SECTION 11

AMPLIFIERS AND OSCILLATORS

Amplifiers serve a number of purposes from allowing us to hear beautiful music to accurately positioning elements of complicated systems using control technologies. Oscillators are found in a number of applications from the watch on your wrist to the transmitter and receiver in your cell phone. We look at audio-frequency amplifiers and oscillators and radio-frequency amplifiers and oscillators.

The most versatile amplifier has to be the operational amplifier (op amp). The key to its success is that it is perhaps the most ideal device in analog electronics. Because of this it is found in a number of amplifier designs.

High-power amplifiers are necessary where significant amounts of power need to be used to accomplish activities such as radio and television broadcasts. Just imagine what a rock concert might sound like without power amplifiers. Microwave amplifiers and oscillators represent a special part of the high-power amplifier and oscillator field. C.A.

In This Section:

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CHAPTER 11.1
AMPLIFIER AND OSCILLATOR PRINCIPLES OF OPERATION

G. Burton Harrold

AMPLIFIERS: PRINCIPLES OF OPERATION

Gain

In most amplifier applications the prime concern is gain. A generalized amplifier is shown in Fig. 11.1.1. The most widely applied definitions of gain using the quantities defined there are:

$$\text{Voltage gain } A_v = e_{22}/e_{11} \qquad \text{Current gain } A_i = i_2/i_1$$

$$\text{Available power from source } P_{\text{avs}} = \frac{|e_s|^2}{4 \text{ Re } Z_s} \quad \text{where Re} = \text{real part of complex impedance}$$

$$\text{Output load power } P_L = \frac{|e_{22}|^2}{\text{Re } Z_L} \qquad \text{Input power } P_I = \frac{|e_{11}|^2}{\text{Re } Z_{\text{in}}}$$

$$\text{Available power at output } P_{\text{avo}} = \frac{|e_{22}|^2}{4 \text{ Re } Z_{\text{out}}} \qquad \text{Transducer gain } G_T = P_L/P_{\text{avs}}$$

$$\text{Available power gain } G_A = P_{\text{avo}}/P_{\text{avs}} \qquad \text{Power gain } G = P_L/P_I$$

$$\text{Insertion power gain } G_I = \frac{\text{power into load with network inserted}}{\text{power into load with source connected to load}}$$

Bandwidth and Gain-Bandwidth Product

Bandwidth is a measure of the range of frequencies within which an amplifier will respond. The frequency range (passband) is usually measured between the half-power (3-dB) points on the output-response-versus-frequency curve, for constant input. In some cases it is defined at the quarter-power points (6 dB). See Fig. 11.1.2.

The gain-bandwidth product of a device is a commonly used figure of merit. It is defined for a bandpass amplifier as

$$F_a = A_r B$$

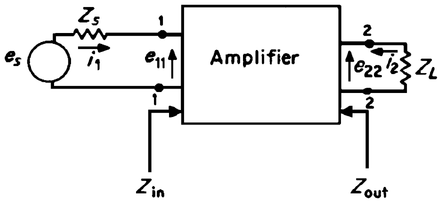


FIGURE 11.1.1 Input and output quantities of generalized amplifier.

where F_a = figure of merit (rad/s)
A_r = reference gain, either the maximum gain or the gain at the frequency where the gain is purely real or purely imaginary
B = 3-dB bandwidth (rad/s)

For low-pass amplifiers

$$F_a = A_r W_H$$

where F_a = figure of merit (rad/s)
A_r = reference gain
W_H = upper cutoff frequency (rad/s)

In the case of vacuum tubes and certain other active devices this definition is reduced to

$$F_a = g_m / C_T$$

where F_a = figure of merit (rad/s)
g_m = transconductance of active device
C_T = total output capacitance, plus input capacitance of subsequent stage

Noise

The major types of noise are illustrated in Fig. 11.1.3. Important relations and definitions in noise computations are:

Noise factor

$$F = \frac{S_i / N_i}{S_o / N_o}$$

where S_i = signal power available at input
S_o = signal power available at output
N_i = noise power available at input at $T = 290$ K
N_o = noise power available at output

Available noise power

$$P_{n,\text{av}} = \frac{e_n^2}{4R} = KTB \quad \text{for thermal noise}$$

where the quantities are as defined in Fig. 11.1.3.

Excess noise factor

$$F - 1 = N_e / N_i$$

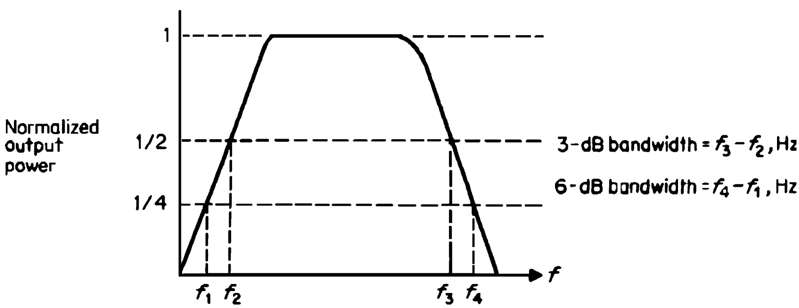
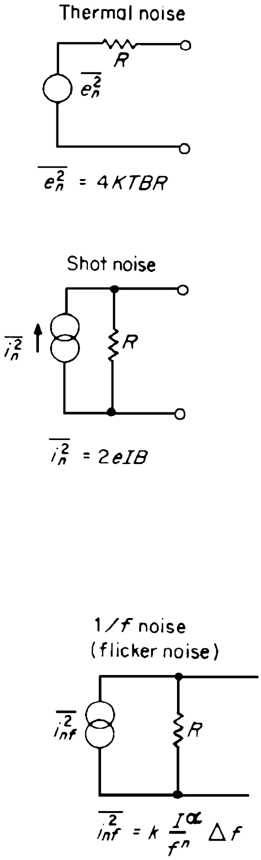


FIGURE 11.1.2 Amplifier response and bandwidth.



$\overline{e_n^2}$ = mean-square open-circuit noise voltage from a resistor R

$K = 1.38 \times 10^{-23}$ J/K

T = temperature, K

B = bandwith, Hz

R = resistance, Ω

$\overline{i_n^2}$ = mean-square short-circuit noise current

$e = 1.6 \times 10^{-19} C$

I = dc current amps through R

R = resistance, Ω

B = bandwith, Hz

$\overline{i_{nf}^2}$ = mean-square short-circuit flicker noise current

R = resistance, Ω

I = dc current

f = frequency, Hz

Δf = frequency interval

k, α, n = empirical constants depending on device and mode of operation

FIGURE 11.1.3 Noise-equivalent circuits.

where $F - 1$ = excess noise factor
N_e = total equivalent device noise referred to input
N_i = thermal noise of source at standard temperature

Noise temperature $$T = P_{n,\text{av}}/KB$$

where $P_{n,\text{av}}$ is the average noise power available.

At a single input-output frequency in a two-port,

Effective input noise temperature $$T_e = 290(F - 1)$$

Noise Factor of Transmission Lines and Attenuators. The noise factor of two ports composed entirely of resistive elements at room temperature (290 K) and an impedance matched loss of $L = 1/G_A$ is $F = L$.

Cascaded noise factor $$F_T = F_1 + (F_2 - 1)/G_A$$

where F_T = overall noise factor
F_1 = noise factor of first stage
F_2 = noise factor of second stage
G_A = available gain of first stage

System Noise Temperature. Space probes and satellite communication systems using low-noise amplifiers and antennas directed toward outer space make use of system noise temperatures. When we define T_A = antenna temperature, L = waveguide numeric loss (greater than 1), T_{E1} = amplifier noise temperature, G_A = amplifier available gain, F = postamplifier noise factor, and B = postamplifier bandwidth, this temperature can be calculated as

$$T_{\text{sys}} = T_A + |L-1|\,290° + LT_{E1} + \frac{(F-1)(290L)}{G_{A1}}$$

The quantity of interest is the output signal-to-noise ratio where S_A is available signal power at the antenna (assuming the antenna is matched to free space)

$$\text{S/N} = S_A / KT_{\text{sys}}B \qquad K = 1.38 \times 10^{-23}$$

Generalized Noise Factor. A general representation of noise performances can be expressed in terms of Fig. 11.1.4. This is the representation of a noisy two-port in terms of external voltage and current noise sources with a correlation admittance. In this case the noise factor becomes

$$F = 1 + \frac{G_u}{G_s} - \frac{R_N}{G_s}[(G_s + G_\gamma)^2 + (B_s + B_\gamma)^2]$$

where F = noise factor
G_s = real part of Y_s
B_s = imaginary part of Y_s
G_u = conductance owing to the uncorrelated part of the noise current
Y_γ = correlation admittance between cross product of current and voltage noise sources
G_γ = real part of Y_γ
B_γ = imaginary part of Y_γ
R_N = equivalent noise resistance of the noise voltage

The optimum source admittance is $Y_{\text{opt}} = G_{\text{opt}} + jB_{\text{opt}}$

$$G_{\text{opt}} = \left(\frac{G_u + R_N G_\gamma^2}{R_N}\right)^{1/2}$$

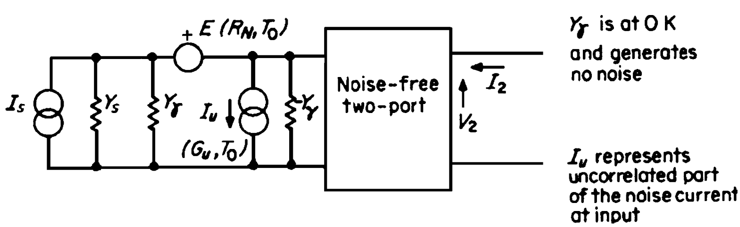


FIGURE 11.1.4 Noise representation using correlation admittance.

where $B_{opt} = -B_{gg}$ and the value of the optimum noise factor F_{opt} is

$$F_{opt} = 1 + 2R_N(G_\gamma + G_0)$$

The noise factor for an arbitrary source impedance is

$$F = F_{opt} + \frac{R_N}{G_s}[(G_s - G_0)^2 + (B_s - B_0)^2]$$

The values of the parameters of Fig. 11.1.4 can be determined by measurement of (1) noise figure versus B_s with G_s constant and (2) noise figure versus G_s with B_s at its optimum value.

Dynamic Characteristic, Load Lines, and Class of Operation

Most active devices have two considerations involved in their operation. The first is the dc bias condition that establishes the operating point (the *quiescent point*). The choice of operating point is determined by such considerations as signal level, uniformity of the device, and temperature of operation.

The second consideration is the ac operating performance, related to the slope of the dc characteristic and to the parasitic reactances of the device. These ac variations give rise to the *small-signal parameters.* The ac parameters may also influence the choice of dc bias point when basic constraints, such as gain and noise performance, are considered.

For frequencies of operation where these parasites are not significant, the use of a load line is valuable. The class of amplifier operation is dependent on its quiescent point, its load line, and input signal level. The types of operation are shown in Fig. 11.1.5.

Distortion

Distortion takes many forms, most of them undesirable. The basic causes of distortion are nonlinearity in amplitude response and nonuniformity of phase response. The most commonly encountered types of distortion are as follows:

Harmonic distortion is a result of nonlinearity in the amplitude transfer characteristics. The typical output contains not only the fundamental frequency but integer multiples of it.

Crossover distortion is a result of the nonlinear characteristics of a device when changing operating modes (e.g., in a push-pull amplifier). It occurs when one device is cut off and the second turned on if the crossover is not smooth between the two modes.

Intermodulation distortion is a spurious output resulting from the mixing of two or more signals of different frequencies. The spurious output occurs at the sum or difference of integer multiples of the original frequencies.

Cross-modulation distortion occurs when two signals pass through an amplifier and the modulation of one is transferred to the other.

Phase distortion results from the deviation from a constant slope of the output-phase–versus–frequency response of an amplifier. This deviation gives rise to echo responses in the output that precede and follow the main response, and a distortion of the output signal when an input signal having a large number of frequency components is applied.

Feedback Amplifiers

Feedback amplifiers fall into two categories: those having positive feedback (usually oscillators) and those having negative feedback. The positive-feedback case is discussed under oscillators. The following discussion is concerned with negative-feedback amplifiers.

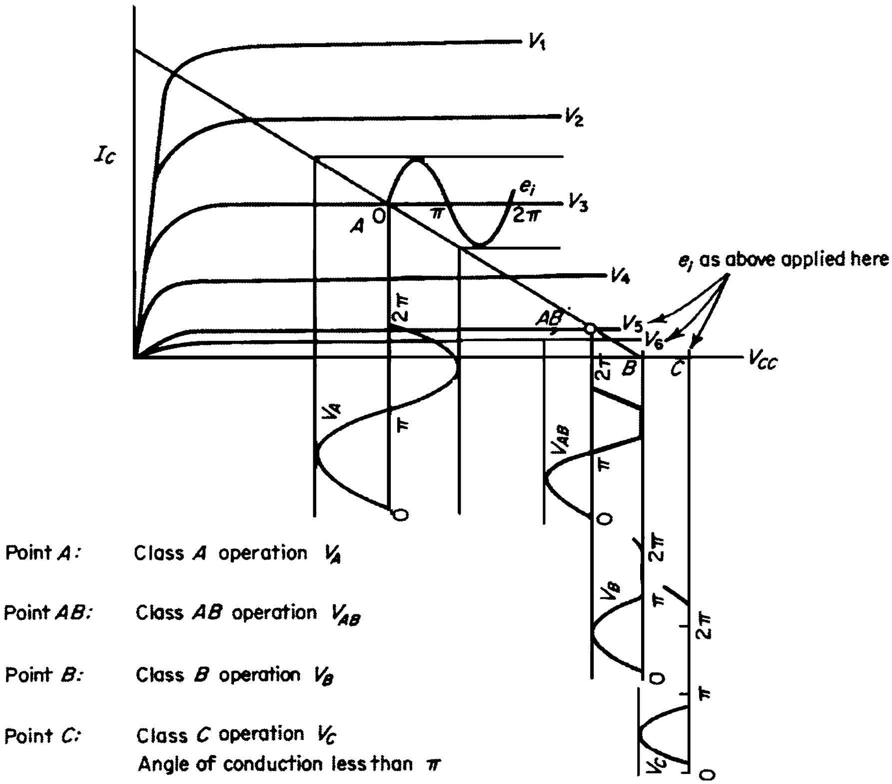


FIGURE 11.1.5 Classes of amplifier operation. Class S operation is a switching mode in which a square-wave output is produced by a sine-wave input.

Negative Feedback

A simple representation of a feedback network is shown in Fig. 11.1.6. The closed-loop gain is given by

$$e_2/e_1 = A/(1 - BA)$$

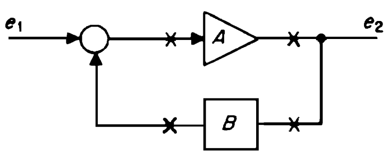


FIGURE 11.1.6 Amplifier with feedback loop.

where A is the forward gain with feedback removed and B is the fraction of output returned to input.

For negative feedback, A provides a 180° phase shift in midband, so that

$$1 - AB > 1 \qquad \text{in this frequency range}$$

The quantity $1 - AB$ is called the *feedback factor*, and if the circuit is cut at any X point in Fig. 11.1.6, the open-loop gain is AB.

It can be shown that for large loop gain AB the closed-loop transfer function reduces to

$$e_2/e_1 \approx 1/B$$

The gain then becomes essentially independent of variations in A. In particular, if B is passive, the closed-loop gain is controlled only by passive components. Feedback has no beneficial effect in reducing unwanted signals at the input of the amplifier, e.g., input noise, but does reduce unwanted signals generated in the amplifier chain (e.g., output distortion).

The return ratio can be found if the circuit is opened at any point X (Fig. 11.1.6) and a unit signal P is injected at that X point. The return signal P' is equal to the return ratio, since the input P is unity. In this case the return ratio T is the same at any point X and is

$$T = -AB$$

The minus sign is chosen because the typical amplifier has an odd number of phase reversals and T is then a positive quantity. The return difference is by definition

$$F = 1 + T$$

It has been shown by Bode that

$$F = \Delta / \Delta^0$$

where Δ is the network determinant with XX point connected and Δ^0 is the network determinant of amplifier when gain of active device is set to zero.

Stability

The stability of the network can be analyzed by several techniques. Of prime interest are the Nyquist, Bode, Routh, and root-locus techniques of analyzing stability.

Nyquist Method. The basic technique of Nyquist involves plotting T on a polar plot as shown in Fig. 11.1.7 for all values $s = j\omega$ for ω between minus and plus infinity. Stability is then determined by the following method:

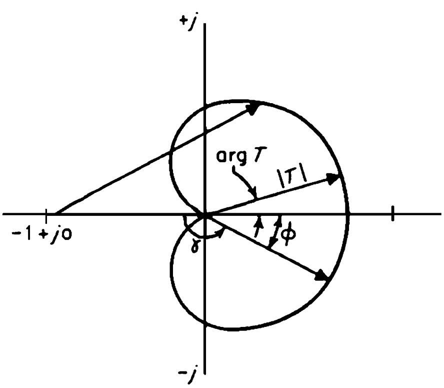


FIGURE 11.1.7 Nyquist diagram for determining stability.

1. Draw a vector from the $-1 + j0$ point to the plotted curve and observe the rotation of this vector as ω varies from $-\infty$ to $+\infty$. Let R be the net number of counterclockwise revolutions of this vector.
2. Determine the number of roots of the denominator of $T = -AB$ which have positive real parts. Call this number P.
3. The system is stable if and only if $P = R$. Note that in many systems A and B are stable by themselves, so that P becomes zero and the net counterclockwise revolution N becomes zero for stability.

Bode's Technique. A technique that has historically found wide use in determining stability and performance, especially in control systems, is the Bode diagram. The assumptions used here for this method are that $T = -AB$, where A and B are stable when the system is open-circuited and consists of minimum-phase networks. It is also necessary to define a phase margin γ such that $\gamma = 180 + \phi$, where ϕ is the phase angle of T and is positive when measured counterclockwise from zero, and γ, the phase margin, is positive when measured counterclockwise from the 180° line (Fig. 11.1.7). The stability criterion under these conditions reads: Systems having a positive phase margin when their return ratio equal to 20 log $|T|$ goes through 0 dB (i.e., where $|T|$ crosses the unit circle in the Nyquist plot) are stable; if a negative γ exists at 0 dB, the system is unstable.

Bode's theorems show that the phase angle of a system is related to the attenuation or gain characteristic as a function of frequency. Bode's technique relies heavily on straight-line approximation.

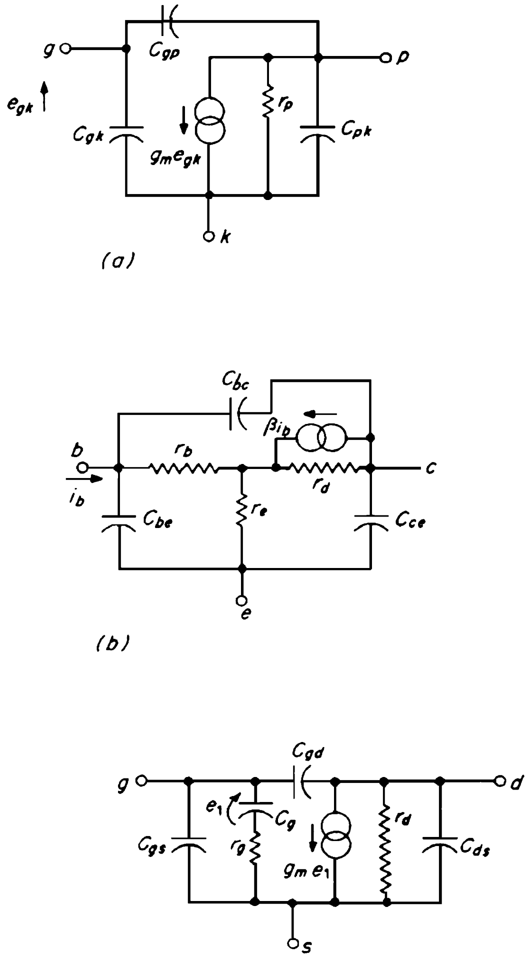


FIGURE 11.1.8 Equivalent circuits of active devices: (*a*) vacuum tube; (*b*) bipolar transistor; (*c*) field-effect transistor (FET).

Routh's Criterion for Stability. Routh's method has also been used to test the characteristic equations or return difference $F = 1 + T = 0$, to determine whether it has any roots that are real and positive or complex with positive real parts that will give rise to growing exponential responses and hence instability.

Root-Locus Method. The root-locus method of analysis is a means of finding the variations of the poles of a closed-loop response as some network parameter is varied. The most convenient and commonly used parameter is that of the gain K. The basic equation then used is

$$F = 1 + KT(s) = 1 - K\frac{(S - S_2)(S - S_4)\cdots}{(S - S_1)(S - S_3)\cdots} = 0$$

This is a useful technique in feedback and control systems, but it has not found wide application in amplifier design. A detailed exposition of the technique is found in Truxal.

AMPLIFIER AND OSCILLATOR PRINCIPLES OF OPERATION

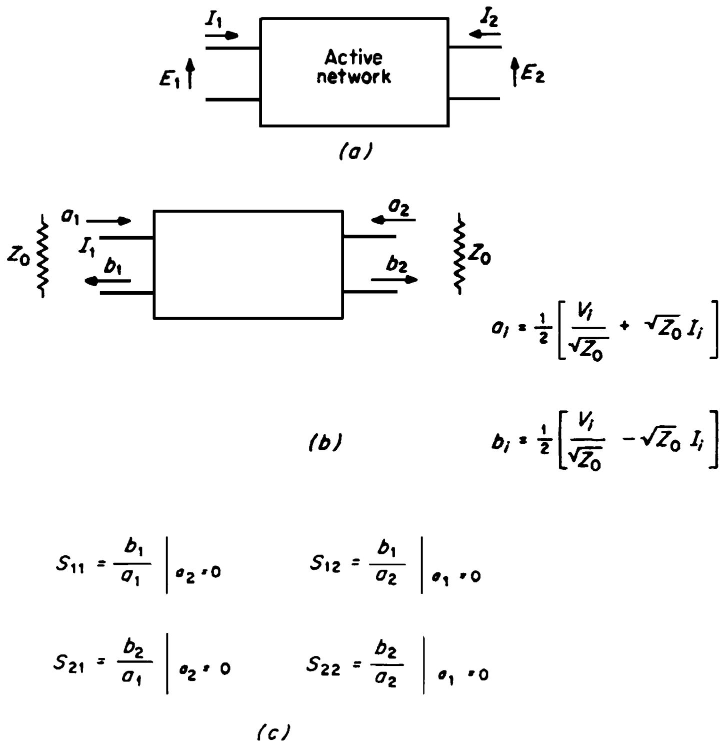


FIGURE 11.1.9 Definitions of active-network parameters: (*a*) general network; (*b*) ratios a_i and b_i of incident and reflected waves (square root of power); (*c*) *s* parameters.

Active Devices Used in Amplifiers

There are numerous ways of representing active devices and their properties. Several common equivalent circuits are shown in Fig. 11.1.8. Active devices are best analyzed in terms of the *immittance* or *hybrid matrices.* Figures 11.1.9 and 11.1.10 show the definition of the commonly used matrices, and their interconnections are shown in Fig. 11.1.11. The requirements at the bottom of Fig. 11.1.11 must be met before the interconnection of two matrices is allowed.

The matrix that is becoming increasingly important at higher frequencies is the *S* matrix. Here the network is embedded in a transmission-line structure, and the incident and reflected powers are measured and reflected coefficients and transmission coefficients are defined.

Cascaded and Distributed Amplifiers

Most amplifiers are cascaded (i.e., connected to a second amplifier). The two techniques commonly used are shown in Fig. 11.1.12. In the cascade structure the overall response is the product of the individual responses: in the distributed structure the response is one-half the sum of the individual responses, since each stage's output is propagated in both directions. In cascaded amplifiers the frequency response and gain are determined by the active device as well as the interstage networks. In simple audio amplifiers these interstage networks may become simple *RC* combinations, while in rf amplifiers they may become critically coupled double-tuned circuits. Interstage coupling networks are discussed in subsequent sections.

$$\begin{bmatrix} E_1 \\ E_2 \end{bmatrix} = \begin{bmatrix} z_{11} & z_{12} \\ z_{21} & z_{22} \end{bmatrix} \times \begin{bmatrix} I_1 \\ I_2 \end{bmatrix} \qquad z_{11} = \left(\frac{E_1}{I_1}\right)_{I_2=0} \quad z_{12} = \left(\frac{E_1}{I_2}\right)_{I_1=0} \quad z_{21} = \left(\frac{E_2}{I_1}\right)_{I_2=0} \quad z_{22} = \left(\frac{E_2}{I_2}\right)_{I_1=0}$$

$$\begin{bmatrix} I_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} \times \begin{bmatrix} E_1 \\ E_2 \end{bmatrix} \qquad y_{11} = \left(\frac{I_1}{E_1}\right)_{E_2=0} \quad y_{12} = \left(\frac{I_1}{E_2}\right)_{E_1=0} \quad y_{21} = \left(\frac{I_2}{E_1}\right)_{E_2=0} \quad y_{22} = \left(\frac{I_2}{E_2}\right)_{E_1=0}$$

$$\begin{bmatrix} E_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \times \begin{bmatrix} I_1 \\ E_2 \end{bmatrix} \qquad h_{11} = \frac{1}{y_{11}} \quad h_{12} = \left(\frac{E_1}{E_2}\right)_{I_1=0} \quad h_{21} = \left(\frac{I_2}{I_1}\right)_{E_2=0} \quad h_{22} = \frac{1}{z_{22}}$$

$$\begin{bmatrix} I_1 \\ E_2 \end{bmatrix} = \begin{bmatrix} l_{11} & l_{12} \\ l_{21} & l_{22} \end{bmatrix} \times \begin{bmatrix} E_1 \\ I_2 \end{bmatrix} \qquad l_{11} = \frac{1}{z_{11}} \quad l_{12} = \left(\frac{I_1}{I_2}\right)_{E_1=0} \quad l_{21} = \left(\frac{E_2}{E_1}\right)_{I_2=0} \quad l_{22} = \frac{1}{y_{22}}$$

$$\begin{bmatrix} E_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix} \times \begin{bmatrix} E_2 \\ -I_2 \end{bmatrix} \qquad a_{11} = \frac{1}{l_{21}} \quad a_{12} = -\frac{1}{y_{21}} \quad a_{21} = \frac{1}{z_{21}} \quad a_{22} = -\frac{1}{h_{21}}$$

$$\begin{bmatrix} E_2 \\ I_2 \end{bmatrix} = \begin{bmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{bmatrix} \times \begin{bmatrix} E_1 \\ -I_1 \end{bmatrix} \qquad b_{11} = \frac{1}{h_{12}} \quad b_{12} = -\frac{1}{y_{12}} \quad b_{21} = \frac{1}{z_{12}} \quad b_{22} = -\frac{1}{l_{12}}$$

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \times \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \qquad S_{11} = \left(\frac{b_1}{a_1}\right)_{a_2=0} \quad S_{12} = \left(\frac{b_1}{a_2}\right)_{a_1=0} \quad S_{21} = \left(\frac{b_2}{a_1}\right)_{a_2=0} \quad S_{22} = \left(\frac{b_2}{a_2}\right)_{a_1=0}$$

FIGURE 11.1.10 Network matrix terms.

In distributed structures (Fig. 11.1.12*b*), actual transmission lines are used for the input to the amplifier, while the output is taken at one end of the upper transmission line. The propagation time along the input line must be the same as that along the output line, or distortion will result. This type of amplifier, noted for its wide frequency response, is discussed later.

OSCILLATORS: PRINCIPLES OF OPERATION

Introduction

An oscillator can be considered as a circuit that converts a dc input to a time-varying output. This discussion deals with oscillators whose output is sinusoidal, as opposed to the relaxation oscillator whose output exhibits abrupt transitions (see Section 14). Oscillators often have a circuit element that can be varied to produce different frequencies.

An oscillator's frequency is sensitive to the stability of the frequency-determining elements as well as the variation in the active-device parameters (e.g., effects of temperature, bias point, and aging). In many instances

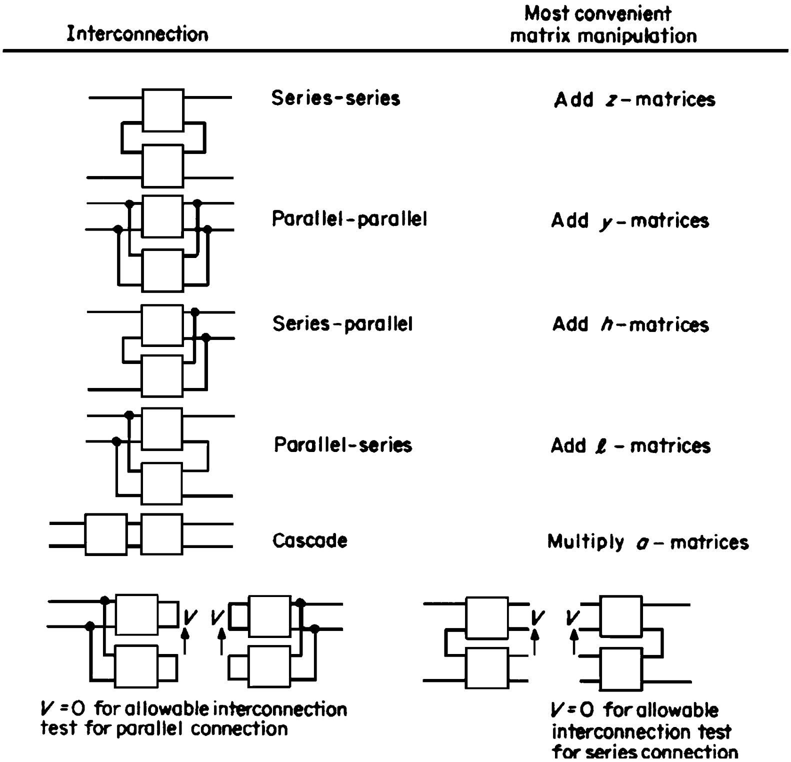


FIGURE 11.1.11 Matrix equivalents of network interconnections.

the oscillator is followed by a second stage serving as a buffer, so that there is isolation between the oscillator and its load. The amplitude of the oscillation can be controlled by automatic gain control (AGC) circuits, but the nonlinearity of the active element usually determines the amplitude. Variations in bias, temperature, and component aging have a direct effect on amplitude stability.

Requirements for Oscillation

Oscillators can be considered from two viewpoints: as using positive feedback around an amplifier or as a one-port network in which the real component of the input immittance is negative. An oscillator must have frequency-determining elements (generally passive components), an amplitude-limiting mechanism, and sufficient closed-loop gain to make up for the losses in the circuit. It is possible to predict the operating frequency and conditions needed to produce oscillation from a Nyquist or Bode analysis. The prediction of output amplitude requires the use of nonlinear analysis.

Oscillator Circuits

Typical oscillator circuits applicable up to ultra high frequencies (UHF) are shown in Fig. 11.1.13. These are discussed in detail in the following subsections. Also of interest are crystal oscillators. In this case the crystal is used as the passive frequency-determining element. The frequency range of crystal oscillators extends from a few hundred hertz to over 200 MHz by use of overtone crystals. The analysis of crystal oscillators is best done using the equivalent circuit of the crystal.

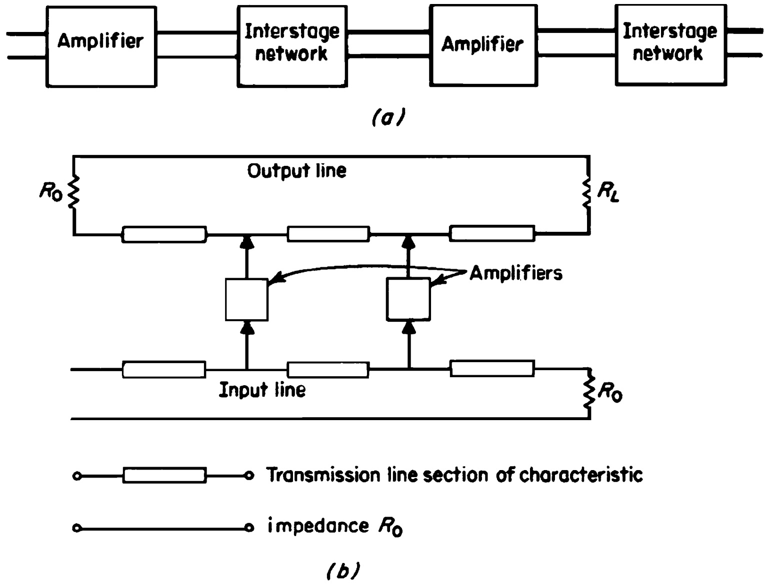


FIGURE 11.1.12 Multiamplifier structures: (*a*) cascade; (*b*) distributed.

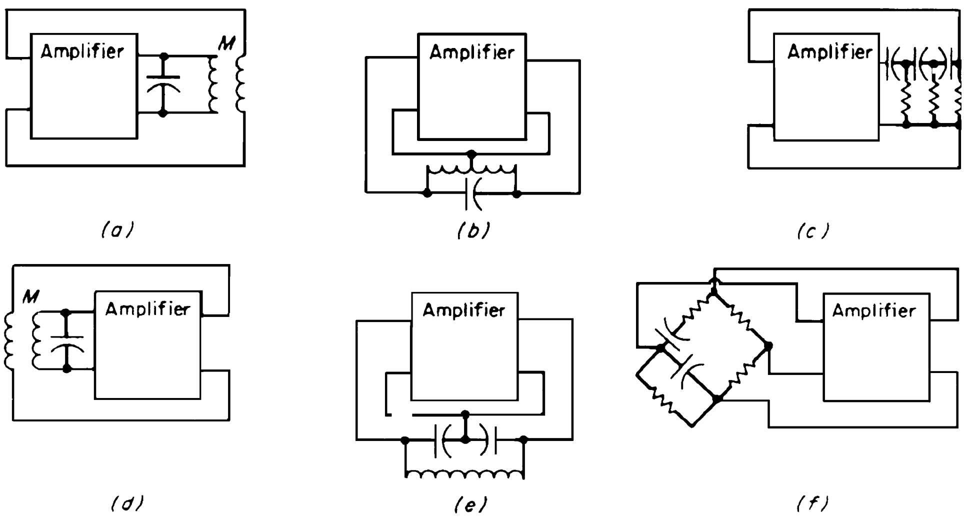


FIGURE 11.1.13 Types of oscillators: (*a*) tuned-output; (*b*) Hartley; (*c*) phase-shift; (*d*) tuned-input; (*e*) Colpitts; (*f*) Wien bridge.

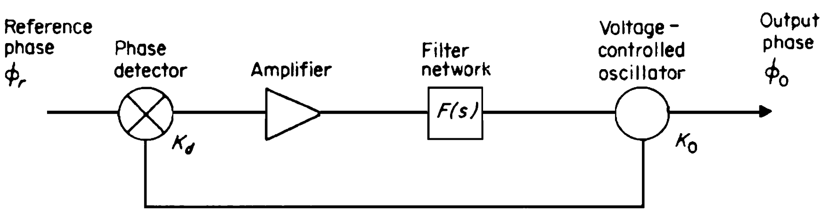


FIGURE 11.1.14 Phase-locked-loop oscillator.

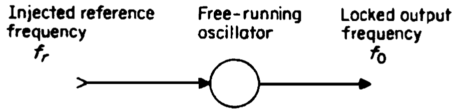


FIGURE 11.1.15 Injection-locked oscillator.

Synchronization

Synchronization of oscillators is accomplished by using phase-locked loops or by direct low-level injection of a reference frequency into the main oscillator. The diagram of a phase-locked loop is shown in Fig. 11.1.14 and that of an injection-locked oscillator in Fig. 11.1.15.

Harmonic Content

The harmonic content of the oscillator output is related to the amount of oscillator output power at frequencies other than the fundamental. From the viewpoint of a negative-conductance (resistance) oscillator, better results are obtained if the curve of the negative conductance (or resistance) versus amplitude of oscillation is smooth and without an inflection point over the operating range. Harmonic content is also reduced if the oscillator's operating point Q is chosen so that the range of negative conductance is symmetrical about Q on the negative conductance-versus-amplitude curve. This can be done by adjusting the oscillator's bias point within the requirement of $|G_C| = |G_D|$ for sustained oscillation (see Fig. 11.1.16).

Stability

The stability of the oscillator's output amplitude and frequency from a negative-conductance viewpoint depends on the variation of its negative conductance with operating point and the amount of fixed positive conductance in the oscillator's associated circuit. In particular, if the change of bias results in vertical translation of the conductance-(resistance)-versus-amplitude curve, the oscillator's stability is related to the change of slope at the point where the circuit's fixed conductance intersects this curve (point Q in Fig. 11.1.16). If the $|G_D|$ curve is of the shape of $|G_D|_2$, the oscillation can stop when a large enough change in bias point occurs for $|G_D|$ to be less than $|G_C|$ for all amplitudes of oscillation. Stabilization of the amplitude of oscillation may occur in the form of modifying G_C, G_D, or both to compensate for bias changes.

Particular types of oscillators and their parameters are discussed later in this section.

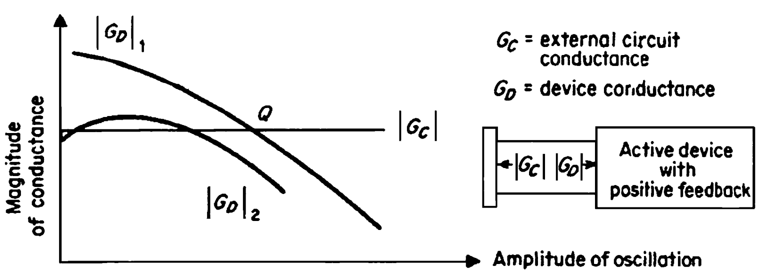


FIGURE 11.1.16 Device conductance vs. amplitude of oscillation.