Feedback and Oscillator Circuits

14

CHAPTER OBJECTIVES

- The concept of negative feedback
- About practical feedback circuits
- Various types of oscillator circuits

14.1 FEEDBACK CONCEPTS

Feedback has been mentioned previously, in particular, in op-amp circuits as described in Chapters 10 and 11. Depending on the relative polarity of the signal being fed back into a circuit, one may have negative or positive feedback. Negative feedback results in decreased voltage gain, for which a number of circuit features are improved, as summarized below. Positive feedback drives a circuit into oscillation as in various types of oscillator circuits.

A typical feedback connection is shown in Fig. 14.1. The input signal V_s is applied to a mixer network, where it is combined with a feedback signal V_f. The difference of these signals V_i is then the input voltage to the amplifier. A portion of the amplifier output V_o is connected to the feedback network (β), which provides a reduced portion of the output as feedback signal to the input mixer network.

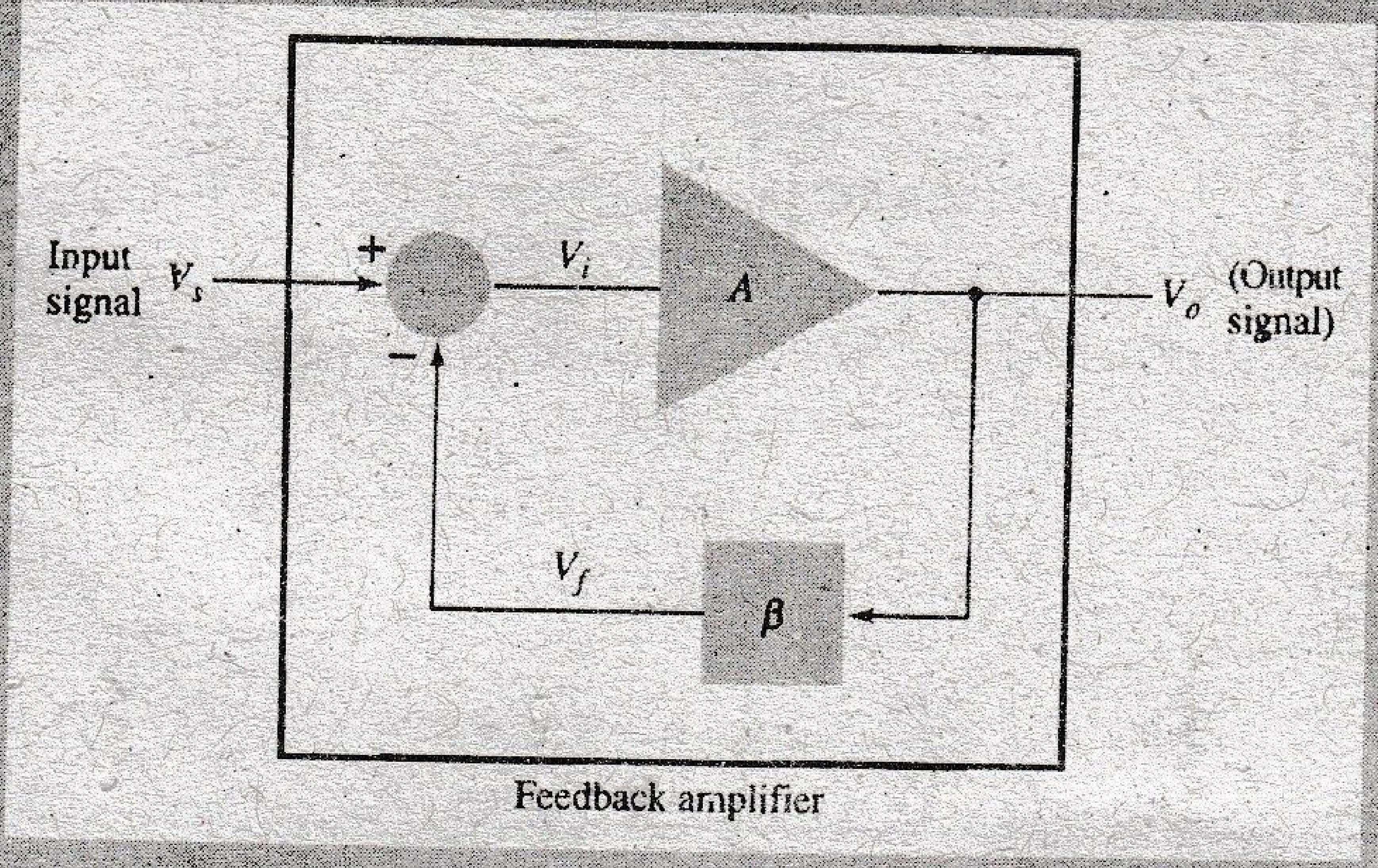


FIG. 14.1
Simple block diagram of feedback amplifier.

If the feedback signal is of opposite polarity to the input signal, as shown in Fig. 14.1, negative feedback results. Although negative feedback results in reduced overall voltage gain, a number of improvements are obtained, among them being:

1. Higher input impedance.
2. Better stabilized voltage gain.
3. Improved frequency response.
4. Lower output impedance.
5. Reduced noise.
6. More linear operation.

14.2 FEEDBACK CONNECTION TYPES

There are four basic ways of connecting the feedback signal. Both *voltage* and *current* can be fed back to the input either in *series* or *parallel*. Specifically, there can be:

1. Voltage-series feedback (Fig. 14.2a).
2. Voltage-shunt feedback (Fig. 14.2b).
3. Current-series feedback (Fig. 14.2c).
4. Current-shunt feedback (Fig. 14.2d).

In the list above, *voltage* refers to connecting the output voltage as input to the feedback network; *current* refers to tapping off some output current through the feedback network. *Series* refers to connecting the feedback signal in series with the input signal voltage; *shunt* refers to connecting the feedback signal in shunt (parallel) with an input current source.

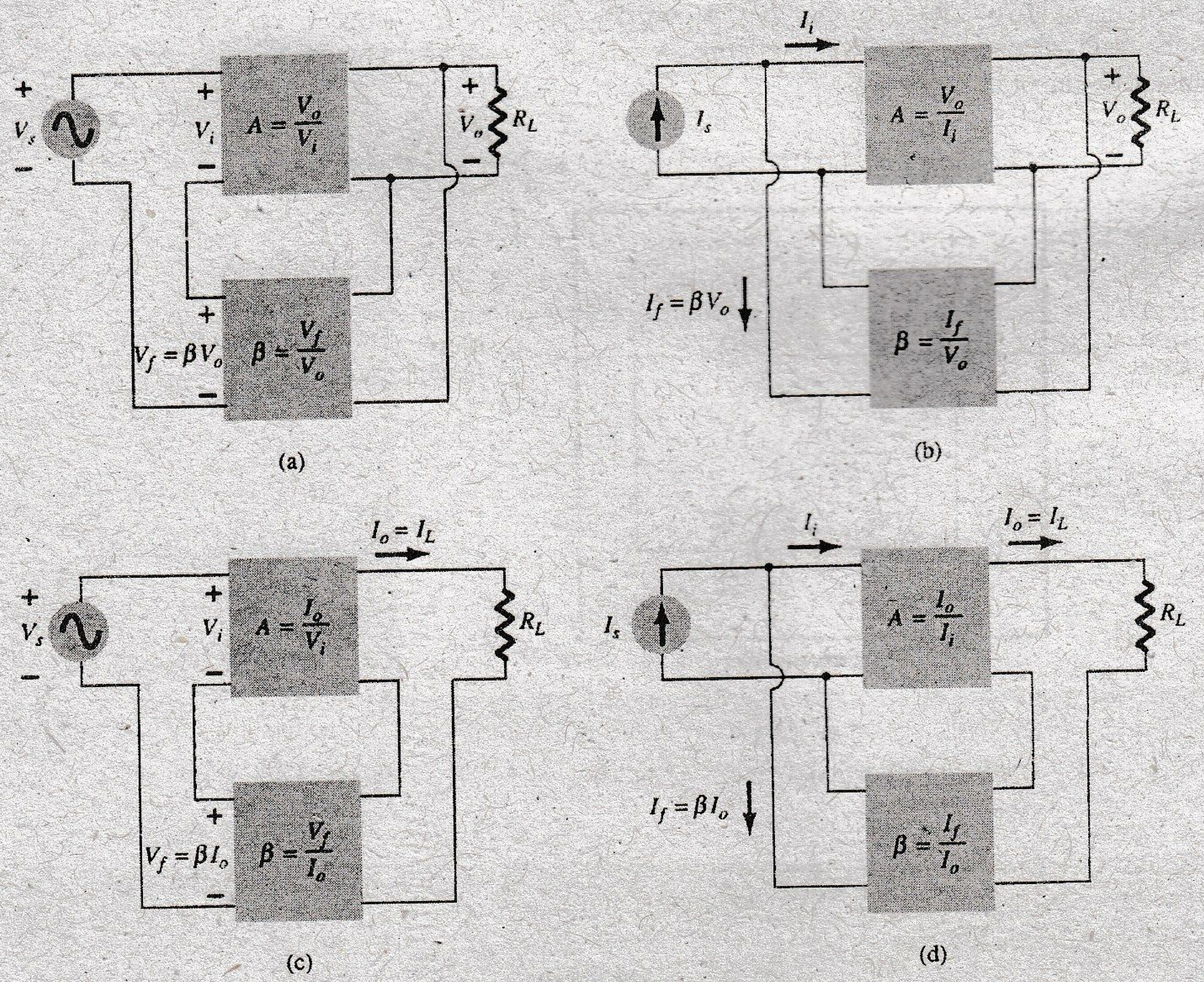


FIG. 14.2

Feedback amplifier types: (a) voltage-series feedback, $A_f = V_o/V_s$; (b) voltage-shunt feedback, $A_f = V_o/I_s$; (c) current-series feedback, $A_f = I_o/V_s$; (d) current-shunt feedback, $A_f = I_o/I_s$.

Series feedback connections tend to *increase* the input resistance, whereas shunt feedback connections tend to *decrease* the input resistance. Voltage feedback tends to *decrease* the output impedance, whereas current feedback tends to *increase* the output impedance. Typically, higher input and lower output impedances are desired for most cascade amplifiers. Both of these are provided using the voltage-series feedback connection. We shall therefore concentrate first on this amplifier connection.

Gain with Feedback

In this section we examine the gain of each of the feedback circuit connections of Fig. 14.2. The gain without feedback, A, is that of the amplifier stage. With feedback β, the overall gain of the circuit is reduced by a factor $(1 + \beta A)$, as detailed below. A summary of the gain, feedback factor, and gain with feedback of Fig. 14.2 is provided for reference in Table 14.1.

TABLE 14.1
Summary of Gain, Feedback, and Gain with Feedback from Fig. 14.2

		Voltage-Series	Voltage-Shunt	Current-Series	Current-Shunt
Gain without feedback	A	$\frac{V_o}{V_i}$	$\frac{V_o}{I_i}$	$\frac{I_o}{V_i}$	$\frac{I_o}{I_i}$
Feedback	β	$\frac{V_f}{V_o}$	$\frac{I_f}{V_o}$	$\frac{V_f}{I_o}$	$\frac{I_f}{I_o}$
Gain with feedback	A_f	$\frac{V_o}{V_s}$	$\frac{V_o}{I_s}$	$\frac{I_o}{V_s}$	$\frac{I_o}{I_s}$

Voltage-Series Feedback Figure 14.2a shows the voltage-series feedback connection with a part of the output voltage fed back in series with the input signal, resulting in an overall gain reduction. If there is no feedback ($V_f = 0$), the voltage gain of the amplifier stage is

$$A = \frac{V_o}{V_s} = \frac{V_o}{V_i} \qquad (14.1)$$

If a feedback signal V_f is connected in series with the input, then

$$V_i = V_s - V_f$$

Since $\quad V_o = AV_i = A(V_s - V_f) = AV_s - AV_f = AV_s - A(\beta V_o)$

then $\quad (1 + \beta A)V_o = AV_s$

so that the overall voltage gain *with* feedback is

$$\boxed{A_f = \frac{V_o}{V_s} = \frac{A}{1 + \beta A}} \qquad (14.2)$$

Equation (14.2) shows that the gain *with* feedback is the amplifier gain reduced by the factor $(1 + \beta A)$. This factor will be seen also to affect input and output impedance among other circuit features.

Voltage-Shunt Feedback The gain with feedback for the network of Fig. 14.2b is

$$A_f = \frac{V_o}{I_s} = \frac{A\,I_i}{I_i + I_f} = \frac{A\,I_i}{I_i + \beta V_o} = \frac{A\,I_i}{I_i + \beta A\,I_i}$$

$$\boxed{A_f = \frac{A}{1 + \beta A}} \qquad (14.3)$$

Input Impedance with Feedback

Voltage-Series Feedback A more detailed voltage-series feedback connection is shown in Fig. 14.3. The input impedance can be determined as follows:

$$I_i = \frac{V_i}{Z_i} = \frac{V_s - V_f}{Z_i} = \frac{V_s - \beta V_o}{Z_i} = \frac{V_s - \beta AV_i}{Z_i}$$

$$I_i Z_i = V_s - \beta AV_i$$

$$V_s = I_i Z_i + \beta AV_i = I_i Z_i + \beta A I_i Z_i$$

$$Z_{if} = \frac{V_s}{I_i} = Z_i + (\beta A)Z_i = Z_i(1 + \beta A) \quad (14.4)$$

The input impedance with series feedback is seen to be the value of the input impedance without feedback multiplied by the factor $(1 + \beta A)$, and applies to both voltage-series (Fig. 14.2a) and current-series (Fig. 14.2c) configurations.

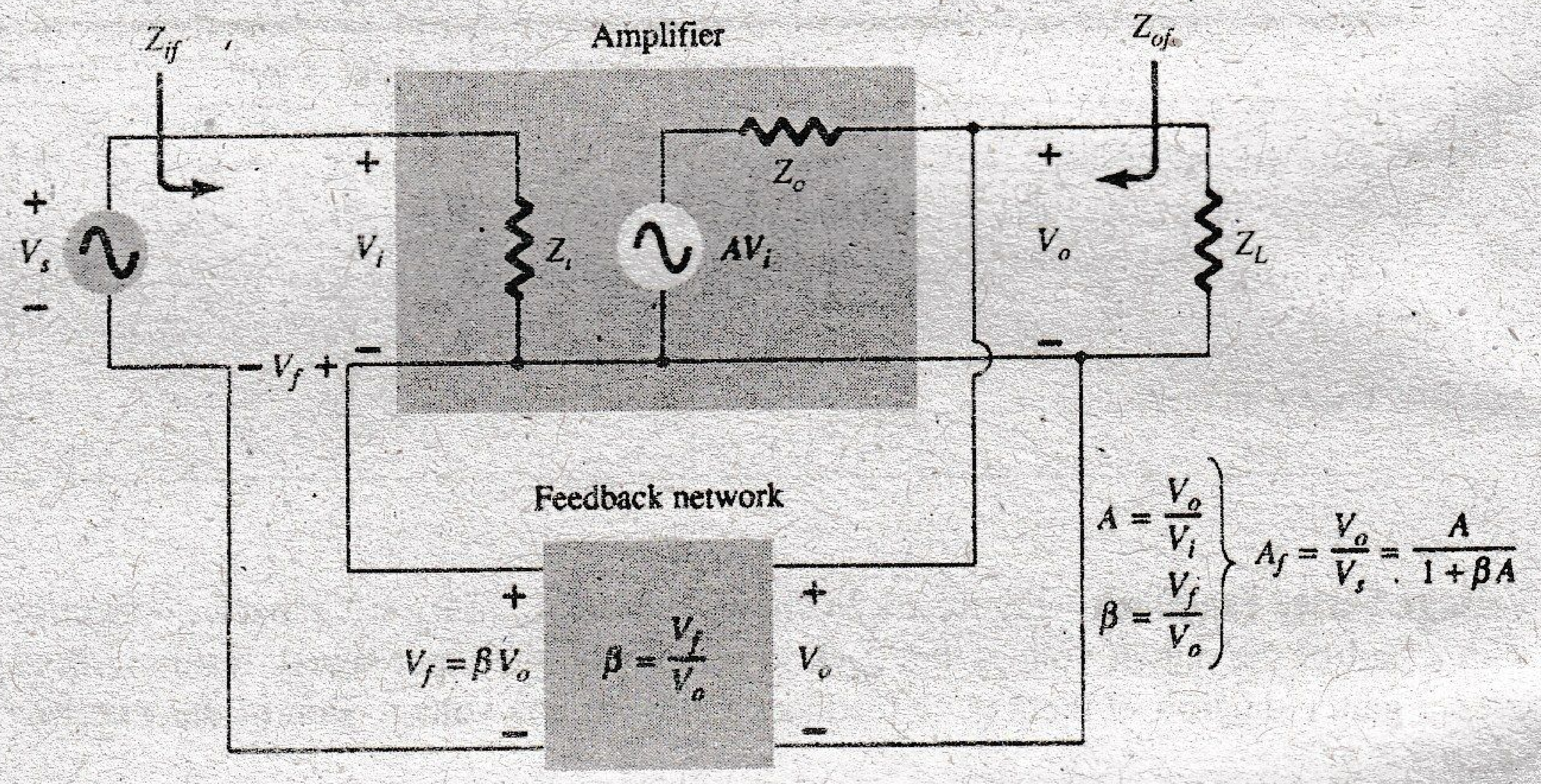


FIG. 14.3
Voltage-series feedback connection.

Voltage-Shunt Feedback A more detailed voltage-shunt feedback connection is shown in Fig. 14.4. The input impedance can be determined to be

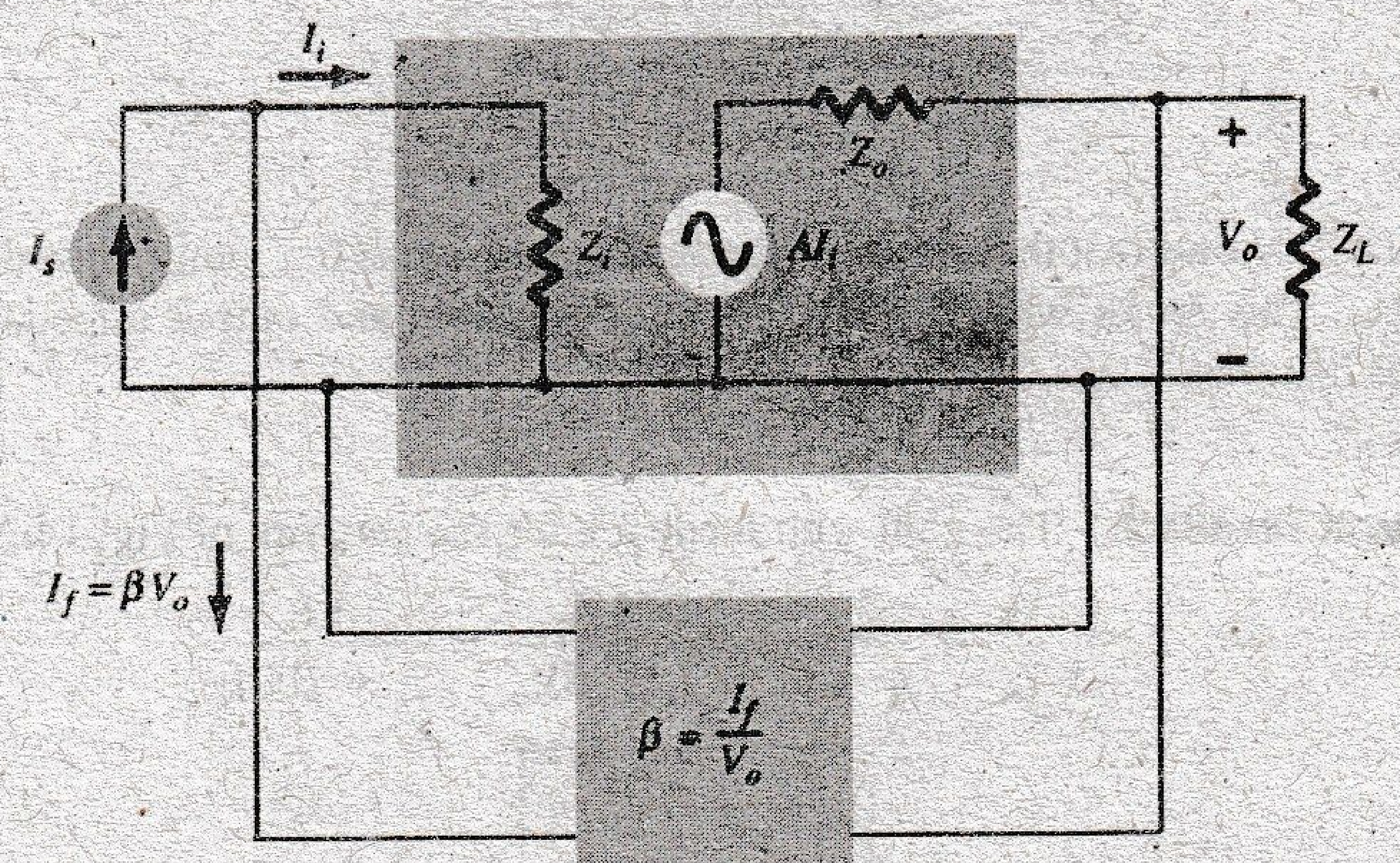


FIG. 14.4
Voltage-shunt feedback connection.

$$Z_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i + I_f} = \frac{V_i}{I_i + \beta V_o}$$

$$= \frac{V_i/I_i}{I_i/I_i + \beta V_o/I_i}$$

$$Z_{if} = \frac{Z_i}{1 + \beta A} \qquad (14.5)$$

This reduced input impedance applies to the voltage-series connection of Fig. 14.2a and the voltage-shunt connection of Fig. 14.2b.

Output Impedance with Feedback

The output impedance for the connections of Fig. 14.2 is dependent on whether voltage or current feedback is used. For voltage feedback, the output impedance is decreased, whereas current feedback increases the output impedance.

Voltage-Series Feedback The voltage-series feedback circuit of Fig. 14.3 provides sufficient circuit detail to determine the output impedance with feedback. The output impedance is determined by applying a voltage V, resulting in a current I, with V_s shorted out ($V_s = 0$). The voltage V is then

$$V = IZ_o + AV_i$$

For $V_s = 0$, $\quad V_i = -V_f$

so that $\quad V = IZ_o - AV_f = IZ_o - A(\beta V)$

Rewriting the equation as

$$V + \beta AV = IZ_o$$

allows solving for the output impedance with feedback:

$$Z_{of} = \frac{V}{I} = \frac{Z_o}{1 + \beta A} \qquad (14.6)$$

Equation (14.6) shows that with voltage-series feedback the output impedance is reduced from that without feedback by the factor $(1 + \beta A)$.

Current-Series Feedback The output impedance with current-series feedback can be determined by applying a signal V to the output with V_s shorted out, resulting in a current I, the ratio of V to I being the output impedance. Figure 14.5 shows a more detailed

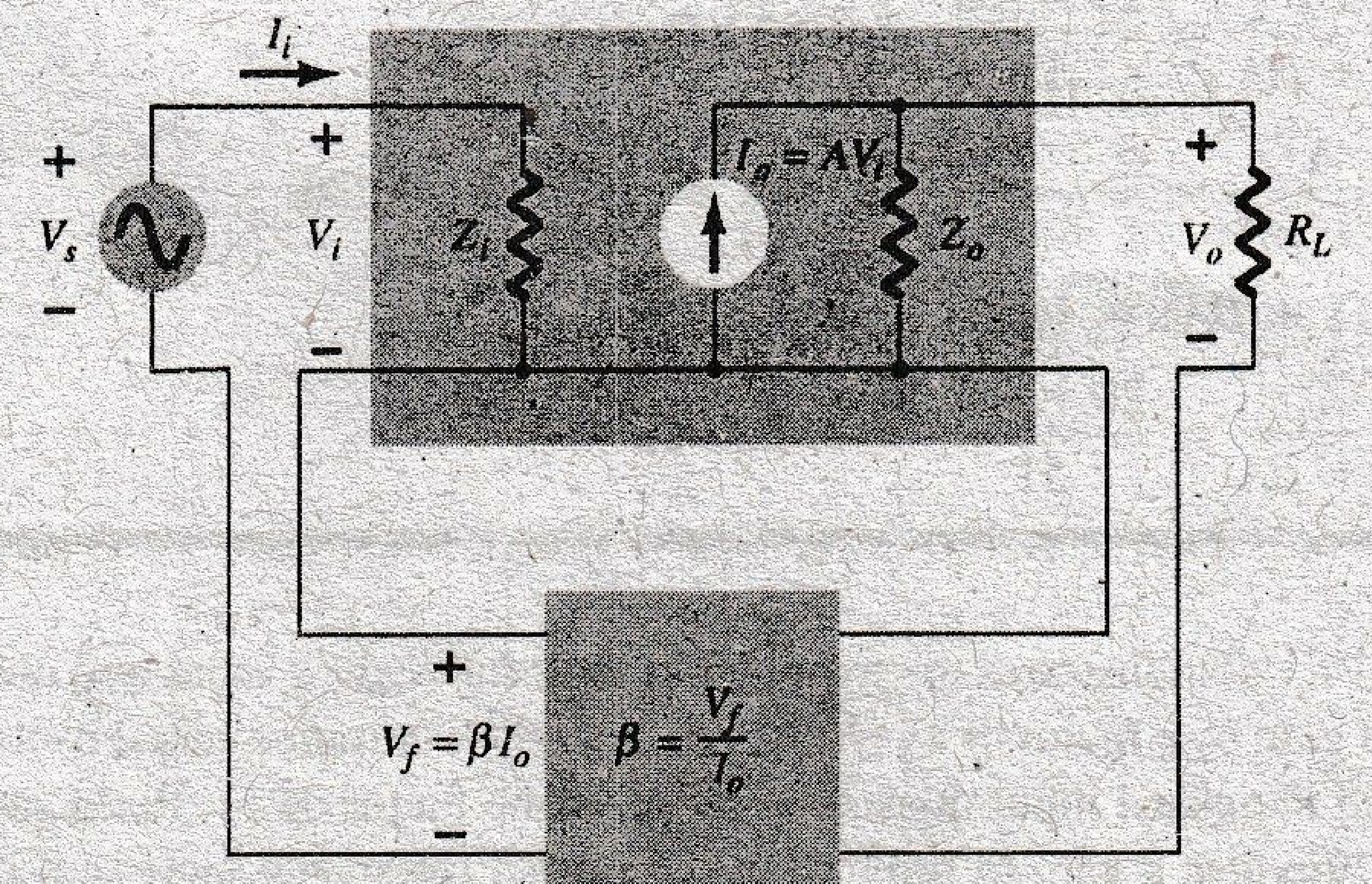


FIG. 14.5
Current-series feedback connection.

connection with current-series feedback. For the output part of a current-series feedback connection shown in Fig. 14.5, the resulting output impedance is determined as follows. With $V_s = 0$,

$$V_i = V_f$$

$$I = \frac{V}{Z_o} - AV_i = \frac{V}{Z_o} - AV_f = \frac{V}{Z_o} - A\beta I$$

$$Z_o(1 + \beta A)I = V$$

$$\boxed{Z_{of} = \frac{V}{I} = Z_o(1 + \beta A)} \quad (14.7)$$

A summary of the effect of feedback on input and output impedance is provided in Table 14.2.

TABLE 14.2
Effect of Feedback Connection on Input and Output Impedance

Voltage-Series	Current-Series	Voltage-Shunt	Current-Shunt
Z_{if} $Z_i(1 + \beta A)$ (increased)	$Z_i(1 + \beta A)$ (increased)	$\frac{Z_i}{1 + \beta A}$ (decreased)	$\frac{Z_i}{1 + \beta A}$ (decreased)
Z_{of} $\frac{Z_o}{1 + \beta A}$ (decreased)	$Z_o(1 + \beta A)$ (increased)	$\frac{Z_o}{1 + \beta A}$ (decreased)	$Z_o(1 + \beta A)$ (increased)

EXAMPLE 14.1 Determine the voltage gain, input, and output impedance with feedback for voltage-series feedback having $A = -100$, $R_i = 10\,\text{k}\Omega$, and $R_o = 20\,\text{k}\Omega$ for feedback of (a) $\beta = -0.1$ and (b) $\beta = -0.5$.

Solution: Using Eqs. (14.2), (14.4), and (14.6), we obtain

a. $$A_f = \frac{A}{1 + \beta A} = \frac{-100}{1 + (-0.1)(-100)} = \frac{-100}{11} = \mathbf{-9.09}$$

$$Z_{if} = Z_i(1 + \beta A) = 10\,\text{k}\Omega\,(11) = \mathbf{110\,k\Omega}$$

$$Z_{of} = \frac{Z_o}{1 + \beta A} = \frac{20 \times 10^3}{11} = \mathbf{1.82\,k\Omega}$$

b. $$A_f = \frac{A}{1 + \beta A} = \frac{-100}{1 + (-0.5)(-100)} = \frac{-100}{51} = \mathbf{-1.96}$$

$$Z_{if} = Z_i(1 + \beta A) = 10\,\text{k}\Omega\,(51) = \mathbf{510\,k\Omega}$$

$$Z_{of} = \frac{Z_o}{1 + \beta A} = \frac{20 \times 10^3}{51} = \mathbf{392.16\,\Omega}$$

Example 14.1 demonstrates the trade-off of gain for desired input and output resistance. Reducing the gain by a factor of 11 (from 100 to 9.09) is complemented by a reduced output resistance and increased input resistance by the same factor of 11. Reducing the gain by a factor of 51 provides a gain of only 2 but with input resistance increased by the factor of 51 (to over 500 kΩ) and output resistance reduced from 20 kΩ to under 400 Ω. Feedback offers the designer the choice of trading away some of the available amplifier gain for other desired circuit features.

Reduction in Frequency Distortion

For a negative-feedback amplifier having $\beta A \gg 1$, the gain with feedback is $A_f \cong 1/\beta$. It follows from this that if the feedback network is purely resistive, the gain with feedback is not dependent on frequency even though the basic amplifier gain is frequency dependent. Practically, the frequency distortion arising because of varying amplifier gain with frequency is considerably reduced in a negative-voltage feedback amplifier circuit.

Reduction in Noise and Nonlinear Distortion

Signal feedback tends to hold down the amount of noise signal (such as power-supply hum) and nonlinear distortion. The factor $(1 + \beta A)$ reduces both input noise and resulting nonlinear distortion for considerable improvement. However, there is a reduction in overall gain (the price required for the improvement in circuit performance). If additional stages are used to bring the overall gain up to the level without feedback, the extra stage(s) might introduce as much noise back into the system as that reduced by the feedback amplifier. This problem can be somewhat alleviated by readjusting the gain of the feedback-amplifier circuit to obtain higher gain while also providing reduced noise signal.

Effect of Negative Feedback on Gain and Bandwidth

In Eq. (14.2), the overall gain with negative feedback is shown to be

$$A_f = \frac{A}{1 + \beta A} \cong \frac{A}{\beta A} = \frac{1}{\beta} \qquad \text{for } \beta A \gg 1$$

As long as $\beta A \gg 1$, the overall gain is approximately $1/\beta$. For a practical amplifier (for single low- and high-frequency breakpoints) the open-loop gain drops off at high frequencies due to the active device and circuit capacitances. Gain may also drop off at low frequencies for capacitively coupled amplifier stages. Once the open-loop gain A drops low enough and the factor βA is no longer much larger than 1, the conclusion of Eq. (14.2) that $A_f \cong 1/\beta$ no longer holds true.

Figure 14.6 shows that the amplifier with negative feedback has more bandwidth (B_f) than the amplifier without feedback (B). The feedback amplifier has a higher upper 3-dB frequency and smaller lower 3-dB frequency.

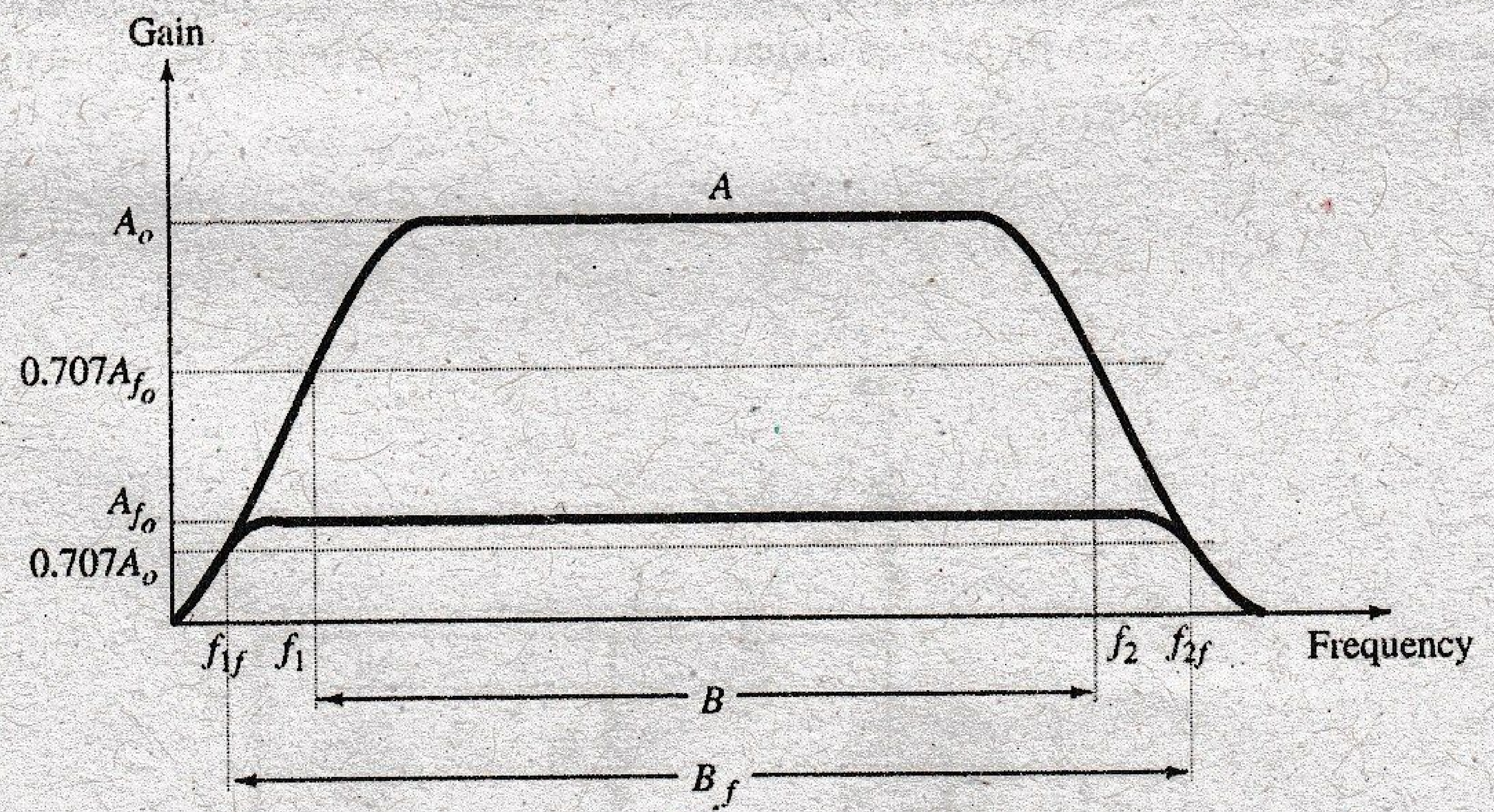


FIG. 14.6
Effect of negative feedback on gain and bandwidth.

It is interesting to note that the use of feedback, although resulting in a lowering of voltage gain, has provided an increase in B and in the upper 3-dB frequency particularly. In fact, the product of gain and frequency remains the same, so that the gain–bandwidth product of the basic amplifier is the same value for the feedback amplifier. However, since the feedback amplifier has lower gain, the net operation was to *trade* gain for bandwidth (we use bandwidth for the upper 3-dB frequency since typically $f_2 \gg f_1$).

Gain Stability with Feedback

In addition to the β factor setting a precise gain value, we are also interested in how stable the feedback amplifier is compared to an amplifier without feedback. Differentiating Eq. (14.2) leads to

$$\left|\frac{dA_f}{A_f}\right| = \frac{1}{|1+\beta A|}\left|\frac{dA}{A}\right| \tag{14.8}$$

$$\left|\frac{dA_f}{A_f}\right| \cong \left|\frac{1}{\beta A}\right|\left|\frac{dA}{A}\right| \quad \text{for } \beta A \gg 1 \tag{14.9}$$

This shows that magnitude of the relative change in gain $\left|\frac{dA_f}{A_f}\right|$ is reduced by the factor $|\beta A|$ compared to that without feedback $\left(\left|\frac{dA}{A}\right|\right)$.

EXAMPLE 14.2 If an amplifier with gain of -1000 and feedback of $\beta = -0.1$ has a gain change of 20% due to temperature, calculate the change in gain of the feedback amplifier.

Solution: Using Eq. (14.9), we get

$$\left|\frac{dA_f}{A_f}\right| \cong \left|\frac{1}{\beta A}\right|\left|\frac{dA}{A}\right| = \left|\frac{1}{-0.1(-1000)}(20\%)\right| = \mathbf{0.2\%}$$

The improvement is 100 times. Thus, whereas the amplifier gain changes from $|A| = 1000$ by 20%, the gain with feedback changes from $|A_f| = 100$ by only 0.2%.

14.3 PRACTICAL FEEDBACK CIRCUITS

Examples of practical feedback circuits will provide a means of demonstrating the effect feedback has on the various connection types. This section provides only a basic introduction to this topic.

Voltage-Series Feedback

Figure 14.7 shows an FET amplifier stage with voltage-series feedback. A part of the output signal (V_o) is obtained using a feedback network of resistors R_1 and R_2. The feedback voltage V_f is connected in series with the source signal V_s, their difference being the input signal V_i.

Without feedback the amplifier gain is

$$A = \frac{V_o}{V_i} = -g_m R_L \tag{14.10}$$

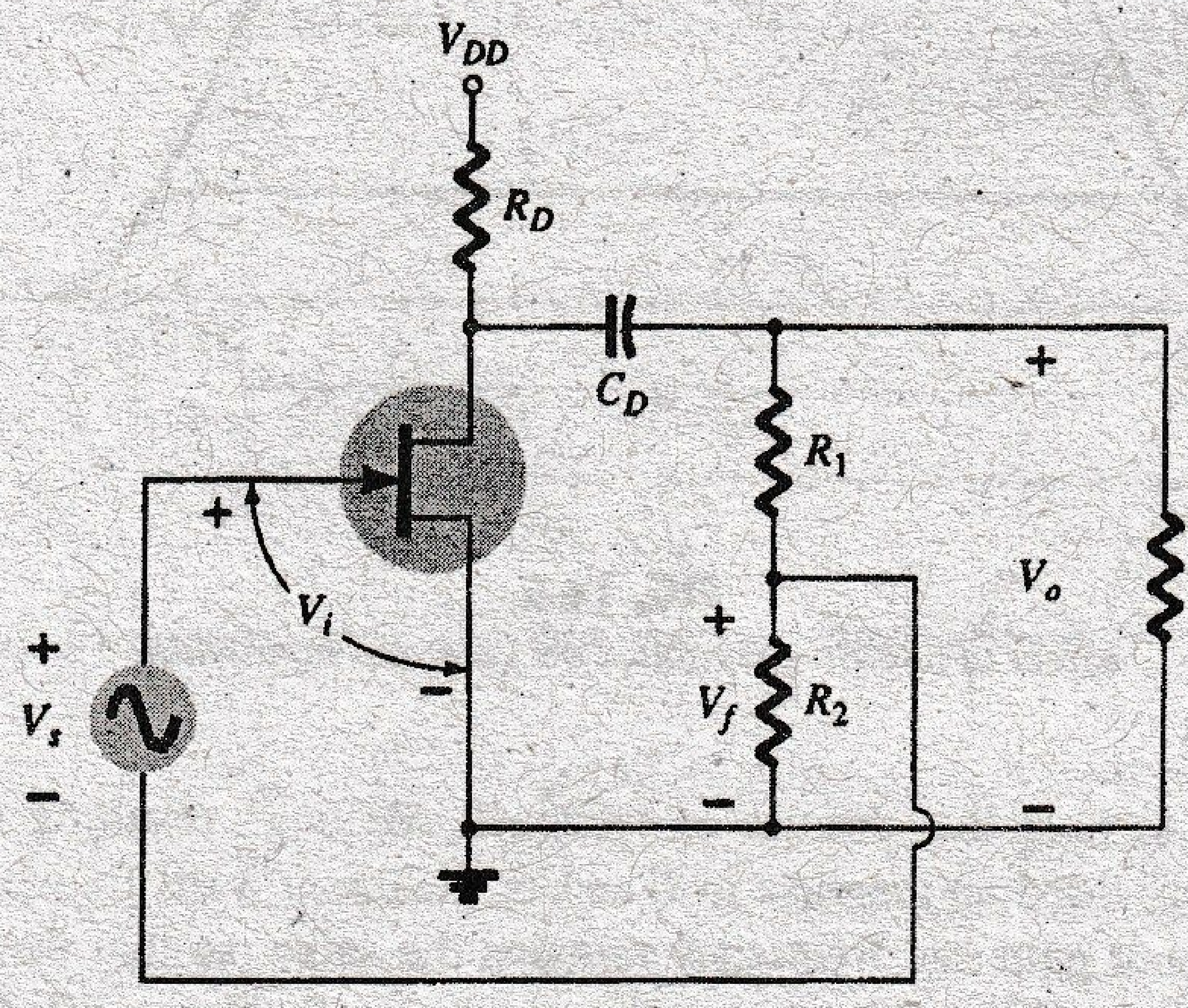


FIG. 14.7
FET amplifier stage with voltage-series feedback.

where R_L is the parallel combination of resistors:

$$R_L = R_D R_o(R_1 + R_2) \tag{14.11}$$

The feedback network provides a feedback factor of

$$\beta = \frac{V_f}{V_o} = \frac{-R_2}{R_1 + R_2} \tag{14.12}$$

Using the values of A and β above in Eq. (14.2), we find the gain with negative feedback to be

$$A_f = \frac{A}{1 + \beta A} = \frac{-g_m R_L}{1 + [R_2 R_L/(R_1 + R_2)]g_m} \tag{14.13}$$

If $\beta A \gg 1$, we have

$$\boxed{A_f \cong \frac{1}{\beta} = -\frac{R_1 + R_2}{R_2}} \tag{14.14}$$

EXAMPLE 14.3 Calculate the gain without and with feedback for the FET amplifier circuit of Fig. 14.7 and the following circuit values: $R_1 = 80\text{ k}\Omega$, $R_2 = 20\text{ k}\Omega$, $R_o = 10\text{ k}\Omega$, $R_D = 10\text{ k}\Omega$, and $g_m = 4000\ \mu\text{S}$.

Solution:

$$R_L \cong \frac{R_o R_D}{R_o + R_D} = \frac{10\text{ k}\Omega(10\text{ k}\Omega)}{10\text{ k}\Omega + 10\text{ k}\Omega} = 5\text{ k}\Omega$$

Neglecting the 100-kΩ resistance of R_1 and R_2 in series gives

$$A = -g_m R_L = -(4000 \times 10^{-6}\ \mu\text{S})(5\text{ k}\Omega) = \mathbf{-20}$$

The feedback factor is

$$\beta = \frac{-R_2}{R_1 + R_2} = \frac{-20\text{ k}\Omega}{80\text{ k}\Omega + 20\text{ k}\Omega} = -0.2$$

The gain with feedback is

$$A_f = \frac{A}{1 + \beta A} = \frac{-20}{1 + (-0.2)(-20)} = \frac{-20}{5} = \mathbf{-4}$$

Figure 14.8 shows a voltage-series feedback connection using an op-amp. The gain of the op-amp, A, without feedback, is reduced by the feedback factor

$$\beta = \frac{R_2}{R_1 + R_2} \tag{14.15}$$

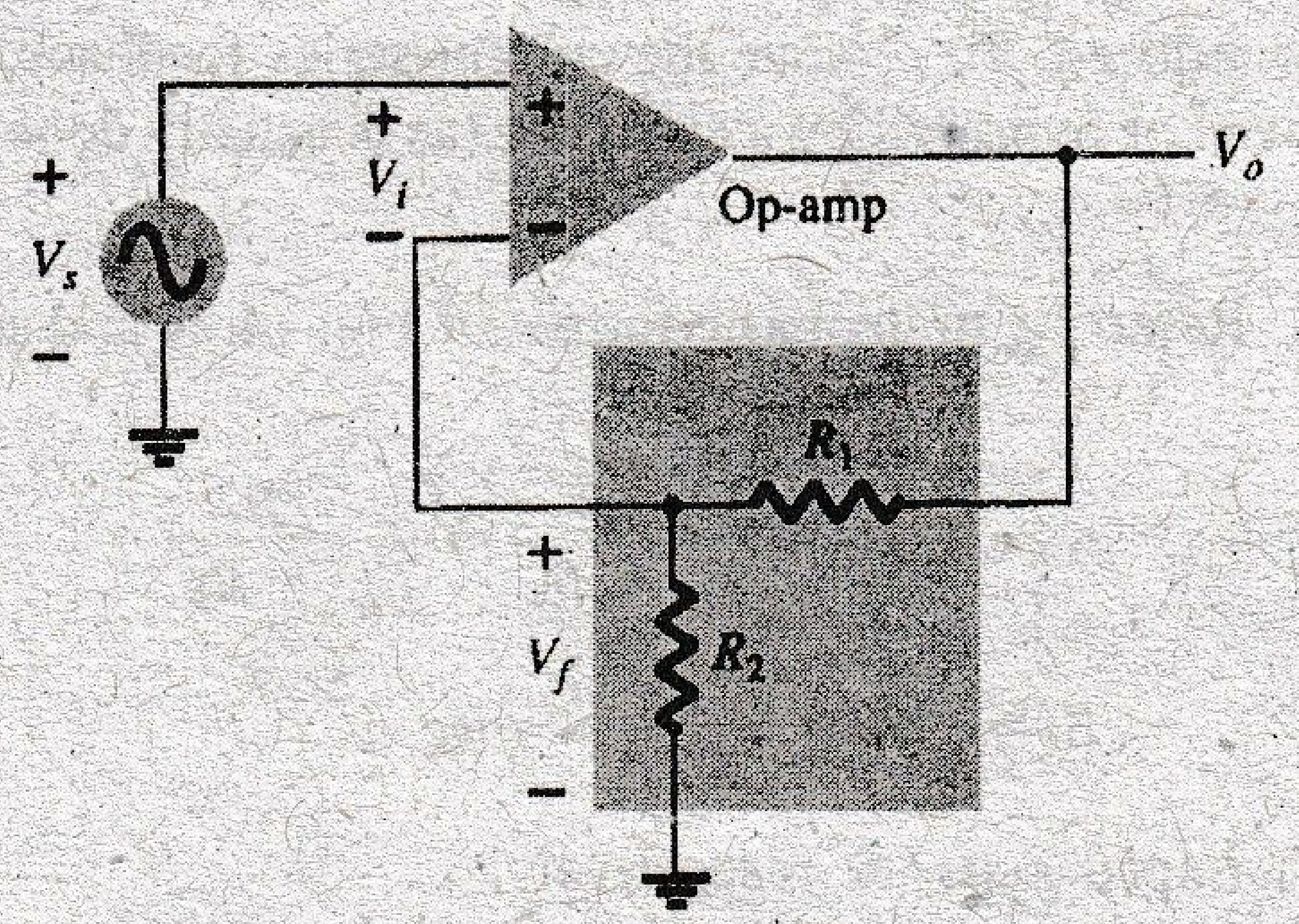


FIG. 14.8
Voltage-series feedback in an op-amp connection.

EXAMPLE 14.4 Calculate the amplifier gain of the circuit of Fig. 14.8 for op-amp gain $A = 100{,}000$ and resistances $R_1 = 1.8\text{ k}\Omega$ and $R_2 = 200\ \Omega$.

Solution:

$$\beta = \frac{R_2}{R_1 + R_2} = \frac{200\ \Omega}{200\ \Omega + 1.8\text{ k}\Omega} = 0.1$$

$$A_f = \frac{A}{1 + \beta A} = \frac{100{,}000}{1 + (0.1)(100{,}000)}$$

$$= \frac{100{,}000}{10{,}001} = 9.999$$

Note that since $\beta A \gg 1$,

$$A_f \cong \frac{1}{\beta} = \frac{1}{0.1} = \mathbf{10}$$

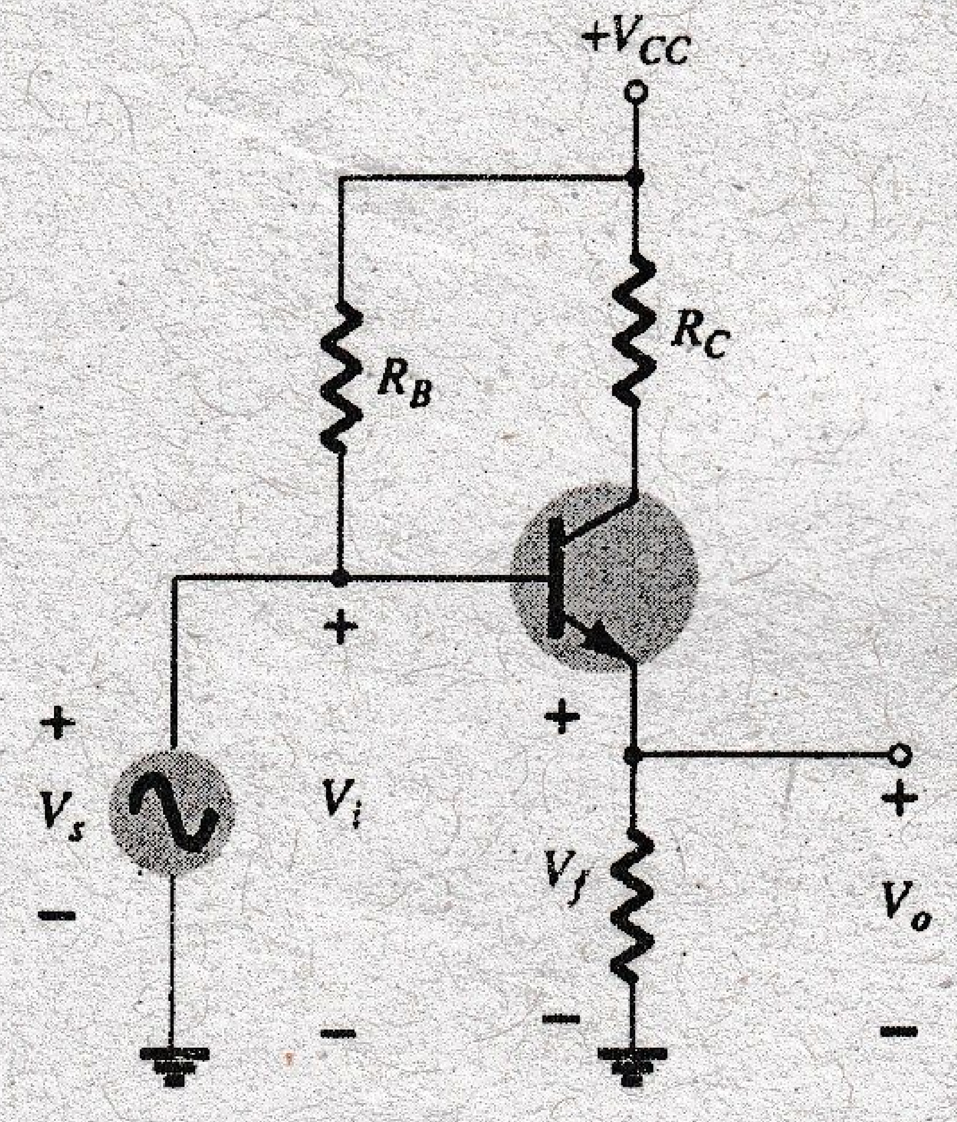


FIG. 14.9
Voltage-series feedback circuit (emitter-follower).

The emitter-follower circuit of Fig. 14.9 provides voltage-series feedback. The signal voltage V_s is the input voltage V_i. The output voltage V_o is also the feedback voltage in series with the input voltage. The amplifier, as shown in Fig. 14.9, provides the operation *with* feedback. The operation of the circuit without feedback provides $V_f = 0$, so that

$$A = \frac{V_o}{V_s} = \frac{h_{fe}I_bR_E}{V_s} = \frac{h_{fe}R_E(V_s/h_{ie})}{V_s} = \frac{h_{fe}R_E}{h_{ie}}$$

and

$$\beta = \frac{V_f}{V_o} = 1$$

The operation with feedback then provides that

$$A_f = \frac{V_o}{V_s} = \frac{A}{1 + \beta A} = \frac{h_{fe}R_E/h_{ie}}{1 + (1)(h_{fe}R_E/h_{ie})}$$

$$= \frac{h_{fe}R_E}{h_{ie} + h_{fe}R_E}$$

For $h_{fe}R_E \gg h_{ie}$,

$$A_f \cong 1$$

Current-Series Feedback

Another feedback technique is to sample the output current I_o and return a proportional voltage in series with the input. Although it stabilizes the amplifier gain, the current-series feedback connection increases input resistance.

Figure 14.10 shows a single transistor amplifier stage. Since the emitter of this stage has an unbypassed emitter, it effectively has current-series feedback. The current through resistor R_E results in a feedback voltage that opposes the source signal applied, so that the output voltage V_o is reduced. To remove the current-series feedback, the emitter resistor must be either removed or bypassed by a capacitor (as is usually done).

Without Feedback Referring to the basic format of Fig. 14.2a and summarized in Table 14.1, we have

$$A = \frac{I_o}{V_i} = \frac{-I_bh_{fe}}{I_bh_{ie} + R_E} = \frac{-h_{fe}}{h_{ie} + R_E} \tag{14.16}$$

$$\beta = \frac{V_f}{I_o} = \frac{-I_oR_E}{I_o} = -R_E \tag{14.17}$$

The input and output impedances are, respectively,

$$Z_i = R_B \| (h_{ie} + R_E) \cong h_{ie} + R_E \tag{14.18}$$

$$Z_o = R_C \tag{14.19}$$

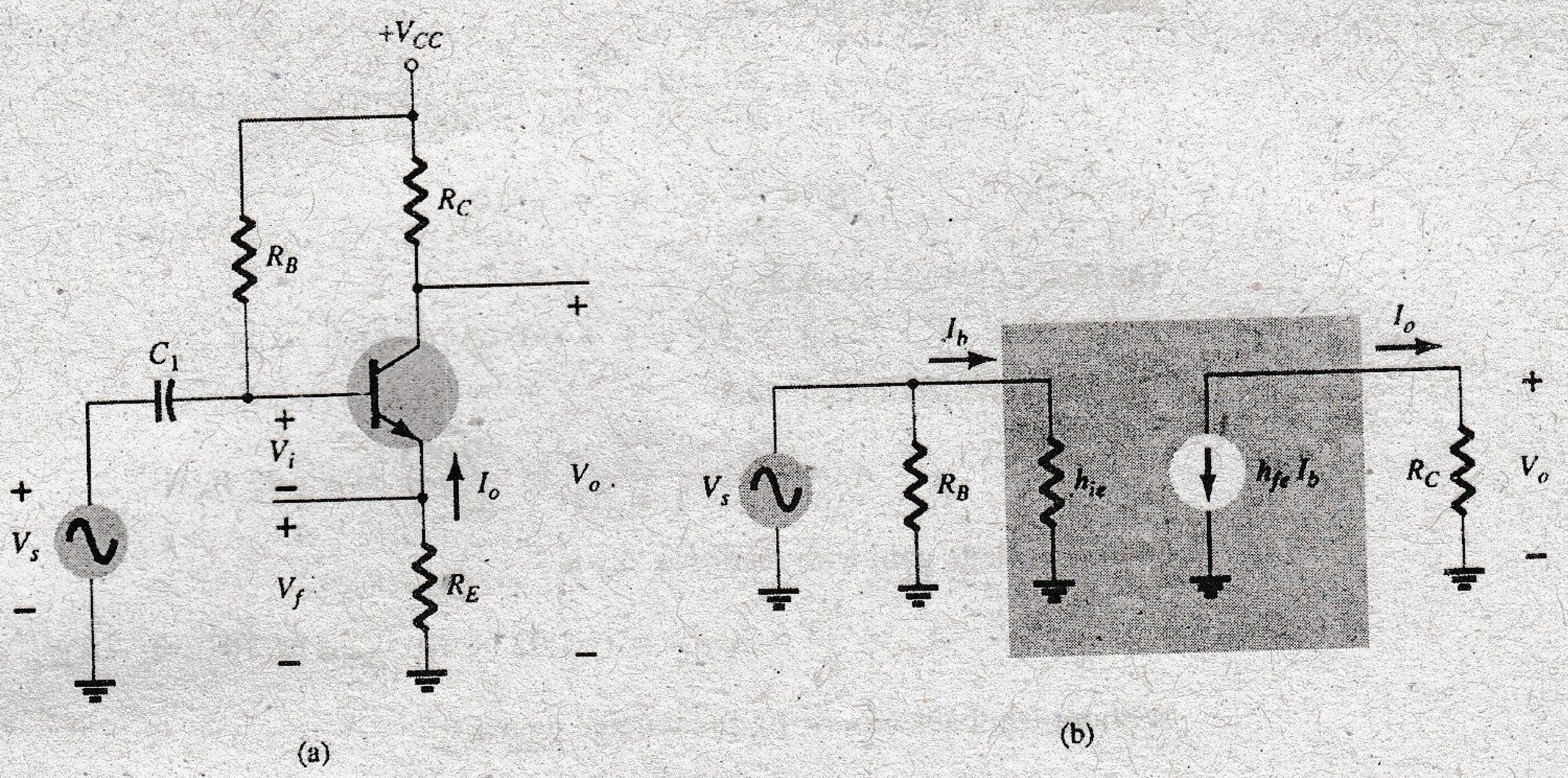


FIG. 14.10
Transistor amplifier with unbypassed emitter resistor (R_E) for current-series feedback: (a) amplifier circuit; (b) ac equivalent circuit without feedback.

With Feedback

$$A_f = \frac{I_o}{V_s} = \frac{A}{1 + \beta A} = \frac{-h_{fe}/h_{ie}}{1 + (-R_E)\left(\dfrac{-h_{fe}}{h_{ie} + R_E}\right)} \cong \frac{-h_{fe}}{h_{ie} + h_{fe}R_E} \quad \textbf{(14.20)}$$

The input and output impedances are calculated as specified in Table 14.2:

$$Z_{if} = Z_i(1 + \beta A) \cong h_{ie}\left(1 + \frac{h_{fe}R_E}{h_{ie}}\right) = h_{ie} + h_{fe}R_E \quad \textbf{(14.21)}$$

$$Z_{of} = Z_o(1 + \beta A) = R_C\left(1 + \frac{h_{fe}R_E}{h_{ie}}\right) \quad \textbf{(14.22)}$$

The voltage gain A with feedback is

$$A_{vf} = \frac{V_o}{V_s} = \frac{I_oR_C}{V_s} = \left(\frac{I_o}{V_s}\right)R_C = A_fR_C \cong \frac{-h_{fe}R_C}{h_{ie} + h_{fe}R_E} \quad \textbf{(14.23)}$$

EXAMPLE 14.5 Calculate the voltage gain of the circuit of Fig. 14.11.

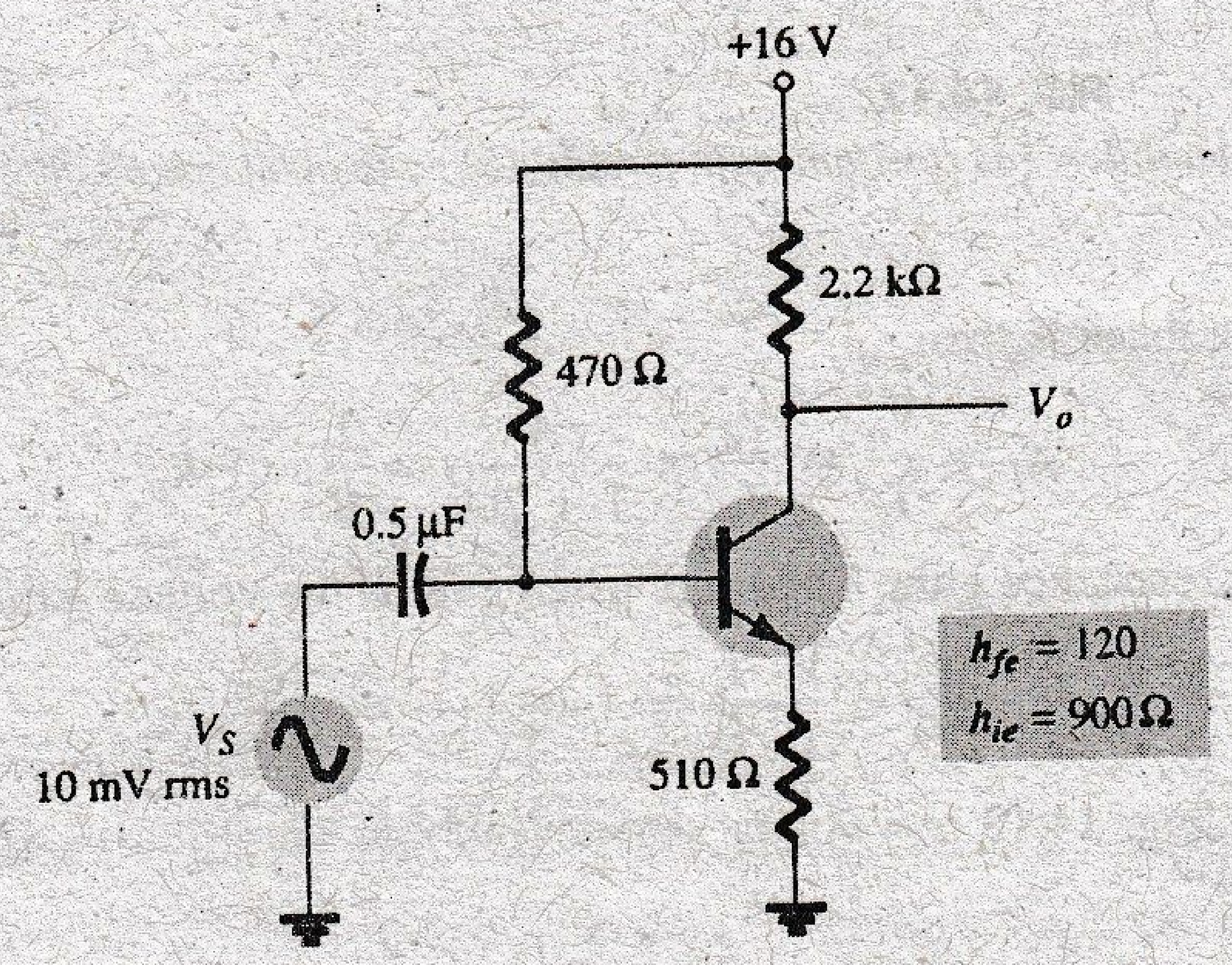


FIG. 14.11
BJT amplifier with current-series feedback for Example 14.5.

Solution: Without feedback,

$$A = \frac{I_o}{V_i} = \frac{-h_{fe}}{h_{ie} + R_E} = \frac{-120}{900 + 510} = -0.085$$

$$\beta = \frac{V_f}{I_o} = -R_E = -510$$

The factor $(1 + \beta A)$ is then

$$1 + \beta A = 1 + (-0.085)(-510) = 44.35$$

The gain with feedback is then

$$A_f = \frac{I_o}{V_s} = \frac{A}{1 + \beta A} = \frac{-0.085}{44.35} = -1.92 \times 10^{-3}$$

and the voltage gain with feedback A_{vf} is

$$A_{vf} = \frac{V_o}{V_s} = A_f R_C = (-1.92 \times 10^{-3})(2.2 \times 10^3) = \mathbf{-4.2}$$

Without feedback ($R_E = 0$), the voltage gain is

$$A_v = \frac{-R_C}{r_e} = \frac{-2.2 \times 10^3}{7.5} = \mathbf{-293.3}$$

Voltage-Shunt Feedback

The constant-gain op-amp circuit of Fig. 14.12a provides voltage-shunt feedback. Referring to Fig. 14.2b and Table 14.1 and the op-amp ideal characteristics $I_i = 0$, $V_i = 0$, and voltage gain of infinity, we have

$$A = \frac{V_o}{I_i} = \infty \tag{14.24}$$

$$\beta = \frac{I_f}{V_o} = \frac{-1}{R_o} \tag{14.25}$$

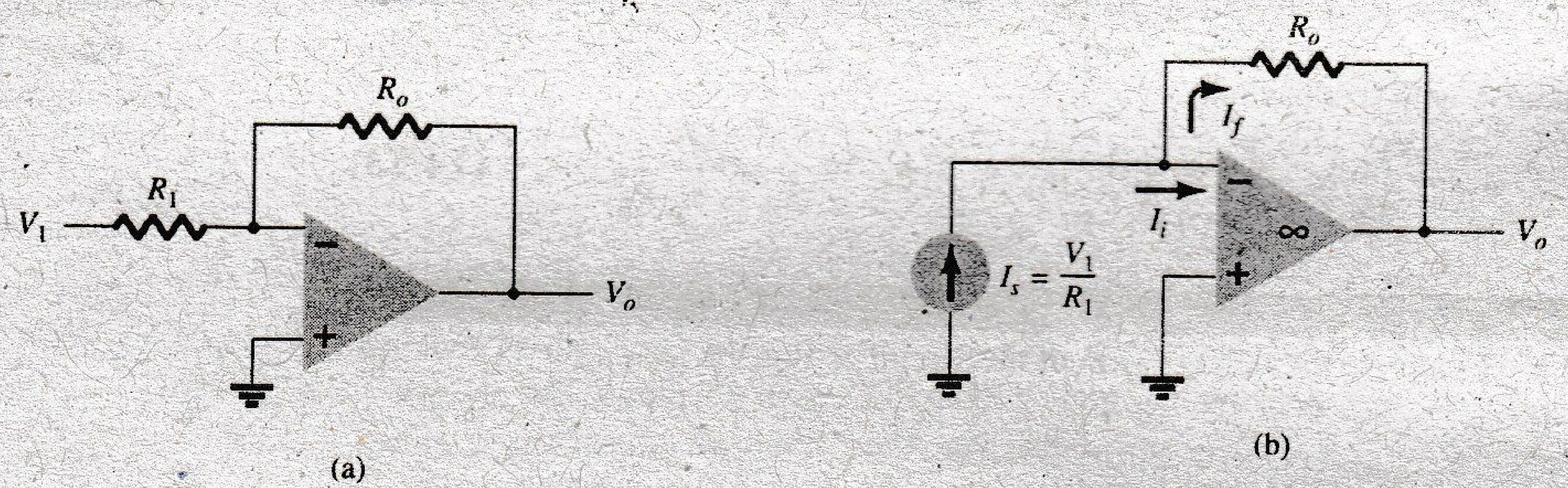


FIG. 14.12
Voltage-shunt negative feedback amplifier: (a) constant-gain circuit; (b) equivalent circuit.

The gain with feedback is then

$$A_f = \frac{V_o}{I_s} = \frac{V_o}{I_i} = \frac{A}{1 + \beta A} = \frac{1}{\beta} = -R_o \tag{14.26}$$

This is a transfer resistance gain. The more usual gain is the voltage gain with feedback,

$$A_{vf} = \frac{V_o}{I_s}\frac{I_s}{V_1} = (-R_o)\frac{1}{R_1} = \frac{-R_o}{R_1} \tag{14.27}$$

The circuit of Fig. 14.13 is a voltage-shunt feedback amplifier using an FET with no feedback, $V_f = 0$.

$$A = \frac{V_o}{I_i} \cong -g_m R_D R_S \tag{14.28}$$

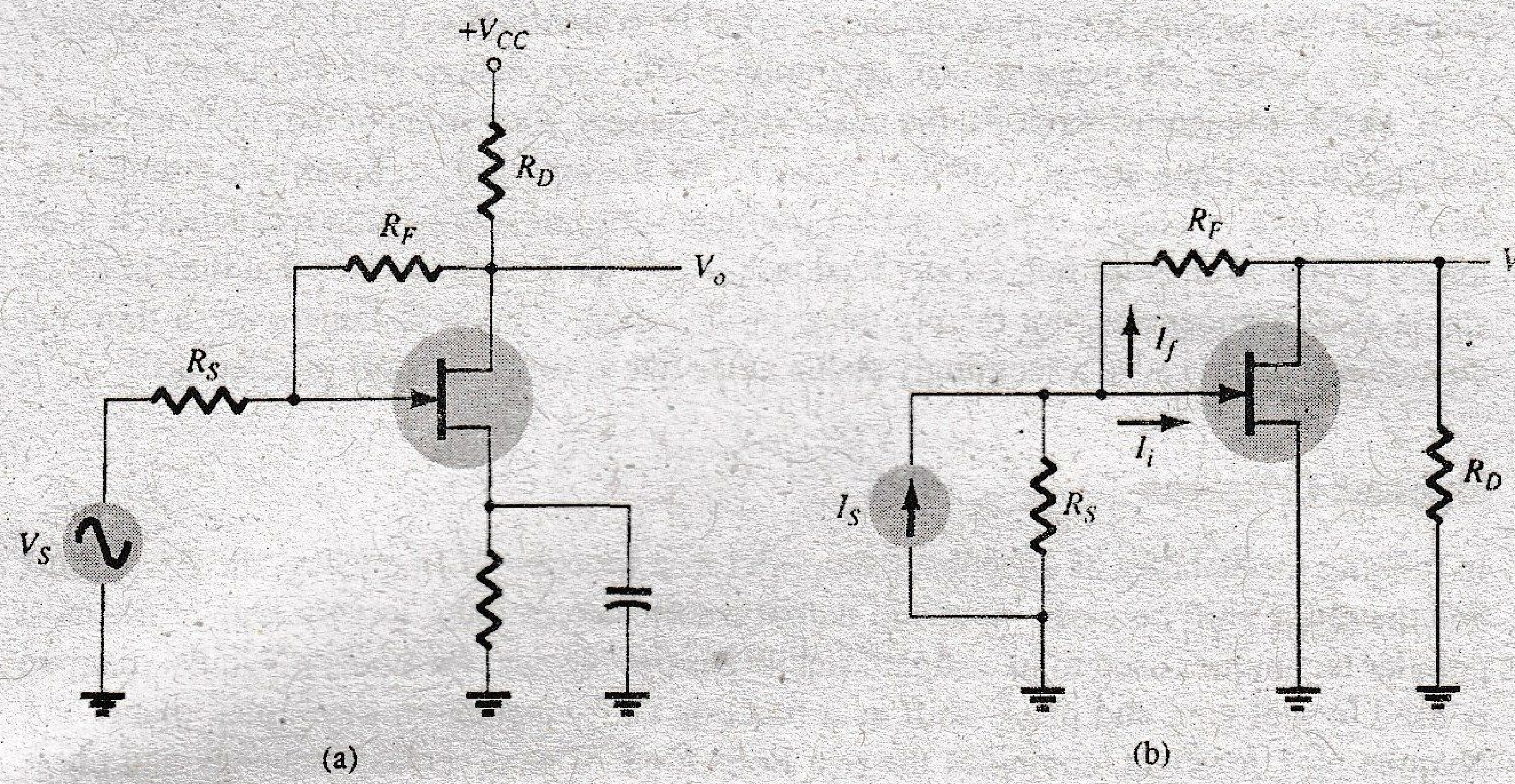


FIG. 14.13
Voltage-shunt feedback amplifier using an FET: (a) circuit; (b) equivalent circuit.

The feedback is

$$\beta = \frac{I_f}{V_o} = \frac{-1}{R_F} \tag{14.29}$$

With feedback, the gain of the circuit is

$$A_f = \frac{V_o}{I_s} = \frac{A}{1 + \beta A} = \frac{-g_m R_D R_S}{1 + (-1/R_F)(-g_m R_D R_S)}$$
$$= \frac{-g_m R_D R_S R_F}{R_F + g_m R_D R_S} \tag{14.30}$$

The voltage gain of the circuit with feedback is then

$$A_{vf} = \frac{V_o}{I_s}\frac{I_s}{V_s} = \frac{-g_m R_D R_S R_F}{R_F + g_m R_D R_S}\left(\frac{1}{R_S}\right)$$
$$= \frac{-g_m R_D R_F}{R_F + g_m R_D R_S} = (-g_m R_D)\frac{R_F}{R_F + g_m R_D R_S} \tag{14.31}$$

EXAMPLE 14.6 Calculate the voltage gain with and without feedback for the circuit of Fig. 14.13a with values of $g_m = 5$ mS, $R_D = 5.1$ kΩ, $R_S = 1$ kΩ, and $R_F = 20$ kΩ.

Solution: Without feedback, the voltage gain is

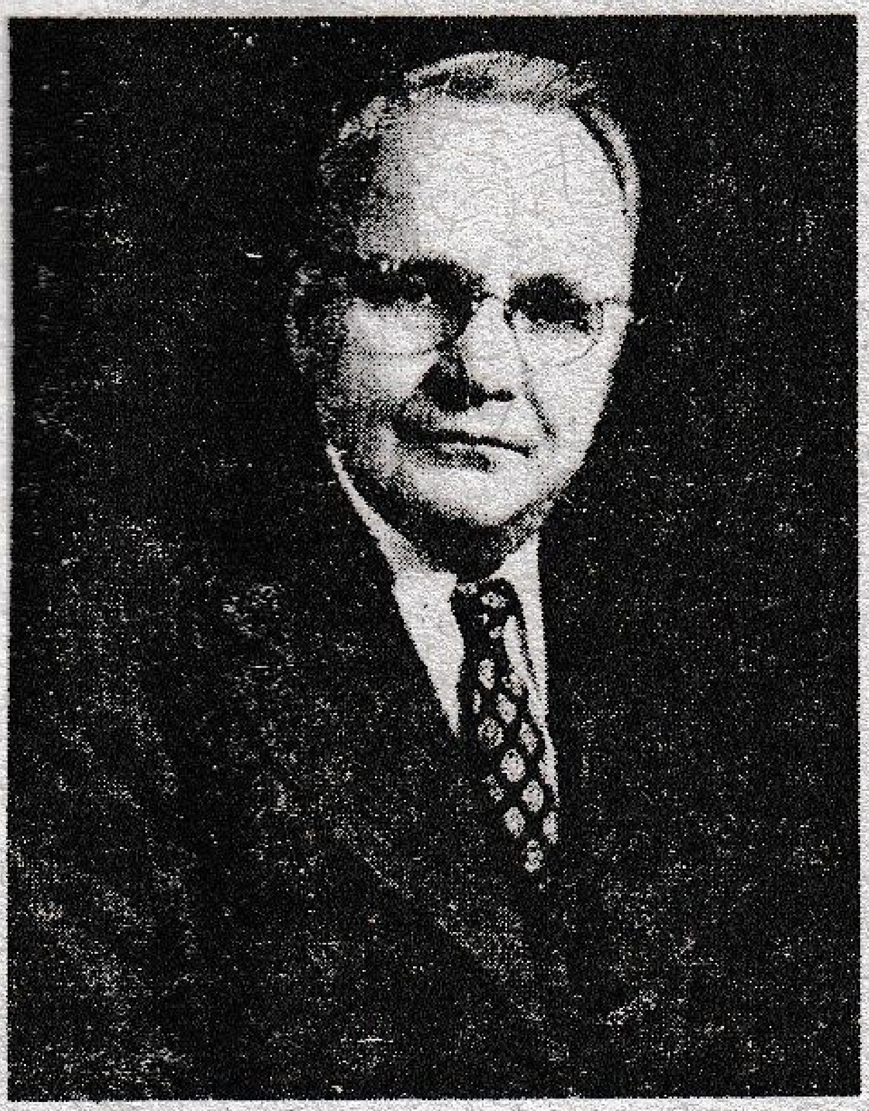
$$A_v = -g_m R_D = -(5 \times 10^{-3})(5.1 \times 10^3) = \mathbf{-25.5}$$

With feedback the gain is reduced to

$$A_{vf} = (-g_m R_D)\frac{R_F}{R_F + g_m R_D R_S}$$
$$= (-25.5)\frac{20 \times 10^3}{(20 \times 10^3) + (5 \times 10^{-3})(5.1 \times 10^3)(1 \times 10^3)}$$
$$= -25.5(0.44) = \mathbf{-11.2}$$

14.4 FEEDBACK AMPLIFIER—PHASE AND FREQUENCY CONSIDERATIONS

So far we have considered the operation of a feedback amplifier in which the feedback signal was *opposite* to the input signal—negative feedback. In any practical circuit this condition occurs only for some mid-frequency range of operation. We know that an amplifier



Harry Nyquist was born in Sweden in 1889. He immigrated to the United States in 1907, and died in Texas in 1976. He received a Ph.D. in physics from Yale University in 1917. He worked at AT&T's Department of Development and Research and at Bell Telephone Laboratories from 1917 until his retirement in 1954. As an engineer at Bell Laboratories, Nyquist did important work on thermal noise, the stability of feedback amplifiers, telegraphy, facsimile, television, and other important communications problems. In 1932, he published a classic paper on stability of feedback amplifiers: The Nyquist stability criterion can now be found in all textbooks on feedback control theory.

(Courtesy of AT&T Archives and History Center)

gain will change with frequency, dropping off at high frequencies from the mid-frequency value. In addition, the phase shift of an amplifier will also change with frequency.

If, as the frequency increases, the phase shift changes, then some of the feedback signal will *add* to the input signal. It is then possible for the amplifier to break into oscillations due to positive feedback. If the amplifier oscillates at some low or high frequency, it is no longer useful as an amplifier. Proper feedback-amplifier design requires that the circuit be stable at *all* frequencies, not merely those in the range of interest. Otherwise, a transient disturbance could cause a seemingly stable amplifier to suddenly start oscillating.

Nyquist Criterion

In judging the stability of a feedback amplifier as a function of frequency, the βA product and the phase shift between input and output are the determining factors. One of the most popular techniques used to investigate stability is the Nyquist method. A Nyquist diagram is used to plot gain and phase shift as a function of frequency on a complex plane. The Nyquist plot, in effect, combines the two Bode plots of gain versus frequency and phase shift versus frequency on a single plot. A Nyquist plot is used to quickly show whether an amplifier is stable for all frequencies and how stable the amplifier is relative to some gain or phase-shift criteria.

As a start, consider the *complex plane* shown in Fig. 14.14. A few points of various gain (βA) values are shown at a few different phase-shift angles. By using the positive real axis as reference (0°), we see a magnitude of $\beta A = 2$ at a phase shift of 0° at point 1. Additionally, a magnitude of $\beta A = 3$ at a phase shift of $-135°$ is shown at point 2 and a magnitude/phase of $\beta A = 1$ at 180° is shown at point 3. Thus points on this plot can represent *both* gain magnitude of βA and phase shift. If the points representing gain and phase shift for an amplifier circuit are plotted at increasing frequency, then a Nyquist plot is obtained as shown by the plot in Fig. 14.15. At the origin, the gain is 0 at a frequency of 0 (for *RC*-type coupling). At increasing frequency, points f_1, f_2, and f_3 and the phase shift increase, as does the magnitude of βA. At a representative frequency f_4, the value of A is the vector length from the origin to point f_4 and the phase shift is the angle ϕ. At a frequency f_5, the phase shift is 180°. At higher frequencies, the gain is shown to decrease back to 0.

The Nyquist criterion for stability can be stated as follows:

The amplifier is unstable if the Nyquist curve encloses (encircles) the –1 point, and it is stable otherwise.

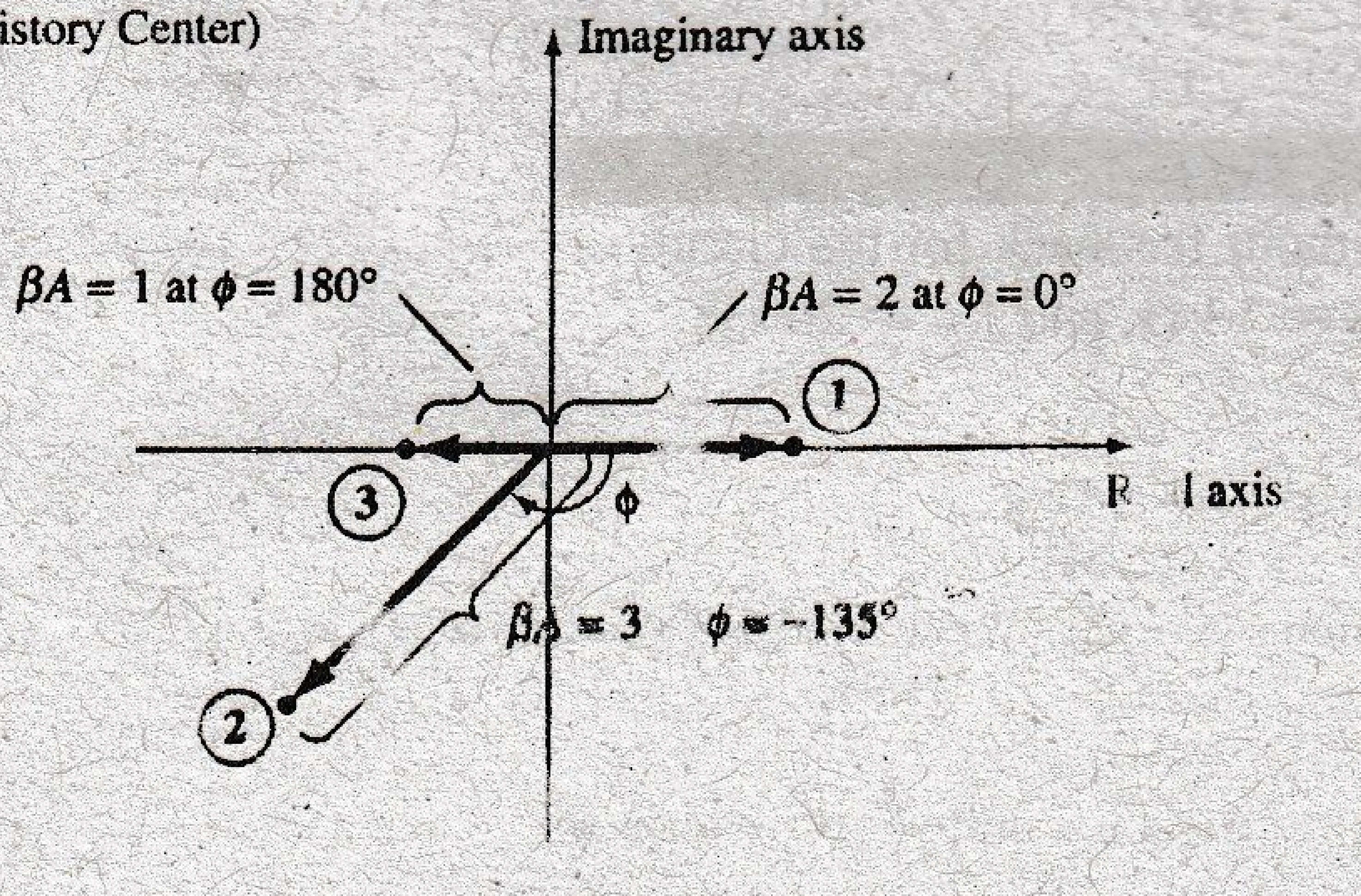


FIG. 14.14
Complex plane showing typical gain-phase points.

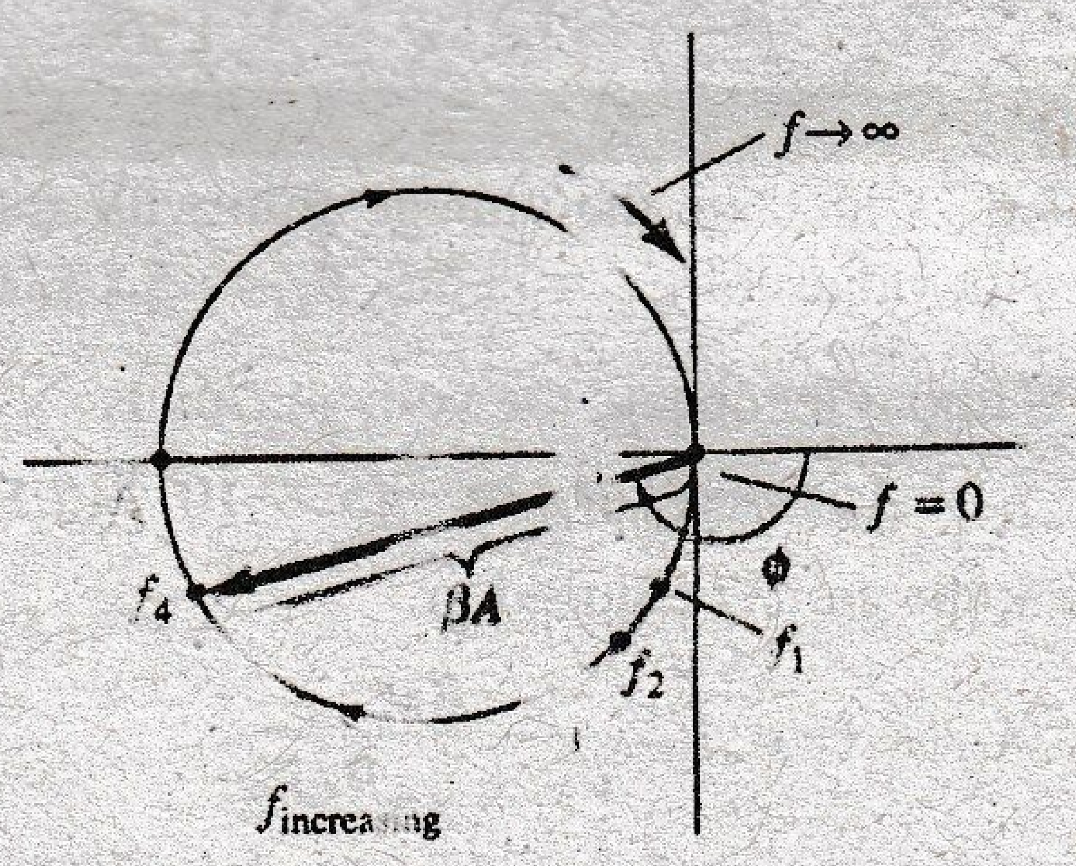


FIG. 14.15
Nyquist plot.

An example of the Nyquist criterion is demonstrated by the curves in Fig. 14.16. The Nyquist plot in Fig. 14.16a is stable since it does not encircle the -1 point, whereas that shown in Fig. 14.16b is unstable since the curve does encircle the -1 point. Keep in mind that encircling the -1 point means that at a phase shift of 180° the loop gain (βA) is greater than 1; therefore, the feedback signal is in phase with the input and large enough to result in a larger input signal than that applied, with the result that oscillation occurs.

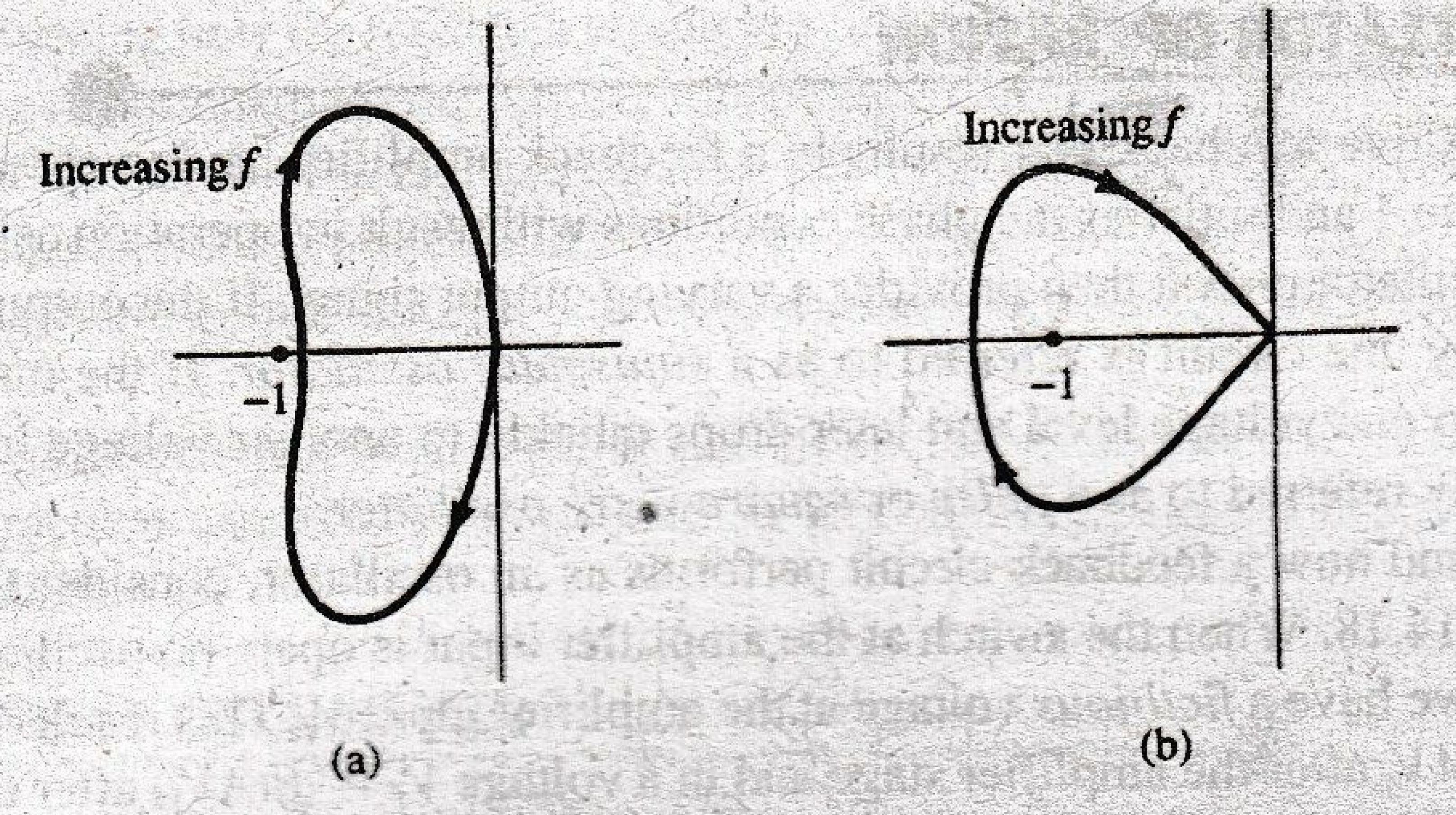


FIG. 14.16
Nyquist plots showing stability conditions: (a) stable; (b) unstable.

Gain and Phase Margins

From the Nyquist criterion, we know that a feedback amplifier is stable if the loop gain (βA) is less than unity (0 dB) when its phase angle is 180°. We can additionally determine some margins of stability to indicate how close to instability the amplifier is. That is, if the gain (βA) is less than unity but, say, 0.95 in value, this would not be as relatively stable as another amplifier having, say, $\beta A = 0.7$ (both measured at 180°). Of course, amplifiers with loop gains 0.95 and 0.7 are both stable, but one is closer to instability, if the loop gain increases, than the other. We can define the following terms:

Gain margin (GM) is defined as the negative of the value of $|\beta A|$ in decibels at the frequency at which the phase angle is 180°. Thus, 0 dB, equal to a value of $\beta A = 1$, is on the border of stability and any negative decibel value is stable. The GM may be evaluated in decibels from the curve of Fig. 14.17.

Phase margin (PM) is defined as the angle of 180° minus the magnitude of the angle at which the value $|\beta A|$ is unity (0 dB). The PM may also be evaluated directly from the curve of Fig. 14.17.

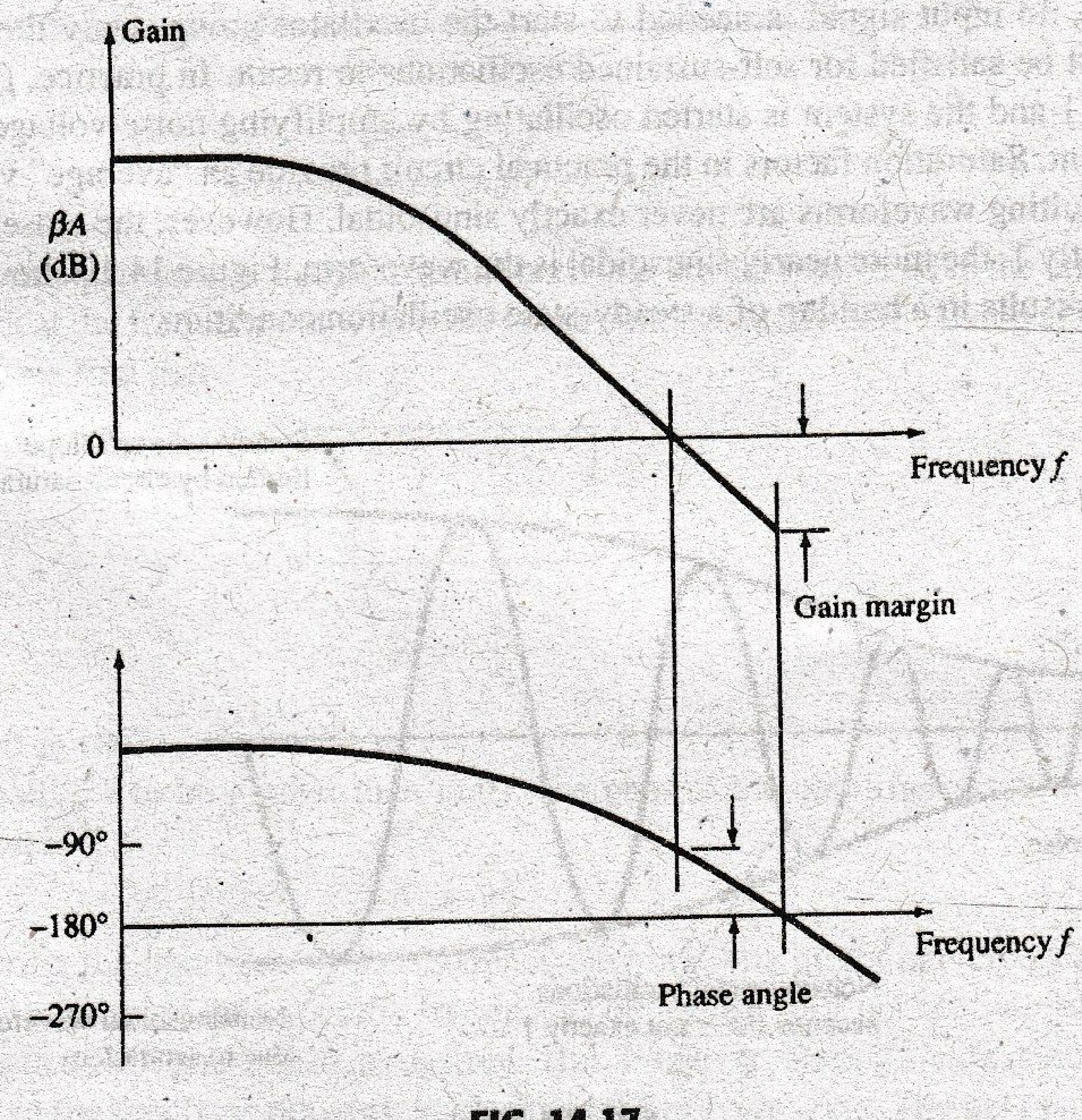


FIG. 14.17
Bode plots showing gain and phase margins.

14.5 OSCILLATOR OPERATION

The use of positive feedback that results in a feedback amplifier having closed-loop gain $|A_f|$ greater than 1 and satisfies the phase conditions will result in operation as an oscillator circuit. An oscillator circuit then provides a varying output signal. If the output signal varies sinusoidally, the circuit is referred to as a *sinusoidal oscillator.* If the output voltage rises quickly to one voltage level and later drops quickly to another voltage level, the circuit is generally referred to as a *pulse* or *square-wave oscillator.*

To understand how a feedback circuit performs as an oscillator, consider the feedback circuit of Fig. 14.18. When the switch at the amplifier input is open, no oscillation occurs. Consider that we have a *fictitious* voltage at the amplifier input V_i. This results in an output voltage $V_o = AV_i$ after the amplifier stage and in a voltage $V_f = \beta(AV_i)$ after the feedback stage. Thus, we have a feedback voltage $V_f = \beta AV_i$, where βA is referred to as the *loop gain.* If the circuits of the base amplifier and feedback network provide βA of a correct magnitude and phase, V_f can be made equal to V_i. Then, when the switch is closed and the fictitious voltage V_i is removed, the circuit will continue operating since the feedback voltage is sufficient to drive the amplifier and feedback circuits, resulting in a proper input voltage to sustain the loop operation. The output waveform will still exist after the switch is closed if the condition

$$\beta A = 1 \tag{14.32}$$

is met. This is known as the *Barkhausen criterion* for oscillation.

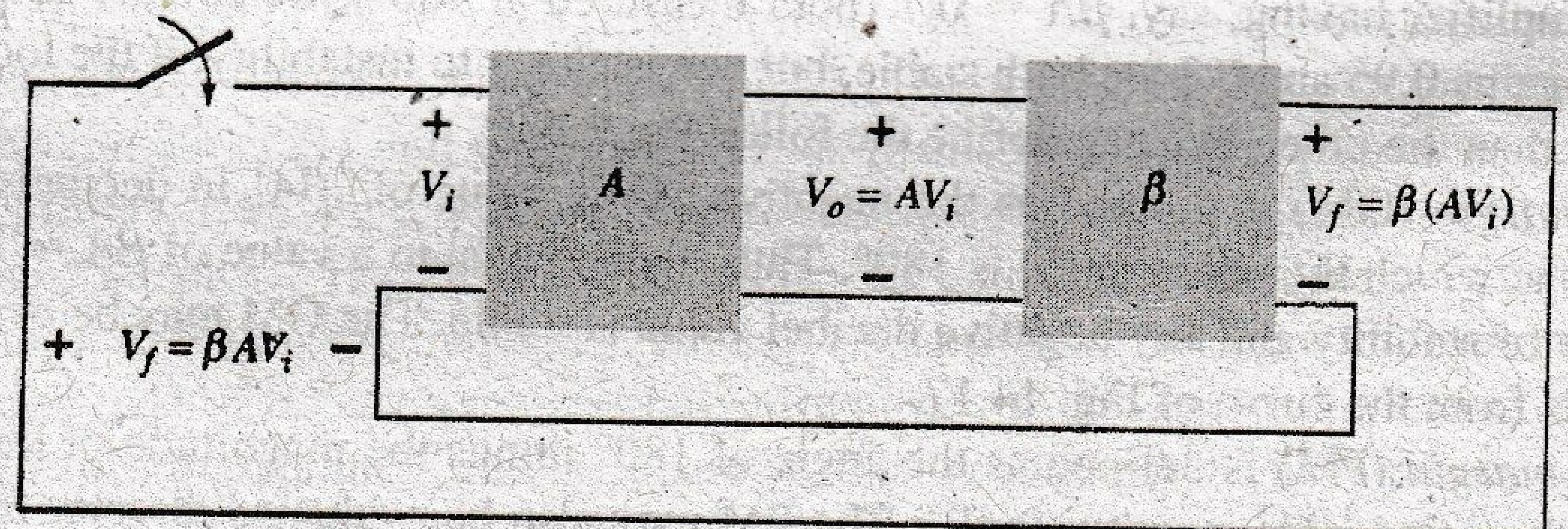


FIG. 14.18
Feedback circuit used as an oscillator.

In reality, no input signal is needed to start the oscillator going. Only the condition $\beta A = 1$ must be satisfied for self-sustained oscillations to result. In practice, βA is made greater than 1 and the system is started oscillating by amplifying noise voltage, which is always present. Saturation factors in the practical circuit provide an "average" value of βA of 1. The resulting waveforms are never exactly sinusoidal. However, the closer the value βA is to exactly 1, the more nearly sinusoidal is the waveform. Figure 14.19 shows how the noise signal results in a buildup of a steady-state oscillation condition.

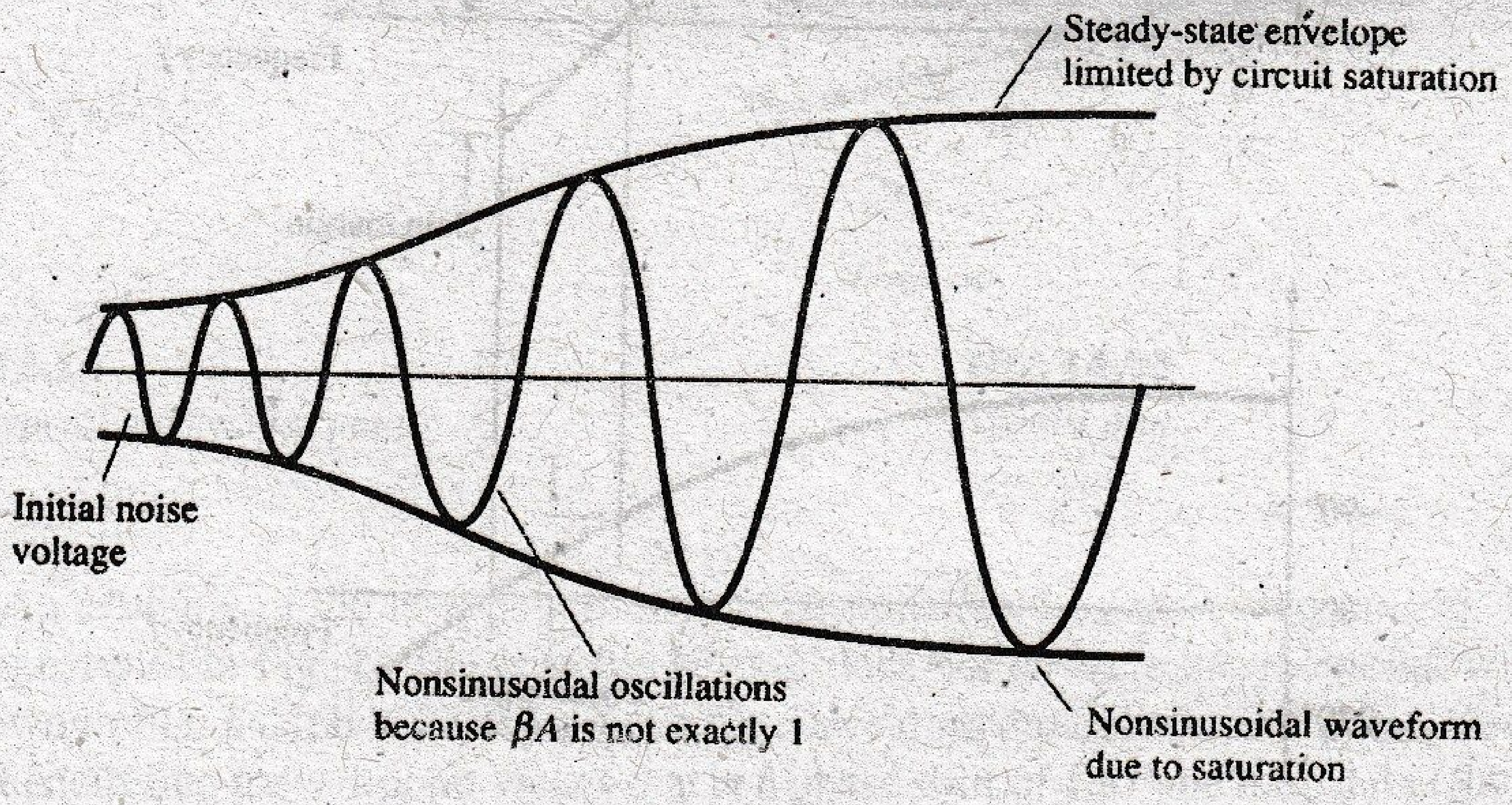


FIG. 14.19
Buildup of steady-state oscillations.

Another way of seeing how the feedback circuit provides operation as an oscillator is obtained by noting the denominator in the basic feedback equation (14.2), $A_f = A/(1 + \beta A)$. When $\beta A = -1$ or magnitude 1 at a phase angle of 180°, the denominator becomes 0 and the gain with feedback A_f becomes infinite. Thus, an infinitesimal signal (noise voltage) can provide a measurable output voltage, and the circuit acts as an oscillator even without an input signal.

The remainder of this chapter is devoted to various oscillator circuits that use a variety of components. Practical examples are included so that workable circuits in each of the various cases are discussed.

14.6 PHASE-SHIFT OSCILLATOR

An example of an oscillator circuit that follows the basic development of a feedback circuit is the *phase-shift oscillator.* An idealized version of this circuit is shown in Fig. 14.20. Recall that the requirements for oscillation are that the loop gain βA is greater than unity *and* that the phase shift around the feedback network is 180° (providing positive feedback). In the present idealization, we are considering the feedback network to be driven by a perfect source (zero source impedance) and the output of the feedback network to be connected into a perfect load (infinite load impedance). The idealized case will allow development of the theory behind the operation of the phase-shift oscillator. Practical circuit versions will then be considered.

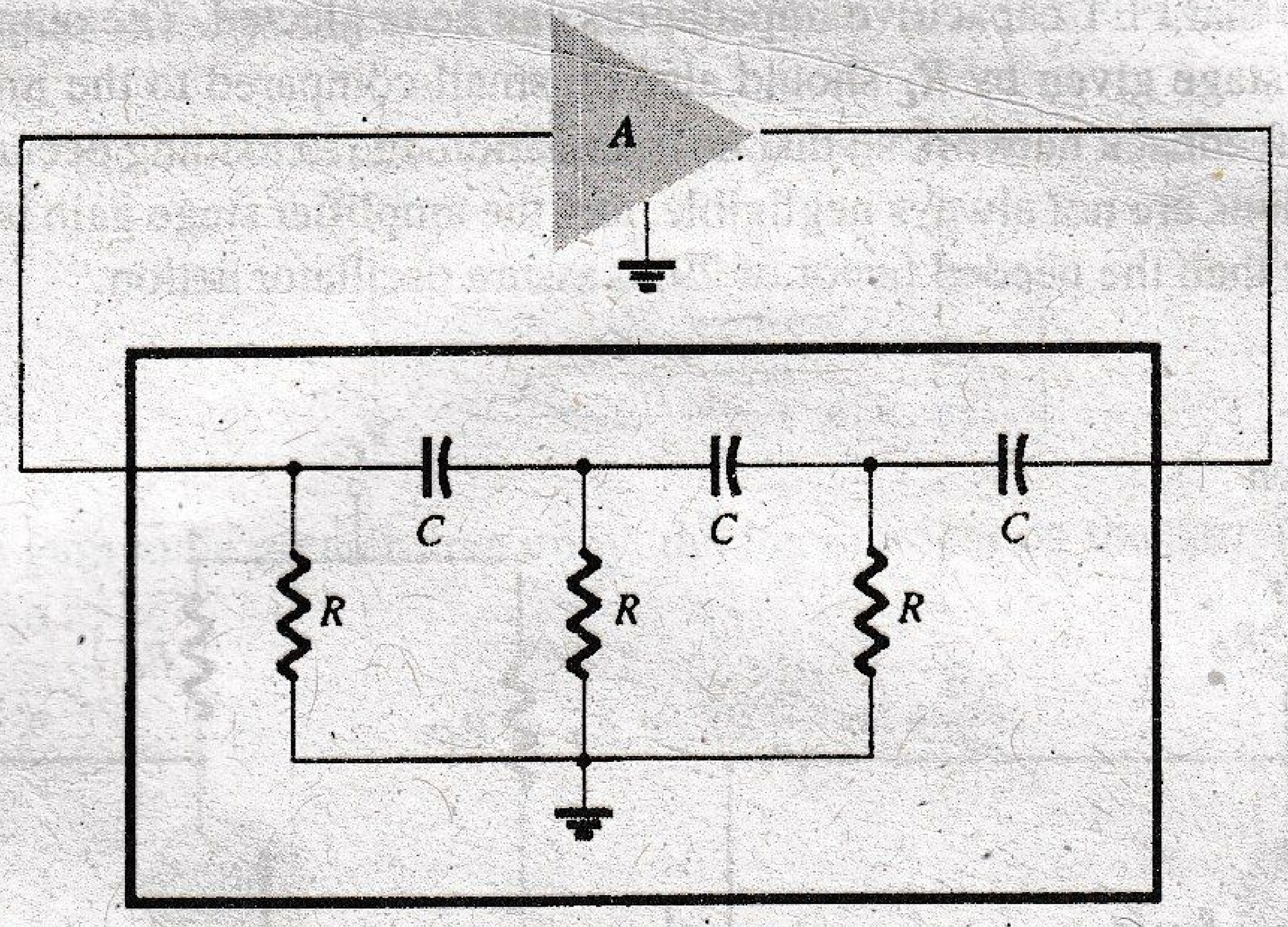


FIG. 14.20
Idealized phase-shift oscillator.

Concentrating our attention on the phase-shift network, we are interested in the attenuation of the network at the frequency at which the phase shift is exactly 180°. Using classical network analysis, we find that

$$f = \frac{1}{2\pi RC\sqrt{6}} \qquad \textbf{(14.33)}$$

$$\beta = \frac{1}{29} \qquad \textbf{(14.34)}$$

and the phase shift is 180°.

For the loop gain βA to be greater than unity, the gain of the amplifier stage must be greater than $1/\beta$ or 29:

$$A > 29 \qquad \textbf{(14.35)}$$

When considering the operation of the feedback network, one might naively select the values of R and C to provide (at a specific frequency) 60°-phase shift per section for three sections, resulting in a 180° phase shift, as desired. This, however, is not the case, since each section of the RC in the feedback network loads down the previous one. The net result that the *total* phase shift be 180° is all that is important. The frequency given by Eq. (14.33) is that at which the *total* phase shift is 180°. If one measured the phase shift per RC section, each section would not provide the same phase shift (although the overall phase shift is 180°). If it were desired to obtain exactly a 60° phase shift for each of three stages, then emitter-follower stages would be needed for each RC section to prevent each from being loaded from the following circuit.

FET Phase-Shift Oscillator

A practical version of a phase-shift oscillator circuit is shown in Fig. 14.21a. The circuit is drawn to show clearly the amplifier and feedback network. The amplifier stage is self-biased with a capacitor bypassed source resistor R_S and a drain bias resistor R_D. The FET device parameters of interest are g_m and r_d. From FET amplifier theory, the amplifier gain magnitude is calculated from

$$|A| = g_m R_L \tag{14.36}$$

where R_L in this case is the parallel resistance of R_D and r_d,

$$R_L = \frac{R_D r_d}{R_D + r_d} \tag{14.37}$$

We shall assume as a very good approximation that the input impedance of the FET amplifier stage is infinite. This assumption is valid as long as the oscillator operating frequency is low enough so that FET capacitive impedances can be neglected. The output impedance of the amplifier stage given by R_L should also be small compared to the impedance seen looking into the feedback network so that no attenuation due to loading occurs. In practice, these considerations are not always negligible, and the amplifier stage gain is then selected somewhat larger than the needed factor of 29 to assure oscillator action.

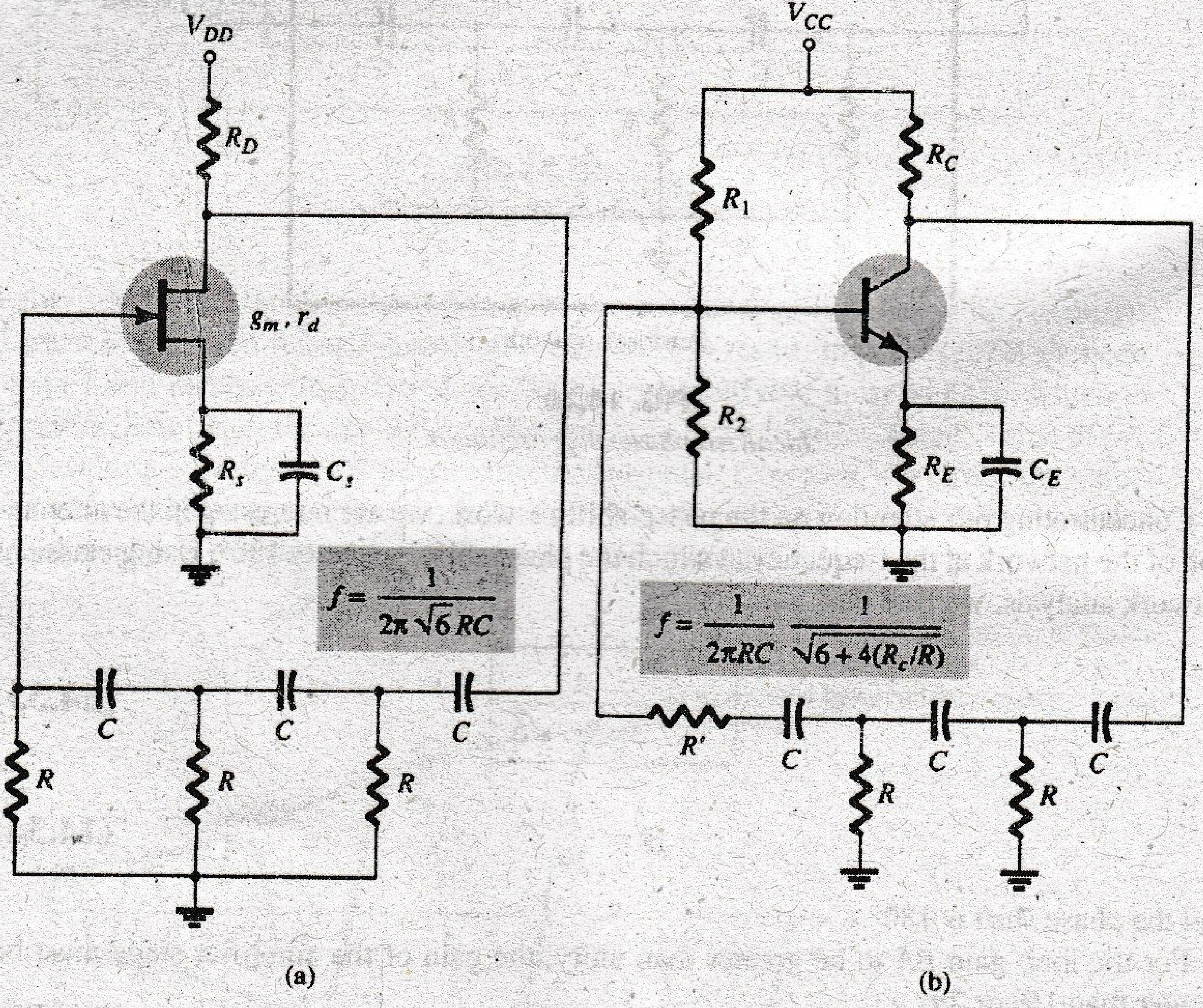


FIG. 14.21
Practical phase-shift oscillator circuits: (a) FET version; (b) BJT version.

EXAMPLE 14.7 It is desired to design a phase-shift oscillator (as in Fig. 14.21a) using an FET having $g_m = 5000\ \mu S$, $r_d = 40\ k\Omega$, and a feedback circuit value of $R = 10\ k\Omega$. Select the value of C for oscillator operation at 1 kHz and R_D for $A > 29$ to ensure oscillator action.

Solution: Equation (14.33) is used to solve for the capacitor value. Since $f = 1/2\pi RC\sqrt{6}$, we can solve for C:

$$C = \frac{1}{2\pi Rf\sqrt{6}} = \frac{1}{(6.28)(10 \times 10^3)(1 \times 10^3)(2.45)} = \mathbf{6.5\ nF}$$

Using Eq. (14.36), we solve for R_L to provide a gain of, say, $A = 40$ (this allows for some loading between R_L and the feedback network input impedance):

$$|A| = g_m R_L$$

$$R_L = \frac{|A|}{g_m} = \frac{40}{5000 \times 10^{-6}} = 8\ \text{k}\Omega$$

Using Eq. (14.37), we solve for $R_D = \mathbf{10\ k\Omega}$.

Transistor Phase-Shift Oscillator

If a transistor is used as the active element of the amplifier stage, the output of the feedback network is loaded appreciably by the relatively low input resistance (h_{ie}) of the transistor. Of course, an emitter-follower input stage followed by a common-emitter amplifier stage could be used. If a single transistor stage is desired, however, the use of voltage-shunt feedback (as shown in Fig. 14.21b) is more suitable. In this connection, the feedback signal is coupled through the feedback resistor R' in *series* with the amplifier stage input resistance (R_i).

Analysis of the ac circuit provides the following equation for the resulting oscillator frequency:

$$f = \frac{1}{2\pi RC}\frac{1}{\sqrt{6 + 4(R_C/R)}} \tag{14.38}$$

For the loop gain to be greater than unity, the requirement on the current gain of the transistor is found to be

$$h_{fe} > 23 + 29\frac{R}{R_C} + 4\frac{R_C}{R} \tag{14.39}$$

IC Phase-Shift Oscillator

As IC circuits have become more popular, they have been adapted to operate in oscillator circuits. One need buy only an op-amp to obtain an amplifier circuit of stabilized gain setting and incorporate some means of signal feedback to produce an oscillator circuit. For example, a phase-shift oscillator is shown in Fig. 14.22. The output of the op-amp is fed to a three-stage RC network, which provides the needed 180° of phase shift (at an attenuation factor of 1/29). If the op-amp provides gain (set by resistors R_i and R_f) of greater than 29,

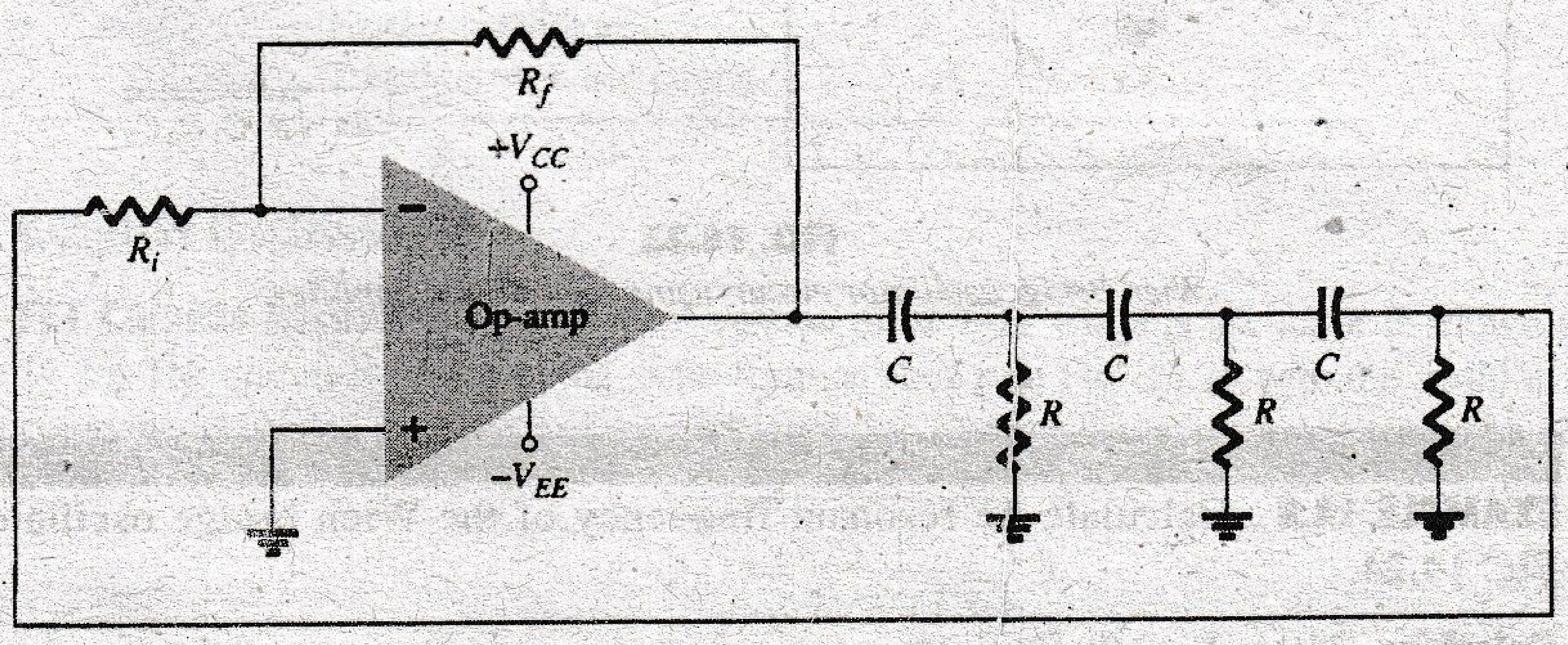


FIG. 14.22
Phase-shift oscillator using an op-amp.

a loop gain greater than unity results and the circuit acts as an oscillator [oscillator frequency is given by Eq. (14.33)].

14.7 WIEN BRIDGE OSCILLATOR

A practical oscillator circuit uses an op-amp and *RC* bridge circuit, with the oscillator frequency set by the *R* and *C* components. Figure 14.23 shows a basic version of a Wien bridge oscillator circuit. Note the basic bridge connection. Resistors R_1 and R_2 and capacitors C_1 and C_2 form the frequency-adjustment elements, and resistors R_3 and R_4 form part of the feedback path. The op-amp output is connected as the bridge input at points *a* and *c*. The bridge circuit output at points *b* and *d* is the input to the op-amp.

Neglecting loading effects of the op-amp input and output impedances, the analysis of the bridge circuit results in

$$\frac{R_3}{R_4} = \frac{R_1}{R_2} + \frac{C_2}{C_1} \tag{14.40}$$

and

$$f_o = \frac{1}{2\pi\sqrt{R_1C_1R_2C_2}} \tag{14.41}$$

If, in particular, the values are $R_1 = R_2 = R$ and $C_1 = C_2 = C$, the resulting oscillator frequency is

$$f_o = \frac{1}{2\pi RC} \tag{14.42}$$

and

$$\frac{R_3}{R_4} = 2 \tag{14.43}$$

Thus a ratio of R_3 to R_4 greater than 2 will provide sufficient loop gain for the circuit to oscillate at the frequency calculated using Eq. (14.42).

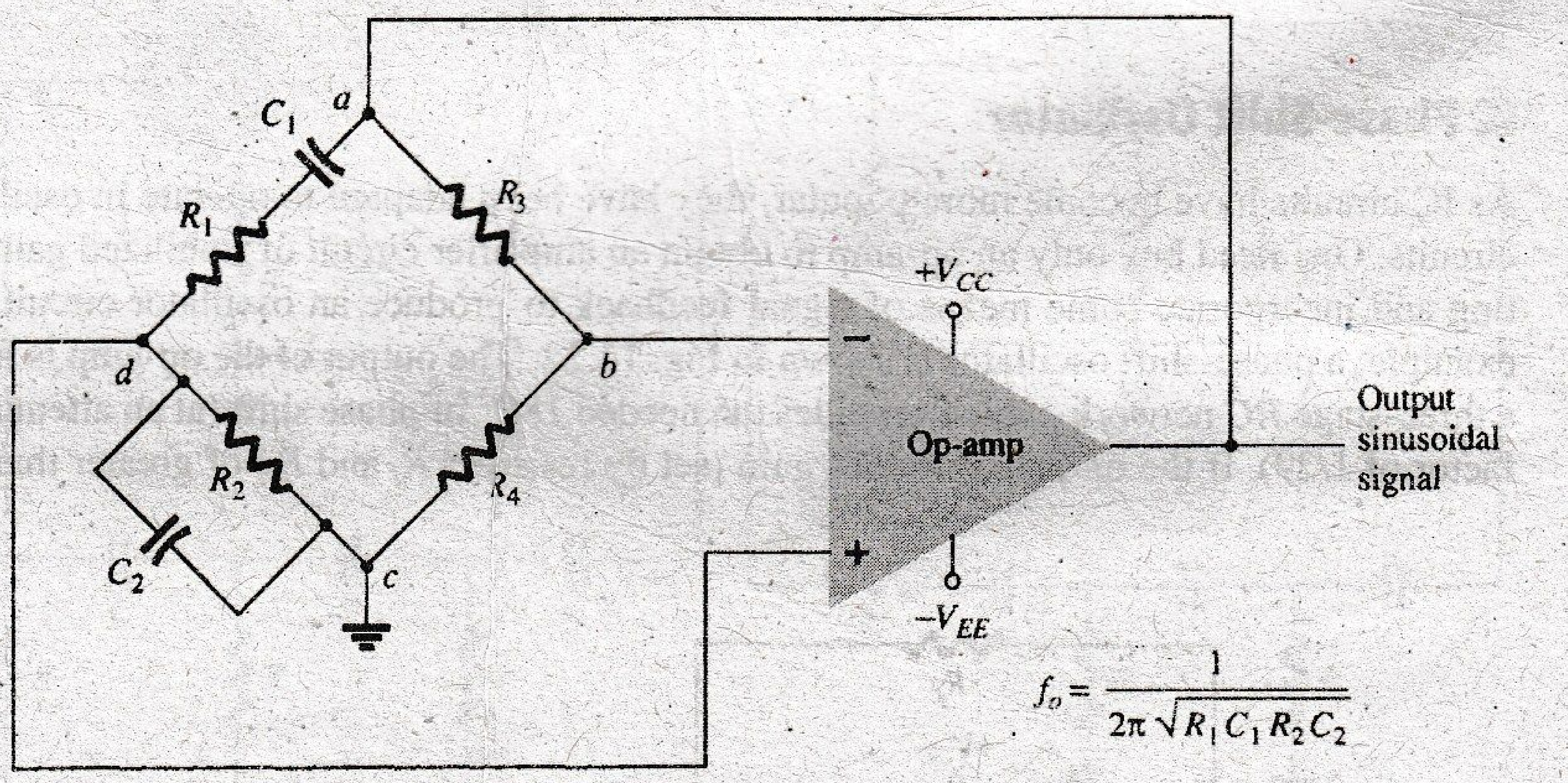


FIG. 14.23
Wien bridge oscillator circuit using an op-amp amplifier.

EXAMPLE 14.8 Calculate the resonant frequency of the Wien bridge oscillator of Fig. 14.24.

Solution: Using Eq. (14.42) yields

$$f_o = \frac{1}{2\pi RC} = \frac{1}{2\pi(51 \times 10^3)(0.001 \times 10^{-6})} = \mathbf{3120.7\ Hz}$$

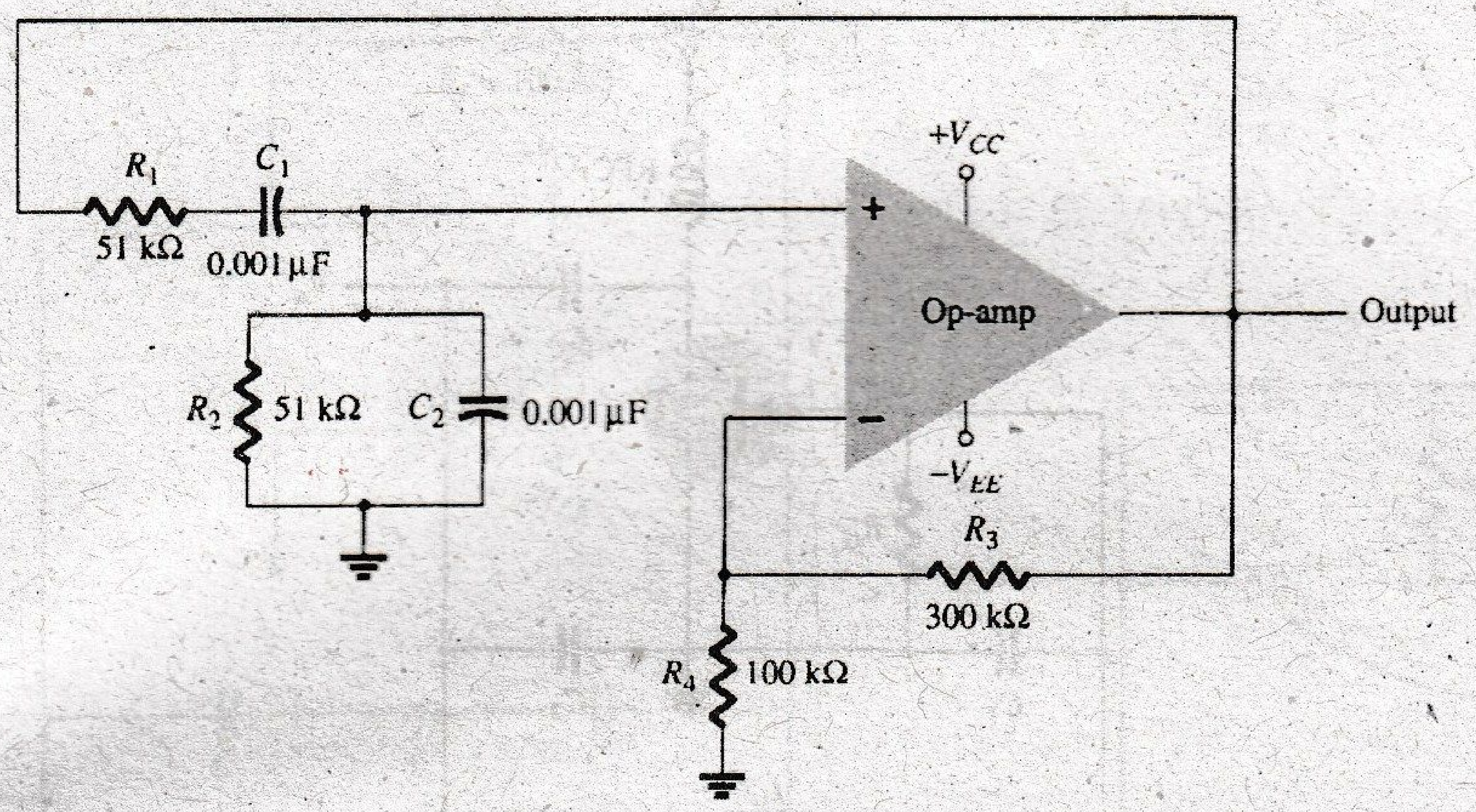


FIG. 14.24
Wien bridge oscillator circuit for Example 14.8.

EXAMPLE 14.9 Design the *RC* elements of a Wien bridge oscillator as in Fig. 14.24 for operation at $f_o = 10$ kHz.

Solution: Using equal values of *R* and *C*, we can select $R = 100\text{ k}\Omega$ and calculate the required value of *C* using Eq. (14.42):

$$C = \frac{1}{2\pi f_o R} = \frac{1}{6.28(10 \times 10^3)(100 \times 10^3)} = \frac{10^{-9}}{6.28} = \mathbf{159\ pF}$$

We can use $R_3 = 300\text{ k}\Omega$ and $R_4 = 100\text{ k}\Omega$ to provide a ratio R_3/R_4 greater than 2 for oscillation to take place.

14.8 TUNED OSCILLATOR CIRCUIT

Tuned-Input, Tuned-Output Oscillator Circuits

A variety of circuits can be built using that shown in Fig. 14.25 by providing tuning in both the input and output sections of the circuit. Analysis of the circuit of Fig. 14.25 reveals that the following types of oscillators are obtained when the reactance elements are as designated:

	Reactance Element		
Oscillator Type	X_1	X_2	X_3
Colpitts oscillator	*C*	*C*	*L*
Hartley oscillator	*L*	*L*	*C*
Tuned input, tuned output	*LC*	*LC*	—

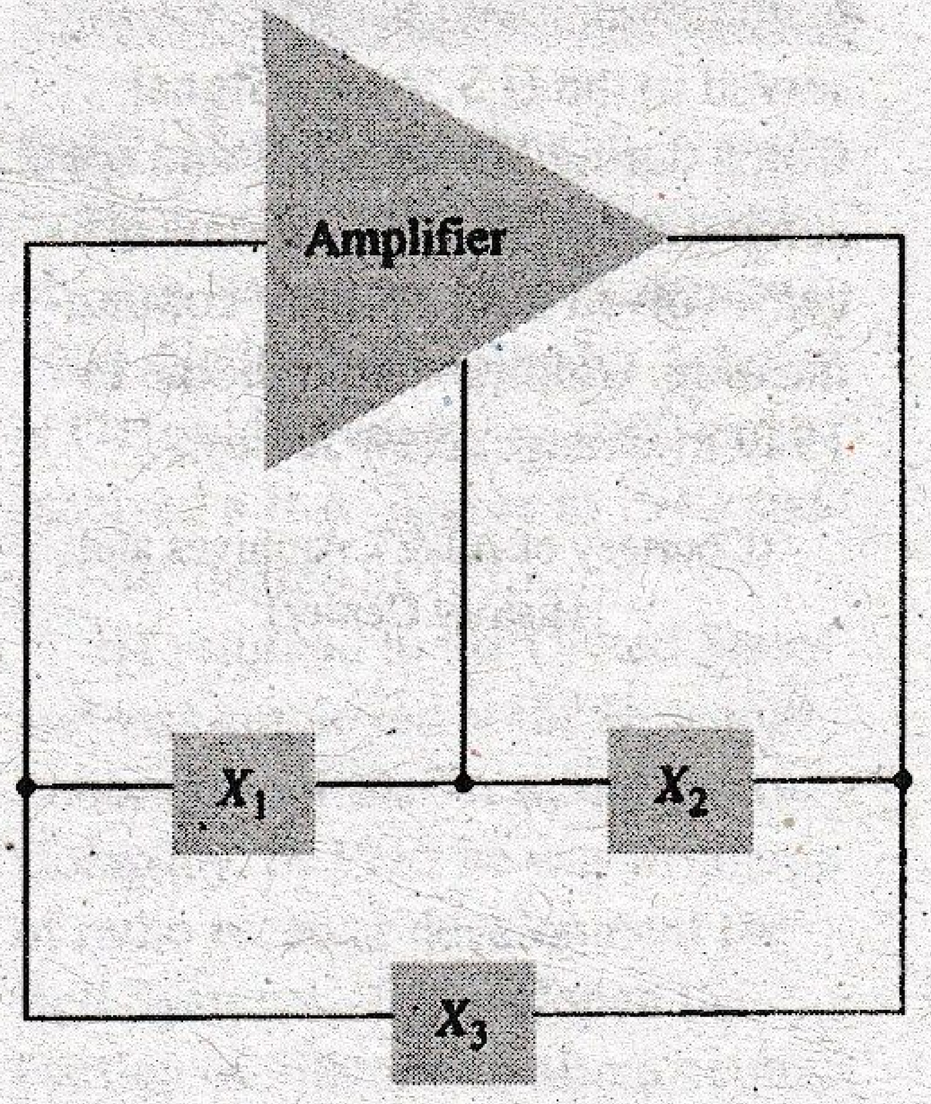


FIG. 14.25
Basic configuration of resonant circuit oscillator.

Colpitts Oscillator

FET Colpitts Oscillator A practical version of an FET Colpitts oscillator is shown in Fig. 14.26. The circuit is basically the same form as shown in Fig. 14.25 with the addition of the components needed for dc bias of the FET amplifier. The oscillator frequency can be found to be

$$f_o = \frac{1}{2\pi\sqrt{LC_{eq}}} \qquad \textbf{(14.44)}$$

where

$$C_{eq} = \frac{C_1C_2}{C_1 + C_2} \qquad \textbf{(14.45)}$$

Edwin Henry Colpitts (1872–1949) was a communications pioneer best known for his invention of the Colpitts oscillator. In 1915, his Western Electric team successfully demonstrated the first transatlantic radio telephone. In 1895 he entered Harvard University where he studied physics and mathematics. He received a B.A. in 1896 and a master's degree in 1897 from that institution. In 1899, Colpitts accepted a position with American Bell Telephone Company. He moved to Western Electric in 1907. His colleague Ralph Hartley invented an inductive coupling oscillator, which Colpitts improved in 1915. Colpitts served in the U.S. Army Signal Corps during World War I and spent some time in France as a staff officer involved with military communication. Colpitts died at home in 1949 in Orange, New Jersey.

(Courtesy of AT&T Archives and History Center)

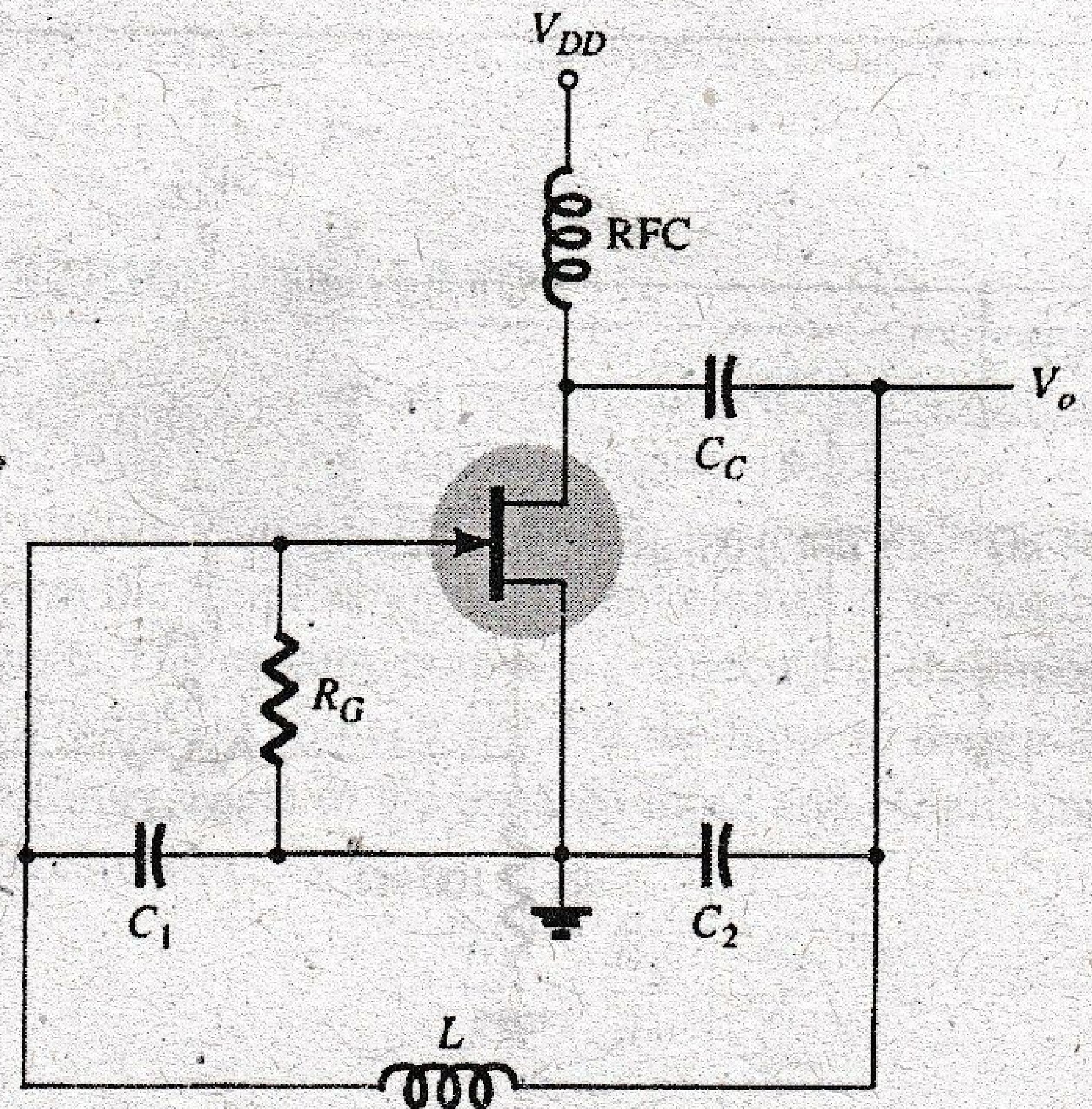


FIG. 14.26
FET Colpitts oscillator.

Transistor Colpitts Oscillator A transistor Colpitts oscillator circuit can be made as shown in Fig. 14.27. The circuit frequency of oscillation is given by Eq. (14.44).

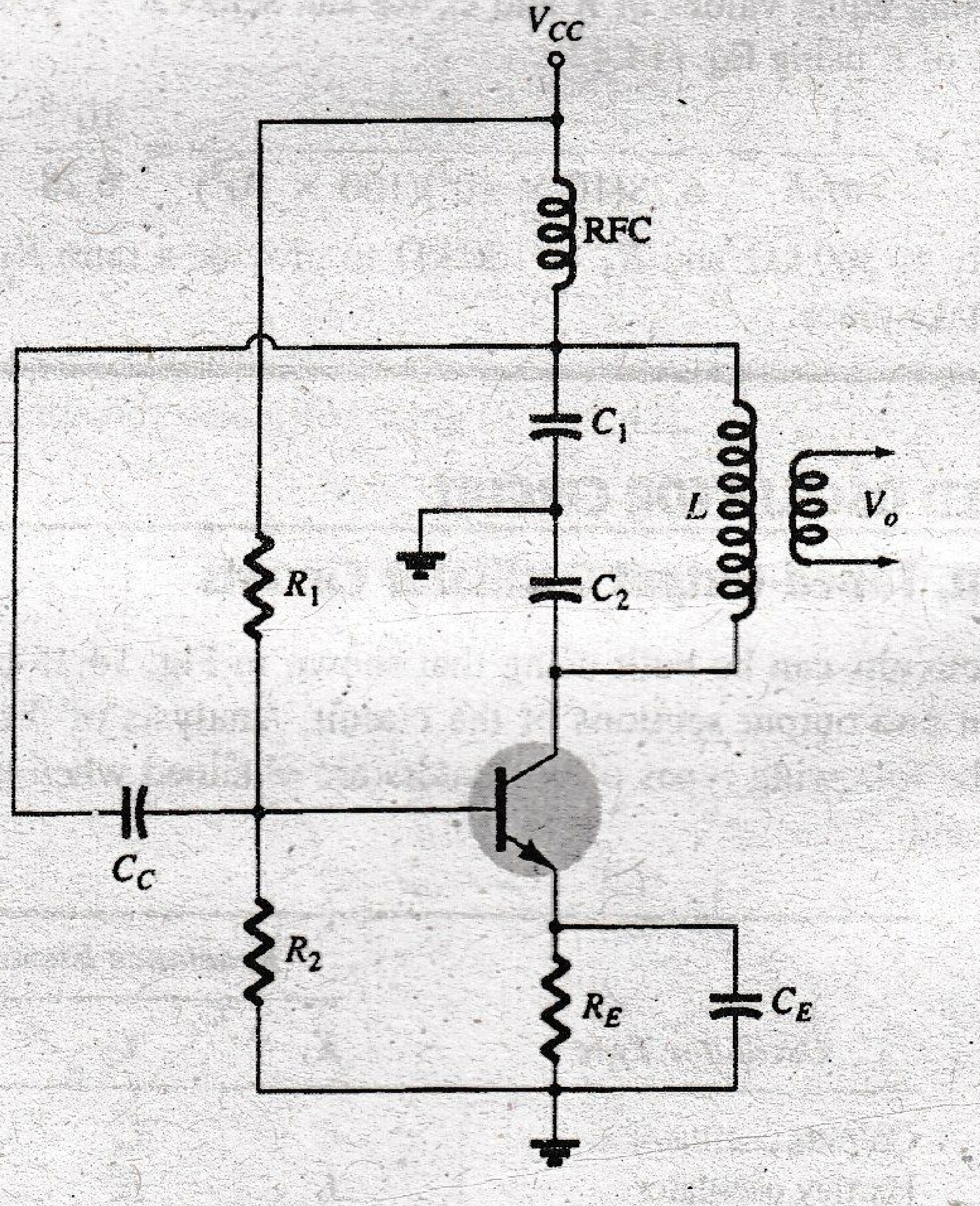


FIG. 14.27
Transistor Colpitts oscillator.

IC Colpitts Oscillator An op-amp Colpitts oscillator circuit is shown in Fig. 14.28. Again, the op-amp provides the basic amplification needed, and the oscillator frequency is set by an *LC* feedback network of a Colpitts configuration. The oscillator frequency is given by Eq. (14.44).

Hartley Oscillator

If the elements in the basic resonant circuit of Fig. 14.25 are X_1 and X_2 (inductors) and X_3 (capacitor), the circuit is a Hartley oscillator.

FET Hartley Oscillator An FET Hartley oscillator circuit is shown in Fig. 14.29. The circuit is drawn so that the feedback network conforms to the form shown in the basic resonant circuit (Fig. 14.25). Note, however, that inductors L_1 and L_2 have a mutual coupling M,

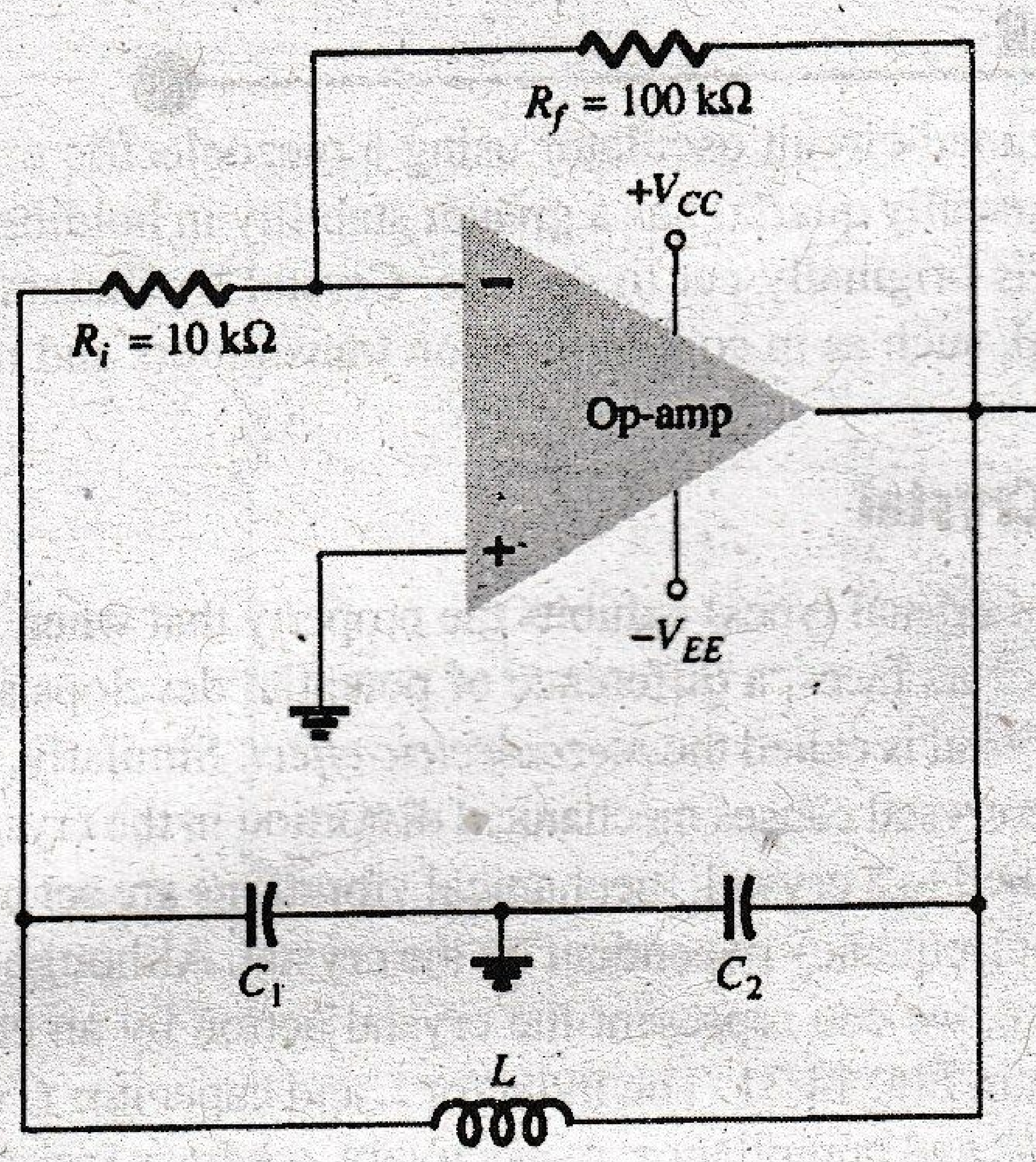


FIG. 14.28
Op-amp Colpitts oscillator.

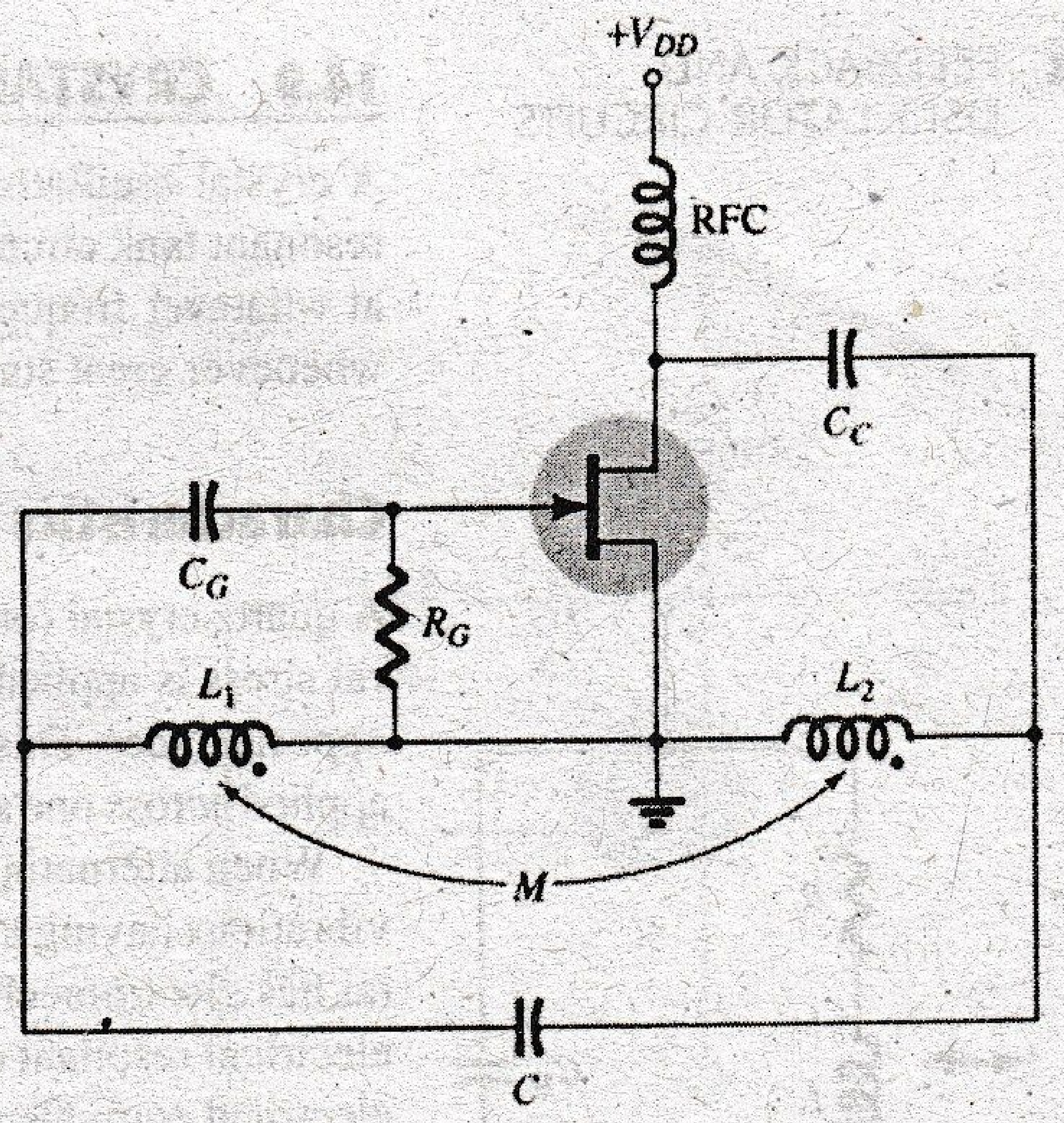


FIG. 14.29
FET Hartley oscillator.

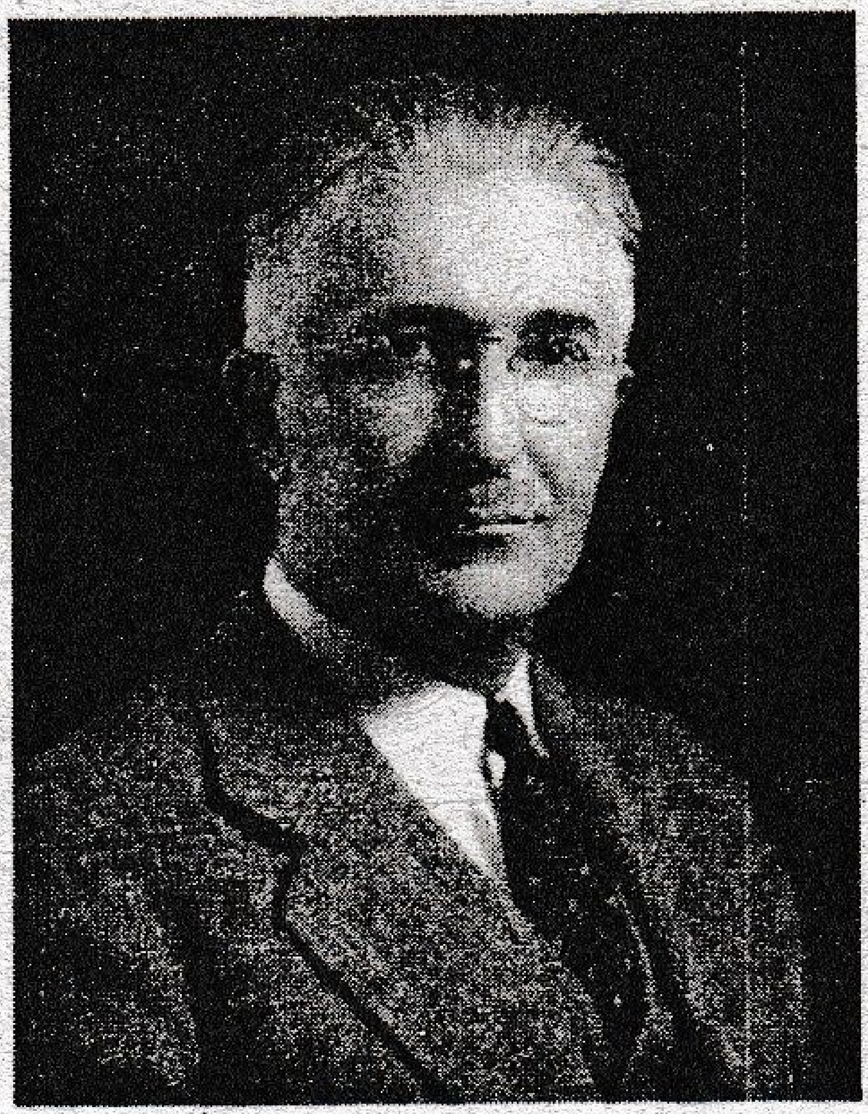
which must be taken into account in determining the equivalent inductance for the resonant tank circuit. The circuit frequency of oscillation is then given approximately by

$$f_o = \frac{1}{2\pi\sqrt{L_{eq}C}} \tag{14.46}$$

with

$$\boxed{L_{eq} = L_1 + L_2 + 2M} \tag{14.47}$$

Transistor Hartley Oscillator Figure 14.30 shows a transistor Hartley oscillator circuit. The circuit operates at a frequency given by Eq. (14.46).



Ralph Hartley was born in Nevada in 1888 and attended the University of Utah, receiving an A.B. degree in 1909. He became a Rhodes Scholar at Oxford University in 1910 and received a B.A. degree in 1912 and a B.Sc. degree in 1913.

He returned to the United States and was employed at the Research Laboratory of the Western Electric Company. In 1915 he was in charge of radio receiver development for Bell Systems. He developed the Hartley oscillator and also a neutralizing circuit to eliminate triode singing resulting from internal coupling. During World War I he established the principles that led to sound-type directional finders. He retired from Bell Labs in 1950 and died on May 1, 1970.

(Courtesy of AT&T Archives and History Center)

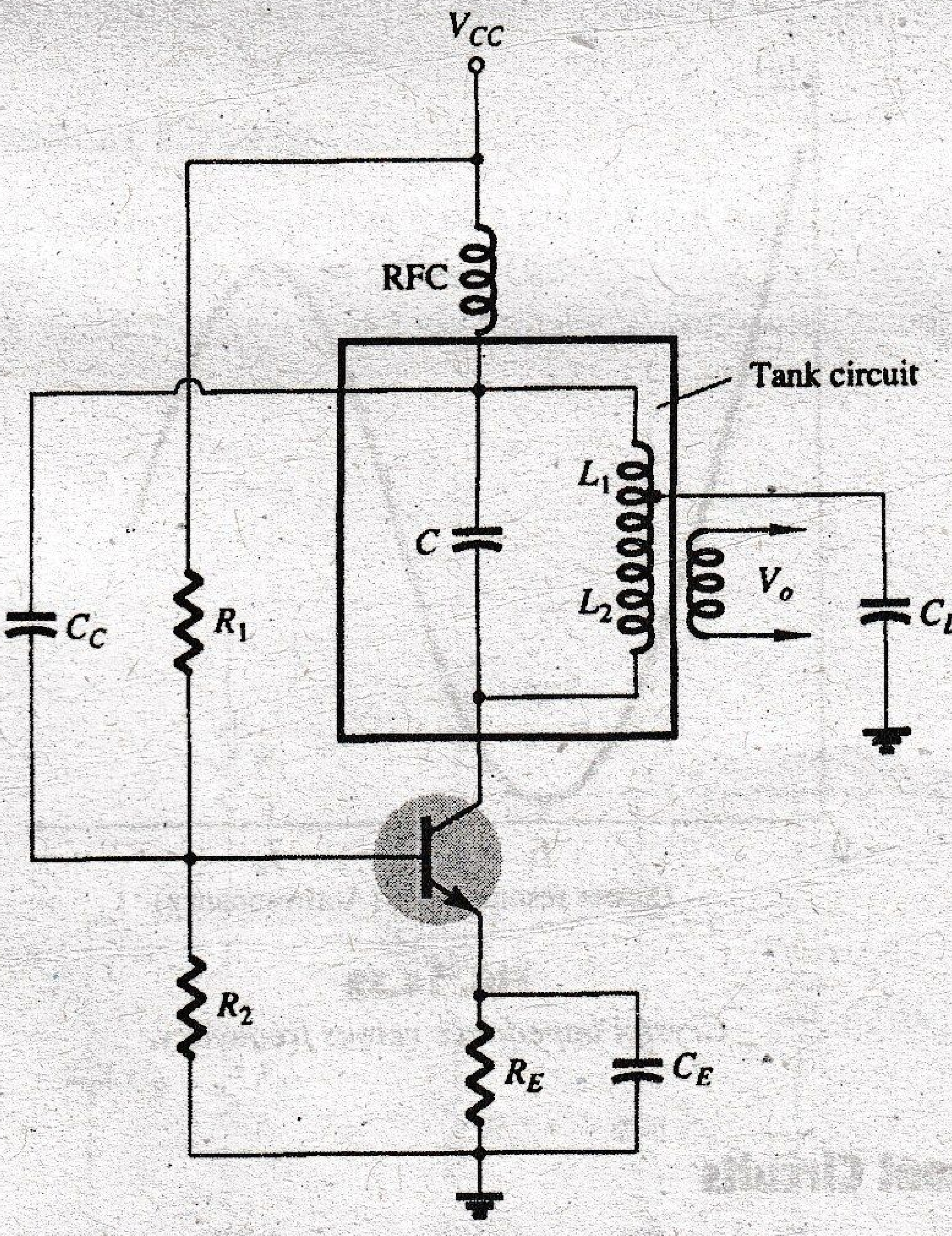


FIG. 14.30
Transistor Hartley oscillator circuit.

14.9 CRYSTAL OSCILLATOR

A crystal oscillator is basically a tuned-circuit oscillator using a piezoelectric crystal as a resonant tank circuit. The crystal (usually quartz) has a greater stability in holding constant at whatever frequency the crystal is originally cut to operate. Crystal oscillators are used whenever great stability is required, such as in communication transmitters and receivers.

Characteristics of a Quartz Crystal

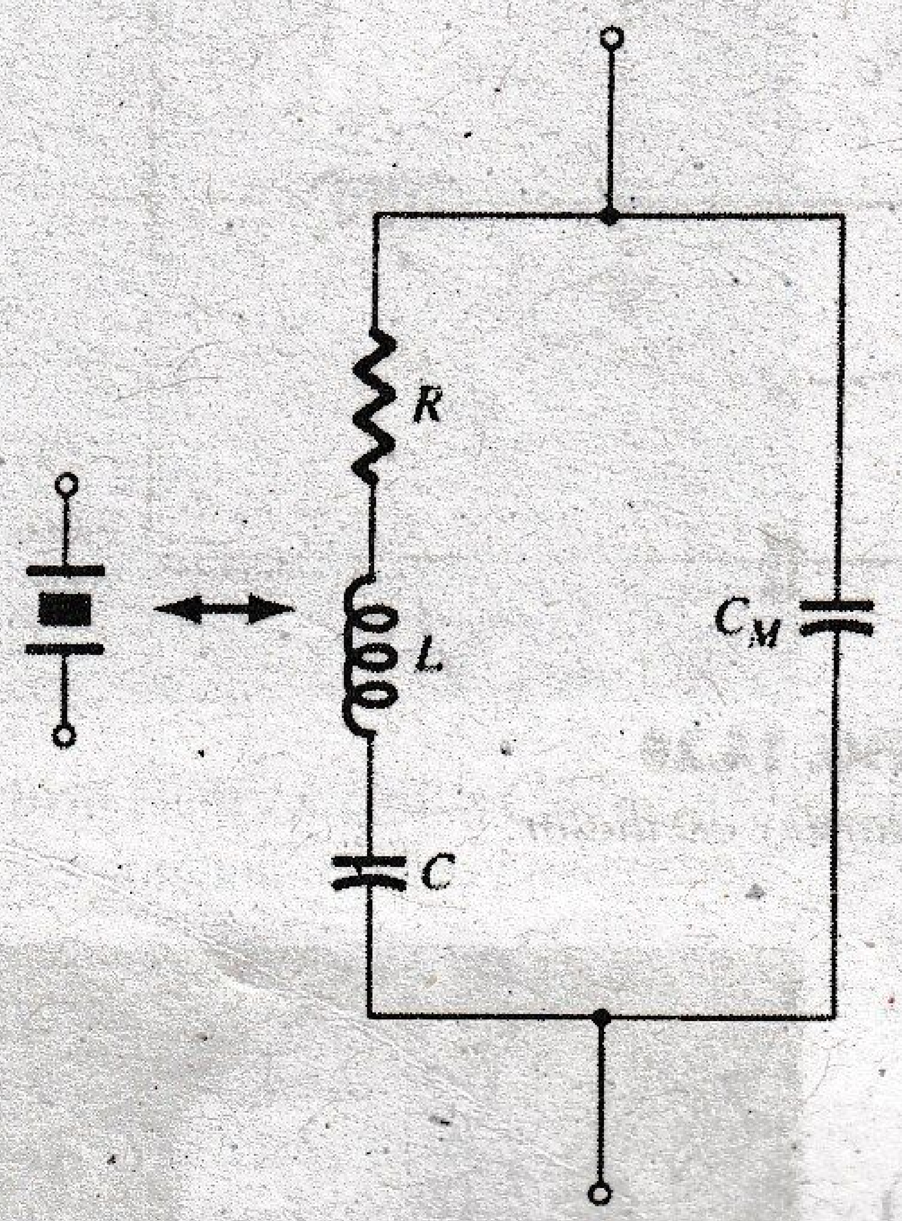


FIG. 14.31
Electrical equivalent circuit of a crystal.

A quartz crystal (one of a number of crystal types) exhibits the property that when mechanical stress is applied across one set of its faces, a difference of potential develops across the opposite faces. This property of a crystal is called the *piezoelectric effect.* Similarly, a voltage applied across one set of faces of the crystal causes mechanical distortion in the crystal shape.

When alternating voltage is applied to a crystal, mechanical vibrations are set up—these vibrations having a natural resonant frequency dependent on the crystal. Although the crystal has electromechanical resonance, we can represent the crystal action by an equivalent electrical resonant circuit as shown in Fig. 14.31. The inductor L and capacitor C represent electrical equivalents of crystal mass and compliance, respectively, whereas resistance R is an electrical equivalent of the crystal structure's internal friction. The shunt capacitance C_M represents the capacitance due to mechanical mounting of the crystal. Because the crystal losses, represented by R, are small, the equivalent crystal Q (quality factor) is high—typically 20,000. Values of Q up to almost 10^6 can be achieved by using crystals.

The crystal as represented by the equivalent electrical circuit of Fig. 14.31 can have two resonant frequencies. One resonant condition occurs when the reactances of the series RLC leg are equal (and opposite). For this condition, the *series-resonant* impedance is very low (equal to R). The other resonant condition occurs at a higher frequency when the reactance of the series-resonant leg equals the reactance of capacitor C_M. This is a parallel resonance or antiresonance condition of the crystal. At this frequency, the crystal offers a very high impedance to the external circuit. The impedance versus frequency of the crystal is shown in Fig. 14.32. To use the crystal properly, it must be connected in a circuit so that its low impedance in the series-resonant operating mode or high impedance in the antiresonant operating mode is selected.

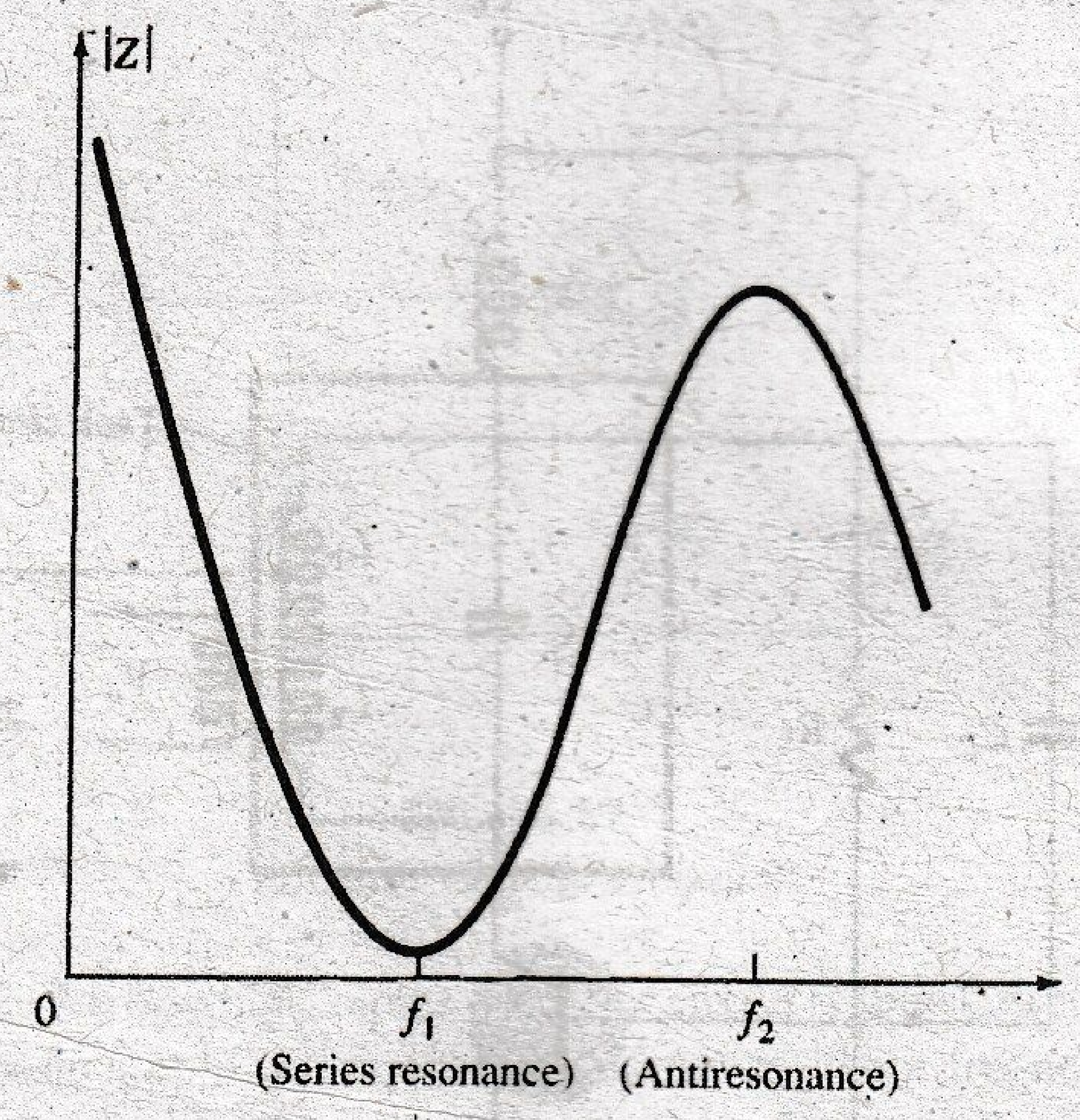


FIG. 14.32
Crystal impedance versus frequency.

Series-Resonant Circuits

To excite a crystal for operation in the series-resonant mode, it may be connected as a series element in a feedback path. At the series-resonant frequency of the crystal, its impedance is smallest and the amount of (positive) feedback is largest. A typical transistor circuit is

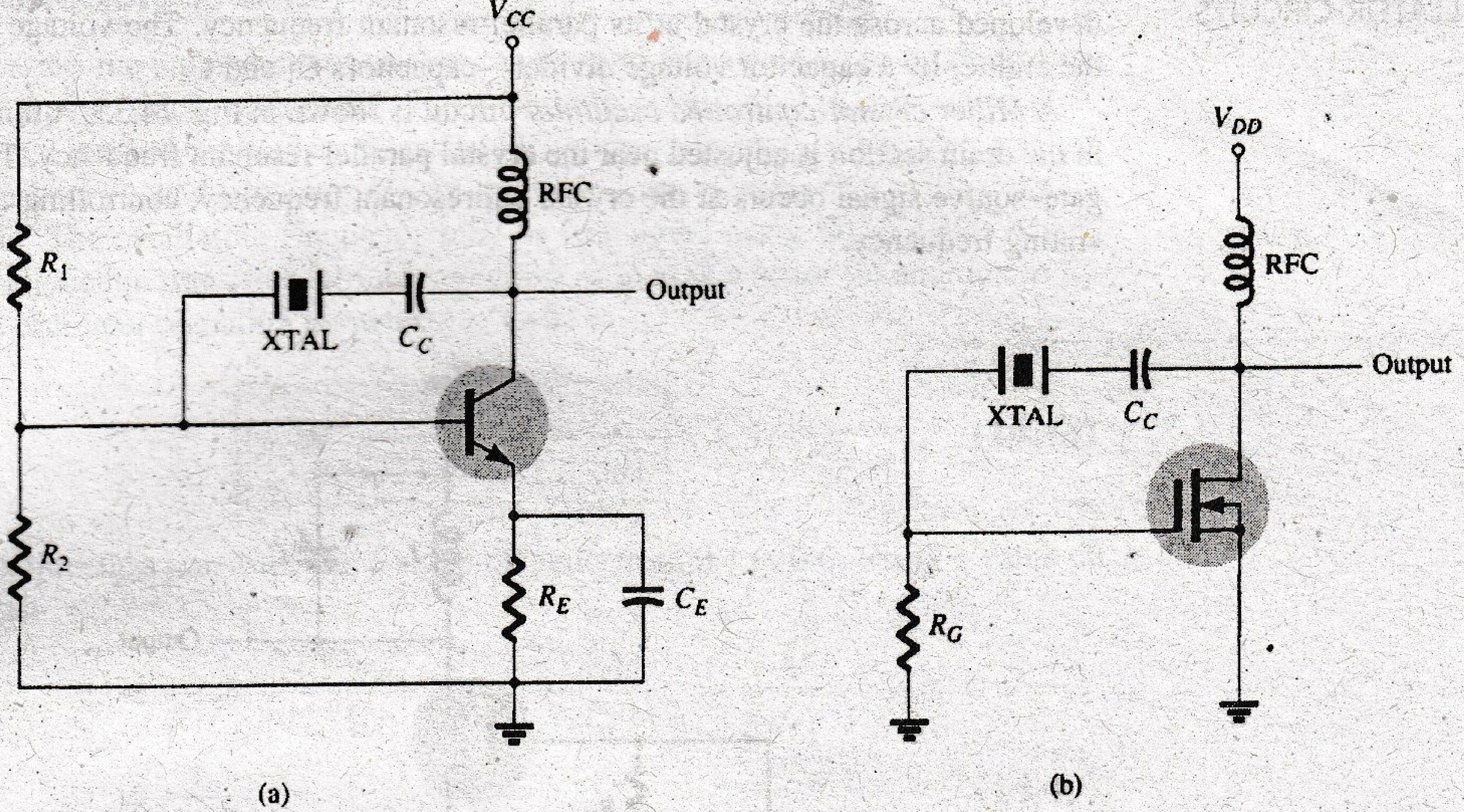


FIG. 14.33
Crystal-controlled oscillator using a crystal (XTAL) in a series-feedback path: (a) BJT circuit; (b) FET circuit.

shown in Fig. 14.33. Resistors R_1, R_2, and R_E provide a voltage-divider stabilized dc bias circuit. Capacitor C_E provides ac bypass of the emitter resistor, and the RFC coil provides for dc bias while decoupling any ac signal on the power lines from affecting the output signal. The voltage feedback from collector to base is a maximum when the crystal impedance is minimum (in series-resonant mode). The coupling capacitor C_C has negligible impedance at the circuit operating frequency but blocks any dc between collector and base.

The resulting circuit frequency of oscillation is set, then, by the series-resonant frequency of the crystal. Changes in supply voltage, transistor device parameters, and so on, have no effect on the circuit operating frequency, which is held stabilized by the crystal. The circuit frequency stability is set by the crystal frequency stability, which is good.

Parallel-Resonant Circuits

Since the parallel-resonant impedance of a crystal is a maximum value, it is connected in shunt. At the parallel-resonant operating frequency, a crystal appears as an inductive reactance of largest value. Figure 14.34 shows a crystal connected as the inductor element in a

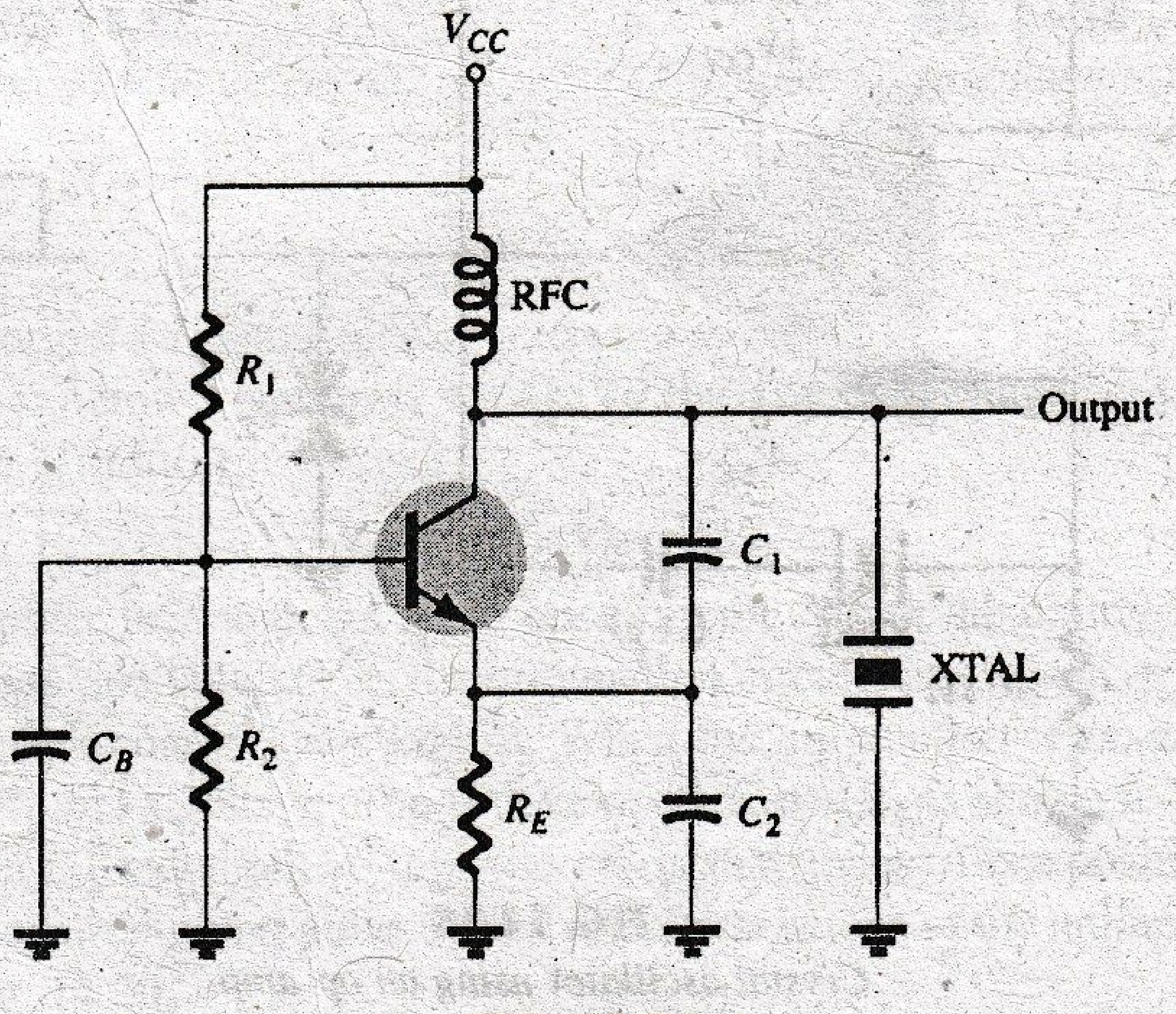


FIG. 14.34
Crystal-controlled oscillator operating in parallel-resonant mode.

modified Colpitts circuit. The basic dc bias circuit should be evident. Maximum voltage is developed across the crystal at its parallel-resonant frequency. The voltage is coupled to the emitter by a capacitor voltage divider—capacitors C_1 and C_2.

A *Miller crystal-controlled oscillator* circuit is shown in Fig. 14.35. A tuned *LC* circuit in the drain section is adjusted near the crystal parallel-resonant frequency. The maximum gate–source signal occurs at the crystal antiresonant frequency, controlling the circuit operating frequency.

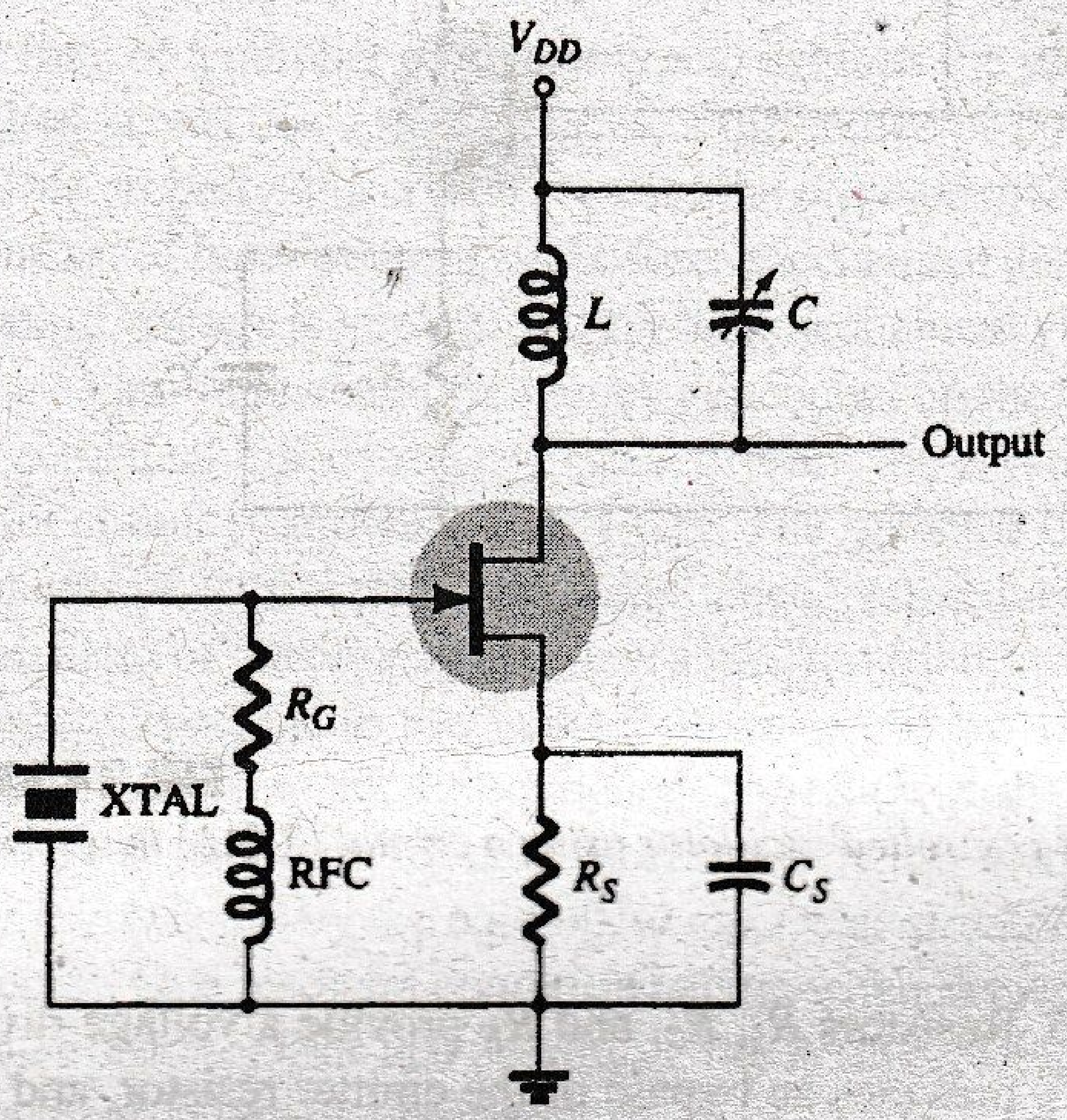


FIG. 14.35
Miller crystal-controlled oscillator.

Crystal Oscillator

An op-amp can be used in a crystal oscillator as shown in Fig. 14.36. The crystal is connected in the series-resonant path and operates at the crystal series-resonant frequency. The present circuit has a high gain, so that an output square-wave signal results as shown in the figure. A pair of Zener diodes is shown at the output to provide output amplitude at exactly the Zener voltage (V_Z).

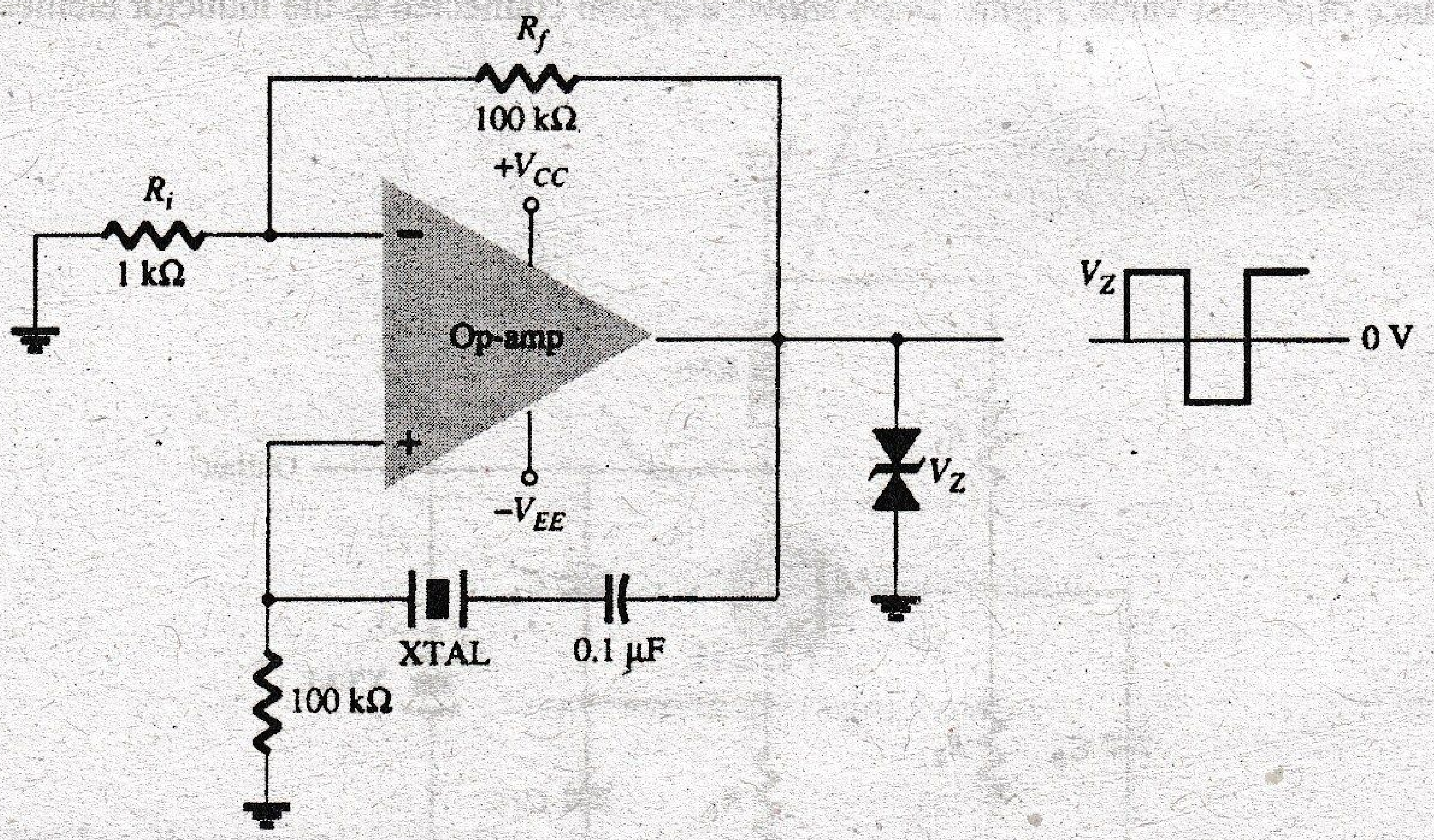


FIG. 14.36
Crystal oscillator using an op-amp.

14.10 UNIJUNCTION OSCILLATOR

A particular device, the unijunction transistor, can be used in a single-stage oscillator circuit to provide a pulse signal suitable for digital-circuit applications. The unijunction transistor can be used in what is called a *relaxation oscillator* as shown by the basic circuit of Fig. 14.37. Resistor R_T and capacitor C_T are the timing components that set the circuit oscillating rate. The oscillating frequency may be calculated using Eq. (14.48), which includes the unijunction transistor *intrinsic stand-off ratio* η as a factor (in addition to R_T and C_T) in the oscillator operating frequency:

$$f_o \cong \frac{1}{R_T C_T \ln[1/(1-\eta)]} \qquad (14.48)$$

Typically, a unijunction transistor has a stand-off ratio from 0.4 to 0.6. Using a value of $\eta = 0.5$, we get

$$f_o \cong \frac{1}{R_T C_T \ln[1/(1-0.5)]} = \frac{1.44}{R_T C_T \ln 2} = \frac{1.44}{R_T C_T}$$
$$\cong \frac{1.5}{R_T C_T} \qquad (14.49)$$

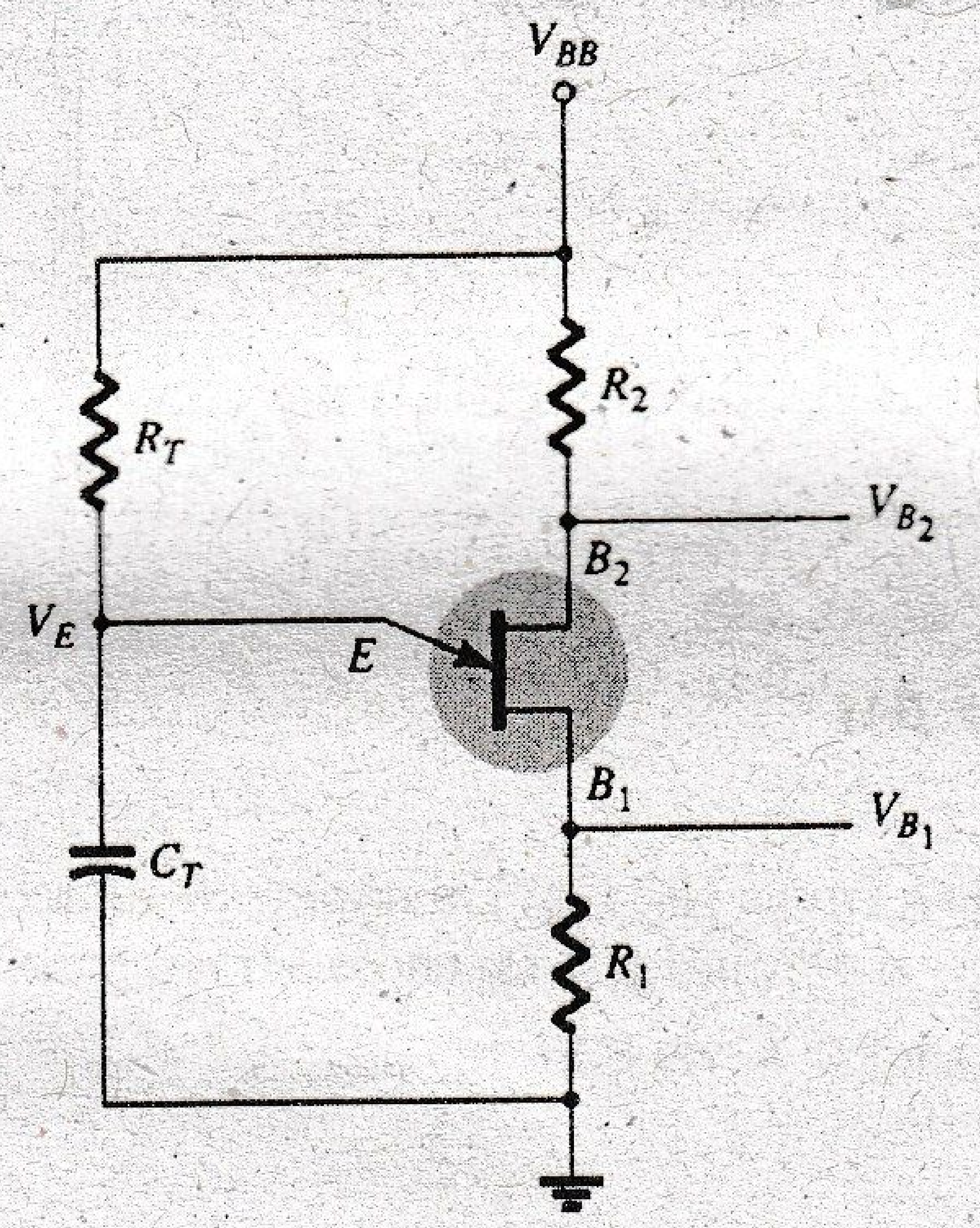


FIG. 14.37
Basic unijunction oscillator circuit.

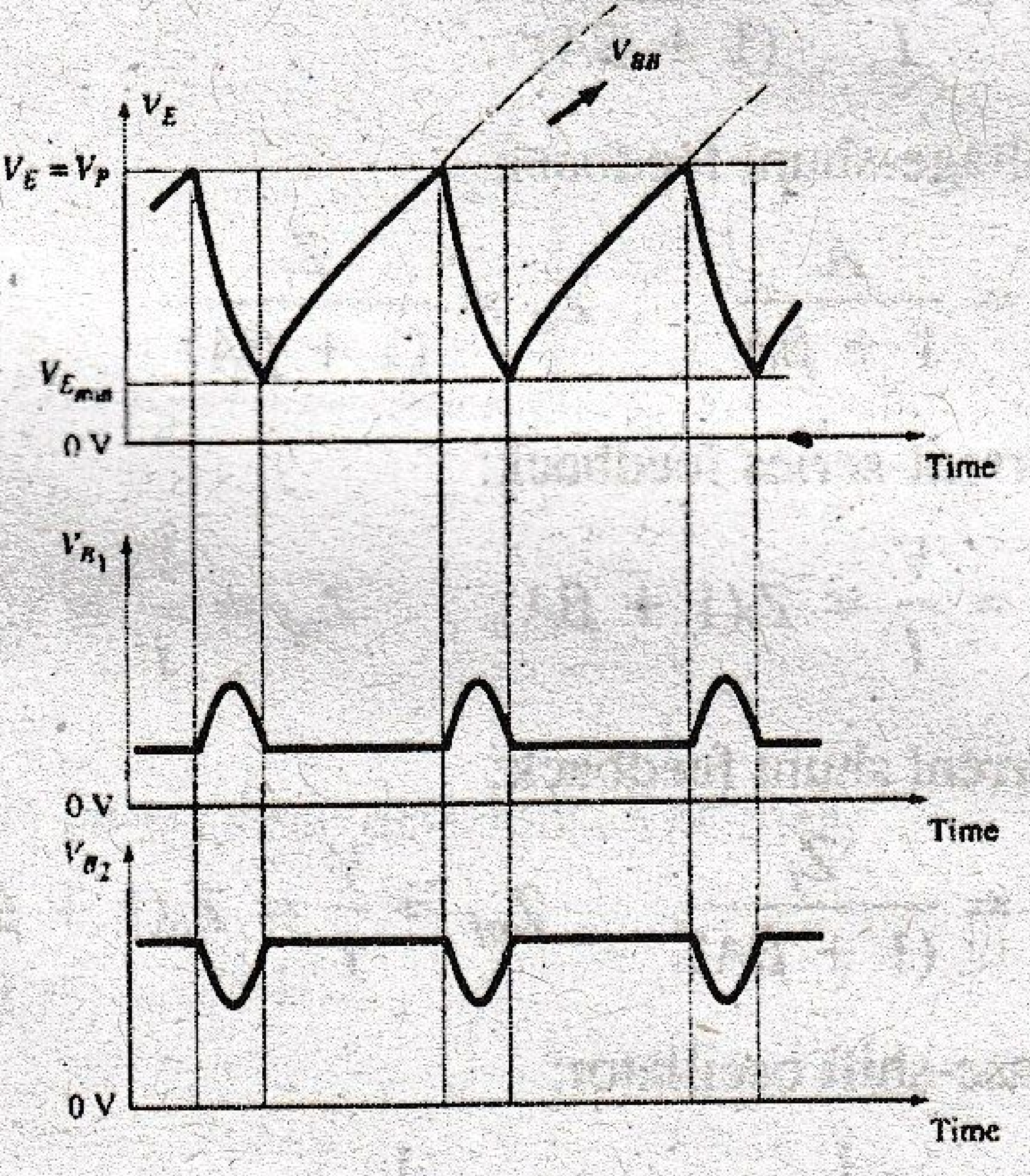


FIG. 14.38
Unijunction oscillator waveforms.

Capacitor C_T is charged through resistor R_T toward supply voltage V_{BB}. As long as the capacitor voltage V_E is below a stand-off voltage (V_P) set by the voltage across $B_1 - B_2$ and the transistor stand-off ratio η,

$$V_P = \eta V_{B_1} V_{B_2} - V_D \qquad (14.50)$$

the unijunction emitter lead appears as an open circuit. When the emitter voltage across capacitor C_T exceeds this value (V_P), the unijunction circuit fires, discharging the capacitor, after which a new charge cycle begins. When the unijunction fires, a voltage rise is developed across R_1 and a voltage drop is developed across R_2 as shown in Fig. 14.38. The signal at the emitter is a sawtooth voltage waveform that at base 1 is a positive-going pulse and at base 2 is a negative-going pulse. A few circuit variations of the unijunction oscillator are provided in Fig. 14.39.

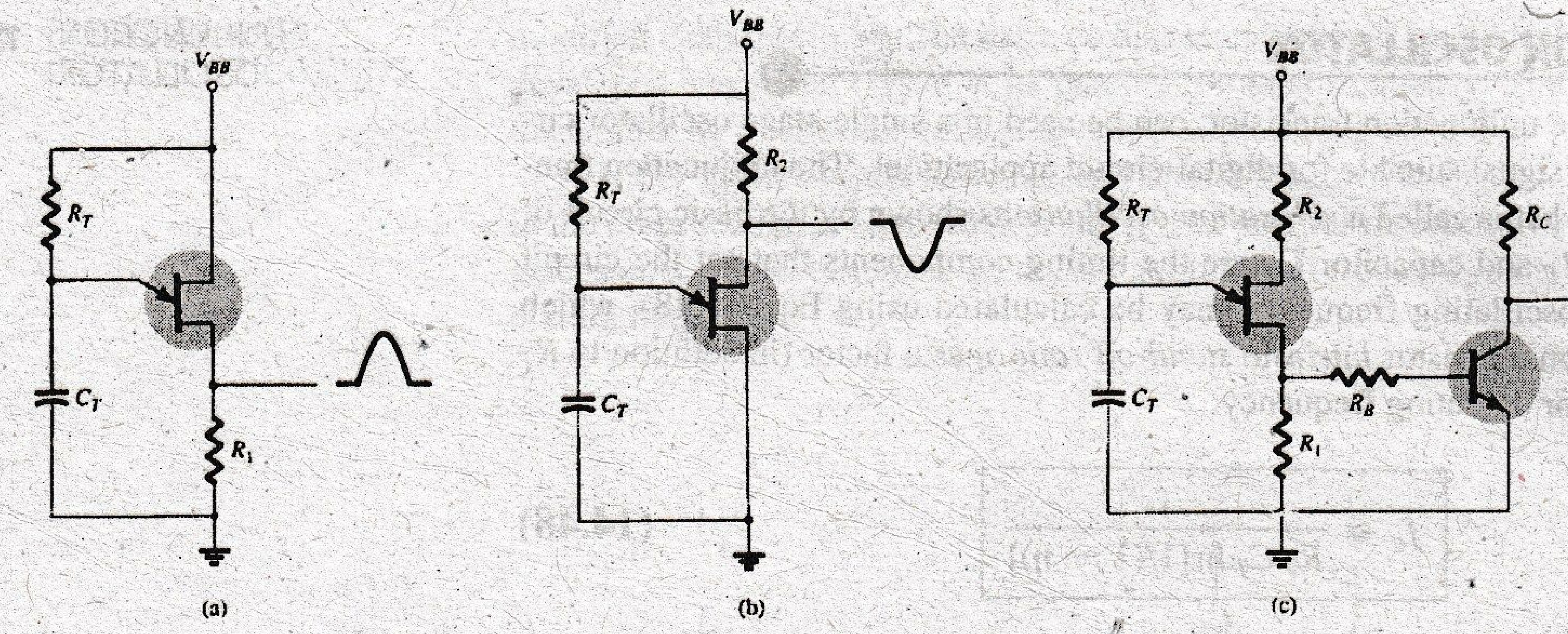


FIG. 14.39
Some unijunction oscillator circuit configurations.

14.11 SUMMARY

Equations

Voltage-series feedback:

$$A_f = \frac{V_o}{V_s} = \frac{A}{1 + \beta A}, \qquad Z_{if} = \frac{V_s}{I_i} = Z_i + (\beta A)Z_i = Z_i(1 + \beta A),$$

$$Z_{of} = \frac{V}{I} = \frac{Z_o}{(1 + \beta A)}$$

Voltage-shunt feedback:

$$A_f = \frac{A}{1 + \beta A}, \qquad Z_{if} = \frac{Z_i}{(1 + \beta A)}$$

Current-series feedback:

$$Z_{if} = \frac{V}{I} = Z_i(1 + \beta A), \qquad Z_{of} = \frac{V}{I} = Z_o(1 + \beta A)$$

Current shunt feedback:

$$Z_{if} = \frac{Z_i}{(1 + \beta A)}, \qquad Z_{of} = \frac{V}{I} = Z_o(1 + \beta A)$$

Phase-shift oscillator:

$$f = \frac{1}{2\pi RC\sqrt{6}}, \qquad \beta = \frac{1}{29}$$

Wien bridge oscillator:

$$f_o = \frac{1}{2\pi\sqrt{R_1C_1R_2C_2}}$$

Colpitts oscillator:

$$f_o = \frac{1}{2\pi\sqrt{LC_{eq}}} \quad \text{where} \quad C_{eq} = \frac{C_1C_2}{C_1 + C_2}$$

Hartley oscillator:

$$f_o = \frac{1}{2\pi\sqrt{L_{eq}C}} \quad \text{where} \quad L_{eq} = L_1 + L_2 + 2M$$

Unijunction oscillator:

$$f_o \cong \frac{1}{R_TC_T\ln[1/(1 - \eta)]}$$

14.12 COMPUTER ANALYSIS

Multisim

Example 14.10—IC Phase-Shift Oscillator Using Multisim, we draw a phase-shift oscillator as shown in Fig. 14.40. The diode network helps the circuit go into self-oscillation, with the output frequency calculated using

$$f_o = 1/(2\pi\sqrt{6}RC)$$
$$= 1/[2\pi\sqrt{6}(20 \times 10^3)(0.001 \times 10^{-6})] = 3{,}248.7 \text{ Hz}$$

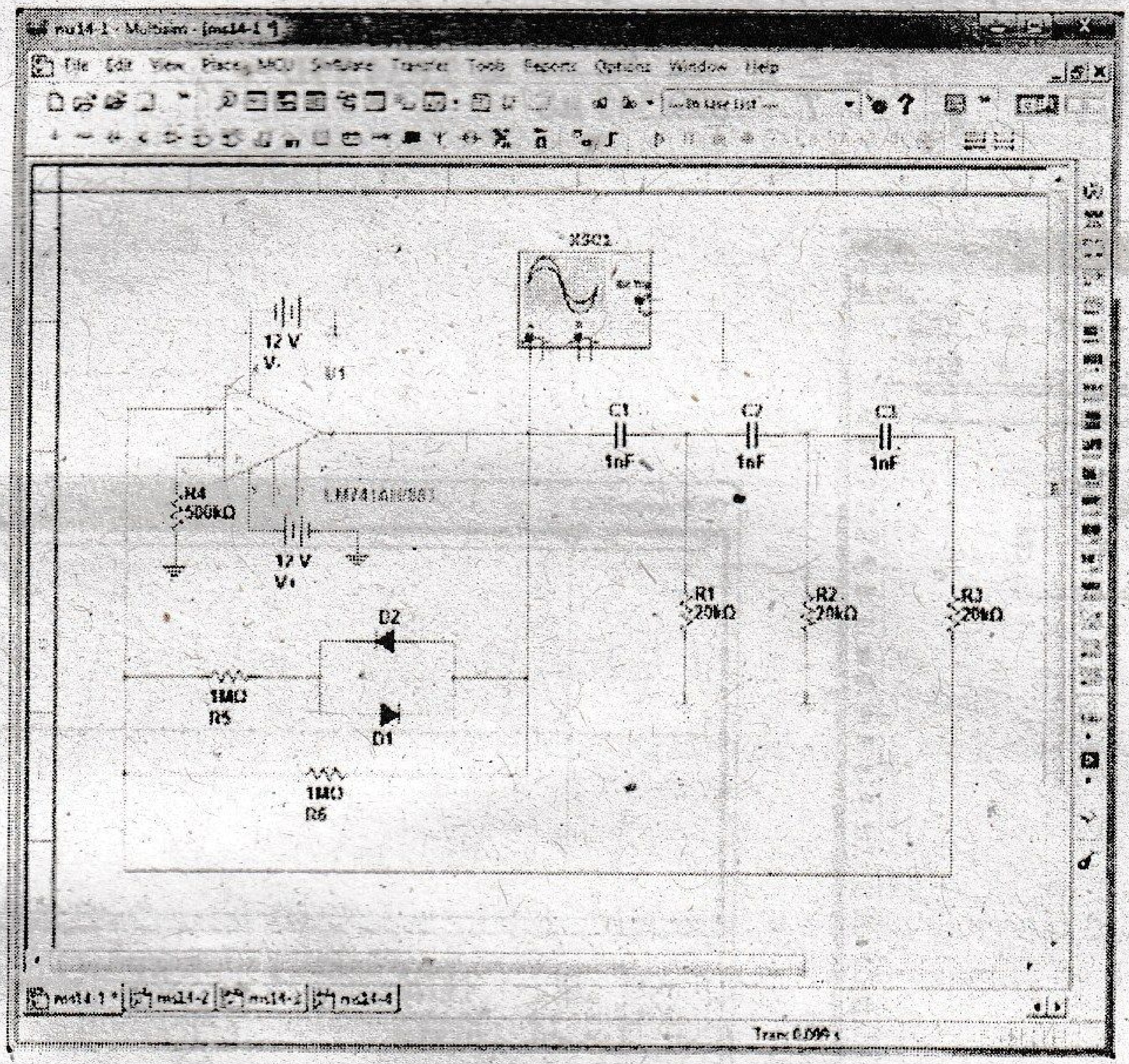


FIG. 14.40
Phase-shift oscillator using Multisim.

The oscilloscope waveform in Fig. 14.41 shows a cycle in about three divisions. The measured frequency for the scope set at 0.1 ms/div is

$$f_{\text{measured}} = 1/(3 \text{ div} \times 0.1 \text{ ms/div}) = 3{,}333 \text{ Hz}$$

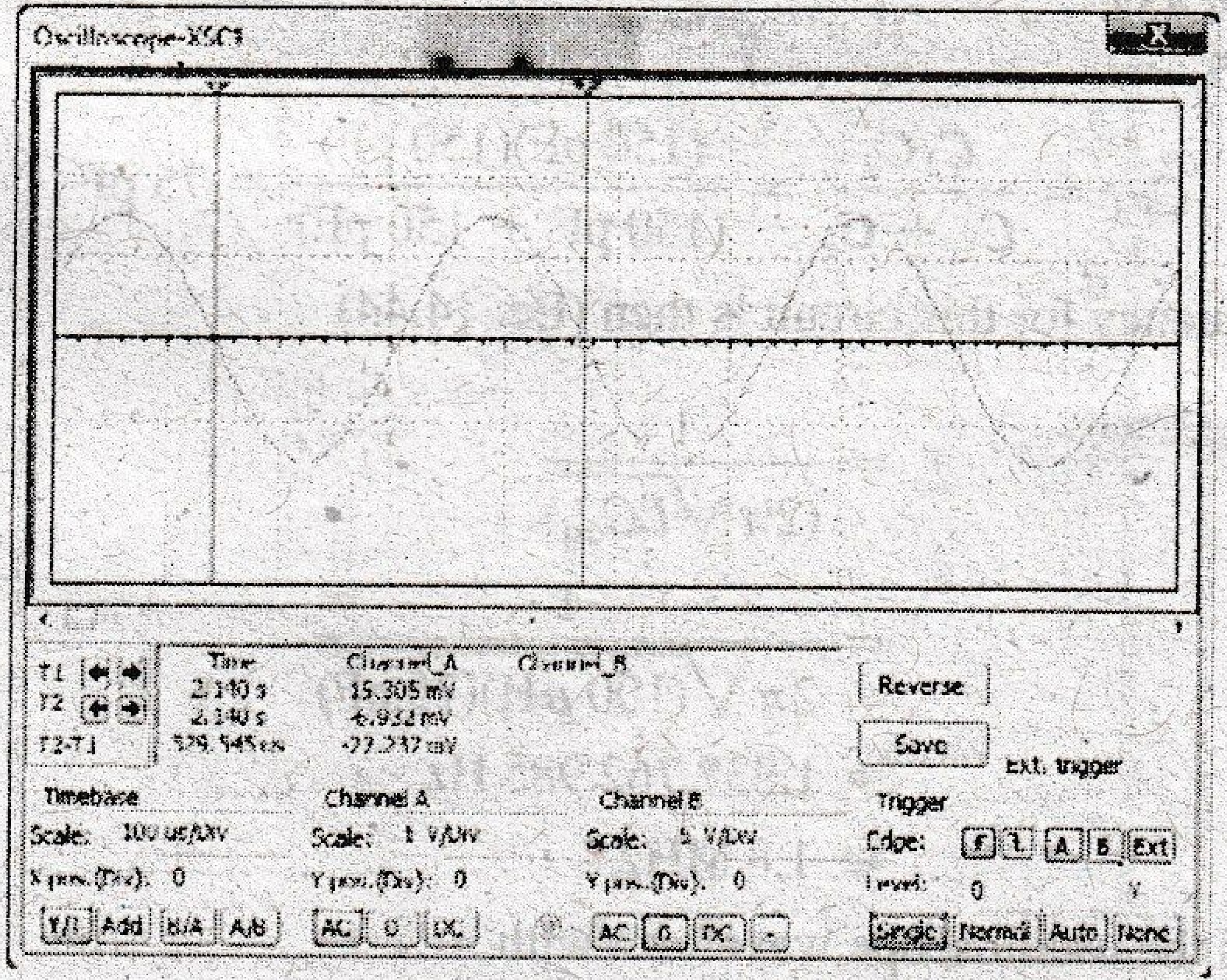


FIG. 14.41
Oscilloscope waveform.

Example 14.11—IC Wien Bridge Oscillator Using Multisim, we construct an IC Wien bridge oscillator as shown in Fig. 14.42a. The oscillator frequency is calculated using

$$f_o = 1/(2\pi\sqrt{R_1C_1R_2C_2})$$

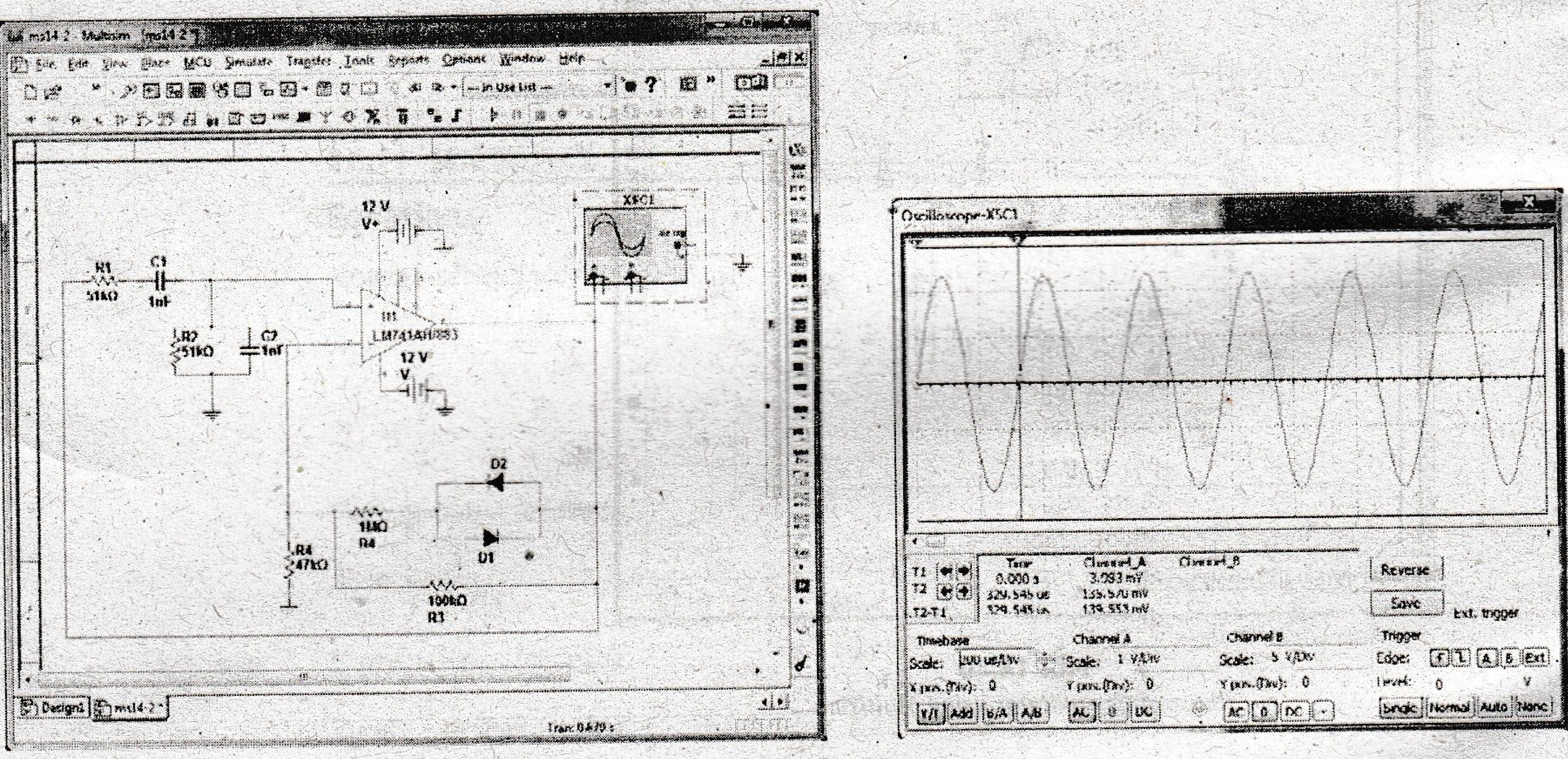
which, for $R_1 = R_2 = R$ and $C_1 = C_2 = C$, is

$$f_o = 1/(2\pi RC) = \frac{1}{2\pi(51\text{ k})(1nf)}$$

$$= 312\text{ Hz}$$

The oscilloscope waveform in Fig. 14.42b shows the resonating waveform with cursors $T2 - T1 = 329.545\ \mu S$, the scope frequency is

$$f = \frac{1}{T} = \frac{1}{329.545\ \mu S} \cong 3{,}034.5\text{ Hz}$$



(a) (b)

FIG. 14.42
(a) Wien bridge oscillator using Multisim; (b) scope waveform.

Example 14.12—IC Colpitts Oscillator Using Multisim, we construct a Colpitts oscillator as shown in Fig. 14.43a.

Using Eq. 14.45

$$Ce_1 = \frac{C_1C_2}{C_1 + C_2} = \frac{(150\text{ pF})(150\text{ pF})}{(150\text{ pF} + 150\text{ pF})} = 75\text{ pF}$$

The oscillator frequency for this circuit is then (Eq. 14.44)

$$f_o = \frac{1}{(2\pi\sqrt{LC_{eq}})}$$

$$= \frac{1}{2\pi\sqrt{(100\ \mu\text{H})(75\text{ pF})}}$$

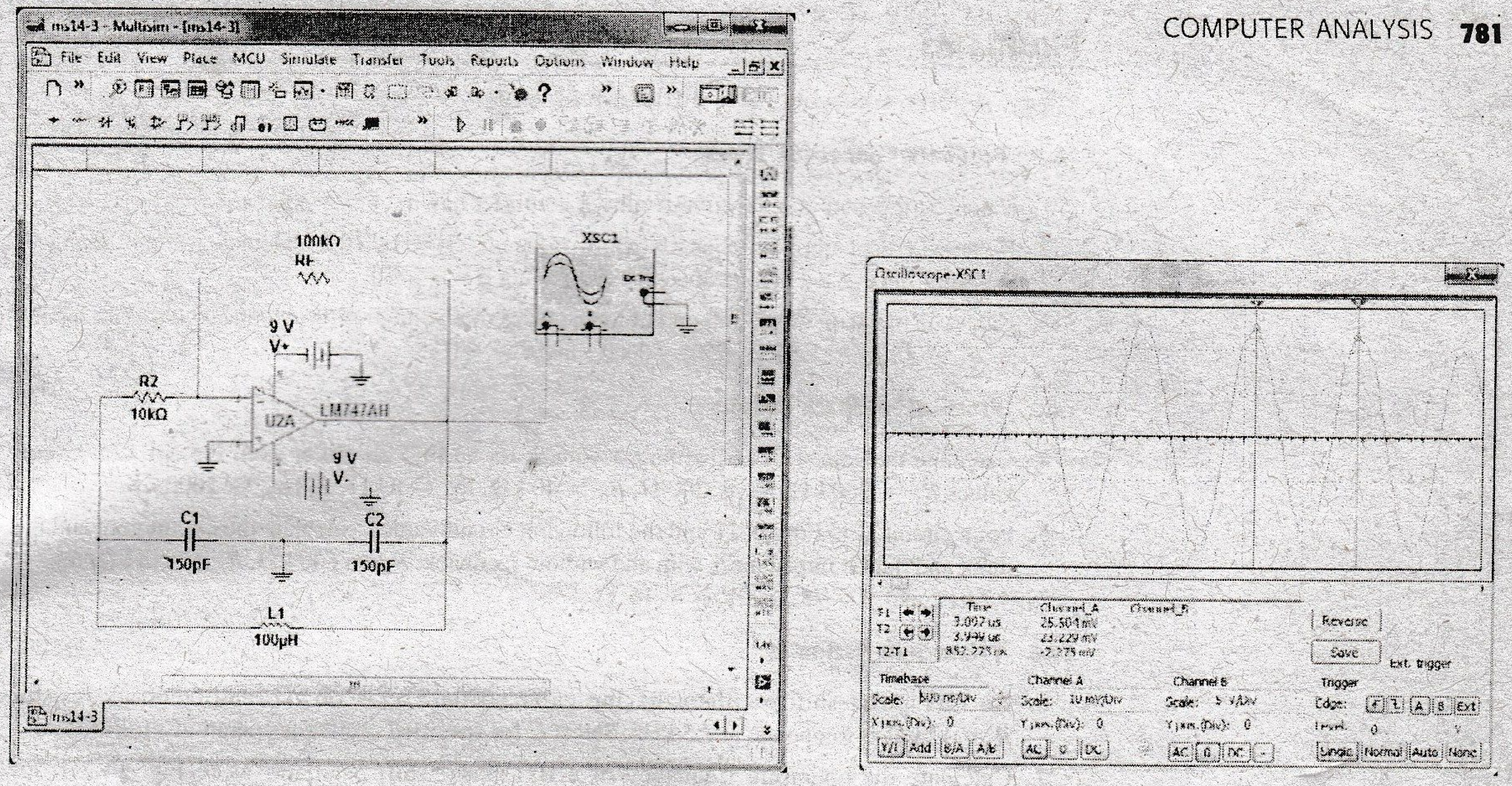
$$= 1{,}837{,}762.985\text{ Hz}$$

$$\cong 1.8\text{ MHz}$$

Fig. 14.43b shows the oscilloscope waveform with

$$f = \frac{1}{T} = \frac{1}{(852.273\ \mu S)}$$

$$\cong 1.2\text{ MHz}$$



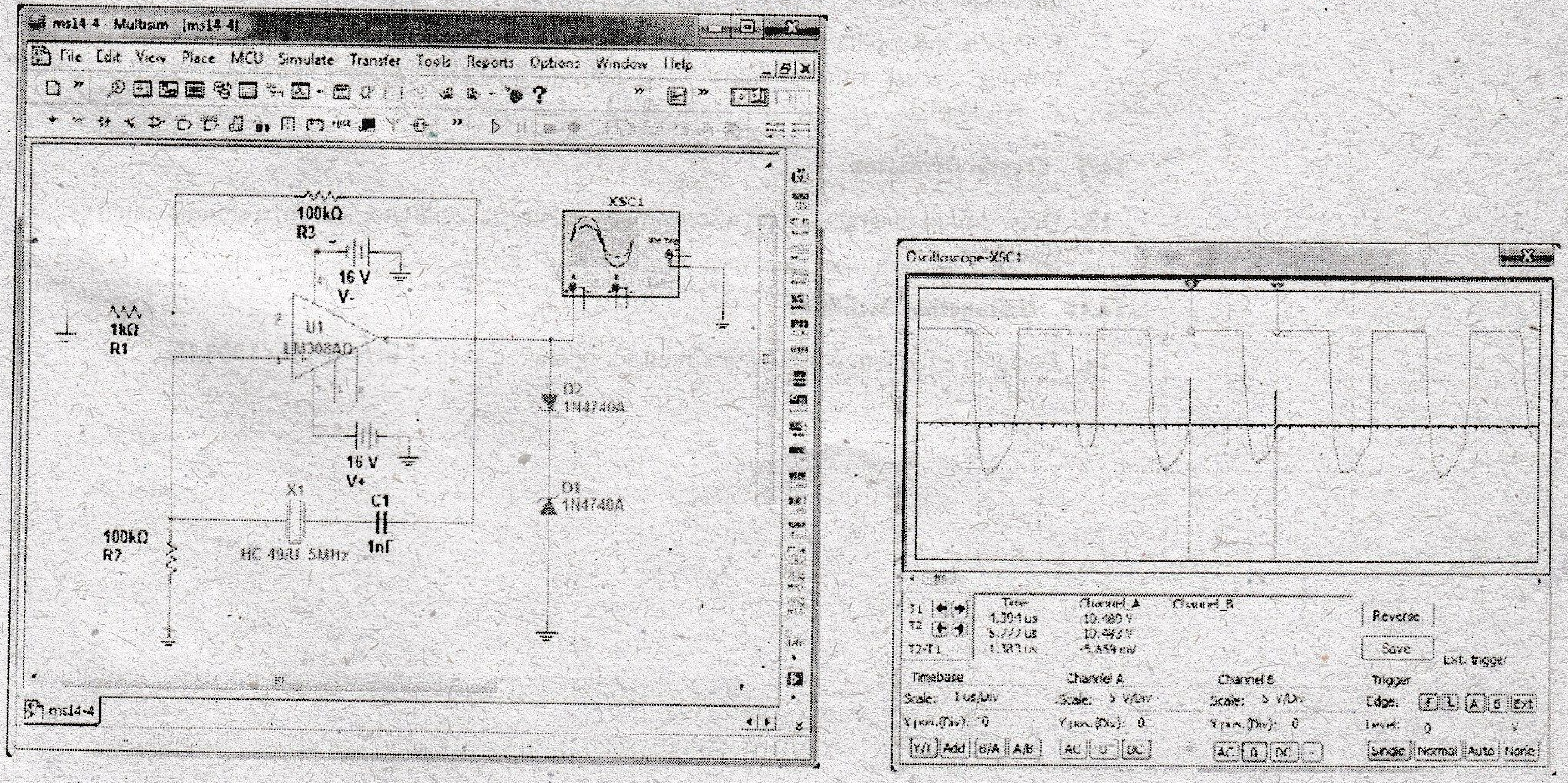
(a) (b)

FIG. 14.43
(a) IC Colpitts oscillator using Multisim; (b) scope waveform.

Example 14.13—Crystal Oscillator Using Multisim, we draw a crystal oscillator circuit as shown in Fig. 14.44a. The oscillator frequency is kept from changing by the crystal. The waveform in Fig. 14.44b shows the period to be about 2.383 μS.

The frequency is then

$$f = 1/T = 1/2.383\ \mu s = 0.42\ \text{MHz}$$



(a) (b)

FIG. 14.44
(a) Crystal oscillator using Multisim; (b) oscilloscope output using Multisim.

PROBLEMS

**Note:* Asterisks indicate more difficult problems.

14.2 Feedback Connection Types

1. Calculate the gain of a negative-feedback amplifier having $A = -2000$ and $\beta = -1/10$.
2. If the gain of an amplifier changes from a value of -1000 by 10%, calculate the gain change if the amplifier is used in a feedback circuit having $\beta = -1/20$.
3. Calculate the gain, input, and output impedances of a voltage-series feedback amplifier having $A = -300$, $R_i = 1.5\ \text{k}\Omega$, $R_o = 50\ \text{k}\Omega$, and $\beta = -1/15$.

14.3 Practical Feedback Circuits

*4. Calculate the gain with and without feedback for an FET amplifier as in Fig. 14.7 for circuit values $R_1 = 800\ \text{k}\Omega$, $R_2 = 200\ \Omega$, $R_o = 40\ \text{k}\Omega$, $R_D = 8\ \text{k}\Omega$, and $g_m = 5000\ \mu\text{S}$.

5. For a circuit as in Fig. 14.11 and the following circuit values, calculate the circuit gain and the input and output impedances with and without feedback: $R_B = 600\ \text{k}\Omega$, $R_E = 1.2\ \text{k}\Omega$, $R_C = 4.7\ \text{k}\Omega$, and $\beta = 75$. Use $V_{CC} = 16$ V.

14.6 Phase-Shift Oscillator

6. An FET phase-shift oscillator having $g_m = 6000\ \mu\text{S}$, $r_d = 36\ \text{k}\Omega$, and feedback resistor $R = 12\ \text{k}\Omega$ is to operate at 2.5 kHz. Select C for specified oscillator operation.
7. Calculate the operating frequency of a BJT phase-shift oscillator as in Fig. 14.21b for $R = 6\ \text{k}\Omega$, $C = 1500$ pF, and $R_C = 18\ \text{k}\Omega$.

14.7 Wien Bridge Oscillator

8. Calculate the frequency of a Wien bridge oscillator circuit (as in Fig. 14.23) when $R = 10\ \text{k}\Omega$ and $C = 2400$ pF.

14.8 Tuned Oscillator Circuit

9. For an FET Colpitts oscillator as in Fig. 14.26 and the following circuit values determine the circuit oscillation frequency: $C_1 = 750$ pF, $C_2 = 2500$ pF, and $L = 40\ \mu$H.
10. For the transistor Colpitts oscillator of Fig. 14.27 and the following circuit values, calculate the oscillation frequency: $L = 100\ \mu$H, $L_{RFC} = 0.5$ mH, $C_1 = 0.005\ \mu$F, $C_2 = 0.01\ \mu$F, and $C_C = 10\ \mu$F.
11. Calculate the oscillator frequency for an FET Hartley oscillator as in Fig. 14.29 for the following circuit values: $C = 250$ pF, $L_1 = 1.5$ mH, $L_2 = 1.5$ mH, and $M = 0.5$ mH.
12. Calculate the oscillation frequency for the transistor Hartley circuit of Fig. 14.30 and the following circuit values: $L_{RFC} = 0.5$ mH, $L_1 = 750\ \mu$H, $L_2 = 750\ \mu$H, $M = 150\ \mu$H, and $C = 150$ pF.

14.9 Crystal Oscillator

13. Draw circuit diagrams of (a) a series-operated crystal oscillator and (b) a shunt-excited crystal oscillator.

14.10 Unijunction Oscillator

14. Design a unijunction oscillator circuit for operation at (a) 1 kHz and (b) 150 kHz.