Microelectronic Circuits

Chapter 14 – Signal Generators and Waveform-Shaping Circuits

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Outline

- Basic Principles of Sinusoidal Oscillators
- Op Amp–RC Oscillator Circuits
- LC and Crystal Oscillators
- Bistable Multivibrators
- Generation of Square and Triangular Waveforms Using Astable Multivibrators
- Generation of a Standardized Pulse—The Mono-stable Multivibrator
- Integrated-Circuit Timers
- Nonlinear Waveform-Shaping Circuits
- Precision Rectifier Circuits

Basic Principles of Sinusoidal Oscillators

- Techniques have been developed by which the design of sinusoidal oscillators can be performed in two steps:
  - The first step is a linear one, and frequency-domain methods of feedback circuit analysis can be readily employed.
  - Subsequently, a nonlinear mechanism for amplitude control can be provided.

The Oscillator Feedback Loop

- A sinusoidal oscillator consists of an amplifier and a frequency-selective network connected in a positive-feedback loop
  - Here the feedback signal x_f is summed with a *positive* sign.
- The gain-with-feedback is given by
  - The loop gain L(s) is given by $L(s) \equiv A(s)\beta(s)$
  - The characteristic equation is $1 - L(s) = 0$

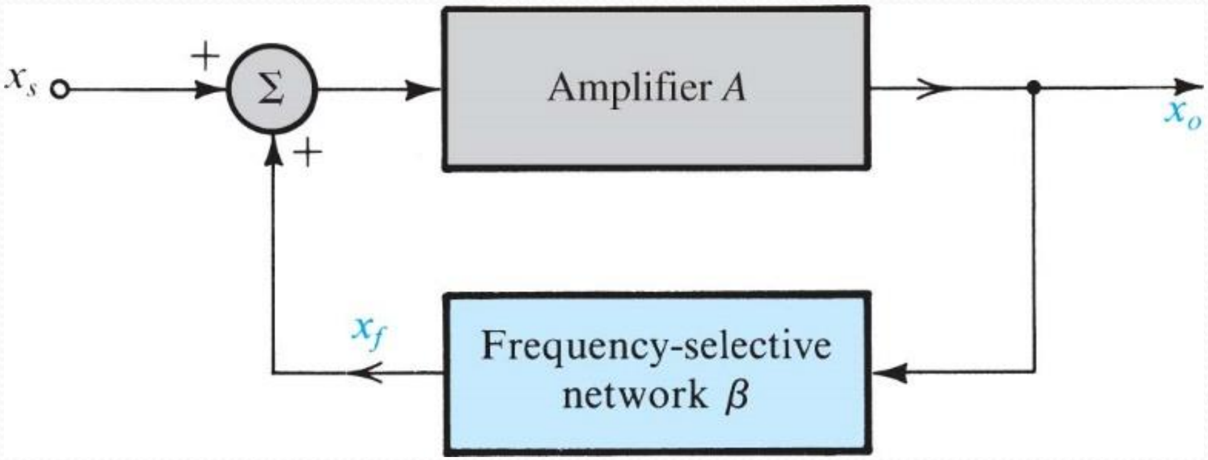


Figure 14.1 The basic structure of a sinusoidal oscillator. A positive-feedback loop is formed by an amplifier and a frequency-selective network. In an actual oscillator circuit, no input signal will be present; here an input signal x_s is employed to help explain the principle of operation.

The Oscillation Criterion

- The condition for the feedback loop of Fig. 14.1 to provide sinusoidal oscillations of frequency ω_0 is $L(j\omega_0) \equiv A(j\omega_0)\beta(j\omega_0) = 1$
- That is, at ω_0 the phase of the loop gain should be zero and the magnitude of the loop gain should be unity. This is known as the Barkhausen criterion.
- In the feedback loop of Fig. 14.1, the Barkhausen criterion means that this loop produces and sustains an output x_o with no input applied ($x_s = 0$), the feedback signal x_f

$$x_f = \beta x_o$$

should be sufficiently large that when multiplied by A it produces x_o, that is,

$$Ax_f = x_o$$

that is, $A\beta x_o = x_o$

which results in $A\beta = 1$

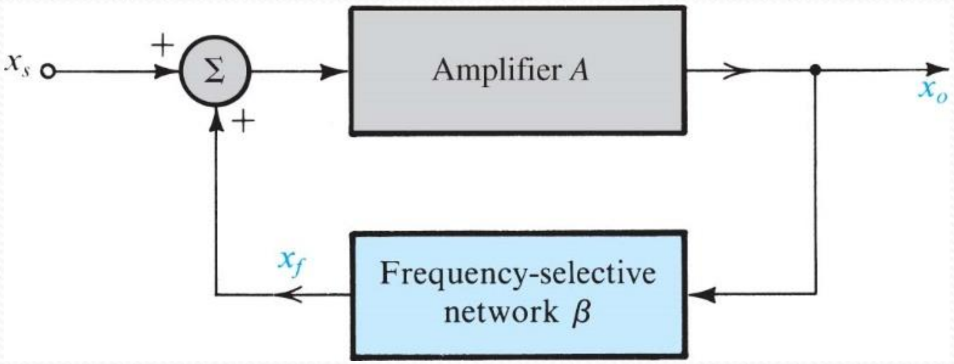
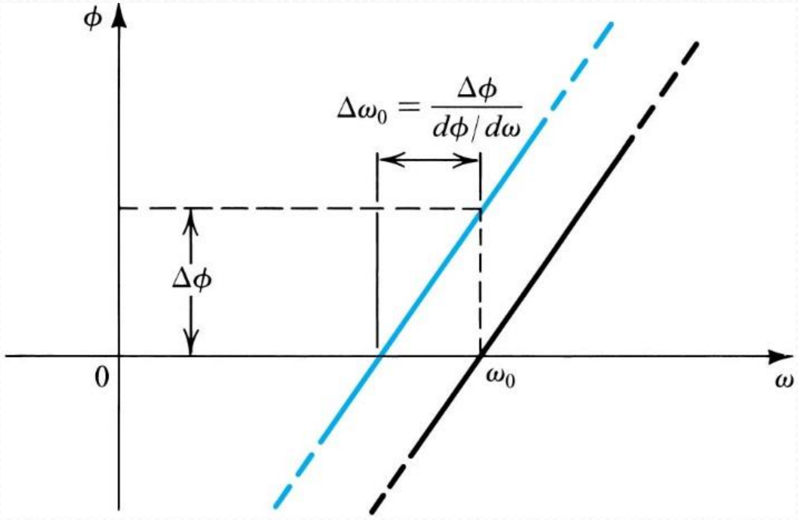


Figure 14.1

The Oscillation Criterion (Cont'd)

- It should be noted that the frequency of oscillation ω_0 is determined solely by the phase characteristics of the feedback loop;
  - the loop oscillates at the frequency for which the phase is zero (or, equivalently, 360°).
  - A "steep" function $\phi(\omega)$ will result in a more stable frequency. This can be seen if one imagines a change in phase $\Delta\phi$ due to a change in one of the circuit components. If *dφ/dω* is large, the resulting change in ω_0 will be small, as illustrated in Fig. 14.2.



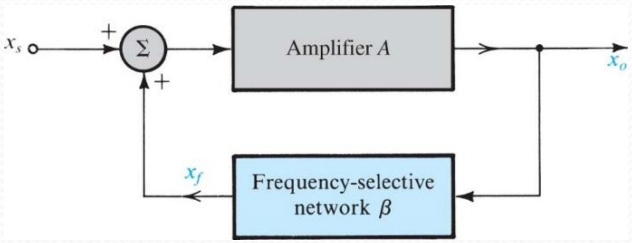
- $1 - L(s) = 0$
- $L(j\omega_0) \equiv A(j\omega_0)\beta(j\omega_0) = 1$

Figure 14.2 Dependence of the oscillator-frequency stability on the slope of the phase response. A steep phase response (i.e., large $d\phi/d\omega$) results in a small $\Delta\omega_0$ for a given change in phase $\Delta\phi$ [resulting from a change (due, for example, to temperature) in a circuit component].

Analysis of Oscillator Circuits

- Analysis of a given oscillator circuit proceeds in three steps:
  1. Break the feedback loop to determine the loop gain $A(s)\,\beta(s)$.
  2. The oscillation frequency ω_0 is found as the frequency for which the phase angle of $A(j\omega_0)\beta(j\omega_0)$ is zero or, equivalently, 360^o.
  3. The condition for the oscillations to start is found from $|A(j\omega_0)\beta(j\omega_0)| \geqq 1$

- $1 - L(s) = 0$
- $L(j\omega_0) \equiv A(j\omega_0)\beta(j\omega_0) = 1$



- Note that making the magnitude of the loop gain slightly greater than unity ensures that oscillations will start.

- An Alternative Analysis Approach
  - The method proceeds as follows: We assume that the circuit is oscillating at the oscillation frequency ω_0. We reduce the equations to a single equation in terms of a single voltage or current variable. The equation can then be manipulated to the form $D(s) = 0$

Example 14.1

Figure 18.3(a) shows a sinusoidal oscillator formed by placing a second-order LCR bandpass filter [see Fig. 17.18(d)] in the feedback path of a positive-gain amplifier. Find the frequency of oscillation ω_0, and the condition for oscillations to start. Assume an ideal op amp.

Solution

To obtain the loop gain, we break the positive-feedback loop at the positive input terminal of the op amp where the input impedance is infinite, apply an input voltage *Vi*, and find the returned voltage *Vr* . This results in the circuit in Fig. 18.3(b). Since the op amp is ideal, we

can find its output voltage *Vo* and hence *A(s)* as $A(s) \equiv \frac{V_o}{V_i} = 1 + \frac{r_2}{r_1}$

$$\beta(s) \equiv \frac{V_r}{V_o} = \frac{s\frac{1}{CR}}{s^2 + s\frac{1}{CR} + \frac{1}{LC}}$$

The transfer function of the frequency-selective network $\beta(s) \equiv Vr/Vo$ can be found utilizing Eq.(17.39) as

While we have found it easy to determine *A(s)* and $\beta(s)$ separately, our interest is in fact in their product,

the loop gain $A(s)\,\beta(s)$,

$$A(s)\beta(s) = \frac{s\frac{1}{CR}\left(1 + \frac{r_2}{r_1}\right)}{s^2 + s\frac{1}{CR} + \frac{1}{LC}}$$

Substituting $s = j\omega$,

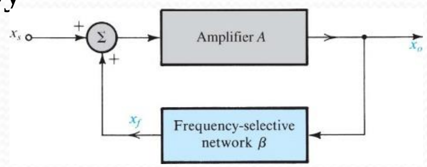
$$A(j\omega)\beta(j\omega) = \frac{j\frac{\omega}{CR}\left(1 + \frac{r_2}{r_1}\right)}{\left(-\omega^2 + \frac{1}{LC}\right) + j\frac{\omega}{CR}}$$

From this expression we see that the phase angle of $A(j\omega)\,\beta(j\omega)$ will be zero at the value of ω that makes the **real part of the denominator zero**, thus $\omega_0 = 1/\sqrt{LC}$

At this frequency, the magnitude of the loop gain is given by

$$|A(j\omega_0)\beta(j\omega_0)| = 1 + \frac{r_2}{r_1}$$

Thus oscillation will start for $r2/r1 \geq 0$



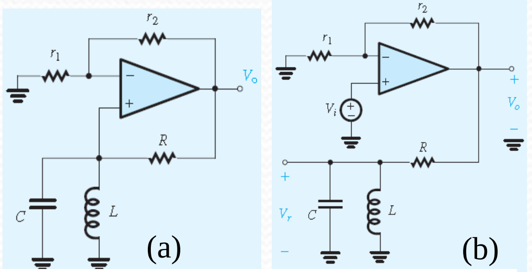


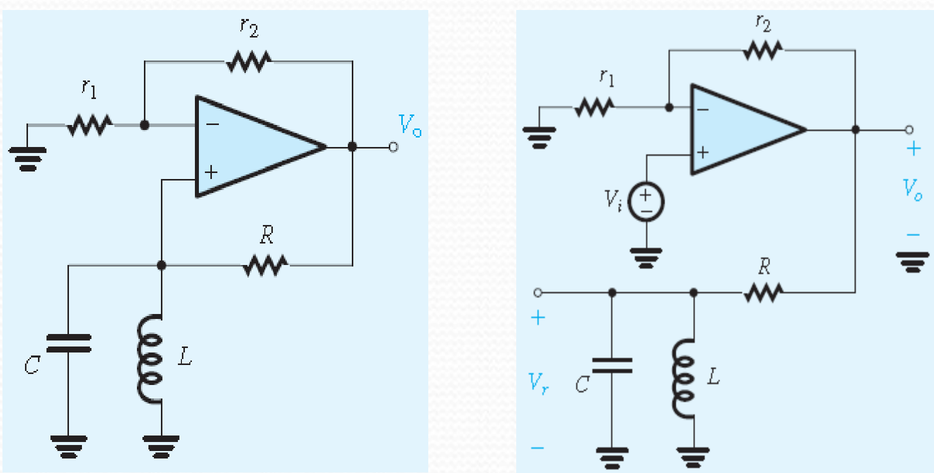
Figure 18.3 **(a)** An oscillator formed by connecting a positive-gain amplifier in a feedback loop with a bandpass RLC circuit. **(b)** Breaking the feedback loop at the input of the op amp to determine $A(s) \equiv Vo(s)/Vi(s)$ and $\beta(s) \equiv Vr(s)/Vo(s)$, and hence the loop gain $A(s)\beta(s)$.

Example 14.1 (Cont'd)

*While $r2/r1=0$ is theoretically sufficient for sustained oscillations, it is usual to design for $r2/r1 > 0$ to ensure that oscillations start.

*Inspection: Since the phase angle of the amplifier gain is zero, we examine the LCR bandpass circuit to determine the frequency at which its phase is zero. This in turn is the frequency at which the LC tank has an infinite impedance, which is the resonance frequency $\omega_0 = 1/\sqrt{LC}$.

*At this frequency, the transmission of the bandpass LCR circuit is unity. Thus for oscillation to start, the amplifier gain must be greater than or equal to unity.



$$\beta(s) \equiv \frac{V_r}{V_o} = \frac{s\dfrac{1}{CR}}{s^2 + s\dfrac{1}{CR} + \dfrac{1}{LC}}$$

$$A(s)\beta(s) = \frac{s\dfrac{1}{CR}\left(1+\dfrac{r_2}{r_1}\right)}{s^2 + s\dfrac{1}{CR} + \dfrac{1}{LC}}$$

$$A(j\omega)\beta(j\omega) = \frac{j\dfrac{\omega}{CR}\left(1+\dfrac{r_2}{r_1}\right)}{\left(-\omega^2 + \dfrac{1}{LC}\right) + j\dfrac{\omega}{CR}}$$

Figure 18.3 **(a)** An oscillator formed by connecting a positive-gain amplifier in a feedback loop with a bandpass RLC circuit. **(b)** Breaking the feedback loop at the input of the op amp to determine $A(s) \equiv Vo(s)/Vi(s)$ and $\beta(s) \equiv Vr(s)/Vo(s)$, and hence the loop gain $A(s)\beta(s)$.

Nonlinear Amplitude Control to achieve robustness

- The parameters of any physical system cannot be maintained constant for any length of time.
- Suppose we work hard to make $|A\beta| = 1$ at $\omega = \omega_0$, and then the temperature changes and $|A\beta|$ becomes slightly less than unity.
  - We therefore need a mechanism for forcing $|A\beta|$ to remain equal to unity at the desired value of output amplitude.
  - This task is accomplished by providing a nonlinear circuit for gain control.
- Two basic approaches to the implementation of the nonlinear amplitude-stabilization mechanism.
  - A limiter circuit. – making $|A\beta| > 1$ and a limiter to maintain oscillation.
  - The other mechanism for amplitude control utilizes an element whose resistance can be controlled by the amplitude of the output sinusoid.

A Popular Limiter Circuit for Amplitude Control

- Consider first the case of a small (close to zero) input signal v_I and a small output voltage v_O, so that v_A is positive and v_B is negative. It can be easily seen that both diodes D1 and D2 will be off.
- Thus all of the input current v_I/R_1 flows through the feedback resistance R_f, and the output voltage is given by $v_O = -(R_f / R_1)v_I$
- This is the linear portion of the limiter transfer characteristic in Fig. 14.4(b).

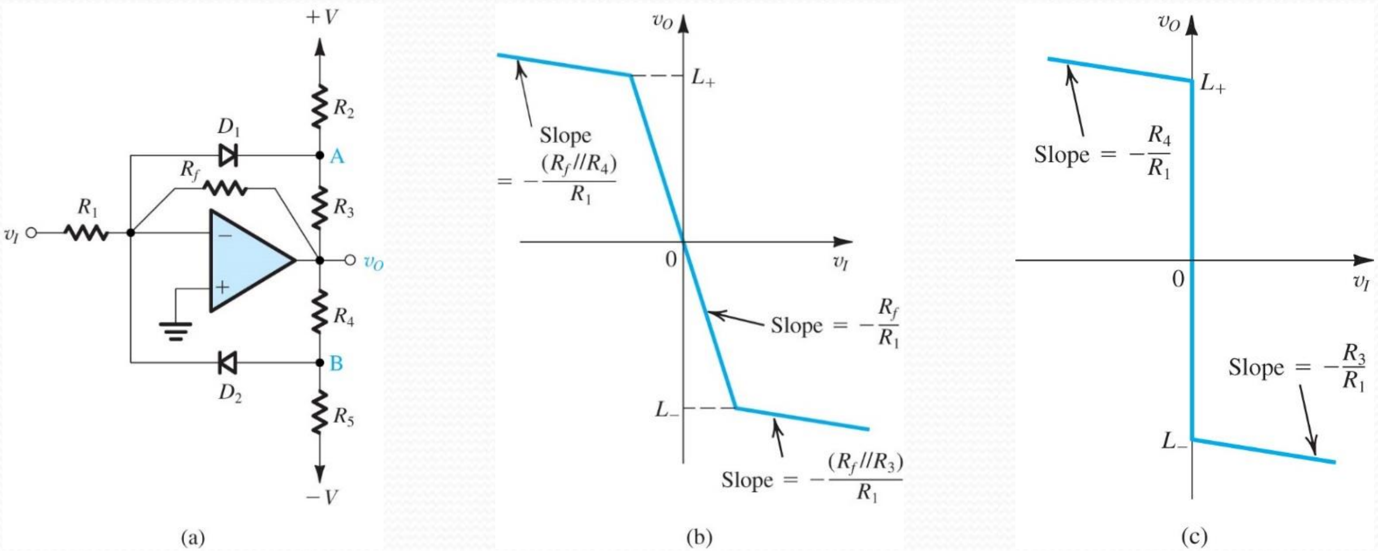
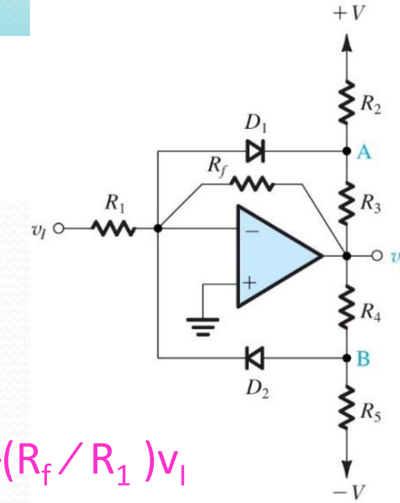


Figure 14.4 (a) A popular limiter circuit. **(b)** Transfer characteristic of the limiter circuit; L_- and L_+ are given by Eqs. (14.8) and (14.9), respectively. **(c)** When R_f is removed, the limiter turns into a comparator with the characteristic shown.

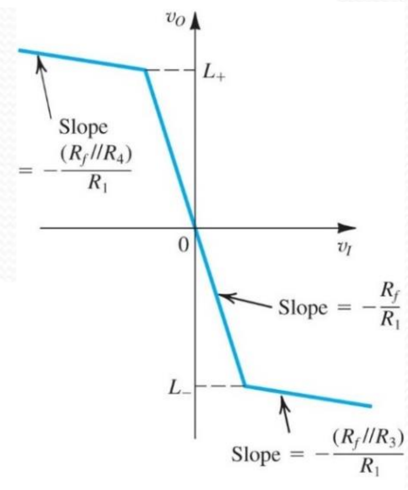
A Popular Limiter Circuit for Amplitude Control (Cont'd)



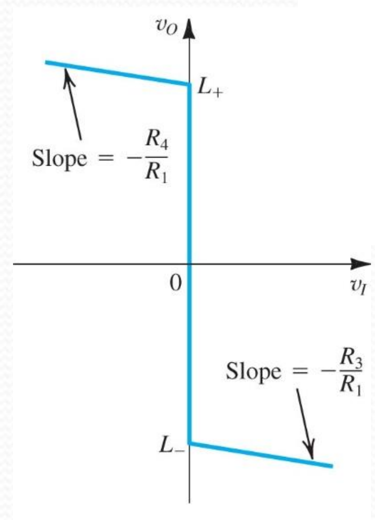
- Via superposition:
  - $$v_A = V\frac{R_3}{R_2+R_3} + v_O\frac{R_2}{R_2+R_3} \quad (14.6)$$
  - $$v_B = -V\frac{R_4}{R_4+R_5} + v_O\frac{R_5}{R_4+R_5} \quad (14.7)$$

$v_O = -(R_f / R_1)v_I$

- A constant-voltage-drop V_D for D_1, the value of v_O at which **D_1 conducts** can be found from Eq. (14.6). This is the negative limiting level, which we denote L_- ,
  - $$L_- = -V\frac{R_3}{R_2} - V_D\left(1+\frac{R_3}{R_2}\right)$$
- It can be easily seen that for negative v_I, we can use Eq. (14.7) to find the positive limiting level L_+,
  - $$L_+ = V\frac{R_4}{R_5} + V_D\left(1+\frac{R_4}{R_5}\right)$$
- We see that the circuit of Fig. 14.4(a) functions as a soft limiter, with the limiting levels L_+ and L_-, and the limiting gains independently adjustable by the selection of appropriate resistor values.
- In the limit, removing R_f results in the transfer characteristic of Fig. 14.4(c), which is that of a comparator.



(b)



(c)

Op Amp–RC Oscillator Circuits: The Wien-Bridge Oscillator

- A Wien-bridge oscillator without the nonlinear gain-control network. The loop gain can be easily obtained by multiplying the transfer function $V_a(s)/V_o(s)$ of the feedback network by the amplifier gain,

$$L(s) = \left[1 + \frac{R_2}{R_1}\right]\frac{Z_p}{Z_p + Z_s} = \frac{1 + R_2/R_1}{1 + Z_s Y_p}$$

- Thus, $$L(s) = \frac{1 + R_2/R_1}{3 + sCR + 1/sCR}$$

- Substituting s = jω results in $$L(j\omega) = \frac{1 + R_2/R_1}{3 + j(\omega CR - 1/\omega CR)}$$

- The loop gain will be a real number (i.e., the phase will be zero) at one frequency given by $$\omega_0 CR = \frac{1}{\omega_0 CR}$$

  That is $\omega_0 = 1/RC$

- Oscillations will start at this frequency if the loop gain is at least unity. This can be achieved by selecting $R_2/R_1 = 2$

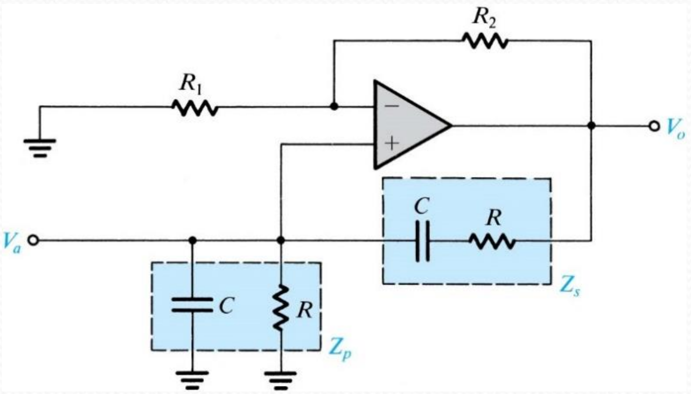


Figure 14.5 A Wien-bridge oscillator without amplitude stabilization.

The Wien-Bridge Oscillator (Cont'd)

- Two different implementations of the amplitude-controlling function are shown in Figs. 14.6 and 14.7

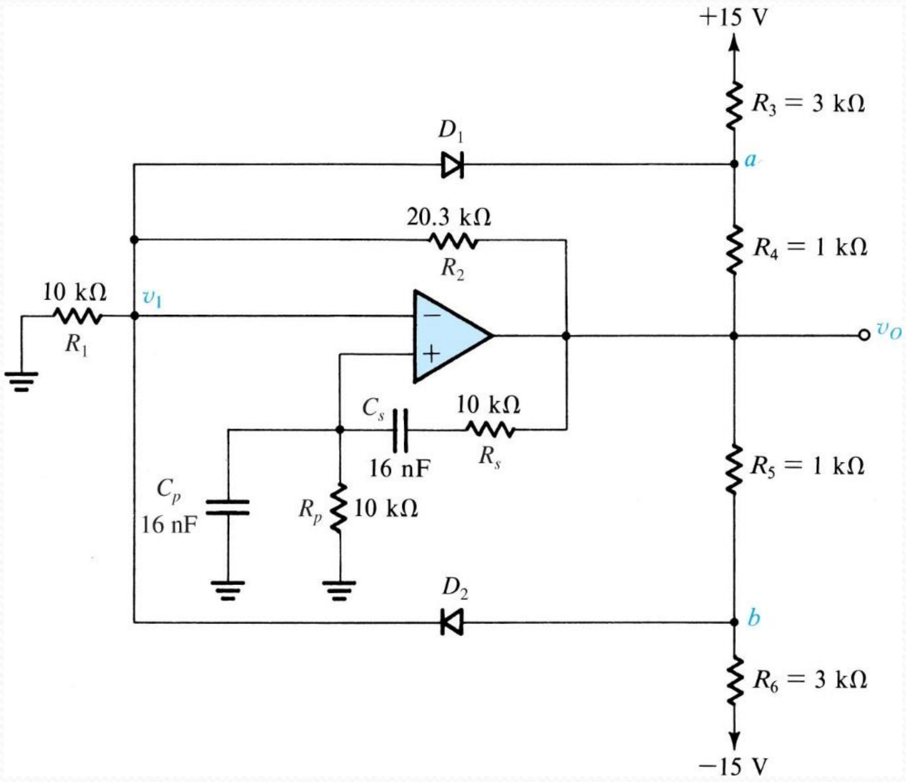


Figure 14.6 A Wien-bridge oscillator with a limiter used for amplitude control.

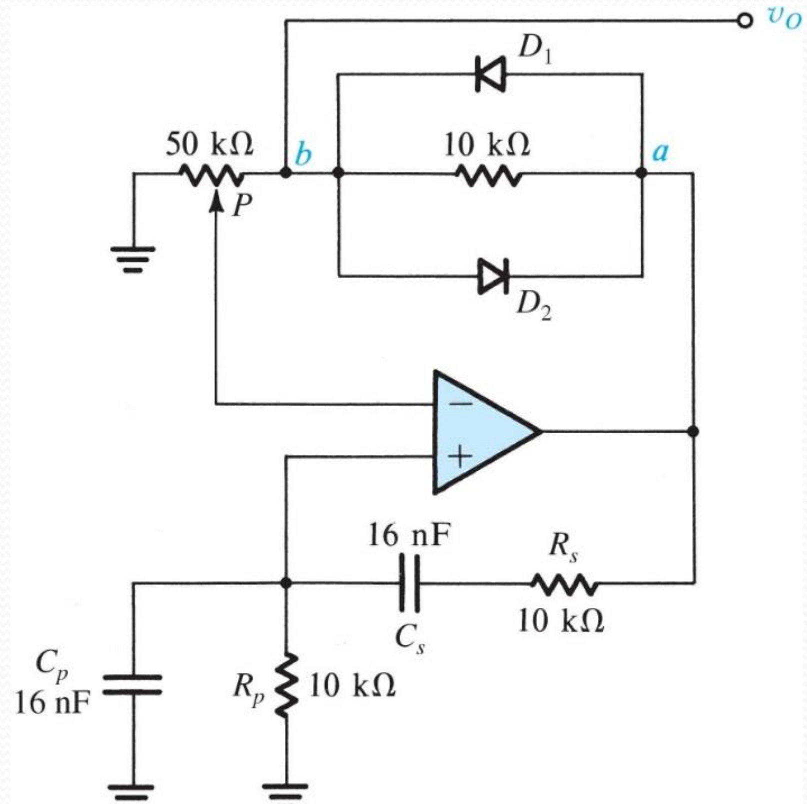


Figure 14.7 A Wien-bridge oscillator with an alternative method for amplitude stabilization.

The Phase-Shift Oscillator

- The phase-shift circuit in Fig. 14.8 will oscillate at the frequency for which the phase shift of the RC network is 180°. Only at this frequency will the total phase shift around the loop be 0° or 360°.
- The reason for using a three-section RC network is that three is the minimum number of sections (i.e., lowest order) that is capable of producing a 180° phase shift at a finite frequency.
- To ensure that oscillations start, the value of K has to be chosen slightly higher than the value that satisfies the unity-loop-gain condition.
- Figure 14.9 shows a practical phase-shift oscillator with a feedback limiter, consisting of diodes D_1 and D_2 and resistors R_1, R_2, R_3, and R_4 for amplitude stabilization. To start oscillations, R_f has to be made slightly greater than the minimum required value.

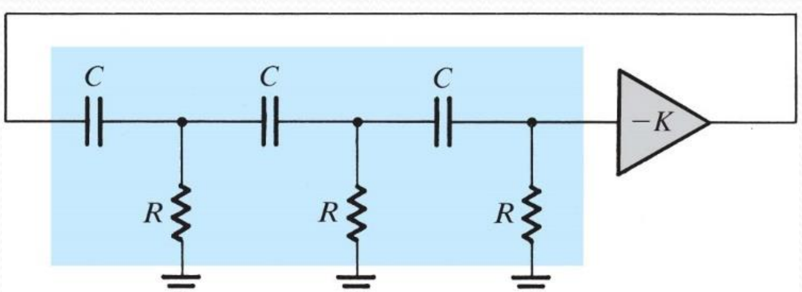


Figure 14.8 A phase-shift oscillator.

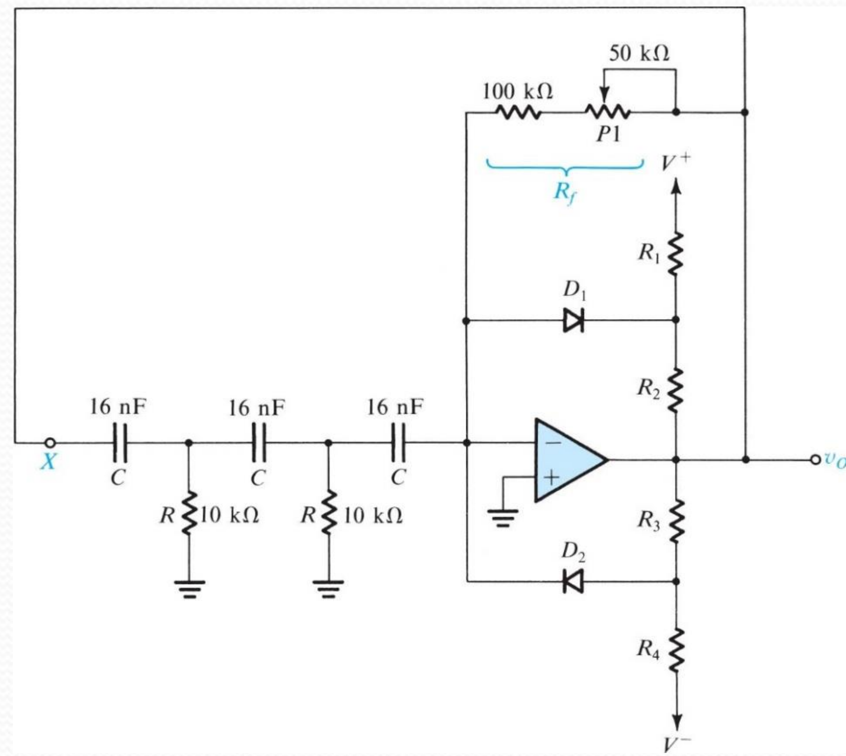


Figure 14.9 A practical phase-shift oscillator with a limiter for amplitude stabilization.

The Quadrature Oscillator

- The quadrature oscillator is based on the two-integrator loop studied in Section 13.7.
- The equivalent circuit shown in Fig. 14.10(b). Nominally, R_f is made equal to 2R, and thus $-R_f$ cancels 2R, and at the input we are left with a current source $v_{O1}/2R$ feeding a capacitor C.
  - $v = \frac{1}{C}\int_0^t \frac{v_{O1}}{2R}\,dt$ and $v_{O2} = 2v = \frac{1}{CR}\int_0^t v_{O1}\,dt.$
  - That is, for $R_f = 2R$, the circuit functions as a perfect noninverting integrator. If, however, R_f is made smaller than 2R, a net negative resistance appears in parallel with C.
- If we disregard the limiter and break the loop at X, the loop gain can be obtained as

$$L(s) \equiv \frac{V_{o2}}{V_x} = -\frac{1}{s^2C^2R^2}$$

- Thus the loop will oscillate at frequency ω_o, given by

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

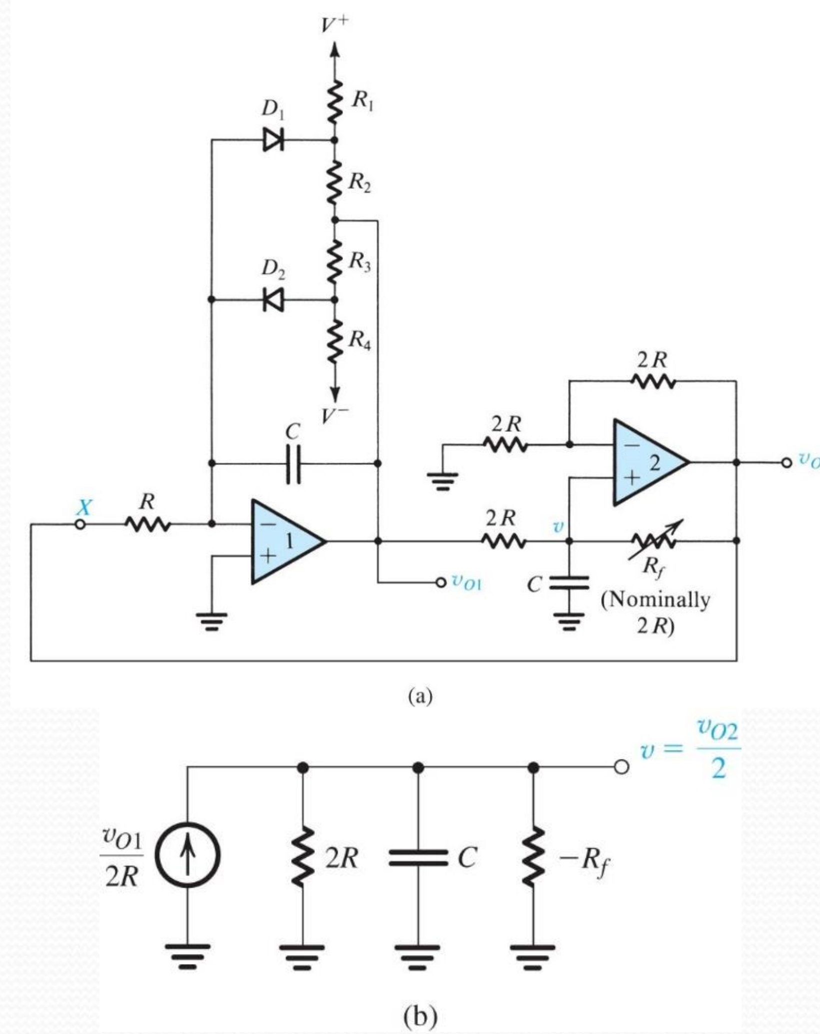


Figure 14.10 (a) A quadrature-oscillator circuit. **(b)** Equivalent circuit at the input of op amp 2.

The Active-Filter-Tuned Oscillator

- The basic principle is illustrated in Fig. 14.11. The circuit consists of a high-Q bandpass filter connected in a positive-feedback loop with a hard limiter. The purity of the output sine wave will be a direct function of the selectivity (or Q factor) of the bandpass filter.
- Figure 14.12 shows one possible implementation of the active-filter-tuned oscillator.

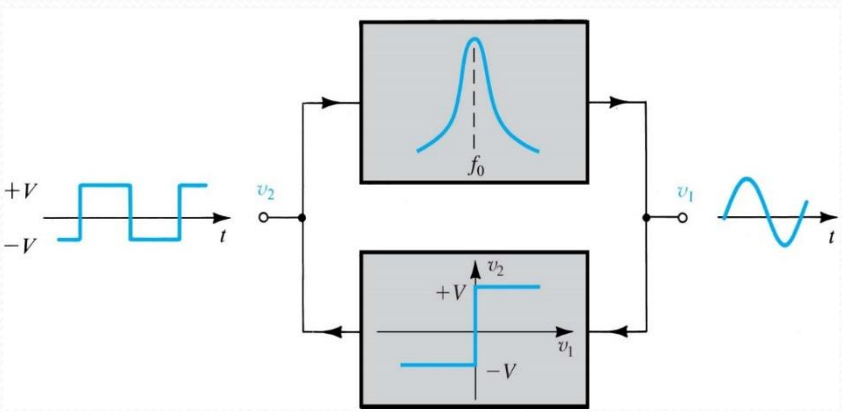


Figure 14.11 Block diagram of the active-filter-tuned oscillator.

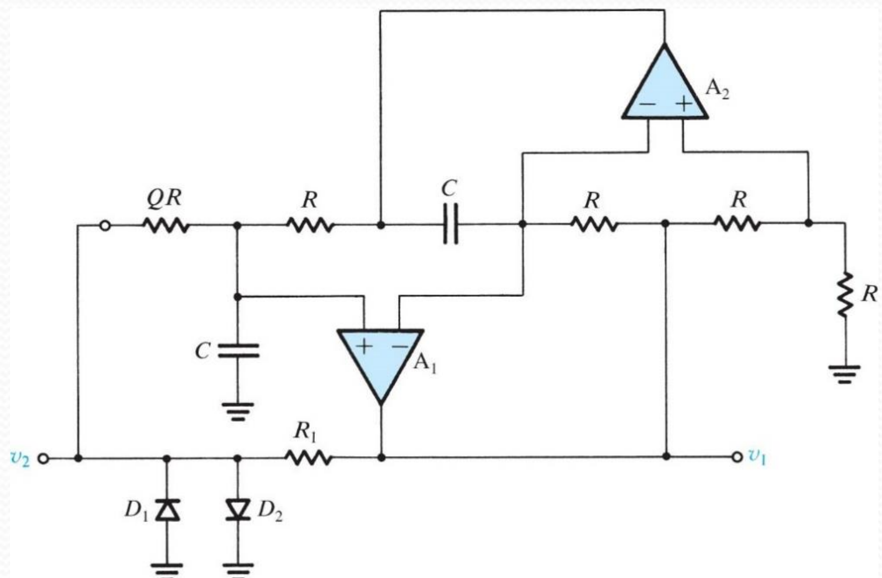


Figure 14.12 A practical implementation of the active-filter-tuned oscillator.

A Final Remark

- The op amp–RC oscillator circuits studied are useful for operation in the range 10 Hz to 100 kHz (or perhaps 1 MHz at most).
- Whereas the lower frequency limit is dictated by the size of passive components required, the upper limit is governed by the frequency-response and slew-rate limitations of op amps.
- For higher frequencies, circuits that employ transistors together with LC-tuned circuits or crystals are frequently used.

LC and Crystal Oscillators

- Oscillators utilizing transistors (FETs or BJTs), with LC-tuned circuits or crystals as feedback elements, are used in the frequency range of 100 kHz to hundreds of megahertz.
- They exhibit higher Q than the RC types.
- However, LC oscillators are difficult to tune over wide ranges, and crystal oscillators operate at a single frequency.

Colpitts and Hartely Oscillators

- Figure 14.13 shows two commonly used configurations of LC oscillators - Colpitts oscillator and Hartley oscillator.
- V_{eb} gives rise to a current I_c in the direction shown, which in turn results in a positive voltage across the LC circuit. Thus, we do have a positive-feedback loop.
- For the Colpitts oscillator we have

$$\omega_0 = 1 \Big/ \sqrt{L\left(\frac{C_1 C_2}{C_1 + C_2}\right)}$$

- and the Hartley oscillator we have

$$\omega_0 = 1/\sqrt{(L_1 + L_2)C}$$

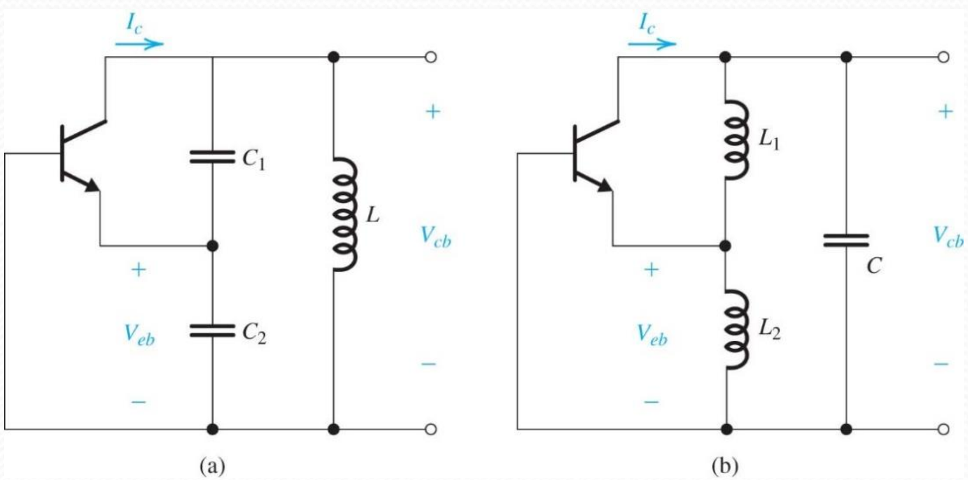


Figure 14.13 Two commonly used configurations of LC-tuned oscillators: **(a)** Colpitts and **(b)** Hartley.

The Colpitts and Hartely Oscillators (Cont'd)

- To determine the oscillation condition for the Colpitts oscillator, we replace the transistor with its equivalent circuit, as shown in Fig. 14.14.
- A node equation at the transistor collector (node C) in the circuit of Fig. 14.14 yields

$$sC_2V_\pi + g_mV_\pi + \left(\frac{1}{R} + sC_1\right)(1 + s^2LC_2)V_\pi = 0$$

- Since $V_\pi \neq 0$(oscillations have started), it can be eliminated, and the equation can be rearranged in the form $s^3LC_1C_2 + s^2(LC_2/R) + s(C_1 + C_2) + \left(g_m + \frac{1}{R}\right) = 0$
- Substituting s = jω gives $\left(g_m + \frac{1}{R} - \frac{\omega^2LC_2}{R}\right) + j[\omega(C_1 + C_2) - \omega^3LC_1C_2] = 0$
- For oscillations to start, both the real and imaginary parts must be zero. Equating the imaginary part to zero gives the frequency of oscillation as $\omega_0 = 1\Big/\sqrt{L\left(\frac{C_1C_2}{C_1 + C_2}\right)}$

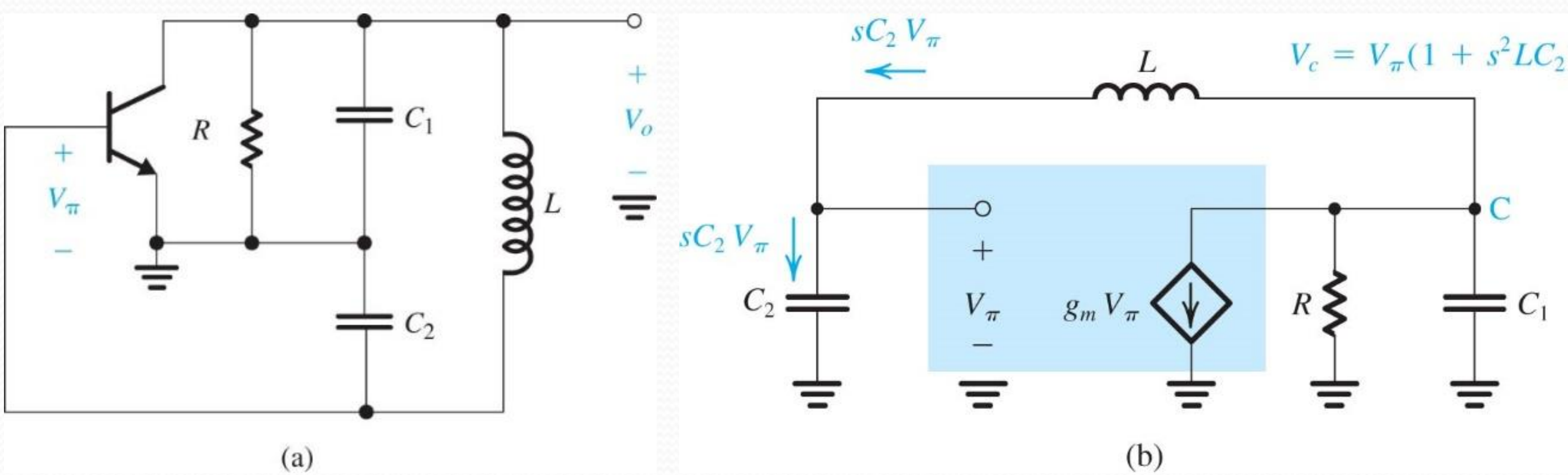


Figure 14.14 Equivalent circuit of the Colpitts oscillator of Fig. 14.13(a). To simplify the analysis, C_μ and r_π are neglected. We can consider C_π to be part of C_2, and we can include r_o in R.

The Colpitts and Hartely Oscillators (Cont'd)

- which is the resonance frequency of the tank circuit, as anticipated. Equating the real part to zero together with Eq. (14.20) gives
  $C2/C1 = gmR$
- For oscillations to start, the loop gain must be made greater than unity, a condition that can be stated in the equivalent form
  $gmR > C2/C1$
- As an example of a practical LC oscillator, we show in Fig. 14.15 the circuit of a Colpitts oscillator, complete with bias details. Here the radio-frequency choke (RFC) provides a high reactance at ω0 but a low dc resistance.

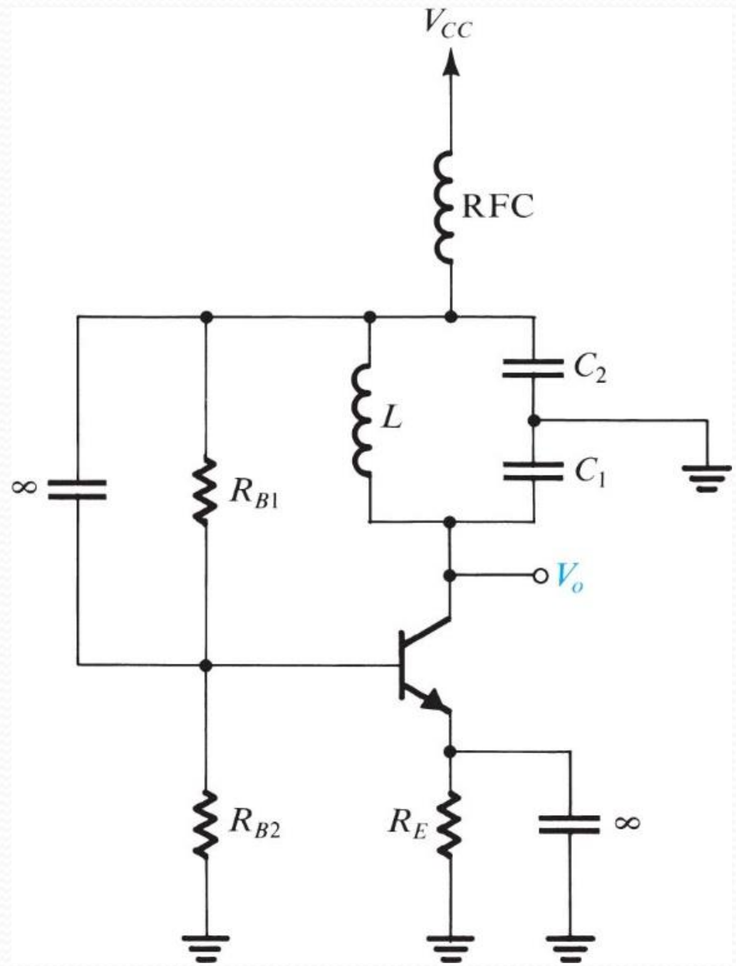


Figure 14.15 Complete circuit for a Colpitts oscillator.

The Cross-Coupled LC Oscillator

- A currently popular LC oscillator circuit suitable for fabrication in IC form and capable of operating at frequencies approaching hundreds of gigahertz (for use in wireless transceivers) is shown in Fig. 14.16(a). The cross coupling simply means that the output of each amplifier drives the input of the other, resulting in the feedback loop shown in Fig. 14.16(b).
- Examination of the feedback loop in Fig. 14.16(b) reveals that at the resonance frequency of each of the two tank circuits (i.e., at $\omega = \omega_0 = 1\sqrt{LC}$), the load of each of Q_1 and Q_2 reduces to a resistance $R_p = \omega_0 LQ$, where Q is the quality factor of the inductance. Taking into consideration the output resistance r_o of each of Q_1 and Q_2, we can write for the gain of each of the two stages at $\omega = \omega_0$, $A_1 = A_2 = -g_m(R_p || r_o)$
- The circuit will provide sustained oscillations at $\omega_0 = 1\sqrt{LC}$ provided $| A_1 A_2 | = g_m(R_p || r_o)]^2 = 1$
- Which reduces to $g_m(R_p || r_o) = 1$

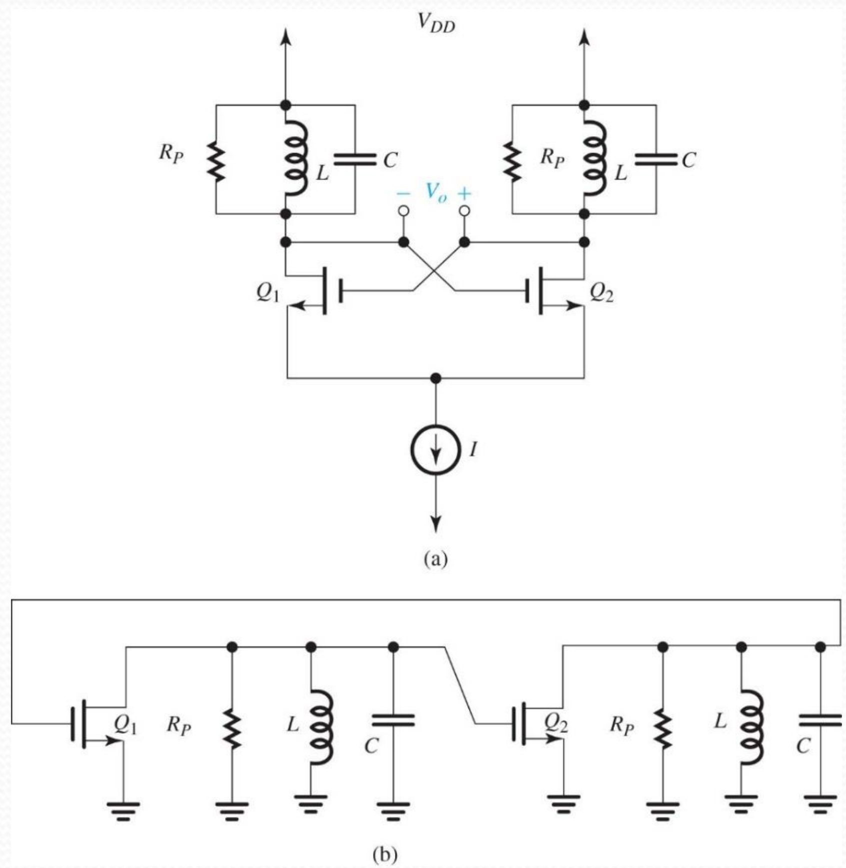


Figure 14.16 (a) The cross-coupled LC oscillator. **(b)** Signal equivalent circuit of the cross-coupled oscillator in **(a)**.

Crystal Oscillators

- A piezoelectric crystal, such as quartz, exhibits electromechanical-resonance characteristics that are very stable (with time and temperature) and highly selective (having very high Q factors). The circuit symbol of a crystal is shown in Fig. 14.17(a), and its equivalent circuit model is given in Fig. 14.17(b).
- Since the Q factor is very high, we may neglect the resistance r and express the crystal impedance as $Z(s) = 1 \Big/ \left[sC_p + \dfrac{1}{sL + 1/sC_s} \right]$ which can be manipulated to the form $Z(s) = \dfrac{1}{sC_p} \dfrac{s^2 + (1/LC_s)}{s^2 + [(C_p + C_s)/LC_sC_p]}$
- The crystal has two resonance frequencies: a series resonance at ω_s $\omega_s = 1/\sqrt{LC_s}$ and a parallel resonance at ω_p $\omega_p = 1 \Big/ \sqrt{L\left(\dfrac{C_sC_p}{C_s + C_p}\right)}$
- Thus for s = jω we can write $Z(j\omega) = -j\dfrac{1}{\omega C_p}\left(\dfrac{\omega^2 - \omega_s^2}{\omega^2 - \omega_p^2}\right)$
- Expressing Z(jω) = jX(ω), the crystal reactance X(ω) will have the shape shown in Fig. 14.17(c). Since C_s is much smaller than the three other capacitances [in Fig. 14.13(a)], it will be dominant and $\omega_0 \simeq 1/\sqrt{LC_s} = \omega_s$

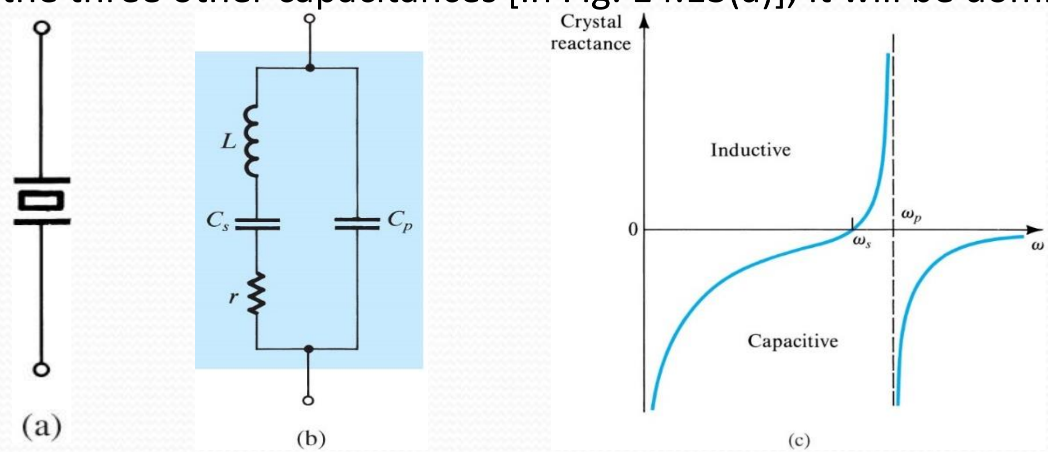


Figure 14.17 A piezoelectric crystal. **(a)** Circuit symbol. **(b)** Equivalent circuit. **(c)** Crystal reactance versus frequency [note that, neglecting the small resistance *r*, $Z_{crystal} = jX(\omega)$].

Crystal Oscillators (Cont'd)

- Figure 14.18 shows a popular configuration (called the Pierce oscillator) utilizing a CMOS inverter (see Section 15.3) as an amplifier. Note that this circuit also is based on the Colpitts configuration.
- The extremely stable resonance characteristics and the very high Q factors of quartz crystals result in oscillators with very accurate and stable frequencies. Unfortunately, however, crystal oscillators, being mechanical resonators, are fixed-frequency circuits.

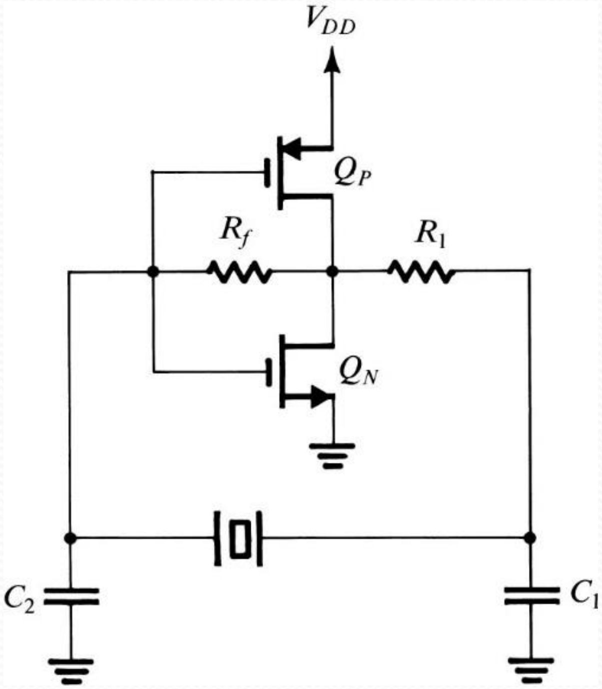


Figure 14.18 A Pierce crystal oscillator utilizing a CMOS inverter as an amplifier.

Bistable Multivibrators
The Feedback Loop

- Nonlinear oscillators or function generators make use of a special class of circuits known as multivibrators. There are three types of multivibrator: bistable, monostable, and astable. The bistable multivibrator has two stable states.
- Bistability can be obtained by connecting a dc amplifier in a positive-feedback loop having a loop gain greater than unity. Such a feedback loop is shown in Fig. 14.19.
- The circuit of Fig. 14.19 has two stable states, one with the op amp in positive saturation and the other with the op amp in negative saturation. Any disturbance, such as that caused by electrical noise, causes the bistable circuit to switch to one of its two stable states. A physical analogy for the operation of the bistable circuit is depicted in Fig. 14.20.

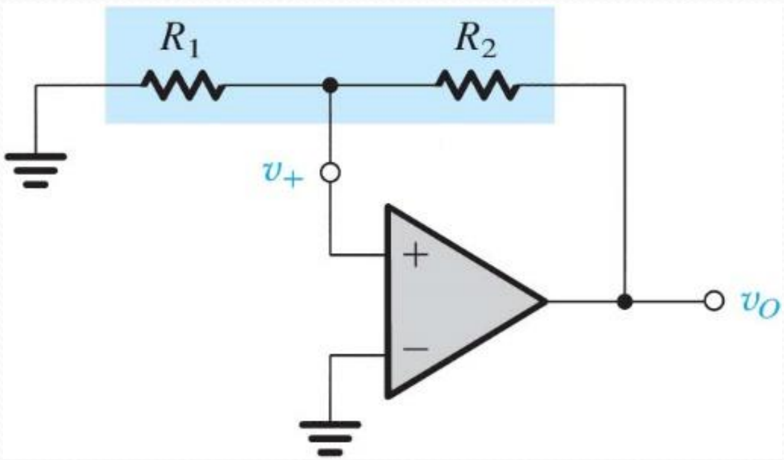


Figure 14.19 A positive-feedback loop capable of bistable operation.

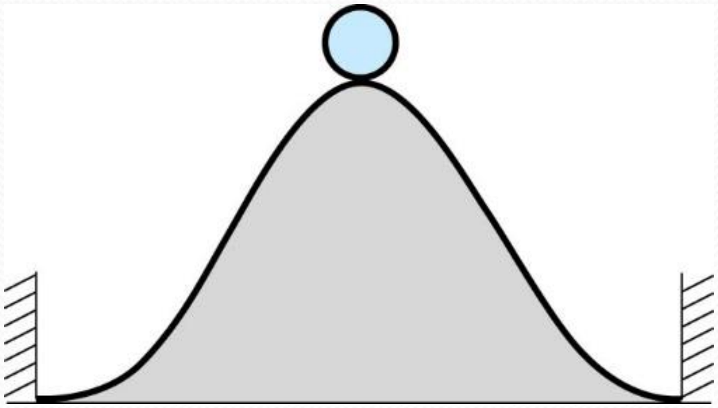


Figure 14.20 A physical analogy for the operation of the bistable circuit. The ball cannot remain at the top of the hill for any length of time (a state of unstable equilibrium or metastability); the inevitably present disturbance will cause the ball to fall to one side or the other, where it can remain indefinitely (the two stable states).

Transfer Characteristics of the Bistable Circuit

- The question naturally arises as to how we can make the bistable circuit of Fig. 14.19 change state.
- Figure 14.21(a) shows the bistable circuit with a voltage v_I applied to the inverting input terminal of the op amp. Figure 14.21(b) shows the transfer characteristic for increasing v_I. The transfer characteristic for decreasing v_I is shown in Fig. 14.21(c). The complete transfer characteristics, v_O–v_I, of the circuit in Fig. 14.21(a) can be obtained by combining the characteristics in Fig. 14.21(b) and (c), as shown in Fig. 14.21(d). As indicated, the circuit changes state at different values of v_I, depending on whether v_I is increasing or decreasing. Thus the circuit is said to exhibit hysteresis ; the width of the hysteresis is the difference between the high threshold V_{TH} and the low threshold V_{TL}. Also note that the bistable circuit is in effect a comparator with hysteresis.

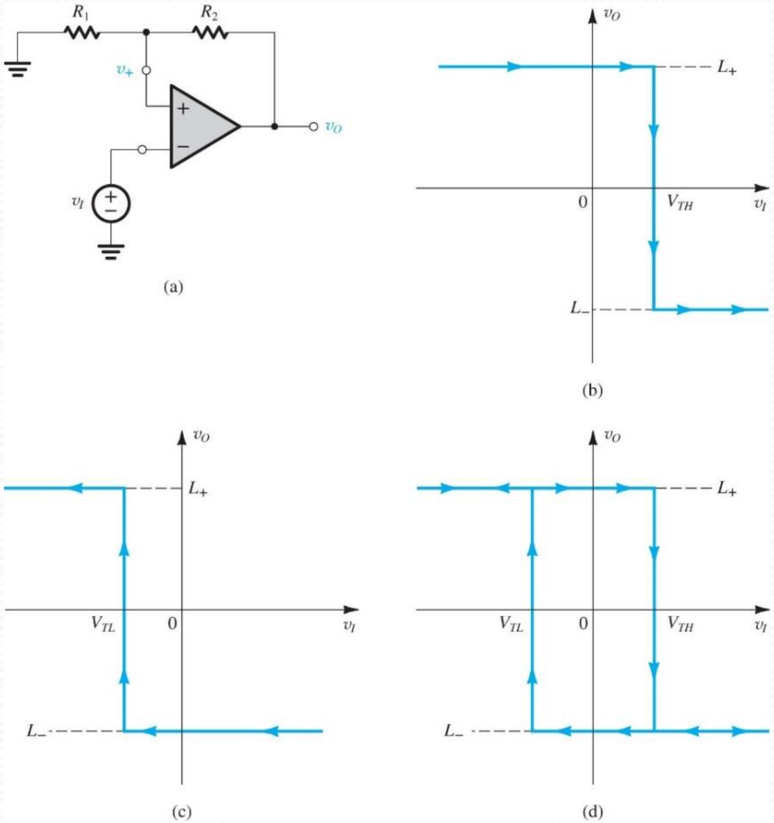


Figure 14.21 (a) The bistable circuit of Fig. 14.19 with the negative input terminal of the op amp disconnected from ground and connected to an input signal v_I. **(b)** The transfer characteristic of the circuit in **(a)** for increasing v_I. **(c)** The transfer characteristic for decreasing v_I. **(d)** The complete transfer characteristics.

Triggering the Bistable Circuit
The Bistable Circuit as a Memory Element

- It is important to note that the input v_I merely initiates or triggers regeneration. Thus we can remove v_I with no effect on the regeneration process. In other words, v_I can be simply a pulse of short duration. The input signal v_I is thus referred to as a trigger signal, or simply a trigger.
- The characteristics of Fig. 14.21(d) indicate also that the bistable circuit can be switched to the positive state ($v_O = L_+$) by applying a negative trigger signal v_I of magnitude greater than that of the negative threshold V_{TL}.
- We observe from Fig. 14.21(d) that for input voltages in the range $V_{TL} < v_I < V_{TH}$, the output can be either L_+ or L_-, depending on the state that the circuit is already in. Thus the circuit exhibits memory. the bistable circuit is also known as a Schmitt trigger.

A Bistable Circuit with Noninverting Transfer Characteristics

- By applying the input signal v_I (the trigger signal) to the terminal of R_1 that is connected to ground in Fig. 14.19, the resulting circuit is shown in Fig. 14.22(a).
- $$v_+ = v_I \frac{R_2}{R_1 + R_2} + v_O \frac{R_1}{R_1 + R_2}$$
- $$V_{TL} = -L_+(R_1/R_2)$$
- $$V_{TH} = -L_-(R_1/R_2)$$
- The complete transfer characteristic of the circuit of Fig. 14.22(a) is displayed in Fig. 14.22(b). Observe that a positive triggering signal v_I (of value greater than V_{TH}) causes the circuit to switch to the positive state (v_O goes from L_- to L_+). Thus the transfer characteristic of this circuit is noninverting.

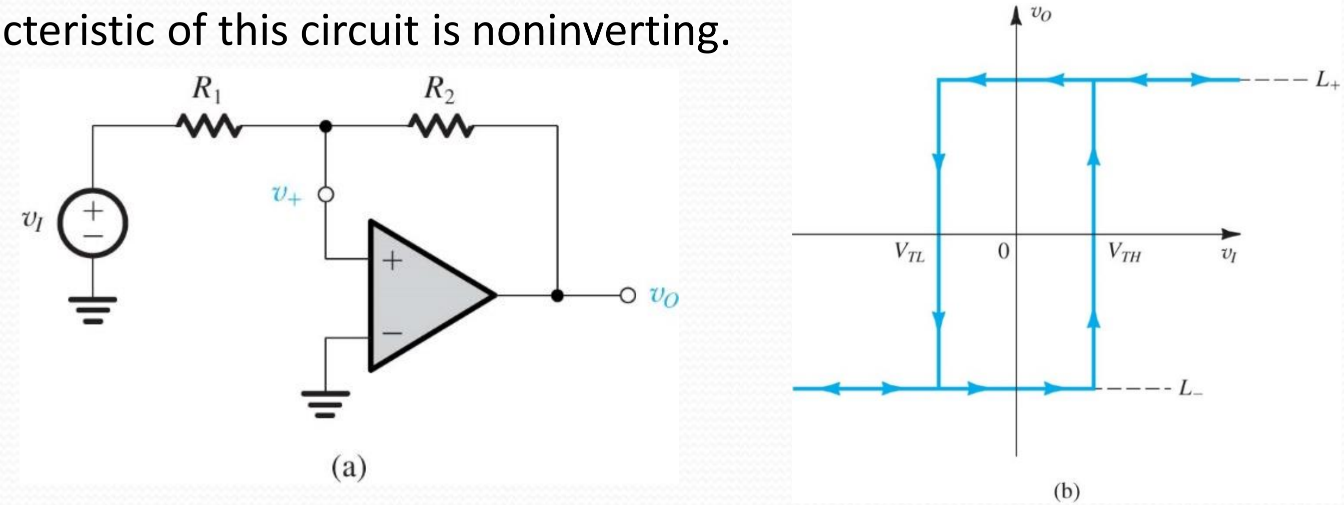
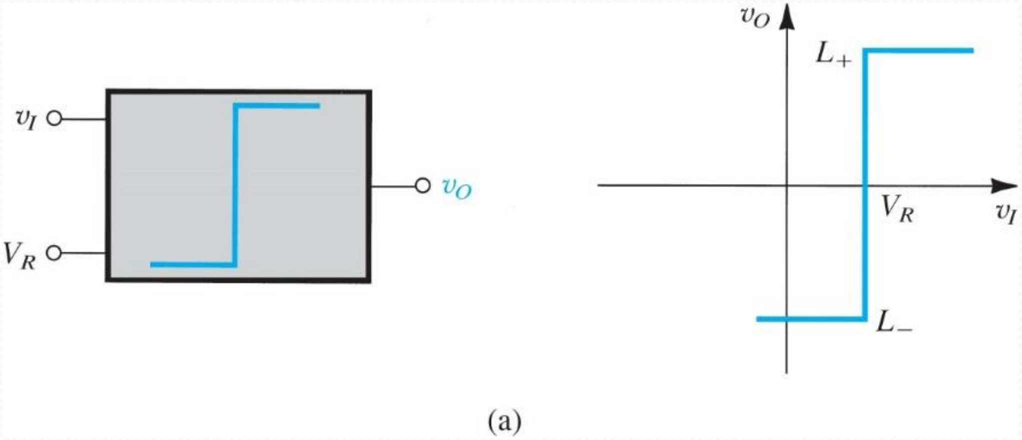


Figure 14.22 (a) A bistable circuit derived from the positive-feedback loop of Fig. 14.19 by applying v_I through R_1. **(b)** The transfer characteristic of the circuit in **(a)** is noninverting. [Compare it to the inverting characteristic in Fig. 14.21(d).]

Application of the Bistable Circuit as a Comparator

- Although one normally thinks of the comparator as having a single threshold value [see Fig. 14.23(a)], it is useful in many applications to add hysteresis to the comparator characteristics as indicated in Fig. 14.23(b).



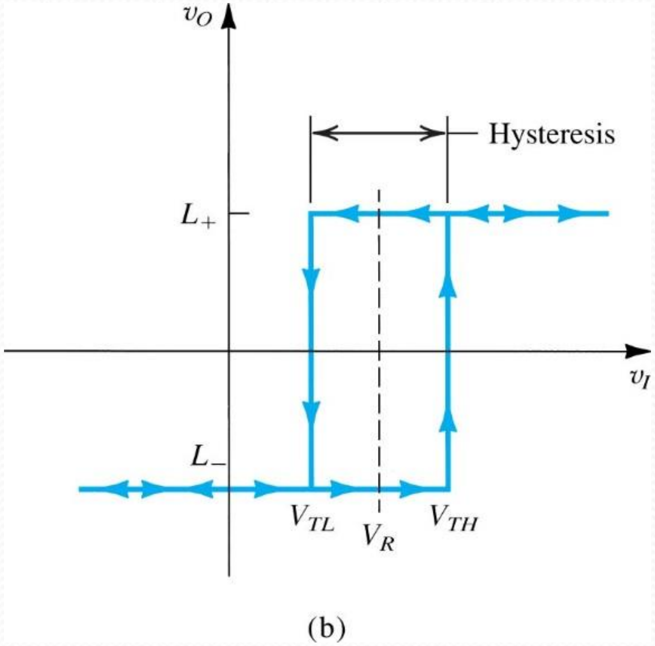


Figure 14.23 (a) Block diagram representation and transfer characteristic for a comparator having a reference, or threshold, voltage V_R. **(b)** Comparator characteristic with hysteresis.

Application of the Bistable Circuit as a Comparator(Cont'd)

- Imagine now what happens if the signal being processed has—as it usually does have—interference superimposed on it, say of a frequency much higher than that of the signal. It follows that the signal might cross the zero axis a number of times around each of the zero-crossing points we are trying to detect, as shown in Fig. 14.24. The situation is illustrated in Fig. 14.24, from which we see that including hysteresis in the comparator characteristics provides an effective means for rejecting interference (thus providing another form of filtering).

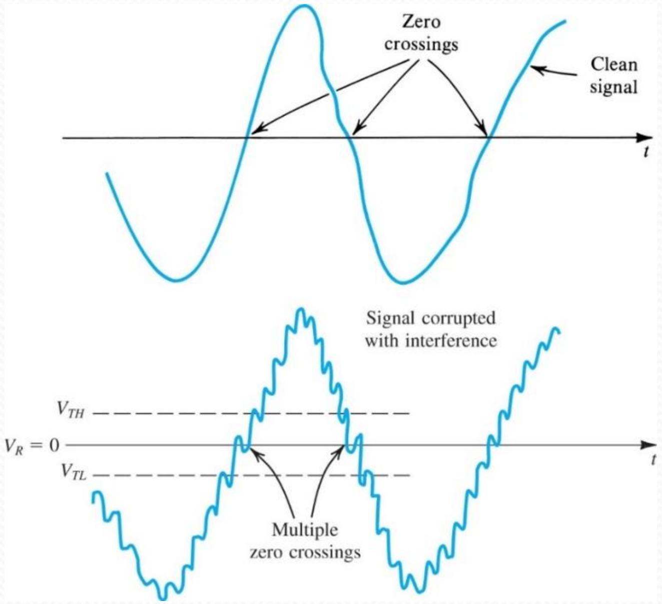


Figure 14.24 Illustrating the use of hysteresis in the comparator characteristics as a means of rejecting interference.

Making the Output Levels More Precise

- The output levels of the bistable circuit can be made more precise than the saturation voltages of the op amp are by cascading the op amp with a limiter circuit. Two such arrangements are shown in Fig. 14.25.

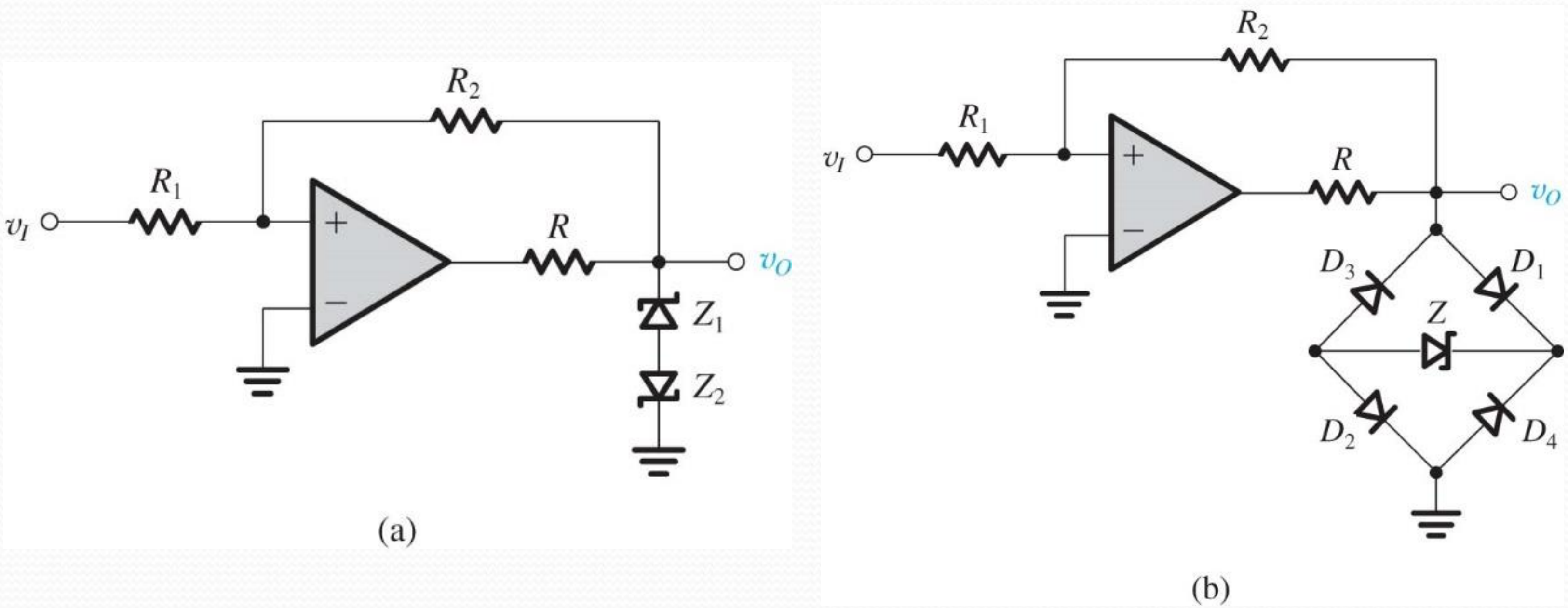
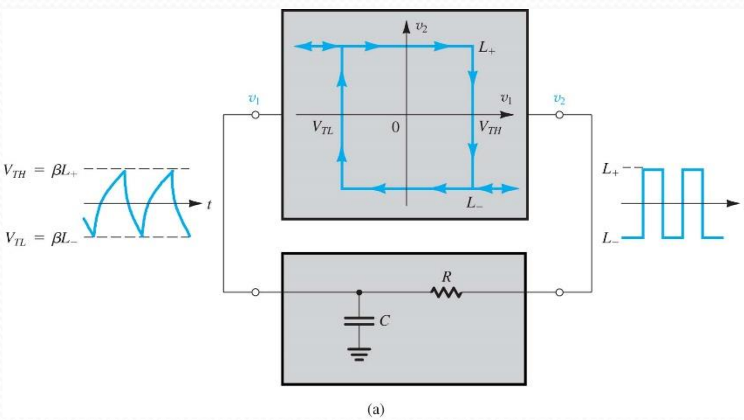
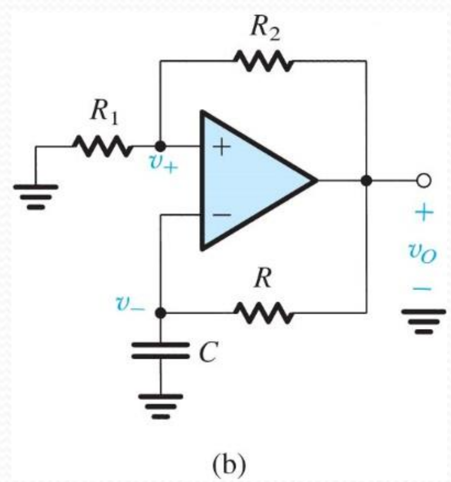


Figure 14.25 Limiter circuits are used to obtain more precise output levels for the bistable circuit. In both circuits the value of *R* should be chosen to yield the current required for the proper operation of the zener diodes. **(a)** For this circuit $L_+ = V_{Z1} + V_D$ and $L_- = -(V_{Z2} + V_D)$, where V_D is the forward diode drop. **(b)** For this circuit $L_+ = V_Z + V_{D1} + V_{D2}$ and $L_- = -(V_Z + V_{D3} + V_{D4})$

Generation of Square and Triangular Waveforms Using Astable Multivibrators

- A square waveform can be generated by arranging for a bistable multivibrator to switch states periodically as shown in Fig. 14.26(a). The bistable multivibrator has an inverting transfer characteristic and can thus be realized using the circuit of Fig. 14.21(a). This results in the circuit of Fig. 14.26(b). We shall show shortly that this circuit has no stable states and thus is appropriately named an astable multivibrator.





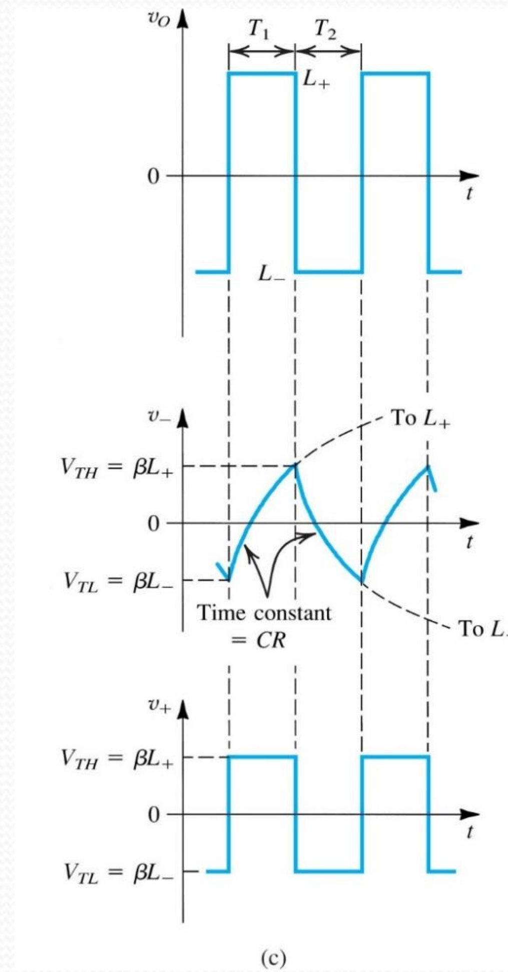


Figure 14.26 (a) Connecting a bistable multivibrator with inverting transfer characteristics in a feedback loop with an RC circuit results in a square-wave generator. (b) The circuit obtained when the bistable multivibrator is implemented with the circuit of Fig. 14.21(a). (c) Waveforms at various nodes of the circuit in (b). This circuit is called anastable multivibrator.

Operation of the Astable Multivibrator

- To see how the astable multivibrator operates, refer to Fig. 14.26(b) and let the output of the bistable multivibrator be at one of its two possible levels, say L_+. From the preceding description we see that the astable circuit oscillates and produces a square waveform at the output of the op amp. This waveform, and the waveforms at the two input terminals of the op amp, are displayed in Fig. 14.26(c). The period T of the square wave can be found as follows:
  - $T_1 = \tau \ln \frac{1-\beta(L_-/L_+)}{1-\beta}$ $\qquad T_2 = \tau \ln \frac{1-\beta(L_+/L_-)}{1-\beta}$
  - T = T1 + T2 , $\qquad T = 2\tau \ln \frac{1+\beta}{1-\beta}$
- Although the astable circuit has no stable states, it has two quasi-stable states and remains in each for a time interval determined by the time constant of the RC network and the thresholds of the bistable multivibrator.

Generation of Triangular Waveforms

- The exponential waveforms generated in the astable circuit of Fig. 14.24 can be changed to triangular by replacing the low-pass RC circuit with an integrator. The resulting circuit is shown in Fig. 14.27(a).
- Let the output of the bistable circuit be at L+. A current equal L+/R to will flow into the resistor R and through capacitor C, causing the output of the integrator to linearly decrease with a slope of -L+/CR as shown in Fig. 14.27(c).
- During the interval T1 we have, from Fig. 14.27(c), $\frac{V_{TH}-V_{TL}}{T_1}=\frac{L_+}{CR}$

  from which we obtain $T_1 = CR\frac{V_{TH}-V_{TL}}{L_+}$
- Similarly, during T2 we have $\frac{V_{TH}-V_{TL}}{T_2}=\frac{-L_-}{CR}$

  from which we obtain $T_2 = CR\frac{V_{TH}-V_{TL}}{-L_-}$

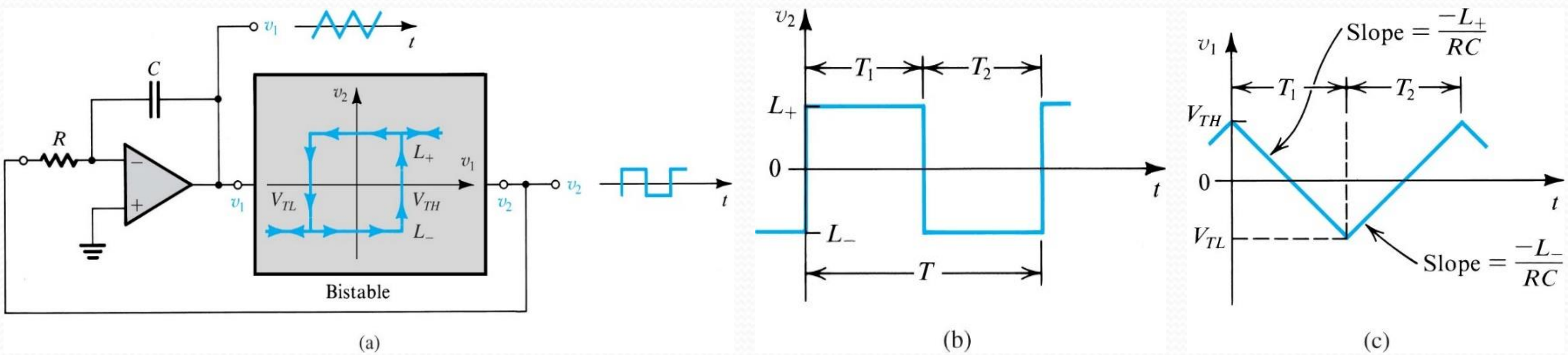


Figure 14.27 A general scheme for generating triangular and square waveforms.

Generation of a Standardized Pulse—The Monostable Multivibrator

- The monostable multivibrator has one stable state in which it can remain indefinitely. It also has a quasi-stable state to which it can be triggered and in which it stays for a predetermined interval equal to the desired width of the output pulse. The action of the monostable multivibrator has given rise to its alternative name, the one shot.
- Figure 14.28(a) shows an op-amp monostable circuit. From Fig. 14.28(b), we observe that a negative pulse is generated at the output during the quasi-stable state. The duration T of the output pulse is determined from the exponential waveform of v_B,

$$v_B(t) = L_- - (L_- - V_{D1})e^{-t/C_1R_3}$$

by substituting $v_B(T) = \beta L_-$, $\beta L_- = L_- - (L_- - V_{D1})e^{-T/C_1R_3}$

which yields $T = C_1R_3 \ln\left(\frac{V_{D1} - L_-}{\beta L_- - L_-}\right)$

- For $V_{D1} \ll |L_-|$, this equation can be approximated by $T \simeq C_1R_3 \ln\left(\frac{1}{1-\beta}\right)$
- Note that the monostable circuit should not be triggered again until capacitor C_1 has been recharged to V_{D1}; otherwise the resulting output pulse will be shorter than normal. This recharging time is known as the recovery period. Circuit techniques exist for shortening the recovery period.

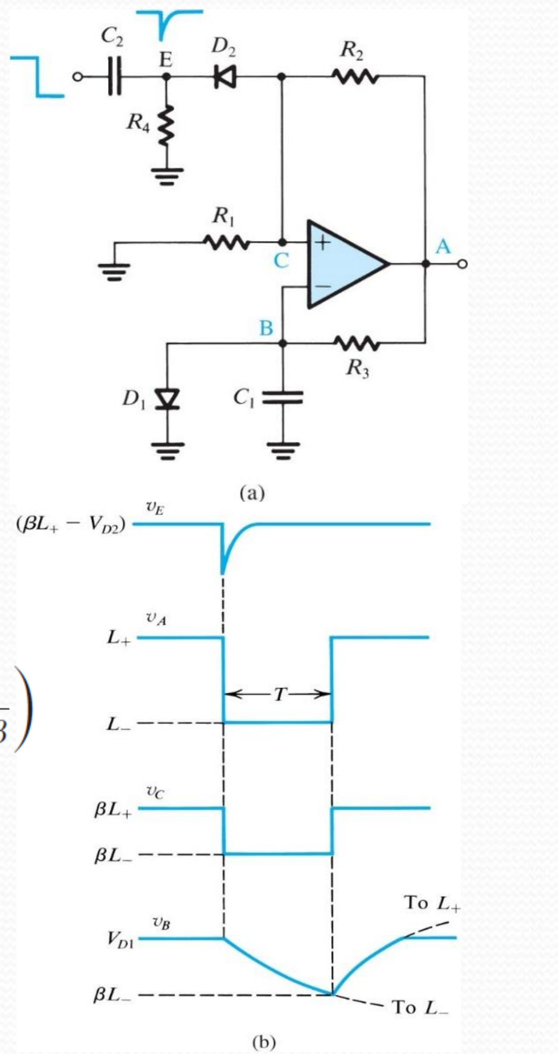


Figure 14.28 (a) An op-amp monostable circuit. **(b)** Signal waveforms in the circuit of **(a)**.

Integrated-Circuit Timers
The 555 Circuit

- In this section we discuss the most popular of such ICs, the 555 timer.
- Figure 14.29 shows a block diagram representation of the 555 timer circuit.

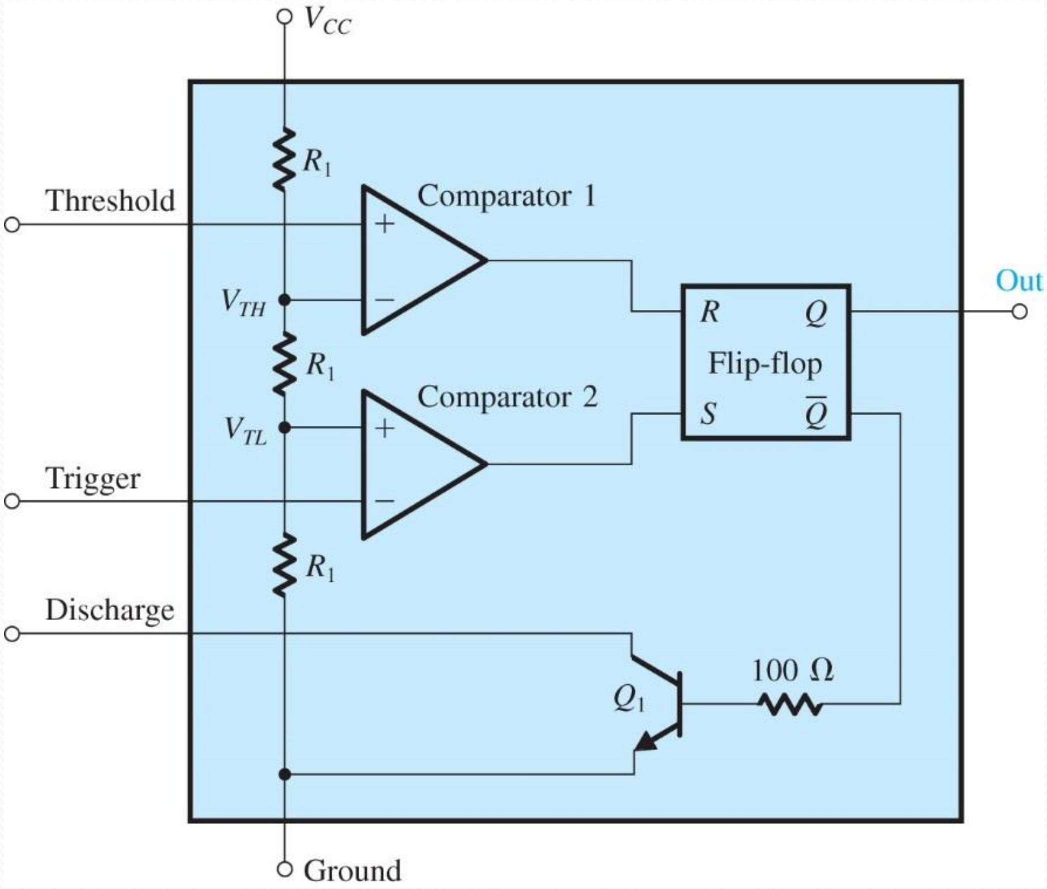
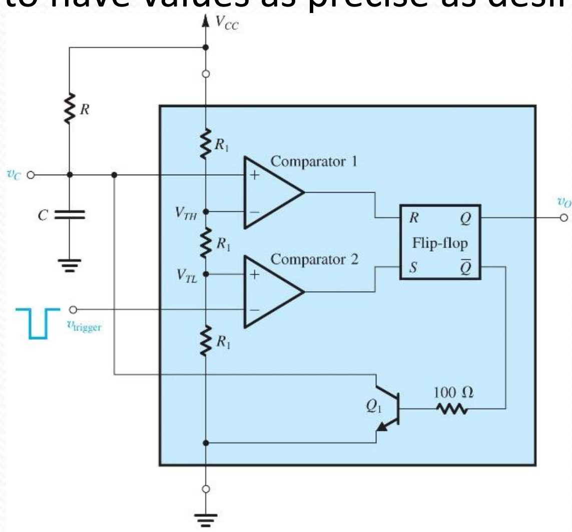


Figure 14.29 A block diagram representation of the internal circuit of the 555 integrated-circuit timer.

Implementing a Monostable Multivibrator Using the 555 IC

- Figure 14.30(a) shows a monostable multivibrator implemented using the 555 IC together with an external resistor R and an external capacitor C. To trigger the monostable multivibrator, a negative input pulse is applied to the trigger input terminal. The monostable multivibrator produces an output pulse v_O as indicated in Fig. 14.30(b). The width of the pulse, T, is the time interval that the monostable multivibrator spends in the quasi-stable state; it can be determined by reference to the waveforms in Fig. 14.30(b) as follows: Denoting the instant at which the trigger pulse is applied as t = 0, the exponential waveform of v_C can be expressed as $v_C = V_{CC}\left(1 - e^{-t/CR}\right)$
- Substituting at $v_C = V_{TH} = \frac{2}{3}V_{CC}$ t = T gives T = CR ln 3 ≅ 1.1CR
- Thus the pulse width is determined by the external components C and R, which can be selected to have values as precise as desired.



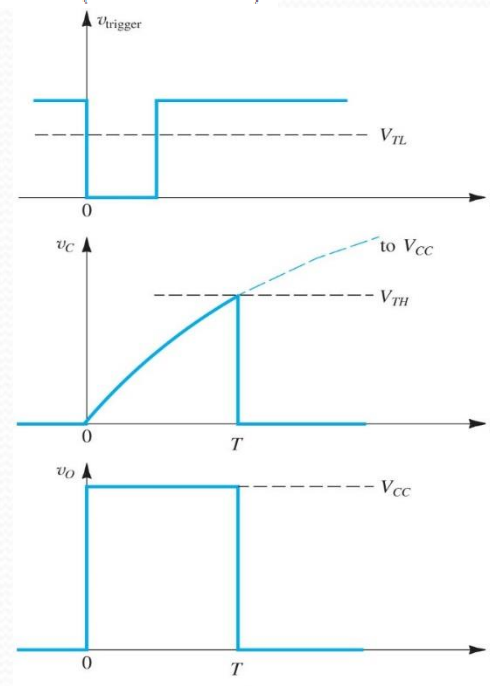


Figure 14.30 **(a)** The 555 timer connected to implement a monostable multivibrator. **(b)** Waveforms of the circuit in **(a)**.

An Astable Multivibrator Using the 555 IC

- Figure 14.31(a) shows the circuit of an astable multivibrator employing a 555 IC, two external resistors, R_A and R_B, and an external capacitor C. To see how the circuit operates, refer to the waveforms depicted in Fig. 14.31(b).
- Reference to Fig. 14.31(b) indicates that the output will be high during the interval T_H, in which v_C rises from V_{TL} to V_{TH}. The exponential rise of v_C can be described by

$$v_C = V_{CC} - (V_{CC} - V_{TL})e^{-t/C(R_A+R_B)}$$

where t = 0 is the instant at which the interval T_H begins.
Substituting results $v_C = V_{TH} = \frac{2}{3}V_{CC}$ at $t = T_H$ and $V_{TL} = \frac{1}{3}V_{CC}$
results in $T_H = C(R_A + R_B)\ln 2 \approx 0.69\, C(R_A + R_B)$

- The exponential fall of v_C can be described by $v_C = V_{TH}e^{-t/CR_B}$

where we have taken t = 0 as the beginning of the interval T_L.
Substituting $v_C = V_{TL} = \frac{1}{3}V_{CC}$ at $t = T_L$ and $V_{TH} = \frac{2}{3}V_{CC}$ results in

$$T_L = C\, R_B \ln 2 \approx 0.69\, C\, R_B$$

$$T = T_H + T_L = 0.69\, C(R_A + 2R_B)$$

- Also, the duty cycle of the output square wave can be found from Eqs. (14.43) and (14.45):

$$\text{Duty cycle} \equiv \frac{T_H}{T_H + T_L} = \frac{R_A + R_B}{R_A + 2R_B}$$

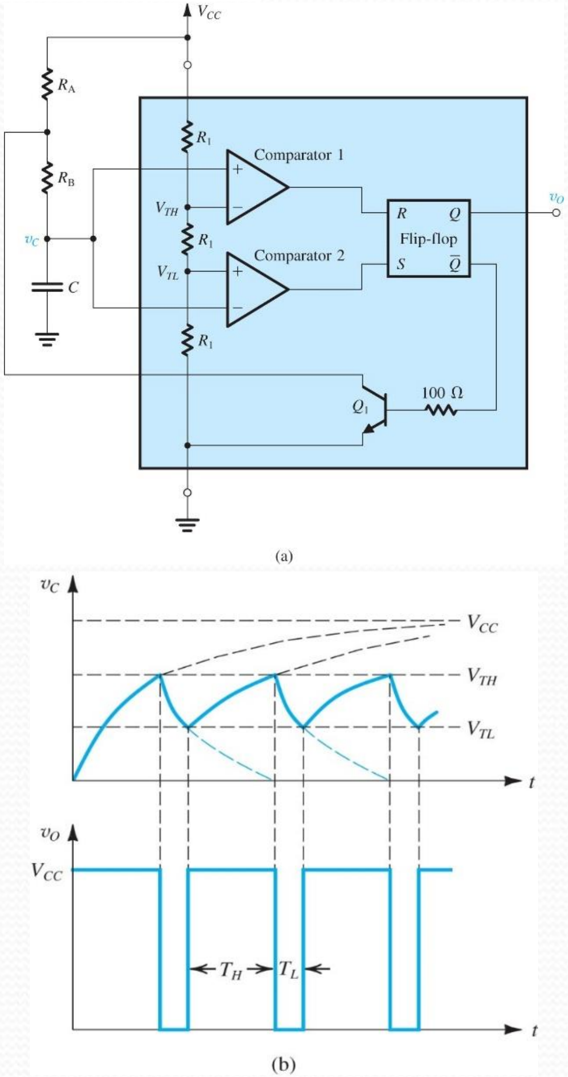


Figure 14.31 (a) The 555 timer connected to implement an astable multivibrator. **(b)** Waveforms of the circuit in **(a)**.

Nonlinear Waveform-Shaping Circuits
The Breakpoint Method

- Diodes or transistors can be combined with resistors to synthesize two-port networks having arbitrary nonlinear transfer characteristics. Such two-port networks can be employed in **waveform shaping**—that is, changing the waveform of an input signal in a prescribed manner to produce a waveform of a desired shape at the output. In this section we illustrate this application by a concrete example: the **sine-wave shaper**.
- In the breakpoint method the desired nonlinear transfer characteristic (in our case the sine function shown in Fig. 14.32) is implemented as a piecewise linear curve. Diodes are utilized as switches that turn on at the various breakpoints of the transfer characteristic, thus switching into the circuit additional resistors that cause the transfer characteristic to change slope.

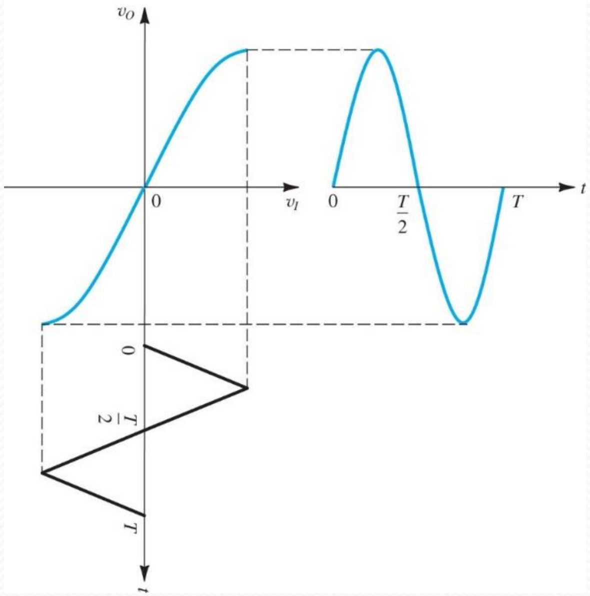


Figure 14.32 Using a nonlinear (sinusoidal) transfer characteristic to shape a triangular waveform into a sinusoid.

The Breakpoint Method (Cont'd)

- Consider the circuit shown in Fig. 14.33(a). Let the input be the triangular wave shown in Fig. 14.33(b)
- For $v_I > V_1$, $v_O = V_1 + (v_I - V_1)\dfrac{R_5}{R_4 + R_5}$
- This implies that as the input continues to rise above V_1, the output follows, but with a reduced slope. This gives rise to the second segment in the output waveform, as shown in Fig. 14.33(b).
- Although the circuit is relatively simple, its performance is surprisingly good. A measure of goodness usually taken is to quantify the purity of the output sine wave by specifying the percentage total harmonic distortion (THD). Although the circuit is relatively simple, its performance is surprisingly good. A measure of goodness usually taken is to quantify the purity of the output sine wave by specifying the percentage total harmonic distortion (THD).

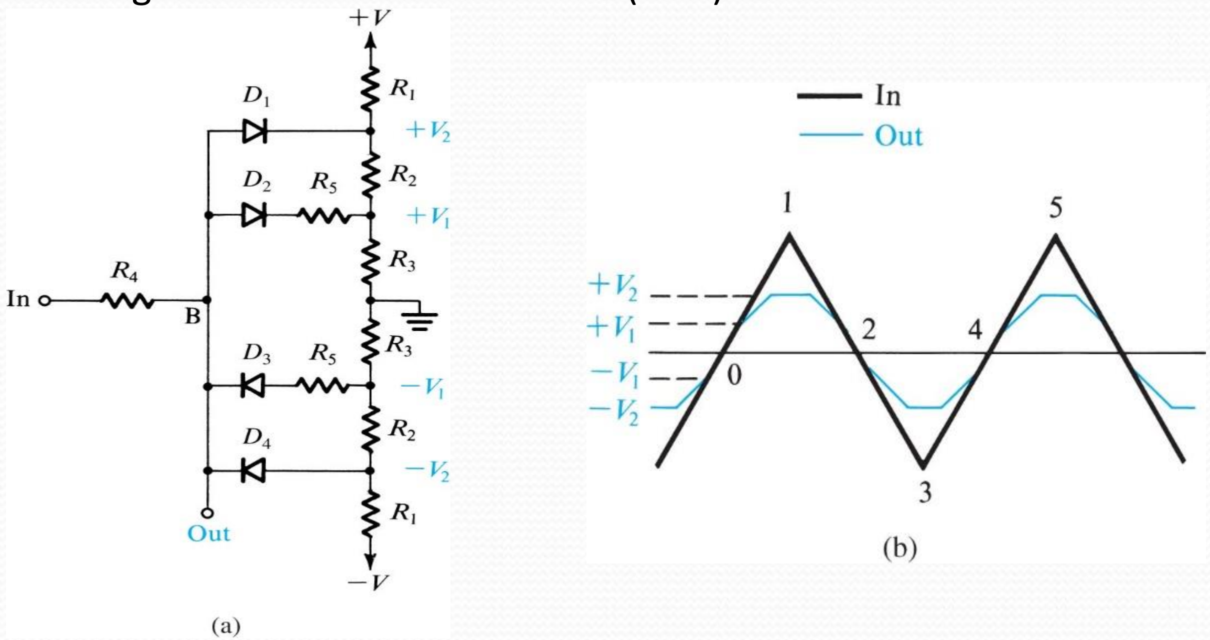


Figure 14.33 (a) A three-segment sine-wave shaper. **(b)** The input triangular waveform and the output approximately sinusoidal waveform.

The Nonlinear-Amplification Method

- The other method we discuss for the conversion of a triangular wave into a sine wave is based on feeding the triangular wave to the input of an amplifier having a nonlinear transfer characteristic that approximates the sine function. One such amplifier circuit consists of a differential pair with a resistance connected between the two emitters, as shown in Fig. 14.34.

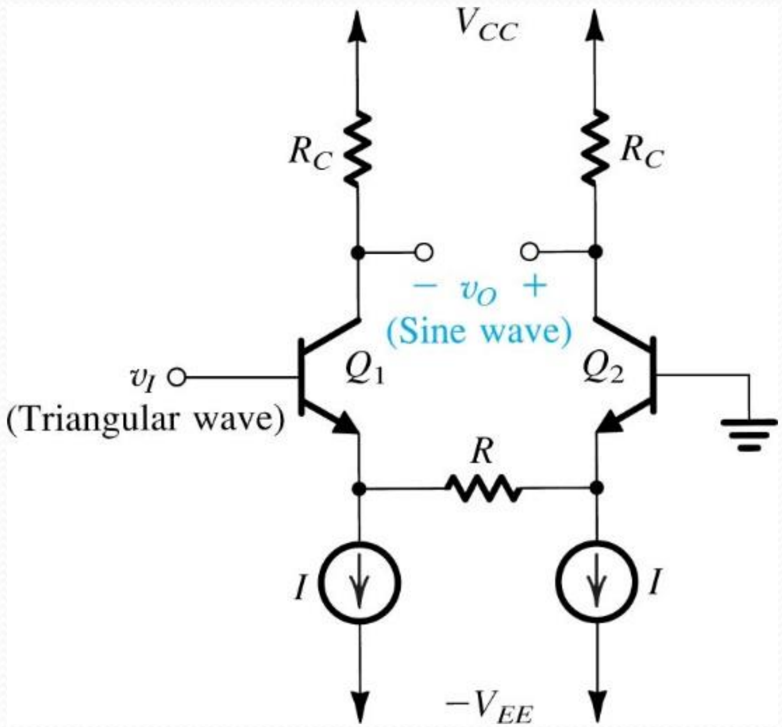


Figure 14.34 A differential pair with an emitterdegeneration resistance used to implement a triangular-wave to sine-wave converter. Operation ofthe circuit can be graphically described by Fig. 14.32.