$$f_o = \frac{1}{2\pi\sqrt{LC}} \quad ...(20.12)$$

20.11. COLPITT'S OSCILLATOR

The Colpitt's oscillator circuit is a superb circuit and is widely used in commercial signal generators up to 100 MHz. The basic circuit of a transistor Colpitt's Oscillator is shown in Fig. 20.12 It basically consists of a single stage inverting amplifier and an L-C phase shift network, as obvious from the circuit diagram shown. The two series capacitors C_1 and C_2 form the potential divider used for providing the feedback voltage—the voltage developed across capacitor C_2 provides the regenerative feedback required for sustained oscillations. Parallel combination of R_E and C_E along with resistors R_1 and R_2 provides the stabilized self bias. The collector supply voltage V_{CC} is applied to the collector through a radio-frequency choke (RFC) which

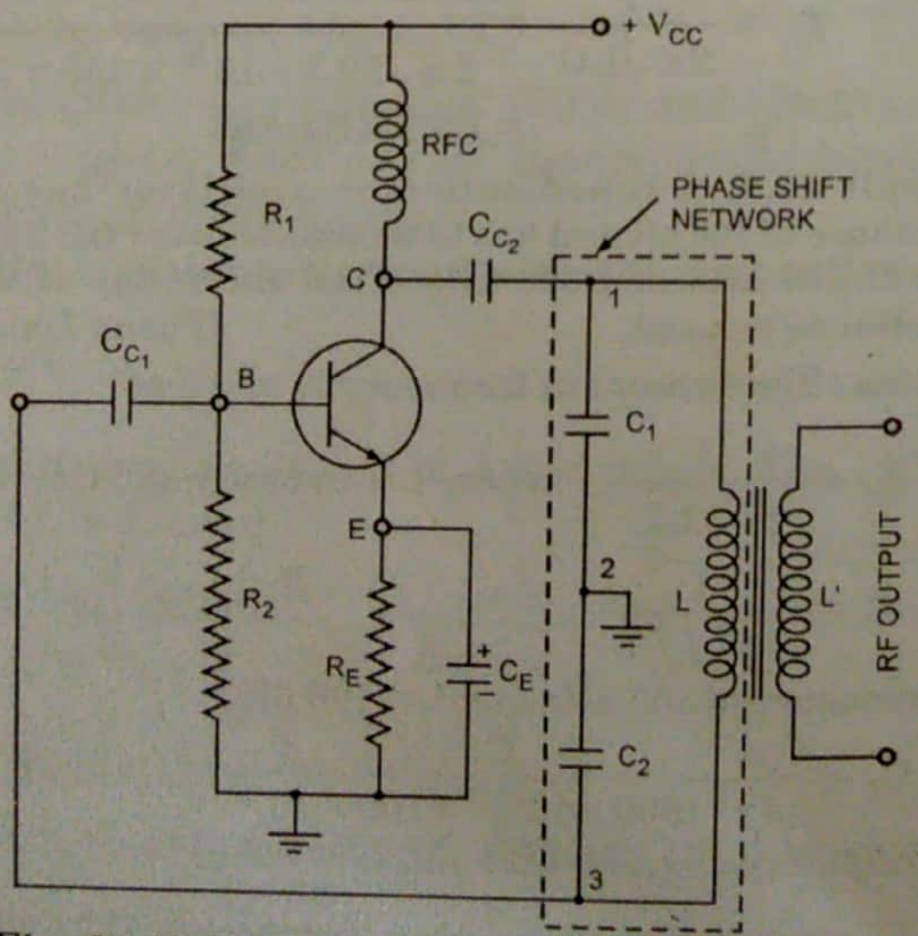


Fig. 20.12. ***Basic Circuit For Transistor Colpitt's Oscillator***

permits an easy flow of direct current but at the same time it offers very high impedance to the high frequency currents. The presence of coupling capacitor C_{C_2} in the output circuit does not permit the dc currents to go to the tank circuit (the flow of dc current in a tank circuit reduces its Q). The radio-frequency energy developed across RFC is capacitively coupled to the tank circuit through the capacitor C_C. The output of the phase-shift L-C network is coupled from the junction of L and C_2 to the amplifier input at base through coupling capacitor C_{C_1} which blocks dc but provides path to ac. Transistor itself produces a phase shift of 180° and another phase shift of 180° is provided by the capacitive feedback. Thus a total phase shift of 360° is obtained which is an essential condition for developing oscillations. The output voltage is derived from a secondary winding L' coupled to the inductance L. The frequency is determined by the tank circuit and is varied by gang-tuning the two capacitors C_1 and C_2. It is to be noted that capacitors C_1 and C_2 are ganged. As the tuning is varied, values of both capacitors vary simultaneously, the ratio of the two capacitances remaining the same.

Working. When the collector supply voltage V_{CC} is switched on, the capacitors C_1 and C_2 are charged. These capacitors C_1 and C_2 discharge through the coil L, setting up oscillations of frequency $f = \frac{1}{2\pi}\sqrt{\frac{1}{LC_1} + \frac{1}{LC_2}}$.

The oscillations across capacitor C_2 are applied to the base-emitter junction and appear in the amplified form in the collector circuit. Of course, the amplified output in the collector circuit is of the same frequency as that of the oscillatory circuit. This amplified output in the collector circuit is supplied to the tank circuit in order to meet the losses. Thus the tank circuit is getting continuously energy from the circuit to make up for the losses occurring in it and, therefore, ensures undamped oscillations. The energy supplied to the tank circuit is of correct phase, as already explained, and if Aβ exceeds unity, oscillations are sustained in the circuit.

Frequency of Oscillation. First of all let us develop a general theory for Colpitt's and Hartley oscillators. The equivalent circuits (Fig. 20.13) is drawn with the following two assumptions :

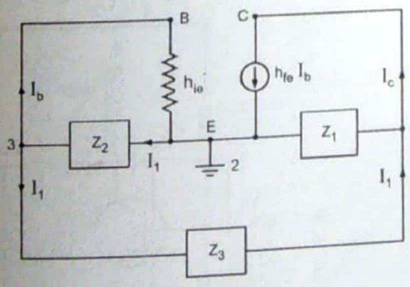


Fig. 20.13

(i) h_{re} of transistor is negligibly small and, therefore, the feedback source $h_{re}\, V_{out}$ is negligible.

(ii) h_{oe} of the transistor is very small *i.e.* the output resistance $\frac{1}{h_{oe}}$ is very large and, therefore, $\frac{1}{h_{oe}}$ is omitted from the equivalent circuit.

Let us determine the load impedance between output terminals 1 and 2. Here Z_2 and h_{ie} are in parallel and their resultant impedance is in series with impedance Z_3.

The equivalent impedance is in parallel with impedance Z_1.

Thus load impedance between output terminals is given as

$$Z_L = Z_1 \,||\, [Z_3 + (Z_2 \,||\, h_{ie})]$$
$$= Z_1 \,||\left[Z_3 + \frac{Z_2\, h_{ie}}{Z_2 + h_{ie}}\right]$$
$$= Z_1 \,||\left[\frac{Z_3\,(Z_2 + h_{ie}) + Z_2\, h_{ie}}{Z_2 + h_{ie}}\right]$$
$$= Z_1 \,||\left[\frac{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3}{Z_2 + h_{ie}}\right]$$

or
$$\frac{1}{Z_L} = \frac{1}{Z_1} + \frac{Z_2 + h_{ie}}{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3}$$
$$= \frac{h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2 + Z_2\, Z_3}{Z_1\,[h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3]}$$

or
$$Z_L = \frac{Z_1\,[h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3]}{h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2 + Z_2\, Z_3} \qquad \text{...(20.13)}$$

The voltage gain of a CE amplifier without feedback is given as

$$A = \frac{-h_{fe}\, Z_L}{h_{ie}} \qquad \text{...(20.14)}$$

The output voltage between terminals 1 and 2 is given as

$$V_{out} = \left[Z_3 + \frac{Z_2\, h_{ie}}{Z_2 + h_{ie}}\right] I_1 = \left[\frac{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3}{Z_2 + h_{ie}}\right] I_1$$

The voltage feedback to the input terminals 2 and 3 is given as

$$V_f = \frac{Z_2\, h_{ie}}{Z_2 + h_{ie}}\, I_1$$

So feedback fraction,

$$\beta = \frac{V_f}{V_{out}} = \frac{Z_2\, h_{ie}}{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3} \qquad \text{...(20.15)}$$

Applying the criterion of oscillation *i.e.* A β = 1, we have

$$\frac{-h_{fe}\, Z_L}{h_{ie}} \cdot \frac{Z_2\, h_{ie}}{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3} = 1$$

or
$$\frac{h_{fe}\, Z_1\,[h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3]}{h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2 + Z_2\, Z_3} \cdot \left[\frac{Z_2}{h_{ie}\,(Z_2 + Z_3) + Z_2\, Z_3}\right] = -1$$

or
$$\frac{h_{fe}\, Z_1\, Z_2}{h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2 + Z_2\, Z_3} = -1$$

or $h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2 + Z_2\, Z_3 = -h_{fe}\, Z_1\, Z_2$

or $h_{ie}\,(Z_1 + Z_2 + Z_3) + Z_1\, Z_2\,(1 + h_{fe}) + Z_2\, Z_3 = 0$ ...(20.16)

This is the general equation for the oscillator.

In case of the Colpitt's oscillator

$$Z_1 = \frac{1}{j\,\omega\, C_1} = -\frac{j}{\omega\, C_1};\; Z_2 = \frac{1}{j\,\omega\, C_2} = \frac{-j}{\omega\, C_2} \text{ and } Z_3 = j\,\omega\, L$$

Substituting these values in Eq. (20.16) we have

$$h_{ie}\left[\frac{-j}{\omega C_1} - \frac{j}{\omega C_2} + j\omega L\right] + \left[\frac{-j}{\omega C_1} \cdot \frac{-j}{\omega C_2}\right](1 + h_{fe}) + \left[\frac{-j}{\omega C_2}\right] j\omega L = 0$$

or
$$-j\, h_{ie}\left[\frac{1}{\omega\, C_1} + \frac{1}{\omega\, C_2} - \omega\, L\right] - \frac{1 + h_{fe}}{\omega^2\, C_1\, C_2} + \frac{L}{C_2} = 0 \qquad \text{...(20.17)}$$

Equating the imaginary component of the above Eq. (20.17) to zero we have

$$h_{ie}\left[\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right] = 0$$

$$\text{or} \quad \frac{1}{\omega C_1}+\frac{1}{\omega C_2} = \omega L \qquad \because h_{ie} \neq 0$$

$$\text{or} \quad \frac{C_1+C_2}{\omega C_1 C_2} = \omega L$$

$$\text{or} \quad \omega^2 = \frac{C_1+C_2}{L C_1 C_2}$$

$$\text{or} \quad \omega = \sqrt{\frac{C_1+C_2}{L C_1 C_2}} = \sqrt{\frac{1}{L C_1}+\frac{1}{L C_2}}$$

$$\text{or} \quad f = \frac{1}{2\pi}\sqrt{\frac{1}{L C_1}+\frac{1}{L C_2}} \qquad \text{...(20.18)}$$

The above Eq. (20.18) gives the frequency of oscillation. Equating the real component of Eq. (20.19) to zero we have

$$\frac{1+h_{fe}}{\omega^2 C_1 C_2} = \frac{L}{C_2}$$

$$\text{or} \quad 1+h_{fe} = \omega^2 L C_1$$

$$\text{or} \quad 1+h_{fe} = \frac{C_1+C_2}{L C_1 C_2}\times L C_1 = \frac{C_1+C_2}{C_2} = 1+\frac{C_1}{C_2}$$

$$\text{or} \quad h_{fe} = \frac{C_1}{C_2} \qquad \text{...(20.19)}$$

As for other oscillator circuits, the loop gain must be greater than unity to ensure that the circuit oscillates.

$$\text{So } A\beta \geq 1 \text{ or } A \geq \frac{1}{\beta} \geq \frac{C_2}{C_1} \qquad \text{...(20.20)}$$

FET Colpitt's Oscillator. A practical version of an

However, precaution is to be taken in selection of C_3. If capacitor C_3 is made too small, the L-C branch will not have a net inductive reactance and under such condition the circuit will refuse to oscillate.

In a Colpitt's oscillator, the resonant frequency is affected by the transistor and stray capacitances because the capacitors C_1 and C_2 are shunted by the transistor and stray capacitances and so their values are altered. But in a Clapp oscillator, the transistor and stray capacitances have no effect on capacitor C_3, so the oscillation frequency is more stable and accurate. This is the reason that Clapp oscillator is preferred over a Colpitt's oscillator.

High frequency stability can further be obtained by enclosing the entire circuit in a constant temperature chamber and by maintaining the supply voltage constant with the help of a zener diode.

20.13. HARTLEY OSCILLATOR

The transistor Hartley oscillator is as popular as Colpitt's oscillator and is widely used as a local oscillator in radio receivers. The circuit arrangement is shown in Fig. 20.16. Hartley oscillator circuit is similar to Colpitt's oscillator circuit, except that phase-shift network consists of two inductors L_1 and L_2 and a capacitor C instead of two capacitors and one inductor. The output of the amplifier is applied across inductor L_1 and the voltage across inductor L_2 forms the feedback voltage. The coil L_1 is inductively coupled to coil L_2, the combination functions as an auto-transformer. However, because of direct connection, the junction of L_1 and L_2 cannot be directly grounded. Instead, another capacitor C_L is used. The operation of the circuit is similar to that of the Colpitt's oscillator circuit.

Considering the fact that there exists mutual inductance between coils L_1 and L_2 because the coils are wound on the same core, their net effective inductance is increased by mutual inductance M. So in this case effective inductance is given by the equation

$$L = L_1 + L_2 + 2M$$

and resonant or oscillation frequency is given by the equation

$$f = \frac{1}{2\pi\sqrt{[C(L_1 + L_2 + 2M)]}}, \text{ as derived below :}$$

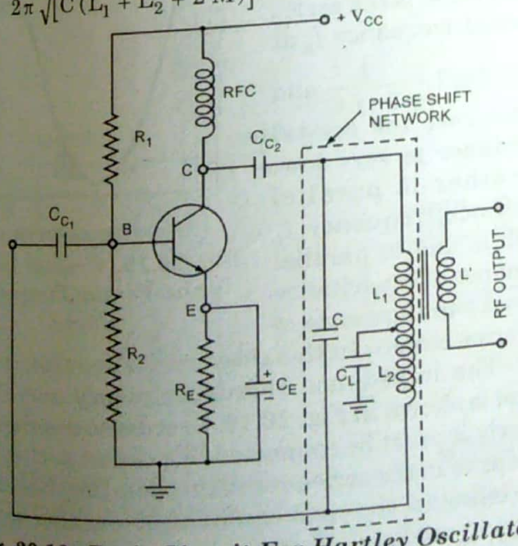


Fig. 20.16. ***Basic Circuit For Hartley Oscillator***

Frequency of Oscillation. The general Eq. (20.16) derived in Art. 20.11 is reproduced here

$$h_{ie}(Z_1 + Z_2 + Z_3) + Z_1 Z_2 (1 + h_{fe}) + Z_2 Z_3 = 0$$

Here $Z_1 = j\omega L_1 + j\omega M$; $Z_2 = j\omega L_2 + j\omega M$ and $Z_3 = \frac{1}{j\omega C} = -j/\omega C$

Substituting these values in general Eq. (20.16) we get

$$h_{ie}\left[(j\omega L_1 + j\omega M) + (j\omega L_2 + j\omega M) - \frac{j}{\omega C}\right]$$
$$+ (j\omega L_1 + j\omega M)(j\omega L_2 + j\omega M)(1 + h_{fe}) + (j\omega L_2 + j\omega M)\left(\frac{-j}{\omega C}\right) = 0$$

$$\text{or} \quad j\omega h_{ie}\left[L_1 + L_2 + 2M - \frac{1}{\omega^2 C}\right]$$
$$-\omega^2 (L_2 + M)\left[(L_1 + M)(1 + h_{fe}) - \frac{1}{\omega^2 C}\right] = 0 \quad \text{....(20.24)}$$

Equating the imaginary part of above Eq. (20.24) to zero we get

$$\omega h_{ie}\left[L_1 + L_2 + 2M - \frac{1}{\omega^2 C}\right] = 0$$

$$\text{or} \quad L_1 + L_2 + 2M - \frac{1}{\omega^2 C} = 0 \qquad \because \omega h_{ie} \neq 0$$

$$\text{or} \quad \omega^2 C = \frac{1}{L_1 + L_2 + 2M}$$

$$\text{or} \quad f = \frac{\omega}{2\pi} = \frac{1}{2\pi\sqrt{C(L_1 + L_2 + 2M)}} \quad \text{...(20.25)}$$

The above Eq. (20.25) gives the frequency of oscillation.

Equating the real component of Eq. (20.24) to zero we get

$$(L_1 + M)(1 + h_{fe}) - \frac{1}{\omega^2 C} = 0$$

$$\text{or} \quad (1 + h_{fe}) = \frac{1}{\omega^2 C (L_1 + M)} = \frac{L_1 + L_2 + 2M}{L_1 + M} = 1 + \frac{L_2 + M}{L_1 + M}$$

$$\text{or} \quad h_{fe} = \frac{L_2 + M}{L_1 + M} \quad \text{...(20.26)}$$

As for other oscillator circuits, the loop gain must be greater than 1 to ensure that circuit oscillates.

$$\text{So } A\beta \geq 1$$

$$\text{or} \quad A \geq \frac{1}{\beta} \geq \frac{L_1 + M}{L_2 + M} \quad \text{....(20.27)}$$

FET Hartley Oscillator. An FET Hartley oscillator circuit is shown in Fig. 20.17. This circuit operates at a frequency given by Eq. (20.25).

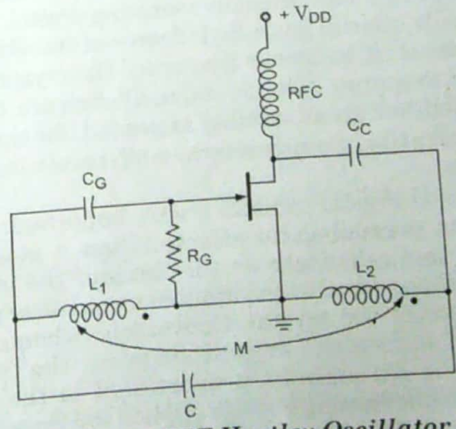


Fig. 20.17. ***FET Hartley Oscillator***

4. It needs high voltage (12 V) battery so as to develop sufficiently large feedback voltage.

Applications. FET phase-shift oscillator is used for generating signals over a wide frequency range. The frequency may be varied from a few Hz to 200 Hz by employing one set of resistors with three capacitors ganged together to vary over a capacitance range in the 1 : 10 ratio. Similarly the frequency ranges of 200 Hz to 2 kHz, 2 kHz to 20 kHz and 20 kHz to 200 kHz can be obtained by using other sets of resistors.

Transistor Phase Shift Oscillator. The circuit arrangement of a phase-shift oscillator using N-P-N transistor in CE configuration is shown in Fig. 20.25. As usual, the voltage divider $R_1 - R_2$ provides dc emitter base bias, R_E and C_E combination provides temperature stability and prevent ac signal degeneration and collector resistor R_C controls the collector voltage. The oscillator output voltage is capacitively coupled to the load by C_C.

In case of a transistor phase shift oscillator, the output of the feedback network is loaded appreciably by the relatively small input resistance (h_{ie}) of the transistor. Hence, instead of employing voltage-series feedback (as used in case of FET phase shift oscillator), voltage-shunt feedback is used for a transistor phase shift oscillator, as shown in Fig. 20.25. In this circuit, the feedback signal is coupled through the feedback resistor R' in series with the amplifier stage input resistance h_{ie}. The value of R' should be such that when added with amplifier stage input resistance h_{ie}, it is equal to R *i.e.*, $R' + h_{ie} = R$.

Operation. The circuit is set into oscillations by any random or variation caused in the base current, that may be either due to noise inherent in the transistor or minor variation in voltage of dc power supply. This variation in base current is amplified in collector circuit. The output of the amplifier is supplied to an R-C feedback network. The R-C network produces a phase shift of 180° between output and input voltages. Since CE amplifier produces a phase reversal of the input signal, total phase shift becomes 360° or 0° which is essential for *regeneration* or for sustained oscillations. The output of this network is thus in the same phase as the originally assumed input to the amplifier and is applied to the base terminal of the transistor. Thus sustained variation in collector current between saturation and cutoff values are obtained. R-C phase shift network is the frequency determining network, as already explained in case of FET phase-shift oscillator.

Frequency of Oscillation. The equivalent circuit for the analysis of a transistor phase shift oscillator (circuit shown in Fig. 20.25) is shown in Fig. 20.26. The equivalent circuit shown in Fig. 20.26 is simplified if the following assumptions are made.

(*i*) h_{re} of the transistor is negligibly small and, therefore, $h_{re} V_{out}$ is omitted from the circuit.

(*ii*) h_{oe} of the transistor is very small *i.e.* $\frac{1}{h_{oe}}$ is much larger than R_C. Thus the effect of h_{oe} can be neglected.

Making above assumptions and replacing current source by equivalent voltage source, the simplified equivalent circuit is shown in Fig. 20.27.

Applying Kirchhoff's voltage law to the three loops shown in Fig. 20.27 we have

$$\left(R + R_C + \frac{1}{j\omega C}\right) I_1 - R I_2 + h_{fe} I_b R_C = 0 \qquad \text{....(20.31)}$$

$$-R I_1 + \left(2R + \frac{1}{j\omega C}\right) I_2 - R I_b = 0 \qquad \text{...(20.32)}$$

$$0 - R I_2 + \left(2R + \frac{1}{j\omega C}\right) I_b = 0 \qquad \text{...(20.33)}$$

As the currents I_1, I_2 and I_b are non-vanishing, the determinant of the coefficients of I_1, I_2 and I_b must be zero. Substituting $\frac{1}{\omega C} = X_C$ we have

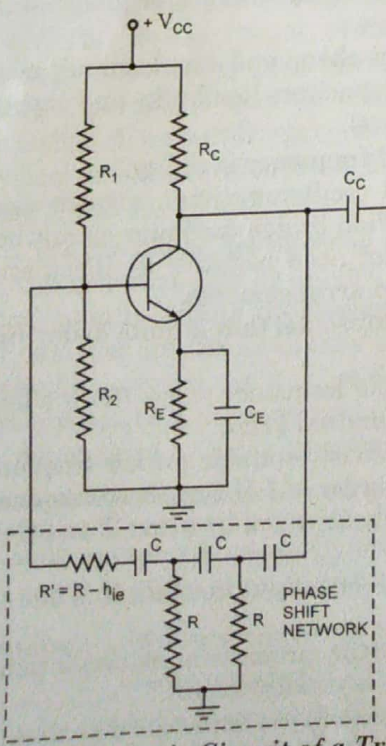


Fig. 20.25. *Basic Circuit of a Transistor Phase Shift Oscillator*

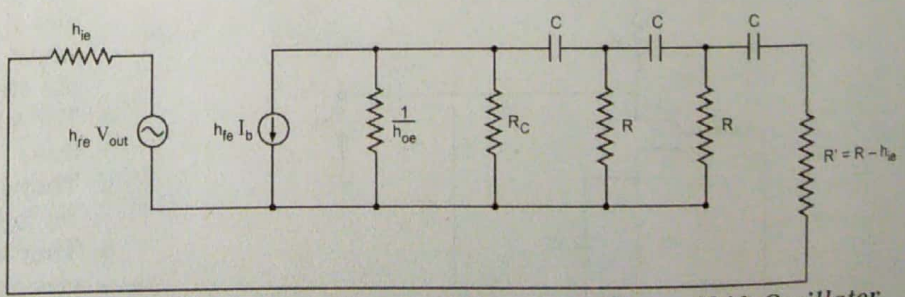


Fig. 20.26. *Equivalent Circuit of Transistor Phase-Shift Oscillator Circuit Shown in Fig. 20.25*

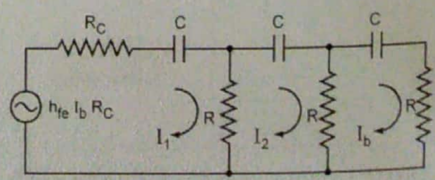


Fig. 20.27

$$\begin{vmatrix} (R + R_C - j X_C) & (-R) & (h_{fe} R_C) \\ (-R) & (2R - j X_C) & (-R) \\ 0 & (-R) & (2R - j X_C) \end{vmatrix} = 0 \quad ...(20.34)$$

$$\text{or } (R + R_C - j X_C)[(2R - j X_C)^2 - R^2] + R[(-R)(2R - j X_C) - h_{fe} R_C (-R)] = 0$$

$$\text{or } (R + R_C - j X_C)(3R^2 - X_C^2 - j 4 R X_C) - R[2R^2 - j R X_C - h_{fe} R_C R] = 0$$

$$\text{or } R^3 + R^2 R_C (3 + h_{fe}) - 5 R X_C^2 - R_C X_C^2 - 6 j R^2 X_C - j 4 R R_C X_C + j X_C^3 = 0 \quad ...(20.35)$$

Equating the imaginary component of the above equation to zero we have

$$6 R^2 X_C + 4 R R_C X_C - X_C^3 = 0$$

$$\text{or } X_C = \sqrt{6 R^2 + 4 R R_C}$$

$$\text{or } 2\pi f C = \frac{1}{\sqrt{6 R^2 + 4 R R_C}}$$

$$\because X_C = \frac{1}{\omega C} = \frac{1}{2\pi f C}$$

$$\text{or } f = \frac{1}{2\pi RC\sqrt{6 + \frac{4 R_C}{R}}} \quad ...(20.36)$$

If $R = R_C$, then

$$f = \frac{1}{2\pi RC\sqrt{10}} \quad ...(20.37)$$

The above equation gives frequency of oscillation. Equating the real component of Eq. (20.35) to zero we have

$$R^3 + R^2 R_C (3 + h_{fe}) - X_C^2 (5R + R_C) = 0$$

$$\text{or } R^3 + R^2 R_C (3 + h_{fe}) - (6 R^2 + 4 R R_C)(5R + R_C) = 0$$

$$\text{or } -29 R^3 - 23 R^2 R_C + h_{fe} R^2 R_C - 4 R R_C^2 = 0$$

$$\text{or } \frac{-29 R}{R_C} - 23 + h_{fe} - 4\frac{R_C}{R} = 0$$

$$\text{or } h_{fe} = 23 + 29\frac{R}{R_C} + 4\frac{R_C}{R} \quad ...(20.38)$$

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

$$h_{fe} > 23 + 29\frac{R}{R_C} + 4\frac{R_C}{R} \quad ...(20.39)$$

If $R = R_C$, then $\quad h_{fe} > (23 + 29 + 4)$ *i.e.* $> 56 \quad ...(20.40)$

Applications. The phase shift oscillator is well suited to the range of frequencies from several hertz to several hundred kilohertz (20 Hz to 200 kHz), and so includes the audio frequency range (up to 20 kHz). For generating different audio-frequencies, variable air capacitors are employed as circuit elements in the phase-shift network. It is possible to vary the frequency in the range of about 1 : 10 because the range of capacitors can be varied in the ratio of 10 : 1 (typically from 40 pF to 450 pF). For variations of frequency over a large range, the three capacitors are usually ganged so as to vary the capacitance of the three capacitors simultaneously. Such a variation keeps the input impedance to the phase-shift network constant and also keeps constant the magnitudes of β and Aβ. Thus the amplitude of oscillations will remain unaffected as the frequency is adjusted. The phase-shift oscillator is operated in class A so as to keep distortion to the minimum. Frequency range from 20 Hz to 200 Hz, 200 Hz to 2 kHz, 2 kHz to 20 kHz and 20 kHz to 200 kHz can be obtained by using different set of resistors.

Phase-shift oscillators are not suitable for higher frequency operation because at higher frequency, the internal phase shift of the transistor and reduction in h_{fe} cause difficulties in designing the circuit. The frequency of the oscillator cannot be changed easily.

Example 20.9. In a phase shift oscillator shown in Fig. 20.24, $R_1 = R_2 = R_3 = 800$ kΩ and $C_1 = C_2 = C_3 = 100$ pF. Determine the frequency of oscillation.

Solution : $R_1 = R_2 = R_3 = R = 800\text{ k}\Omega = 8 \times 10^5\ \Omega$

$C_1 = C_2 = C_3 = C = 100\text{ pF} = 10^{-10}$ F

Frequency of oscillation,

$$f_o = \frac{1}{2\pi RC\sqrt{6}} = \frac{1}{2\pi \times 8 \times 10^5 \times 10^{-10} \times \sqrt{6}}$$

$$= 812 \text{ Hz } \textbf{Ans.}$$

20.17. WIEN BRIDGE OSCILLATOR

It is one of the most popular type of oscillators used in audio and sub-audio frequency ranges (20 – 20 kHz). This type of oscillator is simple in design, compact in size, and remarkably stable in its frequency output. Furthermore, its output is relatively free from distortion and its frequency can be varied easily. However, the maximum frequency output of a typical Wien bridge oscillator is only about 1 MHz.

This is also, in fact, a phase-shift oscillator. It employs two transistors, each producing a phase shift of 180°, and thus producing a total phase-shift of 360° or 0°.

The circuit diagram of Wien bridge oscillator is shown in Fig. 20.28. It is essentially a two-stage amplifier with an R-C bridge circuit. R-C bridge circuit (Wien bridge) is a lead-lag network. The phase-shift across the network lags with increasing frequency and leads with decreasing frequency. By adding Wien bridge feedback network, the oscillator becomes sensitive to a signal of only one particular frequency. This particular frequency is that at which Wien bridge is balanced and for which the phase shift is 0°. If the Wien-bridge feedback network is not employed and output of transistor Q_2 is fedback to transistor Q_1 for providing regeneration required for producing oscillations, the transistor Q_1 will amplify signals over a wide range of frequencies and thus direct coupling would result in poor frequency stability. Thus by employing Wien-bridge feedback network frequency stability is increased.

In the bridge circuit R_1 in series with C_1, R_3, R_4 and R_2 in parallel with C_2 form the four arms.

From the analysis of the bridge circuit it is obvious that the bridge will be balanced only when

$$R_3\left[\frac{R_2}{1 + j\omega C_2 R_2}\right] = R_4\left(R_1 - \frac{j}{\omega C_1}\right)$$

$$\text{or } R_2 R_3 = R_4 (1 + j\omega C_2 R_2)(R_1 - j/\omega C_1)$$

$$\text{or } R_2 R_3 - R_4 R_1 - \frac{C_2}{C_1} R_2 R_4 + \frac{j R_4}{\omega C_1} - j \omega C_2 R_2 R_1 R_4 = 0$$

Separating real and imaginary terms we have

$$R_2 R_3 - R_4 R_1 - \frac{C_2}{C_1} R_2 R_4 = 0$$

$$\text{or } \frac{C_2}{C_1} = \frac{R_3}{R_4} - \frac{R_1}{R_2} \qquad ...(20.41)$$

$$\text{and } \frac{R_4}{\omega C_1} - \omega C_2 R_2 R_1 R_4 = 0$$

$$\text{or } \omega^2 = \frac{1}{C_1 C_2 R_1 R_2}$$

$$\text{or } \omega = \frac{1}{\sqrt{C_1 C_2 R_1 R_2}}$$

$$\text{and } f = \frac{1}{2\pi \sqrt{R_1 C_1 R_2 C_2}} \qquad ...(20.42)$$

if $C_1 = C_2 = C$ and $R_1 = R_2 = R$, then $f = \frac{1}{2\pi CR}$ ...(20.43)

and $R_3 = 2 R_4$ ...(20.44)

Thus we see that in a bridge circuit the output will be in phase with the input only when the bridge is balanced *i.e.*, at resonant frequency given by Eq. (20.43). At all other frequencies the bridge is off-balance *i.e.* the voltage fedback and output voltage do not have the correct phase relationship for sustained oscillations.

So this bridge circuit can be used as feedback network for an oscillator, provided that the phase shift through the amplifier is zero. This requisite condition is achieved by using a two stage amplifier, as illustrated in Fig. 20.28. In this arrangement the output of the second stage is supplied back to the feedback network and the voltage across the parallel combination C_2 R_2 is fed to the input of the first stage. Transistor Q_1 serves as an oscillator and amplifier whereas the transistor Q_2 as an invertor to cause a phase shift of 180°. The circuit uses positive and negative feedbacks. The positive feedback is through R_1, C_1, R_2, C_2 to transistor Q_1 and negative feedback is through the voltage divider to the input of transistor Q_1. Resistors R_3 and R_4 are used to stabilize the amplitude of the output.

The two transistors Q_1 and Q_2 thus cause a total phase shift of 360° and ensure proper positive feedback. The negative feedback is provided in the circuit to ensure constant output over a range of frequencies. This is achieved by taking resistor R_4 in the form of a temperature sensitive lamp, whose resistance increases with the increase in current. In case the amplitude of the output tends to increase, more current would provide more negative feedback. Thus the output would regain its original value. A reverse action would take place in case the output tends to fall.

The amplifier voltage gain,

$$A = \frac{R_3 + R_4}{R_4} = \frac{R_3}{R_4} + 1 = 3 \quad ...(20.45) \quad \because \text{ from Eq. (20.44) } R_3 = 2R_4$$

The above corresponds with the feedback network attentuation of 1/3. Thus, in this case, voltage gain A must be equal to or greater than 3, to sustain oscillations.

To have a voltage gain of 3 is not difficult. On the other hand, to have a gain as low as 3 may be difficult. For this reason also negative feedback is essential.

Operation. The circuit is set in oscillation by any random change in base current of transistor Q_1, that may be due to noise inherent in the transistor or variation in voltage of dc supply. This variation in base current is amplified in collector circuit of transistor Q_1 but with a phase-shift of 180°. The output of transistor Q_1 is fed to the base of second transistor Q_2 through capacitor C_4. Now a still further amplified and twice phase-reversed signal appears at the collector of the transistor Q_2. Having been inverted twice, the output signal will be in phase with the signal input to the base of transistor Q_1. A part of the output signal at transistor Q_2 is fedback to the input points of the bridge circuit (points A-C). A part of this feedback signal is applied to emitter resistor R_4 where it produces degenerative effect (or negative feedback). Similarly, a part of the feedback signal is applied across the base-bias resistor R_2 where it produces regenerative effect (or positive feedback). At the rated frequency, effect of regeneration is made slightly more than that of degeneration so as to obtain sustained oscillations.

The continuous frequency variation in this oscillator can be had by varying the two capacitors C_1 and C_2 simultaneously. These capacitors are variable air-gang capacitors. We can change the frequency range of the oscillator by switching into the circuit different values of resistors R_1 and R_2.

FET Wein bridge oscillator is shown in Fig. 20.29.

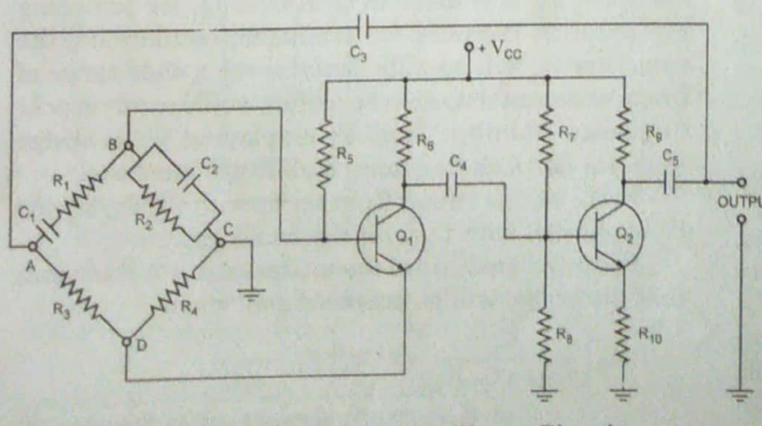


Fig. 20.28. *Wien Bridge Oscillator Circuit*

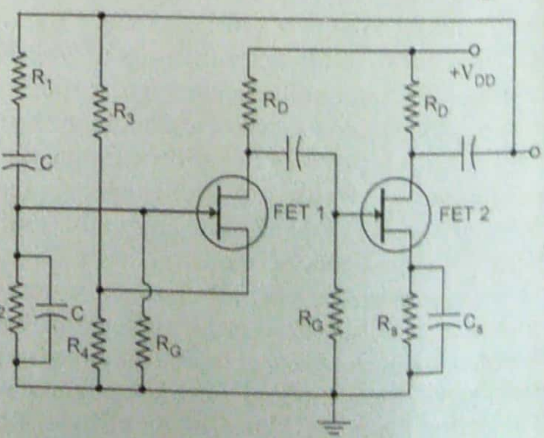
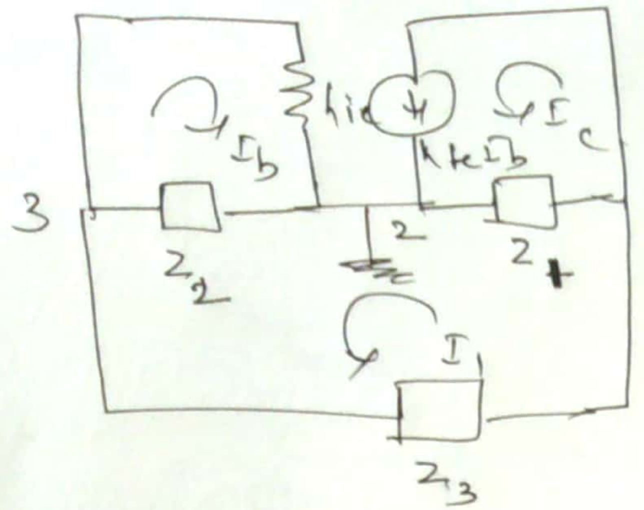
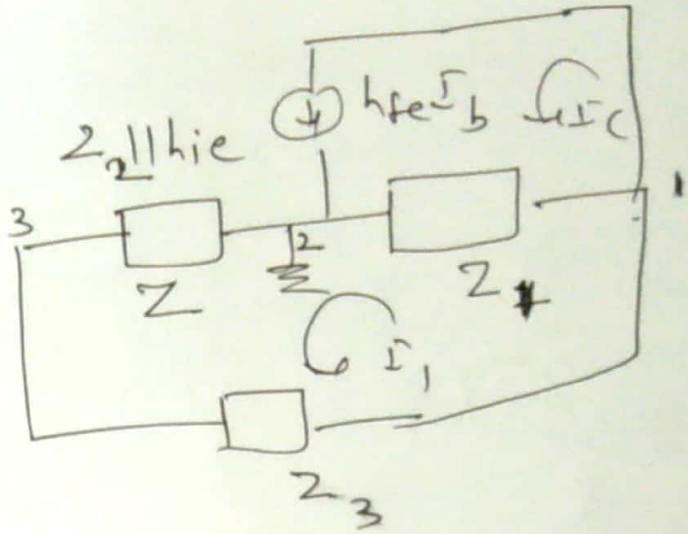


Fig. 20.29. *FET Wein Bridge Oscillator*



$\simeq$



V_{12} is voltage across Z_1

or V_{12} is also voltage across Z_3 & Z

Z_L is load betN terminal 1 & 2 ie O/P terminal

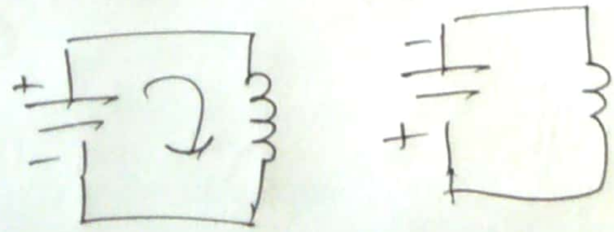
ie $(Z_3 + Z) \| Z_1$

Final equation has Real & imaginary part

from imaginary part we get freq. of osc

Real - Condition

tank circuit



goes on