Chapter 13. Signal Generators and Waveform-Shaping Circuits

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INTRODUCTION

13.1 Basic Principles of Sinusoidal Osillators

13.2 Op-Amp-RC Oscillator Circuits

13.3 LC and Crystal Oscillators

13.4 Bistable Multivibrators

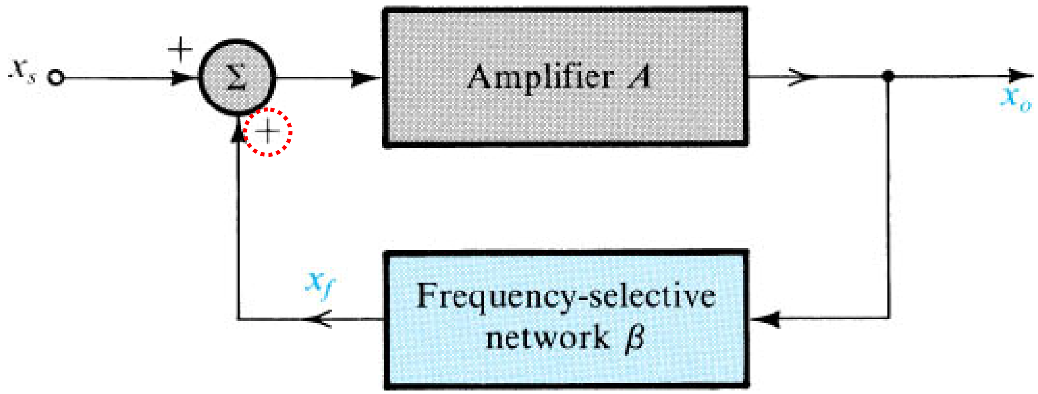
13.5 Generation of Square and Triangular Waveforms Using Astable Multivibrators

13.1 BASIC PRINCIPLES OF SINUSOIDAL OSCILLATORS

- Oscillators
  - **linear oscillator** : sinusoidal signal ← Need frequency selective feedback
  - **nonlinear oscillator or multi-vibrator** : square/triangular/pulse generators ← Op amp construction

- Analysis methods
  - Fundamentally **nonlinear** circuit → cannot use s-transform or any linear theory, in principle.
  - Compromised method : frequency domain linear analysis to start oscillation + nonlinear mechanism for amplitude control.

13.1.1 Oscillator Feedback Loop for Linear Oscillator



• Sinusoidal oscillator = Amplifier, ***A*** + Frequency-selective network, ***ß*** connected in a positive-feedback loop.

• In an actual oscillator circuit, no input signal will be present.

• The loop gain(Chapter 8) of the circuit is –A(s)ß(s). However, for our purposes here it is more convenient to drop the minus sign

$$A_f(s) = \frac{A(s)}{1 - A(s)\beta(s)}$$

$$L(s) \equiv A(s)\beta(s)$$

• The characteristic equation : $1 - L(s) = 0$

13.1.2 The Oscillation Criterion

• Steady-state analysis : If at a specific frequency f_0 the loop gain Aß is equal to unity, A_f will be infinite.

$$A_f(s) = \frac{A(s)}{1 - A(s)\beta(s)}$$

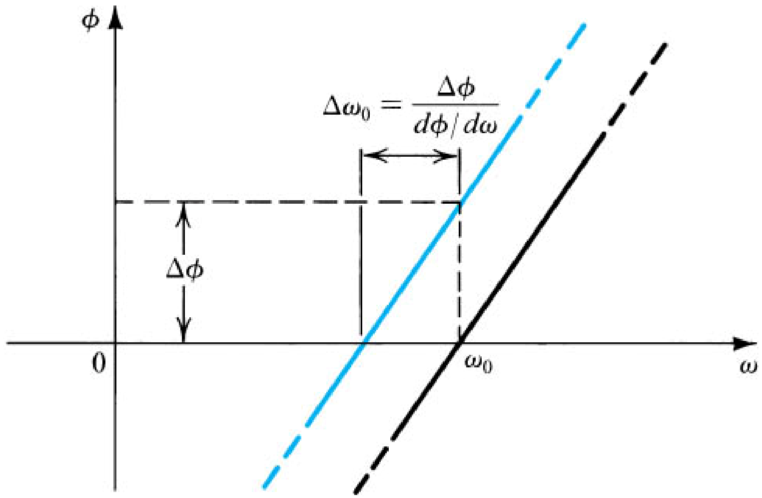
• At this frequency the circuit will have a finite output <u>for zero input signal</u> → Definition of oscillator

• **Barkhausen criterion** : at ω_0 *the phase of the loop gain should be zero and the magnitude of the loop gain should be unity* for zero input signal.

$$\boxed{L(j\omega_0) \equiv A(j\omega_0)\beta(j\omega_0) = 1}$$

* Very important to guarantee this condition be met only at a single frequency

How to Select a Single Oscillation Frequency



- **Barkhausen criterion** $\boxed{L(j\omega_0) \equiv A(j\omega_0)\beta(j\omega_0) = 1}$
  - **Amplitude :** single frequency is impossible (active tuned amplifier with infinite Q)
  - **Phase** : most reactance is a strong function of frequency X(ω)

- Oscillation **Stability**
  - For given ΔΦ fluctuation, steep dΦ/dω results in a smaller ω_0 change

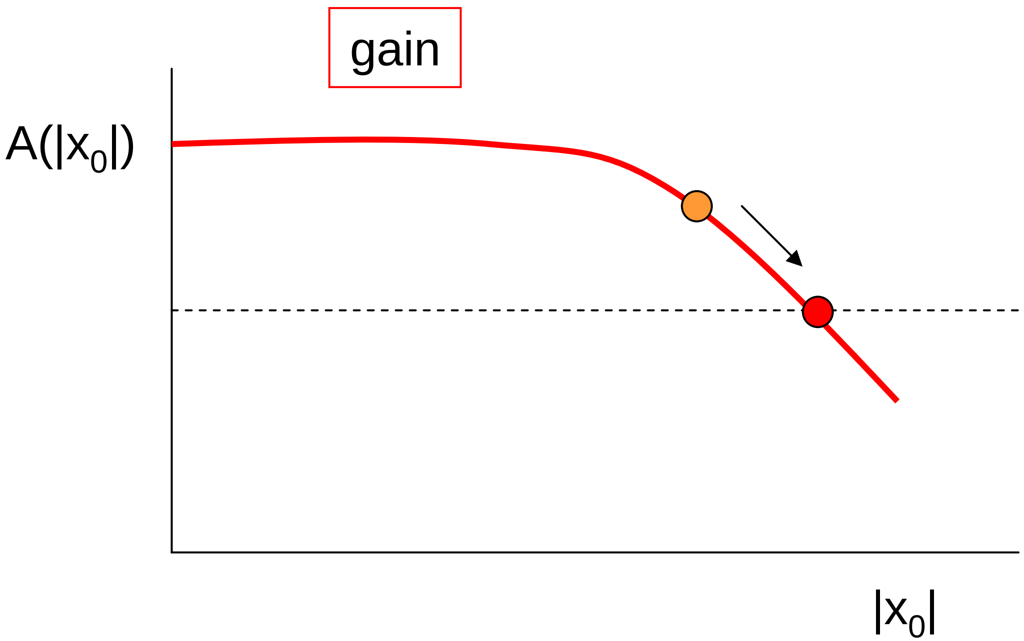
Steady-State Oscillation and Stability

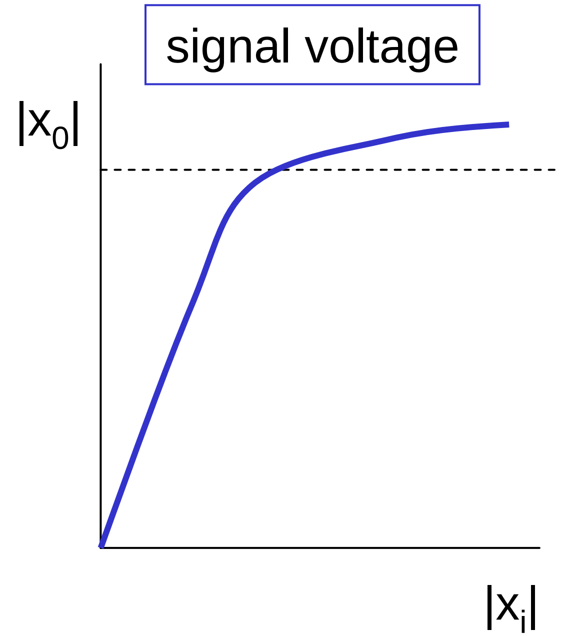
• **Problem** : the parameters of any physical system cannot be maintained constant for any length of time (due, for example, to temperature).

→ Aß becomes slightly less than unity : oscillation will cease.

→ Aß exceeds unity : oscillation will grow in amplitude.

• **Solution** : nonlinear gain





13.1.3 Nonlinear Amplitude Control

• **Implementation** of the nonlinear amplitude-stabilization mechanism

1. Limiter circuit (Chapter 3, p184~187)
2. Controlled resistance element (JFET, diode) in the feedback circuit
3. Inherent gain saturation of transistor

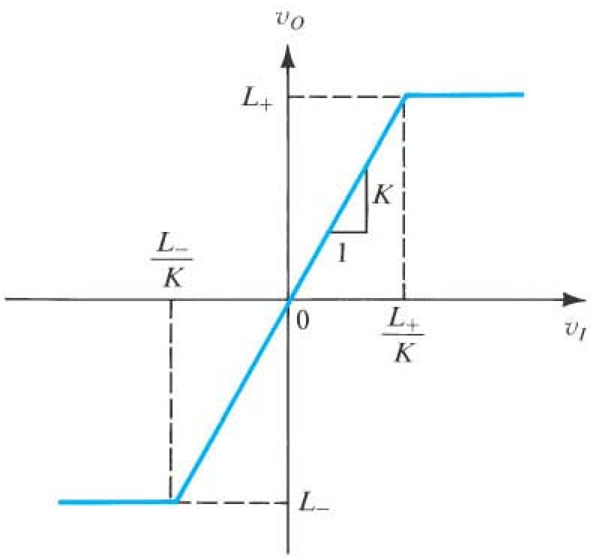


Figure 3.32 General transfer characteristic for a limiter circuit.

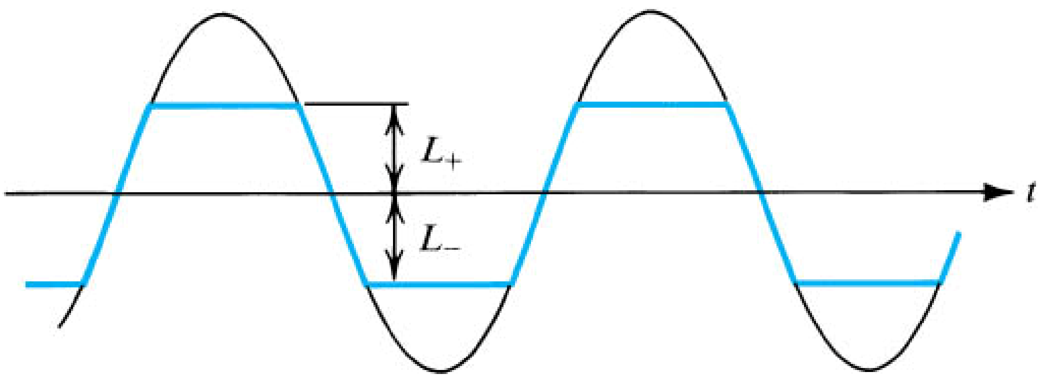


Figure 3.33 Applying a sine wave to a limiter can result in clipping off its two peaks.

double limiter & hard limiter

Nonlinear Amplitude Control by Limiter using Diodes

- **Implementation** of the nonlinear amplitude-stabilization mechanism
  1. Limiter circuit(Chapter 3, p184~187)

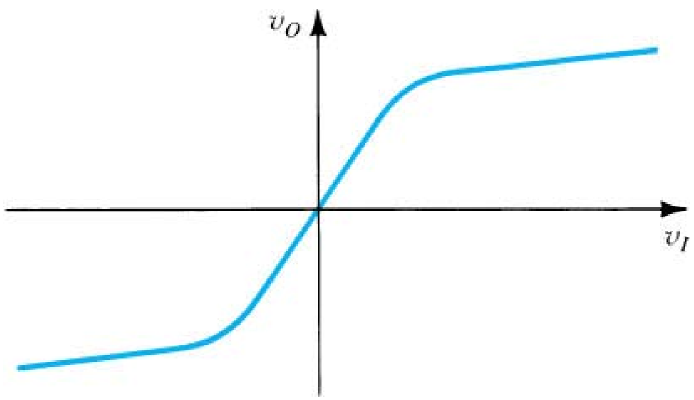


Figure 3.34 Soft limiting.

double limiter & soft limiter

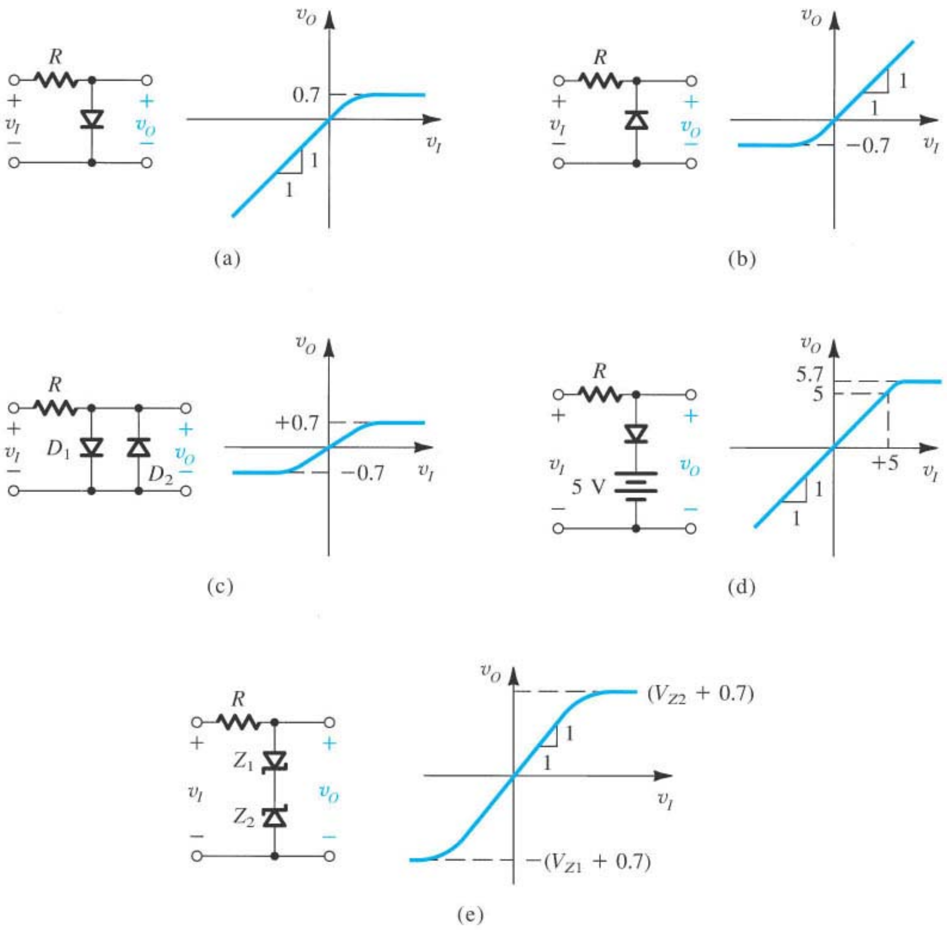


Figure 3.35 A variety of basic limiting circuits.

“Hard” vs “Soft” Limiting

- Hard limiting
  - **Very high stability**
  - **But, excessive distortion**

- Signal recovery
  - Filtering action of frequency-selective feedback network
  - Ultimate sinusoidal signal required → need bandpass filter with highest Q

13.1.4 A Popular Limiter Circuit

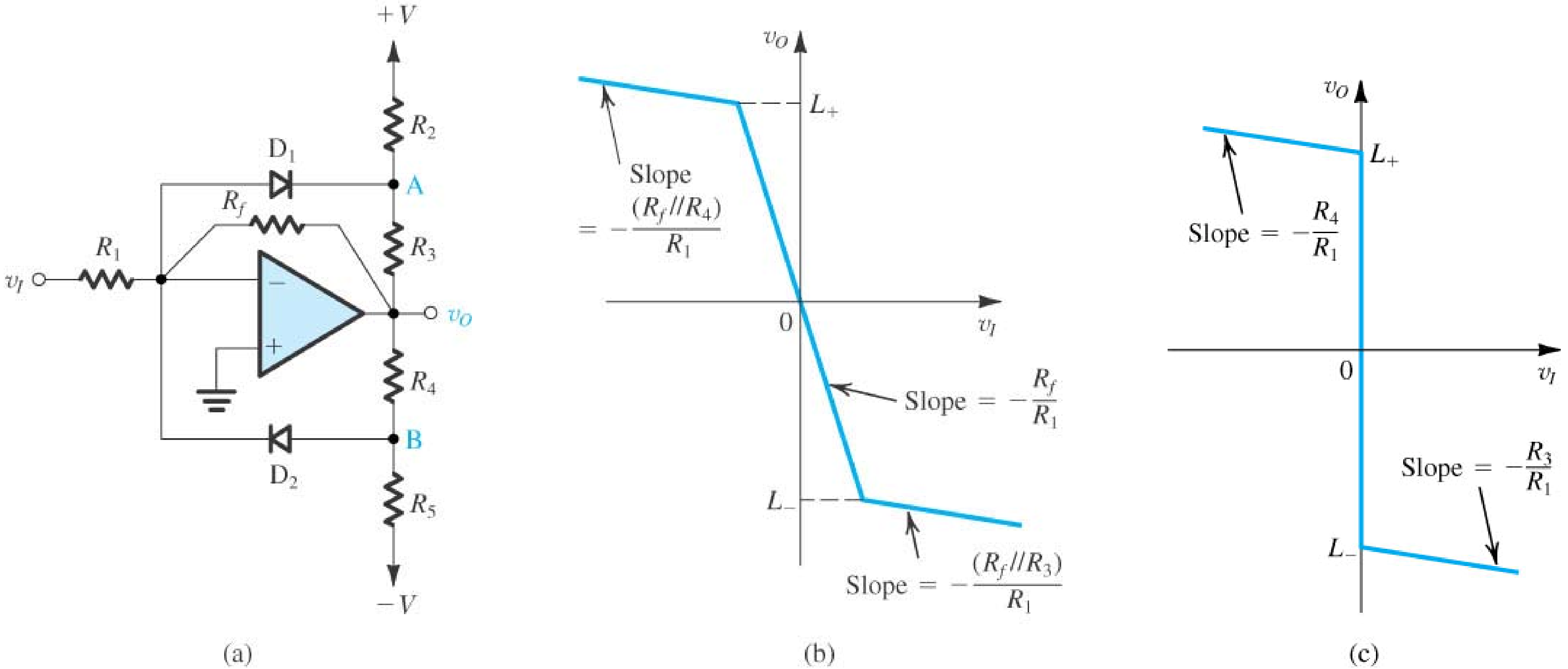
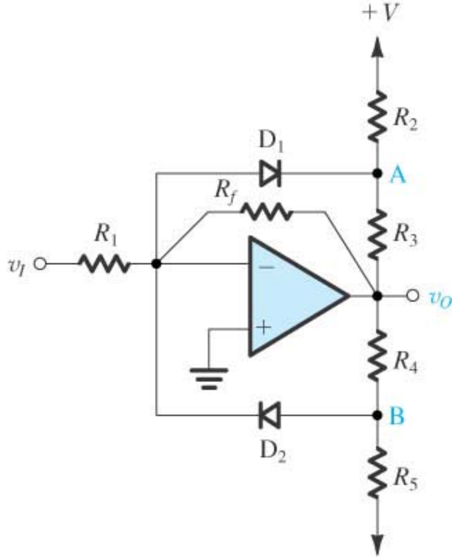
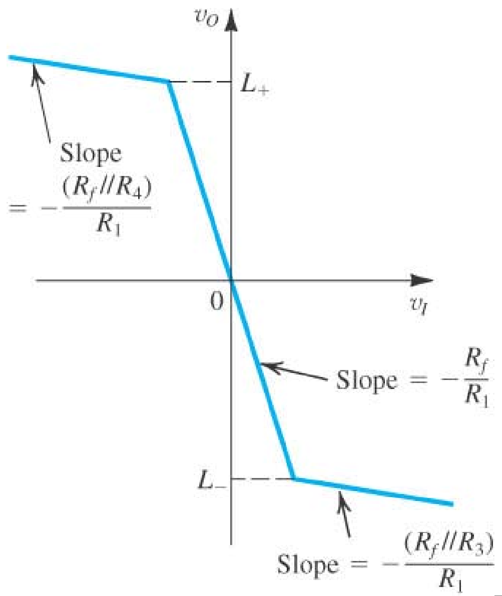


Figure 13.3 **(a)** A popular limiter circuit. **(b)** Transfer characteristic of the limiter circuit; L_- and L_+ are given by Eqs. (13.8) and (13.9), respectively. **(c)** When R_f is removed, the limiter turns into a comparator with the characteristic shown.

• The circuit is more precise and versatile than those presented in Chapter 3.

A Popular Limiter Circuit : Analysis





(b)

■ Transfer characteristic

Initial operation : linear amplification with diodes turned off

$$v_O = -(R_f / R_1) v_I$$

$$v_A = V \frac{R_3}{R_2 + R_3} + v_O \frac{R_2}{R_2 + R_3}$$

$$v_B = -V \frac{R_4}{R_4 + R_5} + v_O \frac{R_5}{R_4 + R_5}$$

Diode D1 turns on when V_A reaches 0.7V ($-V_D$)

$$L_- = -V \frac{R_3}{R_2} - V_D \left(1 + \frac{R_3}{R_2}\right)$$

Assuming R_{on} of diode=0, amplifier gain reduces to

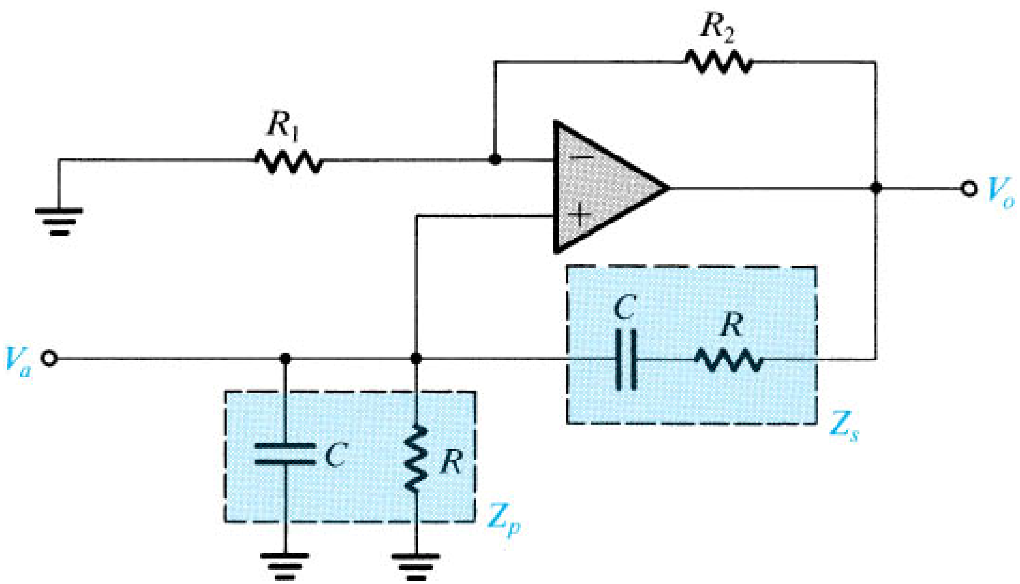
$$\frac{v_O}{v_I} = -\frac{R_f // R_3}{R_1}$$

When Rf is removed, it becomes a comparator

R_f determines linear range

R_2, R_3 R_4, R_5 determines limiter levels

Linear Osc 1 : The Wien-Bridge Oscillator



- **A** circuit : an op amp connected in the noninverting configuration with a closed-loop gain of $1+R_2/R_1$.
- **β** circuit : RC network → frequency selectivity

- The loop gain : L=Aβ

$$L(s) = \left[1+\frac{R_2}{R_1}\right]\frac{Z_P}{Z_P+Z_S} = \frac{1+R_2/R_1}{3+sCR+1/sCR}$$

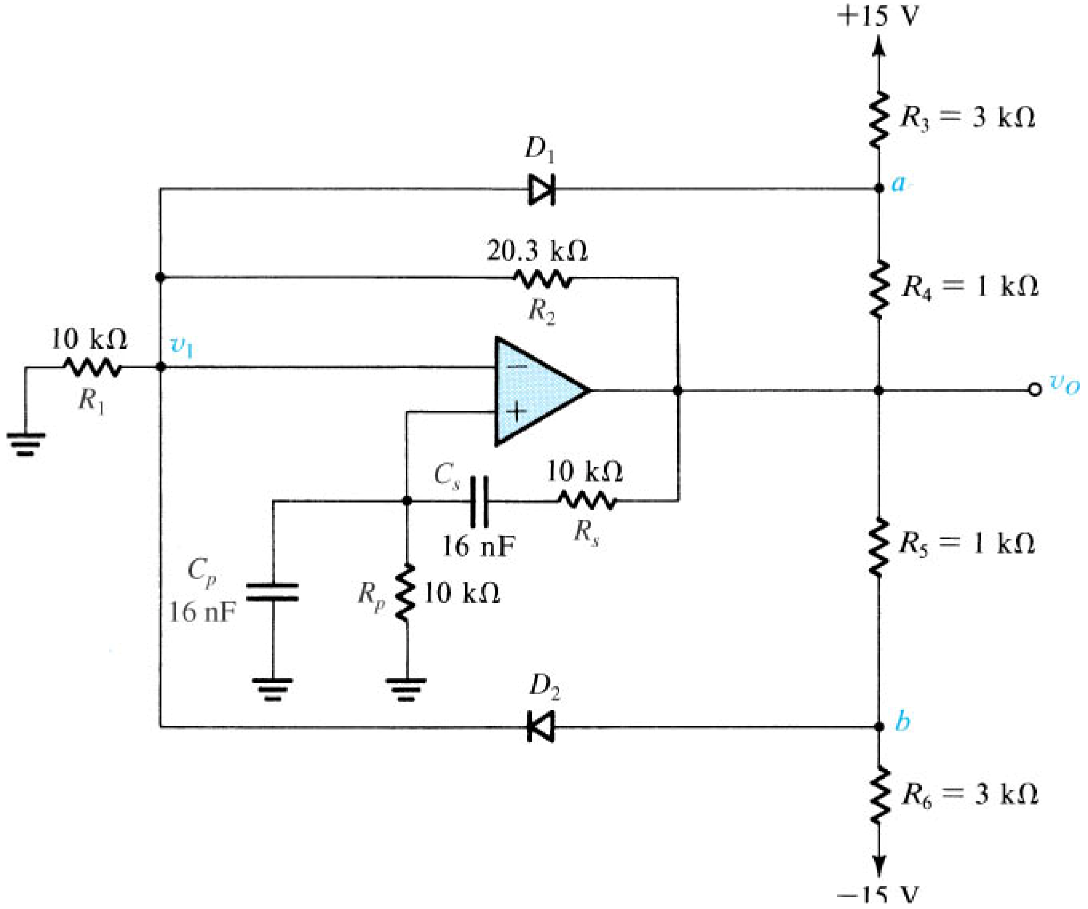
$$L(j\omega) = \frac{1+R_2/R_1}{3+j(\omega CR-1/\omega CR)}$$

- Oscillation Condition
  - Phase : $\omega_0 = 1/CR$
  - Magnitude : $R_2/R_1 = 2$
  - Start magnitude :

$$R_2/R_1 = 2+\delta$$

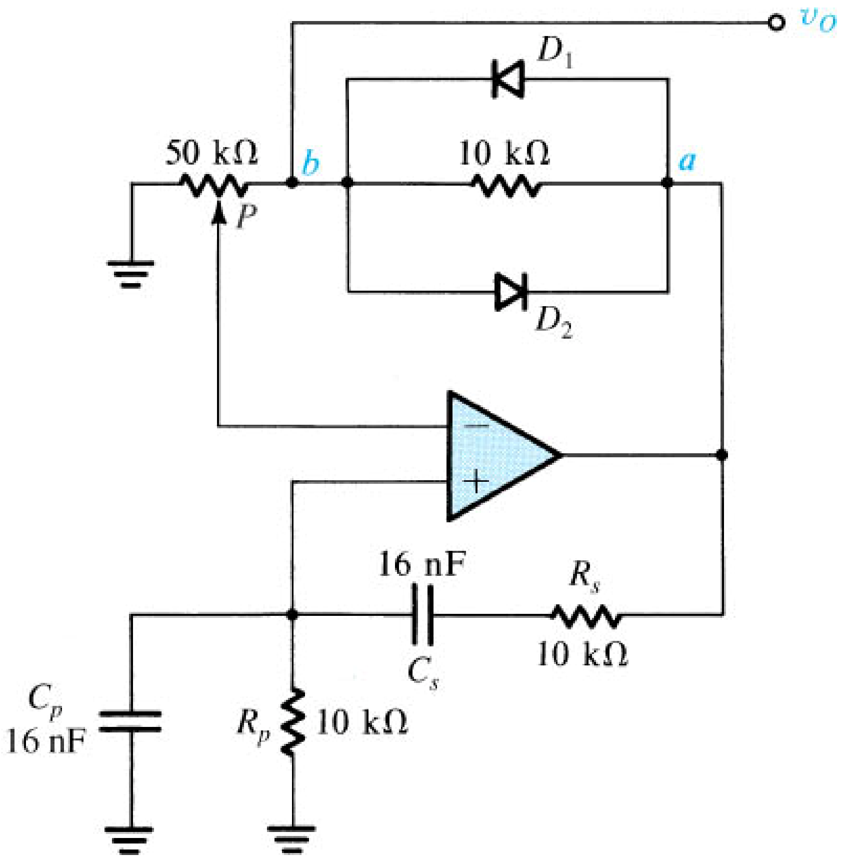
Poles pulled back to jω axis

The Wien-Bridge Oscillator with Limiter



- To obtain a symmetrical output waveform,
  - R_3 is chosen equal to R_6
  - R_4 is chosen equal to R_5.

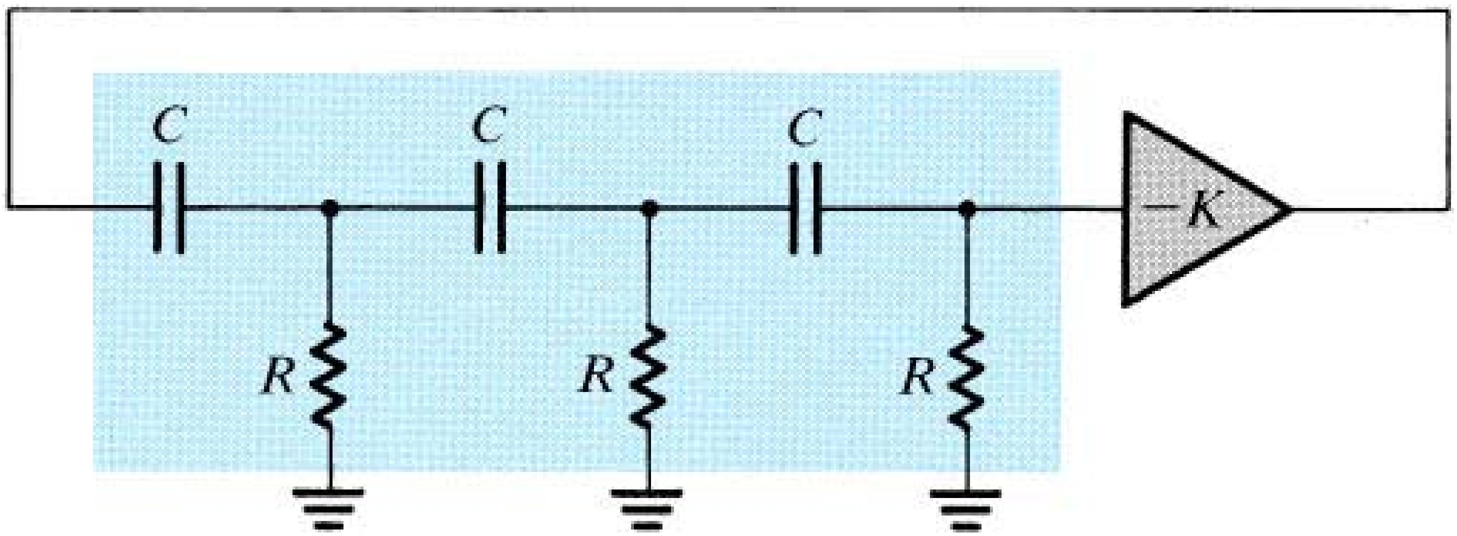
Inexpensive Implementation of Wien-Bridge Oscillator



- Potentiometer P is adjusted until oscillations just start to grow.
- The output amplitude can be varied by adjusting potentiometer P.
- The output is taken at point b rather than at the op-amp output terminal.

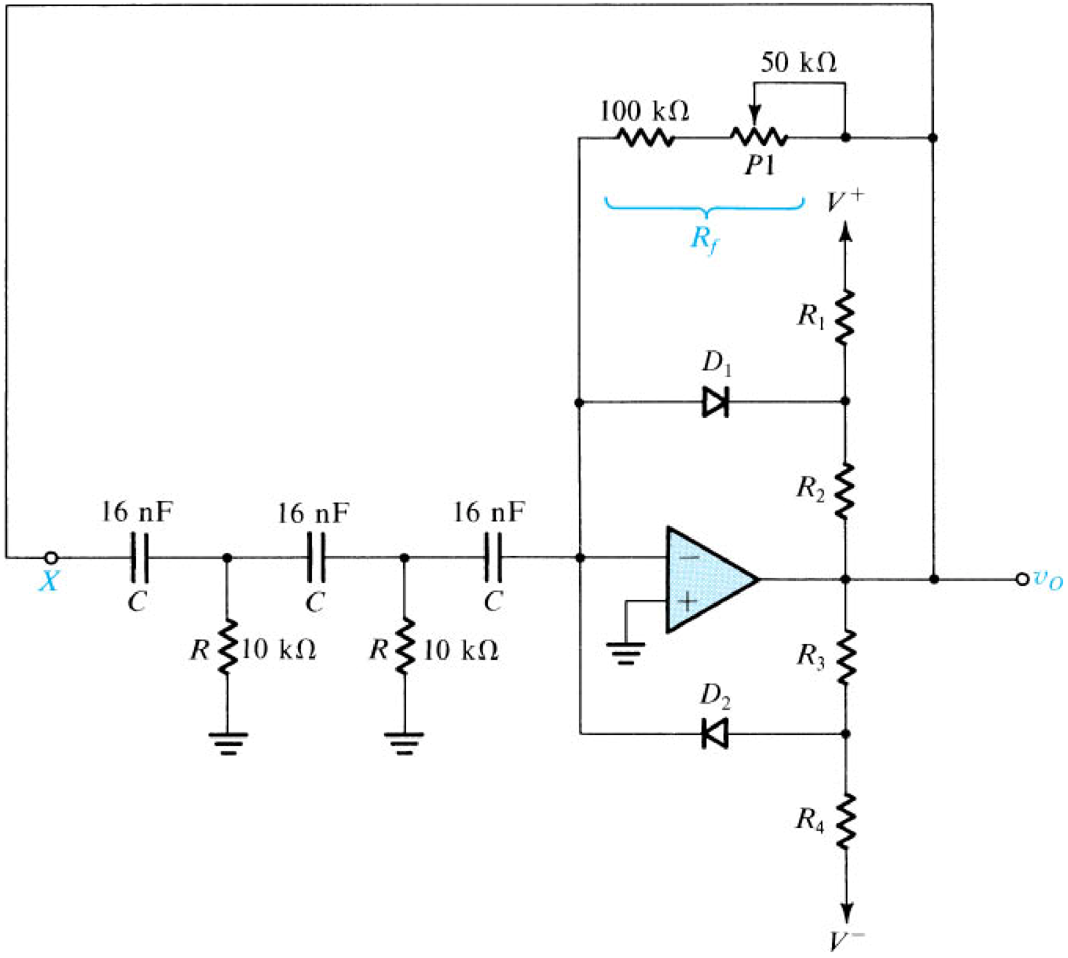
(∵ Signal at b has lower distortion than that at a.) → need buffer amplifier

Linear Osc 2 : Phase-Shift Oscillator



- Negative gain amplifier(-K) + three-section (three-order) RC ladder network.
- Oscillate at the frequency for which the phase shift of the RC network is 180°.
- Three is the minimum number of RC network that is capable of producing a 180° phase shift at a finite frequency.

Implementation of Phase-Shift Oscillator



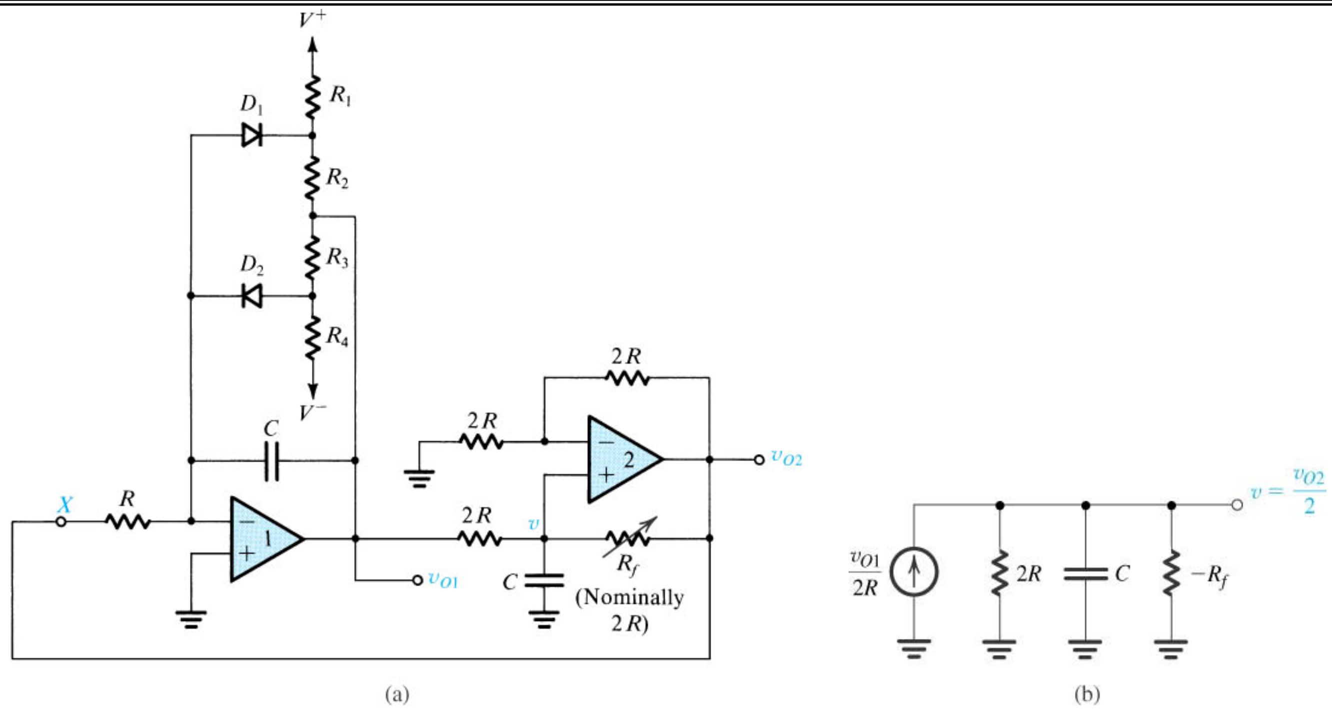
- To start oscillations, R_f has to be made slightly greater than the minimum required value.

(+) The circuit stabilizes more rapidly

(+) Provides sine waves with more stable amplitude

(-) The price paid is an increased output distortion.

Linear Osc 3 : Quadrature Oscillator



- Idea : one inverting integrator generates $v_o = -\frac{\int \frac{v}{R} dt}{C} \longrightarrow v_o = -\frac{v_i}{sCR}$ → 90^o phase shift

- Need non-inverting integrator to make net loop phase = 0 → v_{o2}=2v, current through R_f = $-(v-v_{o2})/R_f = -v/R_f$ → Negative resistance is viewed. →
- Oscillation start : set $|-R_f| > 2R$

13.2.3 The Quadrature Oscillator(2)

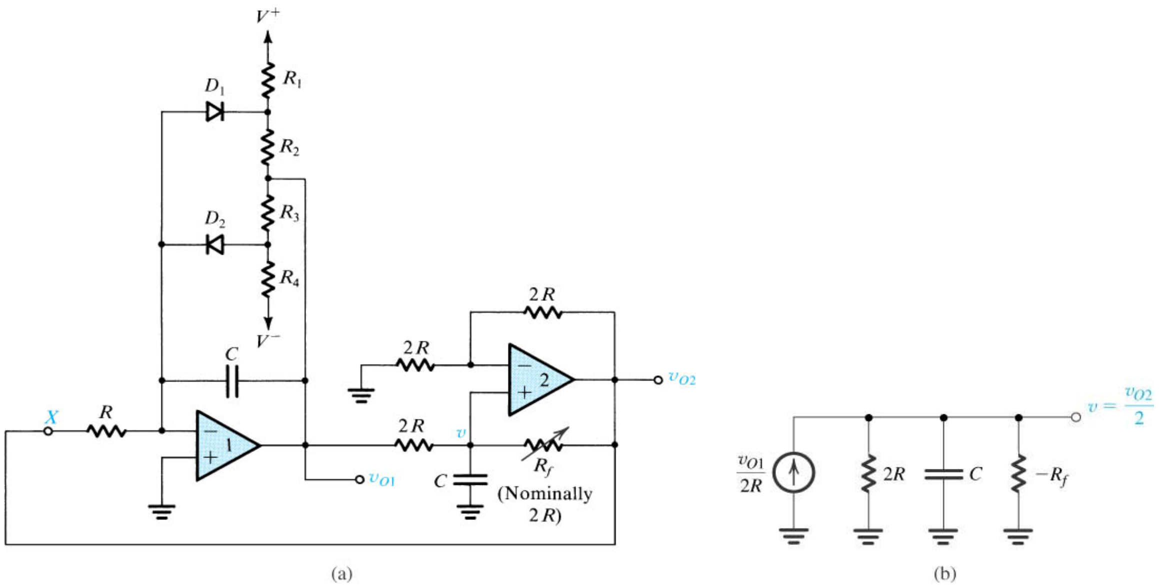


Figure 13.9 **(a)** A quadrature-oscillator circuit. **(b)** Equivalent circuit at the input of op amp 2.

- ❑ Amplifier 1 is connected as an inverting Miller integrator with a limiter in the feedback for amplitude control.

- ❑ Amplifier 2 is connected as a noninverting integrator.

13.2.3 The Quadrature Oscillator(3)

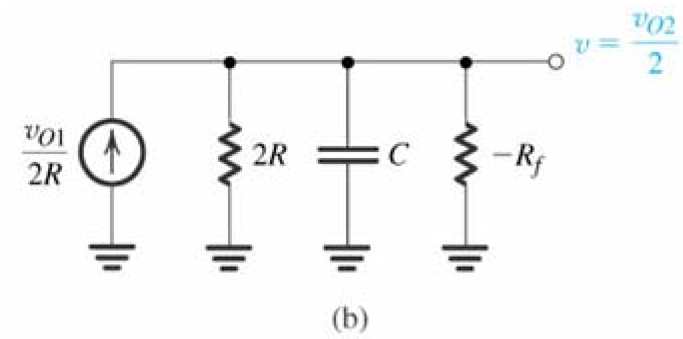


Figure 13.9 (b) Equivalent circuit at the input of op amp 2.

- The integrator input voltage v_{O1} and the series resistance 2

  → The Norton equivalent composed of a current source $v_{O1}/2R$ and a parallel resistance 2R.

- Since $v_{O2} = 2v$,
  the current through R_f: $(2v - v)/R = v/R$ (the direction from output to input).

- $-R_f$ cancels 2R, and $v_{O1}/2R$ feeding a capacitor *C*.

- The result is $v = \frac{1}{C}\int_0^t \frac{v_{O1}}{2R}dt$ and $v_{O2} = 2v = \frac{1}{CR}\int_0^t v_{O1}dt$ (noninverting integrator).

13.2.3 The Quadrature Oscillator(4)

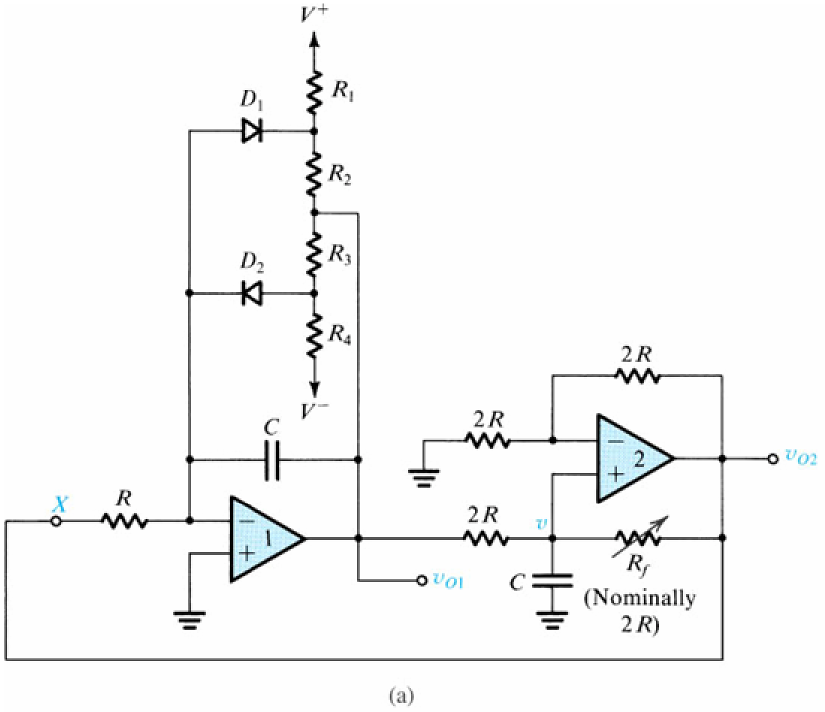


Figure 13.9 **(a)** A quadrature-oscillator circuit.

- The resistance R_f in the positive-feedback path is made variable.
- Decreasing the value of R_f ensures that the oscillations start.
- The loop gain

$$L(s) \equiv \frac{V_{O2}}{V_x} = -\frac{1}{s^2C^2R^2}$$

- The loop will oscillate at frequency

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

13.2.4 The Active-Filter-Tuned Oscillator(1)

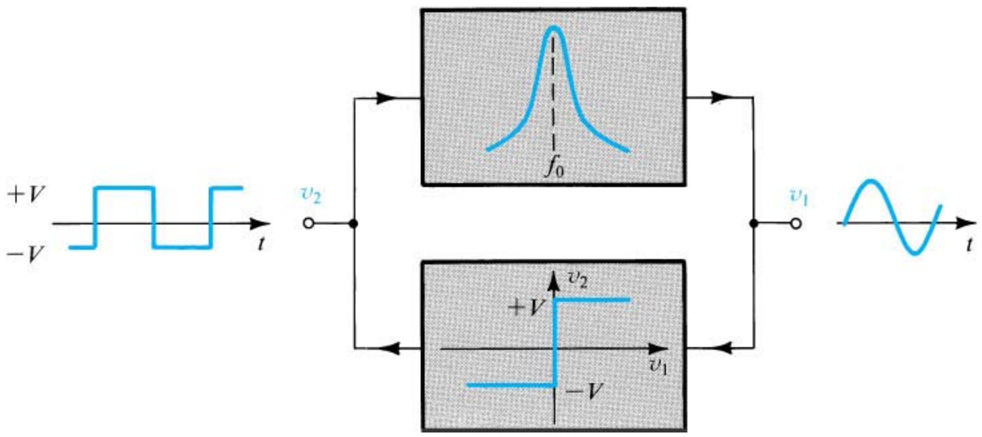


Figure 13.10 Block diagram of the active-filter-tuned oscillator.

- The circuit consists of a high-*Q* bandpass filter connected in a positive-feedback loop with a hard limiter.
- Assume that oscillations have already started.

- The output of the bandpass filter will be a sine wave whose frequency is f_0.
- The sine-wave signal v_1 is fed to the limiter.
- The square wave is fed to the bandpass filter.
- Independent control of frequency and amplitude as well as of distortion of the output sinusoid.

13.2.4 The Active-Filter-Tuned Oscillator(2)

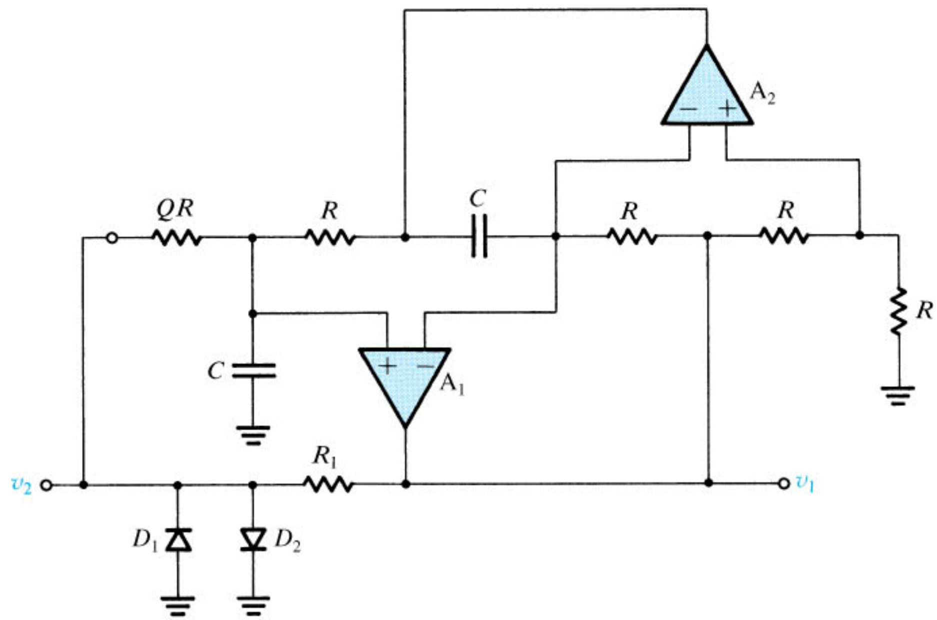


Figure 13.11 A practical implementation of the active-filter-tuned oscillator.

- Resistor R_2 and capacitor C_4 make the output of the lower op amp directly proportional to the voltage across the resonator.

- Limiter : resistance R1 and two diodes.

13.2.5 A Final Remark

- Useful for operation in the range 10Hz to 100kHz(or perhaps 1MHz at most).
- 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,
  transistors together with LC tuned circuits or crystals are frequently used.

13.3 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 **100kHz to hundreds of megahertz**.

• They exhibit **higher Q** than the RC types

• LC oscillators are difficult to tune over wide ranges, and crystal oscillators operate a single frequency.

13.3.1 LC-Tuned Oscillators(1)

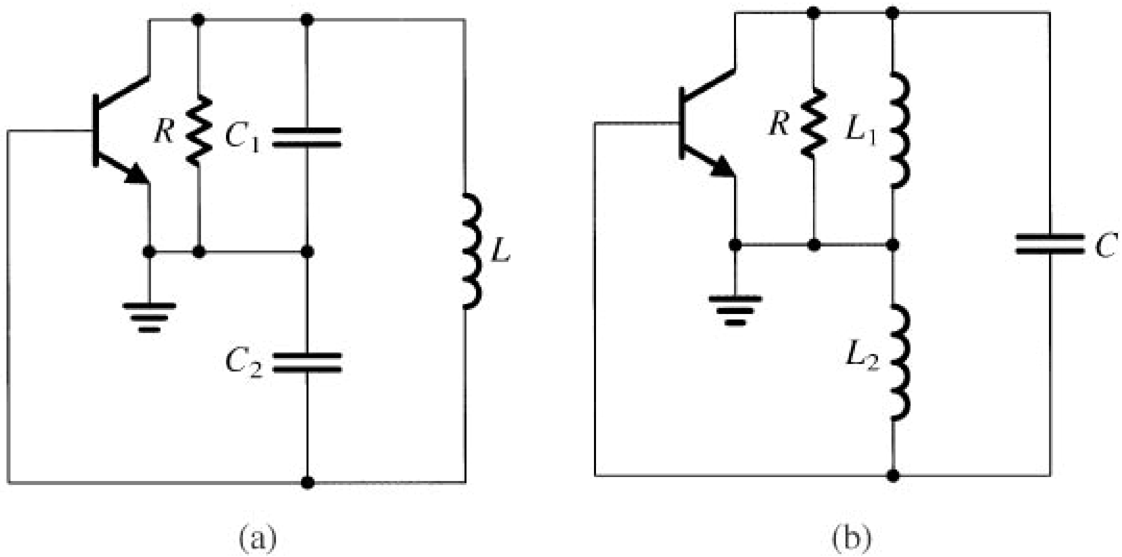


Figure 13.12 Two commonly used configurations of LC-tuned oscillators: **(a)** Colpitts and **(b)** Hartley.

- They are known as the **Colpitts oscillator(a)** and the **Hartley oscillator(b)**.
- This feedback is achieved by way of a capacitive divider in the Colpitts oscillator and by way of an inductive divider in the Hartley circuit.
- The resistor R models the combination of the losses of the inductors, the load resistance of the oscillator, and the output resistance of the transistor.

13.3.1 LC-Tuned Oscillators(2)

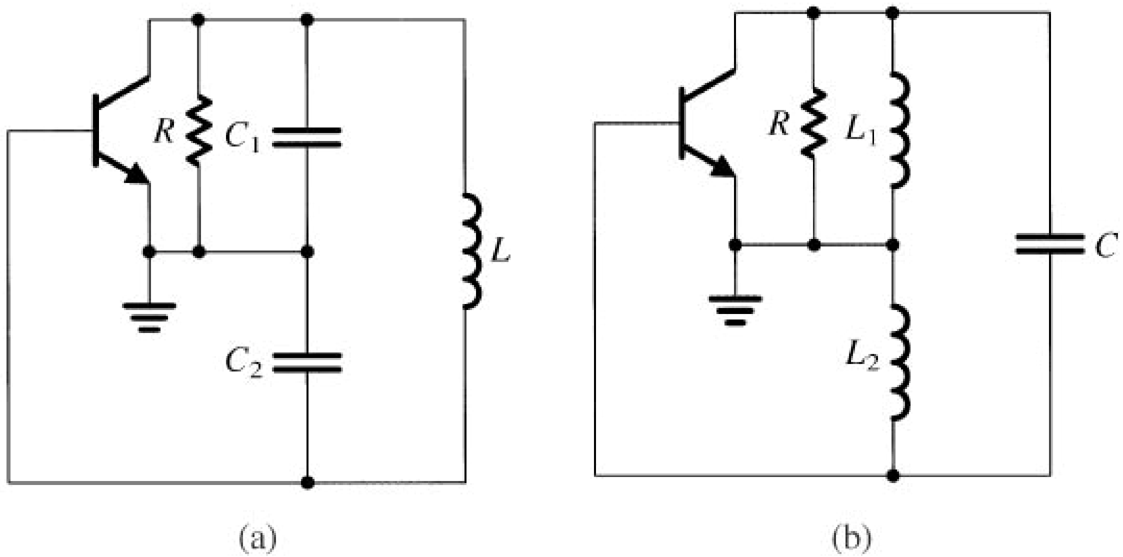


Figure 13.12 Two commonly used configurations of LC-tuned oscillators: **(a)** Colpitts and **(b)** Hartley.

• If the frequency of operation is sufficiently low that we can neglect the transistor capacitances, the frequency of oscillation will be determined by the resonance frequency of the parallel-tuned circuit

■ The Colpitts oscillator

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

■ The Hartley oscillator

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

13.3.1 LC-Tuned Oscillators(3)

■ Colpitts oscillator

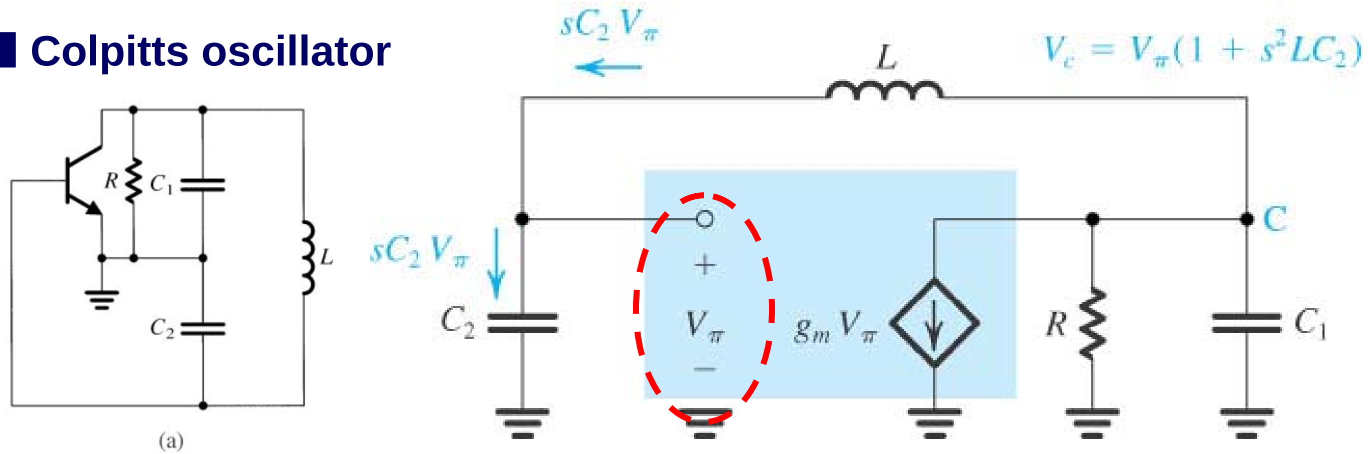


Figure 13.13 Equivalent circuit of the Colpitts oscillator of Fig. 13.12(a). To simplify the analysis, C_μ and r_π are neglected. We can consider C_π to be part of C_2, and we can include r_o in R.

- The ratio L_1/L_2 or C_1/C_2 determines the feedback factors.
- Capacitance C_μ is neglected & capacitance C_π is included in C_2
- Input resistance r_π is neglected assuming that at the frequency of oscillation $r_\pi \gg (1/\omega C_2)$.
- Resistance R includes r_0 of the transistor.
- **To find the loop gain:** break the loop at the transistor base, apply an input voltage Vπ and find the returned voltage that appears across the input terminals of the transistor.
- **To analyze the circuit:** eliminate all current and voltage variables, and thus obtain one equation.
- The resulting equation will give us the conditions for oscillation.

13.3.1 LC-Tuned Oscillators(4)

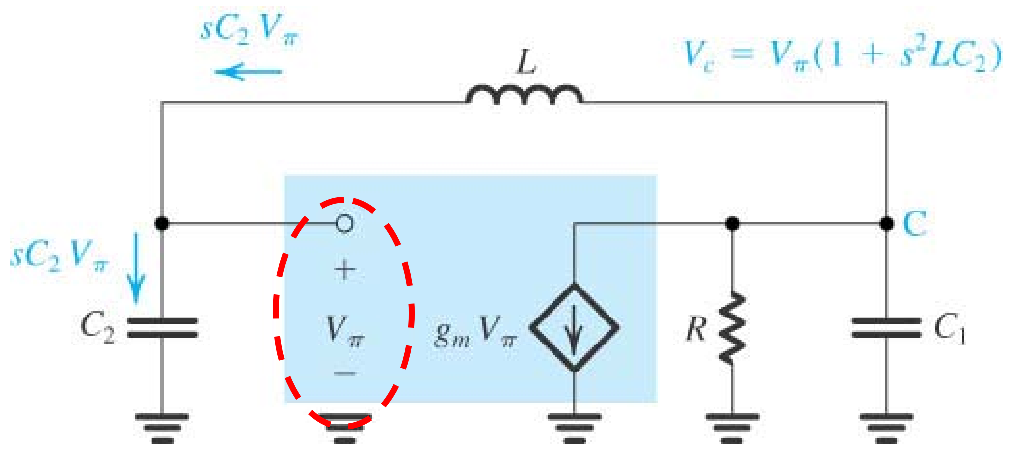


Figure 13.13 Equivalent circuit of the Colpitts oscillator of Fig. 13.12(a). To simplify the analysis, C_μ and r_π are neglected. We can consider C_π to be part of C_2, and we can include r_o in R.

- A node equation at node C is

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

- Since $V_\pi \neq 0$(oscillations have started), it can be eliminated,

$$s^3LC_1C_2 + s^2(LC_2/R) + s(C_1 + C_2) + \left(g_m + \frac{1}{R}\right) = 0$$

- Substituting s=jω gives,

$$\left(g_m + \frac{1}{R} - \frac{\omega^2 LC_2}{R}\right) + j[\omega(C_1 + C_2) - \omega^3 LC_1C_2] = 0$$

13.3.1 LC-Tuned Oscillators(5)

• For oscillations to start, both the real and imaginary parts must be zero.

$$\omega(C_1+C_2)-\omega^3 LC_1C_2=0$$

$$\omega_0=\frac{1}{\sqrt{L\left(\frac{C_1C_2}{C_1+C_2}\right)}}$$

• Substituting s=jω gives,

$$\left(g_m+\frac{1}{R}-\frac{\omega^2 LC_2}{R}\right)=0$$

$$g_mR+1-\frac{1}{L\left(\frac{C_1C_2}{C_1+C_2}\right)}LC_2=0$$

$$g_mR+1-\frac{C_1+C_2}{C_1}=0$$

$$\therefore \boxed{C_2/C_1=g_mR}$$

• For sustained oscillations, the magnitude of the gain from base to collector (g_mR) must be equal to the inverse of the voltage ratio provided by the capacitive divider($v_{eb}/v_{ce}=C_1/C_2$).

■ For oscillations to start, the loop gain must be greater than unity.

$$g_mR>C_2/C_1$$

• As oscillations grow in amplitude, the transistor's **nonlinear characteristic reduces** the effective value of $\mathbf{g_m}$ and reduce the loop gain to unity.

13.3.1 LC-Tuned Oscillators(6)

❑ The Hartley circuit analysis(Exercise 13.8)

· At high frequencies, more accurate transistor models must be used.

: The y parameters(the short-circuit admittance) of the transistor can be measured at the intended frequency ω_0, and the analysis can then be carried out using the y-parameter model(Appendix B).

: This is usually simpler and more accurate, especially at frequencies above about 30% of the transistor f_T.

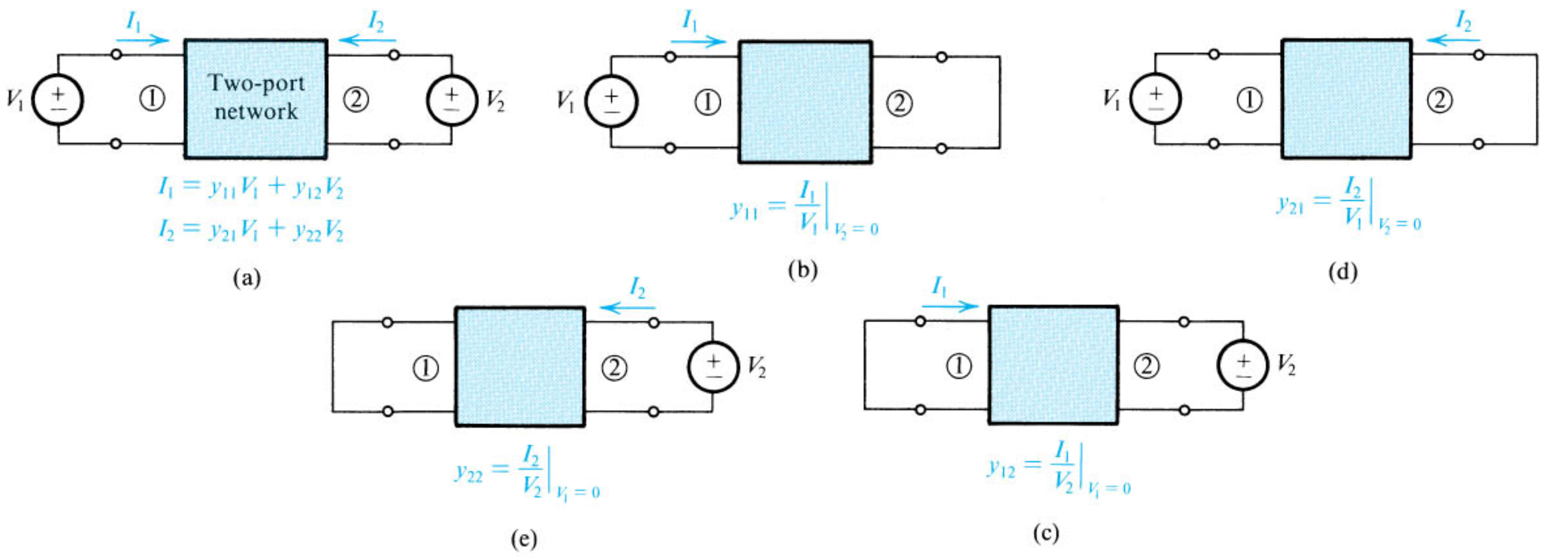


Figure B.2 Definition and conceptual measurement circuits for *y* parameters.

13.3.1 LC-Tuned Oscillators(7)

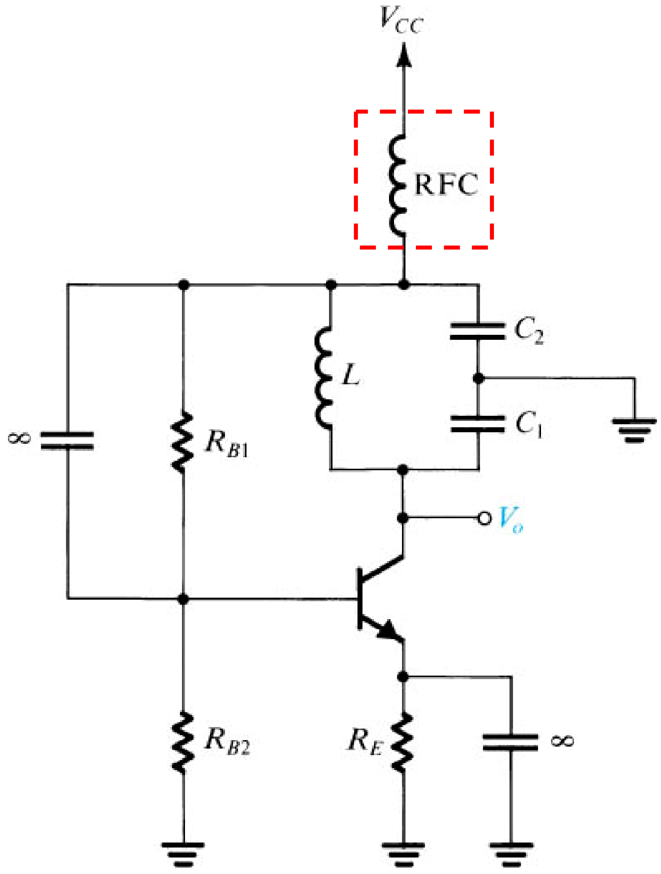


Figure 13.14 Complete circuit for a Colpitts oscillator.

• An example of a practical LC oscillator(Colpitts)

• The radio-frequency choke(RFC) provides a high reactance at ω_0 but a low dc resistance.

13.3.1 LC-Tuned Oscillators(9)

❑ Determining the amplitude of oscillation

· Unlike the op-amp oscillators that incorporate special amplitude-control circuitry, LC-tuned oscillators utilize the nonlinear i_C-v_{BE} characteristics of the BJT(*self-limiting oscillators*).

→ As the oscillations grow in amplitude, the effective gain of the transistor is reduced below its small-signal value.

→ Eventually, an amplitude is reached at which the effective gain is reduced to the point that the Barkhausen criterion is satisfied exactly.

→ The amplitude then remains constant at this value.

13.3.2 Crystal Oscillators(1)

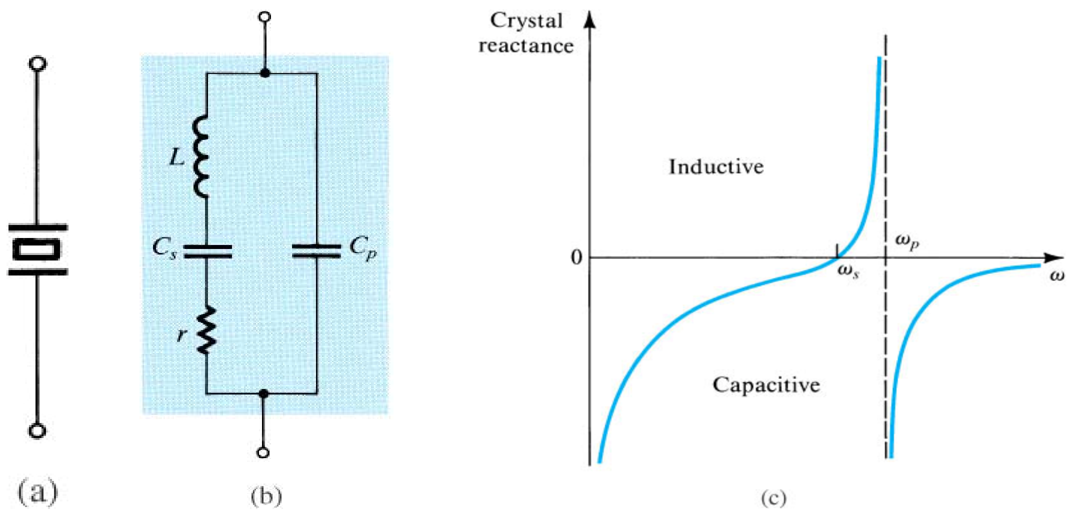


Figure 13.15 A piezoelectric crystal. **(a)** Circuit symbol. **(b)** Equivalent circuit. **(c)** Crystal reactance versus frequency [note that, neglecting the small resistance r, $Z_{\text{crystal}} = jX(\omega)$].

- A piezoelectric crystal(quartz) exhibits electromechanical-resonance characteristics that are very stable(with time and temperature) and highly selective(having very high Q factors).
- The resonance properties are characterized by

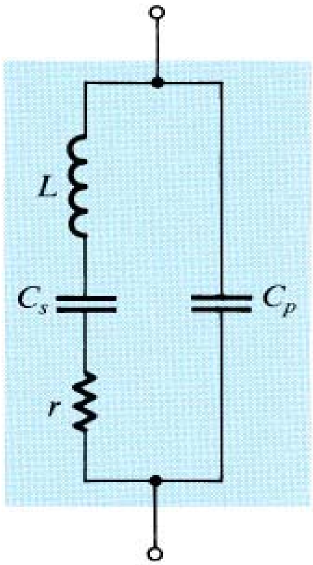
: large inductance L(as high as hundreds of henrys), very small series capacitance C_S(as small as 0.0005pF), series resistance *r* representing a Q factor $\omega_0 L/r$(can be as high as a few hundred thousand) and parallel capacitance C_P(a few pF, $C_P \gg C_S$)

13.3.2 Crystal Oscillators(2)

- Since the Q factor is very high, we may neglect the resistance *r*.
  The crystal impedance is

$$Z(s) = 1/\left[sC_P + \frac{1}{sL + 1/sC_S}\right]$$

$$= \frac{1}{sC_P}\frac{s^2 + \left(1/LC_S\right)}{s^2 + [\left(C_P + C_S\right)/LC_SC_P]} \quad ...Eq.(13.23)$$



(b)

- From Eq.(13.23) and from Fig. 13.15(b) we see that the crystal has two resonance frequencies.

ⅰ. Series resonance at ω_S

$$\omega_S = 1/\sqrt{LC_S} \quad ...Eq.13.24$$

ⅱ. Parallel resonance at ω_P

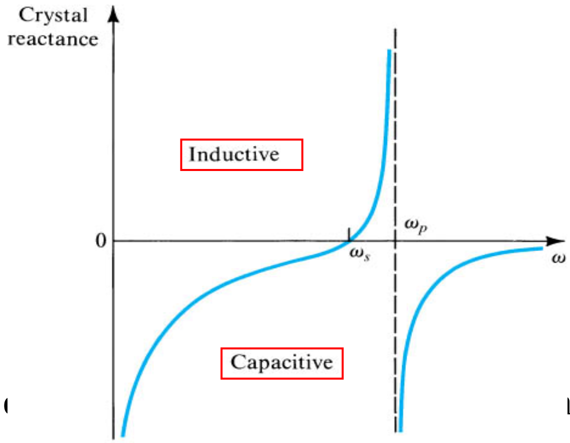
$$\omega_P = 1/\sqrt{L\left(\frac{C_SC_P}{C_S + C_P}\right)} \quad ...Eq.13.25$$

- For s=jω

$$Z(j\omega) = -j\frac{1}{\omega C_P}\left(\frac{\omega^2 - \omega_S{}^2}{\omega^2 - \omega_P{}^2}\right)$$

13.3.2 Crystal Oscillators(3)

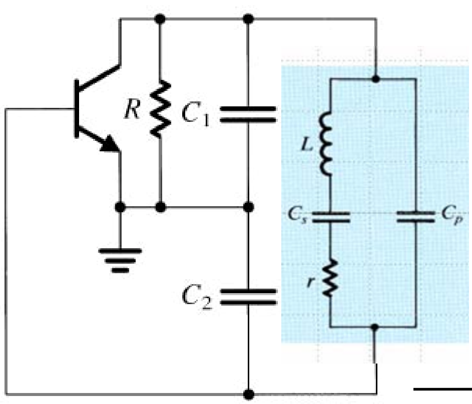
• Expressing Z(jω)=jX(ω), the crystal reactance X(ω) will have the shape,



(c)

$$Z(j\omega) = jX(\omega) = -j\frac{1}{\omega C_P}\left(\frac{\omega^2 - \omega_S{}^2}{\omega^2 - \omega_P{}^2}\right)$$

• We (ductive over the very narrow frequency band between ω_S and ω_P.



(a)

• For a given crystal, this frequency band is well defined. Thus we may use the crystal to replace the inductor of the Colpitts oscillator.

• The resulting circuit will oscillate at the resonance frequency of the crystal inductance L with the series equivalent of C_S and $(C_P + C_1C_2/(C_1+C_2))$.

• Since C_S is much smaller than the three other capacitances,

$$\omega_0 \approx 1/\sqrt{LC_S} = \omega_S$$

13.3.2 Crystal Oscillators(4)

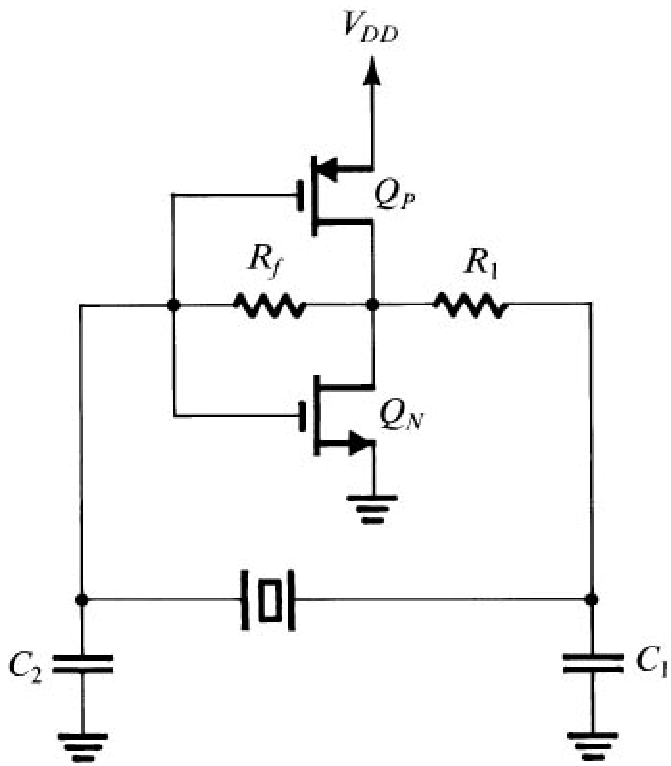


Figure 13.16 A Pierce crystal oscillator utilizing a CMOS inverter as an amplifier.

❑Pierce oscillator

· Utilizing CMOS inverter(Section 4.10) as amplifier

· Resistor R_f determines a dc operating point in the high-gain region of the CMOS inverter

· Resistor R_1 and capacitor C_1 provide a low-pass filter that discourages the circuit from oscillating at a higher harmonic of the crystal frequency

13.4 BISTABLE MULTIVIBRATORS

- Multivibrators.

  Bistable.
  Monostable.
  Astable.

- Bistable vibrator
  has two stable states.
  ① can remain in stable state indefinitely.
  ② moves to the other stable state only when appropriately triggered.

13.4.1 The Feedback Loop(1)

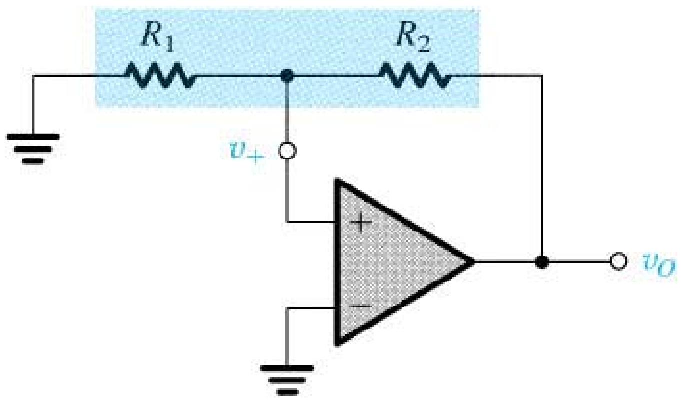


Figure 13.17 A positive-feedback loop capable of bistable operation.

- Consists of an op amp and a resistive voltage divider in the positive-feedback path.
- Assume that the electrical noise causes a small positive increment in the voltage v_+.

$$\beta \equiv R_1/(R_1+R_2)$$

① Positive increment occrrred in v+.

$$v_O = L_+ \qquad v_+ = L_+ R_1/(R_1+R_2)$$

② Negative increment occrrred in v+.

$$v_O = L_- \qquad v_+ = L_- R_1/(R_1+R_2)$$

13.4.1 The Feedback Loop(2)

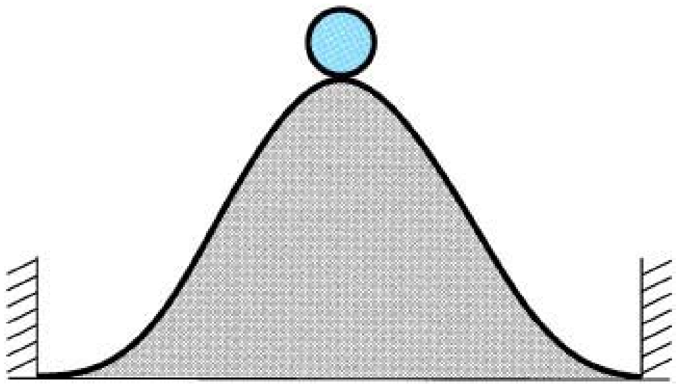
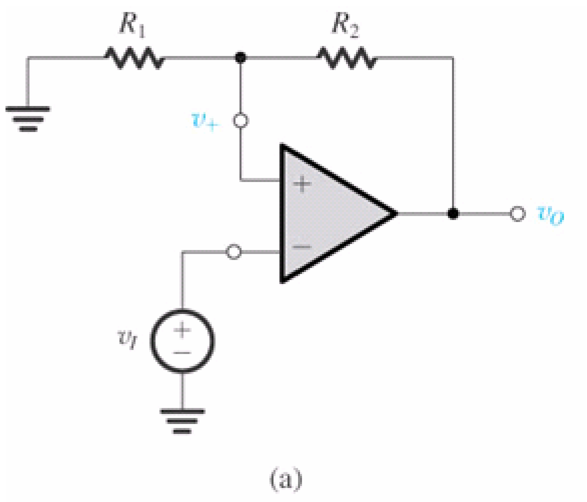


Figure 13.18 A physical analogy for the operation of the bistable circuit. The ball cannot remain at the top of the hill for any length of time (a state of unstable equilibrium or metastability); the inevitably present disturbance will cause the ball to fall to one side or the other, where it can remain indefinitely (the two stable states).

- The circuit cannot exist in the state for which $v_+ = 0$ and $v_O = 0$ (state of unstable equilibrium, metastable state) for any length of time.

- Any disturbance (electrical noise) causes the bistable circuit to switch to one of its two stable states (positive saturation or negative saturation).

13.4.2 Transfer Characteristics of the Bistable Circuit(1)



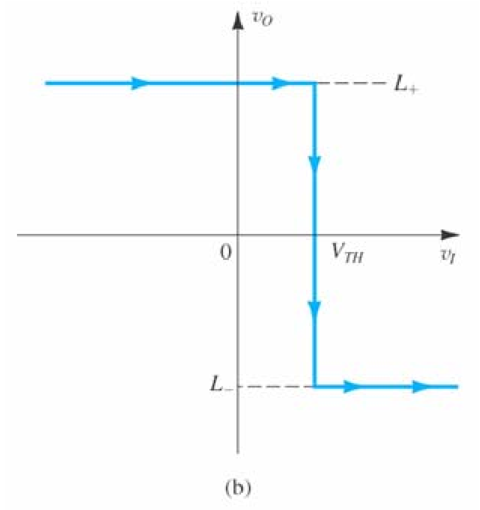
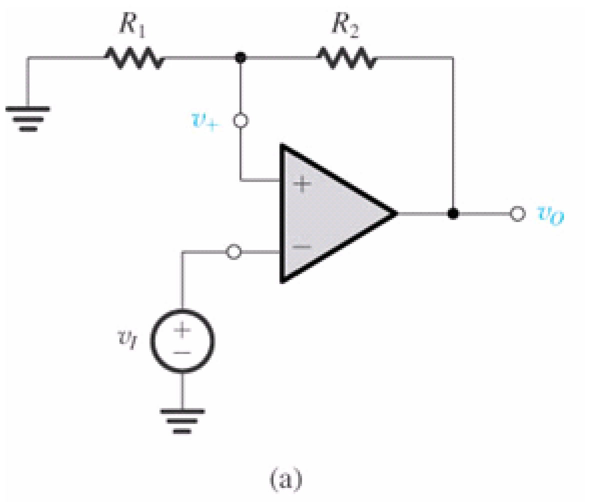


Figure 13.19 (a) The bistable circuit of Fig. 13.17 with the negative input terminal of the op amp disconnected from ground and connected to an input signal v_I. **(b)** The transfer characteristic of the circuit in (a) for increasing v_I.

① Assume that v_I is increased from 0V, $v_O = L_+$ and $v_+ = \beta L_+$

- As v_I begins to exceed v_+, a net negative voltage develops between the input terminals of the op amp and thus v_O goes negative.

13.4.2 Transfer Characteristics of the Bistable Circuit(2)



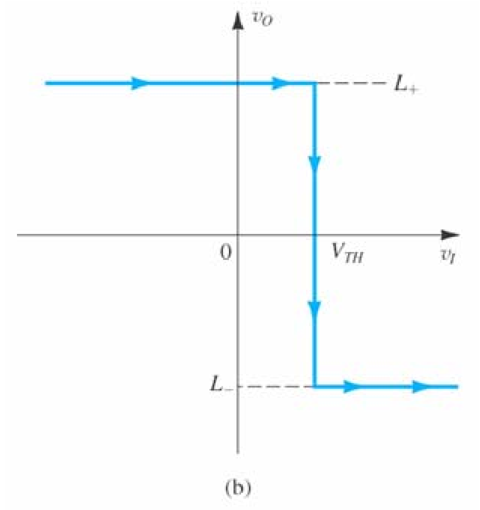


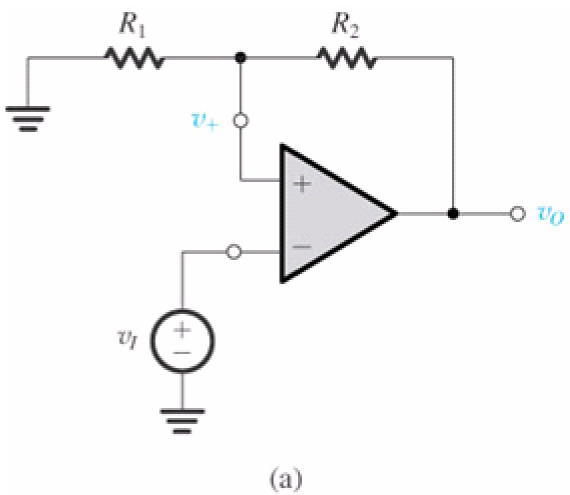
Figure 13.19 (a) The bistable circuit of Fig. 13.17 with the negative input terminal of the op amp disconnected from ground and connected to an input signal v_I. **(b)** The transfer characteristic of the circuit in (a) for increasing v_I.

- v_+ goes negative, increasing the net negative input to the op amp.
- The process culminates in the op amp saturating in the negative direction.

$$v_O = L_- \text{ and } v_+ = \beta L_-$$

- Threshold voltage : $V_{TH} = \beta L_+$

13.4.2 Transfer Characteristics of the Bistable Circuit(3)



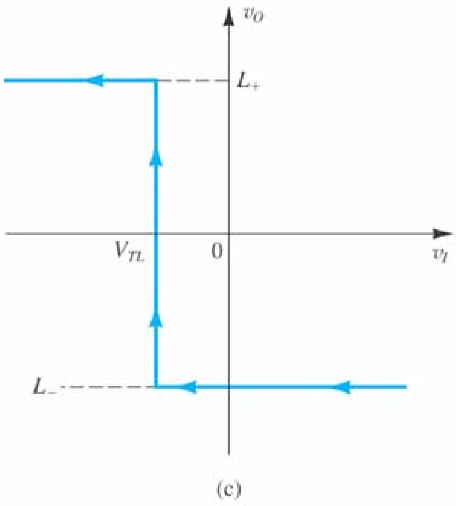
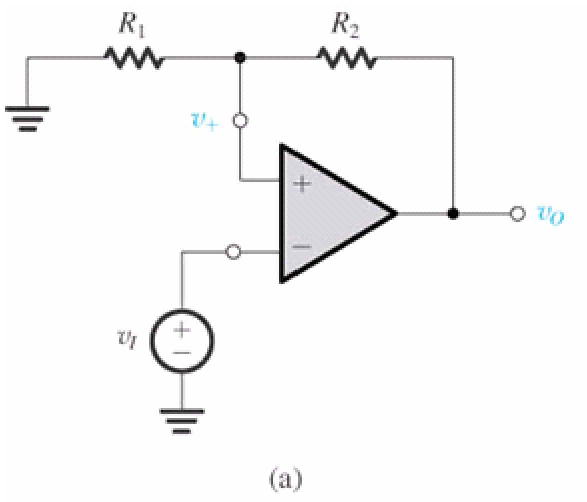


Figure 13.19 **(a)** The bistable circuit of Fig. 13.17 with the negative input terminal of the op amp disconnected from ground and connected to an input signal v_I. **(c)** The transfer characteristic for decreasing v_I.

② Consider v_I is decreased.

- Circuit remains in the negative-saturation state until $v_I \geq \beta L_-$
- $v_I < \beta L_- \longrightarrow$ Net positive voltage appears between the op amp's input terminals $\longrightarrow$ Positive-saturation state
- Threshold voltage : $V_{TL} = \beta L_-$

13.4.2 Transfer Characteristics of the Bistable Circuit(4)



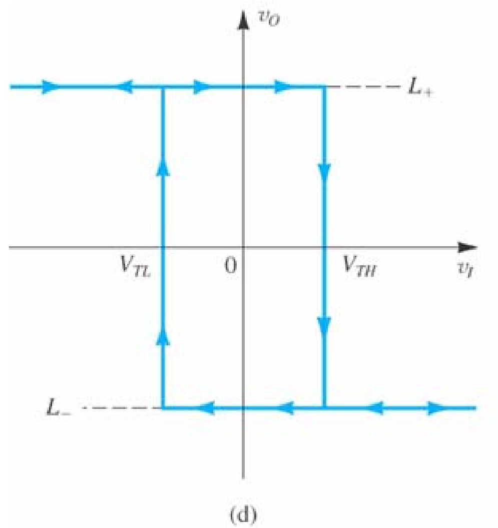


Figure 13.19 **(a)** The bistable circuit of Fig. 13.17 with the negative input terminal of the op amp disconnected from ground and connected to an input signal v_I. **(d)** The complete transfer characteristics.

- The circuit changes state at different values of v_I, depending on whether v_I is increasing or decreasing.
- The width of the *hysteresis* is the difference between the high threshold V_{TH} and the low threshold V_{TL}.
- Inverting circuit.

13.4.3 Triggering the Bistable Circuit

- If the circuit is in the L_+ state.

→ Applying an input v_I of value greater than $V_{TH} \equiv \beta L_+$

→ The circuit can be switched to the L_- state.

- If the circuit is in the L_- state.

→ Applying an input v_I of value smaller than $V_{TL} \equiv \beta L_-$

→ The circuit can be switched to the L_+ state.

$\therefore$ v_I : trigger signal.

13.4.4 The Bistable Circuit as a Memory Element

- For certain input range, the output is determined by the previous value of the tirgger signal.

- The bistable multivibrator is the basic *memory* element of digital systems.

13.4.5 A Bistable Circuit with noninverting Transfer Characteristics(1)

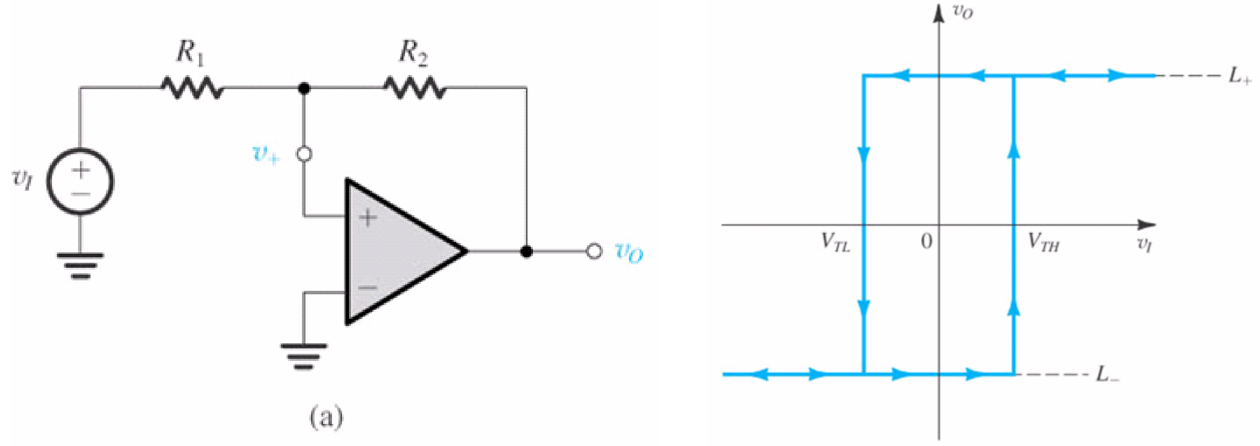


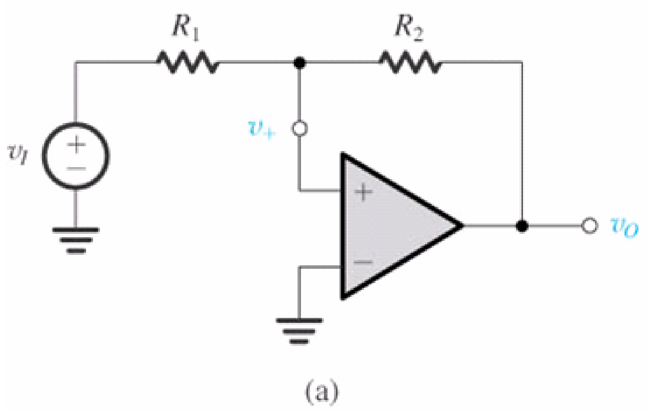
Figure 13.20 **(a)** A bistable circuit derived from the positive-feedback loop of Fig. 13.17 by applying v_I through R_1. **(b)** The transfer characteristic of the circuit in (a) is noninverting. (Compare it to the inverting characteristic in Fig. 13.19d.)

- Transfer characteristics,

$$v_+ = v_I \frac{R_2}{R_1 + R_2} + v_O \frac{R_1}{R_1 + R_2}$$

- If the circuit is in the positive stable state, $v_I = V_{TL} = -L_+(R_1 / R_2)$ will trigger the circuit into the L_- state.

13.4.5 A Bistable Circuit with noninverting Transfer Characteristics(2)



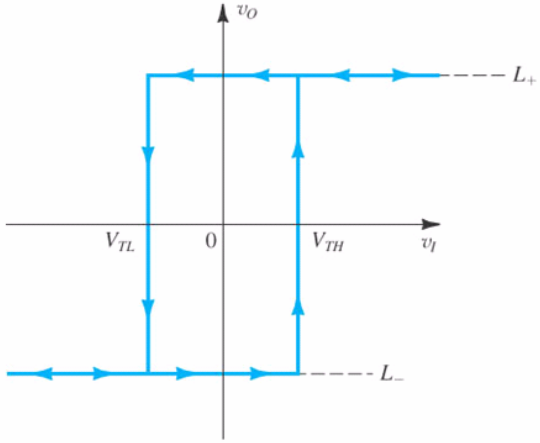


Figure 13.20 **(a)** A bistable circuit derived from the positive-feedback loop of Fig. 13.17 by applying v_I through R_1. **(b)** The transfer characteristic of the circuit in (a) is noninverting. (Compare it to the inverting characteristic in Fig. 13.19d.)

- If the circuit is in the negative stable state, $v_I = V_{TH} = -L_-(R_1/R_2)$ will trigger the circuit into the L_+ state.

- Negative triggering signal $\longrightarrow$ Negative state.
- Positive triggering signal $\longrightarrow$ Positive state.

$\therefore$ The transfer characteristic of this circuit is noninverting.

13.4.6 Application of the Bistable Circuit as a Comparator(1)

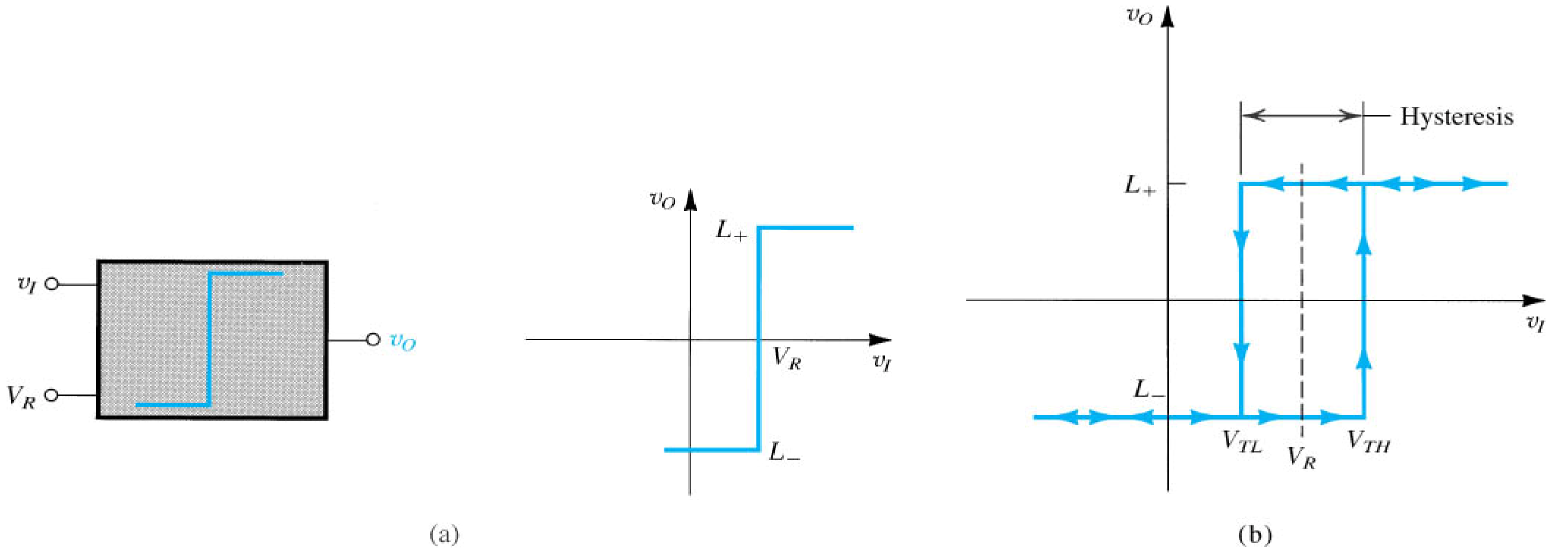


Figure 13.21 (a) Block diagram representation and transfer characteristic for a comparator having a reference, or threshold, voltage V_R. **(b)** Comparator characteristic with hysteresis.

- It is useful in many applications to add hysteresis to the comparator characteristics.
- The comparator exhibits two threshold values, V_{TL} and V_{TH}.
- Usually V_{TH} and V_{TL} are separated by a small amount(100mV).

13.4.6 Application of the Bistable Circuit as a Comparator(2)

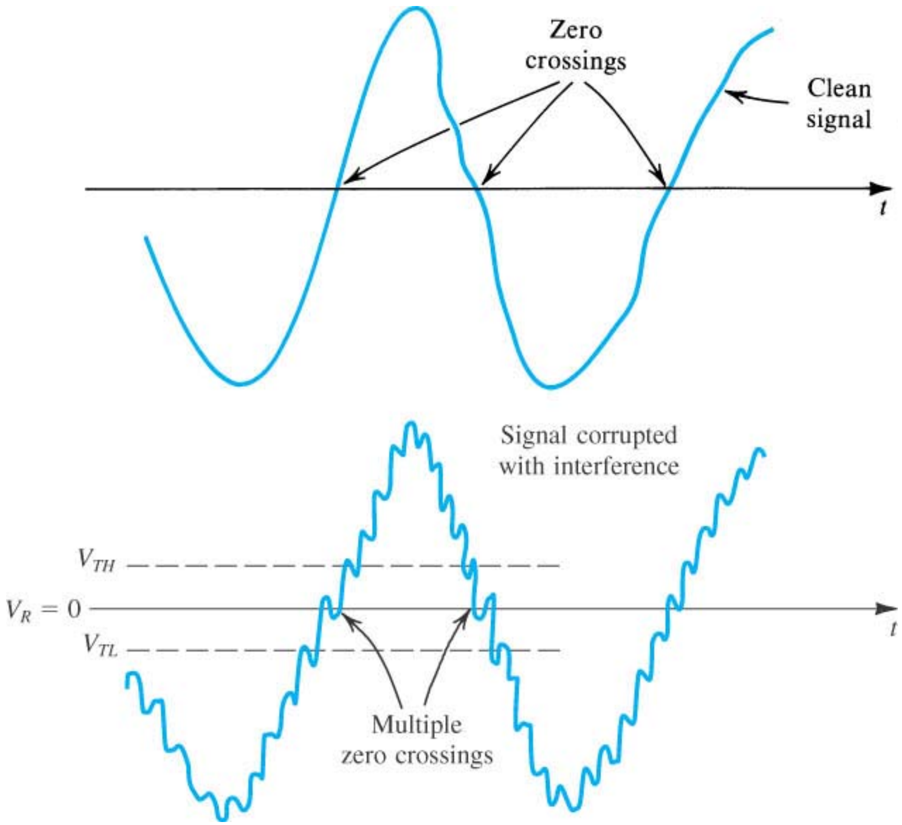


Figure 13.22 Illustrating the use of hysteresis in the comparator characteristics as a means of rejecting interference.

- ❑ To design a circuit that detects and counts the zero crossings of an arbitrary waveform.
- ❑ The comparator provides a step change at its output every time a zero crossing occurs.
- ❑ If the signal being processed has interference superimposed on it.

→ Solved by introducing hysteresis of appropriate width in the comparator characteristics.

13.4.7 Making the Output Levels more Precise

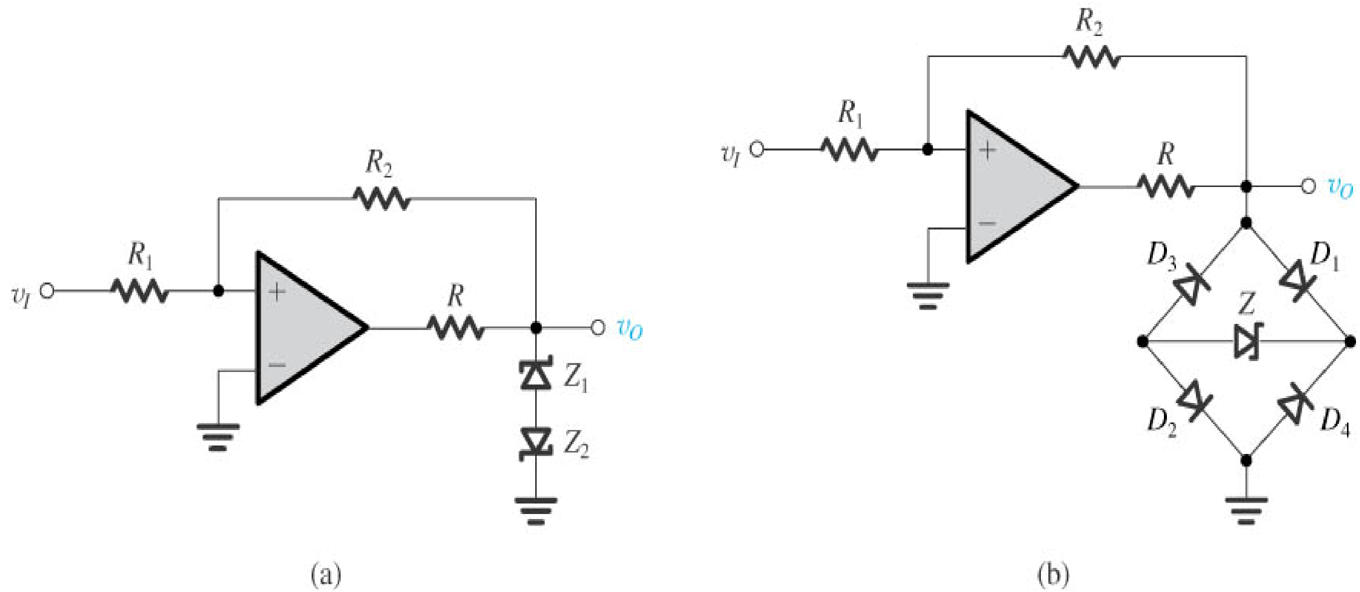


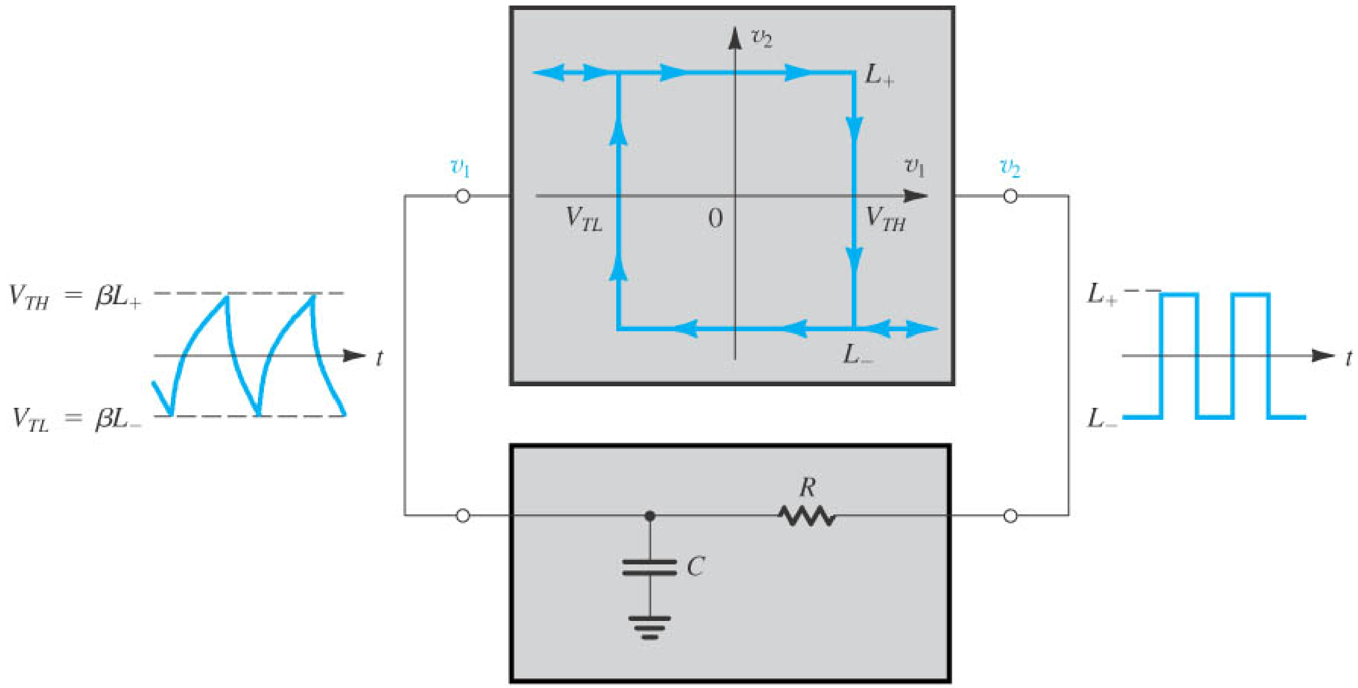
Figure 13.23 Limiter circuits are used to obtain more precise output levels for the bistable circuit. In both circuits the value of R should be chosen to yield the current required for the proper operation of the zener diodes. **(a)** For this circuit $L_+ = V_{Z_1} + V_D$ and $L_- = -(V_{Z_2} + V_D)$, where V_D is the forward diode drop. **(b)** For this circuit $L_+ = V_Z + V_{D_1} + V_{D_2}$ and $L_- = -(V_Z + V_{D_3} + V_{D_4})$.

- By cascading the op amp with a limiter circuit.

→ The output levels of the bistable circuit can be made more precise.

Generation of Square and Triangular Waveforms Using Astable Multivibrators

❑ Operation of the Astable Multivibrator



– The bistable multivibrator with inverting transfer characteristics in a feedback loop with an RC circuit results in a square-wave generator.

– The circuit has no stable state → Astable multivibrator

Operation of the Astable Multivibrator

- During the charging interval $T_1(\tau = RC)$

$$v_- = L_+ - (L_+ - \beta L_-)e^{-t/\tau}$$

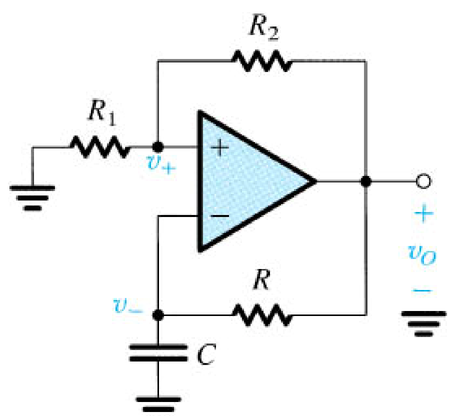
$$T_1 = \tau \ln \frac{1-\beta(L_-/L_+)}{1-\beta}$$

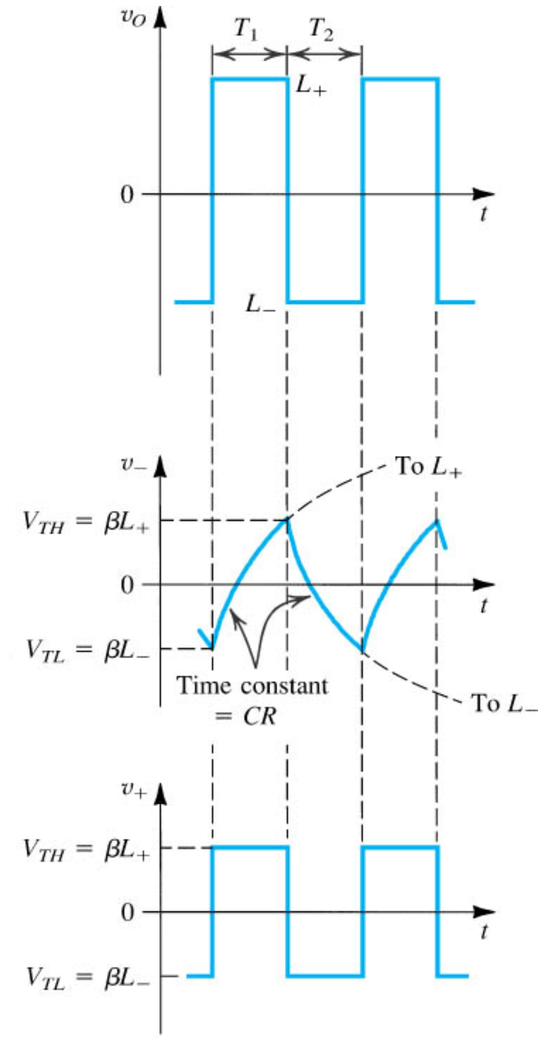
- Similarly T_2

$$v_- = L_- - (L_- - \beta L_+)e^{-t/\tau}$$

$$T_2 = \tau \ln \frac{1-\beta(L_+/L_-)}{1-\beta}$$

$$\therefore T = T_1 + T_2 = 2\tau \ln \frac{1+\beta}{1-\beta}$$





Generation of Triangular waveforms

- Curing the internal T_1

$$\frac{V_{TH}-V_{TL}}{T_1}=\frac{L_+}{CR}$$

$$T_1 = CR\frac{V_{TH}-V_{TL}}{L_+}$$

- Similary

$$\frac{V_{TH}-V_{TL}}{T_2}=\frac{-L_-}{CR}$$

$$T_2 = CR\frac{V_{TH}-V_{TL}}{-L_-}$$