CHAPTER 12

NOISE IN VACUUM TUBES

12.1. Noise as a Limiting Factor in the Ultimate Sensitivity of Electronic Devices. Vacuum-tube amplifiers making use of triodes and pentodes are capable of giving extremely large amplification of power and voltage. In fact, it may be said that an amplification of any desired magnitude may be achieved by the use of vacuum tubes. At first glance this would seem to imply that arbitrarily small signals could be detected. Ultimately, however, it is found that there is a limit determined by the noise generated by the random motion of electrons at the input of the circuit. Any signal whose level is appreciably less than that of the electron noise will be masked by it. The order of the equivalent voltage of the electron noise is extremely small, of the order of millimicrovolts, but many electronic devices have enough amplification to bring this up to a detectable level.

Electron noise shows up in both resistors and in vacuum tubes. Even in a passive resistor, the molecular and the electronic agitation is evident with sufficient amplification. Here the noise is referred to as "thermal-agitation noise." In vacuum tubes the random emission and fluctuation of space-charge-limited currents contribute a similar noise. In temperature-limited tubes this noise is called "shot noise" and is due to random emission. In space-charge-limited emission tubes the noise is much less and is called "reduced shot noise." Both types of noise are characterized by a uniform distribution of energy over the frequency spectrum. Depending upon the application, the noise from either the tubes or the resistors at the input of an electronic device may predominate.

Needless to say, resistor and tube noise is an exclusive concern of electronic devices. No other type of device can have sufficient sensitivity to be limited by random electron motion. Resistor and tube noise set the ultimate sensitivity of high-gain amplifiers, receivers, phototube input circuits, and television pickup tubes. Although resistor and tube noise can never be avoided, much can be done by circuit design and selection of tubes to approach the minimum attainable noise.

Although the formulas for various types of electron noise and their application are quite simple, their derivation is dependent upon some

intimate aspects of thermodynamics and the kinetic theory of gases that have not been developed in this book. For this reason, only the basis of the development will be given, and emphasis will be upon the interpretation and application of the formulas.

12.2. Noise in Resistors. Noise in resistors is due to the random motion of electrons within them. The noise energy is proportional to the resistance, the absolute temperature, and the frequency band width over which the noise is observed and is independent of the material of which the resistor is made. The noise energy increases with absolute temperature because the molecular agitation is proportional to this. The noise is probably made up of extremely short and sharp pulses resulting from the impact of the electrons with the molecules. These pulses are probably so short and sharp that they are made up of a continuous distribution of frequency components of equal amplitude up to the highest frequencies in use today. As a result, the noise energy is uniformly distributed over the useful r-f spectrum.

The mean-square thermal-agitation noise voltage $\bar{e}^2$ across the terminals of a resistor R at an absolute temperature T, associated with a frequency band width B, is

$$\bar{e}^2 = 4kT_rRB \tag{12.1}$$

where e is rms value of the noise voltage, volts
k is Boltzmann's constant, 1.3805×10^{-23} watt-second per °K
R is resistance, ohms
T_r is room temperature, °K (°C + 273)
B is frequency band width, cycles per sec[1,2]

If room temperature is taken as 290°K (63°F), the expression for the rms noise voltage becomes

$$e_{\text{rms}} = 126.0\sqrt{RB} \quad \text{micromicrovolts} \tag{12.2}$$

A nomograph of this equation is shown in Fig. 12.1. For the sample construction line shown, the rms value of the noise voltage across a 1,000-ohm resistor in a frequency band of 10 mc is 12.6 microvolts.

The effect of thermal-agitation noise can be expressed either as an emf in series with the resistor considered noiseless or as a constant-current generator in parallel with the resistor considered noiseless. This follows from application of Thévénin's and Norton's theorems. The two equivalent circuits of a noisy resistor are shown in Fig. 12.2. The circle in the figure indicates a zero-impedance constant-voltage generator

[1] JOHNSON, J. B., Thermal Agitation of Electricity in Conductors, *Phys. Rev.*, vol. 32, pp. 97–109, July, 1928.

[2] NYQUIST, H., Thermal Agitation of Electric Charge in Conductors, *Phys. Rev.*, vol. 32, pp. 110–113, July, 1928.

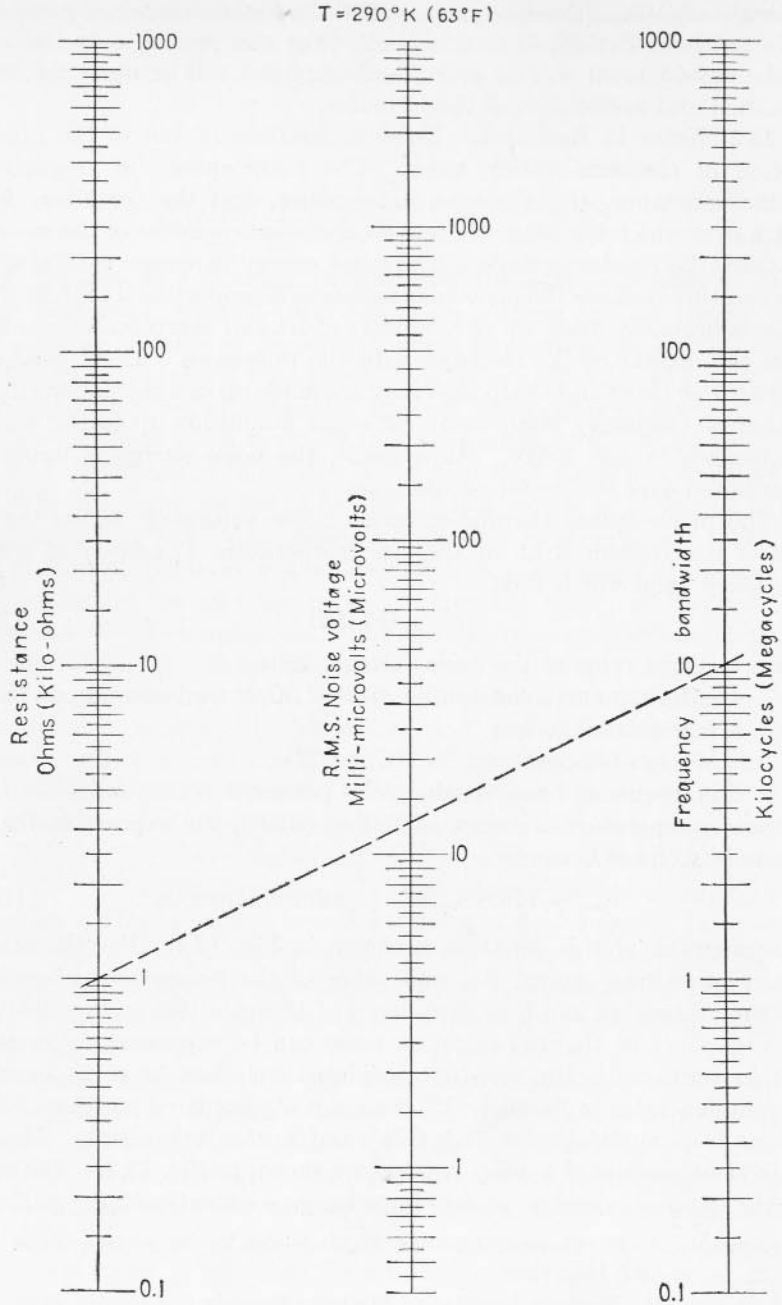


FIG. 12.1.—Nomographic chart of rms noise voltage across a resistor.

of the value given by Eq. (12.2). The square in the figure indicates an infinite-impedance constant-current generator whose output is

$$i_{\text{rms}} = 126.0\sqrt{\frac{B}{R}} \quad \text{micromicroamperes} \tag{12.3}$$

at room temperature. This is obtained by dividing the expression for the rms noise voltage by the resistance. The general form of the above equation is

$$i_{\text{rms}} = \sqrt{\frac{4kTB}{R}} \quad \text{amperes} \tag{12.4}$$

where all the symbols have their previous significance. A nomograph of the rms noise-current equivalent of a noisy resistor to fit the right circuit of

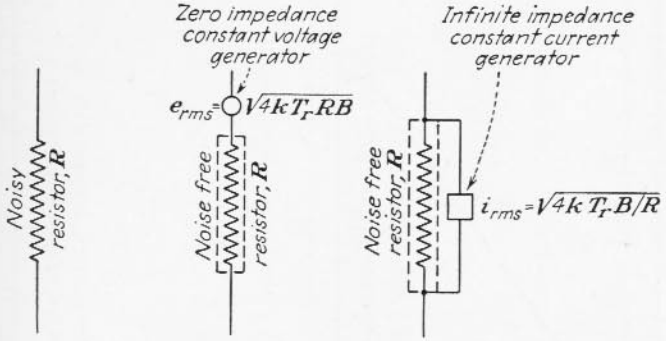


FIG. 12.2.—Equivalent circuits of a noisy resistor.

Fig. 12.2 is given in Fig. 12.3. For the sample construction line shown, the equivalent rms noise current of a 1,000-ohm resistor is 4 milli-microamperes for a frequency band width of 1 mc.

The noise power associated with thermal-agitation noise in a resistor is i^2R, or

$$N = 4kTB \quad \text{watts} \tag{12.5}$$

Note that this is independent of the value of resistance. The *available* noise power that can be obtained from a resistor by perfect matching is just one-fourth of this value. This follows immediately from maximizing the power obtainable from the middle circuit of Fig. 12.2 by varying the resistance load on such a generator. Maximum output power is obtained when the load resistance equals the generator resistance and has a value of $\frac{e^2}{4R}$. Thus the *available* noise power is

$$N_a = kTB \quad \text{watts} \tag{12.6}$$

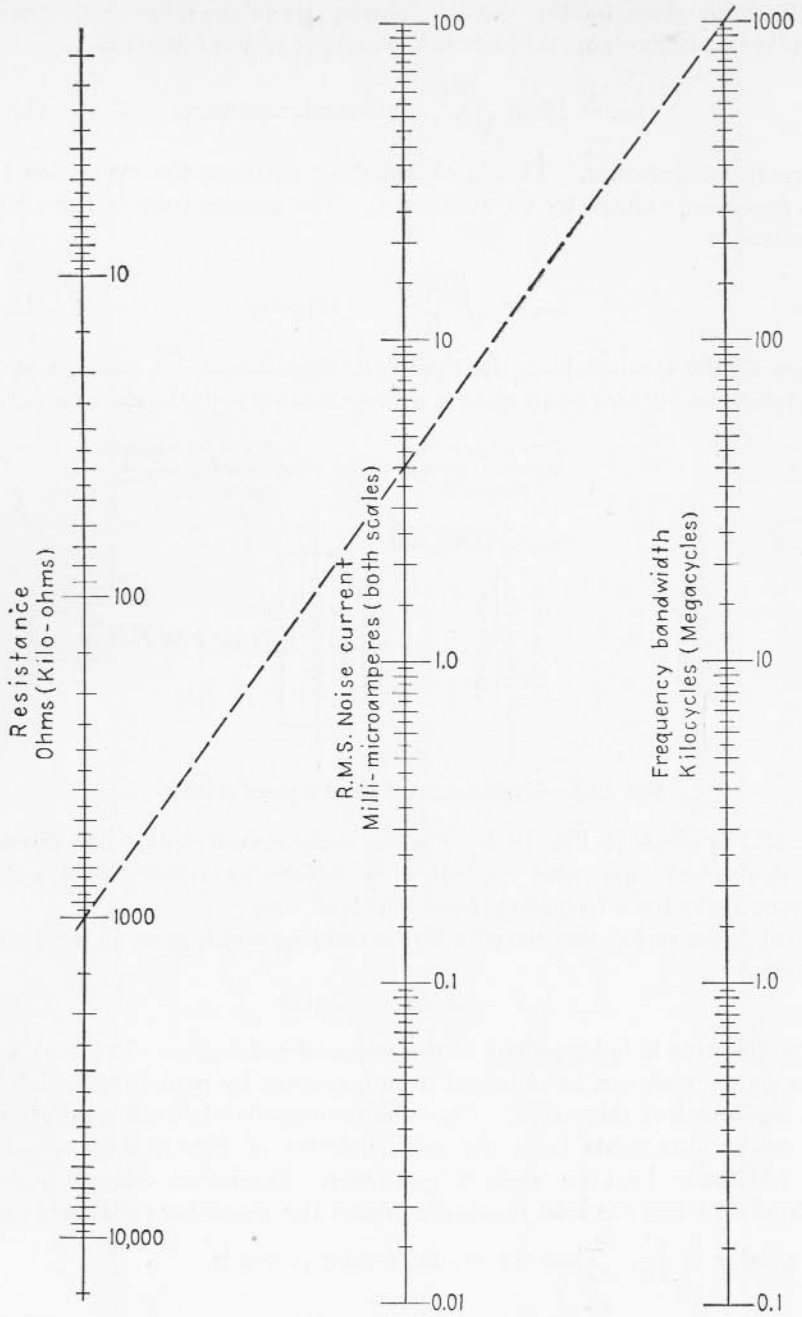


FIG. 12.3.—Nomographic chart of equivalent rms current through a noisy resistor.

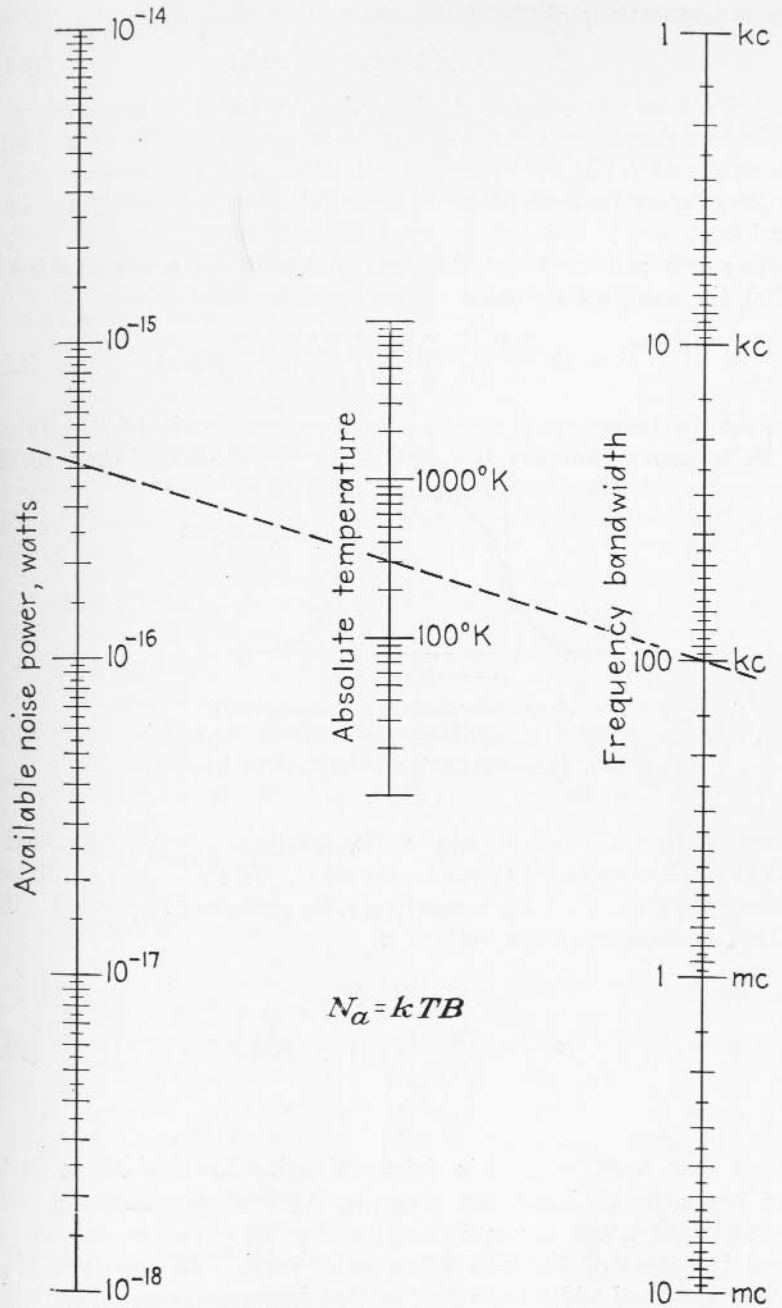


FIG. 12.4.—Nomographic chart of noise power available from a resistor.

At room temperature of 290°K this has a value of

$$N_a = 4.0 \times 10^{-21} B \qquad \text{watts} \tag{12.7}$$

A nomograph of this equation is shown in Fig. 12.4. A temperature of 300°K and a frequency band width of 10 kc is seen to give an available noise power of 4.1×10^{-16} watt. *It is convenient to remember that at room temperature the available noise power for a* 1*-mc band width is* 144 *db below* 1 *watt.*

When two resistances at different temperatures are connected in parallel, the mean-square noise voltage becomes

$$\bar{e}^2 = 4k \frac{R_1 R_2 (R_2 T_1 + R_1 T_2)}{(R_1 + R_2)^2} B \qquad \text{volts}^2 \tag{12.8}$$

where the resistances are in ohms and the temperatures in degrees Kelvin, and Boltzmann's constant is 1.3805×10^{-23} watt-sec. When several

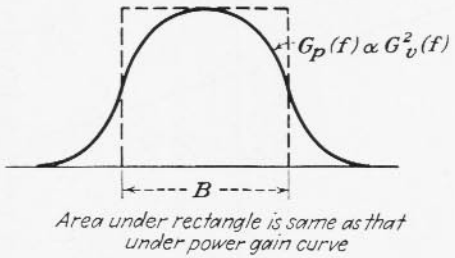


FIG. 12.5.—Definition of equivalent band width.

resistances are connected in parallel, the results are better expressed in terms of conductances. Let conductances $G_1, G_2, G_3, \ldots, G_n$ at temperatures $T_1, T_2, \ldots, T_n$, respectively, be connected in parallel. The resulting mean-square noise voltage is

$$\bar{e}^2 = 4k \frac{\sum_1^n G_n T_n}{\left(\sum_1^n G_n\right)^2} B \qquad \text{volts}^2 \tag{12.9}$$

Some care must be used in determining the band width to fit the above expressions. Since the concepts involved are basically those concerned with power, the equivalent band width of any device must be defined in terms of the power-frequency curve. In the case of an amplifier the band width is defined as that frequency interval for which a power gain equal to the gain at mid-band would transmit the same noise

energy as does the actual power-gain–frequency curve. Analytically, this becomes

$$B = \frac{\int_0^\infty G_p(f)\, df}{G_p(f_0)} \tag{12.10a}$$

where G_p is the power gain. All this amounts to is finding the width of a rectangular power-gain–frequency curve of height equal to the mid-frequency power gain of the actual curve as shown in Fig. 12.5. In terms of voltage amplification, the equivalent band width is

$$B = \frac{\int_0^\infty G_v^2(f)\, df}{G_v^2(f_0)} \tag{12.10b}$$

where G_v is the voltage gain.

12.3. Sources of Noise in Tubes. Noise can occur in vacuum tubes from a rather imposing list of sources. The principal sources of noise in tubes are

1. Shot effect (temperature-limited emission).
2. Reduced shot effect.
3. Flicker effect.
4. Collision ionization.
5. Random division of current between electrodes.
6. Induced noise at ultra-high frequencies.
7. Faulty tube construction.
   a. Hum.
   b. Poor insulation.
   c. Vibration.
   d. Varying wall charges.

Shot effect is the noise associated with random emission in a tube whose emission is temperature-limited. This is probably the loudest of the electronic tube noises but not the most serious, for tubes are seldom operated so that their emission is temperature-limited.

The so-called "reduced shot effect" is observed in tubes whose emission is space-charge-limited. The magnitude of the noise is much less than in tubes whose emission is temperature-limited. In this case the space charge exerts a smoothing action upon the true shot effect, and the noise is principally due to variations in the space-charge currents.

Flicker effect is observed with oxide cathodes. This effect is associated with variations in the activity of the emitting surface. The effect is much more noisy than the true shot effect for temperature-limited

emission. When the emission is space-charge-limited, the magnitude of the noise is greatly reduced.

Noise in tubes is raised by the presence of an appreciable amount of gas. This is due to the fact that gas molecules are ionized by collision with emitted electrons and the positive ions formed subsequently liberate little bursts of electrons as they penetrate the virtual cathode in front of the emitting surface. Gas noise is inappreciable unless the positive-ion gas current is more than a few hundredths of a microampere. Such gas noise appears mostly below 10 mc.

Random division of current between electrodes contributes to the noise of multielectrode tubes and makes pentodes three to five times as noisy as the same tube connected as a triode. It may be said that multielectrode tubes will always be noisier than triodes because of this additional factor contributing to the tube noise.

The ultra-high-frequency components of the random fluctuations of space charge in a tube will induce voltages in the grid circuit, which in turn will react back upon the space-charge flow. This effect is important only for frequency components above 30 mc.

Noise due to faulty tube construction is always present to a degree. If the filament is not sufficiently noninductive, hum will result. If insulation is poor at any point in the tube, there will be leakage currents, which will generally create noise because of nonconstant leakage resistance. Vibration may be a factor in an electromechanical feedback circuit. Dirt on the glass inside of a tube may give rise to varying wall charges, which will influence the tube current in a noisy manner. All these items can, however, with sufficient care in construction be held to a very low level.

Items 1, 2, 4, 5, and 6 listed above can never be removed entirely. They are, however, subject to an analysis that shows how their effects may be minimized. These items will be the subject of the subsequent sections. It has been found in most cases that it is convenient to express tube-noise effects in terms of equivalent noisy resistors. These resistors in turn are considered to have internal-noise emfs.

12.4. Shot Noise in Diodes with Temperature-limited Emission. Noise in diodes was probably the first form of tube noise ever detected. It is generally referred to as shot noise but also as "*Schrot* noise" and "Schottky noise," after the scientist who first analyzed the effect.[1,2] The noise is due to the random emission and arrival of electrons

[1] SCHOTTKY, W., Spontaneous Current Fluctuations in Electron Streams, *Ann. Physik*, vol. 57, pp. 541–567, Dec. 20, 1918.

[2] See also FRY, T. C., The Theory of the Schroteffekt, *Jour. Franklin Inst.*, vol. 199, February, 1925.

at the plate. It cannot be explained in terms of individual electron emission or arrival alone, for if the electrons were emitted at a uniform rate the lowest frequency component of noise would be above the highest frequency that vacuum tubes can handle. Thus a current of 1 ma corresponds to a flow of approximately 10^{16} electrons per sec. If these did flow at a uniform rate, there would be no noise components below 10^{16} cycles per sec. The electron stream evidently exhibits rather pronounced variations in density caused by the electrons arriving in groups. The mean square of the fluctuation components of current is

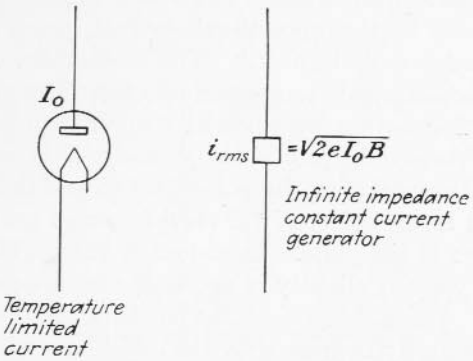


FIG. 12.6.—Constant-current-generator equivalent of a diode with temperature-limited current.

found to depend only upon the magnitude of the emitted current and the frequency band width

where e is electronic charge, 1.6020×10^{-19} coulomb

$$\bar{i}^2 = 2eI_0B \quad \text{amperes}^2 \tag{12.11}$$

I_0 is emission current, amperes

B is band width, cycles per sec

This expression may more conveniently be written

$$\bar{i}^2 = 3.2041 \times 10^{-19} I_0 B \quad \text{amperes}^2 \tag{12.12}$$

If the current from a diode with temperature-limited emission is put through a resistor R, the diode effectively puts a noise power of value $\bar{i}^2R$ into the resistor.

The above relations have been verified experimentally and they are reproducible to a high degree of accuracy. This property makes the diode with temperature-limited emission valuable as a standard noise source for such purposes as receiver and amplifier sensitivity measurement.

The diode with temperature-limited emission acts like a constant-

current generator as far as noise energy is concerned and will put a noise current given by the above equations through a resistor of any size. The equivalent circuit of the diode with temperature-limited emission is shown in Fig. 12.6.

12.5. Reduced Shot Effect in Diodes with Space-charge-limited Emission. In diodes in which the emission is space-charge-limited, the shot noise is much less than in the same diode passing the same current when its emission is temperature-limited.[1–4] The noise power is of the order of 10 per cent of that encountered for the same current when the emission is temperature-limited. The space charge thus has a very definite "smoothing" action upon the shot effect, giving rise to what may be called the "reduced shot effect." The mechanism of the smoothing action of the space charge is something like this: The virtual cathode in front of the emitting surface has a potential lower than that of the emitter by a value determined by the mean velocity of emission. Electrons with all velocities are storming this potential hill, and those with velocities greater than the mean velocity will *on the average* get past the virtual cathode and go on to the plate. Occasionally there will come a group of electrons with a velocity slightly in excess of that needed to get past the virtual cathode. When this occurs, the potential minimum at the virtual cathode is momentarily depressed by the additional space charge and as a result a few electrons that normally would have got past the potential minimum fail to do so and are returned to the emitter. This means that for every burst of electrons which might give rise to noise there is a compensating current set up in the opposite direction which tends to cancel the noise produced by the burst. The net result is an over-all reduction in noise that is considerable. The resulting noise levels are, however, high enough still to be of concern in the design of electronic equipment.

By considering the action of each increment of the velocities encountered in the process of emission some fairly good theoretical expressions for the reduced shot noise may be obtained. If the ratio of the noise

[1] RACK, A. J., Effect of Space Charge and Transit Time on the Shot Noise in Diodes, *Bell Sys. Tech. Jour.*, vol. 17, pp. 1–28, October, 1938.

[2] NORTH, D. O., Fluctuations in Space-charge-limited Currents at Moderately High Frequencies, *RCA Rev.*, vol. 4, Part II, pp. 441–473, April, 1940; vol. 5, pp. 244–260, July, 1940.

[3] WILLIAMS, F. C., Fluctuations of Space Charge Limited Currents in Diodes, *Jour. I.E.E.*, vol. 89, Part III, pp. 219–229, December, 1941.

[4] BELL, D. A., Fluctuations in Space Charge Limited Currents, *Jour. I.E.E.*, vol. 89, Part III, pp. 207–212, December, 1942.

See also SCHOTTKY, W., *Wiss. Veröffent. Siemens-Werken*, vol. 14, p. 15, 1937.

power of a diode passing a given current with and without space-charge limitation of current be designated by Γ^2, then, for a plane-electrode diode with a large ratio of plate voltage V_p to average emission voltage (velocity equivalent) V_e,

$$\Gamma^2 = \frac{9V_e}{V_p} \tag{12.13}$$

seems to be functionally correct.[1,2] The mean-square noise current can be expressed by an equation similar to that for the diode with temperature-limited emission as

$$\bar{i}^2 = \Gamma^2 2eI_0B \qquad \text{amperes}^2 \tag{12.14}$$

by analogy to Eq. (12.11). The mean-square noise current can also be written in the form

$$\bar{i}^2 = \theta \frac{4kT_cB}{R_1} \qquad \text{amperes}^2 \tag{12.15}$$

by analogy to Eq. (12.4). In this form, θ is a dimensionless parameter that has an asymptotic value of 0.644 for large ratios of plate to minimum potential, T_c is the absolute cathode temperature, and R_1 is the a-c diode resistance. Theoretically, the parameter θ is within a fraction of a per cent of the asymptotic value as long as the plate current is less than 80 per cent of the emission current and the plate voltage is greater than 2 volts for normal oxide operating temperatures.[3–5] Experimentally determined values of diode noise are 50 per cent higher than predicted by the theoretical expressions of Eqs. (12.13) and (12.15), so that θ assumes a value of unity for diodes. *The significance of Eq.* (12.15) *is that the noise power from a diode whose emission is space-charge-limited is the same as that from a resistor at the cathode temperature equal to the a-c diode resistance.*[6] The equivalent circuit for this case is given in Fig. 12.7.

[1] WILLIAMS, *op. cit.*

[2] BELL, *op. cit.*

[3] PEARSON, G. L., Shot Effect and Thermal Agitation in an Electron Current Limited by Space Charge, *Physics*, vol. 6, pp. 6–9, January, 1935.

[4] RACK, *op. cit.*

[5] NORTH, *op. cit.*

[6] Combination of Eqs. (12.14) and (12.15) and the use of the fact that $R_1 = \frac{2}{3}\frac{V_p}{I_0}$ for the diode with space charge suggests that Γ^2 has the value $(0.644)\frac{3kT_c}{eV_p}$, which is consistant with the observed relation of Eq. (12.13).

12.6. Reduced Shot Effect in Triodes with Space-charge-limited Current. Triodes, too, are noisy. In fact, the noise in triodes, as pointed out before, may be a limiting factor in the sensitivity of an electronic device. The noise is due to the effects just observed in diodes and shows

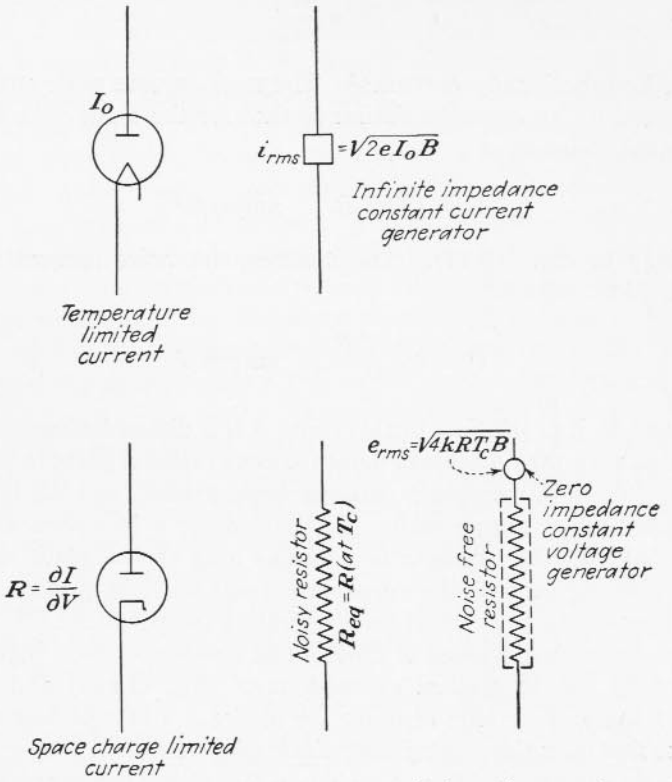


FIG. 12.7.—Resistor equivalent to a diode with space-charge-limited current.

as a fluctuation in plate current. It is convenient to interpret the noise in a triode as being due to a noisy resistor in series with the grid of the tube considered free from noise. The value of this noisy resistor in series with the grid is

$$R_{eq} = \frac{\theta T_c G}{G_m T_r G_m} \quad \text{ohms} \tag{12.16}$$

where θ is the effective cathode-temperature ratio of approximate value two-thirds, G_m is the mutual conductance of the triode, G is the a-c conductance of the diode equivalent of the triode, T_c is absolute cathode temperature, and T_r is absolute room temperature. The ratio of equivalent-diode conductance to triode mutual conductance is the inverse-square ratio of equivalent-diode spacing to the cathode-grid spacing as discussed in Chaps. 7 and 8. For the plane-electrode diode, from Eq. (8.45), this ratio is

$$\frac{G}{G_m} = \left[1 + \frac{1}{\mu}\left(\frac{d_{cp}}{d_{cg}}\right)^{4/3}\right]^{3/2} \tag{12.17}$$

The corresponding ratio for cylindrical triodes is difficult to express

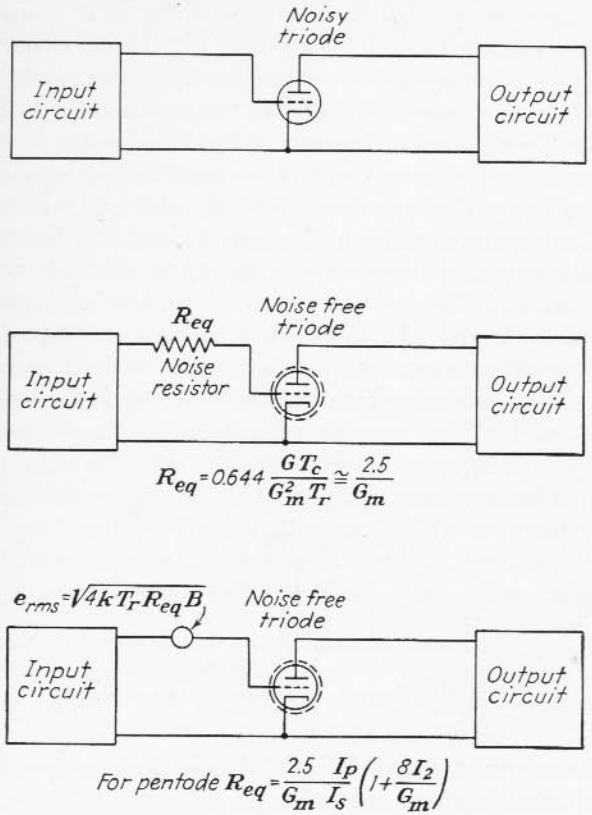


FIG. 12.8.—Equivalent circuit of a noisy triode.

exactly but is given approximately by

$$\frac{G}{G_m} = 1 + \frac{1}{\mu}\left[1 + \frac{2}{3}\ln\frac{r_p}{r_g}\right] \tag{12.18}$$

The value of this ratio will generally lie between 1 and 2. Assuming that a typical value of the conductance ratio is 1.25 and that the cathode

temperature is 3.33 times the room temperature, Eq. (12.16) reduces to

$$R_{eq} = \frac{2.5}{G_m} \quad \text{ohms} \tag{12.19}$$

which is sufficiently accurate for a good many purposes.[1] The value of equivalent resistance thus given is that resistance which if inserted in series with the grid of the triode considered noise-free would cause as much noise current in the plate circuit of the tube as does the reduced shot effect. The equivalent circuits of triode in terms of a noise-generating resistor and a noise-free tube is given in Fig. 12.8. Observed values of noise in a triode agree very closely with the values predicted by the above equations, much more so than was the case with diodes.

12.7. Noise Due to Gas in Tubes. When there is gas in tubes, there is an extra component of noise due to the electrons and ions liberated by collision ionization. The electrons passed by the grid will collide with some of the gas molecules, forming positive ions and liberating more electrons. The liberated electrons will pass on to the plate and give rise to some extra noise. The positive ions will be attracted to the negative grid and flowing through the external impedance will cause a voltage in the grid that will also give rise to noise. The noise is proportional to the number of ions formed, which in turn is proportional to the normal space current and to the number of gas molecules, or gas pressure. Fortunately, the positive-ion grid current is a measure of the number of ions formed per second and can be used to determine the noise without knowing the gas pressure.[2,3] As with other components of noise, the noise can be described in terms of a resistor in series with the control grid of the tube considered noise-free. The equivalent noise-generating grid resistor is

$$R_{eq} = \left(20R_g^2 + 4 \times 10^4 \frac{I_p}{G_m^3}\right) I_1 \quad \text{ohms} \tag{12.20}$$

where R_g is shunt resistance of the grid circuit, ohms
G_m is mutual conductance, mhos
I_p is plate current, amperes
I_1 is control-grid current, amperes

[1] HARRIS, W. A., Space Charge Limited Current Fluctuations in Vacuum Tube Amplifiers and Input Systems, *RCA Rev.*, vol. 5, pp. 505–524, April, 1941; vol. 6, pp. 114–124, July, 1941.

[2] *Ibid.*

[3] THOMPSON, B. J., and D. O. NORTH, Fluctuations in Space-charge-limited Currents Caused by Collision Ionization, *RCA Rev.*, vol. 5, pp. 371–388, January, 1941.

The first term in this expression is due to the flow of positive-ion current through the external grid impedance. The second term is due to the electrons liberated upon ionization that are attracted to the plate.

As an example, consider the case of a gassy tube for which the positive-ion control-grid current is 0.01 microampere. Let the mutual conductance of the tube be $5{,}000 \times 10^{-6}$ mho, the plate current 1 ma, and the shunt resistance of the grid circuit 100,000 ohms. Then the first term of Eq. (12.20) contributes a noise resistance of 2,000 ohms, and the second term contributes a noise resistance of 3.20 ohms. The second term of Eq. (12.20) is usually much smaller than the first term, as in this example, and can ordinarily be neglected.

12.8. Reduced Shot Effect in Multielectrode Tubes with Space-charge-limited Currents. Pentodes are even noisier than triodes—by a considerable factor. In fact, by comparison, triodes are relatively quiet. The additional noise in pentodes is due to the random division of the fluctuation noise between the electrodes. The individual groups of electrons that burst through the virtual cathode are quite local in their impingement upon electrodes in their subsequent travel, but the compensating currents due to the displacement of the virtual cathode are more or less uniformly distributed between the electrodes in the ratio of the direct currents. As a result, the smoothing, or compensating, action in a pentode is much less pronounced than in a triode.

As with the triode, the noise of a pentode can be expressed as being due to a noisy resistor in series with the control grid of the tube considered noise-free. The equivalent resistor in this case is[1,2]

$$R_{eq} = \frac{2.5}{G_m} \frac{I_p}{I_s} \left(1 + 8 \frac{I_s}{G_m}\right) \tag{12.21}$$

where all the symbols have their customary significance. Currents must be expressed in amperes and conductances in mhos to yield equivalent resistance in ohms. The first factor of Eq. (12.21) is seen to be the equivalent resistance for a triode. The remainder of the expression generally increases the value of the triode equivalent resistance by a factor of three to five. For screen currents much smaller than the plate current the noise in both the plate and screen circuits is approximately equal to the true shot effect for a current equal to the screen current. The value of the mean-square noise current in the plate circuit is readily obtained from this by means of Eq. (12.12) and converted to an equivalent noisy resistor *in the plate circuit* by means of the nomograph of Fig. 12.3.

[1] NORTH, D. O., Fluctuations in Space Charge Limited Currents in Multi-collectors, *RCA Rev.*, vol. 5, pp. 244–260, October, 1940.

[2] HARRIS, *op. cit.*

12.9. Noise in Mixer Tubes. Mixers are also noisy. The factors that contribute to the noise of triodes and pentodes also contribute to mixer noise. In a mixer a large voltage from the local oscillator is applied to the tube so that the current and the mutual conductance swing over a large value. Since the noise of tubes is known as a function of current and mutual conductance, it is not too difficult to evaluate the mixer noise.

In a mixer it is the noise in the intermediate-frequency band that is of importance. The noise in this band will vary periodically over the local oscillator cycle as the mutual conductance and current vary with the voltage applied at the local oscillator frequency. The total intermediate-frequency noise can be obtained by summing the plate noise over the local oscillator cycle.

$$\bar{\imath}_{i-f}^2 = \frac{1}{2\pi}\int_0^{2\pi} \bar{\imath}_{pn}^2\, d(\omega t) \tag{12.22}$$

where $\bar{\imath}_{pn}^2$ is the mean-square noise current in the plate at any instant t and ω is 2π times the local oscillator frequency. It is convenient (except in the case of the diode mixer) to express the mixer noise in terms of an equivalent noise resistance in series with the control grid of the tube considered as noise-free. Thus

$$\bar{e}_n^2 = \frac{\bar{\imath}_{i-f}^2}{G_c^2} \tag{12.23}$$

where G_c is the conversion transconductance of the tube. Corresponding to this value of input noise, the equivalent input noisy resistor is

$$R_{eq} = \frac{\bar{e}_n^2}{4kTB} \tag{12.24}$$

$$R_{eq} = \frac{\bar{\imath}_{i-f}^2}{4kTG_c^2B} \tag{12.25}$$

Upon applying the above ideas there are obtained the results shown in Fig. 12.9 for a fictitious pentode tube connected in accordance with the four most common mixer connections.[1] In this figure, G_0 is the maximum value which the mutual conductance assumes over the local oscillator cycle, generally that corresponding to zero grid voltage. The quantity G_x is the maximum value of the screen-plate transconductance. The quantity I_x is the maximum value of plate current when the signal is

[1] HEROLD, E. W., Superheterodyne Converter Considerations in Television Receivers, *RCA Rev.*, vol. 4, pp. 324–337, January, 1940.

See also the summarizing articles by HEROLD, E. W., and L. MALTER, Some Aspects of Radio Reception at Ultra-high Frequency, *Proc. I.R.E.*, vol. 31, pp. 423–438, 491–510, 567–582.

injected into the screen circuit and the local oscillator into the control-grid circuit.

It is seen that the triode is the best converter and the pentode with signal and local oscillator applied to the control grid is the next best. The others are too noisy for high-sensitivity applications.

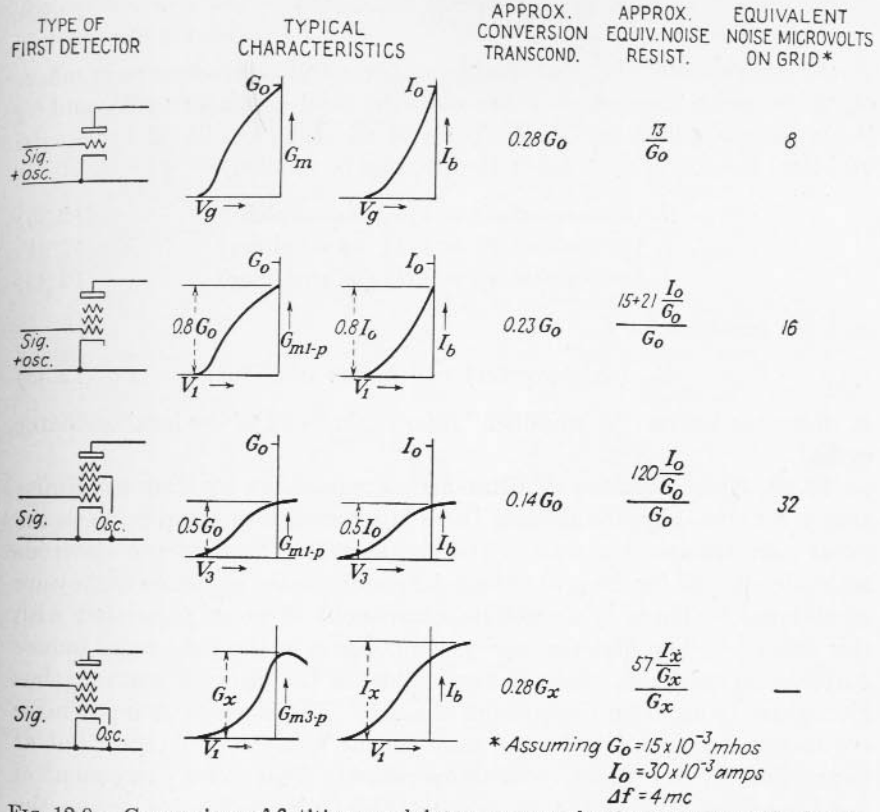


FIG. 12.9.—Comparison of fictitious modulators presumed to have similar cathode and first-grid structures. (*Herold.*)

When the values required in the tabulation of Fig. 12.9 are not available, the following formulas will be found sufficiently accurate for most purposes:

For triode mixers,

$$R_{eq} = \frac{4}{G_0} \tag{12.26}$$

or

$$R_{eq} = \frac{2.5\bar{G}_m}{G_0^2} \tag{12.27}$$

For pentode mixers

$$R_{eq} = \frac{I_p}{I_s}\left(\frac{4}{G_0} + \frac{20I_2}{G_0^2}\right) \tag{12.28}$$

or

$$R_{eq} = \frac{I_p}{I_s}\left(\frac{2.5\bar{G}_m}{G_0^2} + \frac{20I_2}{G_0^2}\right) \tag{12.29}$$

where currents must be expressed in amperes and conductances in mhos. $\bar{G}_m$ is the mean transconductance over the local oscillator cycle, and G_0 is the transconductance at the peak of the local oscillator cycle. In addition, the following rules of thumb may be applied,

$$G_c \text{ (as converter)} = \tfrac{1}{4}G_m \text{ (as amplifier)} \tag{12.30}$$
$$I_p \text{ (as converter)} = \tfrac{1}{4}I_p \text{ (as amplifier)} \tag{12.31}$$
$$I_2 \text{ (as converter)} = \tfrac{1}{4}I_2 \text{ (as amplifier)} \tag{12.32}$$

and, for pentodes only,

$$R_{eq} \text{ (as converter)} = 4R_{eq} \text{ (as amplifier)} \tag{12.33}$$

in which the values "as amplifier" refer to the peak of the local oscillator cycle.[1]

12.10. Noise Induced at Ultra-high Frequencies by Random Emission. At ultra-high frequencies there is a conductive component to the input admittance of a tube. The finite transit time of the electrons makes it possible for the grid to transfer energy to the electrons as they are accelerated.[2] There is a separate component of noise associated with this effect.[3,4] The high-frequency components in the tube noise induce currents on the grid, which in turn influence the electron current, thus giving rise to an extra component of noise. The equivalent input noise conductance is found to be the same as the input conductance, but at approximately five times room temperature. This extra component of noise may be represented as a constant-current generator across the electronic component of grid conductance across the input circuit. To a first order of approximation the induced noise is independent of the normal noise component, which may be added in series with the grid as

[1] HARRIS, *op. cit.*

See also articles by HEROLD, *Proc. I.R.E.*, for some typical values.

[2] FERRIS, W. R., Input Resistance of Tubes as Ultra-high Frequency Amplifiers, *Proc. I.R.E.*, vol. 24, pp. 82–107, January, 1936.

[3] NORTH, D. O., and W. R. FERRIS, Fluctuations Induced in Vacuum Tube Grids at High Frequencies, *Proc. I.R.E.*, vol. 29, pp. 49–50, February, 1941.

[4] BALLANTINE, STUART, Schrot Effect in High Frequency Circuits, *Jour. Franklin Inst.*, vol. 206, pp. 159–168 August, 1928.

has been done before. The equivalent circuit for this effect is shown in Fig. 12.10. At ultra-high frequencies there is a component of grid-input conductance due to feedback through the cathode-lead inductance, as well as the component due to electron-transit-time effects. Both components vary as the square of frequency and thus are hard to separate. The feedback component has noise associated with it too, but as a resistor at room temperature.

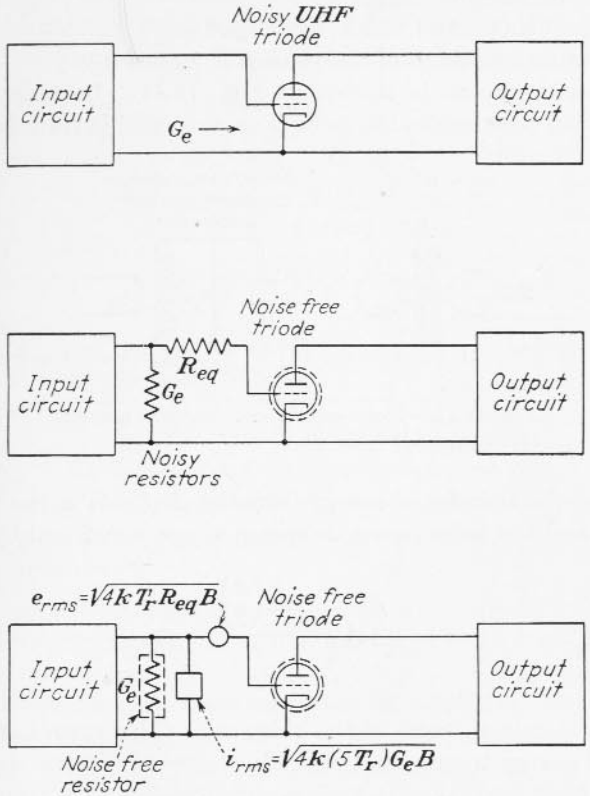


FIG. 12.10.—Equivalent circuit for induced noise in a tube at ultra-high frequencies.

In general, the noise of amplifiers does not change much with feedback. This is because the amplification and input impedance are ordinarily changed by the same factor.

12.11. Noise in Velocity-modulation Tubes. In a velocity-modulation tube a beam of electrons passes through the two grids of a resonator, between which there appears a high shunt resistance. Noise is generated, but there is evidently very little space-charge smoothing action in

this case. This is expected from the fact that there is no virtual cathode between the grids of the resonator. As a result, the noise is nearly the full shot noise of the beam, and therefore the noise power delivered to the resonator is given closely by

$$N = 2eI_0BR_{sh} \quad \text{watts} \tag{12.34}$$

where e is charge of the electron, 1.6020×10^{-19} coulomb
I_0 is beam current, amperes
B is frequency band width, cycles per sec
R_{sh} is shunt resistance of the resonator, ohms
The equivalent circuit is shown in Fig. 12.11. In many cases the electron transit time across the grids is an appreciable fraction of a cycle,

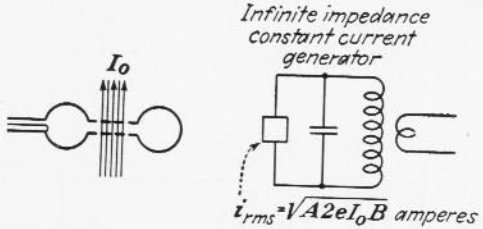


FIG. 12.11.—Noise-equivalent circuit of a velocity-modulated tube.

in which case the transfer of energy from the electrons to the resonator is not perfect and the noise power delivered above is reduced by the factor

$$A = \frac{\sin\left(\frac{\theta}{2}\right)}{\frac{\theta}{2}} \tag{12.35}$$

where θ is the transit angle of the electrons. This expression gives the efficiency of energy transfer between the electrons and the resonator and is developed in Chap. 17. It applies only to tubes with fine grids.

12.12. Noise in Phototubes. High-vacuum phototubes produce the same noise as do hot-cathode tubes with temperature-limited emission. This is true shot effect, giving rise to a mean-square fluctuation current of the value

$$\bar{i}_n^2 = 2eI_0B = 3.2040 \times 10^{-19}I_0B \quad \text{amperes}^2 \tag{12.36}$$

This is the same value of current as is produced by a resistance at room temperature of value

$$R_{eq} = \tfrac{1}{20}I_0 \quad \text{ohms} \tag{12.37}$$

The noise may be represented by a constant-current generator in parallel with the tube considered noise-free. The noise from the tube will often be of the same order as that produced by the large-value resistor used to develop voltage. Thus, if a 10-megohm resistor is used to develop voltage, the noise from the tube will be the same as the noise from the resistor when the current is 0.005 microampere. At higher currents the phototube noise predominates. The equivalent circuit of a typical phototube input circuit is shown in Fig. 12.12.

Gas is sometimes used in phototubes to increase the plate current by cumulative ionization. When this is done, the noise is increased by

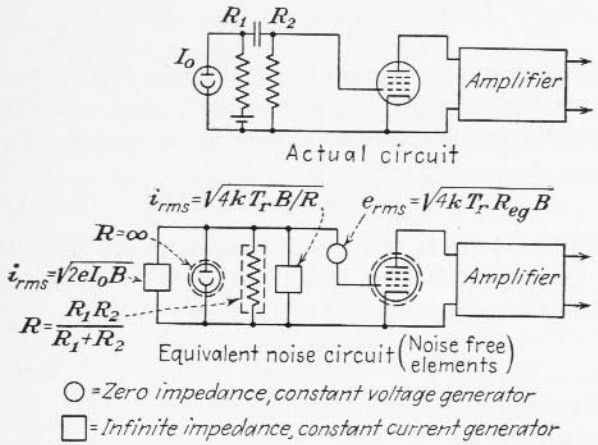


FIG. 12.12.—Noise-equivalent circuit of a phototube input circuit.

about the same amount as the signal but the signal-to-noise ratio is in general improved because the contributions of noise from other sources becomes relatively less.

12.13. Noise in Secondary-emission Multipliers. Another type of electronic amplifier in which noise may be a consideration is the secondary-emission multiplier. In such a tube there is a series of electrodes at successively higher potentials, each coated with a material that emits a large ratio of secondary to primary electrons. Electrons that strike the first anode give rise to S times as many electrons, which are attracted to the second anode, where they give rise to S times as many as the striking electrons, or S^2 times as many as struck the first anode. After n such impacts the current is S^n times as great as it was originally, S being the ratio of secondary to primary electrons—a number that can be made as high as 9 or 10. It might be thought that with such a system tremendous amplifications could be obtained. True, they can, but no improvement in signal-to-noise ratio can be achieved.

Experimental results indicate that the following laws govern the noise associated with secondary emission in such a device:

1. Secondary emission from any anode follows the shot-effect law, $\bar{i}_n^2 = 2eBI_0$, where I_0 is the emitted secondary current.
2. Shot noise from any anode is multiplied by subsequent stages in the same way as the signal is.[1]

Consider the action in a few successive stages of secondary-emission amplification. Let the current from a first anode be I_0; then the mean-square noise current associated with this is

$$\bar{i}_n^2 = 2eBI_0 \qquad (n = 0) \qquad (12.38)$$

When the direct current I_0 strikes the next diode, it gives rise to a direct current SI_0 and the corresponding mean-square noise current is multiplied by S^2, so that the mean-square noise current associated with the current SI_0 is

$$\bar{i}_n^2 = S^2 2eBI_0 + 2eBSI_0 \qquad (n = 1) \qquad (12.39)$$

in which the first term is the amplifier noise power from the previous anode and the second is that associated with the liberated secondary current. The above expression is more simply written

$$\bar{i}_n^2 = 2eBI_0(S^2 + S) \qquad (n = 2) \qquad (12.40)$$

At the next anode the liberated secondary current is S^2I_0, and the mean-square noise current is

$$\bar{i}_n^2 = S^2 2eBI_0(S^2 + S) + 2eBS^2I_0 \quad (n = 2) \qquad (12.41)$$

which is equal to

$$\bar{i}_n^2 = 2eBI_0(S^4 + S^3 + S^2) \qquad (n = 2) \qquad (12.42)$$

Extension of this process to n stages yields

$$\bar{i}_n^2 = 2eBI_0 \frac{S^n(S^{n+1} - 1)}{S - 1} \qquad (n = n) \qquad (12.43)$$

This means that the ratio of output to input noise power is

$$\frac{\text{Noise power out}}{\text{Noise power in}} = \frac{S^n(S^{n+1} - 1)}{S - 1} \qquad (12.44)$$

Correspondingly, the ratio of signal power out to signal power in is

$$\frac{\text{Signal power out}}{\text{Signal power in}} = S^{2n} \qquad (12.45)$$

[1] ZWORYKIN, V. K., G. A. MORTON, and L. MALTER, The Secondary Emission Multiplier, *Proc. I.R.E.*, vol. 24, pp. 351–375, March, 1936.

Upon taking the quotient of the last two equations, the relative change in the ratio of signal to noise power is

$$\frac{\text{Signal-power-to-noise ratio in}}{\text{Signal-power-to-noise ratio out}} = \frac{S^n(S^{n+1} - 1)}{S^{2n}(S - 1)} \qquad (12.46)$$

This ratio is slightly greater than 1 but approaches this value as the secondary-emission ratio S is increased. This simply means that signal and noise are amplified about the same amount in a secondary-emission multiplier, and as a result there is no gain on the signal-to-noise ratio. There is, however, an advantage to using secondary-emission multiplication in that resistor noises are virtually eliminated. Thus a phototube with secondary-emission multiplication has a lower signal-to-noise ratio than a phototube-resistor-amplifier combination at low levels of illumination. A further discussion of this specific case is given in Sec. 19.20.

12.14. Definition of Noise Figure. From all the preceding sections it is seen that there are inherent limitations to electronic devices determined by unavoidable noise. Most electronic devices will, in fact, be noisier than simple theory predicts because of an accumulation of various effects. The smallest amount of noise that an electronic device can possibly exhibit is the available noise power from the thermal agitation of a resistor in the frequency band considered, as given by Eq. (12.6). Usually the noise will be more than this. It is therefore convenient to use as a figure of merit for an electronic device the ratio of the actual noise power at its output terminals to that which it would have if the noise were limited to the minimum noise from thermal agitation. This figure of merit is called the *noise figure* of the device. Basically, the *noise figure is an excess noise ratio.*

A rigorous definition of noise figure involves a consideration of the gain of the device and the available input noise power and the output power. The gain of a device, invariably a four-terminal network, is defined as the ratio of the available signal power at the output to the available signal power at the output of the signal generator,

$$\text{Power gain} = \frac{S_{\text{out}}}{S_{\text{in}}} = G \qquad (12.47)$$

This definition of gain is independent of the output impedance of the device, but it does depend upon the impedance of the signal generator, which is taken as the nominal input impedance of the device. In many applications there is no difficulty in using a signal-generator impedance that is equal to the input impedance of the device under consideration, but in some applications, such as extremely broad-band devices, it is extremely difficult to maintain a constant input impedance. Hence

it is logical that the figure of merit should include nonconstancy of the input impedance. The available noise power between two terminals is defined as the noise power that would be absorbed by a matched output circuit. The available input noise power is simply that given from the Johnson noise formula of Eq. (12.6),

$$N_{in} = kTB \tag{12.48}$$

The noise figure is defined in terms of the factor of most importance in the ultimate sensitivity of an electronic device, the ratio of output signal to noise power. The maximum value this can have is the ratio of available input signal to noise power if there are no other noise sources in the device and if all impedances are properly matched. Some four-terminal networks that consist of passive elements only, say a transformer or a transmission line, have no noise sources present in them, but electronic devices always have some extra sources of noise. *The noise figure F of the device is defined as the ratio of the available signal-to-noise ratio at the signal-generator (input) terminals to the available signal-to-noise ratio at the output terminals*[1,2]

$$F = \frac{\dfrac{S_{in}}{kTB}}{\dfrac{S_{out}}{N_{out}}} \tag{12.49}$$

The noise figure of an electronic device is always greater than unity. The reference temperature is invariably taken as 290°K (63°F).

Some rearrangements of Eq. (12.49) are useful. Upon utilizing the power-gain definition of Eq. (12.47) the noise figure may be written

$$F = \frac{N_{out}}{GkTB} \tag{12.50}$$

From this the available noise output power is

$$N_{out} = FGkTB \quad \text{watts} \tag{12.51}$$

The available output noise due only to noise sources in the network is

$$N_{out} - GkTB = (F - 1)GkTB \quad \text{watts} \tag{12.52}$$

Example: It is desired to calculate the noise figure of a 250-ohm-input 30-mc intermediate-frequency amplifier having a band width of 2 mc and using 6AC7

[1] FRIIS, H. T., Noise Figures of Radio Receivers, *Proc. I.R.E.*, vol. 32, pp. 419–422, July, 1944.

[2] NORTH, D. O., Absolute Sensitivity of Radio Receivers, *RCA Rev.*, vol. 6, pp. 332–343, January, 1942.

pentodes. Coupling between stages is provided with two inductively coupled tuned circuits. Let the tube constants be

$$\begin{aligned} I_p &= 10 \times 10^{-3} && \text{ampere} \\ I_2 &= 2.5 \times 10^{-3} && \text{ampere} \\ G_m &= 9 \times 10^{-3} && \text{mho} \end{aligned}$$

Then, from Eq. (12.21), the equivalent noisy resistor in series with the control grid is

$$\begin{aligned} R_{eq} &= \frac{2.5}{9 \times 10^{-3}} \frac{10 \times 10^{-3}}{12.5 \times 10^{-3}} \left(1 + \frac{8 \times 2.5 \times 10^{-3}}{9 \times 10^{-3}}\right) \\ &= 715 \quad \text{ohms} \end{aligned}$$

From Fig. 12.1 this corresponds to an rms noise voltage of 4.85 microvolts. Since the noise is developed across the input impedance in series with the equivalent noise resistor, while the signal is developed only across the input impedance, the noise figure is

$$F = \frac{715 + 250}{250} = 3.86$$

In cases in which the stage amplification is low the effect of the noise from the following stage must be considered. In this case consider that the coupling network is a unity-ratio impedance transformer with an input impedance of 50,000 ohms. If the second stage is identical to the first in operating characteristics, then there is a noise voltage of 4.85 microvolts developed in series with its grid. This voltage corresponds to a current of 97.0 micromicroamperes through the plate load of the first stage. Upon referring this current back to the control grid of the first stage, it corresponds to a voltage of $\frac{97 \times 10^{-12}}{9 \times 10^{-3}}$, or 0.01075 microvolt. This is small compared with the 4.85 microvolts due to the first tube and so can be neglected in this case.

Noise Figure for Two Networks in Cascade. Usually the sources contributing to the excess noise in an electronic device are principally in the input circuit of the device. However, when the first stage of amplification has insufficient gain, the noise sources effectively located in the input circuit of the second stage of amplification contribute appreciably to the over-all noise as well. Noise from subsequent stages is generally so small compared with the amplified noise from the earlier extra sources that it may be neglected. When the condition cited above is the case, then the following relations hold: Let the first and second stages of amplification be designated by subscripts 1 and 2, respectively. Let over-all characteristics be designated by the subscript 12. The over-all power gain G_{12} is equal to the product of the gains of the first and second stages, G_1G_2. Let the band width be that of the over-all charac-

teristic as defined by Eq. (12.9). The available noise power at the output of the first stage of amplification is

$$N_1 = F_1 G_1 kTB \quad \text{watts} \tag{12.53}$$

from Eq. (12.51). The noise power in the output of the second stage due only to sources in that stage is, from Eq. (12.52),

$$N_2 - G_2 kTB = (F_2 - 1)G_2 kTB \tag{12.54}$$

The total available noise power at the output of the second stage is G_2 times the quantity in Eq. (12.53) plus the quantity in Eq. (12.54),

$$N_{12} = G_2 F_1 G_1 kTB + (F_2 - 1)G_2 kTB \tag{12.55}$$

or

$$N_{12} = \left[F_1 + \left(\frac{F_2 - 1}{G_1}\right)\right] G_1 G_2 kTB \tag{12.56}$$

Now, upon applying the general formula [Eq. (12.51)] to the over-all situation,

$$N_{12} = F_{12} G_{12} kTB \tag{12.57}$$

The values of available output noise over-all, from Eqs. (12.56) and (12.57), must be equal; therefore

$$F_{12} = F_1 + \frac{F_2 - 1}{G_1} \tag{12.58}$$

This is the important relation that has been sought. It gives the over-all noise figure of two amplifiers in terms of the separate noise figures and the gain of the first amplifier. Needless to say, the expression is not limited to amplifiers but may be applied to mixers and networks in general.

Sometimes noise figures of circuit elements are expressed in terms of equivalent temperatures, simply the room temperature multiplied by the noise figure. The symbol t is often used to designate the ratio of the actual noise to the available thermal-agitation noise of Eq. (12.6). This is most frequently done with passive elements that produce noise, such as crystal detectors. The noise figure of a crystal input receiver is

$$F = L_x(t_x - 1 + F_a) \tag{12.59}$$

where L_x is the conversion loss of the crystal detector corresponding to the reciprocal of the conversion gain of a tube mixer, t_x is the equivalent temperature ratio of the crystal, and F_a is the noise figure of the amplifier into which the output of the crystal detector is fed. If the crystal were noiseless, t_x would equal unity.

The design of electronic equipment with regard to obtaining a low noise figure is a rather complex problem, which will only be touched upon here.[1] The problem is a combination one, involving consideration of circuits and noise sources. It is possible in some cases that the minimum noise figure will be obtained with a condition of mismatched impedances owing to the fact that a mismatch will reduce the noise more than it will the signal. With receivers it is found practical to use r-f amplification before the mixer only up to a certain frequency. This frequency is of the order of 600 mc at this time. Beyond this frequency the noise figure of amplifiers is so great that the signal-to-noise ratio is increased rather than reduced upon amplification. There is also an upper frequency limit at which vacuum-tube mixers are practical. At present, this limit occurs at about 1,000 mc. It is quite possible that these frequency limits will be extended with time. The limits occur because the effective mutual conductance of all tubes decreases with increasing frequency owing to transit-time effects, thus raising the noise. The ultra-high-frequency induced noise effect also contributes to increasing noise.

12.15. Measurement of Noise and Noise Figure. It is not difficult to measure the noise and noise figure of an electronic device when its gain is sufficiently high so that an appreciable output noise power is obtained. Square-law devices such as thermocouples are preferred in noise measurements because the output indication is directly proportional to power. Crystal detectors may also be used at low levels of power. If the rectified crystal current is kept below a few microamperes, the crystal is almost certain to be a square-law device and as such is extremely sensitive. In this case the rectified crystal current is proportional to the input power. If a satisfactory power-output indicating device is used, the noise figure of a device can be measured by simply introducing signal input power until the output indication from the noise alone is doubled. The input power is then equal to that generated by the internal noise sources, and the noise figure is given by

$$F = \frac{S_{in}}{kTB} \tag{12.60}$$

where S_{in} is the signal power in. For any other adjustment of signal input power the noise figure is given directly from the defining relation of Eq. (12.49), it being remembered that the output-power indication is the sum of the noise and signal power out.

Diodes operating with temperature-limited emission may be used as a standard source of noise. Such diodes preferably have either a tungsten

[1] See the survey articles by HEROLD and MALTER, *op. cit.*

or a thoriated tungsten filament. It is difficult to keep the emitted current from an oxide-emitting surface constant under temperature-limited conditions of emission. The noise, of course, is due to shot effect, with the mean-square noise current given by Eq. (12.12). Diodes may successfully be operated as a standard noise source up to 100 mc. Beyond that frequency the impedance transformation introduced by the leads cannot be determined very accurately. Undoubtedly, special diodes can be built for noise measurements at higher frequencies. A standard noise source for measurements of the noise figure of an intermediate-frequency amplifier would consist of the diode in a shielded can with leads brought in through properly by-passed chokes. Across the diode there can be placed a coil that tunes the capacity of the diode to the center of the band of interest. The tuned circuit thus formed should be shunted by the nominal input resistance of the amplifier to be tested. Output leads are then brought from across the tuned circuit to the amplifier under test. Let the nominal input resistance of the amplifier be R; then the noise power delivered to the resistor of value R shunted across the diode is $2eI_0RB$, and the available power into the receiver is one-fourth of this. Let the diode current I_0 be adjusted until the normal noise output power of the receiver is doubled when the standard diode noise is connected to its input. Under these conditions the noise figure, from Eq. (12.60), is given by $\frac{eI_0RB}{2kTB}$; or since $\frac{e}{2kT}$ has the numerical value of 20, the noise figure is given simply by

$$F = 20I_0R \tag{12.61}$$

where I_0 is the diode current in amperes and R is the nominal resistance of the receiver in ohms.

12.16. Typical Tube-noise Values. In the table on the next page are given some typical operating conditions and associated noise values of representative triodes, pentodes, and mixers.

TABLE VI
TUBE-NOISE VALUES*

Type	Application	Voltages			Currents			Trans-conductance, micromhos	Noise-equivalent resistance		Noise-equivalent input, voltage** microvolts
		Plate volts	Screen volts	Bias volts	Plate ma	Screen ma	Cathode ma		Calculated ohms	Measured ohms	
6SK7	Pentode amplifier	250	100	−3	9.2	2.4	11.6	2,000	10,500	9,400–11,500	0.94
6SJ7	Pentode amplifier	250	100	−3	3	0.8	3.8	1,650	5,800	5,800	0.70
6SG7	Pentode amplifier	250	125	−1	11.8	4.4	16.2	4,700	3,300	.....	0.53
6AC7/1852	Pentode amplifier	300	150	−2	10	2.5	12.5	9,000	720	600–760	0.25
956	Pentode amplifier	250	100	−3	5.5	1.8	7.3	1,800	9,400	.....	0.90
1T4	Pentode amplifier	90	45	0	2.0	0.65	2.65	750	20,000	.....	1.3
6SA7	Frequency converter	250	100	0	3.4	8.0	11.9	450§	240,000	210,000	4.5
6K8	Frequency converter	250	100	−3	2.5	6.0	8.5‡	350§	290,000	.....	4.9
1R5	Frequency converter	90	45	0	0.8	1.8	2.75	250§	170,000	.....	3.8
6L7	Pentagrid mixer	250	100	−3	2.4	7.1	9.5	375§	255,000	210,000	4.6
6J5	Triode amplifier	250	...	−8	9.0	...	.....	2,600	960	1,250	0.28
955	Triode amplifier	180	...	−5	4.5	...	.....	2,000	1,250	.....	0.32
6AC7/1852	Triode amplifier	150	150	−2	...	...	12.5	11,200	220	200	0.14
6AC7/1852	Pentode mixer	300	150	−1†	5.2	1.3	6.5	3,400§	2,750	3,000	0.48
6SG7	Pentode mixer	250	125	−1†	3.0	1.1	4.1	1,180†	13,000	.....	1.0
956	Pentode mixer	250	100	−1†	2.3	0.8	3.1	650†	33,000	.....	1.7
6J5	Triode mixer	100	...	−1†	2.1	...	.....	620§	6,500	.....	0.74
6AC7/1852	Triode mixer	150	150	−1†	...	...	6.5	4,200§	950	.....	0.28
955	Triode mixer	150	...	−1‡	2.8	...	.....	660§	6,100	.....	0.72

* Reproduced from HARRIS, W. A., Space Charge Limited Current Fluctuations in Vacuum Tube Amplifiers and Input Systems, *RCA Rev.*, vol. 5, pp. 505–524, 6, pp. 114–124, April, July, 1941.

† At peak of oscillator cycle. ‡ Hexode section only. Triode section takes its current from a separate part of the cathode.

§ Conversion transconductance value. ** For effective bandwith of 5,000 cycles.