4

Noise

4.1 Introduction

Noise, as commonly understood, is a disturbance one "hears," but in telecommunications the word noise is also used as a label for the electrical disturbances that give rise to audible noise in a system. These electrical disturbances also appear as interference in video systems, for example, the white flecks seen on a television picture when the received signal is weak, referred to as a "noisy picture."

Noise can arise in a variety of ways. One obvious example is when a faulty connection exists in a piece of equipment, which, if it is a radio receiver, results in an intermittent or "crackling" type of noise at the output. Such sources of noise can, in principle anyway, be eliminated. Noise also occurs when electrical connections that carry current are made and broken, as, for example, at the brushes of certain types of motors. Again in principle, this type of noise can be suppressed at the source.

Natural phenomena that give rise to noise include electric storms, solar flares, and certain belts of radiation that exist in space. Noise arising from these sources may be more difficult to suppress, and often the only solution is to reposition the receiving antenna to minimize the received noise, while ensuring that reception of the desired signal is not seriously impaired.

Noise is mainly of concern in receiving systems, where it sets a lower limit on the size of signal that can be usefully received. Even when precautions are taken to eliminate noise from faulty connections or that arising from external sources, it is found that certain fundamental sources of noise are present within electronic equipment that limit the receiver sensitivity. One might think that any signal, however small, could simply be amplified up to any desired level. Unfortunately, adding amplifiers to a receiving system also adds noise, and the signal-to-noise ratio, which is the significant quantity, may be degraded by the addition of the amplifiers. Thus the study of the fundamental sources of noise within equipment is essential if the effects of the noise are to be minimized.

4.2 Thermal Noise

It is known that the free electrons within an electrical conductor possess kinetic energy as a result of heat exchange between the conductor and its surroundings. The kinetic energy means that the electrons are in motion, and this motion in turn is randomized through collisions with imperfections in the structure of the conductor. This process occurs in all real conductors and is what gives rise to the conductors' resistance. As a result, the electron density throughout the conductor varies randomly, giving rise to a randomly varying voltage across the ends of the conductor (Fig. 4.2.1). Such a voltage may sometimes be observed in the flickerings of a very sensitive voltmeter. Since the noise arises from thermal causes, it is referred to as *thermal noise* (and also as *Johnson noise,* after its discoverer).

The average or mean noise voltage across the conductor is zero, but the root-mean-square value is finite and can be measured. (It will be recalled that a similar situation occurs for sinusoidal voltage, which has a mean value of zero and a finite rms value.) It is found that the mean-square value of the noise voltage is proportional to the resistance of the conductor, to its absolute temperature, and to the frequency bandwidth of the device measuring (or responding to) the noise. The rms voltage is of course the square root of the mean-square value.

Consider a conductor that has resistance R, across which a true rms measuring voltmeter is connected, and let the voltmeter have an ideal band-pass frequency response of bandwidth B_n as shown in Fig. 4.2.2. The subscript n signifies noise bandwidth, which for the moment may be assumed to be the same as the bandwidth of the ideal filter. The relationship between noise bandwidth and actual frequency response will be developed more fully later. The mean-square voltage measured on the meter is found to be

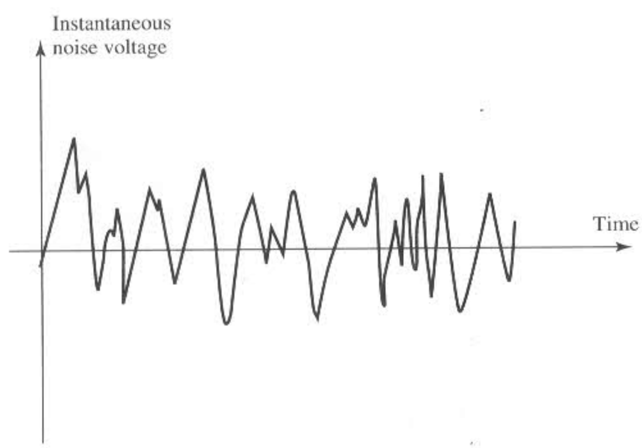


Figure 4.2.1 Thermal noise voltage.

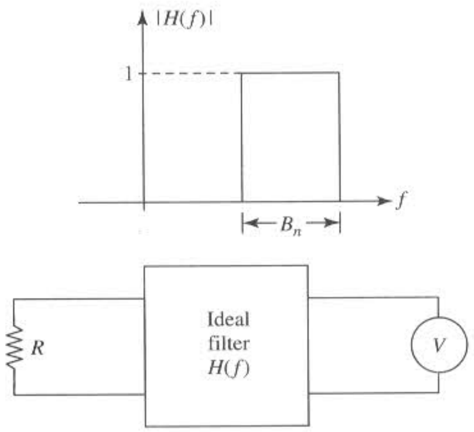


Figure 4.2.2 Measurement of thermal noise.

$$E_n^2 = 4RkTB_n \tag{4.2.1}$$

where E_n = root-mean-square noise voltage, volts
R = resistance of the conductor, ohms
T = conductor temperature, kelvins
B_n = noise bandwidth, hertz
k = Boltzmann's constant
$= 1.38 \times 10^{-23}$ J/K

The equation is given in terms of mean-square voltage rather than root mean square, since this shows the proportionality between the noise power (proportional to E_n^2) and temperature (proportional to kinetic energy).

The rms noise voltage is given by

$$E_n = \sqrt{4RkTB_n} \tag{4.2.2}$$

The presence of the mean-square voltage at the terminals of the resistance R suggests that it may be considered as a generator of electrical noise power. Attractive as the idea may be, thermal noise is not unfortunately a free source of energy. To abstract the noise power, the resistance R would have to be connected to a resistive load, and in thermal equilibrium the load would supply as much energy to R as it receives.

The fact that the noise power cannot be utilized as a free source of energy does not prevent the power being calculated. In analogy with any electrical source, the *available average power* is defined as the maximum average power the source can deliver. For a generator of emf E volts (rms) and internal resistance R, the available power is $E^2/4R$. Applying this to Eq. (4.2.1) gives for the available thermal noise power:

$$P_n = kTB_n \tag{4.2.3}$$

EXAMPLE 4.2.1

Calculate the thermal noise power available from any resistor at room temperature (290 K) for a bandwidth of 1 MHz. Calculate also the corresponding noise voltage, given that $R = 50\ \Omega$.

SOLUTION For a 1-MHz bandwidth, the noise power is

$$P_n = 1.38 \times 10^{-23} \times 290 \times 10^6$$
$$= 4 \times 10^{-15}\ \mathrm{W}$$
$$E_n^2 = 4 \times 50 \times 1.38 \times 10^{-23} \times 290$$
$$= 810^{-13}$$
$$\therefore E_n = \mathbf{0.895\ \mu V}$$

The noise power calculated in Example 4.2.1 may seem to be very small, but it may be of the same order of magnitude as the signal power present. For example, a receiving antenna may typically have an induced signal emf of 1 μV, which is of the same order as the noise voltage.

The thermal noise properties of a resistor R may be represented by the equivalent voltage generator of Fig. 4.2.3(a). This is one of the most useful representations of thermal noise and is widely used in determining the noise performance of equipment. It is best to work initially in terms of E_n^2 rather than E_n, for reasons that will become apparent shortly.

Norton's theorem may be used to find the equivalent current generator and this is shown in Fig. 4.2.3(b). Here, using conductance G (= $1/R$), the rms noise current I_n is given by

$$I_n^2 = 4GkTB_n \tag{4.2.4}$$

It will be recalled that the bandwidth is that of the external circuit, not shown in the source representations, and this must be examined in more detail. Suppose the resistance is left open circuited; then the bandwidth ideally would be infinite, and Eq. (4.2.3) suggests that the open-circuit noise voltage would also be infinite! Two factors prevent this from happening. The first relates to the derivation of the noise energy, which is based on classical thermodynamics and ignores quantum mechanical effects. The quantum mechanical derivation shows that the energy drops off with increasing frequency, and this therefore sets a fundamental limit to the noise power available. However, quantum mechanical effects only become important at

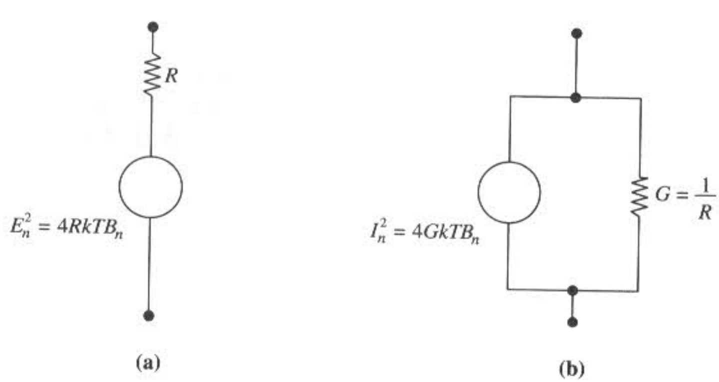


Figure 4.2.3 Equivalent sources for thermal noise: (a) voltage source and (b) current source.

frequencies well into the infrared region. The second and more significant practical factor from the circuit point of view is that *all* real circuits contain reactance (for example, self-inductance and self-capacitance), which sets a finite limit on bandwidth. In the case of the open-circuited resistor, the self-capacitance sets a limit on bandwidth, a situation that is covered in more detail later.

Resistors in Series

Let R_{ser} represent the total resistance of the series chain, where $R_{ser} = R_1 + R_2 + R_3 + \cdots$; then the noise voltage of the equivalent series resistance is

$$\begin{aligned} E_n^2 &= 4R_{ser}kTB_n \\ &= 4(R_1 + R_2 + R_3 + \cdots)kTB_n \\ &= E_{n1}^2 + E_{n2}^2 + E_{n3}^2 + \cdots \end{aligned} \tag{4.2.5}$$

This shows that the total noise voltage *squared* is obtained by summing the mean-square values. Hence the noise voltage of the series chain is given by

$$E_n = \sqrt{E_{n1}^2 + E_{n2}^2 + E_{n3}^2 + \cdots} \tag{4.2.6}$$

Note that simply adding the individual noise voltages would have given the wrong result.

Resistors in Parallel

With resistors in parallel it is best to work in terms of conductance. Thus let G_{par} represent the parallel combination where $G_{par} = G_1 + G_2 + G_3 + \cdots$; then

$$\begin{aligned} I_n^2 &= 4G_{par}kTB_n \\ &= 4(G_1 + G_2 + G_3 + \cdots)kTB_n \\ &= I_{n1}^2 + I_{n2}^2 + I_{n3}^2 + \cdots \end{aligned} \tag{4.2.7}$$

Again, it is to be noted that the mean-square values are added to obtain the total mean-square noise current. Usually, it is more convenient to work in terms of noise voltage rather than current. This is most easily done by first determining the equivalent parallel resistance from $1/R_{par} = 1/R_1 + 1/R_2 + 1/R_3 + \cdots$ and using

$$E_n^2 = 4R_{par}kTB_n \tag{4.2.8}$$

EXAMPLE 4.2.2

Two resistors of 20 and 50 kΩ are at room temperature (290 K). For a bandwidth of 100 kHz, calculate the thermal noise voltage generated by (a) each resistor, (b) the two resistors in series, and (c) the two resistors in parallel.

SOLUTION (a) For the 20-kΩ resistor

$$E_n^2 = 4 \times (20 \times 10^3) \times (4 \times 10^{-21}) \times (100 \times 10^3)$$
$$= 32 \times 10^{-12}\ \text{V}^2$$
$$\therefore E_n = \mathbf{5.66\ \mu V}$$

The voltage for the 50-kΩ resistor may be found by simple proportion:

$$E_n = 5.66 \times \sqrt{\frac{50}{20}}$$
$$= \mathbf{8.95\ \mu V}$$

(b) For the series combination, $R_{ser} = 20 + 50 = 70$ kΩ. Hence

$$E_n = 5.66 \times \sqrt{\frac{70}{20}}$$
$$= \mathbf{10.59\ \mu V}$$

(c) For the parallel combination, $R_{par} = \dfrac{20 \times 50}{20 + 50} = 14.29$ kΩ.

$$\therefore E_n = 5.66 \times \sqrt{\frac{14.29}{20}} = \mathbf{4.78\ \mu V}$$

Reactance

Reactances do not generate thermal noise. This follows from the fact that reactance cannot dissipate power. Consider an inductive or capacitive reactance connected in parallel with a resistor R (Fig. 4.2.4). In thermal equilibrium, equal amounts of power must be exchanged; that is, if the resistor supplies thermal noise power P_2 to the reactance, the reactance must supply thermal noise power $P_1 = P_2$ to the resistor. But since the reactance cannot dissipate power, the power P_2 must be zero, and hence P_1 must also be zero.

The effect of reactance on the noise bandwidth must, however, be taken into account, as shown in the next section.

Spectral Densities

Thermal noise falls into the category of power signals as described in Section 2.17, and hence it has a spectral density. As pointed out previously, the bandwidth B_n is a property of the external measuring or receiving system and is

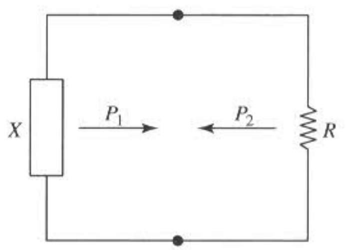


Figure 4.2.4 Power exchange between a reactance and a resistance is $P_1 = P_2 = 0$.

assumed flat so that, from Eq. (4.2.3), the available power spectral density, in watts per hertz, or joules, is

$$G_a(f) = \frac{P_n}{B_n} = kT \tag{4.2.9}$$

The spectral density for the mean-square voltage is also a useful function. This has units of volts^2 per hertz and is given by

$$G_v(f) = \frac{E_n^2}{B_n} = 4RkT \tag{4.2.10}$$

The spectral densities are flat, that is, independent of frequency, as shown in Fig. 4.2.5, and as a result thermal noise is sometimes referred to as *white noise,* in analogy to white light, which has a flat spectrum. When white noise is passed through a network, the spectral density will be altered by the shape of the network frequency response. The total noise power at the output is found by summing the noise contributions over the complete frequency range, taking into account the shape of the frequency response.

Consider a power spectral response as shown in Fig. 4.2.6. At frequency f_1, the available noise power for an infinitesimally small bandwidth δf about f_1 is $\delta P_{n1} = S_p(f_1)\delta f$. This is so because the bandwidth δf may be assumed flat about f_1, and the available power is given as the product of spectral density (watts/hertz) × bandwidth (hertz). The available noise power is therefore seen to be equal to the area of the shaded strip about f_1. Similar arguments can be applied at frequencies $f_2, f_3, \ldots$, and the total power, given by the sum of all these contributions, is equal to the sum of all these small areas, which is the total area under the curve. More formally, this is equal to the integral of the spectral density function over the frequency range $f = 0$ to $f = \infty$.

A similar argument can be applied to mean-square voltage. The spectral density curve in this case has units of V^2/Hz, and multiplying this by bandwidth δf Hz results in units of V^2, so the area under the curve gives the total mean-square voltage.

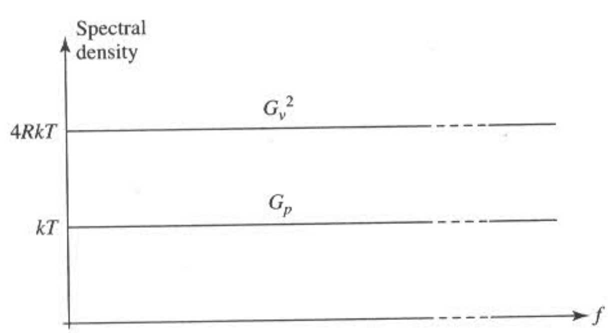


Figure 4.2.5 Thermal noise spectral densities.

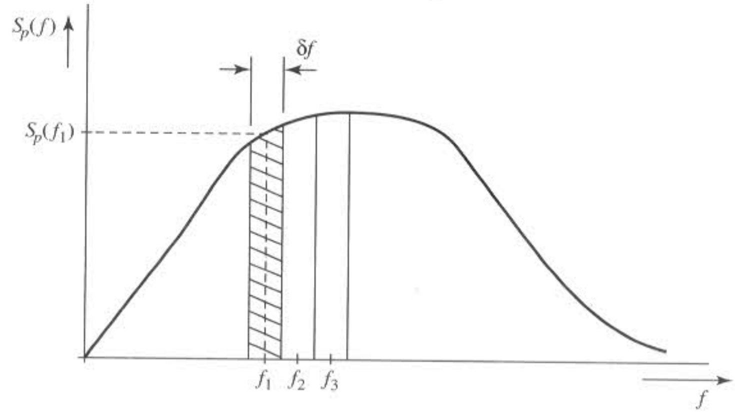


Figure 4.2.6 Nonuniform noise spectral density.

Equivalent Noise Bandwidth

Suppose that a resistor R is connected to the input of an LC filter, as shown in Fig. 4.2.7(a). This represents an input generator of mean-square voltage spectral density $4RkT$ feeding a network consisting of R and the LC filter. Let the transfer function of the network including R be $H(f)$, as shown in Fig. 4.2.7(b). The spectral density for the mean-square output voltage is therefore $4RkT|H(f)|^2$. This follows since $H(f)$ is the ratio of output to input voltage, and here mean-square values are being considered.

From what was shown previously, the total mean-square output voltage is given by the area under the output spectral density curve

$$\begin{aligned} V_n^2 &= \int_0^\infty 4RkT|H(f)|^2 df \\ &= 4RkT \times (\text{area under } |H(f)|^2 \text{ curve}) \end{aligned} \tag{4.2.11}$$

Now the total mean-square voltage at the output can be stated as $V_n^2 = 4RkTB_n$, and equating this with Eq. (4.2.11) gives, for the equivalent noise bandwidth of the network,

$$\begin{aligned} B_n &= \int_0^\infty |H(f)|^2 df \\ &= (\text{area under } |H(f)|^2 \text{ curve}) \end{aligned} \tag{4.2.12}$$

As a simple example consider the circuit of Fig. 4.2.8, which consists of a resistor in parallel with a capacitor. The capacitor may in fact be the self-capacitance of the resistor, or an external capacitor, for example, the input capacitance of the voltmeter used to measure the noise voltage across R.

The transfer function of the RC network is

$$|H(f)| = \frac{1}{\sqrt{1 + (\omega CR)^2}} \tag{4.2.13}$$

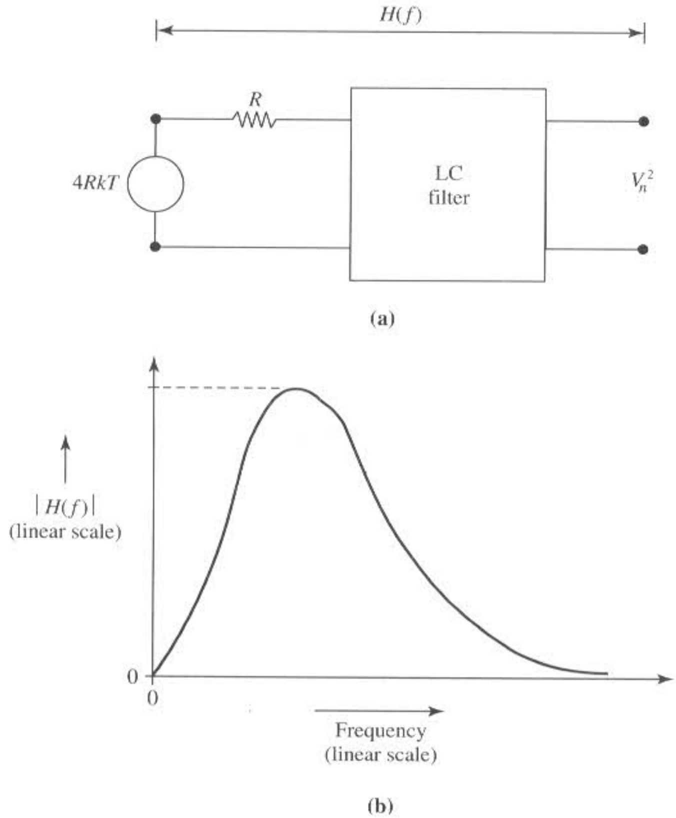


Figure 4.2.7 (a) Filtered noise and (b) the transfer function of the filter including R.

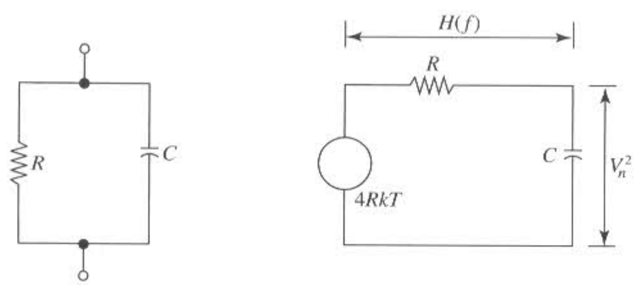


Figure 4.2.8 *RC* network and its transfer function used in determining noise bandwidth.

The equivalent noise bandwidth of the *RC* network is found using Eq. (4.2.12) as

$$B_n = \int_0^\infty |H(f)|^2 df$$

$$= \frac{1}{4RC} \tag{4.2.14}$$

(Details of the integration are left as an exercise for the reader.) The mean-square output voltage is given by

$$\begin{aligned} V_n^2 &= 4RkT \times \frac{1}{4RC} \\ &= \frac{kT}{C} \end{aligned} \tag{4.2.15}$$

This is a surprising result. It shows that the mean-square output voltage is independent of R, even though it originates from R, and it is inversely proportional to C, even though C does not generate noise.

A second example is that of the tuned circuit shown in Fig. 4.2.9. Here the capacitor is assumed lossless, and the inductor has a series resistance r that generates thermal noise.

The transfer function in this case is

$$|H(f)| = \left|\frac{X_c}{Z_s}\right| \tag{4.2.16}$$

where $Z_s = r(1 + jyQ)$ is the impedance of the series tuned circuit as given by Eq. (1.3.10) and $X_c = 1/j\omega C$ is the reactance of C. As before, the equivalent noise bandwidth is found by solving Eq. (4.2.12).

Consider first the situation where the circuit is resonant at f_0, and the noise is restricted to a small bandwidth $\Delta f \ll f_0$ about the resonant frequency. The transfer function is then approximated by $|H(f)| \cong 1/\omega_0 Cr = Q$, and the area under the $|H(f)|^2$ curve over a small constant bandwidth δf is $Q^2\delta f$. Hence the mean-square noise voltage is

$$\begin{aligned} V_n^2 &= 4rkTB_n \\ &= 4rQ^2kT\delta f \\ &= 4R_D kT\Delta f \end{aligned} \tag{4.2.17}$$

Here, use is made of the relationship $Q^2 r = R_D$ developed in Section 1.4. This is an important result, because the bandwidth is often limited in practice to some small percentage about f_0. An example will illustrate this.

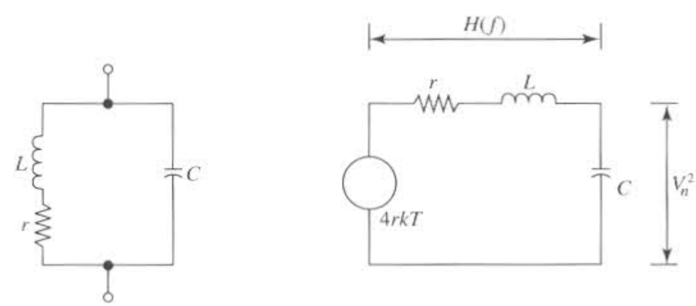


Figure 4.2.9 Tuned circuit and its transfer function used in determining noise bandwidth.

EXAMPLE 4.2.3

The parallel tuned circuit at the input of a radio receiver is tuned to resonate at 120 MHz by a capacitance of 25 pF. The Q-factor of the circuit is 30. The channel bandwidth of the receiver is limited to 10 kHz by the audio sections. Calculate the effective noise voltage appearing at the input at room temperature.

SOLUTION

$$R_D = \frac{Q}{\omega_o C}$$

$$= \frac{30}{2 \times \pi \times 120 \times 10^6 \times 25 \times 10^{-12}}$$

$$= 1.59 \text{ k}\Omega$$

$$\therefore V_n = \sqrt{4 \times 1.59 \times 10^3 \times 4 \times 10^{-21} \times 10^4}$$

$$= \mathbf{0.5\ \mu V}$$

Where the complete frequency range 0 to ∞ has to be taken into account, the integral becomes much more difficult to solve, and only the result will be given here. This is

$$B_n = \frac{1}{4R_D C} \tag{4.2.18}$$

where R_D is the dynamic resistance of the tuned circuit.

The noise bandwidth can be expressed as a function of the −3-dB bandwidth of the circuit. From Eq. (1.3.17), $B_{3\text{ dB}} = f_o/Q$, and from Eq. (1.4.4) $R_D = Q/\omega_o C$. Combining these expressions along with that for the noise bandwidth gives

$$B_n = \frac{\pi}{2} B_{3\text{ dB}} \tag{4.2.19}$$

By postulating that the noise originates from a resistor R_D and is limited by the bandwidth B_n, the mean-square voltage at the output can be expressed as

$$V_n^2 = 4R_D kT \times \frac{1}{4R_D C}$$

$$= \frac{kT}{C} \tag{4.2.20}$$

In the foregoing, to simplify the analysis it was assumed that the Q-factor remained constant, independent of frequency. This certainly would not be true for the range zero to infinity, but the end result still gives a good indication of the noise expected in practice.

For most radio receivers the noise is generated at the front end (antenna input) of the receiver, while the output noise bandwidth is determined by the audio sections of the receiver. The equivalent noise bandwidth is equal to the area under the normalized power-gain/frequency curve for the low-frequency sections. By normalized is meant that the curve is scaled such that the maximum value is equal to unity. Usually this information is available in the form of a frequency response curve showing output in decibels relative to maximum and with frequency plotted on a logarithmic scale, as sketched in Fig. 4.2.10(a).

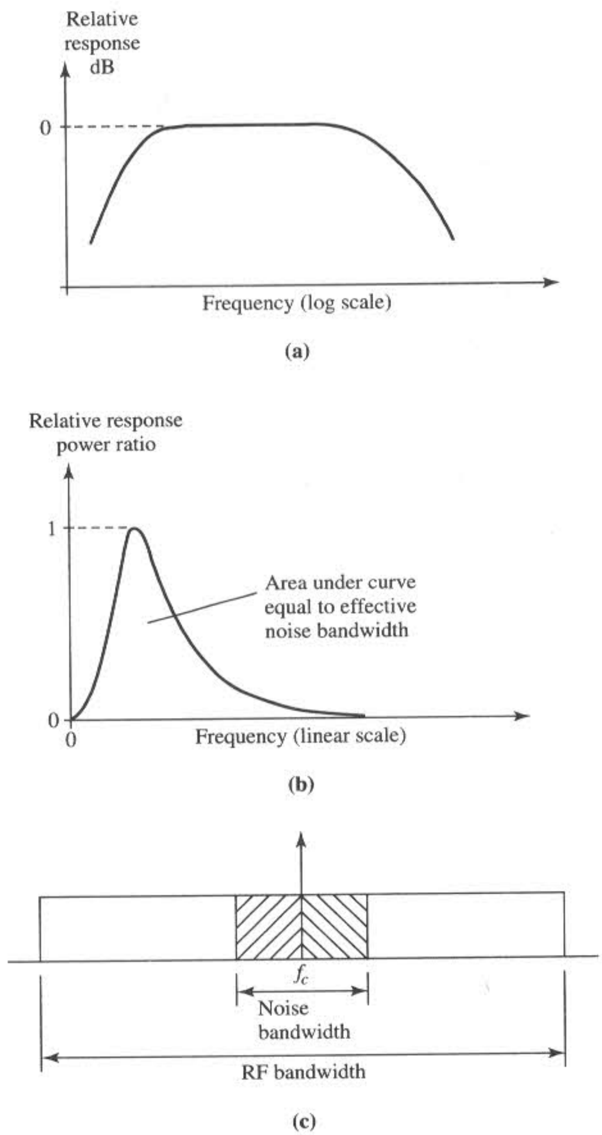


Figure 4.2.10 (a) Amplifier frequency response curve. (b) Curve of (a) using linear scales. (c) Noise bandwidth of a double-sideband receiver.

Before determining the area under the curve, the decibel axis must be converted to a linear power-ratio scale and the frequency axis to a linear frequency scale, as shown in Fig. 4.2.10(b). The equivalent noise bandwidth is then equal to the area under this curve for a single-sideband receiver. Where the receiver is of the double-sideband type, then the noise bandwidth appears on both sides of the carrier and is effectively doubled. This is shown in Fig. 4.2.10(c).

4.3 Shot Noise

Shot noise is a random fluctuation that accompanies any direct current crossing a potential barrier. The effect occurs because the carriers (holes and electrons in semiconductors) do not cross the barrier simultaneously, but rather with a random distribution in the timing for each carrier, which gives rise to a random component of current superimposed on the steady current. In the case of bipolar junction transistors, the bias current crossing the forward biased emitter–base junction carries shot noise. With vacuum tubes the electrons emitted from the cathode have to overcome a potential barrier that exists between cathode and vacuum. The name *shot noise* was first coined in connection with tubes, where the analogy was made between the electrons striking the plate and lead shot from a gun striking a target.

Although it is always present, shot noise is not normally observed during measurement of direct current because it is small compared to the dc value; however, it does contribute significantly to the noise in amplifier circuits. The idea of shot noise is illustrated in Fig. 4.3.1.

Shot noise is similar to thermal noise in that its spectrum is flat (except in the high microwave frequency range). The mean-square noise component is proportional to the dc flowing, and for most devices the mean-square, shot-noise current is given by

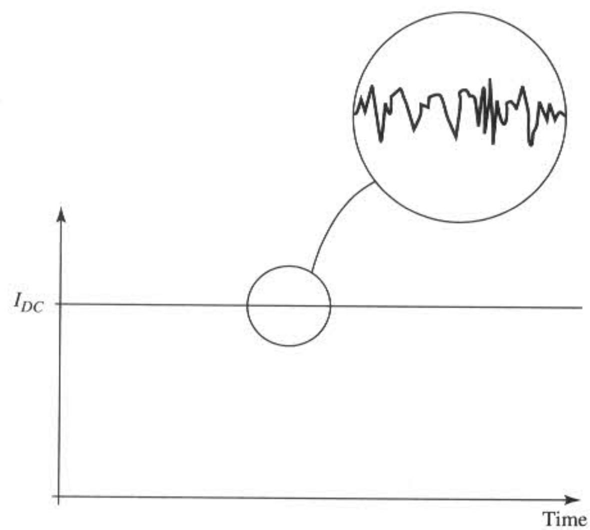


Figure 4.3.1 Shot noise.

$$I_n^2 = 2I_{dc}q_eB_n \quad \text{amperes}^2 \tag{4.3.1}$$

where I_{dc} is the direct current in amperes, q_e the magnitude of electron charge ($= 1.6 \times 10^{-19}$ C), and B_n is the equivalent noise bandwidth in hertz.

EXAMPLE 4.3.1

Calculate the shot noise component of current present on a direct current of 1 mA flowing across a semiconductor junction, given that the effective noise bandwidth is 1 MHz.

SOLUTION

$$\begin{aligned} I_n^2 &= 2 \times 10^{-3} \times 1.6 \times 10^{-19} \times 10^6 \\ &= 3.2 \times 10^{-16}\ \text{A}^2 \\ \therefore I_n &= \mathbf{18\ nA} \end{aligned}$$

4.4 Partition Noise

Partition noise occurs wherever current has to divide between two or more electrodes and results from the random fluctuations in the division. It would be expected therefore that a diode would be less noisy than a transistor (other factors being equal) if the third electrode draws current (such as base or gate current). It is for this reason that the input stage of microwave receivers is often a diode circuit, although, more recently, gallium arsenide field-effect transistors, which draw zero gate current, have been developed for low-noise microwave amplification. The spectrum for partition noise is flat.

4.5 Low Frequency or Flicker Noise

Below frequencies of a few kilohertz, a component of noise appears, the spectral density of which increases as the frequency decreases. This is known as *flicker noise* (and sometimes as 1/*f* noise). In vacuum tubes it arises from slow changes in the oxide structure of oxide-coated cathodes and from the migration of impurity ions through the oxide. In semiconductors, flicker noise arises from fluctuations in the carrier densities (holes and electrons), which in turn give rise to fluctuations in the conductivity of the material. It follows therefore that a noise voltage will be developed whenever direct current flows through the semiconductor, and the mean-square voltage will be proportional to the square of the direct current. Interestingly enough, although flicker noise is a low-frequency effect, it plays an important part in limiting the sensitivity of microwave diode mixers used for Doppler radar systems. This is because, although the input frequencies to the mixer are in the microwave range, the Doppler frequency output is in the low (audio-frequency) range, where flicker noise is significant.

4.6 Burst Noise

Another type of low-frequency noise observed in bipolar transistors is known as *burst noise,* the name arising because the noise appears as a series of bursts at two or more levels (rather like noisy pulses). When present in an audio system, the noise produces popping sounds, and for this reason is also known as "popcorn" noise. The source of burst noise is not clearly understood at present, but the spectral density is known to increase as the frequency decreases.

4.7 Avalanche Noise

The reverse-bias characteristics of a diode exhibit a region where the reverse current, normally very small, increases extremely rapidly with a slight increase in the magnitude of the reverse-bias voltage. This is known as the *avalanche region* and comes about because the holes and electrons in the diode's depletion region gain sufficient energy from the reverse-bias field to ionize atoms by collision. The ionizing process means that additional holes and electrons are produced, which in turn contribute to the ionization process, and thus the descriptive term *avalanche.*

The collisions that result in the avalanching occur at random, with the result that large noise spikes are present in the avalanche current. In diodes such as zener diodes, which are used as voltage reference sources, the avalanche noise is a nuisance to be avoided. However, avalanche noise is put to good use in noise measurements, as described in Section 4.19. The spectral density of avalanche noise is flat.

4.8 Bipolar Transistor Noise

Bipolar transistors exhibit all the sources of noise discussed previously, that is, thermal, shot, partition, flicker, and burst noise. The thermal noise is generated by the bulk or extrinsic resistances of the electrodes, but the only significant component is that generated by the extrinsic base resistance. It should be emphasised at this point that the small-signal equivalent resistances for the base – emitter and the base – collector junctions do not generate thermal noise, but they do enter into the noise calculations made using the small-signal equivalent circuit for the transistor.

The bias currents in the transistor show shot noise and partition noise, and, in addition, the flicker and burst noise components are usually associated with the base current.

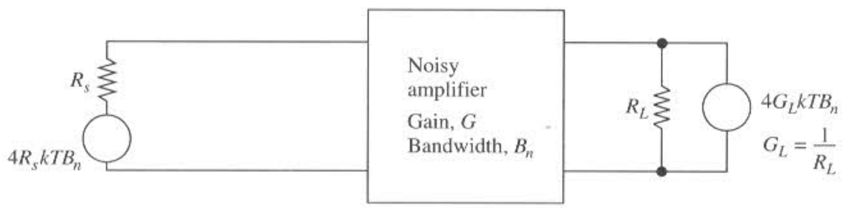
4.9 Field-effect Transistor Noise

In field-effect transistors (both JFETs and MOSFETs), the main source of noise is the thermal noise generated by the physical resistance of the drain–source channel. Flicker noise also originates in this channel. Additionally, there will be shot noise associated with the gate leakage current. This will develop a noise component of voltage across the signal–source impedance and is only significant where this impedance is very high (in the megohm range).

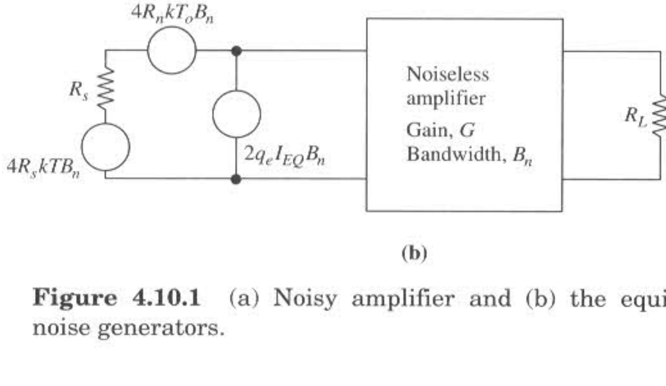
4.10 Equivalent Input Noise Generators and Comparison of BJTs and FETs

An amplifier may be represented by the block schematic of Fig. 4.10.1(a), in which a noisy amplifier is shown and where the source and load resistances generate thermal noise. The circuit may be redrawn as shown in Fig. 4.10.1(b) in which the amplifier itself is considered to be noiseless, the amplifier noise being represented by *fictitious noise generators* $V_{na} = \sqrt{4R_n k T_o B_n}$ and $I_{na} = \sqrt{2q_e I_{EQ} B_n}$ at the input. Here, B_n is the equivalent noise bandwidth of the amplifier in hertz, T_o is room temperature in kelvins, k is Boltzmann's constant $= 1.38 \times 10^{-23}$ J/K, and $q_e = 1.6 \times 10^{-19}$ C is the magnitude of the electron charge. These terms have all been defined previously. What is new here is the *fictitious* resistance R_n ohms, known as the *equivalent input noise resistance* of the amplifier, and I_{EQ} amperes, the *equivalent input shot noise current.* Both these parameters have to be calculated or specified for a transistor under given operating conditions.

The noise generated by the load resistance R_L is generally very small compared to the other sources and is assumed to be negligible, so this is dropped from the equivalent circuit. The thermal noise generated by the signal–source resistance R_s is generally significant and must be taken into account as shown in Fig. 4.10.1(b).



(a)



(b)

Figure 4.10.1 (a) Noisy amplifier and (b) the equivalent input noise generators.