Chapter 16

Oscillator Circuits and Applications

16.0 Introduction

Oscillator contains circuit that generates an output signal without necessity of an input signal. It is a circuit that produces a repetitive waveform on its output with only dc supply as input. The oscillator can be sinusoidal or non-sinusoidal type. They can be used in many applications such as communication and digital system.

Oscillator operation is based on positive feedback whereby portion β of the output signal V_{out} is feedback without phase ϕ change. This shall mean that there is no phase difference between the input and feedback signal.

Many oscillator circuits can be designed using operational amplifier circuit. It can generate various types of waveform with no input other than dc supply. These are known as signal generators or oscillators.

16.1 Principles of Oscillator

With exception such as relaxation oscillator, the operation of oscillator is based on principle of positive feedback where portion of the output signal is feedback into input without phase change. Thus, it reinforces the input and sustains the continuous sinusoidal output. Beside this, the phase shift of feedback signal must be either 0^o or 360^o. The last requirement is the loop gain T of amplifier must be equal to one, which is also named as *Barkhausen criterion*. Thus mathematically, the loop gain T is

$$T = A_V\beta = 1 \tag{16.1}$$

where A_V is the voltage gain of the amplifier and $\beta = \dfrac{V_f}{V_{out}}$ is the feedback portion of output voltage. If A_V is equal to 10 then the feedback portion β should be 1/10. The principles of the oscillator are illustrated in Fig. 16.1. The transfer function of the circuit shall be $A_f = \dfrac{A_V}{1 - A_V\beta}$.

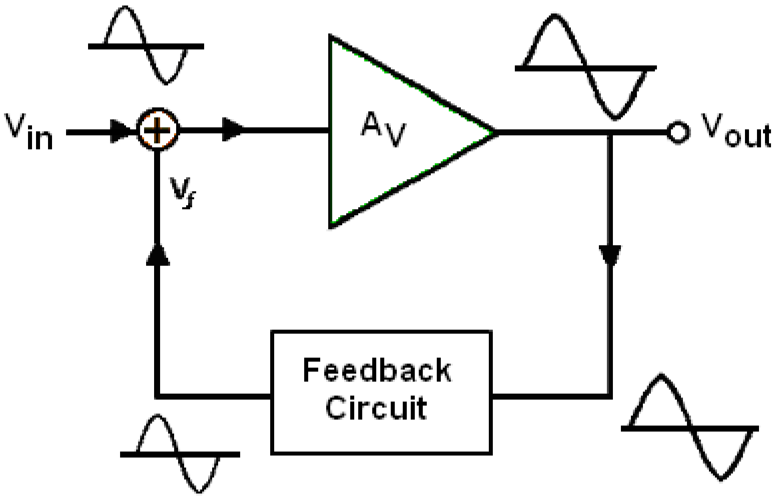


Figure 16.1: Principle of sinusoidal oscillator

16.2 Non Sinusoidal Oscillator

Triangular wave, square wave, and waveform from voltage controlled oscillator VCO are examples of non-sinusoidal waveforms. Generally, they can be designed using operational amplifier, resistor, and capacitor.

16.2.1 Triangular Wave Oscillator

A basic triangular wave generator is shown in Fig. 16.2 and its waveform is shown in Fig. 16.3. When the switch is at position 1, which is at negative voltage, the output of the operational amplifier will ramp from negative to positive voltage. Likewise, when the switch is at position 2, which is at positive voltage, the output of the op-amp will ramp from positive to negative voltage.

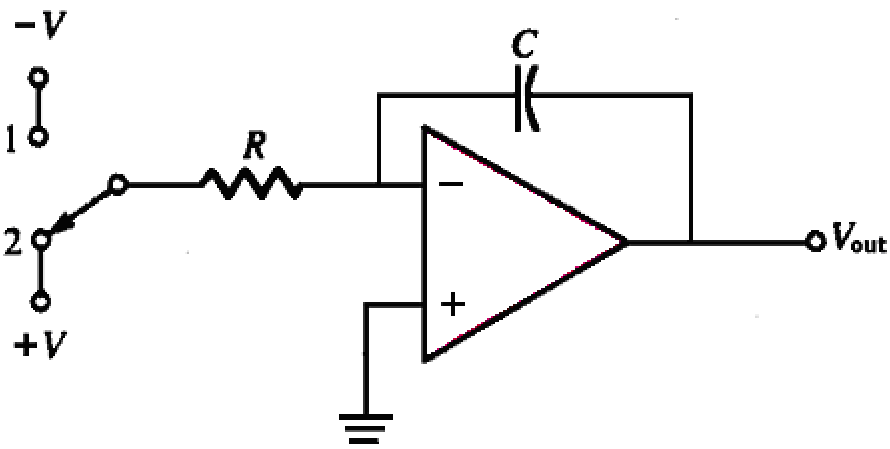


Figure 16.2: A basic triangular waveform generator

The waveform of the basic triangular waveform is shown in Fig. 16.3, which is derived from integrator $V_{out} = -\frac{1}{RC}\int_{t_1}^{t_2} V dt + V_C(t=0)$, where $V_c(t = 0)$ is the voltage of capacitor at time t = 0.

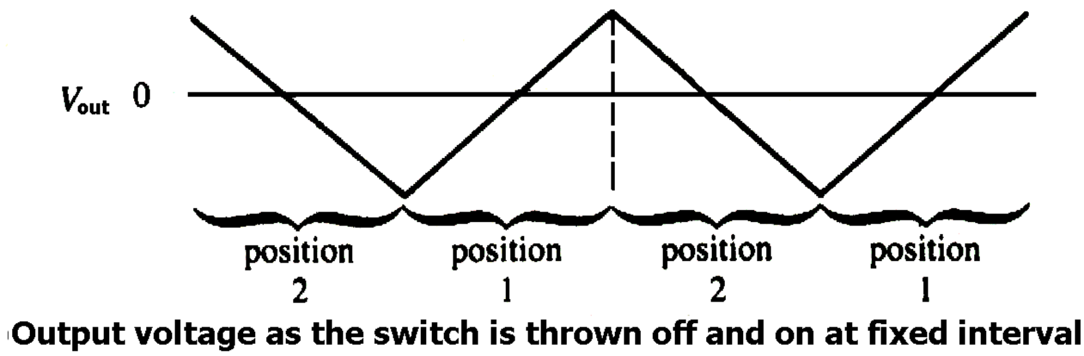


Figure 16.3: The basic principle of a triangular wave generator

A practical triangular wave generator is shown in Fig. 16.4 whereby its positive and negative peak voltage and period can be specified.

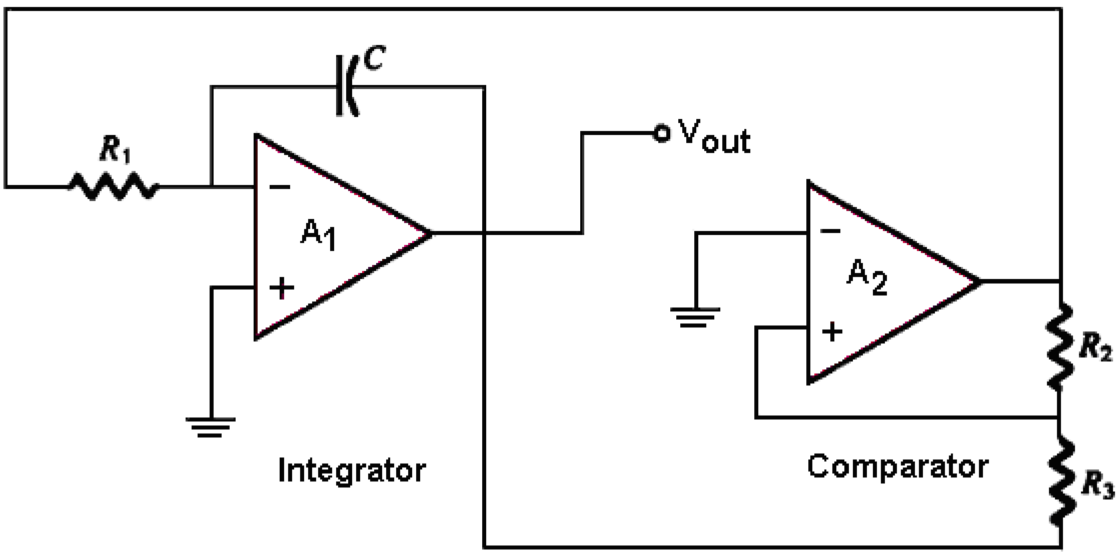


Figure 16.4: A practical triangular wave generator

The peak-to-peak output voltage $V_{out(PP)}$ is,

$$V_{out(PP)} = V_{UTP} - V_{LTP} \qquad (16.2)$$

where $V_{UTP} = \frac{R_3}{R_2}(+V_{out(max)})$ and $V_{LTP} = \frac{R_3}{R_2}(-V_{out(max)})$. $+V_{out(max)}$ and $-V_{out(max)}$ are the positive and negative output swing of the comparator. The resonant frequency f_r of this triangular wave is

$$f_r = \frac{1}{4R_1C} \cdot \left(\frac{R_2}{R_3}\right) \tag{16.3}$$

Based on equation (16.3), by varying the value of resistor R_1 will change the frequency of the oscillator but not the peak-to-peak voltage, which is governed by equation (16.2).

16.2.2 Saw tooth Wave Oscillator

A saw tooth wave generator utilizes the concept of voltage-controlled oscillator VCO. It can be designed by using a programmable unijunction transistor PUT and an operational amplifier integrator arranged as shown in Fig. 16.5.

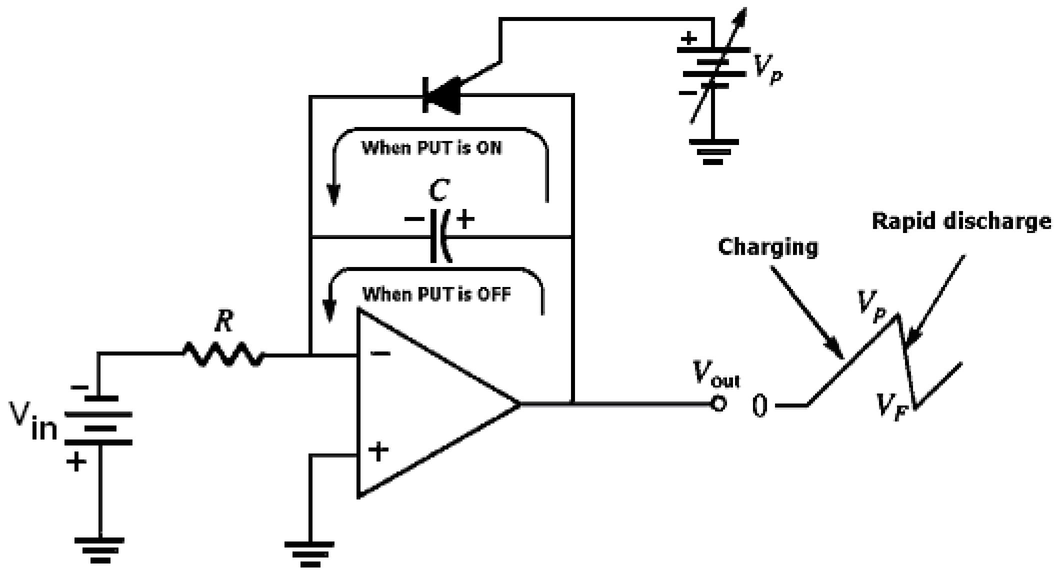


Figure 16.5: A saw tooth wave generator circuit using PUT

A negative input voltage V_{in} is used to establish a positive ramping voltage whereby at this stage the capacitor C is charging and the output V_{out} is ramping up. As soon as the output voltage reaches the programmed voltage V_P of programmable unijunction transistor PUT plus the forward voltage V_F of the diode, which is 0.7V, the PUT is conducting causing the capacitor to discharge and output voltage drop abruptly to the forward voltage V_F of the programmable unijunction transistor PUT.

The period T of the wave is

$$T = \frac{V_P - V_F}{|V_{in}|RC} \tag{16.4}$$

where $|V_{in}|RC$ is ramping rate of output voltage.

Saw tooth wave can also be designed using the triangle wave circuit with inclusion of reference voltage V_{ref} at the non-inverting input pin of the integrating operational amplifier. The circuit of the design is shown in Fig. 16.6.

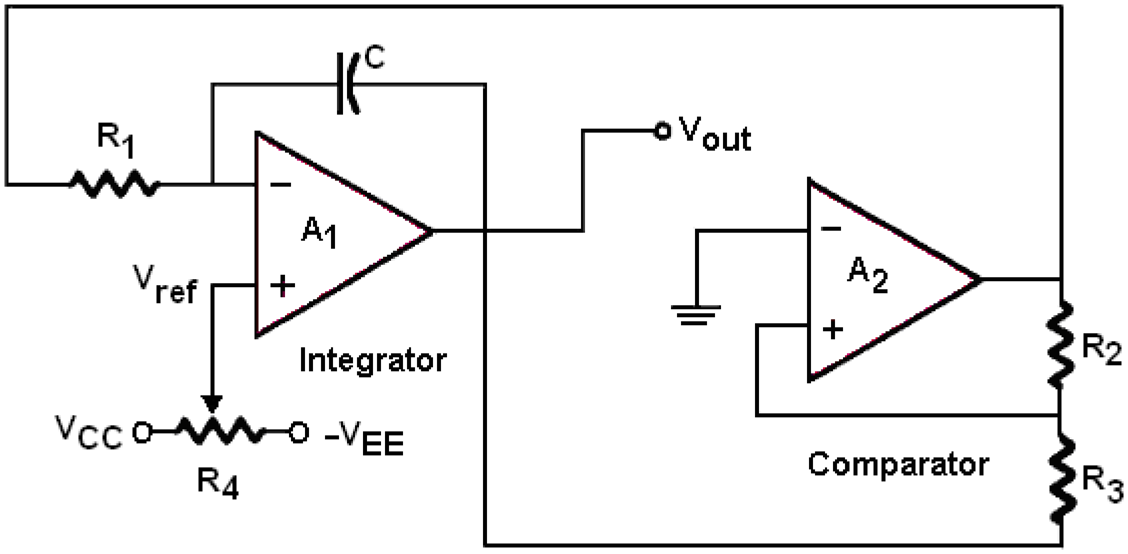


Figure 16.6: A saw tooth wave generator circuit design using square wave circuit

R_4 is the potentiometer that is used to provide V_{ref} voltage to the integrator. By control V_{ref} voltage, shape of the saw tooth wave can be varied.

The time t_1 of the positive ramping of the wave is $t_1 = \frac{2R_1C}{(+V_{out(max)}) - V_{ref}} \cdot \frac{R_3}{R_2}(+V_{out(max)})$. The time t_2 for the negative ramping of the wave is $t_2 = \frac{2R_1C}{(-V_{out(max)}) + V_{ref}} \cdot \frac{R_3}{R_2}(-V_{out(max)})$. If we assume that the saturation voltages of the operational amplifier is the same. i.e. $(+V_{out(max)}) = (-V_{out(max)}) = V_{out(max)}$, then the period T of the saw tooth wave is equal to

$$T = \frac{4R_1C(V_{out(max)})^2}{(V_{out(max)})^2 - V_{ref}^2} \cdot \frac{R_3}{R_2} \tag{16.5}$$

The duty cycle of the saw tooth wave is equal to $t_1/T = \frac{2R_1C}{(V_{out(max)}) - V_{ref}} \cdot \frac{R_3}{R_2}(V_{out(max)}) \cdot \frac{(V_{out(max)})^2 - V_{ref}^2}{4R_1C(V_{out(max)})^2} \cdot \frac{R_2}{R_3} = \frac{1}{2}\left[1 + \frac{V_{ref}}{V_{out(max)}}\right]$. Fixing duty cycle and knowing $V_{out(max)}$, V_{ref} can be known and adjusted using R_4.

16.2.3 Square Wave Oscillator

The square wave oscillator is designed based on the principle of charging and discharge of the capacitor and comparator, which is shown in Fig. 16.7.

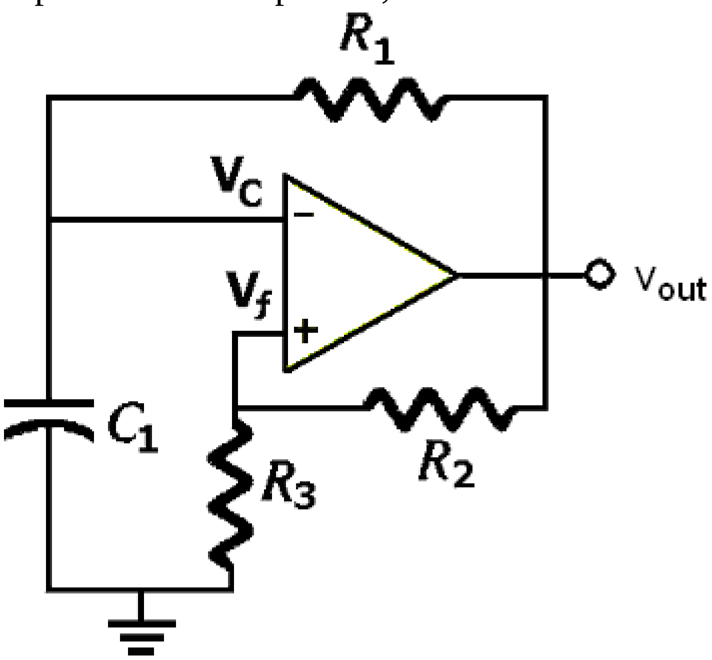


Figure 16.7: Square wave oscillator

When the circuit is turned on, the voltage at inverting input is zero. Thus, the output of the op-amp is swung to maximum positive value, which should be the positive saturation voltage of the operational amplifier. The capacitor begins to charge until the voltage V_C is just greater than the feedback voltage V_f, which is $\frac{R_3}{R_2+R_3}V_{out}$. At this point, the output of the op-amp is swung to minimum negative value, which is the negative saturation voltage of the op-amp. This would cause the capacitor to discharge. When the voltage V_C reaches negative V_f, the output of op-amp swings to positive again. This process repeats and generates square wave. The period T of the oscillation can be shown to follow equation (16.6).

$$T = 2R_1C_1\ln(1 + 2R_3/R_2) \qquad (16.6)$$

It can be calculated using charging equation $V_{C1} = V_{out} + (-V_f - V_{out})\exp(-t/R_1C_1)$ for $V_{C1} = V_f$ and discharging equation $V_{C1} = -V_{out} + (V_f + V_{out})\exp(-t/R_1C_1)$ for $V_{C1} = -V_f$.

If the value of resistor R_2 is $R_2 = 1.1639R_3$, then the period T is $T = 2R_1C_1$.

16.3 Oscillators with RC Feedback Circuit

Many oscillators are designed utilizing RC network and resistor divider circuits. RC network is basically used to determine the resonant frequency of the oscillator, whilst resistor divider circuit is used to provide attenuated feedback. Generally oscillator with RC network can provide frequency not more than 1MHz. For oscillator that can provide frequency more than 1MHz, LC network is used.

16.3.1 Wien-Bridge Oscillator

Wien-Bridge oscillator is an oscillator that meets the principle of oscillator. Its circuit is shown in Fig. 16.8.

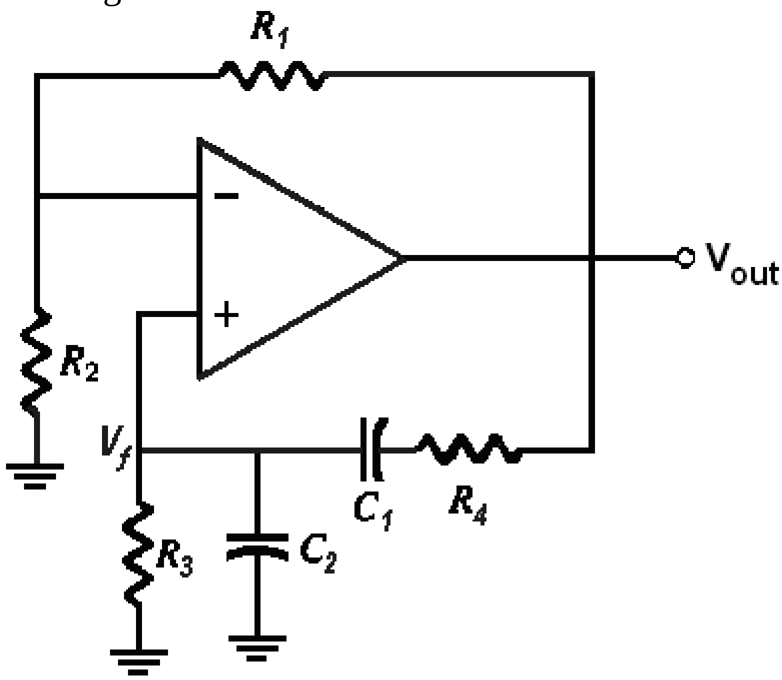


Figure 16.8: Wien-Bridge oscillator

There is a lead-lag RC network whereby C_1 and R_3 leads and R_4 and C_2 lags. Reactance χ_{C_1} of capacitance C_1 is significantly affecting the V_{IN+} at low frequency, whilst reactance χ_{C_2} of capacitor C_2 equal to $1/j\omega C_2$ is significant affecting at high frequency. If $C_1 = C_2$ and $R_4 = R_3$, there will be no phase-shift because the phase lead is compensated by phase lag.

From the analysis of the circuit, the portion of the output feedback to input $\beta = \frac{V_f}{V_{out}}$ shall be

$$\beta = \frac{V_f}{V_{out}} = \frac{j\omega RC}{3j\omega RC + (1-\omega^2R^2C^2)} \tag{16.7}$$

Applying *Barkhausen* principle of oscillator design this shall mean that overall loop gain A $= \dfrac{j\omega RC}{3j\omega RC + (1-\omega^2R^2C^2)} A_V = 1$. From real part and imaginary part of the complex number solution, the resonant frequency f_r of the oscillator shall be

$$f_r = \frac{1}{2\pi RC} \tag{16.8}$$

The voltage gain A_V of the amplifier should be 3. This shall mean that feedback portion β is 1/3.

Since the voltage gain of the amplifier A_V should be 3, this shall mean that

$$A_V = \frac{R_1 + R_2}{R_2} = 3 \tag{16.9}$$

In order for Wien-Bridge oscillator to start oscillating, the initial voltage gain A_V should be slightly more than three. Adding an extra circuit, which is called *stability circuit* as shown in Fig. 16.9, will provide self-start and sustain the oscillation.

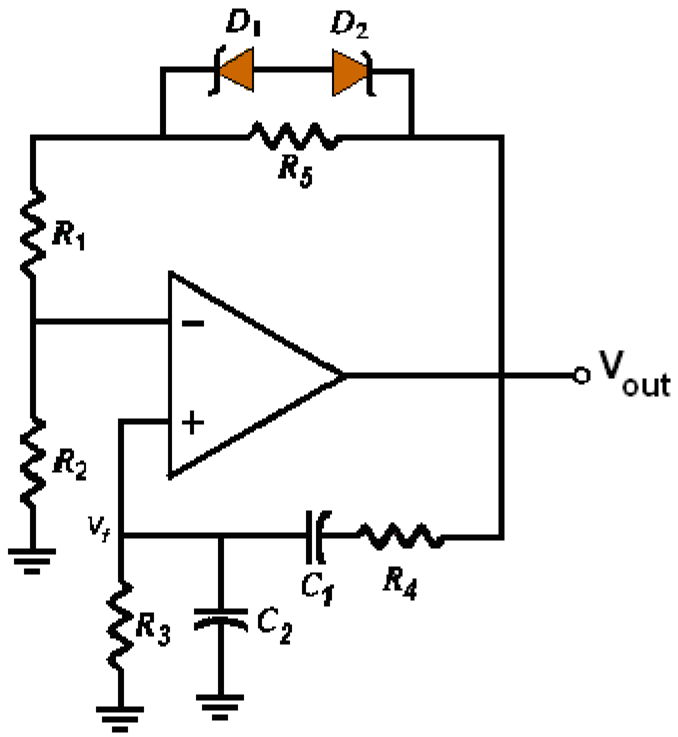


Figure 16.9: Self-starting and sustaining Wien-Bridge oscillator

Before zener diode D_1 and D_2 conduct, the voltage gain A_V of the amplifier is

$$A_V = \frac{R_1 + R_2}{R_2} + \frac{R_5}{R_2} = 3 + \frac{R_5}{R_2} \tag{16.10}$$

Thus, it meets the conditions of closed loop gain greater than 1 for self-starting of the oscillator.

When the output voltage reaches the zener breakdown voltage plus 0.7V, the zener diode conducts. Its forward resistance would be much smaller than the resistance value of resistor R_5. Thus, the closed loop gain will be back to 1 and oscillation would be sustained.

16.3.2 Operational Amplifier Phase-Shift Oscillator

The operational amplifier phase-shift oscillator is another oscillator type that meets the principles of oscillator design. Its circuit is shown in Fig. 16.10.

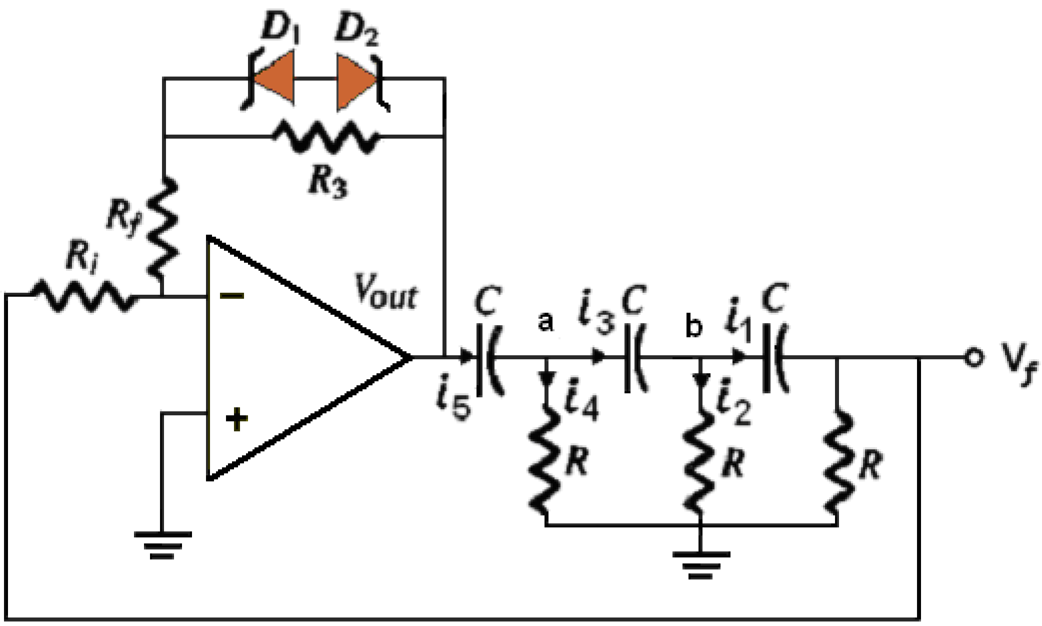


Figure 16.10: Operational amplifier phase-shift oscillator

The feedback portion of the oscillator can be derived by applying Kirchhoff's current law at node a and node b respectively. Current i_1 is $i_1 = \frac{V_b sC}{1 + sRC}$, $i_2 = \frac{V_b}{R}$,

$i_3 = (V_a - V_b)sC$, $i_4 = \frac{V_a}{R}$, and $i_5 = (V_{out} - V_a)sC$. Voltage at node b is also equal to $V_b = \frac{1+sRC}{sRC} \cdot V_f$.

At node b, current $i_3 = i_1 + i_2$; $(V_a - V_b)sC = \frac{V_b sC}{1+sRC} + \frac{V_b}{R}$. This implies that $V_a = V_b\left(\frac{1}{1+sRC} + \frac{1}{sRc} + 1\right)$.

At node a, $i_5 = i_1 + i_2 + i_4$; $(V_{out} - V_a)sC = \frac{V_b sC}{1+sRC} + \frac{V_b}{R} + \frac{V_a}{R}$. This implies that $V_{out}sC = \frac{V_b sC}{1+sRC} + \frac{V_b}{R} + \frac{V_a}{R} + V_a sC$. Substituting the expression of V_a and V_b into the above equation yields equation (16.10), which is the feedback portion β.

$$\beta = \frac{V_f}{V_{out}} = \frac{j^3\omega^3R^3C^3}{j^3\omega^3R^3C^3 + 6j^2\omega^2R^2C^2 + 5j\omega RC + 1} \tag{16.11}$$

After applying *Barkhausen* principle of oscillator design, from real part of the complex number denominator, which is $(1\text{-}6\omega^2R^2C^2) = 0$, the resonant frequency f_r of the oscillator shall be

$$f_r = \frac{1}{2\pi\sqrt{6}RC} \tag{16.12}$$

From the imaginary part of the complex number, which is $\frac{j^3\omega^3R^3C^3A_V}{j^3\omega^3R^3C^3 + 5j\omega RC} = 1$, the voltage gain A_V of the amplifier shall be

$$A_V = 1 - \frac{5}{\omega^2R^2C^2} \tag{16.13}$$

From equation (16.12), this shall mean that $\omega^2 = \frac{1}{6R^2C^2}$. Substituting this expression into equation (16.13), it yields $A_V = -29$. This shall mean that the attenuation of the three-section RC feedback is 29. Therefore, the value of $|R_f/R_i|$ should be 29.

16.3.3 Quadrature Oscillator

Quadrature oscillator as shown Fig. 16.11 generates two signals both sine and cosine that are in quadrature meaning out of phase by 90^0. The sine and cosine outputs can be arbitrary assigned. It is not necessary to assign output of operational amplifier 1 as sine output and output of operational amplifier 2 as cosine output.

The feedback portion β is V_{in1}/V_{O2} = $\frac{1}{1+sRC}$. The gain of operational amplifier 1 is A_{VO1} = $\frac{1+sRC}{sRC}$. Similarly, the gain of operational amplifier 2 is $A_{VO2} = -\frac{1}{sRC}$. The overall gain A_V is $-\frac{1+sRC}{(sRC)^2}$. Applying *Barkhausen* principle of oscillator design, the loop-gain of the oscillator is $A_V\beta = -\frac{1}{1+sRC}\cdot\frac{1+sRC}{(sRC)^2} = -\frac{1}{(sRC)^2}$ and is equal to one. This shall imply that $s^2R^2C^2 = -1$. Based on this equation, frequency f of the oscillator is

$$f = \frac{1}{2\pi RC} \tag{16.14}$$

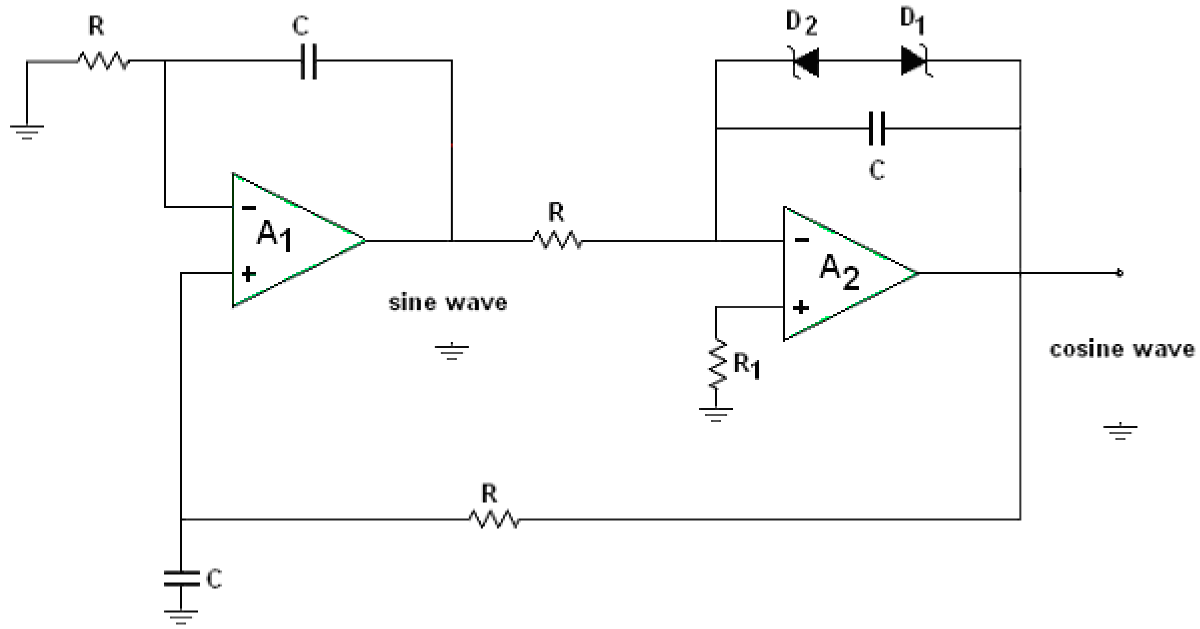


Figure 16.11: Quadrature oscillator

Substituting equation (16.14) into the equation of feedback portion β, which is V_{in1}/V_{o2} = $\frac{1}{1+sRC}$, the feedback portion β is $\frac{1}{1+j} = \frac{1}{\sqrt{2}}\angle -45^0$. This shall mean that its magnitude is $\frac{1}{\sqrt{2}}$. This implies that overall open loop gain of the oscillator A_V is $\sqrt{2} = 1.41$. The gain function is A_{Vo1} = $\frac{1+sRC}{sRC}$, after substituting ω with equation (16.14), the gain function becomes A_{VO1} = $-j(1+j)$ = $\sqrt{2}\angle -45^0$. Similarly the gain function A_{Vo2} is A_{Vo2} = $-\frac{1}{sRC}$ = $1\angle 90^0$. The total phase shift is $-45^0 - 45^0 + 90^0 = 0$ that meets the design principle.

16.3.4 Three-Phase Oscillator

Three-phase oscillator as shown Fig. 16.12 generates three sinusoidal voltages of equal magnitude but displaced by 120^0 from each other.

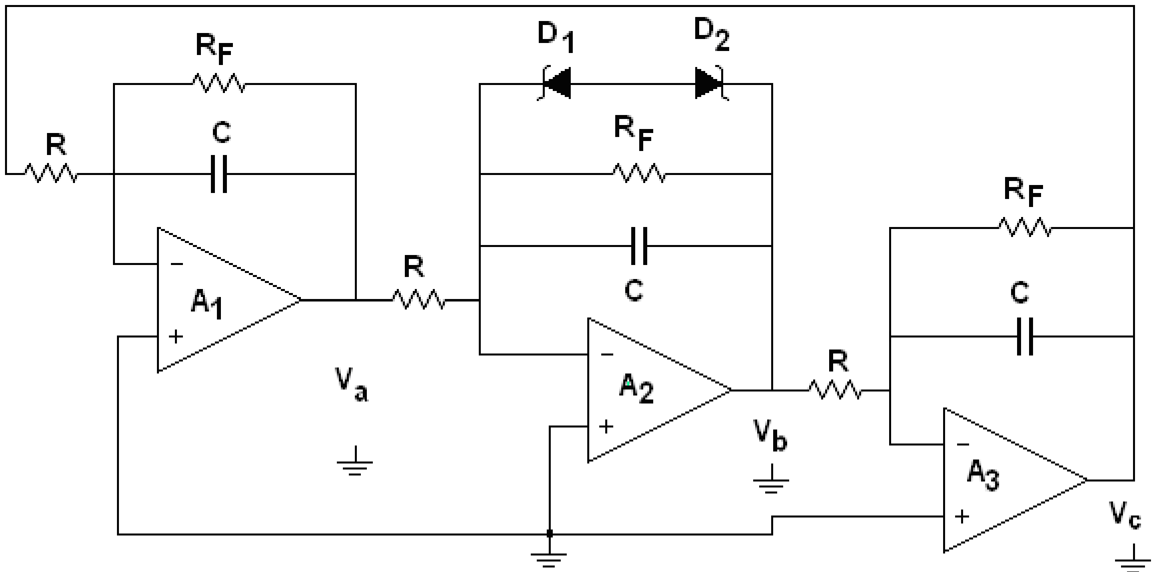


Figure 16.12: Three-phase oscillator

The feedback portion β is equal to one. The gain of each operational amplifier is A_{Vi} = $-\frac{R_F \parallel (1/sC)}{R} = -\frac{R_F/R}{1+sR_FC}$. The overall gain A_V of the oscillator is A_V = $\left(-\frac{R_F/R}{1+sR_FC}\right)^3$. Applying *Barkhausen* principle of oscillator design, the loop-gain of the oscillator shall be $A_V\beta = \left(-\frac{R_F/R}{1+sR_FC}\right)^3 =$ 1. This implies that $(sR_FC)^3 + 3(sR_FC)^2 + 3(sR_FC) + 1 + \left(\frac{R_F}{R}\right)^3 = 0$. Equating the imaginary part of the equation, it gives rise to the frequency of the oscillator to be

$$f = \frac{\sqrt{3}}{2\pi R_F C} \tag{16.15}$$

Equating the real part of the equation, it gives rise to the gain of individual integrator to be $\frac{R_F}{R} = 2$. Substituting the gain equation and frequency equation into $A_{Vi} = -\frac{R_F/R}{1+sR_FC}$, it produces gain function $-\frac{2}{1+j\sqrt{3}} = 1\angle 120^0$.

16.4 Oscillators with LC Feedback Circuit

LC element is used for oscillators that generate more than 1.0 MH frequency. Also because of frequency limitation of most operational amplifier, discrete transistor is used as gain element of LC oscillator.

16.4.1 Colpitts Oscillator

Colpitts oscillator utilizing operational amplifier is shown in Fig. 16.13 and a discrete bipolar junction transistor version is shown in Fig. 16.14. The oscillator uses LC network in the feedback loop. The combination of C_1 and C_2 and L act as a parallel resonant circuit. Using Kirchhoff's current law and ac model as shown in Fig. 16.15 for circuit, the current at output node is $g_mV_{be} + I_1 + I_2 + I_3 = 0$ i.e. $g_mV_{be} + \frac{V_{out}}{R_3} + \frac{V_{out}}{1/j\omega C_1} + \frac{V_{out}}{j\omega L + 1/j\omega C_2} = 0$ and using voltage divider law, base-to-emitter voltage V_{be} equals to $V_{be} = \frac{1/j\omega C_2}{j\omega L + 1/j\omega C_2} V_{out}$. From these two equations, it yields expression $\left(g_m + \frac{1}{R_3} - \frac{\omega^2 LC_2}{R_3}\right) + j\omega\left[(C_1 + C_2) - \omega^2 LC_1C_2\right] = 0$. Thus, the resonant frequency f_r is obtained from the imaginary part of the equation, which is $j\omega\left[(C_1 + C_2) - \omega^2 LC_1C_2\right] = 0$ and it is expressed as

$$f_r = \frac{1}{2\pi\sqrt{L\frac{C_1C_2}{C_1+C_2}}} \tag{16.16}$$

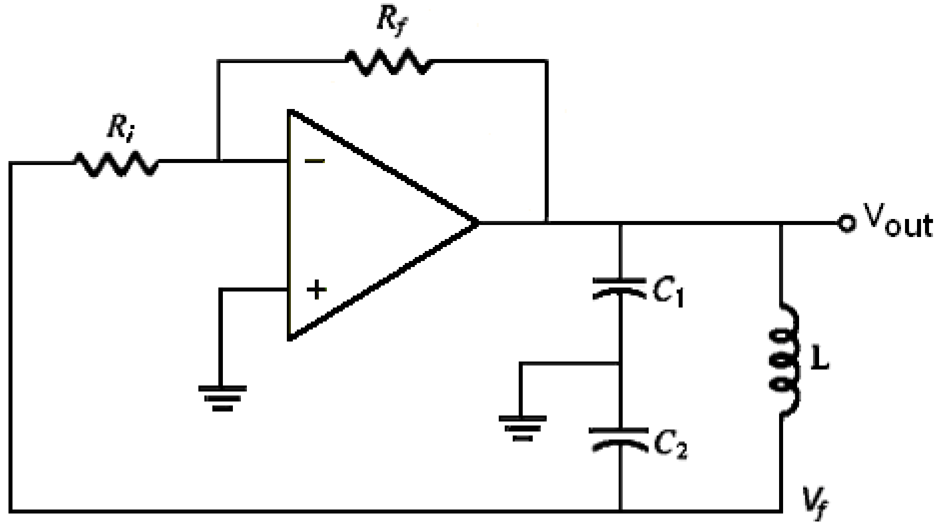


Figure 16.13: Colpitts oscillator

From the real part of equation $\left(g_m + \frac{1}{R_3} - \frac{\omega^2 LC_2}{R_3}\right) + j\omega\left[(C_1 + C_2) - \omega^2 LC_1C_2\right] = 0$, it yields $\frac{\omega^2 LC_2}{R_3} = g_m + \frac{1}{R_3}$. Combining this equation with equation (16.16), it yields gain $A_V = g_mR_3 = C_2/C_1$. Thus, the feedback portion shall be $\beta = C_1/C_2$.

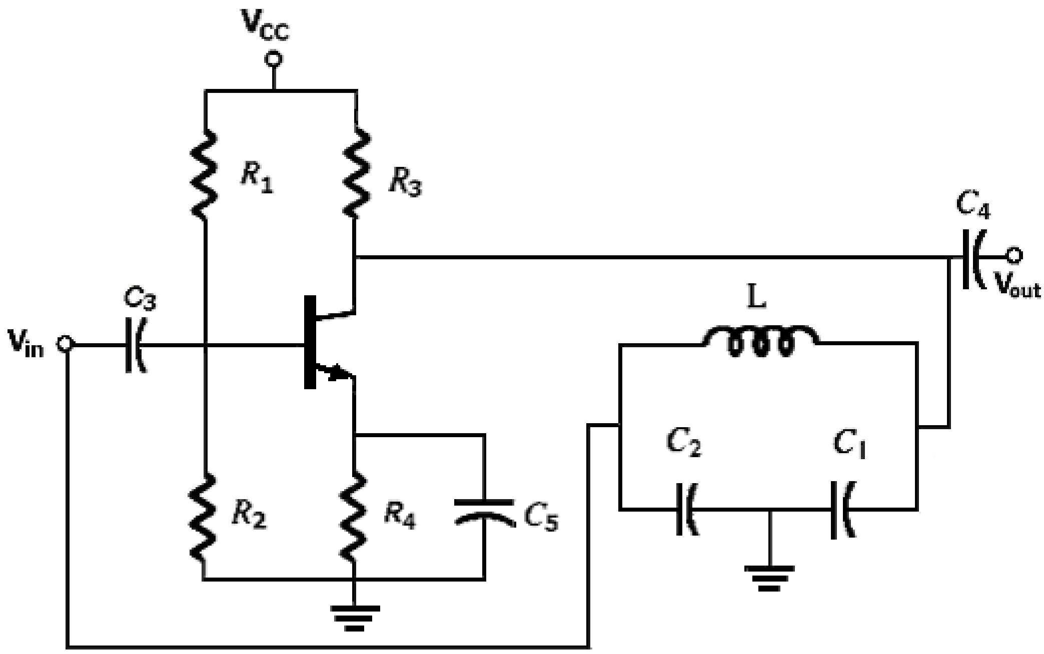


Figure 16.14: A discrete Colpitts oscillator

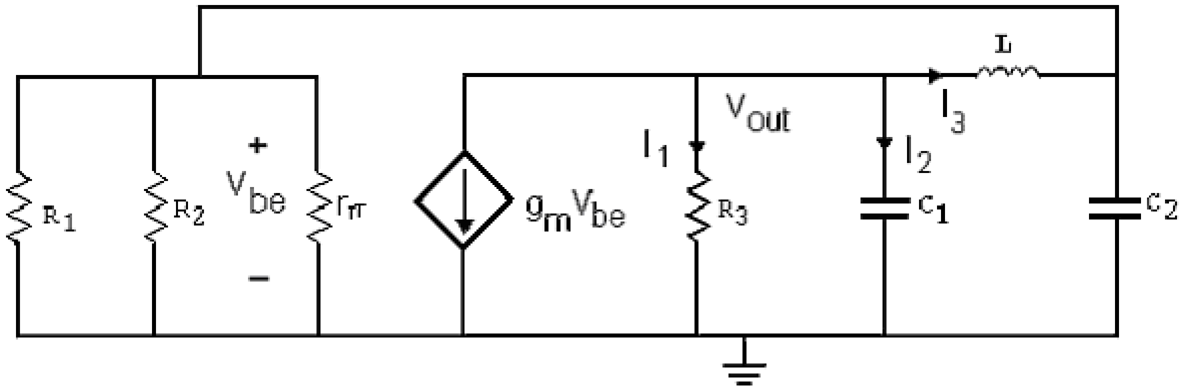


Figure 16.15: ac model of a discrete Colpitts oscillator

The impedance of the transistor will act as the load on the resonant circuit and reduced the quality factor $Q = \frac{\omega_0}{\omega_H - \omega_L}$ of the circuit thus reduced the resonant frequency of the circuit. The equation of resonant frequency for Colpitts oscillator including Q factor is

$$f_r = \frac{1}{2\pi\sqrt{L\frac{C_1C_2}{C_1+C_2}}} \cdot \sqrt{\frac{Q^2}{Q^2+1}} \qquad (16.17)$$

If the Q factor is less than 10, the resonant frequency is significant reduced. On the other hand, if the Q factor is greater than 10, the factor $\sqrt{\frac{Q^2}{Q^2+1}}$ is approximately equal to one, which shall mean it does not affect the resonant frequency of the circuit.

16.4.2 Hartley Oscillator

The Hartley oscillator is shown in Fig. 16.16. Using the similar approach like the way how gain and resonant frequency are derived for Colpitts oscillator, the resonant frequency of Hartley oscillator can be shown equal to

$$f_r = \frac{1}{2\pi\sqrt{(L_1 + L_2)C}} \qquad (16.18)$$

where inductor L_1, L_2, and capacitor C act as resonant circuit.

The feedback portion β of this oscillator is L_1/L_2. Therefore, the gain of this amplifier should be slightly greater than L_2/L_1 for self-starting. This shall mean that the ratio L_2/L_1 is also equal to ratio R_f/R_i, which is the gain A_V of the amplifier.

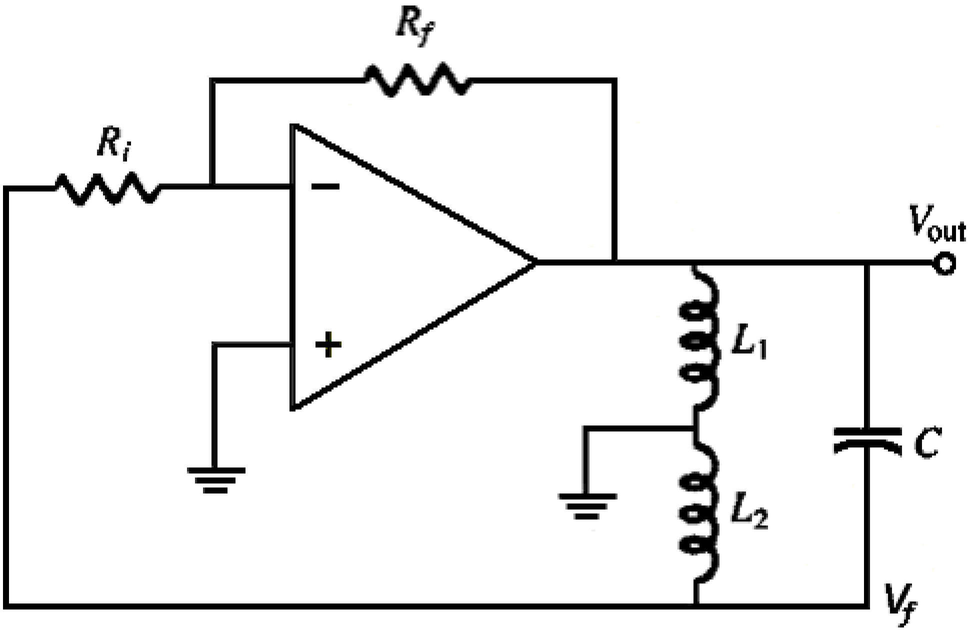


Figure 16.16: Hartley oscillator

16.4.3 Clapp Oscillator

The Clapp oscillator is a variant of the Colpitts oscillator. As shown in Fig. 16.16, the equivalent capacitance C_{eq} is equal

$$C_{eq} = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}} \tag{16.19}$$

If the Q factor is greater than 10, then the resonant frequency of the oscillator is equal to

$$f_r = \frac{1}{2\pi\sqrt{LC_{eq}}} \tag{16.20}$$

If capacitor C_3 is very much smaller than C_1 and C_2, then the equivalent capacitor C_{eq} is $C_{eq} \cong C_3$.

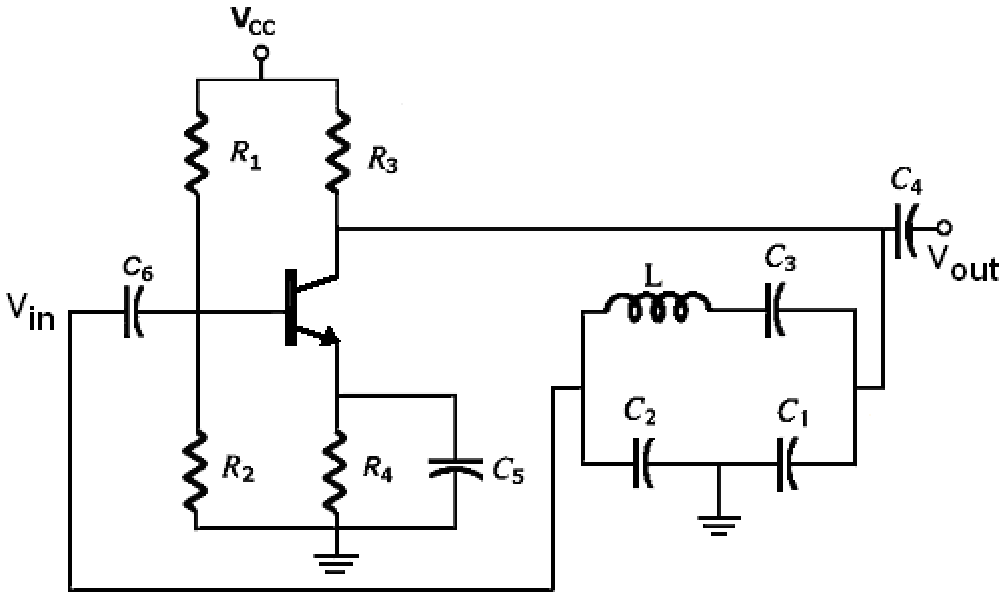


Figure 16.17: Clapping oscillator

16.4.4 Armstrong Oscillator

Armstrong oscillator uses the transformer and the feedback is via the secondary coil of the transformer. This oscillator is also called "tickler" oscillator because of the secondary coil is as feedback. The circuit of the oscillator is shown in Fig. 16.18. The resonant frequency f_r of the circuit is

$$f_r = \frac{1}{2\pi\sqrt{L_1 C_1}} \qquad (16.21)$$

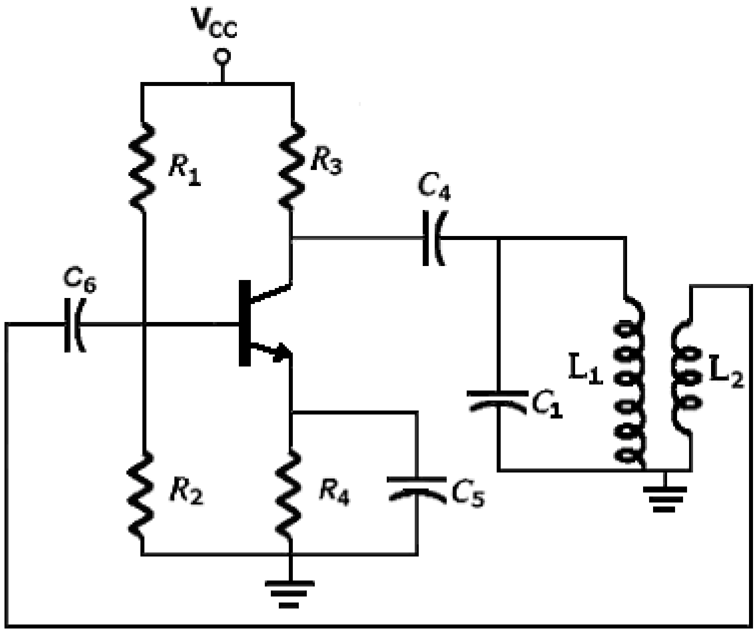
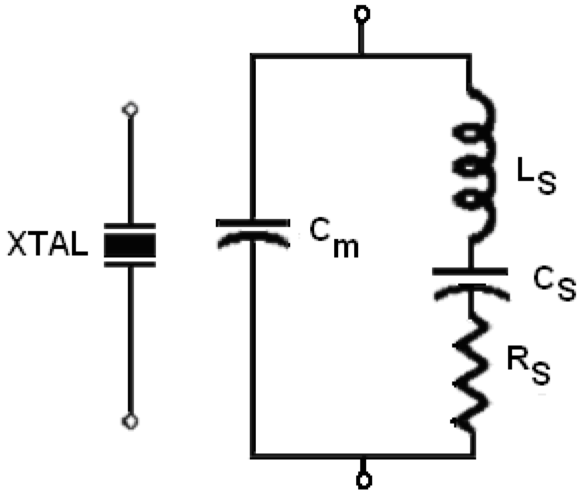


Figure 16.18: Armstrong oscillator

16.5 Crystal Controlled Oscillator

The most accurate and state of arts oscillator is the one that uses piezoelectric crystal in the feedback loop to control the frequency. Quartz is one type of crystalline substance that found in nature that exhibits piezoelectric effect. When a changing mechanical stress is applied to the crystal, it vibrates and a voltage is developed at the frequency of mechanical vibration. If ac voltage is applied, it vibrates at the frequency of applied voltage. The crystal has its nature resonant frequency which is determined by its physical dimension and the way the crystal is cut. A quartz crystal can be represented by the symbol and circuit shown in Fig. 16.19.



(a) Symbol (b) Electrical Equivalent circuit
Figure 16.19: Symbol and equivalent circuit of quartz crystal

Piezoelectric crystal can oscillate in two modes, which are the fundamental or overtone modes. The fundamental mode is the lowest frequency, which is its natural frequency. The fundamental frequency is basically depended on the crystal's dimension, type of cut, and is inversely proportional to the thickness of crystal slab. Most crystal can operate up to 20MHz, thus it is necessary to overtone in the odd multiple of the fundamental mode frequency. Assuming that R_s is small, the impedance $Z(j\omega)$ of the crystal shall be $Z(j\omega)=\dfrac{1}{j\omega C_m}\cdot\dfrac{-\omega^2+(1/L_sC_s)}{-\omega^2+[C_m+C_s)/(L_sC_sC_m)]}$.

A basic crystal oscillator is shown in Fig. 16.20. Capacitor C_C is used to fine tune the frequency of the oscillator.

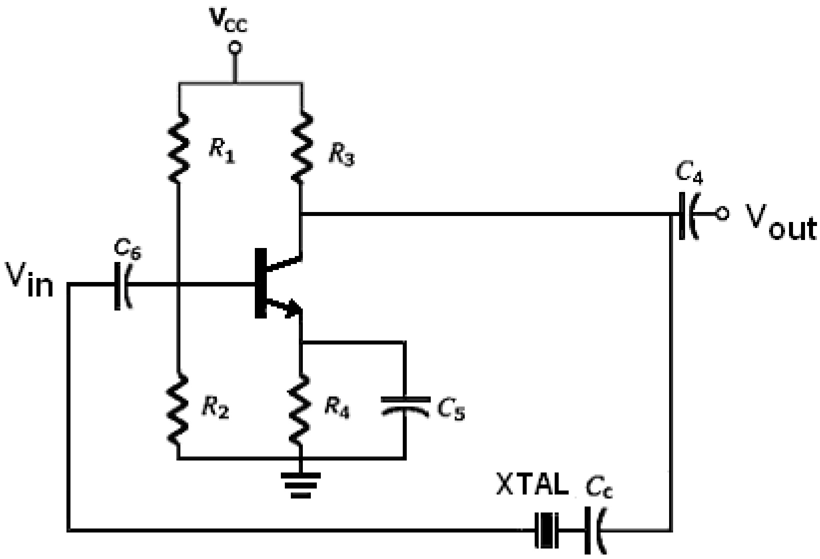


Figure 16.20: Crystal oscillator

16.6 The NE/SE 555 Timer

The NE/SE 555 timer has two comparators, a RS flip-flop, a discharge transistor Q_d, and relative voltage divider formed by resistor R_A, R_B, and R_C as shown in Fig. 16.21. The relative voltage divider can also be externally control using the voltage control pin (5).

NE/SE 555 timer can be configured to form many applications such as monostable multivibrator, voltage control oscillator, square wave generator, astable multivibrator, and etc.

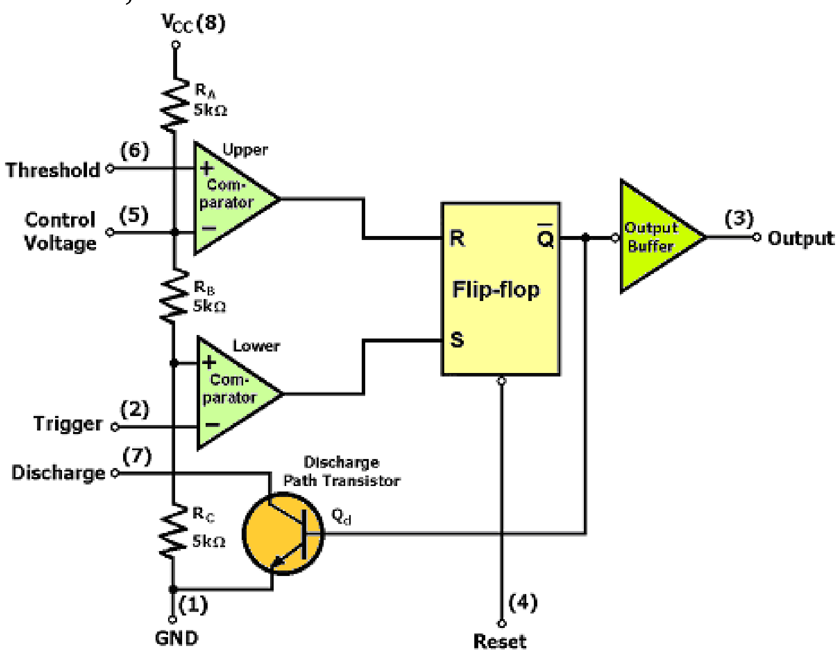


Figure 16.21: The internal structure of NE/SE 555 timer

16.6.1 Astable Multivibrator

The relative voltage divider of NE/SE 555 timer has three resistors of equal value. This would create a $2/3V_{CC}$ reference for the upper comparator and $1/3V_{CC}$ reference for the lower comparator. The output of the comparator is then used to control the output-state of the RS flip-flop. The threshold input is normally connected to an external RC timing circuit, which is shown in Fig. 16.22. When the voltage of external capacitor exceeds $2/3V_{CC}$, the upper comparator reset the flip-flop. This will cause the $\overline{Q}$ state of the RS flip-flop to switch high. This in term causes the transistor Q_d to switch on and allows the capacitor to discharge through it via resistor R_2. When the voltage of the transistor reached $1/3V_{CC}$, this causes the lower comparator to set the RS flip-flop. The $\overline{Q}$ state of flip-flop is set to zero state, switches-off the transistor Q_d and the capacitor will begin to charge up. This process creates a square pulse at the output of the timer.

The frequency of the oscillation is depending on the value of resistor R_1 and R_2 and external capacitor C_{ext} according to the equation (16.22).

$$f_r = \frac{1.44}{(R_1 + 2R_2)C_{ext}} \tag{16.22}$$

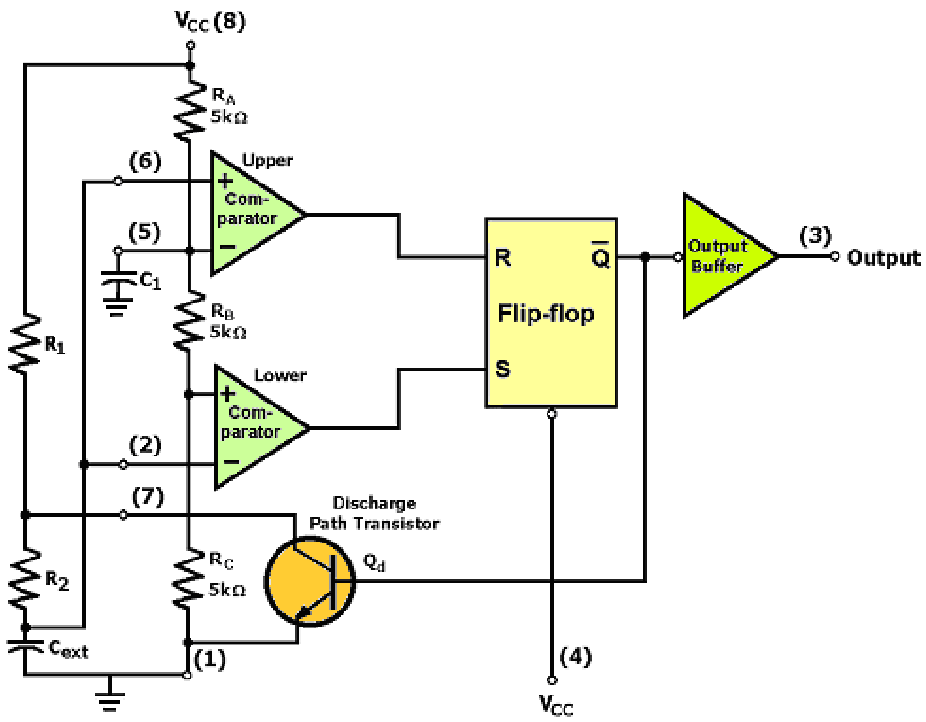


Figure 16.22: NE/SE 555 timer connected as an astable multivibrator

The duty cycle of the pulse at output is not symmetrical because the external capacitor C_{ext} is charged through resistor R_1 and R_2 and discharged through resistor R_2 only. Moreover, the charging occurs from voltage $1/3V_{CC}$ to $2/3V_{CC}$ and the discharging occurs from voltage 2/3 V_{CC} to $1/3V_{CC}$. The charging of external capacitor constitutes the time duration t_H for the output is high and it follows equation (16.23).

$$t_H = 0.693(R_1 + R_2)C_{ext} \tag{16.23}$$

The time t_L is the time duration when the output is low. It follows equation (16.24).

$$t_L = 0.693R_2C_{ext} \tag{16.24}$$

The period T of the output waveform is equal to the sum of t_H and t_L, which is

$$T = 0.693(R_1 + 2R_2)C_{ext} \tag{16.25}$$

The reciprocal of period T is the frequency of the output waveform, which is also equation (16.22). The charging time t_H can be obtained from charging equation $V_{Cext} = V_{CC} + \left(\frac{V_{CC}}{3} - V_{CC}\right)e^{-t/(R_1+R_2)C_{ext}}$ by setting $V_{Cext} = 2V_{CC}/3$, which yields $t_H = (R_1 + R_2)C_{ext} \ln 2$, whilst the discharge time can be obtained from equation $V_{Cext} = \frac{2V_{CC}}{3}e^{-t/R_2C_{ext}}$ by setting $V_{Cext} = V_{CC}/3$, which yields $t_L = R_2C_{ext}$ ln 2. The duty cycle of the waveform follows equation (16.26).

$$\text{Duty cycle} = \frac{t_H}{t_L + t_H}\text{x}100\% = \left(\frac{R_1 + R_2}{R_1 + 2R_2}\right)\text{x}100\% \tag{16.26}$$

To obtain closed to 50% duty cycle, the value of resistor R_2 has to be much greater than the value of resistor R_1. This will make the charging and discharging times of the external capacitor C_{ext} almost the same since $R_1 + R_2$ is approximately equal to R_2. If the pulse is properly negatively offset, then a square wave is obtained.

If the duty cycle of less than 50% is desired, besides selecting value of $R_1 < R_2$, a diode D_1 is connected in parallel with resistor R_2 as shown in Fig. 16.23. This will reduce the effective charging time via resistor R_1 and forward resistor of diode R_F, which is small as compared to value of resistor R_2, to be much smaller than discharge time via resistor R_2. Since, the charge time is much shorter than discharge time, the duty cycle is less than 50%. Duty cycle is now governed by equation (16.27).

$$\text{Duty cycle} = \left(\frac{R_1 + R_{diode}}{R_1 + R_{diode} + R_2} \right) x100\% \qquad (16.27)$$

R_{diode} is the forward resistance of the diode, which can be assumed to be zero as compared with the value of R_1 and R_2.

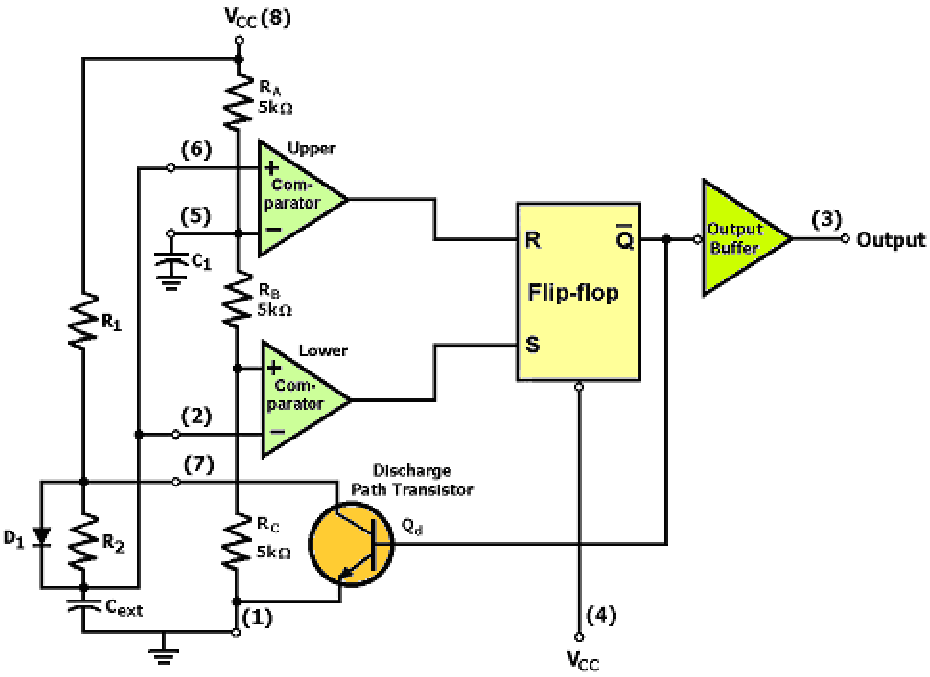


Figure 16.23: NE/SE 555 timer configured for less than 50% duty cycle

16.6.2 Monostable Multivibrator

NE/SE 555 timer can be configured as a monostable multivibrator as shown in Fig. 16.24. In the stable state, the RS flip-flop is in reset state. The output of the timer will be at low state. The voltage V_C at capacitor C is zero volt because the transistor Q_d is switched on. As soon the triggering voltage $V_{trigger}$ is set, which shall mean the voltage is lower than $1/3V_{CC}$. The lower comparator will set the RS flip-flop, which in turn switch-off transistor Q_d and set the output to high state. Since the transistor Q_d is off, the capacitor C begins to charge up. As soon as its voltage reaches $2/3V_{CC}$, the output of upper comparator switches to high state and reset the RS flip-flop. The output of timer shall then switch to low state and the capacitor begins to discharge through the switched-on transistor Q_d. The waveforms of the triggering, voltage at capacitor C, and output voltage are shown in Fig. 16.24.

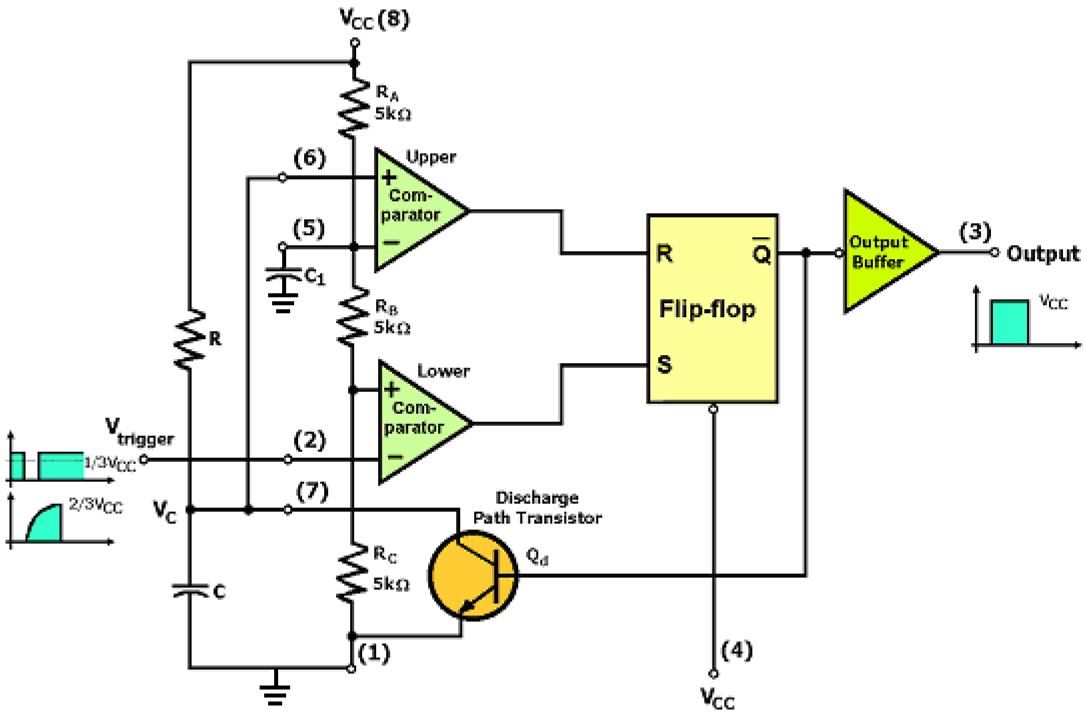


Figure 16.24: NE/SE 555 timer configured as a monostable multivibrator

The pulse width t_{width} of the monostable multivibrator is calculated from the charging time of RC network charging for voltage from zero volt to $2/3V_{CC}$. The pulse width follows equation (16.28).

$$t_{width} = 1.098RC \qquad (16.28)$$

Pulse width t_{width} can be obtained from universal charging/discharging equation $V_C = V_f + (V_i - V_f)\exp(-t/\tau)$, where $V_i = 0$ and $V_f = V_{CC}$ that yields equation $V_C = V_{CC} - V_{CC}\exp(-t_{width}/RC)$. By setting $V_C = 2V_{CC}/3$, the final charge voltage of capacitor, it yields $t_{width} = RC \ln 3$.

If the triggering voltage $V_{trigger}$ is a pulse of known frequency and the time t_{width} is controlled such that it is equal to 1.2 times of the period T of $V_{trigger}$ pulse then the monostable multivibrator is become a divide-by-two frequency divider.

In general, for the monostable multivibrator to be configured as divide-by-*n* frequency divider, the pulse-width t_{width} should be equal to $[0.2 + (n\text{-}1)]T$.

16.6.3 Voltage Controlled Oscillator

The NE/SE 555 timer can also be configured as a voltage controlled oscillator VCO. This can be achieved by connecting an external voltage V_{cont} to voltage control pin (5). The external voltage changes the threshold value of the lower and upper comparator from $1/3V_{CC}$ to $1/2V_{cont}$ and $2/3V_{CC}$ to V_{cont}. An increase of V_{cont} value increases the charging and discharging time and causes the frequency of the output waveform to decrease and vice versa. The period T of voltage controlled oscillator VCO can be obtained from the universal charging/discharging equation $V_{Cext} = V_f + (V_i - V_f)\exp(-t/\tau)$.

During the charging phase, $V_i = \frac{V_{cont}}{2}$ and $V_f = V_{CC}$, thus, the charging equation shall be $V_{Cext} = V_{CC} + \left(\frac{V_{cont}}{2} - V_{CC}\right)e^{-t/(R_1+R_2)C_{ext}}$. Setting $V_{Cext} = V_{cont}$, the charging time t_H shall be

$$t_H = (R_1 + R_2)C_{ext}\ln\left(\frac{V_{CC} - V_{cont}/2}{V_{CC} - V_{cont}}\right) \qquad (16.29)$$

During the discharging phase, $V_i = V_{Cont}$ and $V_f = 0$, thus, the charging equation shall be $V_{Cext} = V_{cont}e^{-t/R_2C_{ext}}$. Setting $V_{Cext} = V_{cont}/2$, the discharging time t_L shall be

$$t_L = R_2C_{ext}\ln 2 \qquad (16.30)$$

The period T of the VCO is

$$T = (R_1 + R_2)C_{ext}\ln\left(\frac{V_{CC} - V_{cont}/2}{V_{CC} - V_{cont}}\right) + R_2C_{ext}\ln 2 \qquad (16.31)$$

The voltage controlled ascillator can be implemented using a integrated circuit NE/SE 566 chip. This intergated circuit VCO produces two simultaneously square wave and triangular wave at the frequency upto 1.0 MHz. The basic block diagram of the VCO is shown in Fig. 16.25.

VCO can also be considered as voltage-to-frequency converter since the input voltage determines the frequency of the output. NE/SE 566 VCO has a current source that consists of the lower current source and upper current source. The schmitt trigger has two threshold voltages namely the lower threshold voltage V_L and upper threshold voltage V_H. The upper current source is used to charge the external capacitor from V_L voltage to V_H voltage. The lower current source is used to discharge the external capacitor from V_H voltage to V_L voltage. Beside providing square wave, the output of the schmitt trigger is also used to control the switching of lower and upper current source. The triangular wave is provided by the charging and discharging of the external capacitor via pin (7). The modulated input V_{CN} is consist of a dc voltage and an ac modulated signal V_{cn} coupled with capacitor.

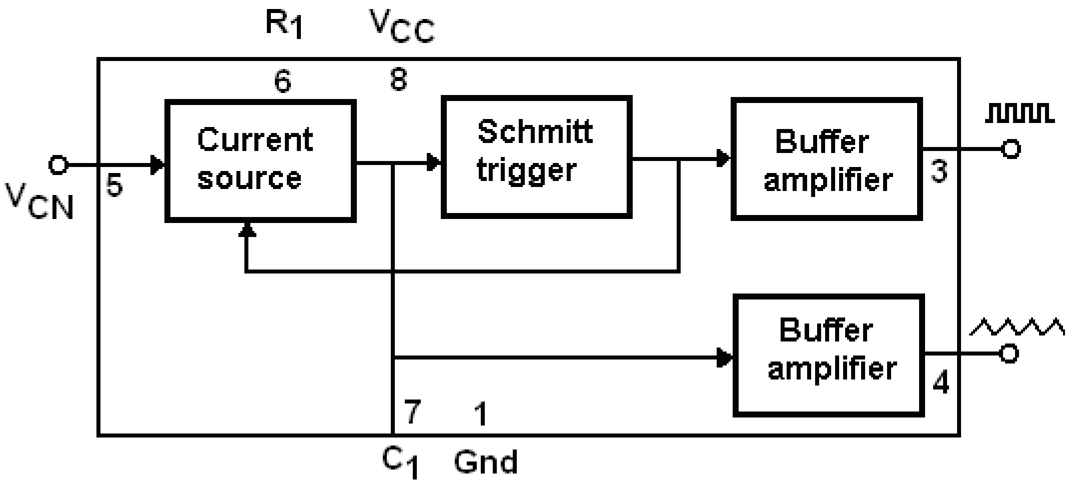


Figure 16.25: Block diagram of NE/SE 566 voltage controlled oscillator

Figure 16.26 shows the circuit connection of a NE/SE 566 VCO. The dc V_{CN} voltage is equal to $\frac{R_3}{R_2 + R_3} \cdot V_{CC}$, in which it must satisfy condition that $\frac{3}{4}V_{CC} \le V_{CN} \le V_{CC}$.

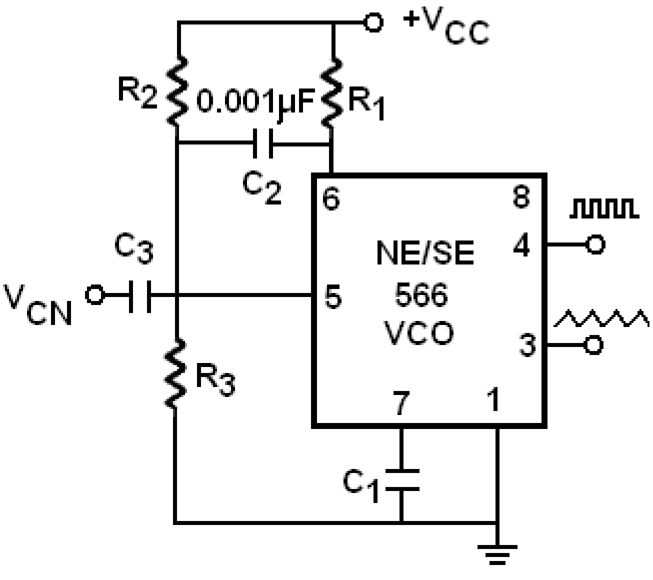


Figure 16.26: Circuit connection of NE/SE 566 voltage controlled oscillator

The modulated input signal V_{cn} must be < $3V_{PP}$, where V_{PP} is the peak-to-peak voltage of the output wave.

The time taken to charge the capacitor C_1 from voltage V_L to V_H is $t_1 = \frac{C_1}{I_Q}(V_H - V_L)$ and time taken to discharge capacitor C_1 from V_H to V_L is $t_2 = \frac{C_1}{I_Q}(V_H - V_L)$, where I_Q is the current source with assumption that both the lower and upper current source are sourcing same current. The period T of the oscillator is equal to the sum of the charging and discharging, which is

$$T = \frac{2C_1}{I_Q} \cdot (V_H - V_L) \tag{16.32}$$

Source current I_Q can be proved to be equal to $I_Q = \frac{V_{CC} - V_{CN}}{R_1}$. Thus, subtituting this equation into equation (16.32) yields the period of the oscillator as equal to

$$T = \frac{2C_1R_1}{(V_{CC} - V_{CN})} \cdot (V_H - V_L) \tag{16.33}$$

Usually the value of R_1 is in the range between 2 kΩ and 20 kΩ and a capacitor C_2 of 0.001 μF is connected to prevent internal oscillation. The value of $(V_H - V_L)$ can be assumed to be $V_{CC}/4$.

One can replace V_{CN} with an ac V_{cn} superimposed on dc V_{CN} to get frequency-modulated output with dc V_{CN} as the control for center frequency and dc $V_{CN} \pm V_{cn}$ as the control for lower and upper band-frequencies.

16.7 Function Generator

A function generator is capable of producing sine wave, square wave, triangular wave. One of popular such device is XR-2206 from EXAR. Beside generating sine wave and square wave, this device can be configured to produce pulse, ramp function, and frequency shift keying function FSK. Student is encourged ro obtain the data sheet of this device to study its capability of function generation.

16.8 Phase-Lock Loop

Let's consider a phase-lock loop, which is often using in the communication circuit as the input frequency locker. There are many other applications of the phase-lock loop PLL such as used in FM stereo decoders, tracking filter FSK decoder, and frequency-synthesized transmitters and receivers.

Phase-lock loop consists of a phase detector, a low-pass filter, and voltage controlled oscillator VCO used to feedback to the phase detector. The block diagram of the phase-lock loop is shown in Fig. 16.27.

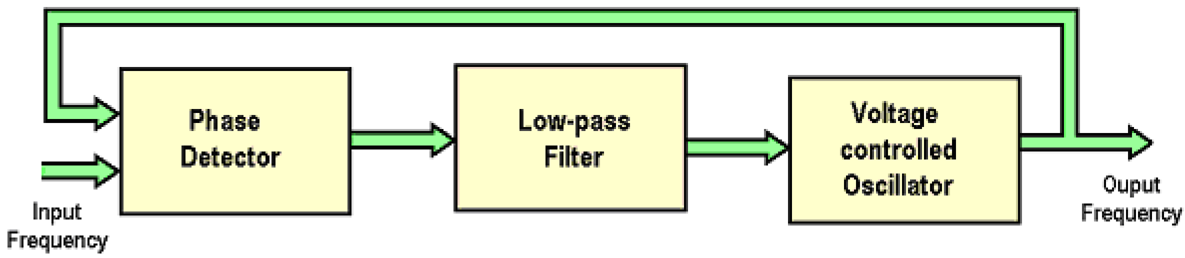


Figure 16.27: Block diagram of a phase-lock loop PLL

When the input frequency is not same as the frequency of voltage controlled oscillator VCO, there is a phase difference detected by the phase detector. The ouput of phase detector and the low-pass filter are proportional to the phase difference of the two signals. This proportional voltage is fed to VCO forcing its frequency moves toward the frequency of incoming signal.

The phase detector circuit in PLL is basically a linear multiplier. If the input signal V_i is $V_i = V_{pi}\sin(2\pi f_i t + \theta_i)$ and output signal of VCO V_o is $V_o = V_{pi}\sin(2\pi f_o t + \theta_o)$, the multipled voltage $V_d = V_{pi}\sin(2\pi f_i t + \theta_i) x V_{pi}\sin(2\pi f_o t + \theta_o)$ of linear multiplier shall be

$$V_d = \begin{array}{l} \dfrac{V_{pi}V_{po}}{2}\cos[(2\pi f_i t+\theta_i)-(2\pi f_o t+\theta_o)] \\ -\dfrac{V_{pi}V_{po}}{2}\cos[(2\pi f_i t+\theta_i)+(2\pi f_o t+\theta_o)] \end{array} \quad (16.34)$$

When the PLL is locked i.e. $f_i = f_o$, equation (16.34) becomes

$$V_d = \frac{V_{pi}V_{po}}{2}\cos(\theta_i-\theta_o)-\frac{V_{pi}V_{po}}{2}\cos(4\pi f_i t+\theta_i+\theta_o) \quad (16.35)$$

The second term of equation (16.35) contains the second harmonic signal, which shall be filtered out by the low-pass filter. Thus, the control voltage V_C to VCO is

$$V_C = \frac{V_{pi}V_{po}}{2}\cos(\theta_i - \theta_o) \qquad (16.36)$$

(θ_i -θ_o) is called *phase error*. Thus, the output of filter is proportional to the phase error.

The phase-lock loop has three modes of operation. When there is no input signal to the phase detector, the voltage controlled oscillator VCO is running at a fixed frequency called *center frequency* f_o corresponds to zero input voltage that has characteristic $f_o(t) = f_o + K_oV_C(t)$, where K_o is the sensitivity of voltage control oscillator VCO. The phase-lock loop is said to be in *free running mode*.

When there is input signal, the phase-lock loop goes into *capture* mode. The frequency of the voltage controlled oscillator changes continuously to match the input frequency. Once the input frequency is equal to the output frequency of voltage controlled oscillator, the phase-lock loop is said to be *phase-lock mode*.

The time taken for a phase-lock loop PLL to capture the incoming signal is called *capture time* or *pull-in-time*. This time depends on the initial frequency and phase difference (θ_i - θ_o) between V_i and V_o as well as the filter and other loop characteristics.

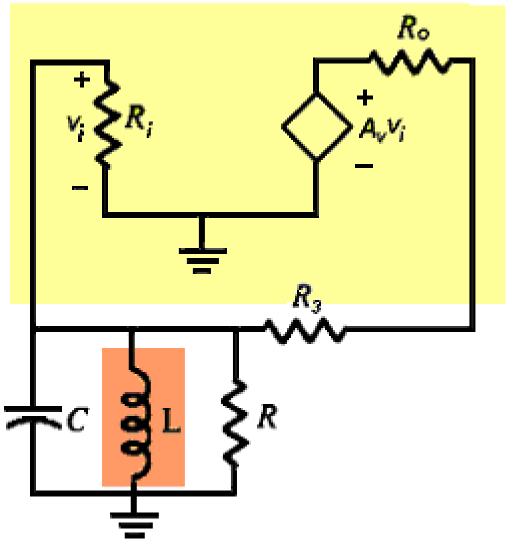
The most commonly used integrated circuit PLL is NE/SE 565. The center frequency f_o of the this chip is $f_o = \frac{1.2}{4R_1C_1}$. R_1 and C_1 are external resistor and capacitor connected to pin (8) and pin (9) of the chip respectively.

The NE/SE 565 PLL typically can track and lock input frequency over bandwidth ±60% of the center frequency f_o. The lock range of the chip f_L is $f_L = \frac{8f_o}{V_{CC} - V_{EE}}$. The capture range of the chip is $f_c = \frac{f_L}{2\pi x3.6x10^3C_2}$, where C_2 is the capacitor connected between V_{CC} and modulated output pin (7). Typically the value of C_2 is 10μF.

Exercises

16.1. State the principles for designing an oscillator and draw a block diagram to illustrate the principles.

16.2. The amplifier shown in the figure has voltage gain A_V = 50, input resistance R_i = 10kΩ and output resistance R_o = 200Ω. Find the resonant frequency f_r and the values of R and R_3 that will sustain the oscillation.



16.3. Describe how a Wien Bridge oscillator can get self-start and draw a diagram to illustrate how it gets self-start?

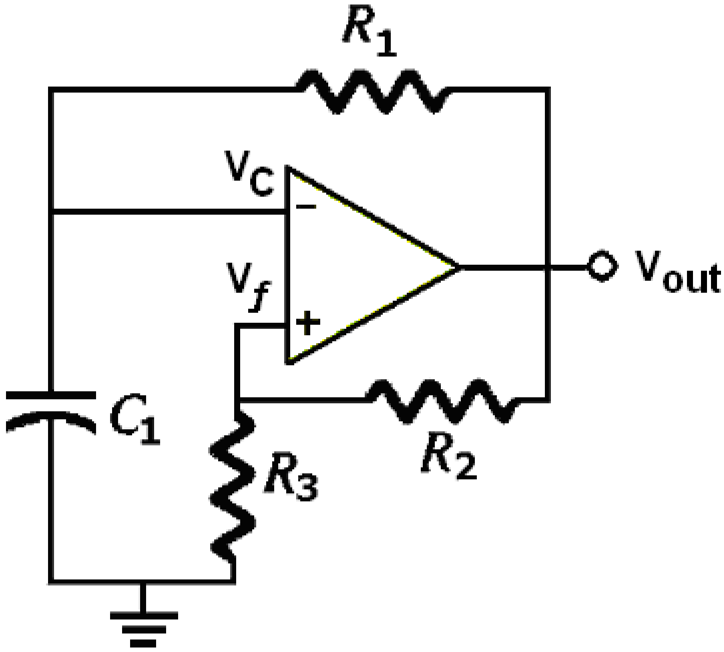
16.4. For the 555 timer circuit, R_1 = 1.0kΩ, R_2 = 2.0kΩ, and C_{ext} = 0.1μF.

i. Calculate the cycle duty of this timer.
ii. Show that the period T of this oscillator circuit is equal to $\frac{(R_1 + 2R_2)C_{ext}}{1.44}$.
iii. Why the duty cycle of this oscillator is more than 50%?
iv. How to achieve duty cycle of less than 50%?

16.5. The given circuit is a square wave generator. The saturation voltages of the operational amplifier are ±10.5 volts, R_1 = 10kΩ and C_1 = 1μF, and V_f is set to be 1/3 of V_{out}. Assume the charging and discharging of the capacitor is linear.

i. Prove that the period T of square wave is $3R_1C_1$.

ii. Find its amplitude.
iii. Find its peak-to-peak voltage.



16.6. Describe how a 555 timer can be configured as a voltage control oscillator.

16.7. Describe how a 555 timer monostable multivibrator works.

16.8. A PLL is locked onto an incoming signal with frequency of 1.0MHz at phase angle of 50^0. The VCO signal is at phase angle of 20^0. The peak amplitude of the incoming signal is 0.5V and that of VCO output signal is 0.7V.

i. What is the VCO frequency?
ii. What is the value of the control voltage being fed back to the VCO at this point?

Bibliography

1. Jacob Millman and Arvin Grabel, "***Microelectronics***", second edition, McGraw-Hill International Editions, 1987.
2. Muhammad H. Rashid, "***Microelectronic Circuits: Analysis and Design***", PWS Publishing Company, 1999.
3. Robert T. Paynter, "***Electronic Devices and Circuits***", fifth edition, McGraw-Hill, 1997.
4. Adel S. Sedra and Kenneth C. Smith, "***Microelectronic Circuits***", fourth edition, Oxford University Press, 1998.