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CHAPTER 11.2
AUDIO-FREQUENCY AMPLIFIERS AND OSCILLATORS

Samuel M. Korzekwa, Robert J. McFadyen

AUDIO-FREQUENCY AMPLIFIERS

Samuel M. Korzekwa

Preamplifiers

General Considerations. The function of a preamplifier is to amplify a low-level signal to a higher level before further processing or transmission to another location. The required amplification is achieved by increased signal voltage and/or impedance reduction. The amount of power amplification required varies with the particular application. A general guideline is to provide sufficient preamplification to ensure that further signal handling adds minimal (or acceptable) signal-to-noise degradation.

Signal-to-Noise Considerations. The design of a preamplifier must consider all potential signal degradation from sources of noise, whether generated externally or within the preamplifier itself.

Examples of externally generated noise are hum and pickup, which may be introduced by the input-signal lines or the power-supply lines. Shielding of the input-signal lines often proves to be an acceptable solution. The preamplifier should be located close to the transmitting source, and the preamplifier power gain must be sufficient to override interference that remains after these steps are taken.

A second major source of noise is that internally generated in the amplifier itself. The noise figure specified in decibels for a preamplifier, which serves as a figure of merit, is defined as the ratio of the available input-to-output signal-to-noise power ratios:

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

where F = noise figure of preamplifier
S_i = available signal input power
N_i = available noise input power
S_o = available signal output power
N_o = available noise output power

Design precautions to realize the lowest possible noise figure include the proper selection of the active device, optimum input and output impedance, correct voltage and current biasing conditions, and pertinent design parameters of devices.

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Low-Level Amplifiers

The low-level designation applies to amplifiers operated below maximum permissible power-dissipation, current, and voltage limits. Thus many low-level amplifiers are purposely designed to realize specific attributes other than delivering the maximum attainable power to the load, such as gain stability, bandwidth, optimum noise figure, and low cost.

In an amplifier designed to be operated with a 24-V power supply and a specified load termination, for example, the operating conditions may be such that the active devices are just within their allowable limits. If operated at these maximum limits, this is not a low-level amplifier; however, if this amplifier also fulfills its performance requirements at a reduced power-supply voltage of 6 V, with resulting much lower internal dissipation levels, it becomes a low-level amplifier.

Medium-Level and Power Amplifiers

The medium-power designation for an amplifier implies that some active devices are operated near their maximum dissipation limits, and precautions must be taken to protect these devices. If power-handling capability is taken as the criterion, the 5- to 100-W power range is a current demarcation line. As higher-power-handling devices come into use, this range will tend to shift to higher power levels.

The amount of power that can safely be handled by an amplifier is usually dictated by the dissipation limits of the active devices in the output stages, the efficiency of the circuit, and the means used to extract heat to maintain devices within their maximum permissible temperature limits. The classes of operation (A, B, AB, C) are discussed relative to Fig. 11.1.5. When single active devices do not suffice, multiple series or parallel configurations can be used to achieve higher voltage or power operation.

Multistage Amplifiers

An amplifier may take the form of a single stage or a complex single stage, or it may employ an interconnection of several steps. Various biasing, coupling, feedback, and other design alternatives influence the topology of the amplifier. For a multistage amplifier, the individual stages may be essentially identical or radically different. Feedback techniques may be used at the individual stage level, at the amplifier functional level, or both, to realize bias stabilization, gain stabilization, output-impedance reduction, and so forth.

Typical Electron-Tube Amplifier

Figure 11.2.1 shows a typical electron-tube amplifier stage. For clarity the signal-source and load sections are shown partitioned. For a multistage amplifier the source represents the equivalent signal generator of the preceding stage. Similarly, the load indicated includes the loading effect of the subsequent stage, if any.

The voltage gain from the grid of the tube to the output can be calculated to be

$$A_{v1} = -\frac{\mu R_1}{r_p + R_l}$$

Similarly, the voltage gain from the source to the tube grid is

$$A_{v2} = \frac{R_1}{(R_1 + R_g) + 1/j\omega C}$$

Combining the above equations gives the composite amplifier voltage gain

$$A_v = \frac{\mu R_1 R_l}{(r_p + R_l)[(R_1 + R_g) + 1/j\omega C]}$$

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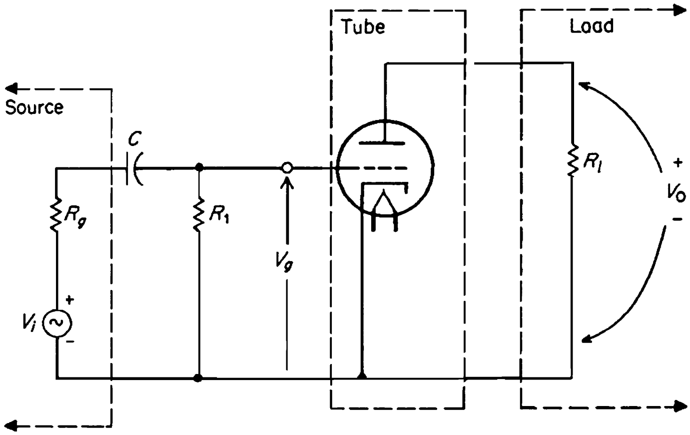

FIGURE 11.2.1 Typical triode electron-tube amplifier stage (biasing not shown).

This example illustrates the fundamentals of an electron-tube amplifier stage. Many excellent references treat this subject in detail.

Typical Transistor Amplifier

The analysis techniques used for electron-tube amplifier stages generally apply to transistorized amplifier stages. The principal difference is that different active-device models are used.

The typical transistor stage shown in Fig. 11.2.2 illustrates a possible form of biasing and coupling. The source section is partitioned and includes the preceding-stage equivalent generator, and the load includes subsequent stage-loading effects. Figure 11.2.3 shows the generalized *h*-equivalent circuit representation for transistors. Table 11.2.1 lists the *h*-parameter transformations for the common-base, common-emitter, and common-collector configurations.

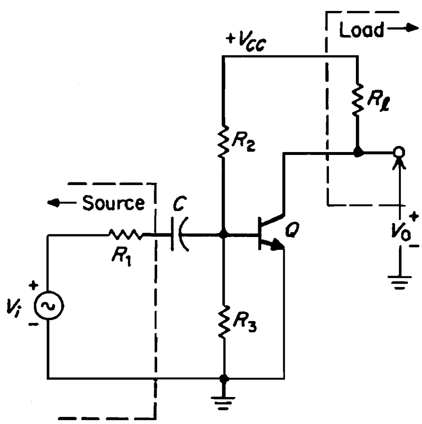

FIGURE 11.2.2 Typical bipolar transistor-amplifier stage.

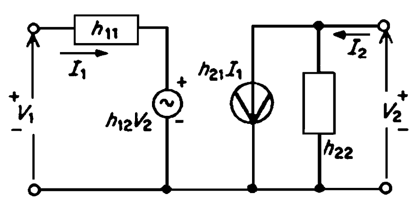

FIGURE 11.2.3 Equivalent circuit of transistor, based on *h* parameters.

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TABLE 11.2.1 *h* Parameters of the Three Transistor Circuit Configurations

	Common-base	Common-emitter	Common-collector
h_{11}	h_{ib}	$h_{ib}(h_{fe}+1)$	$h_{ib}(h_{fe}+1)$
h_{12}	h_{rb}	$h_{ib}h_{ob}(h_{fe}+1)-h_{rb}$	1
h_{21}	h_{fb}	h_{fe}	$-(h_{fe}+1)$
h_{22}	h_{ob}	$h_{ob}(h_{fe}+1)$	$h_{ob}(h_{fe}+1)$

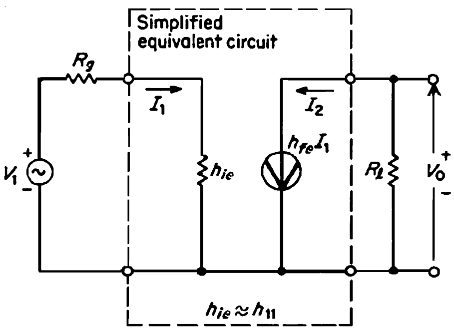

FIGURE 11.2.4 Simplified equivalent circuit of transistor amplifier stage.

While these parameters are complex and frequency-dependent, it is often feasible to use simplifications. Most transistors have their parameters specified by their manufacturers, but it may be necessary to determine additional parameters by test.

Figure 11.2.4 illustrates a simplified model of the transistor amplifier stage of Fig. 11.2.2. The common-emitter *h* parameters are used to represent the equivalent transistor. The voltage gain for this stage is

$$A_v = \frac{V_o}{V_i} = -\frac{h_{fe}R_l}{R_g + h_{ie}}$$

The complexity of analysis depends on the accuracy needed. Currently, most of the more complex analysis is performed with the aid of computers. Several transistor-amplifier-analysis references treat this subject in detail.

Typical Multistage Transistor Amplifier

Figure 11.2.5 is an example of a capacitively coupled three-stage transistor amplifier. It has a broad frequency response, illustrating the fact that an audio amplifier can be useful in other applications. The component values are

$$R_1 = 16{,}000\ \Omega \quad R_2 = 6200\ \Omega \quad R_3 = 1600\ \Omega \quad R_4 = 1000\ \Omega$$
$$R_L = 560\ \Omega \quad Q_1, Q_2, Q_3 = \text{2N1565} \quad C_1 = 10\ \mu F \quad C_2 = 100\ \mu F$$

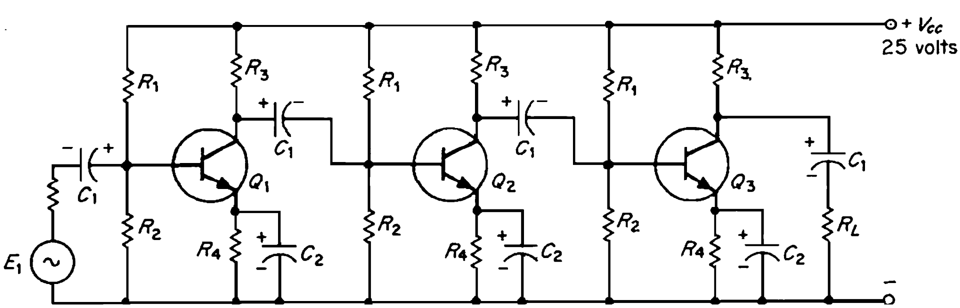

FIGURE 11.2.5 Typical three-stage transistor amplifier.

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This amplifier is designed to operate over a range of −55 to +125°C, with an output voltage swing of 2 V peak to peak and frequency response down 3 dB at approximately 200 Hz and 2 MHz. The overall gain at 1000 Hz is nominally 88 dB at 25°C.

Biasing Methods

The biasing scheme used in an amplifier determines the ultimate performance that can be realized. Conversely, an amplifier with poorly implemented biasing may suffer in performance, and be susceptible to catastrophic circuit failure owing to high stresses within the active devices. In view of the variation of parameters within the active devices, it is important that the amplifier function properly even when the initial and/or end-of-life parameters of the devices vary.

Electron-Tube Biasing

Biasing is intended to maintain the quiescent currents and voltages of the electron tube at the prescribed levels. The tube-plate characteristics represent the biasing relations between the tube parameters.

The principal bias parameters (steady-state plate and grid voltages) can be readily identified by the construction of a load line on the plate characteristic. The operating point *Q* is located at the intersection of the selected plate characteristic with the load line.

Transistor Biasing

Although the methods of biasing a transistor-amplifier stage are in many respects similar to those of an electron-tube amplifier, there are many different types of transistors, each characterized by different curves. Bipolar transistors are generally characterized by their collector and emitter families, while field-effect transistors have different characterizations. The *npn* transistor requires a positive base bias voltage and current (with respect to its emitter) for proper operation; the converse is true for a *pnp* transistor.

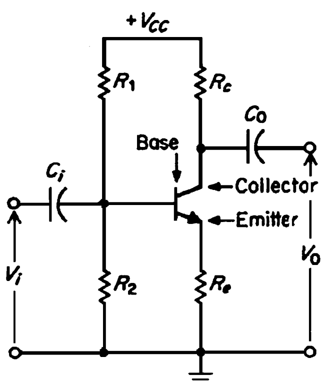

FIGURE 11.2.6 Capacitively coupled *npn* transistor-amplifier stage.

Figure 11.2.6 illustrates a common biasing technique. A single power supply is used, and the transistor is self-biased with the unbypassed emitter resistor R_e. Although a graphical solution of the value of R_e could be found by referring to the collector-emitter curves, an iterative solution, described below, is also commonly used.

Because the performance of the transistors depends on the collector current and collector-to-emitter voltage, they are often selected as starting conditions for biasing design. The unbypassed emitter resistor R_e and collector resistor R_c, the primary voltage-gain-determining components, are determined next, taking into account other considerations such as the anticipated maximum signal level and available power supply V_{cc}. The last step is to determine the R_1 and R_2 values.

Coupling Methods

Transformer coupling and capacitance coupling are commonly used in transistor and electron-tube audio amplifiers. Direct coupling is also used in transistor stages and particularly in integrated transistor amplifiers. Capacitance coupling, referred to as *RC* coupling, is the most common method of coupling stages of an audio amplifier. The discrete-component transistorized amplifier stage shown in Fig. 11.2.6 serves as an example of *RC* coupling, where C_i and C_o are the input and output coupling capacitors, respectively.

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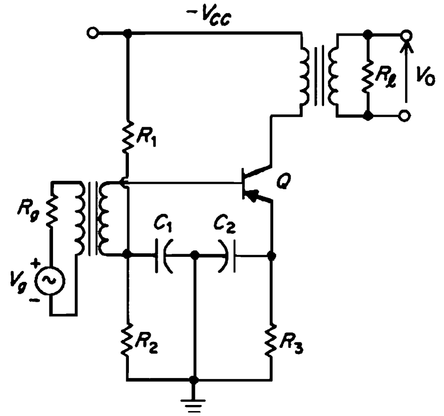

FIGURE 11.2.7 Transformer-coupled *pnp* transistor-amplifier stage.

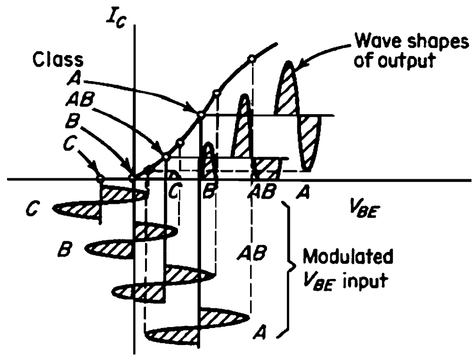

FIGURE 11.2.8 Classes of amplifier operation, based on transistor characteristics.

Transformer coupling is commonly used to match the input and output impedances of electron-tube amplifier stages. Since the input impedance of an electron tube is very high at audio frequencies, the design of an electron-tube stage depends primarily on the transformer parameters. The much lower input impedances of transistors demand that many other factors be taken into account, and the design becomes more complex. The output-stage transformer coupling to a specific load is often the optimum method of realizing the best power match. Figure 11.2.7 illustrates a typical transformer-coupled transistor audio-amplifier stage.

The direct-coupling approach is now also used for discrete-component transistorized amplifiers, and particularly in integrated amplifier versions. The level-shifting requirement is realized by selection from the host of available components, such as *npn* and *pnp* transistors and zener diodes. Since it is difficult to realize large-size capacitors via integrated-circuit techniques, special methods have been developed to direct-couple integrated amplifiers.

Classes A, B, AB, and C Operation

The output or power stage of an amplifier is usually classified as operating class A, B, AB, or C, depending on the conduction characteristics of the active devices (see Fig. 11.1.5). These definitions can also apply to any intermediate amplifier stage. Figure 11.2.8 illustrates relations between the class of operation and conduction using transistor parameters. This figure would be essentially the same for an electron-tube amplifier with the tube plate current and grid voltage as the equivalent device parameters.

Subscripts may be used to denote additional conduction characteristics of the device. For example, the electron-tube grid conduction can also be further classified as A_1, to show that no grid current flows, or A_2, to show that grid-current conduction exists during some portion of the cycle.

Push-Pull Amplifiers

In a single-ended amplifier the active devices conduct continuously. The single-ended configuration is generally used in low-power applications, operated in class A. For example, preamplifiers and low-level amplifiers are generally operated single-ended, unless the output power levels necessitate the more efficient power handling of the push-pull circuit.

In a push-pull configuration there are at least two active devices that alternately amplify the negative and positive cycles of the input waveform. The output connection to the load is most often transformer-coupled. An example of a transformer input and output in a push-pull amplifier is illustrated in Fig. 11.2.9. Direct-coupled push-pull amplifiers and capacitively coupled push-pull amplifiers are also feasible, as illustrated in Fig. 11.2.10.

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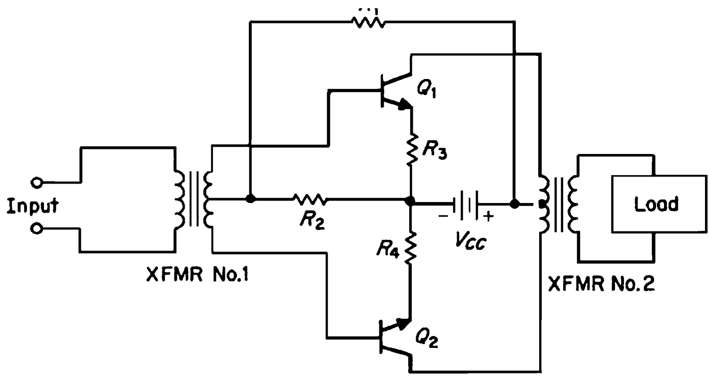

FIGURE 11.2.9 Transformer-coupled push-pull transistor stage.

The active devices in push-pull are usually operated either in class B or AB because of the high power-conversion efficiency. Feedback techniques can be used to stabilize gain, stabilize biasing or operating points, minimize distortion, and the like.

Output Amplifiers

The function of an audio output amplifier is to interface with the preceding amplifier stages and to provide the necessary drive to the load. Thus the output-amplifier designation does not uniquely identify a particular amplifier class. When several different types of amplifiers are cascaded between the signal source and its load, e.g., a high-power speaker, the last-stage amplifier is designated as the output amplifier. Because of the high power requirements, this amplifier is usually a push-pull type operating either in class B or AB.

Stereo Amplifiers

A stereo amplifier provides two separate audio channels properly phased with respect to each other. The objective of this two-channel technique is to enhance the audio reproduction process, making it more realistic

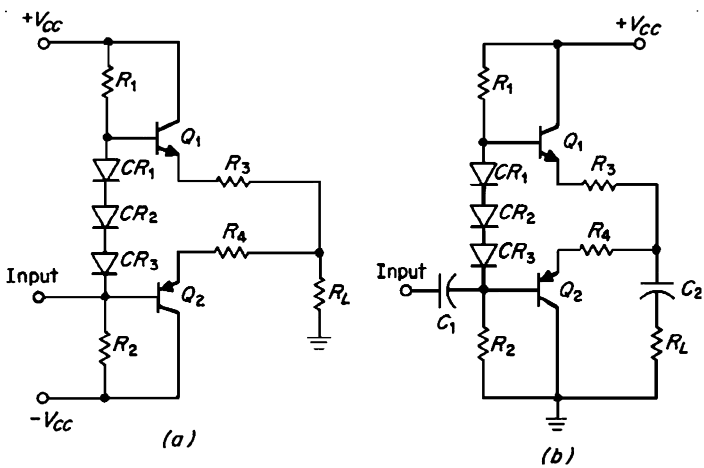

FIGURE 11.2.10 (*a*) Direct- and (*b*) capacitively coupled push-pull stages.

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and lifelike. It is also feasible to extend the system to contain more than two channels of information. A stereo amplifier is a complete system that contains its power supply and other commonly required control functions.

Each channel has its own preamplifier, medium-level stages, and output power stage, with different gain and frequency responses for each mode of operation, e.g., for tape, phonograph, CD, and so forth. The input signal is selected from the phonograph input connection, tape input, or a turner output. Special-purpose trims and controls are also used to optimize performance on each mode. The bandwidth of the amplifier extends to 20 kHz or higher.

AUDIO OSCILLATORS

Robert J. McFadyen

General Considerations

In the strict sense, an audio oscillator is limited to frequencies from about 15 to 20,000 Hz, but a much wider frequency range is included in most oscillators used in audio measurements since knowledge of amplifier characteristics in the region above audibility is often required.

For the production of sinusoidal waves, audio oscillators consist of an amplifier having a nonlinear power gain characteristic, with a path for regenerative feedback. Single- and multistage transistor amplifiers with *LC* or *RC* feedback networks are most often used. The term *harmonic oscillator* is used for these types. *Relaxation oscillators*, which may be designed to oscillate in the audio range, exhibit sharp transitions in the output voltages and currents. Relaxation oscillators are treated in Section 14.

The instantaneous excursions of the operating point in a harmonic oscillator is restricted to the range where the circuit exhibits an impedance with a negative real part. The amplifier supplies the power, which is dissipated in the feedback path and the load. The regenerative feedback would cause the amplitude of oscillation to grow without bound were it not for the fact that the dynamic range of the amplifier is limited by circuit nonlinearities. Thus, in most sine-wave audio oscillators; the operating frequency is determined by passive-feedback elements, whereas the amplitude is controlled by the active-circuit design.

Analytical expressions predicting the frequency and required starting conditions for oscillation can be derived using Bode's amplifier feedback theory, and the stability theorem of Nyquist. Since this analytical approach is based on a linear-circuit model, the results are approximate but usually suitable for design of sinusoidal oscillators. No prediction on waveform amplitude results, since this is determined by nonlinear-circuit characteristics. Estimates of the waveform amplitude can be made from the bias and limiting levels of the active circuits. Separate limiters and AGC techniques are also useful for controlling the amplitude to a prescribed level. Graphical and nonlinear analysis methods can also be used for obtaining a prediction of the amplitude of oscillation.

A general formulation suitable for a linear analysis of almost all audio oscillators can be derived from the feedback diagram in Fig. 11.2.11. Note that the amplifier internal feedback generator has been neglected: that is, y_{12A} is assumed to be zero. This assumption of unilateral amplification is almost always valid in the audio range even for single-stage transistor amplifiers.

The stability requirements for the circuit are derived from the closed-loop-gain expression

$$A_c = A/(1 - A\beta) \tag{1}$$

where the gain A is treated as a negative quantity for an inverting amplifier. Infinite closed-loop gain occurs when AB is equal to unity, and this defines the oscillatory condition. In terms of the equivalent circuit parameters used in Fig. 11.2.1,

$$1 - A\beta = 1 - y_{21A}\frac{y_{12\beta}}{(y_{11A} + y_{11\beta})(y_{22A} + y_{22\beta}) - y_{12\beta}y_{21\beta}} \tag{2}$$

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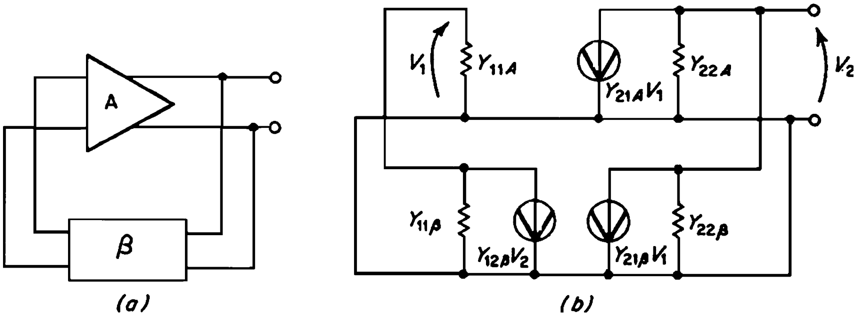

FIGURE 11.2.11 Oscillator representations: (*a*) generalized feedback circuit; (*b*) equivalent *y*-parameter circuit.

In the audio range, y_{21A} remains real, but the fractional portion of the function is complex because β is frequency-sensitive. Therefore, the open-loop gain $A\beta$ can be expressed in the general form

$$A\beta = y_{21A}\frac{A_r + jA_i}{B_r + jB_i} \tag{3}$$

It follows from Nyquist's stability theorem that this feedback system will be unstable if, first, the phase shift of $A\beta$ is zero and, second, the magnitude is equal to or greater than unity. Applying this criterion to Eq. (3) yields the following two conditions for oscillation:

$$A_iB_r - A_rB_i = 0 \tag{4}$$

$$y_{21}^2 \geq \frac{B_r^2 + B_i^2}{A_r^2 + A_i^2} \tag{5}$$

Equation (4) results from the phase condition and determines the frequency of oscillation. The inequality in Eq. (5) is the consequence of the magnitude constraint and defines the necessary condition for sustained oscillation. Equation (5) is evaluated at the oscillation frequency determined from Eq. (4).

A large number of single-stage oscillators have been developed in both vacuum-tube and transistor versions. The transistor circuits followed by direct analogy from the earlier vacuum-tube circuits. In the following examples, transistor versions are illustrated, but the *y*-parameter equations apply to other devices as well.

LC Oscillators

The *Hartley oscillator* circuit is one of the oldest forms: the transistor version is shown in Fig. 11.2.12. With the collector and base at opposite ends of the tuned circuit, the 180° phase relation is secured, and feedback occurs through mutual inductance between the two parts of the coil. The frequency and condition for oscillation are expressed in terms of the transistor *y* parameters and feedback inductance *L*, inductor coupling coefficient *k*, inductance ratio *n*, and tuning capacitance *C*. The frequency of oscillation is

$$\omega^2 = \frac{1}{LC(1 + 2k\sqrt{n} + n) + nL^2(1 - k^2)(y_{11A}y_{22A})}$$

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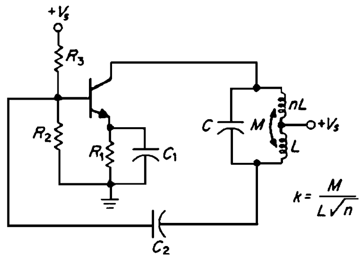

FIGURE 11.2.12 Hartley oscillator circuit.

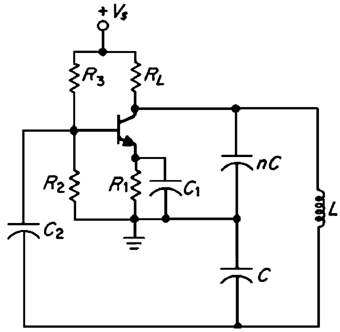

FIGURE 11.2.13 Colpitts oscillator circuit.

The condition for oscillation is

$$y_{21A} \geq \frac{y_{11A} + ny_{22A} + n\omega^2 LC(1-k^2)(y_{11A}y_{22A})}{k\sqrt{n} + n\omega^2 LC(1-k^2)}$$

The admittance parameters of the bias network R_1, R_2, and R_3, as well as the reactance of bypass capacitor C and coupling capacitor C_2, have been neglected. These admittances could be included in the amplifier y parameters in cases where their effect is not negligible.

If

$$\frac{C}{L} >> \frac{n(1-k^2)(y_{11A}y_{22A})}{1 + 2k\sqrt{n} + n} \tag{6}$$

the frequency of oscillation will be essentially independent of transistor parameters.

The transistor version of the *Colpitts oscillator* is shown in Fig. 11.2.13. Capacitors C and nC in combination with inductance L determine the resonant frequency of the circuit. A fraction of the current flowing in the tank circuit is regeneratively fed back to the base through the coupling capacitor C_2. Bias resistors R_1, R_2, R_3, and R_L, as well as capacitors C_1 and C_2, are chosen so as not to affect the frequency or conditions for oscillation. The frequency of oscillation is

$$\omega^2 = \frac{1}{LC}\left(1 + \frac{1}{n}\right) + \frac{1}{nC^2}(y_{11A}y_{22A})$$

The condition for oscillation is

$$y_{21A} \geq \omega^2 LC(nY_{11A} + Y_{22A}) - (y_{11A} + y_{22A})$$

Alternatively, the bias element admittances may be included in the amplifier y parameters.

In the Colpitts circuit, if the ratio of C/L is chosen so that

$$\frac{C}{L} >> \frac{y_{11A}y_{22A}}{1+n} \tag{7}$$

the frequency of oscillation is essentially determined by the tuned-circuit parameters.

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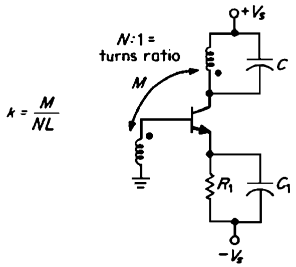

FIGURE 11.2.14 Tuned-collector oscillator.

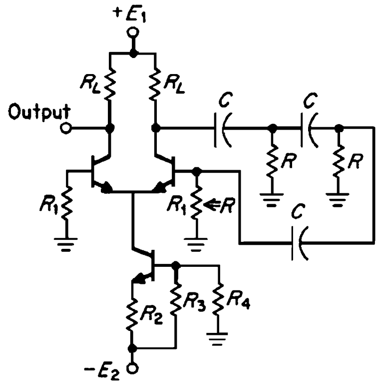

FIGURE 11.2.15 *RC* oscillator with high-pass feedback network.

Another oscillator configuration useful in the audio-frequency range is the tuned-collector circuit shown in Fig. 11.2.14. Here regenerative feedback is furnished via the transformer turns ratio N from the collector to base. The frequency of oscillation is

$$\omega^2 = \frac{1}{LC + N^2L^2 y_{11A} y_{22A}(1-k^2)}$$

The condition for oscillation is

$$y_{22A} \geq \frac{1}{Nk}(N^2 Y_{11A} + Y_{22A}) - \frac{\omega^2 NLCY_{11A}}{k}(1-k^2)$$

If the ratio of C/L is such that

$$\frac{C}{L} >> N^2 y_{11A} y_{22A}(1-k^2) \tag{8}$$

the frequency of oscillation is specified by $\omega^2 = 1/LC$. This circuit can be tuned over a wide range by varying the capacitor C and is compatible with simple biasing techniques.

RC Oscillators

Audio sinusoidal oscillators can be designed using an *RC* ladder network (of three or more sections) as a feedback path in an amplifier. This scheme originally appeared in vacuum-tube circuits, but the principles have been directly extended to transistor design. *RC* phase-shift oscillators can be distinguished from tuned oscillators in that the feedback network has a relatively broad frequency-response characteristic.

Typically, the phase-shift network has three *RC* sections of either a high- or a low-pass nature. Oscillation occurs at the frequency where the total phase shift is 180° when used with an inverting amplifier. Figures 11.2.15 and 11.2.16 show examples of high-pass and low-pass feedback-connection schemes. The amplifier is a differential pair with a transistor current source, a configuration which is common in integrated-circuit amplifiers. The output is obtained at the opposite collector from the feedback connection, since this minimizes external loading on the phase-shift network. The conditions for, and the frequency of, oscillation are derived, assuming that the input resistance of the amplifier, which loads the phase-shift network, has been adjusted to equal the

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resistance R. The load resistor R_L is considered to be part of the amplifier output resistance, and it is included in y_{22A}. The frequency of oscillation for the high-pass case is

$$\omega^2 = \frac{y_{22A}}{2C^2R(2+3Ry_{22A})}$$

The condition for oscillation for the high-pass case is

$$y_{21A} \geq \frac{1}{R}\left(\frac{1+5R/R_L}{\omega^2R^2C^2} - \frac{R}{R_L} - 3\right)$$

The frequency of oscillation for the low-pass case is

$$\omega = \frac{1}{RC}\sqrt{6+4\frac{R}{R_L}}$$

The condition for oscillation for the low-pass case is

$$y_{21A} \geq \frac{1}{R}\left(23\frac{R}{R_L} + 29 + 4\frac{R^2}{R_L^2}\right)$$

Null-Network Oscillators

In almost all respects null-network oscillators are superior to the RC phase-shift circuits described in the previous paragraph. While many null-network configurations are useful (including the bridged-T and twin-T), the Wien bridge design predominates.

The general form for the Wien bridge oscillator is shown in Fig. 11.2.17. In the figure, an ideal differential voltage amplifier is assumed, i.e., one with infinite input impedance and zero output impedance.

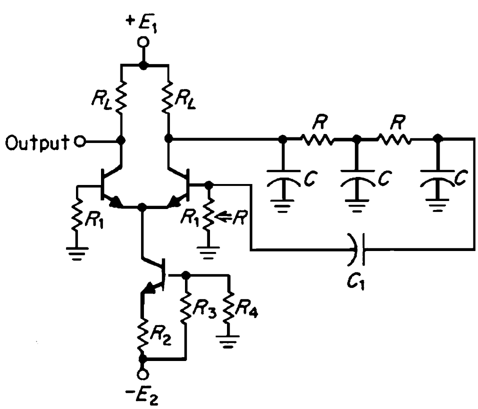
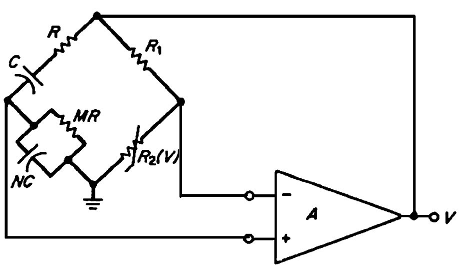


FIGURE 11.2.16 *RC* oscillator with low-pass feedback network.



Frequency of oscillation ($M = N = 1$):

$$\omega_0 = \frac{1}{RC}$$

Condition for oscillation:

$$A \geq 8 = \frac{3(R_1+R_2)}{R_1-2R_2}$$

FIGURE 11.2.17 Wien bridge oscillator circuit.

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An integrated-circuit operational amplifier that has a differential input stage is a practical approximation to this type of amplifier and is often used in bridge-oscillator designs.

The Wien bridge is used as the feedback network, with positive feedback provided through the *RC* branches for regeneration and negative feedback through the resistor divider. Usually the resistor-divider network includes an amplitude-sensitive device in one or both arms which provides automatic correction for variation of the amplifier gain. Circuit elements such as a tungsten lamp, thermistor, and field-effect transistor used as the voltage-sensitive resistance element maintain a constant output level with a high degree of stability. Amplitude variations of less than ±1 percent over the band from 10 to 100,000 Hz are realizable. In addition, since the amplifier is never driven into the nonlinear region, harmonic distortion in the output waveform is minimized. For the connection shown in Fig. 11.2.17, an increase in V will cause a decrease in R_2, restoring V to the original level.

The lamp or thermistor have thermal time constants that set at a lower frequency limit on this method of amplitude control. When the period is comparable with the thermal time constant, the change in resistance over an individual cycle distorts the output waveform. There is an additional degree of freedom with the field-effect transistor, since the control voltage must be derived by a separate detector from the amplifier output. The time constant of the detector, and hence the resistor, are set by a capacitor, which can be chosen commensurate with the lowest oscillation frequency desired.

At ω_0 the positive feedback predominates, but at harmonics of ω_0 the net negative feedback reduces the distortion components. Typically, the output waveform exhibits less than 1 percent total harmonic distortion. Distortion components well below 0.1 percent in the mid-audio-frequency range are also achieved.

Unlike *LC* oscillators, in which the frequency is inversely proportional to the square root of L and C, in the Wien bridge ω_0 varies as $1/RC$. Thus, a tuning range in excess of 10:1 is easily achieved. Continuous tuning within one decade is usually accomplished by varying both capacitors in the reactive feedback branch. Decade changes are normally accomplished by switching both resistors in the resistive arm. Component tracking problems are eased when the resistors and capacitors are chosen to be equal.

Almost any three-terminal null network can be used for the reactive branch in the bridge; the resistor divider network adjusts the degree of imbalance in the manner described. Many of these networks lack the simplicity of the Wien bridge since they may require the tracking of three components for frequency tuning. For this reason networks such as the bridged-T and twin-T are usually restricted to fixed-tuned applications.

Low-Frequency Crystal Oscillators

Quartz-crystal resonators are used where frequency stability is a primary concern. The frequency variations with both time and temperature are several orders of magnitude lower than obtainable in *LC* or *RC* oscillator circuits. The very high stiffness and elasticity of piezoelectric quartz make it possible to produce resonators extending from approximately 1 kHz to 200 MHz. The performance characteristics of crystal depend on both the particular cut and the mode of vibration (see Section 5). For convenience, each “cut-mode” combination is considered as a separate piezoelectric element, and the more commonly used elements have been designated with letter symbols. The audio-frequency range (above 1 kHz) is covered by elements J, H, N, and XY, as shown in Table 11.2.2.

The temperature coefficients vary with frequency, i.e., with the crystal dimensions, and except for the H element, a parabolic frequency variation with temperature is observed. The H element is characterized by a

TABLE 11.2.2 Low-Frequency Crystal Elements

Symbol	Cut	Mode of vibration	Frequency range, kHz
J	Duplex 5°X	Length-thickness flexure	0.9–10
H	5°X	Length-width flexure	10–50
N	NT	Length-width flexure	4–200
XY	XY	XY flexure	8–40

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negative temperature coefficient on the order of –10 ppm/°C. The other elements have lower temperature coefficients, which at some temperatures are zero because of the parabolic nature of the frequency-deviation curve. The point where the zero temperature coefficient occurs is adjustable and varies with frequency. At temperatures below this point the coefficient is positive, and at higher temperatures it is negative. On the slope of the curves the temperature coefficients for the N and XY elements are on the order of 2 ppm/°C, whereas the J element is about double at 4 ppm/°C.

Although the various elements differ in both cut and mode of vibration, the electric equivalent circuit remains invariant. The schematic representation and the lumped constant equivalent circuit are shown in Fig. 11.2.18. As is characteristic of most mechanical resonators, the motional inductance L resulting from the mechanical mass in motion is large relative to that obtainable from coils. The extreme stiffness of quartz makes for very small values of the motional capacitance C, and the very high order of elasticity allows the motional resistance R to be relatively low. The shunt capacitance C_0 is the electrostatic capacitance existing between crystal electrodes with the quartz plate as the dielectric and is present whether or not the crystal is in mechanical motion. Some typical values for these equivalent-circuit parameters are shown in Table 11.2.3.

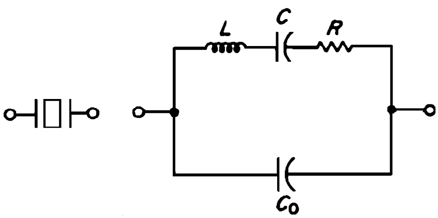

FIGURE 11.2.18 Symbol and equivalent circuit of a quartz crystal.

The H element can have a high Q value when mounted in a vacuum enclosure; however, it then has the poorest temperature coefficient. The N element exhibits an excellent temperature characteristic, but the piezoelectric activity is rather low, so that special care is required when it is used in oscillator circuits. The J and XY elements operate well in low-frequency oscillator designs, the latter having lower temperature drift. For the same frequency the XY crystal is about 40 percent longer than the J element. Where extreme frequency stability is required, the crystals are usually controlled to a constant temperature.

The reactance curve of a quartz resonator is shown in Fig. 11.2.19. The zero occurs at the frequency f_s, which corresponds to series resonance of the mechanical L and C equivalences. The antiresonant frequency f_p is dependent on the interelectrode capacitance C_0. Between f_s and f_p the crystal is inductive and this frequency range is normally referred to as the *crystal bandwidth*

$$\mathrm{BW} = f_s/(2C_0/C) \tag{9}$$

In oscillator circuits the crystal can be used as either a series or a parallel resonator. At series resonance the crystal impedance is purely resistive, but in the parallel mode the crystal is operated between f_s and f_p and is therefore inductive. For oscillator applications the circuit capacitance shunting the crystal must also be included when specifying the crystal, since it is part of the resonant circuit. If a capacitor C_L, that is, a negative reactance, is placed in series with a crystal, the combination will series-resonate at the frequency f_R of zero reactance for the combination.

$$f_R = f_s\left[1 + \frac{1}{(2C_0/C)(1 + C_L/C_0)}\right] \tag{10}$$

TABLE 11.2.3 Typical Crystal Parameter Values

Element	Frequency, kHz	L, H	C, pF	R, kΩ	C_0, pF	Q, approx
J	10	8,000	0.03	50	6	20,000
H	10	2,500	0.1	10	75	20,000
N	10	8,000	0.03	75	30	10,000
XY	10	12,000	0.02	30	20	30,000

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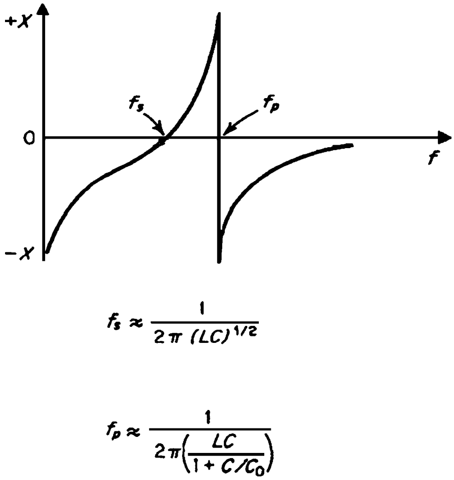

FIGURE 11.2.19 Quartz-crystal reactance curve.

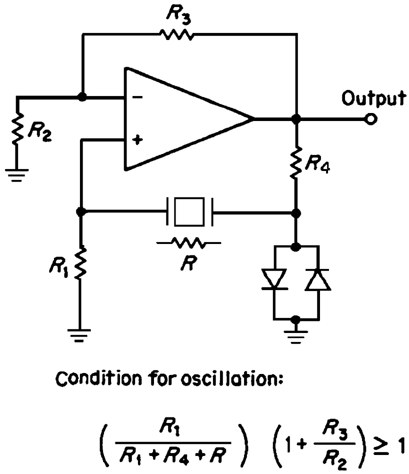

FIGURE 11.2.20 Crystal oscillator using an integrated-circuit operational amplifier.

The operating frequency can vary in value due to changes in the load capacitance, and this variation is prescribed by

$$\Delta f_R = \frac{f_s}{2C_0/C} \frac{\Delta C_L/C_0}{(1+C_L/C_0)^2} \tag{11}$$

This effect can be used to "pull" the crystal for initial alignment, or if the external capacitor is a voltage-controllable device, a VCO with a range of about ±0.01 percent can be constructed. Phase changes in the amplifier will also give rise to frequency shifts since the total phase around the loop must remain at 0° to maintain oscillation.

Although single-stage transistor designs are possible, more flexibility is available in the circuit of Fig. 11.2.20, which uses an integrated-circuit operational amplifier for the gain element. The crystal is operated in the series mode, and the amplifier gain is precisely controlled by the negative-feedback divider R_2 and R_3. The output will be sinusoidal if

$$\frac{V_D R_1}{R_1 + R}\left(1+\frac{R_3}{R_2}\right) < V_{\text{lim}} \tag{12}$$

where V_D is the limiting diode forward voltage drop and V_{lim} is the limiting level of amplifier output.

Low-cost electronic wristwatches use quartz crystals for establishing a high degree of timekeeping accuracy. A high-quality mechanical watch may have a yearly accuracy on the order of 20 min, whereas many quartz watches are guaranteed to vary less than 1 min/year.

Generally the XY crystal is used, but other types are continually being developed to improve accuracy, reduce size, and lower manufacturing cost. The active gain elements for the oscillator are part of the integrated circuit that contains the electronics for the watch functions. The flexure or tuning-fork frequency is set generally to 32,768 Hz, which is 2^{15} Hz. This frequency reference is divided down on the integrated circuit to provide seconds, minutes, hours, day of the week, date, month, and so forth.

A logic gate or inverter is often used as the gain element in the oscillator circuit. A typical configuration is shown in Fig. 11.2.21. The resistor R_1 is used to bias the logic inverter for class A amplifier operation. The resistor R_2 helps reduce both voltage sensitivity of the network and crystal power dissipation. The combination of R_2 and C_2 provides added phase shift for good oscillator startup. The series combination of capacitors

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C_1 and C_2 provides the parallel load for the crystal. C_1 can be made tunable for precise setting of the crystal oscillation frequency. The inverter provides the necessary gain and 180° phase shift. The π network consisting of the capacitors and the crystal provides the additional 180° phase shift needed to satisfy the conditions for oscillation.

Frequency Stability

Many factors contribute to the ability of an oscillator to hold a constant output frequency over a period of time and range from short-term effects, caused by random noise, to longer-term variations, caused by circuit parameter dependence on temperature, bias voltage, and the like. In addition to the temperature and aging effects of the frequency-determining elements, nonlinearities, impedance loading, and amplifier phase variations also contribute to instability.

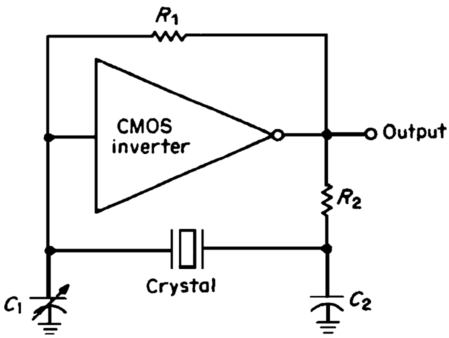

FIGURE 11.2.21 Crystal oscillator using a logic gate for the gain element.

Harmonics generated by circuit nonlinearities are passed through the feedback network, with various phase shifts, to the input of the amplifier. Intermodulation of the harmonic frequencies produces a fundamental frequency component that differs in phase from the amplifier output. Since the condition $A\beta = 1$ must be satisfied, the frequency of oscillation will shift so that the network phase shift cancels the phase perturbation caused by the nonlinearity. Therefore, the frequency of oscillation is influenced by an unpredictable amplifier characteristic, namely, the saturation nonlinearity. This effect is negligible in the Wien bridge oscillator, where automatic level control keeps harmonic distortion to a minimum.

The relationships shown in Fig. 11.2.17 were derived assuming that the amplifier does not load the bridge circuit on either the input or output sides. In the practical sense this is never true, and changes in the input and output impedances will load the bridge and cause frequency variations to occur.

Another source of frequency instability is small phase changes in the amplifier. The effect is minimized by using a network with a large stability factor, defined by

$$S = \left. \frac{d\phi}{d\omega/\omega_0} \right|_{\omega=\omega_0} \tag{13}$$

For the Wien bridge oscillator, which has amplitude-sensitive resistive feedback, the RC impedances can be optimized to provide a maximum stability factor value. As shown in Fig. 11.2.17, this amounts to choosing proper values for M and N. The maximum stability-factor value is $A/4$, and it occurs for $N = 1/2$ and $M = 2$. Most often the bridge is used with equal resistor and capacitor values; that is, $M = N = 1$, in which case the stability factor is $2A/9$. This represents only a slight degradation from the optimum.

Synchronization

It is often desirable to lock the oscillator frequency to an input reference. Usually this is done by injecting sufficient energy at the reference frequency into the oscillator circuit. When the oscillator is tuned sufficiently close to the reference, natural oscillations cease and the synchronization signal is amplified to the output. Thus the circuit appears to oscillate at the injected signal frequency. The injected reference is amplitude-stabilized by the AGC or limiting circuit in the same manner as the natural oscillation. The frequency range over which locking can occur is a linear function of the amplitude of the injected signal. Thus, as the synchronization frequency is moved away from the natural oscillator frequency, the amplitude threshold to maintain lock increases. The phase

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error between the input reference and the oscillator output will also deviate as the input frequency varies from the natural frequency.

Methods for injecting the lock signal vary and depend on the type of oscillator under consideration. For example, *LC* oscillators may have signals coupled directly to the tank circuit, whereas the lock signal for the Wien network is usually coupled into the center of the resistive side of the bridge, i.e., the junction of R_1 and R_2 in Fig. 11.2.17.

If the natural frequency of oscillation can be voltage controlled, synchronization can be accomplished with a phase-locked loop. Replacing both *R*'s with field-effect transistors, or alternatively shunting both *C*'s with varicaps, provides an effective means for voltage controlling the frequency of the Wien bridge oscillator. Although more complicated in structure, the phase-locked loop is more versatile and has many diverse applications.

Piezoelectric Annunciators

Another important class of audio oscillators uses piezoelectric elements for both frequency control and audible-sound generation. Because of their low cost and high efficiency these devices are finding increasing use in smoke detectors, burglar alarms, and other warming devices. Annunciators using these elements typically produce a sound level in excess of 85 dB measured at a distance of 10 ft.

Usually the element consists of a thin brass disk to which a piezoelectric material has been attached. When an electric signal is applied across its surfaces, the piezoceramic disk attempts to change diameter. The brass disk to which it is bonded acts as a spring restraining force on one surface of the ceramic. The brass plate serves as one electrode for applying the electric signal to the ceramic. On the other surface a fired-on silver paste is used as an electrode. The restraining action of the brass disk causes the assembly to change from a flat to a convex shape. When the polarity of the electric signal reverses, the assembly flexes in the other direction to a concave shape. When the device is properly mounted in a suitable horn structure, this motion is used to produce high-level sound waves. One useful method is to clamp the disk at nodal points, i.e., at a distance from the center of the disk where mechanical motion is at a vibrational null.

The piezoelectric assembly will produce sound levels more efficiently when excited near the series-resonant frequency. The simple equivalent circuit used for the quartz crystal (Fig. 11.2.18) also applies to the piezoceramic assembly for frequencies near resonance. Generally the piezoelectric element is used as the frequency-determining element in an audio oscillator. The advantage of this method is that the excitation frequency is inherently near the optimum value, since it is self-excited. A typical piezoceramic 1-in diameter mounted on a $1^3/_4$-in brass disk would have the following equivalent values: $C_0 = 0.02\ \mu$F, $C = 0.0015\ \mu$F, $L = 2$ H, $R = 500\ \Omega$, $Q = 75$, $f_s = 2.9$ kHz, and $f_p = 3.0$ kHz.

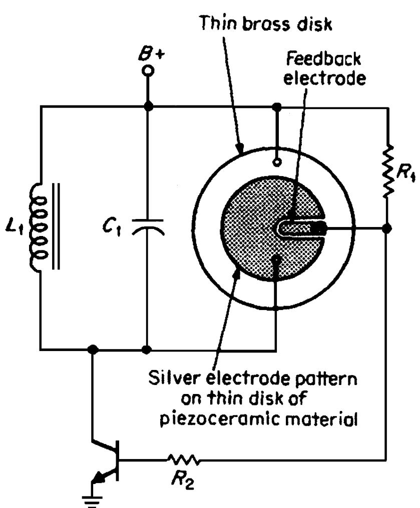

FIGURE 11.2.22 Basic audio annunciator oscillator circuit using a thin-disk piezoelectric transducer.

A basic oscillator, capable of producing high-level sound, is shown in Fig. 11.2.22. The inductor L_1 provides a dc path to the transistor and broadly tunes the parallel input capacitance of the piezoelectric element. C_1 is an optional capacitor which adds to the input shunt capacitance for optimizing the drive impedance to the element. Resistor R_1 provides base-current bias to the transistor so that oscillation can start.

The element has a third small electrode etched in the silver pattern. It is used to derive a feedback signal which, when resistively loaded by R_1, provides an in-phase signal to the base for sustaining circuit oscillation. The circuit operates like a blocking oscillator in that the transistor is switched on and off and the collector voltage can fly above B-plus because of the inductor L_1.

The collector load consisting of L_1 and C_1 can be replaced with a resistor, in which case the audio output will be less.

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