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FLUCTUATION VOLTAGES IN RECEIVER INPUT CIRCUITS

BY JOHN R. RAGAZZINI *

Fluctuation Voltages Due to Thermal Agitation ★ The basic source of fluctuation voltages in circuit elements is a pulsation of the electron gas in the conductors. Such pulses are set up by collisions between the molecules making up the mass of the conductor and the electrons of the free electron gas. The mechanism of this process will be described but first the properties of extremely short pulses will be examined.

A Fourier analysis of a pulse shown in Fig. 1 whose time duration is 2T yields

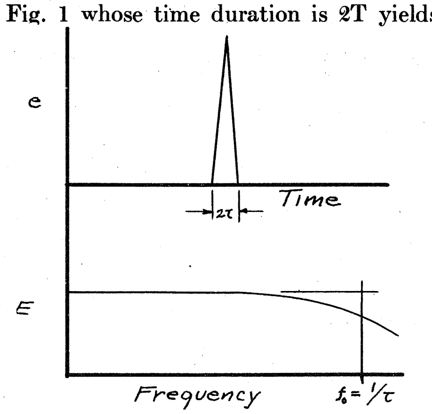


FIG. 1. CURVES SHOWING THE RELATION BETWEEN THE DURATION OF A PULSE, 2T, AND ITS FREQUENCY SPECTRUM

substantial frequency components to an upper frequency of 1/T. If the pulses are of infinitely short duration ($T \rightarrow O$), the harmonic content is uniform at all frequencies up to infinity. Physically, this implies that if an infinitely short pulse of current were applied to a sharply tuned circuit of constant impedance, the voltage developed across this impedance would be constant regardless of the frequency to which the circuit is tuned. The uniformly constant frequency spectrum is a characteristic of extremely short pulses.

Such pulses are spontaneously set up in a circuit due to the thermal agitation of the molecules making up the wire. This process has been studied by various investigators [1,2] and accurate formulas have been derived. However, if some thought is given to the basic phenomena involved, certain simplifications can be presented which make fluctuation noise computations simpler. Such an approach was proposed more recently by Moullin and Ellis [3] but has not been used widely.

Briefly, the analogy between fluctuation voltage effects and the kinetic theory of gases is used. For instance, Brownian movements are an outward manifestation of the gas laws. If a microscopic globule of fat is immersed in water and is observed through a microscope, it is found to have random pulsations or movements in all directions. Closer examination would reveal that the average energy of this globule would rise proportionately to absolute temperature. The source of these pulsations is vibration of the water molecules due to their temperature. As they vibrate, they collide with the globule of fat, which represents a cohesive group of molecules, and impart energy to it. Since the globule is struck at random and from all sides, its movement is a series of aimless pulses in all directions.

At this point, a basic rule called the principle of equipartition of energy is brought into play. Within a gas whose molecules are in a state of agitation, the molecules will, in time, have the same average energy. For instance, if gases at different temperatures were mixed, the mixture will finally reach a uniform temperature. Since temperature is proportional to the average energy of a molecule, it follows that the average energy of all molecules in a gas is the same. The law defining this phenomenon is the principle of equipartition and more careful analysis will show that the average energy per molecule in a gas is:

$$E = \tfrac{1}{2}\ \mathrm{KT} \text{ joules per degree of freedom} \qquad (1)$$

$K = 1.372 \times 10^{-23}$ joules/degree Kelvin
$T = 273 + {}^\circ C$ degrees Kelvin

Equation (1) gives the average energy stored by a molecule in each of its independent modes of energy storage. Thus, if the kinetic energy storage is taken in the three coördinate axes, the resultant value will be three times that given in (1).

Surprisingly enough, this principle holds in relation to cohesive masses of molecules such as the fat globules previously described. Since the suspended globules of fat which exhibit Brownian movements have three degrees of freedom, one for each of the three coördinate axes, the average energy of the globules is 3/2 KT joules. This relation can be proved experimentally.

The electron gas in conductors may now be regarded in the same light as the fat globules randomly pulsing in a fluid. That the electron gas is a cohesive entity may be understood by realizing that as soon as one electron is struck by a molecule and suddenly acquires a velocity, its motion sets up a magnetic field which links all the other electrons in the electron gas. This suddenly generated field causes all the electrons in the circuit to be set in motion also. Thus, in a sense, a tight cohesion exists in the electron gas because one electron cannot be moved without similarly disturbing all the other electrons by induction. This cohesion corresponds to the analogy of the fat globule, since collision by a molecule of the fluid with a molecule of the globule sets the *whole* globule in motion.

In addition, just as the fat globule acquires instantaneous changes of velocity due to collision with molecules, so does the electron gas set up infinitely short pulsations of current in a circuit. The electrons set in motion by a molecular collision set up infinitely short pulses of magnetic field with the resultant infinitely short pulses of voltage by induction. Thus, if the effects of electron gas pulses due to molecular collision were considered as an equivalent generator, this generator would produce voltage pulses whose frequency spectrum is uniform from zero to infinite frequency. This reasoning justifies the replacement of fluctuation voltage effects by an equivalent generator of uniform frequency spectrum.

If this analogy can be carried further, it follows that the electron gas in a circuit can absorb a share of energy equal to that of a single molecule providing a means can be found to store that energy. In other words, a component of energy, ½KT joules, will be stored in the electron gas per degree of energy storage. For instance, in the circuit shown in Fig. 2, a single degree of freedom in energy storage exists in

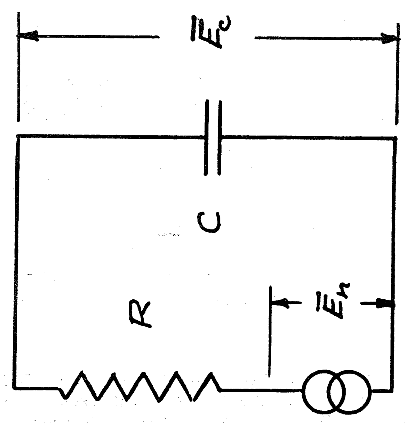


FIG. 2. CIRCUIT USED TO DETERMINE VOLTAGES OF THE EQUIVALENT FLUCTUATION VOLTAGE GENERATOR, E_n

* Electrical Engineering Department, Engineering Building, Columbia University, New York City 27, N. Y.

[1] Johnson, Phys. Rev., 32, 97 (1928).
[2] Nyquist, Phys. Rev., 32, 110 (1928).
[3] Moullin and Ellis, Jour. I. E. E., 74, 323 (1934).

the form of a capacitance C. This will give the following energy relation:

$$\tfrac{1}{2}C\overline{E}_c^2 = \tfrac{1}{2}KT \qquad (2)$$

where

$\overline{E}_c^2$ = mean-square voltage across the condenser

Solving equation (2) for the mean-square voltage across the condenser, the value which results is:

$$\overline{E}_c^2 = KT/C \text{ volts square} \qquad (3)$$

As an example, consider the input circuit of an audio amplifier in which the input capacitance is 25 $\mu\mu$f and the temperature of the circuit is 27 degrees C. Substituting the values yields the result:

$$\overline{E}_c^2 = (1.372)(10^{-23})(273 + 27)/(25)(10^{-12})$$
$$= 165 \times 10^{-12} \text{ volts square}$$
$$E_c = 12.8 \text{ microvolts}$$

This voltage would appear across the input of the amplifier provided the entire frequency spectrum from zero to infinity frequency were considered. A practical amplifier will actually amplify only a fraction of this value because the frequency range is finite but this value of 12.8 microvolts is the maximum value the fluctuation voltage can reach.

It is interesting to note the frequency spectrum of the fluctuation voltages across the condenser. The source of these voltages is a random series of pulse voltages represented by an equivalent generator having a uniform frequency spectrum. It can be shown by consideration of equation (3) that this equivalent generator produces a voltage whose mean-square value over a frequency bandwidth df is:

$$d\overline{E}_n^2 = 4KTR \text{ df volts square} \qquad (4)$$

This is the classical equation for fluctuation voltages generated by a resistance. When these voltages are considered in the circuit of Fig. 3 and are integrated from

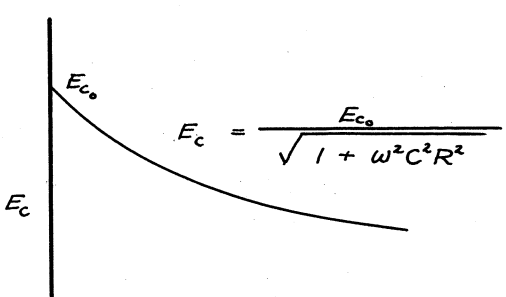


FIG. 3. FREQUENCY SPECTRUM OF THE FLUCTUATION VOLTAGES ACROSS THE CONDENSER OF THE CIRCUIT SHOWN IN FIGURE 2

zero frequency to infinity frequency, the mean-square voltage across the condenser always comes to the value given by equation 3. At the same time, the relative value of the fluctuation voltage content across the condenser as a function of frequency is the same as the frequency characteristic of the circuit. A plot of the frequency spectrum is given in Fig. 3.

It is recognized that equation 4 is the classical fluctuation voltage formula and in most cases is used for computation. Yet in many cases, it is simpler to return to the basic energy concept. At any rate, the simplicity of equation 3 as an outside value of fluctuation voltage makes it very useful.

Fluctuation Voltages in the Tuned Circuit ★ Since the input circuit of a receiver usually has a tuned circuit as shown in Fig. 4 across the

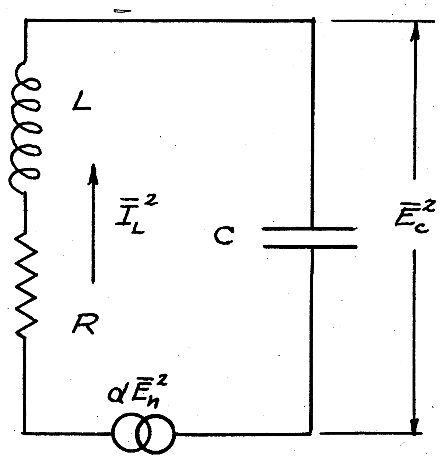


FIG. 4. TUNED CIRCUIT FOR WHICH THE FLUCTUATION VOLTAGES ACROSS THE CONDENSER ARE DETERMINED

grid of the first tube, a study of the fluctuation voltages of this circuit should be examined. Using the basic energy theory, there are *two* independent energy storage modes; the inductance which stores energy by virtue of current flow, and the condenser which stores energy by virtue of the voltages across it. Thus, the relation which holds in this case is:

$$\tfrac{1}{2}L\overline{I}^2 + \tfrac{1}{2}C\overline{E}_c^2 = 2(\tfrac{1}{2}KT) \qquad (4)$$

where: $\overline{E}_c^2$ = mean-square voltage across the condenser

$\overline{I}^2$ = mean-square current in the inductance

These voltages and currents are made up of pulses whose frequency spectra have been altered by the circuit. The pulses of the electron gas are virtually of infinitely short duration and contain all the frequencies of the spectrum. On the other hand the spectra of the voltages and currents included in equation 4 will have the frequency characteristics of the tuned circuit. These spectra are shown in Fig. 5.

In such a circuit, it may be shown that

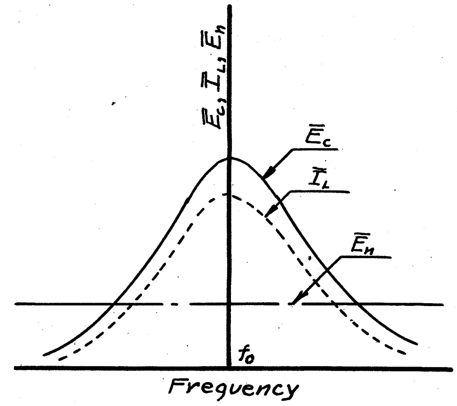


FIG. 5. SPECTRUM OF THE FLUCTUATION VOLTAGES ACROSS THE CONDENSER AND EQUIVALENT GENERATOR AND FLUCTUATION CURRENTS IN THE INDUCTANCE

the energy stored in the inductance is equal to that stored in the condenser if the entire frequency spectrum is considered. Thus:

$$\tfrac{1}{2}L\overline{I}^2 = \tfrac{1}{2}C\overline{E}_c^2 \qquad (5)$$

Then, from equation 4,

$$C\overline{E}_c^2 = KT \qquad (6)$$

and the mean-square voltage across the condenser is,

$$\overline{E}_c^2 = \frac{KT}{C} \text{ volts square} \qquad (7)$$

where K = 1.372(10^{-23}) joules/degree
T = (273 + deg C) degrees Kelvin
C = farads

This extremely simple relation is independent of resistance in the circuit and represents the mean-square value of *all* the frequency components from zero to infinity. More practically, it covers all the frequency components from virtually d-c to frequencies many times the resonant value.

As an example, consider the following constants:

C = 50 $\mu\mu$f
L = 10 μh
R = 60 ohms
T = 27 deg C.

$$\overline{E}_c^2 = \frac{(1.372)(10^{-23})(273 + 27)}{(50)(10^{-12})} \text{volts square}$$
$$\overline{E}_c^2 = 82.3 \text{ microvolts square}$$
$$E_c = 9.08 \text{ microvolts}$$

This value of 9.08 microvolts is the root-mean-square value of the fluctuation voltage across the condenser and contains components at frequencies in relative values as shown in Fig. 6.

In a receiver, the circuits following the input circuit (such as the IF amplifier) are usually sharper than the input circuit. Certainly, they are not so broad as to extend from zero to infinity as is implied in equation (7). Hence a "band-width fac-

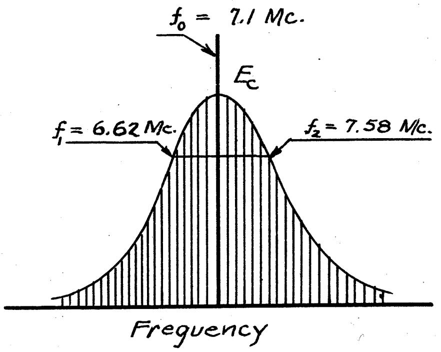


FIG. 6. FREQUENCY SPECTRUM OF THE FLUCTUATION VOLTAGES ACROSS THE CONDENSER IN THE ILLUSTRATIVE PROBLEM. THE BANDWIDTH OF THIS CIRCUIT IS 480 KC. ON EITHER SIDE OF RESONANCE

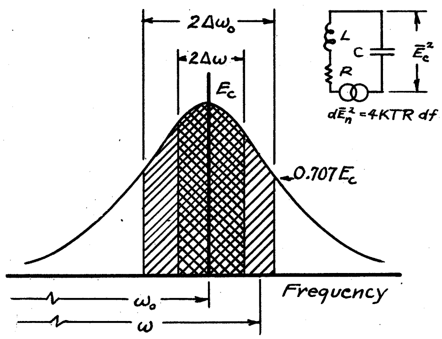


FIG. 7. FREQUENCY SPECTRUM OF FLUCTUATION VOLTAGES SHOWING THE RELATION OF THE R.F. BANDWIDTH ($2\Delta\omega_o$) AND THE BANDWIDTH ($2\Delta\omega$) OF THE INTERMEDIATE FREQUENCY AMPLIFIER FOLLOWING THE RF STAGE

tor" must be added to the expression. Referring to the nominal input circuit band-width as the radian frequency span ($2\,\Delta\omega_o$) over which the resonant voltage drops 3 db and the band-width of the subsequent circuits as $2\Delta\omega$ the effect on the total noise can be computed. The relation between these defined values are shown graphically in Fig. 7. To obtain the equation of the spectrum curve of Fig. 7, the fluctuation voltages will be considered as being produced by a pulse generator ($\overline{E}_n^2$) whose frequency spectrum is uniform. Thus:

$$d\overline{E}_c^2 = \frac{d\overline{E}_n^2 \dfrac{1}{\omega^2 C^2}}{R^2 + \left(\omega L - \dfrac{1}{\omega C}\right)^2} \qquad (8)$$

Defining: $$\omega_o = \frac{1}{\sqrt{LC}} \qquad (9)$$

$$Q = \frac{1}{\omega_o CR} = \frac{\omega_o L}{R} \qquad (10)$$

and noting that:

$$d\overline{E}_n^2 = 4KTR\,df \qquad (4)$$

and assuming that $\Delta\omega$ is reasonably small compared to ω_o (about 10%), a close approximation to this integral is:

$$\overline{E}_c^2 = \frac{KT}{C}\frac{1}{\pi}\int_{\omega_o - \Delta\omega}^{\omega_o + \Delta\omega} \frac{2\,Q/\omega_o\, d\omega}{1 + \dfrac{4Q^2}{\omega_o^2}(\omega - \omega_o)^2} \qquad (11)$$

The useful relations which can be derived from the equivalent circuit given in Fig.

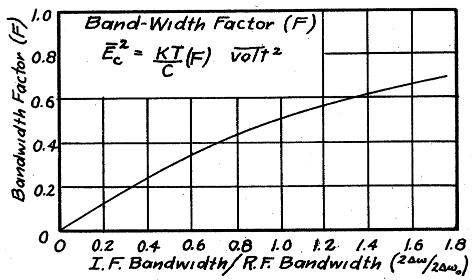


FIG. 8. PLOT OF BANDWIDTH FACTOR, F, VERSUS BANDWIDTH RATIO

9c are the resonant gain of the circuit as well as the effective Q or bandwidth.

$$Q_\epsilon = \frac{\omega L_2}{R_2 + R_a'} = \frac{\omega L_2/R_2}{1 + R_a'/R_2} = \frac{Q_2}{1 + \dfrac{\omega^2 M^2}{R_a R_2}}$$

$$Q_\epsilon = \frac{Q_2}{1 + x^2} = \frac{\omega_o}{2\Delta\omega_o} \qquad (17)$$

where $Q_2 = \omega L_2/R_2 = 1/\omega C_2 R_2 =$ Q of secondary circuit

$x^2 = \dfrac{\omega^2 M^2}{R_a R_2}$ = coupling factor

ω_o = resonant frequency

$2\Delta\omega_o$ = total bandwidth to half-energy points

also,

$$E_2 = \frac{\dfrac{1}{j\omega C_2}\, j\omega M \dfrac{E_a}{R_a}}{R_a' + R_2 + j\left(\omega L_2 - \dfrac{1}{\omega C_2}\right)}$$

(at resonance) (18)

which can be simplified to

$$E_2 = \frac{Q_2}{1 + x^2}\, x\sqrt{\frac{R_2}{R_a}}\, E_a = Q_\epsilon\, x\sqrt{\frac{R_2}{R_a}}\, E_a \qquad (19)$$

Equation 17 gives the effective Q of the system and the resultant RF bandwidth while equation 19 gives the gain of the circuit at the resonant frequency.

The signal-to-noise ratio may now be computed using equation 12b and equations 17 and 19:

$$\frac{\overline{E}_2^2}{\overline{E}_c^2} = \frac{Q_\epsilon^2 x^2 R_2/R_a}{(KT/C_2)(F)}\,\overline{E}_a^2 \qquad (20)$$

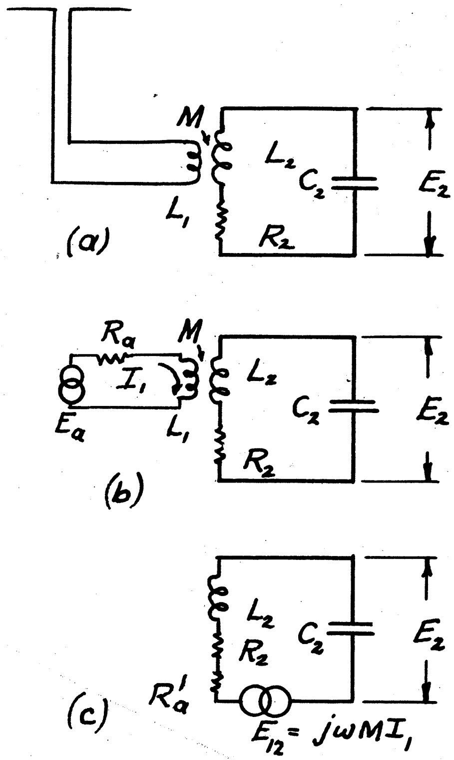


FIG. 9. TYPICAL INPUT CIRCUIT OF A RECEIVER SHOWING THE EQUIVALENT CIRCUITS USED FOR THE COMPUTATION OF THE SIGNAL VOLTAGE, E_2

Simplifying,

$$\frac{\overline{E}_2^2}{\overline{E}_c^2} = \frac{1}{2FKTR_a(2\Delta\omega_o)}\left[\frac{2x^2}{1 + x^2}\right]\overline{E}_a^2 \qquad (21)$$

Now if the following values are assigned:

$K = 1.372 \times 10^{-23}$ joules/degree Kelvin

$T = (273 + 27)$ degrees Kelvin (absolute)

Then by substitution and simplification,

$$\frac{\overline{E}_2}{\overline{E}_c} = \frac{\text{Signal Volts}}{\text{Noise Volts}} = \frac{4400}{\sqrt{(2\Delta f_o)FR_a}}\sqrt{\frac{2\,x^2}{1 + x^2}}\,\overline{E}_{a\mu v} \qquad (22)$$

where

$(2\Delta f_o)$ = Nominal RF bandwidth for 3 db loss of gain

F = Bandwidth factor (See Figure 8)

R_a = Antenna resistance

$x = \dfrac{\omega_o M}{\sqrt{R_a R_2}}$ = Coupling factor

$\overline{E}_{a\mu v}$ = R.M.S. signal voltage in microvolts

A plot of equation 22 is shown in Fig. 10 in which it is observed that the signal-to-noise ratio improves as the coupling is increased. High degrees of coupling are not

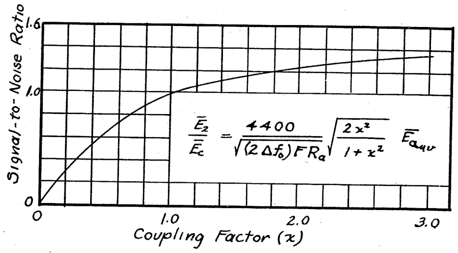


FIG. 10. CURVE OF SIGNAL-TO-NOISE RATIO FOR THE INPUT CIRCUIT OF A RECEIVER PLOTTED VERSUS COUPLING FACTOR, x ($x = \omega M/\sqrt{R_a R_2}$)

always attainable nor desirable. For instance, in the case of receivers which are tuned over a wide range, the antenna impedance varies over considerable limits. In order to prevent these variations from affecting the input circuit of the receiver excessively, the antenna is loosely coupled to the input with a resultant reduction in signal-to-noise ratio. In the case of fixed frequency receivers, however, heavy coupling is desirable from the point of view of noise reduction.

As an illustrative example of the previous conclusions, consider the following requirements:

f_o = Mid-frequency = 50 Megacycles/sec.

$2\Delta f_o$ = RF Bandwidth = 2 Megacycles/sec.

R_a = Antenna Resistance = 100 ohms

F = Bandwidth Factor = 0.5

The bandwidth factor indicates that the I.F. amplifier bandwidth is 2 megacycle/sec. Using equation 22, the signal-to-noise ratio is:

$$\frac{\text{Signal volts}}{\text{Noise volts}} = \frac{4400}{\sqrt{(2)(10^6)(0.5)(100)}} \sqrt{\frac{2x^2}{1+x^2}}\, \bar{E}_{a\mu v}$$

$$= 0.44\sqrt{\frac{2x^2}{1+x^2}}\, \bar{E}_{a\mu v}$$

For instance, if critical coupling ($x = 1$) is used, the signal-to-noise ratio becomes:

$$\frac{\text{Signal Volts}}{\text{Noise Volts}} = 0.44\, E_{a\mu v}$$

Thus, if an antenna signal of 2.28 microvolts is applied to the input of the receiver, the input circuit yields a signal-to-noise ratio of unity.

The choice of the circuit constants making up the input circuit is made as follows (Figure 9a):

whence:

$$\bar{E}_c^2 = \frac{KT}{\pi C}\left[\arctan\frac{2Q}{\omega_o}(\omega - \omega_o)\right]_{\omega_o - \Delta\omega}^{\omega_o + \Delta\omega}$$

$$= \frac{2KT}{\pi C}\arctan\frac{2\Delta\omega}{2\Delta\omega_o} \qquad (12a)$$

$$= \frac{KT}{C}(F) \qquad (12b)$$

where $2\Delta\omega_o$ = Input circuit bandwidth for 3 db loss in gain

$2\Delta\omega$ = Bandwidth of circuits following input stage

Thus, the correction factor which must be applied to the complete spectrum formula (equation 7) is the "bandwidth factor," F:

$$F = \frac{2}{\pi}\arctan\frac{2\Delta\omega}{2\Delta\omega_o} \qquad (13)$$

This factor is plotted in Fig. 8. It is observed that in a "consistent" design in which the input circuit bandwidth is the same as that of the subsequent circuits, this factor is 0.5. Thus taking the illustrative example previously given, the values for a "consistent" condition become:

$\bar{E}_c^2 = 82.3 \times \frac{1}{2} = 41.15$ microvolts square

and $\bar{E}_c = 6.42$ microvolts

Equation 12b is directly applicable to the design of the input circuit of a receiver. Essentially it assumes that the gain of the I.F. amplifier is constant over the pass band although if it is not exactly so, only small error results.

Signal-to-Noise Ratio in Input Circuit ★ In the ultimate analysis, it is the ratio of the useful signal to the fluctuation noise that is important. To make such an analysis, the input circuit will be assumed a simple inductively coupled circuit as shown in Fig. 9. The antenna is assumed matched to a transmission line (or tuned) so that the equivalent antenna impedance is a pure resistance R_a. It is further assumed that the primary reactance (ωL_1) is negligible compared to R_a. If this is true, then it is accurate enough to say that:

$$I_1 = \frac{E_a}{R_a} \qquad (14)$$

Also, referring to Figs. 9b and 9c,

$$E_{12} = j\omega M I_1 = j\omega M\frac{E_a}{R_a} \qquad (15)$$

and the reflected primary impedance in the secondary circuit R'_a is:

$$R'_a = \frac{X_{12}}{R_a} = \frac{\omega^2 M^2}{R_a} \qquad (16)$$

$$Q_\epsilon = \frac{f_o}{2\Delta f_o} = \frac{50}{2} = 25$$

(From equation 17)

and for critical coupling,

$$Q_2 = 2Q_\epsilon = 50$$

(From equation 17)

If the tuning condenser C_2 is 20 $\mu\mu f$, then the inductance should be:

$$L_2 = \frac{1}{\omega_o^2 C_2} = 0.506\ \mu\text{hys}$$

and its effective resistance should be: (to make $Q_2 = 50$)

$$R_2 = \frac{\omega_o L_2}{Q_2} = 3.18\ \text{ohms}$$

The mutual inductance should be: (to make $x = 1$)

$$M = \frac{1}{\omega_o}\sqrt{R_a R_2} = 0.057\ \mu\text{hys}$$

If the coefficient of coupling is made 0.25, the primary or antenna coupling coil inductance is:

$$L_1 = \frac{M^2}{kL_2} = 0.0256\ \mu\text{hys.}$$

Effect of Tube Noise ★ The signal-to-no ratio as given by equation 22 yields the effect of the input circuit only. If the first tube following this circuit is effective in producing gain, the fluctuation noise contributed by the tube may be neglected. On the other hand, if the tube is noisy, its effect must be included in the total signal-to-noise ratio. The subject of the origin and computation of fluctuation voltages in diodes, triodes and pentodes has been covered in a series of papers [1] by Thompson, North, Harris and others. The fluctuation voltages produced by the tube may be simulated by an equivalent generator as shown in Fig. 11 and values of the voltage $\bar{E}_t^2$ may be found in one of the series of papers previously given.[2] It may be pointed out here that the tube fluctuation voltage is dependent on the bandwidth of the circuits following the tube which means that in the analyses previously presented, it would be proportional to the IF bandwidths chosen.

Thus, referring to Fig. 11,

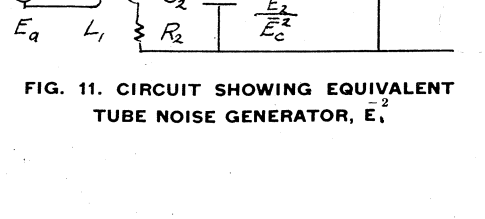


FIG. 11. CIRCUIT SHOWING EQUIVALENT TUBE NOISE GENERATOR, $\bar{E}_t^2$

[1] B. J. Thompson, D. O. North, W. A. Harris, "Fluctuations in Space-Charge-Limited Currents at Moderately High Frequencies," RCA Review. In five parts, 1940 and 1941.

[2] Part V of papers quoted in (1).

$$\overline{E}_n^2 = \overline{E}_t^2 + \overline{E}_c^2 \qquad (23)$$

where

$\overline{E}_t^2$ = Tube noise voltage squared

$\overline{E}_c^2$ = Input circuit noise voltage squared

$\overline{E}_n^2$ = Total noise input voltage squared

Thus, the total signal-to-noise ratio may be written:

$$\frac{\overline{E}_2^2}{\overline{E}_n^2} = \frac{\overline{E}_2^2}{\overline{E}_t^2 + \overline{E}_c^2} = \frac{(\overline{E}_2^2/\overline{E}_c^2)}{1 + (\overline{E}_t^2/\overline{E}_c^2)} \qquad (24)$$

or (Total signal-to-noise ratio)^2 =

$$\left[\frac{(\text{Circuit signal-to-noise ratio})^2}{1 - (\text{Tube noise/Circuit noise})^2}\right] \qquad (25)$$

There are several implications in equations 24 and 25 concerning the relative effects of the circuit and tube noise. For instance:

1. If the tube noise is negligible, equation 24 shows that the total signal-to-noise ratio is the same as the circuit signal-to-noise ratio. For this condition, overcoupling is desirable.

2. If the tube noise is very large as compared to the circuit noise, then equation 24 becomes:

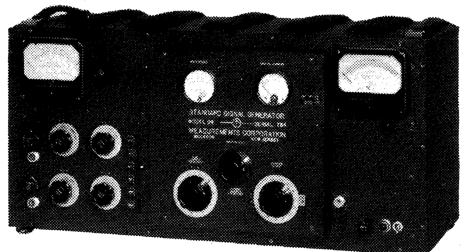
$$\frac{\overline{E}_2^2}{\overline{E}_n^2} = \frac{\overline{E}_2^2}{\overline{E}_t^2} \qquad (26)$$

This equation indicates that the circuit noise need not be considered as a factor in design in those cases where tube noise is large as compared to circuit noise. On the other hand, to increase the signal-to-noise ratio, it is desirable to make the useful signal, E_2, as large as possible. If bandwidth is not a consideration, this may be accomplished by using a value of coupling as close to critical (x = 1) as possible. However, if the bandwidth is to be kept at a fixed value (Q_ϵ = constant), equation 19 indicates that overcoupling will yield the maximum signal voltage, E_2.

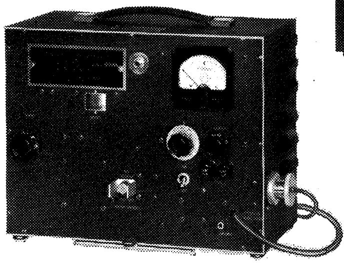
3. If the tube and circuit fluctuation voltage values are of comparable value, equation 24 must be considered in its entirety. If it is necessary that the bandwidths of the RF and IF sections be maintained at a given value, then overcoupling is desirable to raise the signal-to-noise ratio. However, if the RF bandwidth is not considered, it will follow that a degree of coupling somewhere between critical and overcoupling will be optimum. An exact value of the best coupling in this case is not easily determined since it depends on arbitrary choices of circuit constants such as the value of the tuning condenser and best values of Q which are obtainable.

Summary ★ There are a number of points which have been stressed. In the first place, the analogy between Brownian movement and fluctuation voltage phenomena has been reiterated. The simplicity resulting in the use of this concept in calculating the fluctuation voltage across a tuned circuit is of importance. Considering the complete frequency spectrum, the fluctuation voltage squared is given simply as KT/C. If only a portion of this spectrum is amplified by the following intermediate-frequency amplifiers, the bandwidth factor F must also be introduced. Finally, it has been shown that if receiver bandwidth is to be maintained, overcoupling is very desirable in all cases to improve the signal-to-noise ratio. If receiver bandwidth is not of consequence, and input tube noise is high, then it is desirable to use critical coupling.

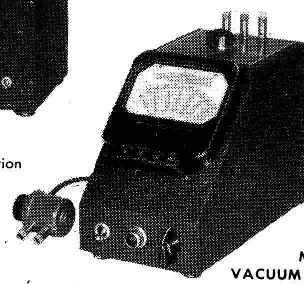
Laboratory Standards



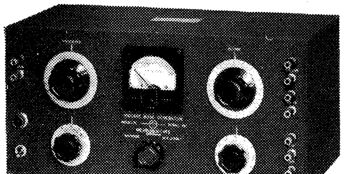
MODEL 84
U.H.F. STANDARD SIGNAL GENERATOR
300 to 1000 megacycles, AM and Pulse Modulation



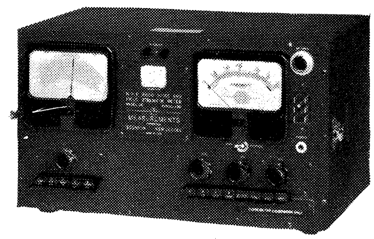
MODEL 78-B STANDARD SIGNAL GENERATOR
Two Frequency Bands between 15 and 250 megacycles



MODEL 62
VACUUM TUBE VOLTMETER
0 to 100 volts AC, DC and RF

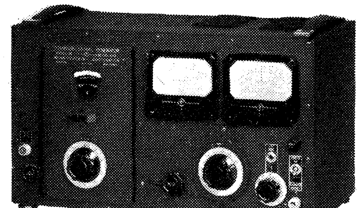


MODEL 71 SQUARE WAVE GENERATOR
5 to 100,000 cycles
Rise Rate 400 volts per microsecond

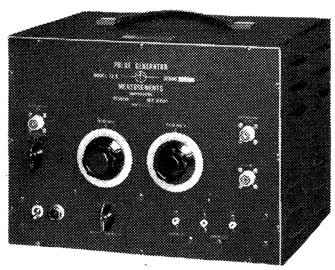


MODEL 58 U.H.F. RADIO NOISE AND FIELD STRENGTH METER
15 to 150 megacycles

MODEL 65-B
STANDARD SIGNAL GENERATOR
75 to 30,000 kilocycles
M.O.P.A., 100% Modulation



MODEL 80
STANDARD SIGNAL GENERATOR
2 to 400 megacycles
AM and Pulse Modulation



MODEL 79-B PULSE GENERATOR
50 to 100,000 cycles
0.5 to 40 microsecond pulse width

Standards are only as reliable as the reputation of their maker.

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