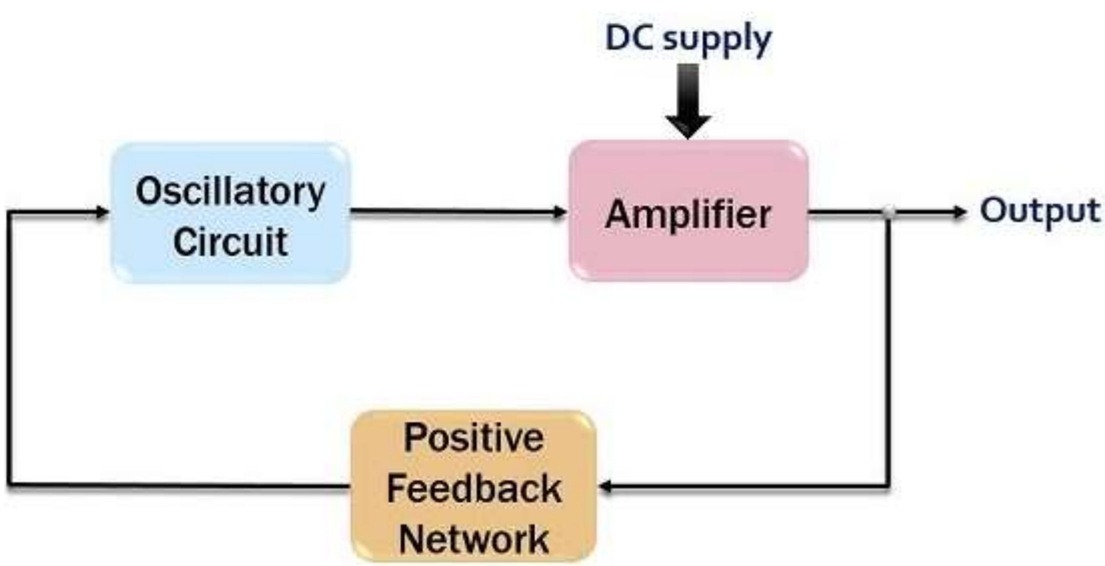
Electronic Circuit

Lecture 6 (10^{th} Week)

Feedback and Oscillator Circuits



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1.1. FEEDBACK CONCEPTS

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

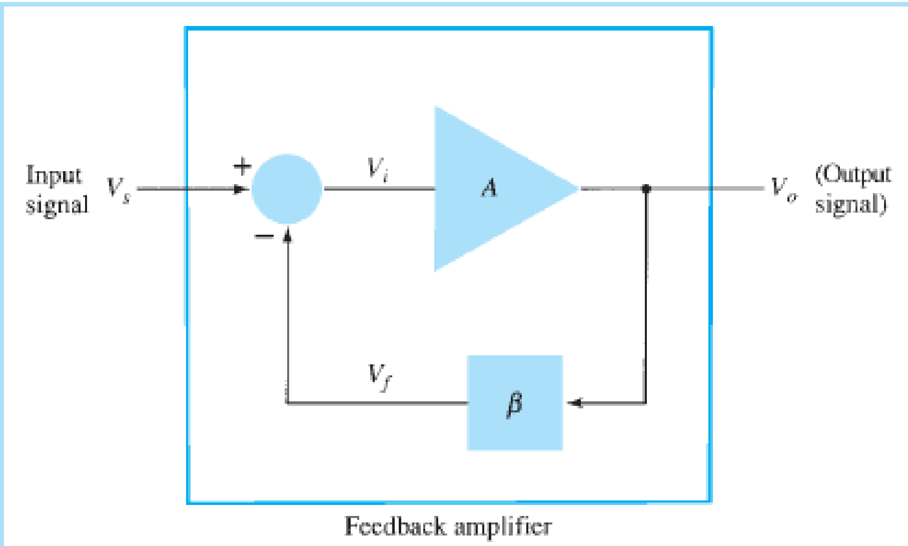

FIG. 14.1
Simple block diagram of feedback amplifier.

1.2. 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:

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

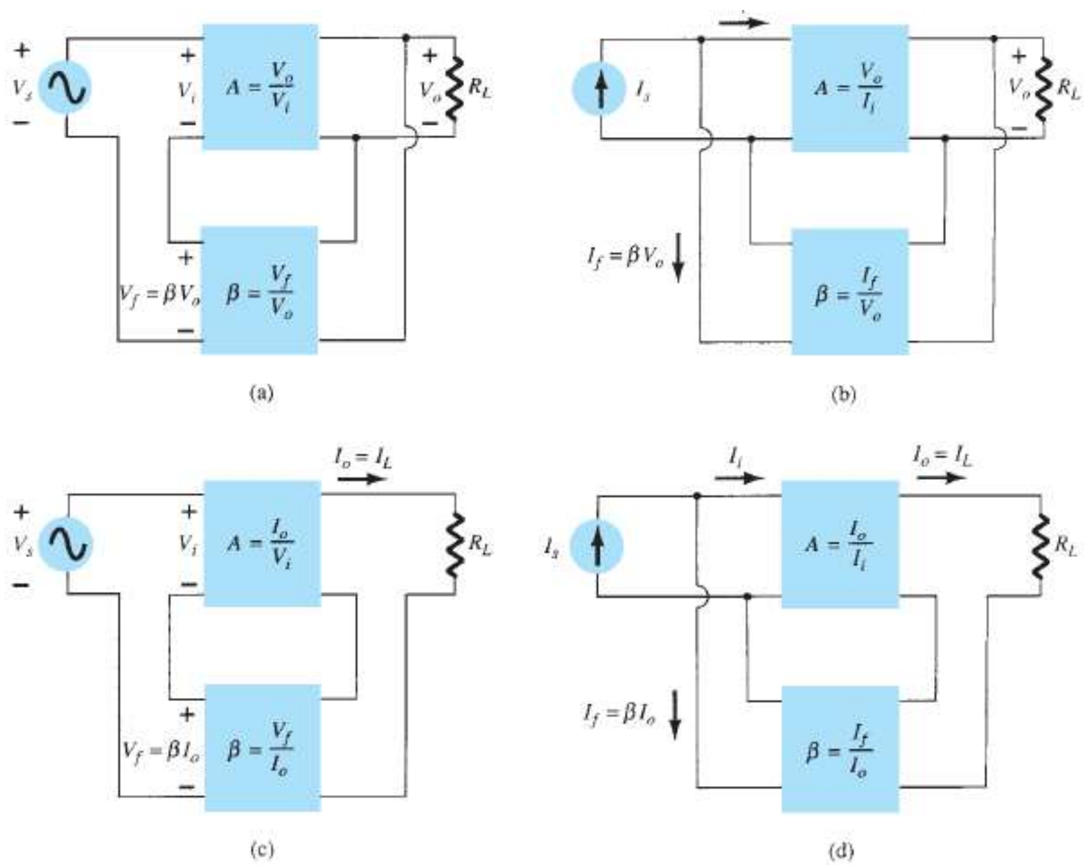

FIG. 14.2
Feedback amplifier types: (a) voltage-series feedback, $A_f = V_o/V_s$; (b) voltage-shunt feedback, $A_f = V_o/I_s$; (c) current-series feedback, $A_f = I_o/V_s$; (d) current-shunt feedback, $A_f = I_o/I_s$.

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.

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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

$$A = \frac{V_o}{V_s} = \frac{V_o}{V_i} \tag{14.1}$$

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

$$V_i = V_s - V_f$$

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

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

so that the overall voltage gain *with* feedback is

$$A_f = \frac{V_o}{V_s} = \frac{A}{1 + \beta A} \tag{14.2}$$

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:

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

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

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

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

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.

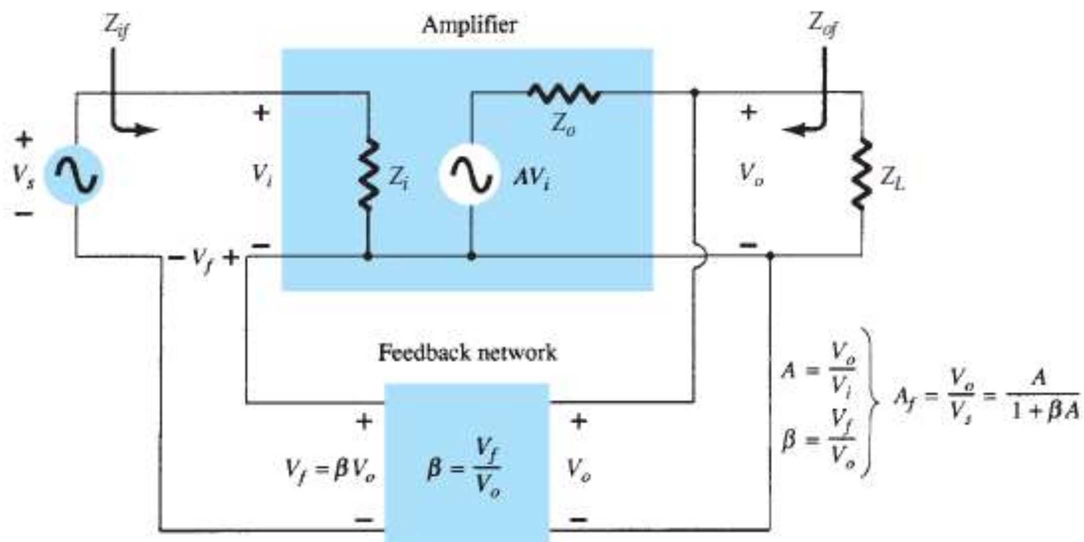

FIG. 14.3
Voltage-series feedback connection.

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 $V_s = 0$,
$$V_i = -V_f$$

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

Rewriting the equation as

$$V + \beta AV = IZ_o$$

allows solving for the output impedance with feedback:

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

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

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

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

1.3. OSCILLATOR OPERATION

The use of positive feedback that results in a feedback amplifier having closed-loop gain |Af| 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 Vo = AVi after the amplifier stage and in a voltage Vf = β(AVi) after the feedback stage. Thus, we have a feedback voltage Vf = βAVi, where A 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, Vf can be made equal to Vi . Then, when the switch is closed and the fictitious voltage Vi 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 βA = 1 (14.32) is met. This is known as the Barkhausen criterion for oscillation.

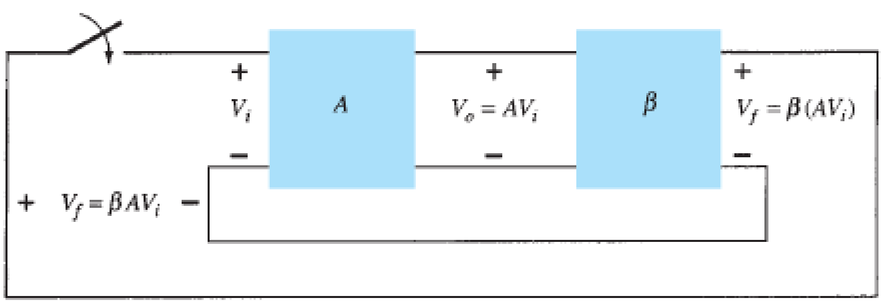

FIG. 14.18
Feedback circuit used as an oscillator.

In reality, no input signal is needed to start the oscillator going. Only the condition βA 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

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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

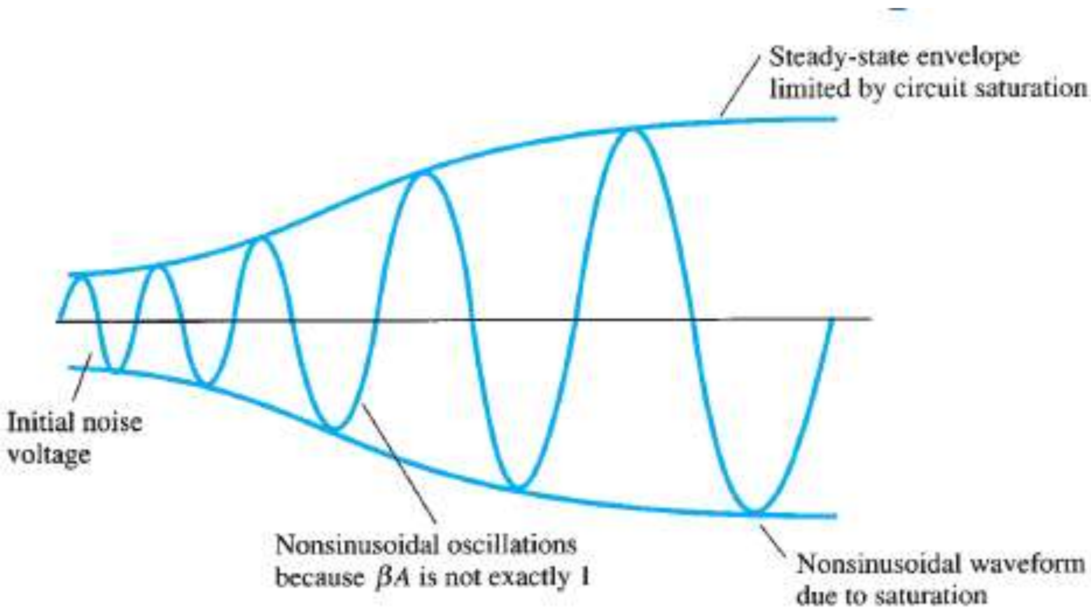

FIG. 14.19
Buildup of steady-state oscillations.

1.4. 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

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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

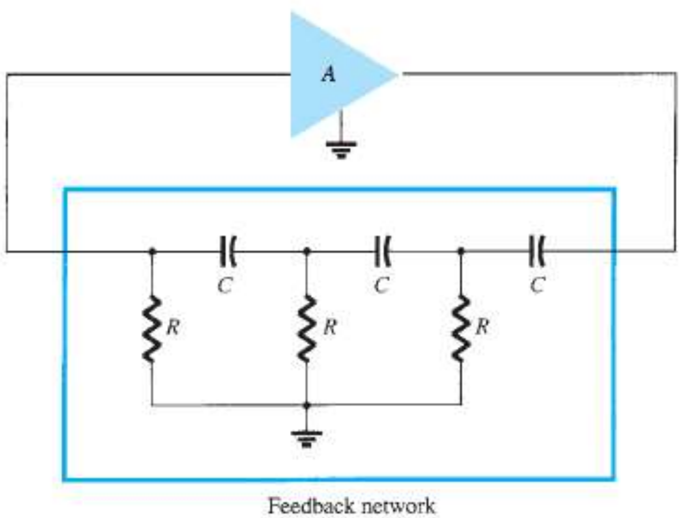

FIG. 14.20
Idealized phase-shift oscillator.

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

$$f = \frac{1}{2\pi RC\sqrt{6}} \tag{14.33}$$

$$\beta = \frac{1}{29} \tag{14.34}$$

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:

$$A > 29 \tag{14.35}$$

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When considering the operation of the feedback network, one might naively select the values of R and C to provide (at a specific frequency) 600-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.

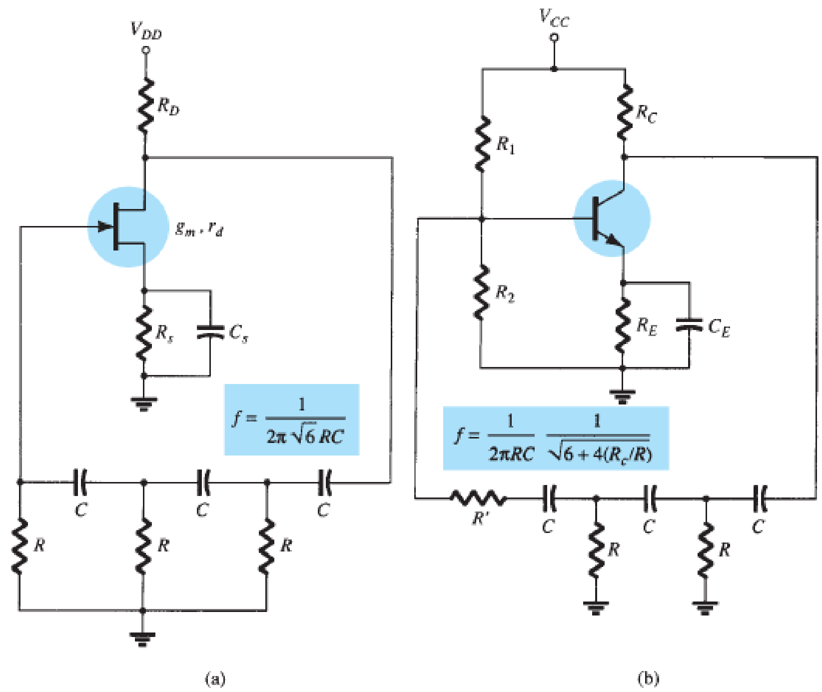

FIG. 14.21
Practical phase-shift oscillator circuits: (a) FET version; (b) BJT version.

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):

$$|A| = g_m R_L$$

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

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

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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,

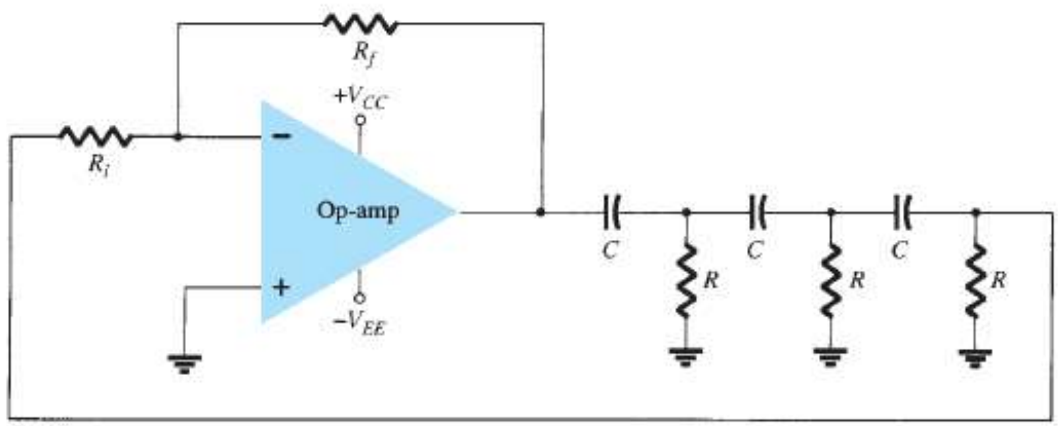

FIG. 14.22
Phase-shift oscillator using an op-amp.

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

$$\frac{R_3}{R_4} = \frac{R_1}{R_2} + \frac{C_2}{C_1} \qquad (14.40)$$

and

$$f_o = \frac{1}{2\pi\sqrt{R_1C_1R_2C_2}} \qquad (14.41)$$

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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).

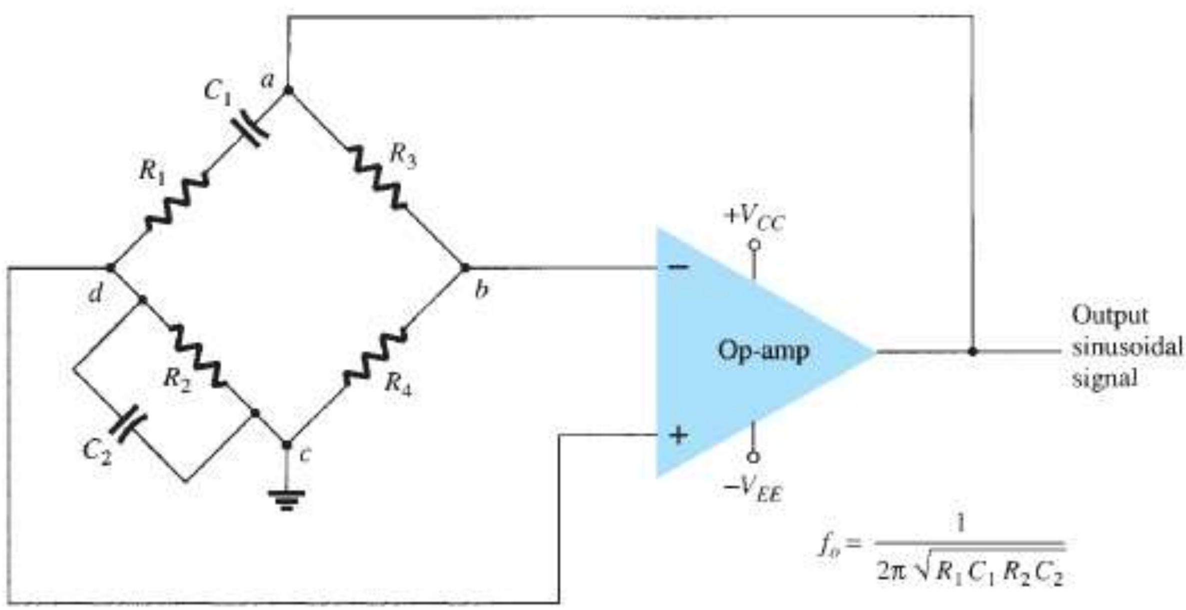

FIG. 14.23
Wien bridge oscillator circuit using an op-amp amplifier.

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

Solution: Using Eq. (14.42) yields

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

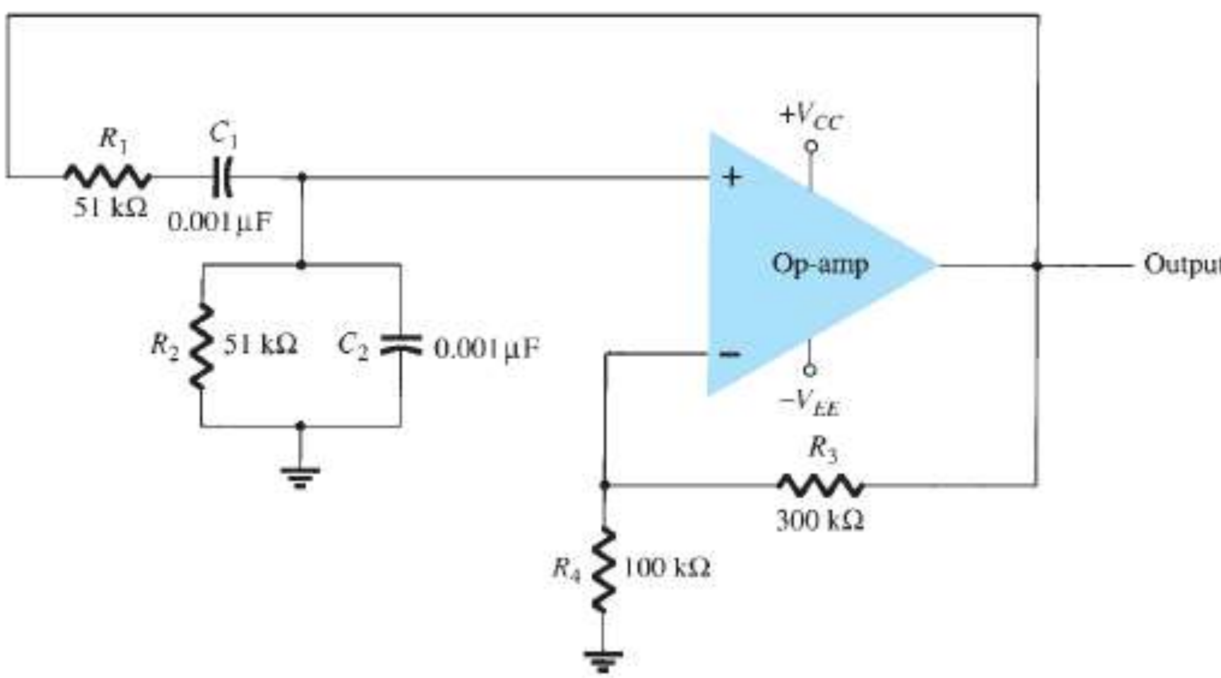

FIG. 14.24
Wien bridge oscillator circuit for Example 14.8.

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