.*—The input admittance of a vacuum-tube amplifier is defined as the current flowing into the control-grid electrode divided by the voltage that is applied between this electrode and the cathode. The input admittance of a vacuum-tube amplifier can be represented by a capacity shunted by a resistance, as shown in Fig. 89. The capacity is called the input capacity, and the resistance is termed either the input resistance or the input conductance.

Input Admittance of Triodes with Particular Reference to Interelectrode-capacity Effects.—The input capacity and resistance of triodes is determined primarily by the capacities existing between the electrodes of the tube. When an alternating signal voltage is applied between grid and cathode, current flows from grid to cathode and from grid to the plate as a result of the electrostatic capacity between these electrodes, as indicated in Fig. 90. The input admittance is equal to this total current divided by the voltage applied to the control grid. Since the part of the current that flows to the plate depends upon the difference between the signal applied to the control grid

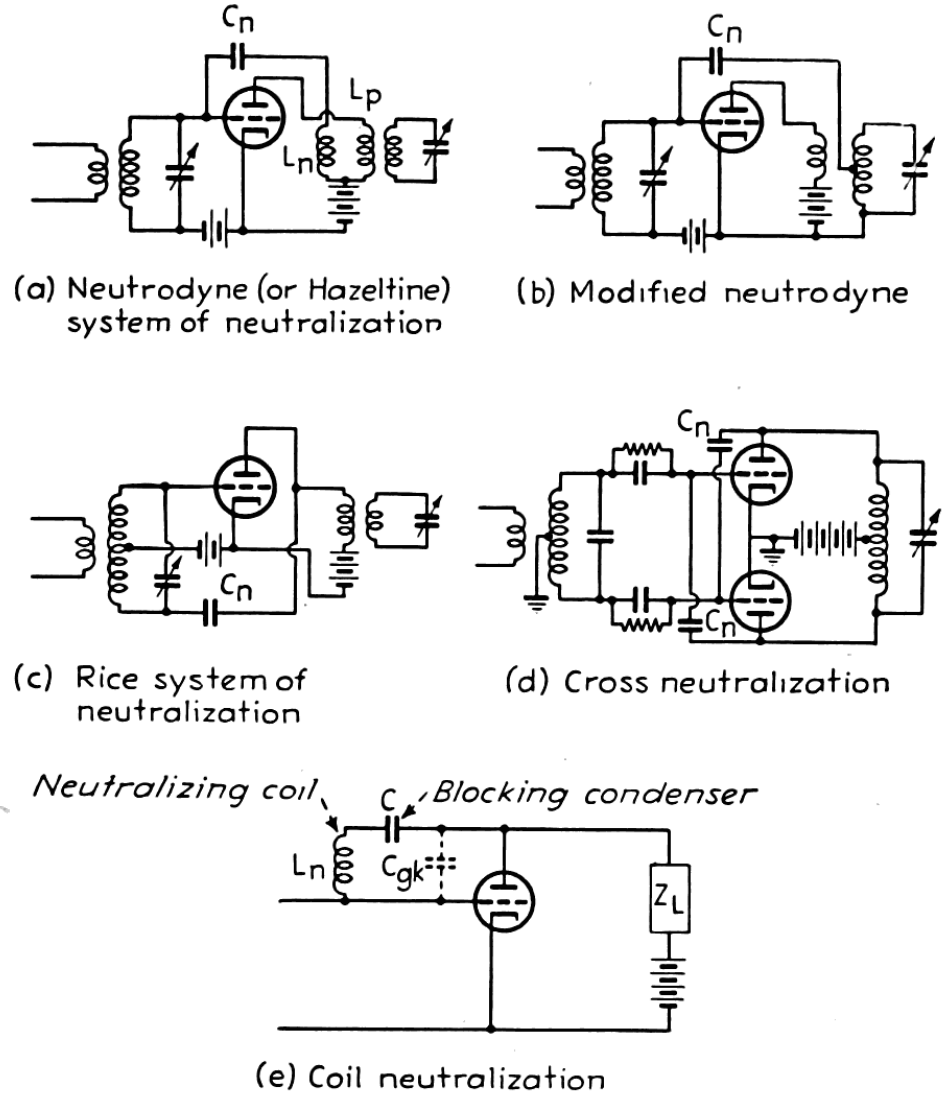


FIG. 91.—Typical neutralizing circuits.

and the amplified voltage developed in the plate circuit, the input admittance is affected by the magnitude and phase of the amplification. The exact relation is[1]

$$\text{Input capacity} = C_{gk} + C_{gp}(1 + A\cos\theta) \tag{106}$$

$$\text{Input resistance} = -\frac{\left(\dfrac{1}{\omega\, C_{gp}}\right)}{A\sin\theta} \tag{107}$$

where C_{gp} = grid-plate tube capacity.

C_{gk} = grid-cathode tube capacity.

A = voltage amplification of the tube alone, not taking into account any step-up in voltage existing in the load impedance, such as is present in the case of transformer coupling.

θ = phase angle of load impedance in the plate circuit, taken positive for inductive loads.

[1] For derivation, see F. E. Terman, "Radio Engineering," 2d ed., p. 232, McGraw-Hill, New York. See also J. M. Miller, Dependence of the Input Impedance of a Three-electrode Vacuum Tube upon the Load in the Plate Circuit, *Bur. Standards Sci. Paper* 351; M. von Ardenne and W. Stoff, On the Values and the Effects of Stray Capacities in Resistance-coupled Amplifiers, *Proc. I.R.E.*, Vol. 15, p. 895, November, 1927; E. L. Chaffee, Equivalent Circuits of an Electron Tube and the Equivalent Input and Output Admittances, *Proc. I.R.E.*, Vol. 17, p. 1633, September, 1929.

The input admittance of a triode amplifier depends largely upon the magnitude A and phase θ of the amplification, because with practical tubes the grid-plate capacity has approximately the same magnitude as the grid-filament capacity, and the amplification is normally considerably greater than unity. Under these conditions the input capacity is approximately proportional to the amplification, and is much greater than the interelectrode capacities themselves.

The input resistance takes into account the energy transferred between the plate and grid electrodes through the grid-plate capacity. With capacitive load impedances in the plate circuit (θ negative), energy is transferred directly from the grid to the plate circuit, with the result that the input resistance is positive. On the other hand, with inductive load impedances in the plate circuit (θ negative), amplified energy is transferred from the plate circuit of the tube back to the grid circuit, with the result that the equivalent input resistance is negative. A resistance load impedance in the plate circuit ($\theta = 0$) gives rise to infinite input resistance, corresponding to zero energy transfer between input and output electrodes. The input resistance for a given phase shift θ is inversely proportional to the product of frequency and amplification.

Neutralization of Grid-plate Capacity in Triode Tubes.—Most of the input capacity and practically all the input resistance of triode tubes results from the current that flows through the grid-plate capacity of the tube. The effect of this capacity can be eliminated or neutralized by the arrangements illustrated in Fig. 91. In the neutrodyne circuit the coil L_n is closely coupled to the primary, and so polarized that it applies a voltage to the neutralizing condenser C_n that is of opposite phase from the a-c voltage between plate and cathode. The current through the neutralizing condenser is hence of opposite phase from the current through the grid-plate capacity C_{gp} of the tube, and by proper adjustment of C_n the energies represented by the two currents can be given the same magnitude. The input capacity of such an amplifier with perfect neutralization is $C_{gk} + C_{gp} + C_n$, and the input resistance is infinite. If the neutralizing coil L_n is closely coupled to its primary and the leads have negligible inductive reactance, the neutralization is substantially independent of frequency over a wide frequency band.[1]

An alternative neutralizing arrangement is the Rice circuit (Fig. 91*c*). Here the neutralizing capacity C_n is adjusted so that the current through it neutralizes the effect of the current through the grid-plate capacity as far as the tuned circuit associated with the grid of the tube is concerned. This form of neutralization is theoretically independent of frequency, just as is the neutrodyne type, but has the disadvantage that only half of the signal voltage developed across the input circuit is applied to the grid of the tube, and that neither side of the tuning condenser in the input circuit can be grounded. Also, if several stages are in cascade, Rice neutralization is more likely to cause trouble from parasitic oscillations.

Cross neutralization is used with push-pull amplifiers and requires the addition of no special circuits other than the neutralizing condensers. It can be thought of as a form of neutrodyne that takes advantage of the fact that the voltages on the two sides of a push-pull amplifier are of opposite polarity, thus giving the phase relations required for neutralizing.

In coil neutralization, the neutralizing inductance L_n is resonated with the grid-plate capacity C_{gp} at the frequency for which neutralization is to be effective. In this way the current flowing from control grid to plate is reduced practically to zero, thus eliminating the grid-plate capacity as far as input admittance effects are concerned. In order that the neutralizing coil will not short-circuit the plate-supply

[1] At very high frequencies the variation in the reactance of the leads cannot be neglected. If neutralization is to be maintained over a wide frequency band under these conditions, it is necessary to use modified circuits (see Fig. 19, Sec. 6).

battery, it is necessary to place the blocking condenser C in series with the inductance. The circuit proportions are then arranged so that the combination L_nC offers the inductive reactance required to resonate with C_{gp} at the frequency of neutralization. Coil neutralization is simpler than the other types, but is effective for only one frequency. It is used extensively with transmitters, such as broadcast transmitters, that operate only at a single frequency.

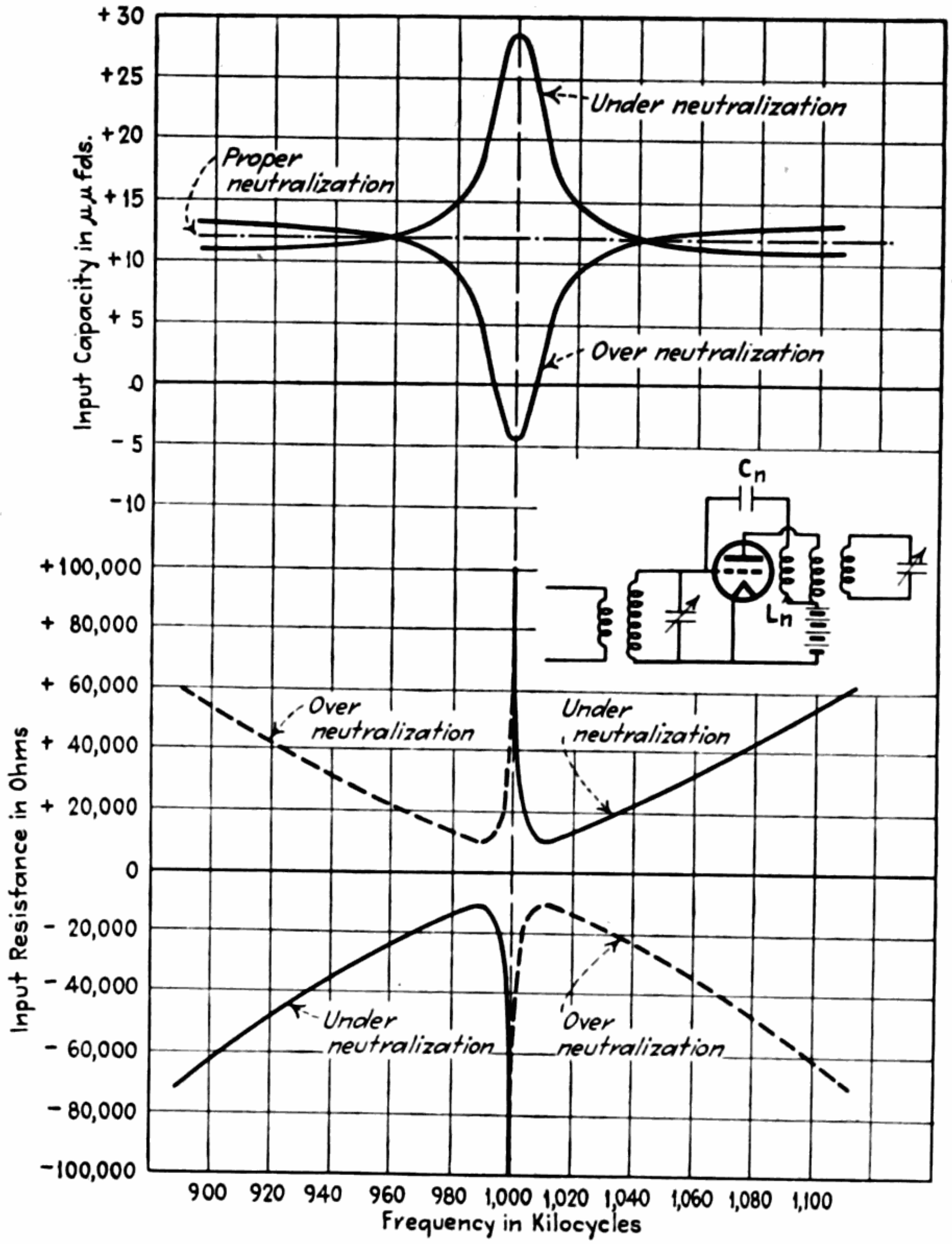


FIG. 92.—Curves showing the effect of an improper neutralizing capacity C_n on input resistance and capacity of a tuned radio-frequency amplifier.

In all neutralizing systems it is necessary to adjust the circuit proportions so that neutralization is exactly correct. Overneutralization produces results similar in character to and just as bad as underneutralization, as illustrated in Fig. 92.

Grounded-grid Self-neutralized Circuit.—The necessity of neutralization can be avoided by means of the grounded-grid circuit of Fig. 93. Here the grid is grounded, the input voltage is applied between cathode and ground, and the output impedance is placed between plate and ground. With this arrangement, capacity currents flowing between plate and grid as a result of the output voltage E_0 developed in the plate circuit do not flow through the input circuit. There is hence no interaction with the source of exciting voltage E_s as a result of grid-plate tube capacity.

Analysis of the voltage and current relations in the grounded-grid circuit shows that since the driving voltage E_s is in series with the external circuit connecting plate and cathode, the tube acts, as far as the output E_0 is concerned, as though it were excited in the usual manner and had an amplification factor $(\mu + 1)$ instead of μ. Also, the exciting voltage delivers some energy directly to the plate circuit, because the amplified plate current flows through E_s in the same polarity as E_s. This places a load on the exciting voltage E_s. This load reduces the gain obtained as compared with a neutrodyne circuit, for example, although appreciable amplification is still possible.

The grounded-grid circuit finds its chief use under conditions where triode tubes must be used, and either (1) it is impractical to obtain the degree of capacity neutralization required for satisfactory operation or (2) the tube or circuit design is such that the natural method of operation is with the grid at ground potential. These conditions are most likely to exist where the tube is to be operated over an enormous frequency range or at a very high frequency, or with special tubes designed for operation at extremely high frequencies.[1]

Input Conductance of Pentodes.—The input conductance of a pentode tube, if negative grid operation is assumed, arises primarily from the transit time of the

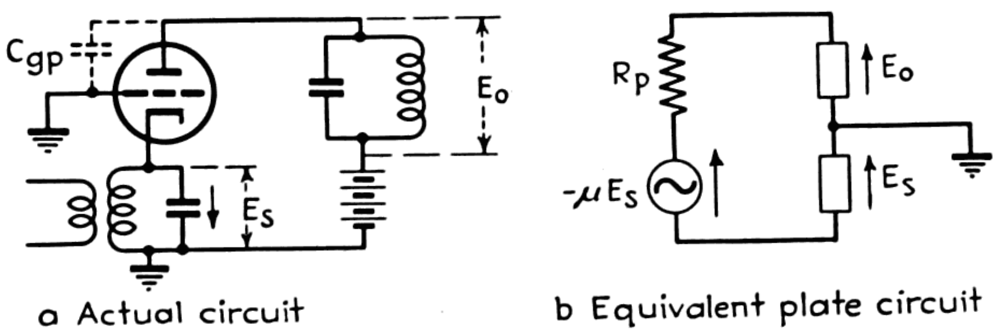


FIG. 93.—Grounded-grid self-neutralized circuit.

electrons and from the fact that the lead from the cathode electrode to the cathode pin on the base of the tube is common to the grid and plate circuits of the tube. Dielectric losses in the grid-cathode and grid-screen capacities are occasionally of importance.[2] The grid-plate capacity contributes very little to the input conductance of a pentode tube, because although this capacity acts the same in pentodes as in triodes, its magnitude is so small (0.005 $\mu\mu$f for a typical voltage amplifier pentode) that the energy transfer taking place between plate and grid is too small to be important.

The input resistance arising from the fact that the electrons require a finite time to travel from cathode to plate is given by[3]

$$\left.\begin{array}{r}\text{Input resistance resulting}\\ \text{from finite transit time}\end{array}\right\} = \frac{1}{Kg_m f^2 \tau^2} \tag{108}$$

where g_m = transconductance of the tube.
f = frequency.
τ = time required for the electron to travel from the cathode to the grid plane.

The constant K is determined by the grid and plate voltages and by the ratio of transit times from cathode to grid plane and grid plane to anode. The component of input resistance arising from transit time is directly proportional to the square root of the electrode voltages and inversely proportional to the square of the linear dimension of the tube.

[1] Discussion of the grounded grid circuits, with particular reference to high-frequency high-power amplifiers, is given by C. E. Strong, The Inverted Amplifier, *Electronics*, Vol. 13, p. 14, July, 1940.
[2] A discussion of dielectric losses is given by C. J. Franks, Measured Input Losses of Vacuum Tubes, *Electronics*, Vol. 8, p. 222, July, 1935.
[3] W. R. Ferris, Input Resistance of Vacuum Tubes as Ultra-high-frequency Amplifiers, *Proc. I.R.E.*, Vol. 24, p. 82, January, 1936. See Par. 9, Sec. 4 for further discussion of transit time effects.

The inductance of the cathode lead causes the voltage existing across the grid-cathode capacity to differ from the signal voltage applied to the tube by an amount equal to the voltage developed across the lead inductance by the amplified plate current (see Fig. 94). This fact causes the current flowing through the grid-cathode capacity to have a component in phase with the applied voltage.[1] The corresponding input resistance, on the assumption that the reactance of the lead inductance L_K is small compared with the grid-cathode capacity C_{gk}, is

$$\left.\begin{array}{l}\text{Input resistance resulting}\\ \quad\text{from inductance of cathode}\\ \quad\text{lead}\end{array}\right\} = \frac{1}{\omega^2 g_m L_K C_{gk}} \tag{109}$$

It will be noted that the input resistance caused by the inductance of the cathode lead varies with frequency and transconductance in exactly the same manner as does the input resistance resulting from transit time action. In ordinary pentode tubes, these two effects are of the same order of magnitude. The combined action of cathode lead inductance and transit time is of great importance at the higher radio frequencies. This is evident from Table 6, which gives results of measurements on input resistance of a pentode tube.

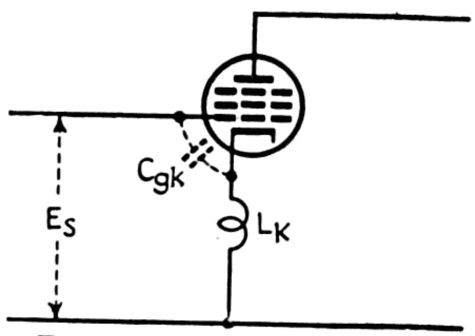


FIG. 94.—The equivalent input circuit of a tube taking into account the inductance of the cathode lead.

The input conductance resulting from the combined effects of cathode-lead inductance and transit time can be neutralized by the arrangement in Fig. 95*a*.[2] Here a small inductance L_n is placed between the cathode terminal on the tube socket and the ground. The grid-return lead is brought directly to the tube socket, and a capacity C_{gg} is connected between grid and ground. Neutralization is obtained when

$$\frac{L_n}{L_k} = \frac{C_{gk}}{C_{gg}} \tag{110}$$

With these proportions, the in-phase component of the current through C_{gg} is equal in magnitude but opposite in phase to the in-phase component of current through C_{gk}, resulting in complete neutralization of the input resistance independently of frequency, provided that $\omega L_k << 1/\omega C_{gk}$. In practical amplifiers, the neutralizing inductance L_n can be a short length of wire of the order of one inch long connecting the cathode terminal of the tube socket to the ground lug, as shown in Fig. 95*b*.

An alternative method of neutralizing the input conductance is to place a small inductance (about $0.1\mu h$) in the screen lead. This method is especially adaptable to converter tubes.

The effect of the cathode lead inductance can be eliminated by using especially constructed tubes having two cathode leads, as shown in Fig. 95*c*.[3] In this way the voltage developed in the cathode lead by the plate current does not appear in series with the voltage applied to the control grid.

Input Capacity of Pentodes, with Particular Reference to the Effect of Control-grid Bias.—The input capacity of a pentode is given by the equation

$$\text{Input capacity} = C_{gk} + C_{gs} + C_{gp}(1 + A\cos\theta) \tag{111}$$

[1] M. J. O. Strutt and A. van der Ziel, The Causes for the Increase of the Admittances of Modern High-frequency Amplifier Tubes on Short Waves, *Proc. I.R.E.*, Vol. 26, p. 1011, August, 1938.
[2] R. L. Freeman, Input Conductance Neutralization, *Electronics*, Vol. 12, p. 22, October, 1939.
[3] Such tubes are discussed by F. Preisach and I. Zakarias, Input Conductance, *Wireless Eng.*, Vol 17, p. 147, April, 1940.

where C_{gk} = grid-cathode capacity.
C_{gs} = grid-screen capacity.
C_{gp} = grid-plate capacity.
A = voltage amplification (magnitude) of tube alone (not including any voltage step-up in plate coupling network).
θ = phase angle of amplification.

The component $C_{gp}(1 + A \cos \theta)$ is the smallest of the three because of the small grid-plate capacity in pentode tubes, but may be of the order of 1$\mu\mu$f in radio-frequency

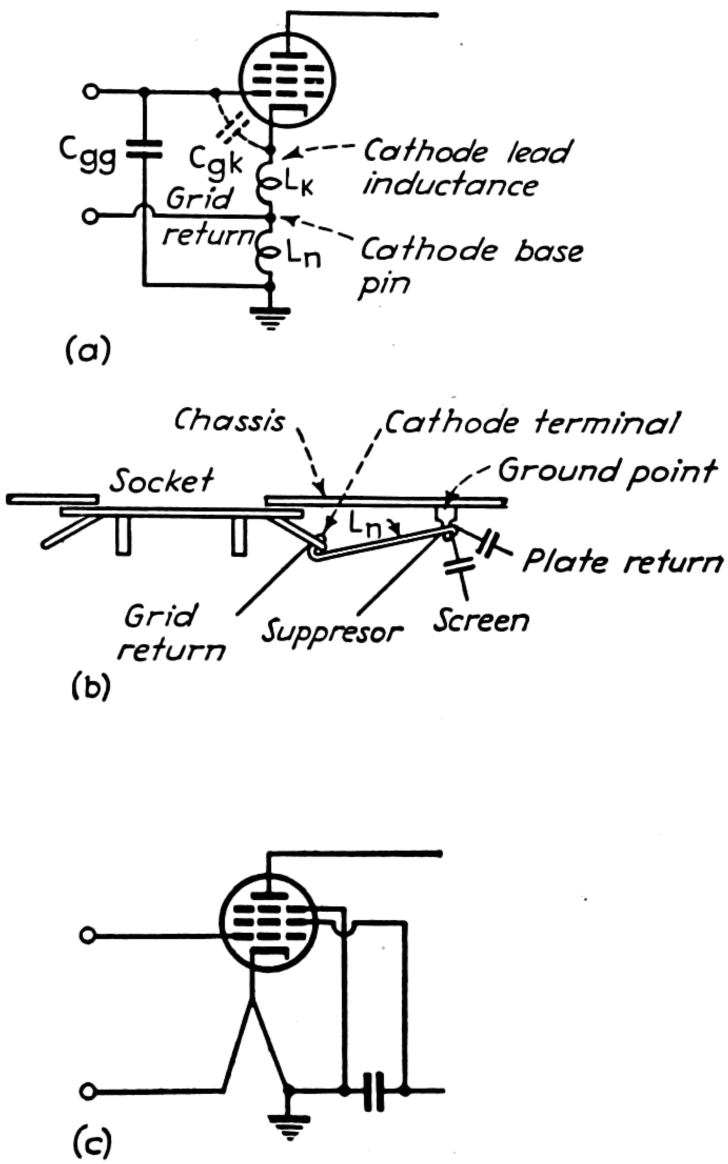


FIG. 95.—Methods of neutralizing the effects of cathode lead inductance and transit time on input resistance.

pentodes when the amplification is large, and somewhat more in audio-frequency pentodes.

The input capacity of a pentode varies somewhat with the grid bias.[1] This is in part due to a change in the component $C_{gp}(1 + A \cos \theta)$ as the grid bias varies the amplification, and in part due to the fact that as the bias varies the plate current the effective position of the space charge surrounding the cathode changes. The total variations produced by varying the grid bias are of the order of 1 to 3 $\mu\mu$f in typical

[1] R. L. Freeman, Use of Feedback to Compensate for Vacuum-tube Input-capacitance Variations with Grid Bias, *Proc. I.R.E.*, Vol. 26, p. 1360, November, 1938; T. Iowerth Jones, The Dependence of the Inter-electrode Capacitances of Valves upon the Operating Conditions, Wireless Section, *I.E.E.*, Vol. 13, p. 11, March, 1938; also, *Jour. I.E.E.*, Vol. 81, p. 658, 1937.

radio-frequency pentodes. This will detune an ordinary 450-kc intermediate-frequency coupling circuit by about 5 kc.

The change in input capacity is approximately proportional to the change in the transconductance of the tube, and so can be neutralized by the arrangements shown in Fig. 96. Here a resistance R_n is placed between cathode and ground and is so related to the capacity C_{gk} between grid and cathode as to satisfy the relation

$$R_n = \frac{\Delta C}{C_{gk} g_m} \tag{112}$$

where g_m is the transconductance for some particular operating point and ΔC is the increase in input capacity of the tube when the transconductance is g_m over the input capacity when the tube biased to cutoff. When the secondary trimmer condenser C_i is returned to ground, as in Fig. 96*b*, the value of resistance R_n required approximates that normally used for self-bias. On the other hand, if C_{gk} is made large by returning the trimmer condenser of the input circuit to the cathode, instead of to ground, as in Fig. 96*a*, the neutralizing resistance R_n is then of the order of 20 ohms with ordinary r-f pentodes used as intermediate-frequency amplifiers.[1]

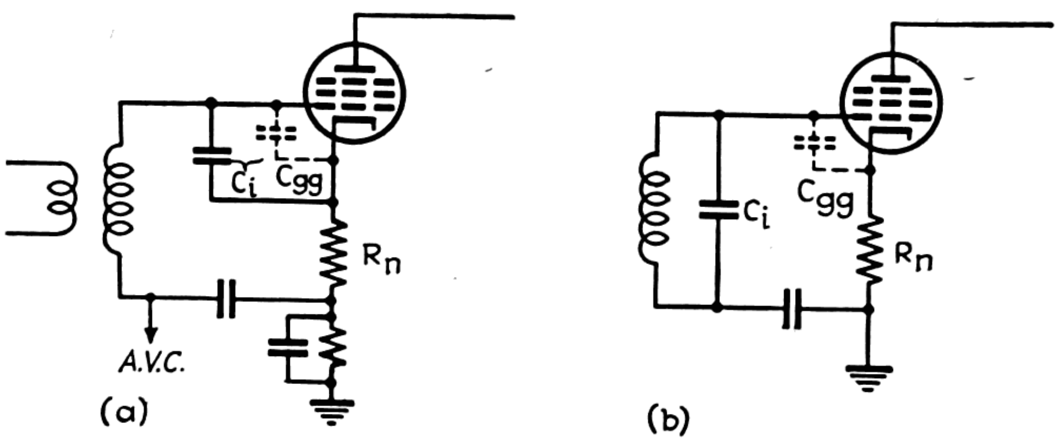


FIG. 96.—Circuits for neutralizing the variation of input capacity of a Class A pentode amplifier with variation in grid bias.

Input Admittance of Amplifier Tube Subjected to a Negative Feedback Voltage.—When a negative feedback voltage is superimposed upon the signal applied to an amplifier, the equivalent input admittance offered by the system to the applied signal voltage is less than in the absence of negative feedback according to the factor

$$\frac{\left\{\begin{matrix}\text{Input admittance with}\\ \text{negative feedback}\end{matrix}\right\}}{\left\{\begin{matrix}\text{Input admittance without}\\ \text{negative feedback}\end{matrix}\right\}} = \frac{1}{1 - A\beta} \tag{113}$$

where $A\beta$ is the feedback factor as defined in connection with Eq. (36). The reduction in input admittance is considerable when $A\beta$ is large, and arises from the fact that most of the signal voltage is used to overcome the negative feedback voltage and only a small fraction is applied across the grid-cathode terminals of the tube.

Practical Importance of Input Admittance in Audio-frequency and Radio-frequency Amplifiers.—In audio-frequency amplifiers, the input capacity is one of the largest shunting capacities in the coupling network in the plate circuit of the preceding ampli-

[1] Another means of compensating for variations in input capacity caused by variation in the control-grid bias by the A.V.C. system is to apply A.V.C. voltage in appropriate amount to the No. 3 grid. See J. F. Farrington, Compensation of Vacuum Tube Input Capacitance Variation by Bias Potential Control, *R.M.A. Eng.*, Vol. 4, p. 13, November, 1937.

fier stage. A pentode is accordingly preferable to a triode tube from the point of view of response obtainable at high frequencies unless the triode is neutralized. In the case of push-pull triodes, neutralization can be easily applied and will reduce the input capacity to the same order of magnitude as that obtained with pentodes, resulting in considerable improvement in high-frequency response.[1] The resistance component of the input admittance is of negligible importance at audio frequencies, since the input resistance is so large as to produce only slight modification of the amplification characteristic, even at the higher audio frequencies with unneutralized triodes.

In tuned radio-frequency amplifiers, the input capacity of the tube is an important factor in fixing the minimum circuit capacity. With unneutralized or partially neutralized triodes, the input capacity varies with frequency, as illustrated in Fig. 92. This dependence of the capacity on frequency can distort the resonance curve of the tuned circuit developing the signal voltage applied to the grid of the tube.

The input resistance at broadcast and lower frequencies is normally caused by the action of the grid-plate tube capacity. With a load impedance supplied by a resonant circuit, the input resistance depends primarily upon frequency, being infinite at resonance and minimum at the 70.7 per cent points of the amplification curve. This minimum value for the case of a simple tuned load circuit is

$$\text{Minimum input resistance} = \frac{(1/\omega C_{gp})}{(A_0/2)} \tag{114}$$

where $1/\omega C_{gp}$ is a reactance of the grid-plate capacity C_{gp} and A_0 is the amplification from grid to plate electrodes at resonance.[2] The input resistance is negative at frequencies below resonance and positive for frequencies greater than resonance, and so tends to distort the resonance curve of the input tuned circuit. If the input resistance is low, oscillations will be produced when the input tuned circuit is resonant at a slightly lower frequency than the output circuit. The effects of the input resistance are so great in tuned amplifiers using triode tubes that such amplifiers must always be neutralized. With pentode tubes, however, the minimum input resistance at broadcast frequencies is so high because of the small grid-plate capacity that it can ordinarily be ignored unless it becomes cumulative through a series of stages.

In the high-frequency and ultra-high-frequency bands, the input resistance is determined primarily by the transit time and cathode lead effects, since these produce effects proportional to the square of frequency, whereas the action of the grid-plate capacity is proportional to frequency. The input resistance at very high frequencies is low enough to produce serious loading, with consequent loss of gain and selectivity.

Output Impedance of Vacuum Tubes.—The output impedance of a vacuum tube is defined as the impedance that the plate circuit of the tube offers to an external voltage applied between the plate and cathode electrodes of the tube. This impedance can be represented by a resistance shunted by a capacity and is the impedance seen by a load looking toward the tube.

In triode tubes the resistance component of the output impedance approximates the plate resistance of the tube up to very high radio frequencies. The output capacity also approximates the sum of the grid-plate and plate-cathode interelectrode capacities. These approximations neglect the effect of dielectric losses, inductance in leads, and energy transfer between the input and output circuits of the tube through the grid-plate capacity, but give satisfactory results for the usual triode operating conditions. In particular, the energy transfer between input and output circuits

[1] P. W. Klipsch, Applying Neutralization to A. F Amplifiers, *Electronics*, Vol. 7, p. 252, August, 1934.

[2] When the tuned load circuit is coupled to the plate circuit of the tube, as with transformer coupling, the value of A_0 is less than the actual voltage amplification of the stage by the factor M/L_s, where M is the mutual inductance between primary and secondary and L_s, the inductance of the secondary.

has a relatively small effect upon the output circuit as compared with the effect on the input circuit because of the difference in power levels.

In pentode and beam tubes the output capacity is equal to the sum of the plate-cathode capacity and the plate-screen capacity. The output resistance, however, depends very greatly upon frequency, being equal to the plate resistance of the tube only at audio and the lower radio frequencies. As the frequency is increased, the output resistance becomes less, first because of dielectric losses associated with the inter-electrode capacities of the tube, and at still higher frequencies as a result of the inductance of the lead wires from the electrodes to the base pins of the tube and capacities between electrodes. The equivalent shunting resistance as measured for a particular pentode tube is shown in Table 6,[1] and it is seen that the output resistance at the higher radio frequencies is far less than the plate resistance of the tube.

26. Tube and Circuit Noise in Amplifiers.—The term *noise* as applied to vacuum-tube amplifiers is commonly used to designate spurious voltages of a random character that represent energy more or less uniformly distributed over an appreciable frequency band. In the case of audio-frequency amplifiers, such noise produces a characteristic hiss.

Thermal Agitation Noise.[2]—The random motion of free electrons in a conductor causes small potential differences to be developed across the terminals of the conductor.

TABLE 6.—INPUT AND OUTPUT RESISTANCE OF A PENTODE AMPLIFIER*

Wave length, meters	Input resistance		Wave length, meters	Output resistance	
	Tube cold, megohms	Normal, megohms		Tube cold, megohms	Normal, megohms
230.0	3.9	3.3	62.5	0.75	0.43
39.5	1.6	0.38	20.4	0.35	0.19
21.2	0.74	0.11	5.05	0.045	0.022
12.4	0.40	0.036			
5.6	0.19	0.0086			

* Data from M. J. O. Strutt and A. Van der Ziel, The Causes for the Increase of the Admittances of Modern High-frequency Amplifier Tubes on Short Waves, *Proc. I.R.E.*, Vol. 26, p. 1011, August, 1938.

This action is termed *thermal agitation*, and the resulting voltage is

$$\left.\begin{array}{l}\text{Square of effective value of}\\ \quad\text{voltage components lying be-}\\ \quad\text{tween frequencies } f_1 \text{ and } f_2\end{array}\right\} = E^2 = 4kT\int_{f_1}^{f_2} R\,df \qquad (115)$$

where k = Boltzmann's constant = 1.374×10^{-23} joule per °K.

T = absolute temperature, °K.

R = resistance component of impedance across which the thermal agitation is developed (a function of frequency).

f = frequency.

[1] From Strutt and van der Ziel, *loc. cit.*

[2] J. B. Johnson, Thermal Agitation of Electricity in Conductors, *Phys. Rev.*, Vol. 32, p. 97, July, 1928; H. Nyquist, Thermal Agitation of Electronic Change in Conductors, *Phys. Rev.*, Vol. 32, p. 110, July, 1928; J. B. Johnson and F. B. Llewellyn, Limits to Amplification, *Elec. Eng.*, Vol. 53, p. 1449, November, 1934; F. C. Williams, Thermal Fluctuations in Complex Networks, Wireless Section, *I.E.E.*, Vol. 13, p. 53, March, 1938, *Jour. I.E.E.*, Vol. 81, p. 751, 1937; F. C. Williams, Coexistent Thermal and Thermionic Fluctuations in Complex Networks, Wireless Section, *I.E.E.*, Vol. 13, p. 327, September. 1938; *Jour. I.E.E.*, Vol. 83, p. 76, 1938.

In the special case where the resistance component of the impedance is constant over the range of frequencies from f_1 to f_2, Eq. (114) reduces to the much simpler form

$$E^2 = 4kTR(f_2 - f_1) \tag{116}$$

In a network involving two or more conductors at different temperatures, the combined thermal agitation effect can be obtained by considering that each impedance involved acts as a generator having the mean square voltage specified by Eq. (115) in series with the impedance.[1] The mean square thermal agitation voltage is by Eq. (116) proportional to the *resistance component* of the impedance across which the voltage is developed, and to be proportional to the band width. The peak amplitude is of the order of 3 to 4 times the equivalent rms value.[2] The energy represented by the thermal agitation voltages developed across a given resistance is uniformly distributed over the entire frequency spectrum from zero frequency to frequencies well above the highest used in communication.

The voltages developed by thermal agitation set a limit to the smallest voltage that can be amplified without being lost in a background of noise. This limit is determined by the band width being amplified and not by the position in the frequency spectrum at which the band is located. The magnitude of the effect can be estimated from the fact that the rms thermal agitation voltage developed across a ½-megohm resistance at 300°K is 6.4 μv for a frequency band of 5,000 cycles.

Noise from Granular Resistances.[3]—Resistances composed of carbon granules (such as an ordinary carbon resistor) generate noise far in excess of the thermal agitation noise when a direct current is passed through the resistance. This high noise arises from fluctuations in the contact resistance between adjacent granules. The noise voltage that results is proportional to the current, and also tends to increase somewhat faster than the resistance.

This effect makes the ordinary carbon resistor unsatisfactory as a plate-coupling resistance in low-level stages of a resistance-coupled amplifier. Carbon resistors are, however, always suitable for bias and screen-grid voltage dropping resistors where by-pass condensers are used.

Tube Noise.—Noise voltages are generated within tubes as a result of a number of actions, the most important of which are (1) random variations in electron emission from the cathode; (2) random variations in the current division between the plate and other positive electrodes, such as the screen grid (this effect is absent in triodes operated with the grid negative); (3) variations in the grid current resulting from positive-ion current. These effects are discussed in Pars. 5 and 13, Sec. 4, where formulas for calculating their magnitudes are also given. Under practical conditions the smallest voltage that can be amplified is sometimes limited by tube noise rather than thermal agitation.

27. Hum.—The term hum is applied to alternating currents appearing in the output of an amplifier as a result of the effect of power-frequency voltages, currents, and fields.

Hum in Audio-frequency Amplifiers.—In audio-frequency amplifiers, hum results from the introduction into the amplifier circuits of currents of the power frequency and its harmonics that are amplified directly by the amplifier. Hum is particularly

[1] F. C. Williams, The Representation and Computation of Fluctuation Voltages, Wireless Section, *I.E.E.*, Vol. 14, p. 325, September, 1939, *Jour. I.E.E.*, Vol. 85, p. 280, 1939.

[2] V. D. Landon, A Study of the Characteristic of Noise, *Proc. I.R.E.*, Vol. 24, p. 1514, November, 1936; The Distribution of Amplitude with Time in Fluctuation Noise, *Proc. I.R.E.*, Vol. 29, p. 50, February, 1941.

[3] C. J. Christensen and G. L. Pearson, Spontaneous Fluctuations in Carbon Microphones and Other Granular Resistances, *Bell System Tech. Jour.*, Vol. 15, p. 181, April, 1936.

ing across C_{gp}. This voltage difference in turn depends upon the amplification E_p/E_g, and will commonly be much larger than the applied signal voltage E_g. As a result, the input admittance of a triode amplifier will be greater than the admittance of $C_{gk} + C_{pk}$; moreover, the input admittance will have a magnitude and phase that depend upon the vector value of the amplification, and hence upon the load impedance in the plate circuit of the tube.

A quantitative analysis of the relations involved in Fig. 12-15*a* leads to the following expression for the input capacitance and input resistance as shown in Fig. 12-15*b*:[1]

$$\text{Input resistance} = R_g = -\frac{1/\omega C_{gp}}{A \sin \theta} \tag{12-27}$$

$$\text{Input capacitance} = C_g = C_{gk} + C_{gp}(1 + A \cos \theta) \tag{12-28}$$

The notation is the same as in Fig. 12-15*a* with the following additions:

$A = |E_p/E_g|$ = magnitude of amplification, not including any voltage transformation that may exist in the load impedance

θ = angle by which the voltage E_p across the load impedance will lead the equivalent voltage $-\mu E_g$ that acts in the plate circuit (θ positive for inductive load impedances)

Examination of Eq. (12-28) shows that the input capacitance of the triode amplifier depends only on the amplification A, the phase shift θ, and the tube capacitances.[2] The input capacitance is independent of the frequency except as frequency affects the magnitude A and phase angle θ

[1] In deriving these relations, one starts by using the voltage E_g as a phase reference and regards the symbols E_g and E_p as representing magnitude only. Then if the angle by which E_p leads $-\mu E_g$ is θ, E_p leads E_g by the angle $(\theta + 180°)$. With these definitions the vector value of the voltage across the grid-plate tube capacitance C_{gp} is $(E_g - E_p\underline{/\theta + 180°})$, and the current flowing from grid to plate as a result of this voltage across C_{gp} is $j\omega C_{gp}(E_g - E_p\underline{/\theta + 180°})$. The current flowing from grid to cathode through the grid capacitance C_{gk} is $j\omega C_{gk}E_g$, so that the total grid current, which is the sum of these, is

$$\begin{aligned}\text{Total grid current} &= j\omega C_{gk}E_g + j\omega C_{gp}(E_g - E_p\underline{/\theta + 180°}) \\ &= \omega C_{gk}E_g\underline{/90°} + \omega C_{gp}(E_g\underline{/90°} - E_p\underline{/\theta + 270°})\end{aligned}$$

This total current divided by the voltage E_g gives the admittance of the grid, which is therefore

$$\text{Admittance of grid} = \omega C_{gk}\underline{/90°} + \omega C_{gp}[1\underline{/90°} - (E_p/E_g)\underline{/\theta + 270°}]$$

The real part of this admittance represents the input conductance (i.e., the reciprocal of the input resistance), while the quadrature part is the input susceptance, which when divided by ω gives the input capacitance. Equations (12-27) and (12-28) are merely these two components of the input admittance with $|E_p/E_g|$ denoted by the symbol A.

[2] The influence of amplification on the input capacitance of a triode is sometimes termed the Miller effect, after John M. Miller, who first recognized and studied this phenomenon.

In some types of cathodes there is a form of shot effect, termed flicker effect, that arises as a result of random changes of emission over small cathode areas. With oxide-coated cathodes the flicker effect is normally greater than the true shot effect.

Positive ions produced in the tube as a result of ionization of residual gas, or as a result of the occasional emission of positive ions by the cathode, also give rise to a modified form of shot effect. With tubes having a good vacuum the noise introduced in this way with ordinary space charges is of the same order of magnitude as the thermal-agitation noise occurring in the plate resistance of the tube.

Miscellaneous Sources of Noise.—In addition to the factors discussed above, there are a number of additional ways by which noise can be created. Thus poor contacts, leaky condensers, faulty resistances, etc., very commonly result in the introduction of noise.

Carbon resistors carrying direct current are also particularly troublesome sources of noise. Such noise arises from fluctuations in the contact resistance between adjacent granules and is similar in character to the "hiss" occurring in carbon microphones. The magnitude of the effect is so great that carbon resistors cannot be used as plate-coupling resistances in amplifiers that must amplify even moderately small voltages.

46. Input Admittance of Amplifiers and the Neutralization of Grid-plate Capacitance.—The input admittance of an amplifier is defined as the admittance that is seen when looking toward the control-grid electrode of the tube. This is the admittance across which the voltage to be amplified is applied.

With screen-grid and pentode tubes arranged to have complete electrostatic shielding between control-grid and plate electrodes, the input admittance is supplied by the sum of the grid-screen and grid-cathode capacitances. This is commonly 3 to 10 $\mu\mu$f and is independent of the conditions existing in the plate circuit of the amplifier.

In tubes having direct capacitance between control grid and plate, such as exists in triodes and also in screen-grid and pentode tubes having incomplete shielding, the input admittance is also influenced by the conditions in the plate circuit of the tube. This is because the amount of current that flows from the control-grid to the plate electrode through the direct capacitance between these electrodes is determined by the potential difference between grid and plate, and this potential difference obviously depends upon the amplified voltage developed across the load impedance in the plate circuit. Since the amplified voltage developed in the plate circuit is normally much greater than the signal voltage, it is apparent that the potential difference between grid and plate will be very high as compared with the signal voltage actually applied to the grid. This situation causes the current flowing from grid to plate to be unusually large. Direct capacitance between grid and plate is hence of very great importance in determining the input admittance.

When the load impedance in the plate circuit is a resistance, the amplified voltage developed in the plate circuit is exactly 180° out of phase with the signal voltage applied to the grid. Under these conditions the voltage difference between grid and plate is $E_s + E_p = E_s(1 + A)$, where A is the voltage amplification between the grid and plate electrodes of the tube (but does not include any transformer step-up). The current that flows from grid to plate is equal to this voltage divided by the capacitive reactance of the grid-plate capacitance and so is $E_s(1 + A)\omega C_{gp}$. Since this current is supplied by the signal voltage E_s, it represents an effective input capacitance of $(1 + A)C_{gp}$.

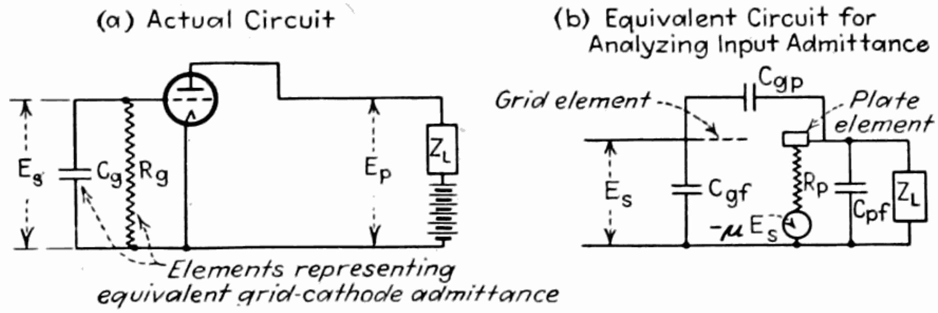


FIG. 73.—Equivalent circuit for analyzing amplifier input admittance. The input admittance is the ratio of current entering the grid electrode to the signal voltage applied to the grid and depends upon the load impedance in the plate circuit.

When the load impedance in the plate circuit is not a resistance, the analysis must be modified to take into account this fact. If the tube is considered as offering to the signal an input capacitance C_g shunted by a resistance R_g, as shown in Fig. 73, one then has

$$\text{Input resistance} = R_g = -\frac{1/\omega C_{gp}}{A \sin \theta} \tag{67}$$

$$\text{Input capacitance} = C_g = C_{gf} + C_{gp}(1 + A \cos \theta) \tag{68}$$

where

C_{gp} = grid-plate tube capacitance
C_{gf} = grid-cathode tube capacitance
A = ratio of voltage developed across load impedance in plate circuit to applied signal (*i.e.*, A is the amplification of tube alone, not taking into account any step-up of voltage in the load)
θ = angle by which voltage across load impedance leads equivalent voltage acting in plate circuit (θ positive for inductive load impedance).

The input resistance in Eq. (67) takes into account the fact that when θ is not zero, energy is transferred directly between grid and plate circuits through the grid-plate capacitance. When the plate load impedance has a capacitive component, the input resistance is positive, meaning that the signal voltage transfers energy directly to the plate circuit. On the other hand, when the plate load impedance has an inductive component, the input resistance is negative. The input circuit then receives some of the amplified energy present in the plate circuit. The magnitude of the input resistance decreases as the frequency becomes greater, because the higher the frequency the greater will be the current flowing through the grid-plate capacitance with a given potential difference. As a consequence, the effects of the input resistance become more important as the frequency is raised.

In audio-frequency amplifiers the input capacitance tends to spoil the high-frequency response. This is particularly the case with high-mu triode tubes, since the large value of A obtained with such tubes makes the input capacitance very high. The resistance component of the input admittance is ordinarily too high to be of much importance at audio frequencies.

In radio-frequency amplifiers the resistance component of the input admittance is particularly important. This is because the input resistance of the tube is in shunt with the tuned circuit supplying the signal voltage that is applied to the tube. In case the input resistance is positive, the effective Q of the tuned circuit will be greatly reduced. On the other hand, if the input resistance is negative, energy will be supplied to the tuned circuit, and oscillations can be expected. It is for this reason that radio-frequency voltage amplifiers ordinarily employ pentode or screen-grid tubes so designed as to give practically perfect electrostatic shielding between grid and plate electrodes. When triode tubes are employed for radio-frequency amplification, it is necessary to neutralize the effect of the energy transfer through the grid-plate capacitance.

Neutralization of the Input Admittance of a Vacuum-tube Amplifier.—The effect of the current that flows through the grid-plate capacitance of a triode tube can be neutralized by arrangements that provide for the flow of an equal and opposite current through an auxiliary (or neutralizing) condenser. The most common methods of doing this are the Neutrodyne and Rice circuits illustrated in Fig. 74. In the form of Neutrodyne circuit shown in Fig. 74, the inductance L_N is closely coupled to the coil L_P but is wound in the opposite direction. The voltage developed by L_N is accordingly proportional to the voltage developed in the plate circuit of the tube but is of opposite polarity. The neutralizing condenser C_N is then adjusted so that the current that flows through it will just neutralize the effect of the current flowing in the opposite direction through the grid-plate capacitance of the tube.

In the Rice system of neutralization the input circuit is split into two parts. By properly adjusting the capacitance of the neutralizing condenser, a voltage developed in the output between plate and ground sends currents into the two parts of the input circuit that cancel each other's effects.

47. Multistage Amplifiers with Particular Reference to Regeneration.—In a multistage amplifier it is possible for energy to be transferred between stages by the grid-plate tube capacitance as discussed above, by stray couplings, or by an impedance common to more than one stage. Such energy transfer is termed *regeneration*, and either modifies the amplification or when of sufficient magnitude will produce oscillations.

In resistance-coupled amplifiers such oscillations are of the multivibrator type (see Sec. 67) and have such a low frequency that when heard in a loud speaker they produce a characteristic "put-put" sound that is termed *motorboating*.

Regeneration in Audio-frequency Amplifiers.—The most important cause of regeneration in audio-frequency amplifiers is energy transfer

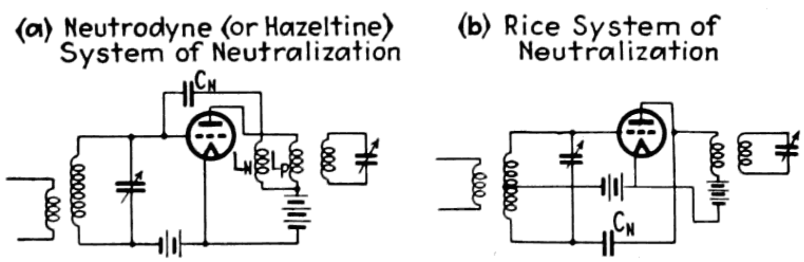


FIG. 74.—Typical circuit arrangements for neutralizing the effect of energy transfer between grid and plate circuits through the grid-plate tube capacitance by an equal and opposite energy transfer through a neutralizing condenser C_n.

between stages as a result of impedance in the source of plate voltage. Thus in Fig. 75*a*, amplified currents in the plate circuit of the final tube will produce a voltage drop in the internal impedance of the plate-supply system. This voltage then acts in the screen and plate circuits of the first tube, resulting in a transfer of energy that modifies the behavior.

The regeneration produced by a common plate impedance is of importance to the extent that the voltage transferred back to the plate and screen circuits of the first tube is of appreciable magnitude compared with the output voltage developed by the first stage in the absence of regeneration. It is not ordinarily necessary to consider regeneration between other tubes than the first and last because of the smaller difference in power levels involved. Regeneration arising from a common plate impedance is most troublesome in amplifiers having very high gain, because then the voltage drop produced in the common plate impedance is largest in proportion to the output voltage of the first tube.

Regeneration from a common plate impedance can be minimized in a number of ways. A large condenser in shunt with the power-supply system will reduce the common impedance, particularly at radio frequency being amplified, the amplifier can be regarded as being located within a waveguide attenuator such as discussed in Sec. 5-8. If the proportions of the enclosing walls or waveguide are such that the attenuation for the dominant mode in a distance equal to the spacing between amplifier stages exceeds the stage gain, then the stray electric and magnetic fields from coils, high potential leads, etc., will attenuate more rapidly with distance than the amplifier gain increases with distance. Under these conditions no further shielding, and no particular care in the detailed arrangement of circuit components, is required.

12-10. Input Admittance of Triode Amplifiers. The input admittance of a vacuum-tube amplifier is defined as the current flowing into the control-grid electrode divided by the voltage that is applied between this

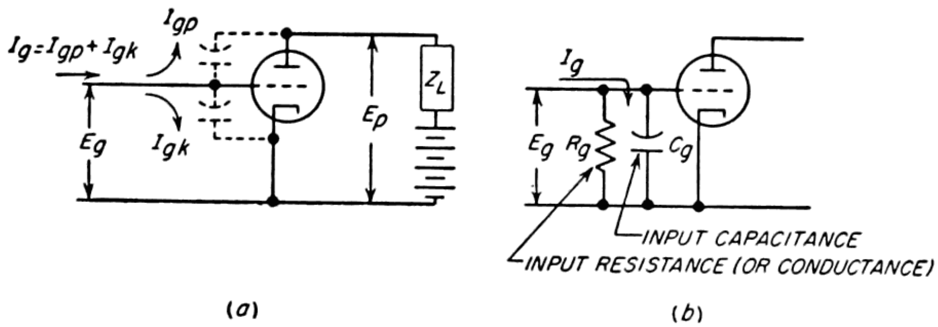


FIG. 12-15. Diagrams illustrating input admittance effects in a vacuum-tube amplifier.

electrode and the cathode. Thus, the input admittance is the admittance that is observed between the grid and cathode terminals when looking toward the tube. The input admittance takes into account the capacitance between the control grid and other electrodes, the transfer of energy between grid and plate circuits through couplings within the tube, and interaction between the potential applied to the control-grid electrode and the electron stream. Furthermore, if the grid is driven positive and so collects electrons, the input admittance is thereby altered, although this action is generally considered as a separate effect.

The input admittance of a vacuum-tube amplifier is normally represented as an equivalent capacitance shunted by an equivalent resistance, as illustrated in Fig. 12-15*b*. This capacitance is called the *input capacitance*, and the resistance is termed either the *input resistance* or *input conductance*. The parallel combination of input capacitance and conductance forms the input admittance.

Analysis of Input Admittance of Triodes. The input admittance of a triode is determined by the currents I_{gk} and I_{gp} that flow through the grid-cathode and grid-plate capacitances, respectively, when a voltage is applied to the control grid, as illustrated in Fig. 12-15*a*. Here I_{gk} represents the current flowing through capacitance C_{gk} as a result of the applied voltage E_g, and requires no discussion. However, the current component I_{gp} representing the current flowing from grid to plate through the grid-plate capacitance C_{gp} depends upon the voltage difference $E_g - E_p$ exist-of the amplification. The maximum possible input capacitance is

$$\left.\begin{array}{l}\text{Maximum possible triode}\\ \text{input capacitance}\end{array}\right\} = C_{gk} + (1 + \mu)C_{gp} \tag{12-29}$$

Here μ is the amplification factor of the tube. This situation results when the load impedance is very much greater than the plate resistance; one then has $A \approx \mu$, and $\theta \approx 0$. The minimum possible input capacitance occurs when the load impedance is zero ($A = 0$), and is

$$\left.\begin{array}{l}\text{Minimum possible triode}\\ \text{input capacitance}\end{array}\right\} = C_{gk} + C_{gp} \tag{12-30}$$

The actual input capacitance will be much larger than this minimum value if the amplification is appreciable.

The input resistance of the vacuum tube may be either positive or negative, as seen from Eq. (12-27). A positive input resistance results when the load impedance in the plate circuit is a capacitive reactance, while a negative resistance is obtained with an inductive load in the plate circuit. *A positive input resistance means that energy is transferred from the grid to the plate through the grid-plate capacitance, while a negative input resistance indicates that the phase relations are such that energy is transferred from the output or plate circuit of the tube to the grid circuit.* The value of input resistance for a given amplification A and phase shift θ varies inversely as the frequency, and may be very low at high frequencies.

Practical Importance of the Input Admittance in Triode Audio and Video Amplifiers. The principal reason that pentode tubes are preferred to high-mu triodes for resistance-coupled amplification and as video amplifiers is that triodes have high input capacitance when the amplification is at all appreciable. In contrast, the input capacitance of pentodes always has a value equivalent to the minimum specified by Eq. (12-30), as discussed below in Sec. 12-12. This high input capacitance of the triode voltage amplifier adversely affects the high-frequency response of the preceding amplifier stage, and so is a serious disadvantage in video voltage amplifiers, and in most resistance-coupled audio-frequency voltage amplifiers as well.

Input capacitance effects do not, however, place triode audio-frequency power amplifiers at an especially serious disadvantage with respect to beam and pentode powes tubes. This is because triode tubes with low to moderate amplification factors are used in power amplification in order to achieve a large plate current with small voltage drop in the tube without driving the grid appreciably positive. Under these conditions the amplification A is not large, and the input capacitance does not exceed the minimum value to anything like the extent that it does in triode voltage amplifiers.

Input Impedance of Tuned Amplifiers. In a tuned amplifier the magnitude A and the phase shift θ of the amplification will vary greatly with frequency. This causes the input conductance and the input capacitance of the triode tuned amplifier to go through corresponding changes, as illustrated in Fig. 12-16 for the case of single tuning. It is to be noted that the input conductance is negative at frequencies below resonance where the plate load impedance is inductive, is zero at resonance where the plate load impedance is resistive, and is positive above resonance where the plate load impedance is capacitive.

The maximum value of input capacitance in Fig. 12-16 is the value given by Eq. (12-28) for resonance, i.e., when $A = A_0$ and $\theta = 0$; thus the "resonant rise" above the base line $C_{gk} + C_{pg}$ is A_0C_{gp}. The peaks of the conductance curve in Fig. 12-16 occur at the 70.7 per cent response points of the amplification curve, and correspond to values of input resistance given by the relation

$$\left.\begin{array}{l}\text{Minimum input}\\\text{resistance}\end{array}\right\} = \pm\frac{1/\omega C_{gp}}{A_0/2} \qquad (12\text{-}31)$$

where A_0 is the amplification of resonance, not including any voltage transformation that may exist in the plate load impedance.[1] The resistance given by Eq. (12-31) can have quite low values at radio frequencies. For example, if $C_{gp} = 2\ \mu\mu\text{f}$ and $A_0 = 15$, then at 1500 kc the minimum input resistance is 7070 ohms. This is a low value compared with the parallel impedance of a resonant circuit that it might shunt.

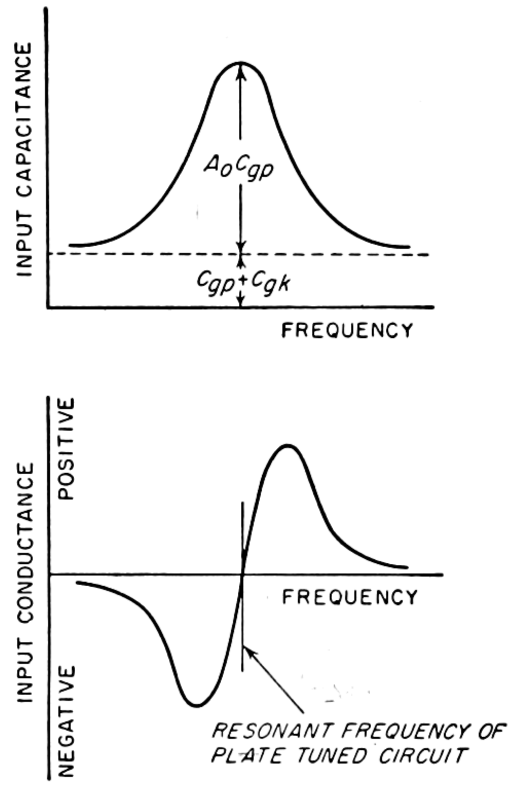


FIG. 12-16. Variation of input capacitance and input conductance of a triode amplifier having a load impedance that is a single resonant circuit.

If a tuned circuit is connected between the grid and cathode of an amplifier tube having an input admittance such as shown in Fig. 12-16, the resonance curve of this tuned circuit is considerably distorted if its resonance frequency is the same (taking into account the input capacitance at resonance) as that of the tuned load impedance in the plate circuit of the amplifier. Alternatively, if the grid tuned circuit is resonant at a slightly lower frequency, then the negative input resistance shunting the grid resonant circuit at its resonant frequency will ordinarily cause the system to break into oscillation. These effects of input admit-

[1] When the tuned load circuit is coupled to the plate circuit of the tube, as with transformer coupling, the value of A_0 is less than the actual voltage amplification of the stage by the factor $(M/L_s)^2$, where M is the mutual inductance between primary and secondary, and L_s is the inductance of the secondary.

tance in tuned amplifiers are so serious that pentode, beam, or screen-grid tubes are employed in tuned amplifiers in preference to triodes whenever possible. If circumstances require the use of triode tubes in tuned amplifiers, it is then necessary either to neutralize the energy transfer taking place between grid and plate circuits of the tuned triode amplifier, or to employ a grounded-grid circuit.

12-11. Neutralization of Input Admittance of Vacuum-tube Amplifiers —Grounded-grid Systems. The effects produced by the transfer of energy between the grid and plate circuits of a vacuum-tube amplifier

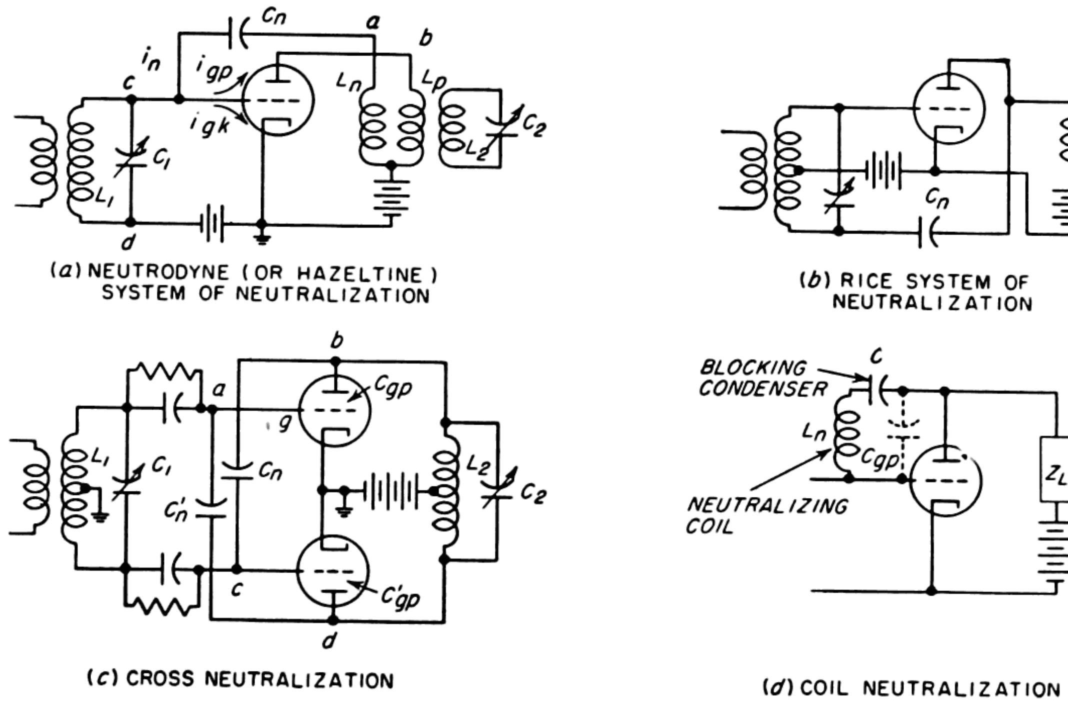
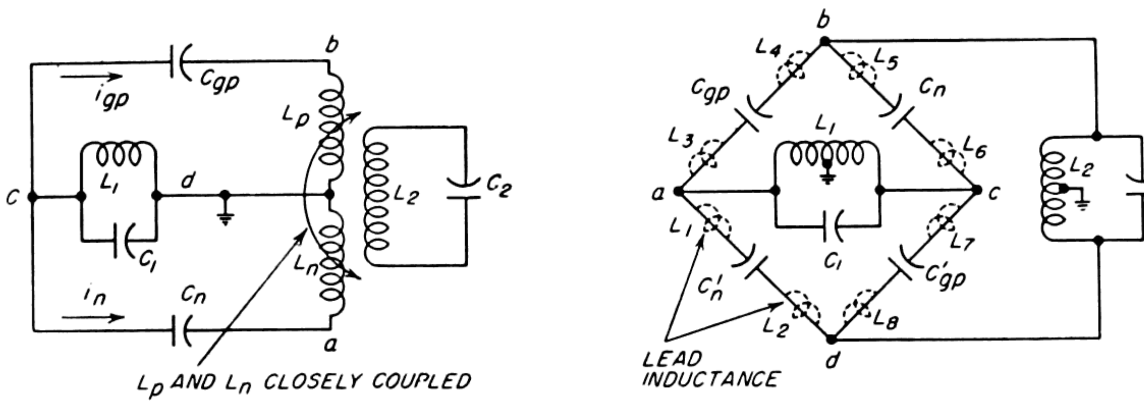


FIG. 12-17. Typical circuits for neutralizing the coupling between the input and output circuits of a triode amplifier arising from the grid-plate capacitance effects.

through the grid-plate tube capacitance can be neutralized by an electrical network that transfers an equal amount of energy in the opposite direction. Examples of such circuits are given in Fig. 12-17*a*, *b*, and *c*.

In each of these a neutralizing capacitor C_n connects the input (i.e., grid) circuit to the output (i.e., plate) circuit in such a way that the current passing through C_n is of the proper amplitude and phase to neutralize exactly the transfer of energy between the input and output circuits of the amplifier *via* the grid-plate tube capacitance. Thus consider Fig. 12-17*a*. This is an ordinary transformer-coupled single-tuned radio-frequency amplifier to which there has been added a neutralizing inductance L_n closely coupled to L_p and connected with a polarity such that the voltage at the end *a* of this coil is in phase opposition to the voltage at the corresponding end *b* of the primary inductance L_p. By making C_n the proper size, the currents i_n and i_{gp} that flow through C_n and the grid-plate capacitance, respectively, are then oppositely affected by the voltages developed in L_n and L_p. The result is complete neutralization of the energy transfer that would otherwise take place between the input and output tuned circuits, through the tube capacitance. The input capacitance of such an amplifier with perfect neutralization is $C_{gk} + C_{gp} + C_n$, and the input resistance is infinite. If the coupling between L_n and L_p is very close, the value of C_n required to give complete neutralization is nearly independent of frequency.

In the circuit of Fig. 12-17*b*, the inductance of the input resonant circuit is provided with a center tap and arranged as shown. If $C_n = C_{gp}$, the current flowing through the neutralizing capacitor C_n from the input circuit to the plate circuit will then be equal and opposite to the corresponding current through C_{gp}. In push-pull amplifiers, neutralization can be accomplished as illustrated in Fig. 12-17*c*. Here the fact that all



(*a*) EQUIVALENT BRIDGE CIRCUIT APPLYING TO FIG 12-17 *a* (*b*) EQUIVALENT BRIDGE CIRCUIT APPLYING TO FIG. 12-17 *c*

FIG. 12-18. Equivalent bridge circuits of neutralized amplifiers of Fig. 12-17*a* and *c*.

circuits are symmetrical with respect to ground makes neutralization particularly simple and effective.

Each of the systems *a*, *b*, and *c* of Fig. 12-17 can be considered as a bridge in which the output and input circuits are connected across the opposite diagonals. Thus Fig. 12-18*a* and *b* shows bridge equivalents of Fig. 12-17*a* and *c*, respectively. When the neutralization is so adjusted as to balance the bridge, the input tuned circuit receives no energy from the output tuned circuit because the two are in electrically neutral locations with respect to each other.

If a system is adjusted to give perfect neutralization at a particular frequency, it is found that the neutralization becomes less effective for frequencies that depart increasingly from the frequency at which the adjustment was made. This is because the inductances of the leads in series with the neutralizing capacitors, shown dotted in Fig. 12-18*b*,[1] cause the equivalent bridge represented by the neutralizing system to be frequency sensitive, particularly at high frequencies. Wideband neutralization can be obtained by adjusting lead lengths and connecting the input

[1] Thus L_1 is the inductance of the lead from junction *a* to C'_n, L_2 the inductance of the lead from C'_n to junction *d*, L_3 the inductance from *a* to the grid electrode, L_4 the inductance from the plate electrode to junction *b*, and so on.

and output leads to points on the system such that these various lead inductances are so distributed in the equivalent bridge circuit that $(L_1 + L_2)/(L_3 + L_4) = (L_7 + L_8)/(L_5 + L_6) = C_{gp}/C'_n = C_n/C'_{gp}$. In this case the bridge balance in Fig. 12-18*c* is not affected by frequency, and the neutralization is hence maintained over a wide frequency band.

Coil Neutralization. When neutralization need be achieved only at a single frequency, the coil neutralizing system of Fig. 12-17*d* is useful. Here the neutralizing inductance L_n is of such size as to resonate with the grid-plate capacitance C_{gp} at the frequency for which neutralization is to be effective. In this way, the current flowing from control grid to plate is reduced practically to zero, thus eliminating the grid-plate capacitance as far as input admittance effects are concerned. In order that the neutralizing coil will not ground the plate-supply potential, it is necessary to place a blocking capacitor C in series with the inductance. Coil neutralization is used extensively in broadcast transmitters.

Grounded-grid Circuits. The necessity of neutralizing a triode amplifier can be avoided by using the grounded-grid circuit of Fig. 12-19*a*. Here the control grid acts as a grounded shield between the output terminals (plate to ground) and the input terminals (cathode to ground). As a result, energy transfer between input and output circuits through the tube capacitances is avoided to the extent that the plate and cathode electrodes and associated leads are so arranged that there is no direct capacitance between them.[1]

The equivalent plate circuit of the grounded-grid amplifier is shown in Fig. 12-19*b*,[2] where the arrows are drawn on the basis that a positive value of input voltage E_g makes the cathode negative with respect to ground, corresponding to the grid positive with respect to the cathode. Study of the relations in this circuit shows that since the signal voltage E_g acts between cathode and ground, it is also in series with the plate circuit of the tube. As a result, the amplified output E_L is the same as though the grid of the tube were excited in the normal manner with a voltage E_g, and the amplification factor had the value $(\mu + 1)$ instead of μ, as shown in Fig. 12-19*c*.

The input resistance of the grounded-grid amplifier is relatively low. This is because the input voltage E_g, in addition to producing a voltage difference between the cathode and grid, also acts directly in the plate circuit and thereby supplies energy that is in part delivered to the plate tuned circuit, and in part dissipated in the plate resistance of the tube.

[1] The effect of any residual plate-cathode capacitance that is present can be eliminated by employing a neutralizing system; see J. J. Muller, Cathode-excited Linear Amplifiers, *Electrical Commun.*, vol. 23, p. 297, September, 1946.

[2] For the sake of simplicity these equivalent circuits neglect the effects of transit time, lead inductance, etc., such as discussed in Sec. 12-12, which can be important at the very high frequencies where grounded-grid amplifiers are of practical importance.

The equivalent load impedance presented by the tube to the input voltage E_g and circuit L_1C_1 is shown in Fig. 12-19d,[1] where Z_L is the impedance of the plate tuned circuit L_2C_2. It is apparent that the impedance offered to E_g is relatively low; this fact limits the power amplification that can be obtained from the grounded-grid circuit to a value much less than that obtainable with a neutralized grounded-cathode amplifier.[2]

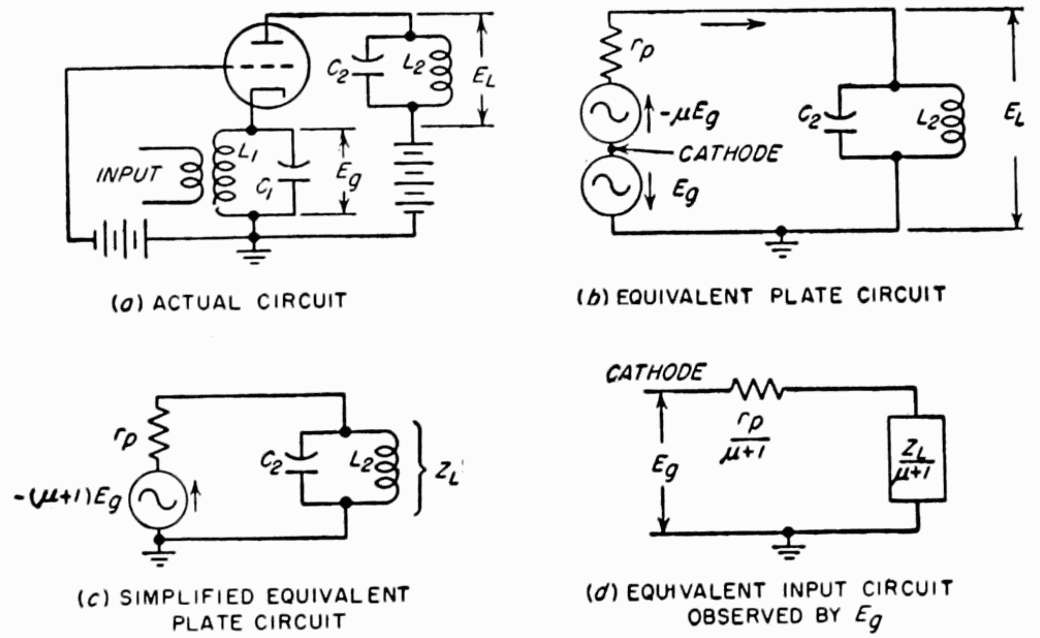


FIG. 12-19. Circuit of grounded-grid amplifier, together with equivalent plate and input circuits.

The grounded-grid triode amplifier finds use as a tuned voltage amplifier at frequencies too high to permit the use of pentode or beam tubes.[3] In these circumstances disk-seal, pencil, or similar closely spaced triodes (see Fig. 6-35) are employed in a grounded-grid circuit using coaxial transmission lines to supply the resonant circuits. Grounded-grid systems using ordinary high-frequency triodes are also often used as Class C and

[1] This equivalent circuit results from the fact that the radio-frequency current flowing through the input circuit L_1C_1 is seen from Fig. 12-19c to be $\mu + 1$ times as great as the current that would be produced by the voltage E_g acting in a circuit consisting of r_p and Z_L in series.

[2] The ratio of amplified to input power, i.e., the power amplification of the grounded-grid circuit, for the case of a resistive load ($Z_L = R_L$) can be readily shown to be

$$\left.\begin{array}{r}\text{Power amplification of}\\ \text{grounded-grid amplifier}\end{array}\right\} = (\mu + 1)\frac{R_L}{R_L + r_p} \qquad (12\text{-}32)$$

[3] Further information on grounded-grid amplifiers in such applications is given by Milton Dishal, Theoretical Gain and Signal-to-noise Ratio Obtained with the Grounded-grid Amplifier at Ultra-high Frequencies, *Proc. IRE*, vol. 32, p. 276, May, 1944; M. C. Jones, Grounded-grid Radio-frequency Voltage Amplifiers, *Proc. IRE*, vol. 32, p. 423, July, 1944; J. Foster, Grounded-grid Amplifier Valves for Very Short Waves, *J. IEE* (*Radiolocation Conv.*), vol. 93, pt. IIIA, p. 868, March–May, 1946; A. E. Bowen and W. W. Mumford, a New Microwave Triode: Its Preformance as a Modulator and as an Amplifier, *Bell System Tech. J.*, vol. 29, p. 531, October, 1950.

linear power amplifiers at high frequencies in preference to neutralized arrangements employing the same tube type.[1] The use of grounded-grid systems at high frequencies is in part because neutralizing systems are increasingly critical of adjustment as the frequency is increased, and in part because the elimination of the neutralizing capacitors reduces the effective input and output capacitances of the tube. This last feature is of particular importance in wideband systems such as linear amplifiers for television transmitters, as it increases the output power and efficiency obtainable from a particular tube for a given bandwidth.[2]

12-12. Input Admittance of Pentode, Beam, and Screen-grid Tubes. *Input Capacitance.* In analogy with Eq. (12-28), the input capacitance of pentode and similar tubes is

$$\text{Input capacitance} = C_{gk} + C_{gs} + C_{gp}(1 + A \cos \theta) \qquad (12\text{-}33)$$

where C_{gs} is the grid-screen capacitance, and the remaining notation is as in Eq. (12-28). It is to be noted, however, that because of the shielding action of the screen grid, the grid-plate capacitance C_{gp} is now a residual stray capacitance much smaller in magnitude than C_{gp} in triodes.

The input capacitance of a tube with a screen grid varies slightly with grid bias. This is in part due to the change in $C_{gp}(1 + A \cos \theta)$ as the grid bias varies the amplification A, and in part due to the fact that the variation in the plate current by grid bias alters the effective position of the space charge surrounding the cathode, and thereby affects C_{gk}. The total variation in input capacitance that can be produced in this way may reach as much as 1 to 3 $\mu\mu$f in some tubes, which is sufficient to produce a noticeable detuning of many tuned amplifiers.[3]

Input Conductance of Pentodes and Similar Tubes. The input conductance of a negative-grid pentode at low and moderate radio frequencies is practically zero, but at very high frequencies cannot be ignored. Coupling between the plate and control-grid circuits through the stray grid-plate capacitance C_{gp} can contribute to the input conductance in accordance with Eq. (12-27). However, the inductance of the cathode lead[4]

[1] For detailed discussions of grounded-grid amplifiers for such purposes, see E. E. Spitzer, Grounded-grid Amplifier, *Electronics*, vol. 19, p. 138, April, 1946; C. E. Strong, The Inverted Amplifier, *Electronics*, vol. 13, p. 14, July, 1940; J. J. Muller, *loc. cit.*

[2] A discussion of the use of grounded-grid power amplifiers for television is given by P. A. T. Bevan, Earthed-grid Power Amplifiers, *Wireless Eng.*, vol. 26, p. 182, June, 1949.

[3] This effect can be neutralized if desired by placing a suitable small resistance, not by-passed, between cathode and ground; see R. L. Freeman, The Use of Feedback to Compensate for Vacuum-tube Input-capacitance Variations, *Proc. IRE*, vol. 26, p. 1360, November, 1938.

[4] The transit-time and cathode-lead inductance effects are also present in triode tubes. However, in such tubes they are so overshadowed by the input admittance effects introduced by the grid-plate capacitance as to be of comparatively little practical importance.

and the transit time of the electron stream are ordinarily the most important causes of input conductance in pentode tubes and also in other tubes having screen grids.

The inductance L_k of the cathode return lead must carry the amplified plate current in a pentode tube, and so has a voltage E_k developed across it. This voltage acts in the input circuit between cathode and control grid as shown in Fig. 12-20, and so makes the voltage difference acting across the grid-cathode capacitance differ from the applied signal voltage E_g by the amount of the voltage drop E_k in the cathode inductance. Since E_k is 90° out of phase with E_g, the result is to produce a component of current flowing through the grid-cathode capacitance that is in phase with the applied voltage E_g, and accordingly causes the input admittance of the tube to have a conductance component. The magnitude of this conductance is[1]

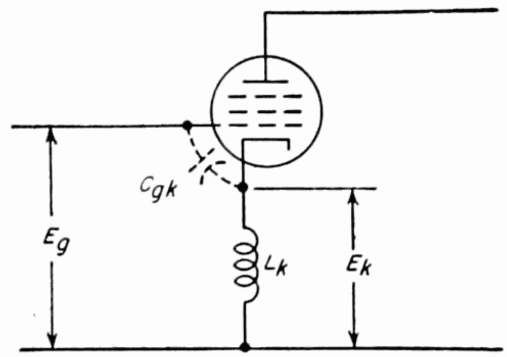


FIG. 12-20. The input circuit of a pentode tube taking into account the inductance of the cathode lead and the capacitance between control grid and cathode.

$$\left.\begin{array}{l}\text{Input conductance resulting}\\ \text{from inductance of cathode lead}\end{array}\right\} = \omega^2 g_m L_k C_{gk} \qquad (12\text{-}34)$$

At very high frequencies the fact that the extremely small length of time it takes an electron to travel from cathode to screen is not negligible compared with the time represented by a cycle of an extremely high-frequency wave causes power to be consumed by the grid of the tube even when the grid is biased negatively and attracts no electrons. Discussed on page 215, this produces an effect equivalent to shunting a conductance [in addition to that of Eq. (12-34)] between control grid and cathode. This transit-time conductance is given by Eq. (6-31) and, like the conductance due to cathode-lead inductance, is proportional to the transconductance and the square of the frequency.

In practical tubes these two effects are commonly of the same order of magnitude. At broadcast frequencies they are negligible but become

[1] Equation (12-34) assumes that the reactance of the lead inductance L_k is small compared with the reactance of the grid-cathode capacitance C_{gk}, and that voltage E_k across L_k is small compared with the applied signal E_g. The derivation is as follows: The voltage E_k developed across L_k by the amplified plate current has a magnitude $g_m\omega L_k E_g$, where g_m is the transconductance of the tube. The voltage E_k acting on the capacitance C_{gk} produces a current $g_m\omega L_k E_g(\omega C_{gk})$. This current is in phase with the applied voltage E_g because the voltage across L_k is 90° out of phase, and there is another 90° phase shift arising from the capacitive reactance of C_{gk}. The resulting input conductance associated with this effect is the current in phase with E_g divided by the grid voltage E_g; Eq. (12-34) is a mathematical expression of this ratio.

increasingly important at higher frequencies, since the conductance varies with the square of the frequency. Thus, a typical general-purpose pentode at 1 Mc may have an input resistance of 20 megohms, but at 100 Mc this becomes only 2000 ohms.[1]

A further factor that can affect the input conductance of pentodes at very high frequencies is the inductance of the screen-grid lead. Amplified screen current flowing through this inductance develops a voltage drop between screen and ground. This voltage in turn introduces a negative component to the input conductance in the same way as does the grid-plate current in a triode possessing an inductive plate load impedance, as discussed above. The resulting negative conductance is proportional to the square of the frequency. By properly adjusting the inductance in the screen circuit it is hence possible to neutralize the positive input conductance arising from cathode-lead inductance and transit time as long as the inductance required to achieve this result is not so large as to resonate with the tube capacitances.

Input Admittance of Amplifier Tubes Subjected to Negative Feedback. When a negative feedback voltage is superimposed upon the signal applied to an amplifier, the equivalent input admittance offered by the system to the applied voltage is less than in the absence of negative feedback according to the factor[2]

$$\frac{\left.\begin{array}{l}\text{Input admittance with}\\ \text{negative feedback}\end{array}\right\}}{\left.\begin{array}{l}\text{Input admittance}\\ \text{without negative feedback}\end{array}\right\}} = \frac{1}{1 - A\beta} \tag{12-35}$$

where $A\beta$ is the feedback factor, as discussed on page 374. The reduction in input admittance is considerable when $A\beta$ has a large negative value, and arises from the fact that most of the signal voltage is used to overcome the feedback voltage, and only a small fraction is applied across the grid-cathode terminals of the tube to produce current flowing to the grid electrode.

12-13. Circuit and Tube Noise. The weakest signal that can be usefully amplified is limited by randomly varying voltages and currents existing in the circuits and tubes of the amplifier.

Resistance Noise. Every electrical conductor produces an irregularly varying voltage across its terminal as a result of the random motion of the free electrons in the conductor caused by thermal action. This effect

[1] When the combined effects of transit time and cathode-lead inductance are not too great, it is possible to neutralize them by methods given by R. L. Freeman, Input Conductance Neutralization, *Electronics*, vol. 12, p. 22, October, 1939; or see F. E. Terman, "Radio Engineers' Handbook," p. 472, McGraw-Hill Book Company, Inc., New York, 1943.

[2] This relation applies to triode amplifiers, as well as those employing pentode and similar tubes.