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Analog

Transistors have 3 mode

1. Common Emitter
   Input → Base
   Output → Collector.
   Ground → Emitter.
2. Common Base
   Input → Emitter
   Output → Collector.
   Ground → Base.

3. Common collector
   Input → Base
   Output → Emitter
   Collector → Ground.

For CE configuration —

1. DC Source
2. AC Source.

Activity of DC source
i) It helps to determine the operation point.
ii) It helps to provide faithful amplication.

Features of Voltage Amplifier —

1. Input impedence is very high.
2. Output " " " low.
3. Voltage gain is high.

Features of Current Amplifier —

1. Input of Current Amplifier.
2. Output impedence is low.
3. Current gain is high.

@Orunayan Bhattacharya@

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Power Amplifier

(1) It able to handle large signal voltage.
(2) Rating is very high, rather voltage amplifier.

→ Reason or (3)rd point

→ Doping concentration is very high rather voltage amplifier.

If the signal 0 to 360° passes through a transistor its efficiency will decrease, because of no relaxation time for the transistor & the power consumption is very high.

Analog (~~Mukher~~ Banerjee).

Types of region in transistor :-

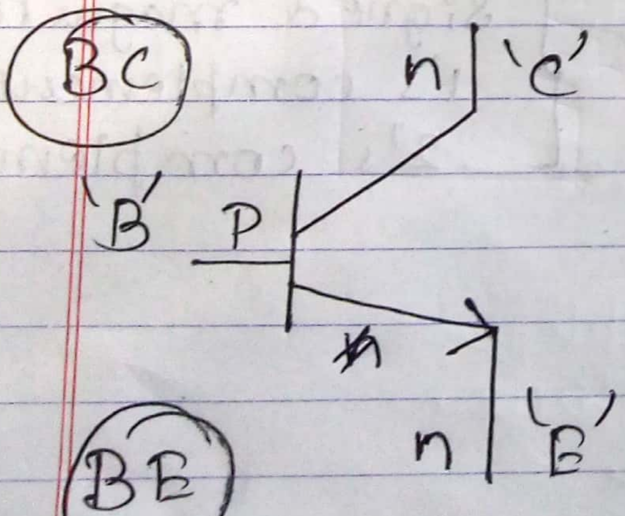
1. Active region

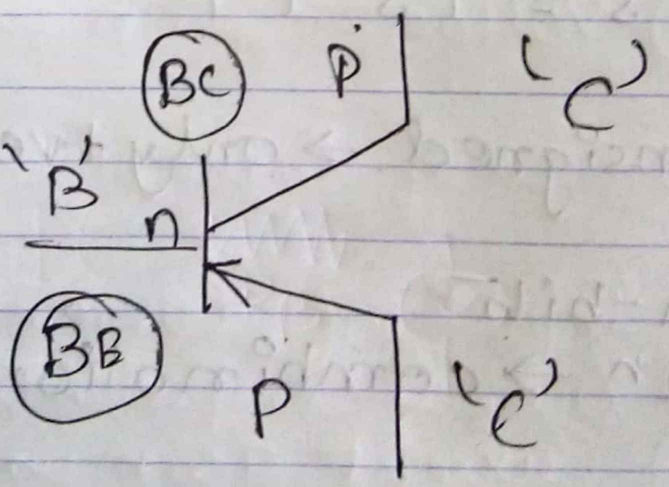
1.1 ~~Active~~ Forward active region (BC).
1.2 Reverse Active region (BE)

2. Cuttoff region (BE → Rev., BC → Rev.).

3. Saturation region.
(BE → Forward, BC → Forward).

Circuit Symbol of transistor





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Forward active region
→ act as typical active region.

Note – In active region transistor act as typical amplifier. In cutoff and saturation region act as ~~the~~ switch.

Application of transistor :–

1) Transistor act as Switch.
2) " " " amplifier.
3) " " " Oscillator.
4) " " " Filter.

For designing amplifier, needed two types of Sources.

e.g. as – DC Source, AC source

DC is biasing voltage.

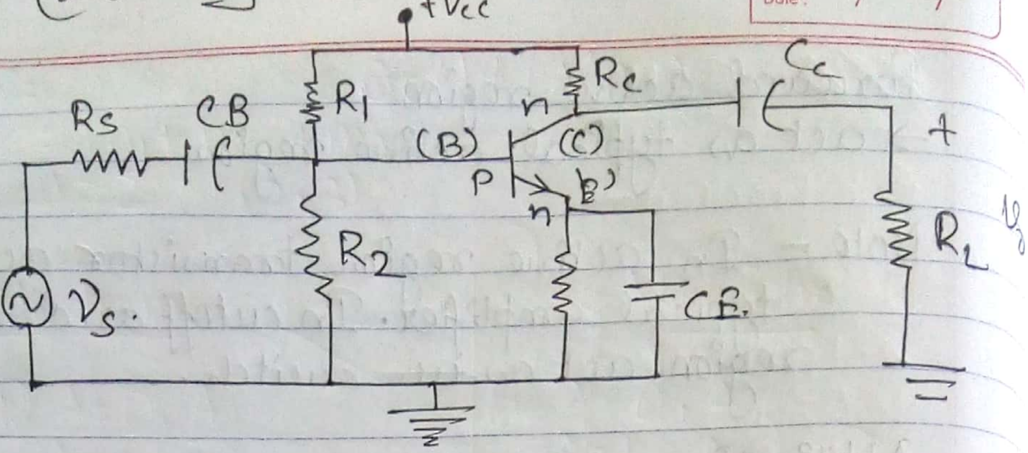
Types of biasing circuit :–

1. Self biased circuit.
2. Collector to base biased circuit.
3. Voltage divider biased ~~b~~ circuit.

Voltage divider biased is – obvious because its stability factor (S) is very low.

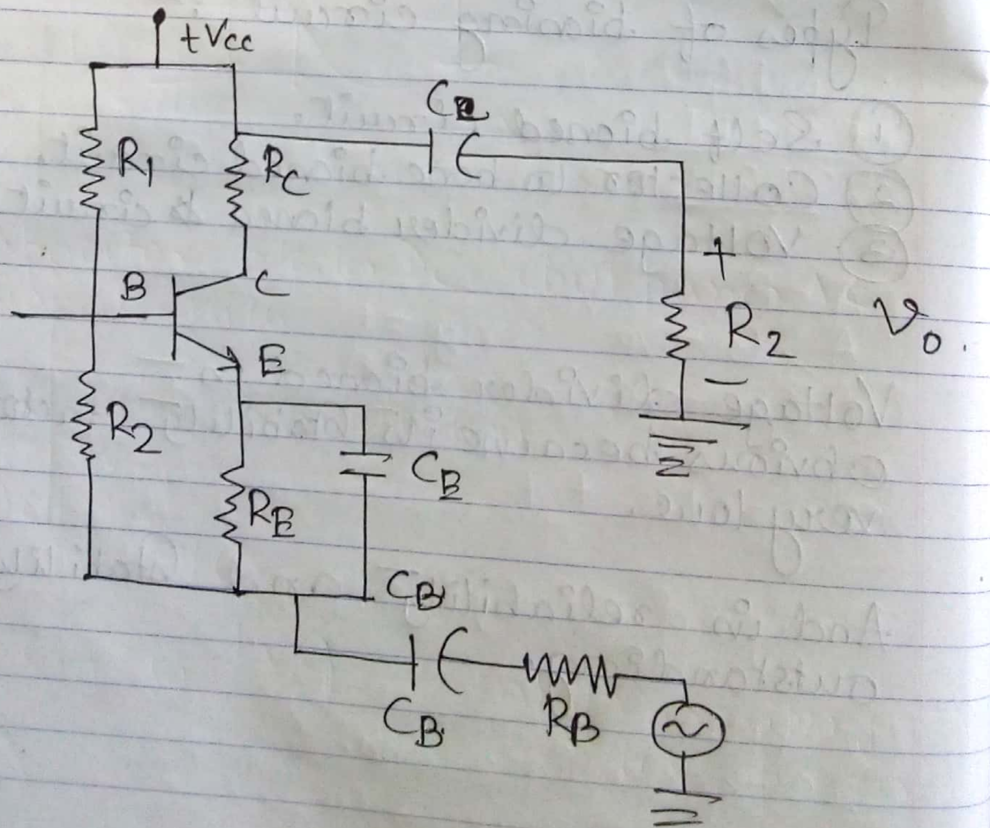
And is reliability and stability is outstanding.

@Omnayan@

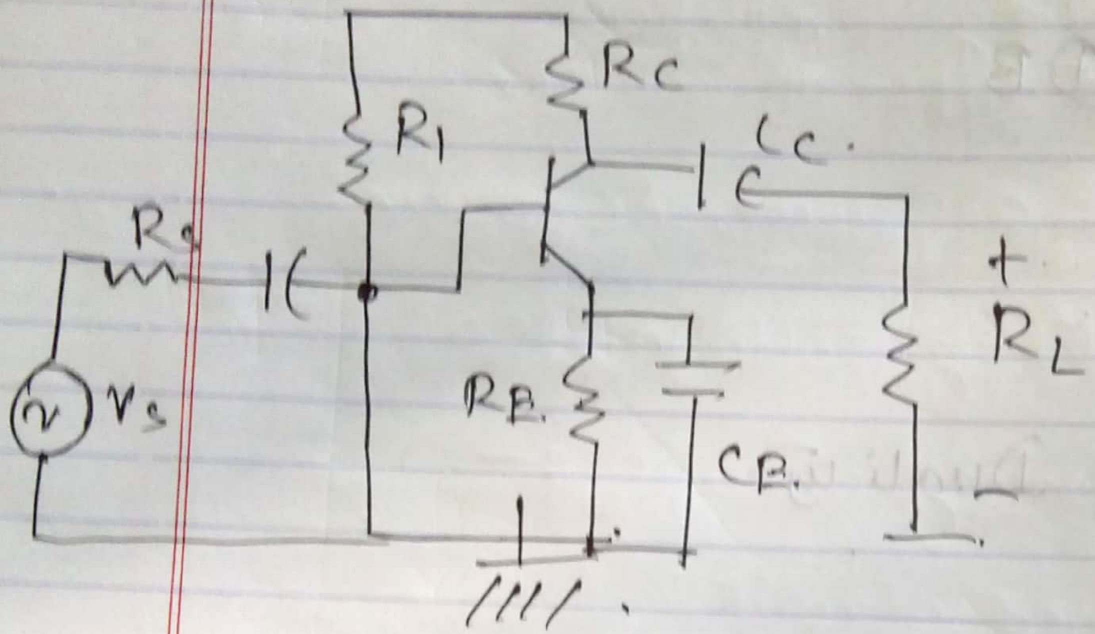


V_s = source voltage.
R_s = Source resistor.
C_B = Blocking capacitor.
R_1, R_2 = Stabilized resistor.
R_C = Collector "
R_E = Emitter "
C_E = Bypass capacitor
C_C = Coupling ".
R_L = Load resistance.
V_0 = Output voltage.

$$\boxed{\frac{v_0}{v_s} = \text{Gain}}$$



Analog (~~Mukherjee~~ Baner).

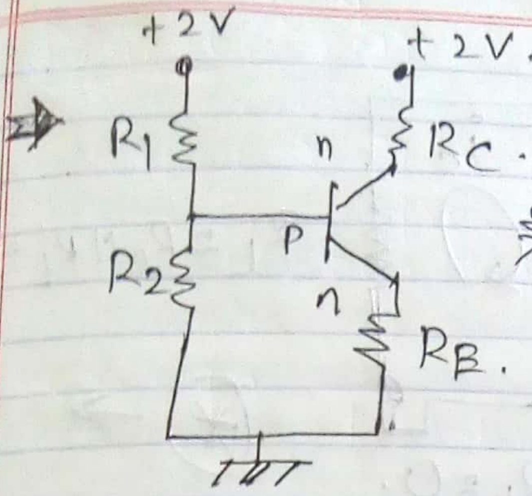


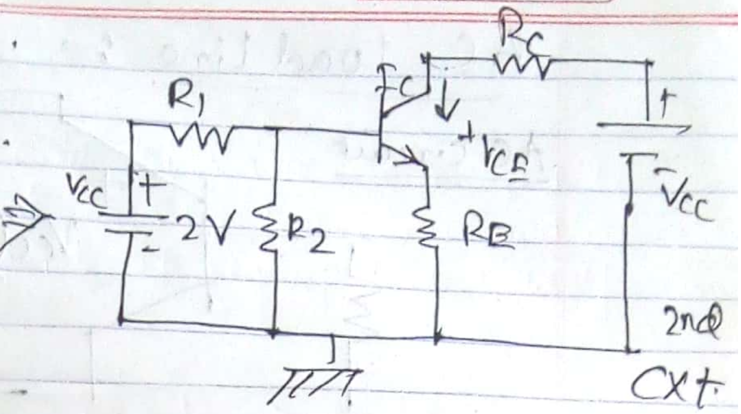
Capacitor blocks DC but passes AC signal easily

- D C load line Analysis

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

D C, $f = 0$, $|X_C| = \frac{1}{0} = \infty$ [Capacitor just like open circuit





$I_E = I_B + I_C$

$I_E = \frac{I_C}{\beta} + I_C$.

$-V_{CC} + I_C \times R_C + V_{CE} + I_E \times R_E = 0$

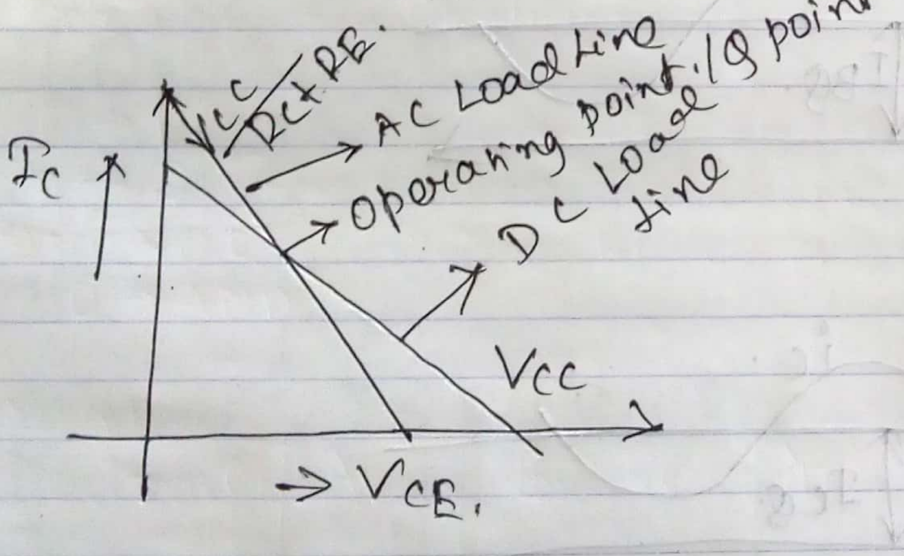
$-V_{CC} + I_C \times R_C + V_{CE} + I_C \times R_E = 0$.

$-V_{CC} + I_C (R_C + R_E) + V_{CE} = 0$

$$\boxed{I_C (R_C + R_E) + V_{CE} = V_{CC}.}$$

Straight Line

→ DC Load Line.



When, $I_C = 0$, $V_{CE} = V_{CC}$.

$V_{CE} = 0$, $I_C = \frac{V_{CC}}{R_L + R_C}$.

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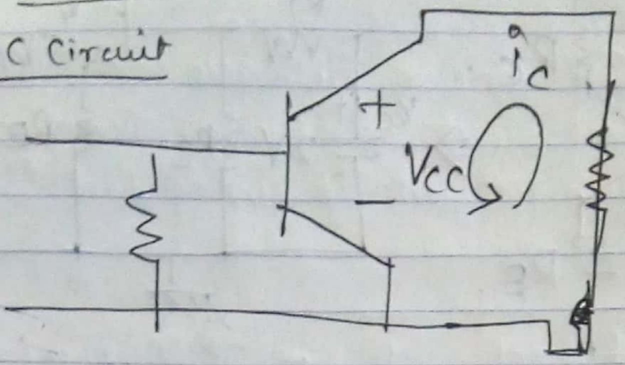
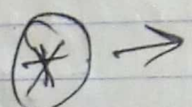
AC Load Line :-

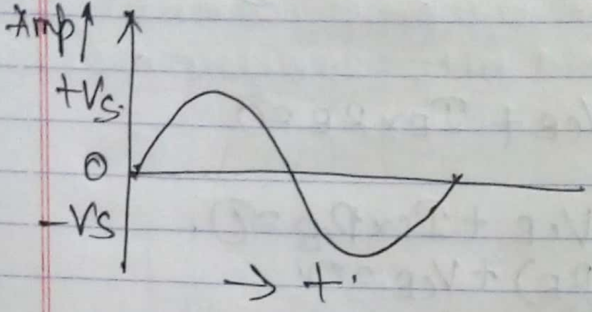
AC Circuit

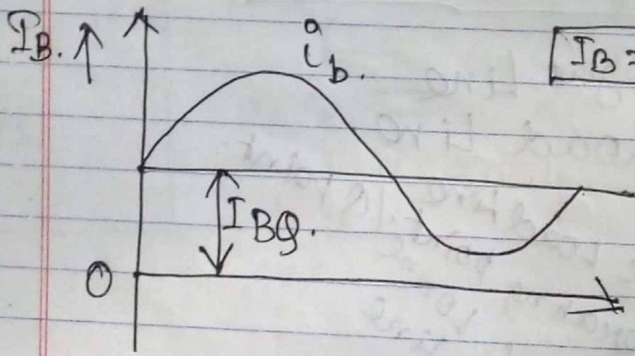


$R_L = R_C \parallel R_L + V_0$



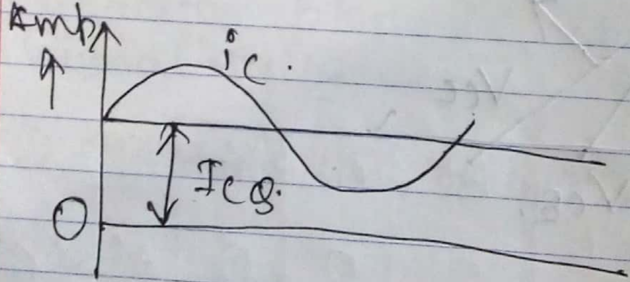
$+V_{ce} + i_c \times R_L = 0.$



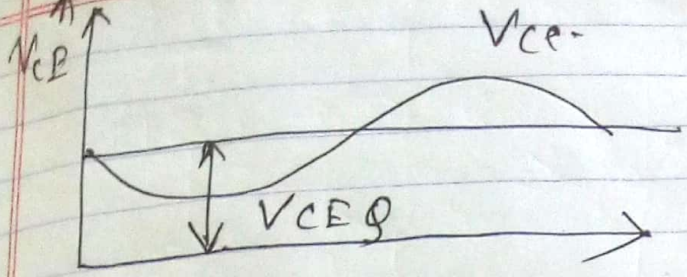


(DC) + (AC)

$\boxed{I_B = I_{BQ} + i_b}$



$\boxed{I_C = I_{CQ} + i_c}$



$$\boxed{V_{CE} = V_{CEQ} + v_{ce}}$$

$$(V_{CE} - V_{CEQ}) + (I_C - I_{CQ}) \times R'_L = 0$$

$$\boxed{V_{CE} + I_C \times R'_L = V_{CEQ} + V_{CEQ} * R'_L}$$

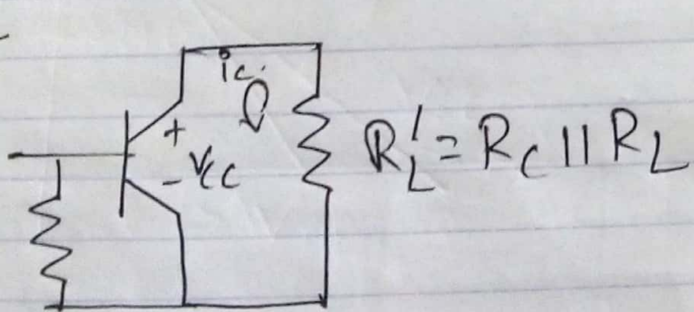
Case-1, $I_C = 0$, $V_{CE} = \cancel{Q_1}\, V_{CEQ} \times R'_L$.

Case-2, $V_{CE} = 0$, $I_C = \frac{V_{CEQ}}{R'_L} + V_{CEQ}$.

① ~~Class-A~~ ~~$(\theta = 360^\circ)$~~

~~②ClassAB~~

* ac ckt



$$+V_{CE} - I_C R'_L = V_{CEQ} + V_{CEQ} \times R'_L$$

Case 1, $I_C = 0$, $V_{CE} = V_{CEQ} + V_{CEQ} + R'_L$.

$V_{CE} = 0$, $I_C = V_{CEQ} + V_{CEQ}/R'_L$

used.

9.6 CLASS A POWER AMPLIFIER

The circuit diagram of a CE-mode class A power amplifier using an *n-p-n* transistor is shown in Fig. 9.8. Of all the three possible transistor configurations, the CE-mode gives the largest power gain. In Fig. 9.8, the amplifier is directly coupled to the load resistance R_L. As the operation is class A, the collector current flows over the entire cycle of the input signal. The blocking capacitor C_1 prevents the interaction between the dc voltage of the base circuit and the ac signal voltage.

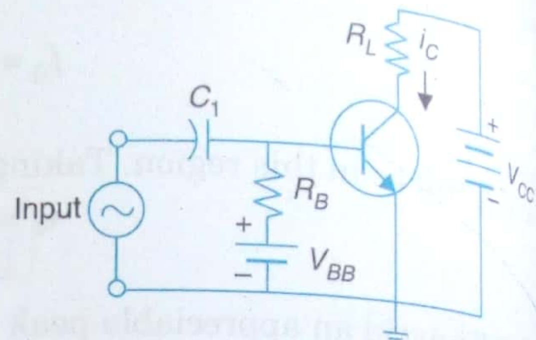


Fig. 9.8 A simple class A power amplifier directly coupled to the load resistance.

The circuit of Fig. 9.8 employs two batteries. A circuit using a single battery and a voltage divider bias is shown in Fig. 8.5. It is assumed here that the battery can deliver the desired power.

Analysis

The analysis uses the output characteristics of the transistor (Fig. 9.9). The transistor must not dissipate a power greater than the allowable limit specified by the manufacturers. For a given ambient temperature, the *maximum power dissipation* is given by

$$V'_C I'_C = \text{constant}, \tag{9.32}$$

where V'_C is the collector voltage and I'_C is the collector current. This equation represents a hyperbola marked "maximum collector dissipation curve" in the output characteristics of Fig. 9.9. The load line must lie below this curve so that the power dissipation in the transistor is within the safe limit.

The load line for the supply voltage V_{CC} and the load resistance R_L (Fig. 9.8) is shown in Fig. 9.9. The quiescent point Q is chosen near the centre of the load line to achieve a maximum ac power output. For the maximum input signal, the operating point moves along the load line between cutoff and saturation. At cutoff, the collector voltage attains a maximum value $V_{C_{\max}}$, and the collector current attains a minimum value $I_{C_{\min}}$. At saturation, on the other hand, the collector voltage has a minimum value $V_{C_{\min}}$ and the collector current a maximum value $I_{C_{\max}}$. As the operating point swings equally on either side of the Q-point, the amplitude of the ac collector current I_m and that of the ac collector voltage V_m are given by

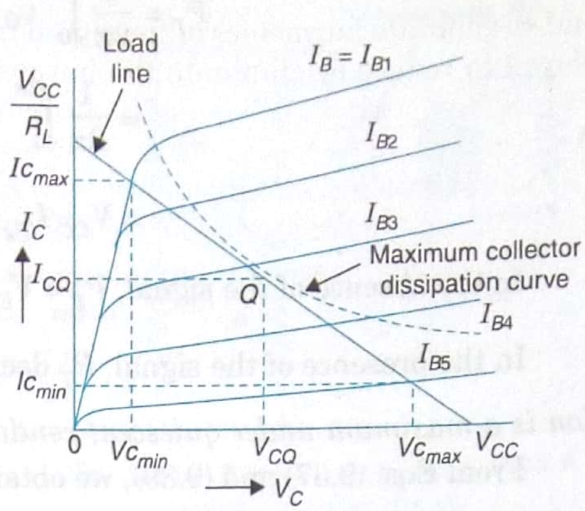


Fig. 9.9 Output characteristics for the analysis of a direct-coupled class A power amplifier.

$$I_m = \frac{I_{C\max} - I_{C\min}}{2} \tag{9.33}$$

and

$$V_m = \frac{V_{C_{\max}} - V_{C_{\min}}}{2} = I_m R_L \tag{9.34}$$

Calculation of Power

The instantaneous collector current is

$$i_C(t) = I_{CQ} + I_m \sin \omega t, \tag{9.35}$$

where ω is the angular frequency of the ac signal. The instantaneous collector-emitter voltage drop is

$$v_C(t) = V_{CC} - i_C R_L = V_{CC} - (I_{CQ} + I_m \sin \omega t) R_L \tag{9.36}$$

The instantaneous power dissipated in the load resistance is $P_L(t) = i_C^2 R_L$. The average power in the load is

$$P_L = \frac{1}{2\pi} \int_0^{2\pi} i_C^2 R_L \, d(\omega t) = I_{CQ}^2 R_L + \frac{1}{2} I_m^2 R_L, \tag{9.37}$$

where Eq. (9.35) is used. In the absence of the signal, $P_L = I_{CQ}^2 R_L$, which is the dc power dissipated in R_L. In the presence of the signal, P_L increases by $\frac{1}{2} I_m^2 R_L$, which represents the ac power P_{ac} in the load. Thus,

$$P_{ac} = \frac{1}{2} I_m^2 R_L = \frac{1}{2} I_m V_m = I_{rms} V_{rms}, \tag{9.38}$$

where I_{rms} and V_{rms} are the rms or effective values of the ac collector current and voltage, respectively. The instantaneous power dissipated in the transistor is $P_T(t) = v_C(t)\, i_C(t)$. The average power dissipation in the transistor is

$$P_T = \frac{1}{2\pi}\int_0^{2\pi} v_C(t)\, i_C(t)\, d(\omega t).$$

$$= \frac{1}{2\pi}\int_0^{2\pi} (V_{CC} - I_{CQ} R_L - I_m R_L \sin \omega t)(I_{CQ} + I_m \sin \omega t)\, d(\omega t)$$

$$= V_{CC} I_{CQ} - I_{CQ}^2 R_L - \frac{1}{2} I_m^2 R_L \tag{9.39}$$

In the absence of the signal, $P_T = V_{CC} I_{CQ} - I^2_{CQ} R_L = V_{CQ} I_{CQ}$.

In the presence of the signal, P_T decreases by $\frac{1}{2} I_m^2 R_L = P_{ac}$. Thus, *the collector dissipation is a maximum under quiescent conditions*.

From Eqs. (9.37) and (9.39), we obtain

$$P_L + P_T = V_{CC} I_{CQ} = P_s, \tag{9.40}$$

where P_s is the dc power supplied by the collector battery. In the presence of the signal, P_L increases by P_{ac} and P_T decreases by P_{ac}, so that the total power $(P_L + P_T)$ is a constant equal to the power supplied by the source under quiescent conditions. This is a manifestation of the *principle of conservation of energy*.

Collector Efficiency

The output power is

$$P_{ac} = \frac{1}{2} I_m V_m = \frac{1}{8}(I_{C\max} - I_{C\min})(V_{C\max} - V_{C\min}), \tag{9.41}$$

where Eqs. (9.33) and (9.34) have been used. The collector efficiency or the conversion efficiency is

$$\eta = \frac{P_{ac}}{P_s} = \frac{(I_{C\max} - I_{C\min})(V_{C\max} - V_{C\min})}{8 V_{CC} I_{CQ}} \tag{9.42}$$

For ideal collector characteristics,

$V_{C\min} = 0$, $I_{C\max} = 2 I_{CQ}$, $V_{C\max} = V_{CC}$, and $I_{C\min} = 0$. Hence, Eq. (9.42) gives

$$\eta = \frac{2 I_{CQ} V_{CC}}{8 I_{CQ} V_{CC}} = 0.25 \tag{9.43}$$

Thus, the maximum efficiency obtainable from a class A power amplifier directly coupled to the load resistance is 25%. The overall efficiency is still smaller since the power taken by the base is not accounted for in the above analysis.

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(*i*) In a *double-tuned amplifier*, two parallel resonant circuits tuned to the same frequency and having equal effective Q values are used. One tuned circuit is connected to the collector of the first transistor and the other tuned circuit is connected to the base of the second transistor. The two tuned circuits are magnetically coupled and the two transistors are in the CE-mode (Fig. 9.20A).

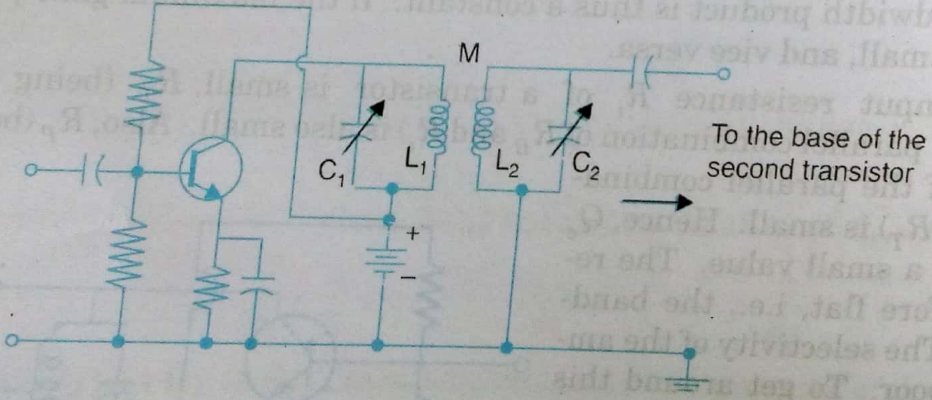

Fig. 9.20A. A double-tuned amplifier.

curves ... (a) and (b), respectively ... depicted in Figs. 9.22

9.10 CLASS B AND CLASS C TUNED POWER AMPLIFIERS

The output efficiency of an amplifier increases as the operation shifts from class A to class C through class AB and class B. As the output power of a radio transmitter is high and the efficiency is of prime concern, class B and class C amplifiers are used at the output stages in transmitters.

The operations of class B and class C amplifiers are nonlinear since the amplifying elements remain cutoff during a part of the input signal cycle. The nonlinearity generates at the output of the amplifier harmonics of the signal frequency. In the push-pull arrangement where the bandwidth requirement is not limited, these harmonics can be eliminated or reduced. When a narrow bandwidth is desirable, a resonant circuit is employed in class B and class C tuned RF power amplifiers to eliminate the harmonics.

The basic circuit of any class of tuned amplifiers is the same; the value of the base-emitter (BE) junction bias voltage distinguishes between class A, class B or class C operation. For class A operation, the BE junction is forward-biased. In a class B amplifier, the BE junction bias voltage just cancels the barrier potential difference across the junction. In class C operation, the BE junction is reverse-biased.

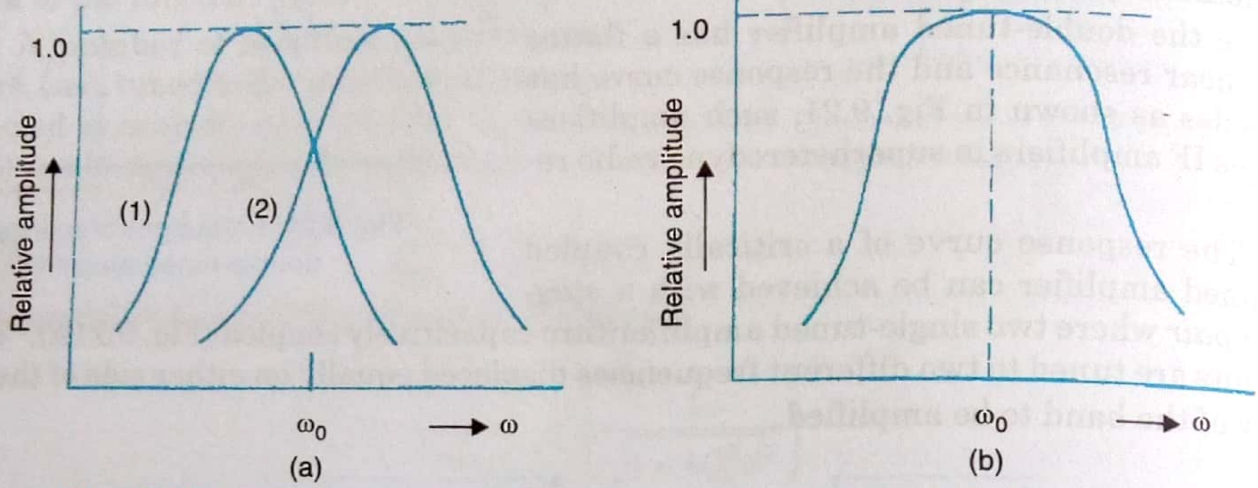

Fig. 9.22 Amplitude response of a stagger-tuned pair: (a) individual responses: curve (1) for the first stage, curve (2) for the second stage : (b) overall response.

Fig. 9.23 depicts the circuit of a class C tuned amplifier in the CE configuration. The source voltage V_{BB} biases the BE junction of the transistor considerably below cutoff. The capacitor C' acts as an RF short. RFC is a radio frequency choke whose high impedance prevents the DC source V_{BB} from shorting the RF input voltage v_i. The signal voltage v_i must have a sufficiently large amplitude so that the BE junction is forward-biased for less than 180° of the input signal cycle. The transistor starts conducting when v_i exceeds V_L where V_L is V_{BB} plus the transistor cutin voltage. The conduction stops when v_i drops below V_L. The collector current therefore consists of a series of pulses, there being one pulse for each cycle of the input signal. The peaks of these pulses resemble the peaks of the input sine wave and the duration of each pulse is less than half the period of the input signal. Clearly, harmonics of the input signal are present in the collector current. As the parallel resonant circuit, providing the load impedance, is tuned to the input frequency and has a high Q, the load impedances at the harmonic frequencies are negligibly small compared to that at the resonant frequency. Thus only the fundamental component of the load voltage corresponding to the input frequency is significant at the output. The output voltage across the tuned circuit is sinusoidal and has the same frequency as that of the input signal.

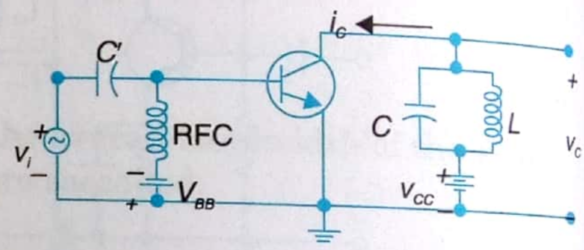

Fig. 9.23 A class C tuned amplifier

The instantaneous value of the collector voltage is a minimum when the collector current is a maximum. So, the ac component of the collector voltage is sinusoidal and 180° out of phase with the input signal voltage. The waveforms for a sinusoidal input signal voltage in a class C amplifier is shown in Fig. 9.24.

A class C tuned amplifier operates at the resonant frequency f_0 of the tank circuit, where $f_0 = \dfrac{1}{2\pi\sqrt{LC}}$. The bandwidth is given by $B = \dfrac{f_0}{Q_e}$, where Q_e is the effective Q of the tank circuit.

If I_l is the peak value of the fundamental component of the collector current, the *ac* ou[tput] power for maximum drive is $P_{ac} = V_{cc}\ I_1/2$. The *dc* power supplied by the collector sup[ply is] $P_s = I_o V_{CC}$ where I_o is the average value of the collector current. The efficiency is $\eta = P_{ac}/P_s =$ $I_1/(2I_o)$. Since I_o is quite small, η is very high (about 90%).

A class C tuned amplifier can be used as a *frequency multiplier* if the resonant cir[cuit is] tuned to a harmonic of the input signal.

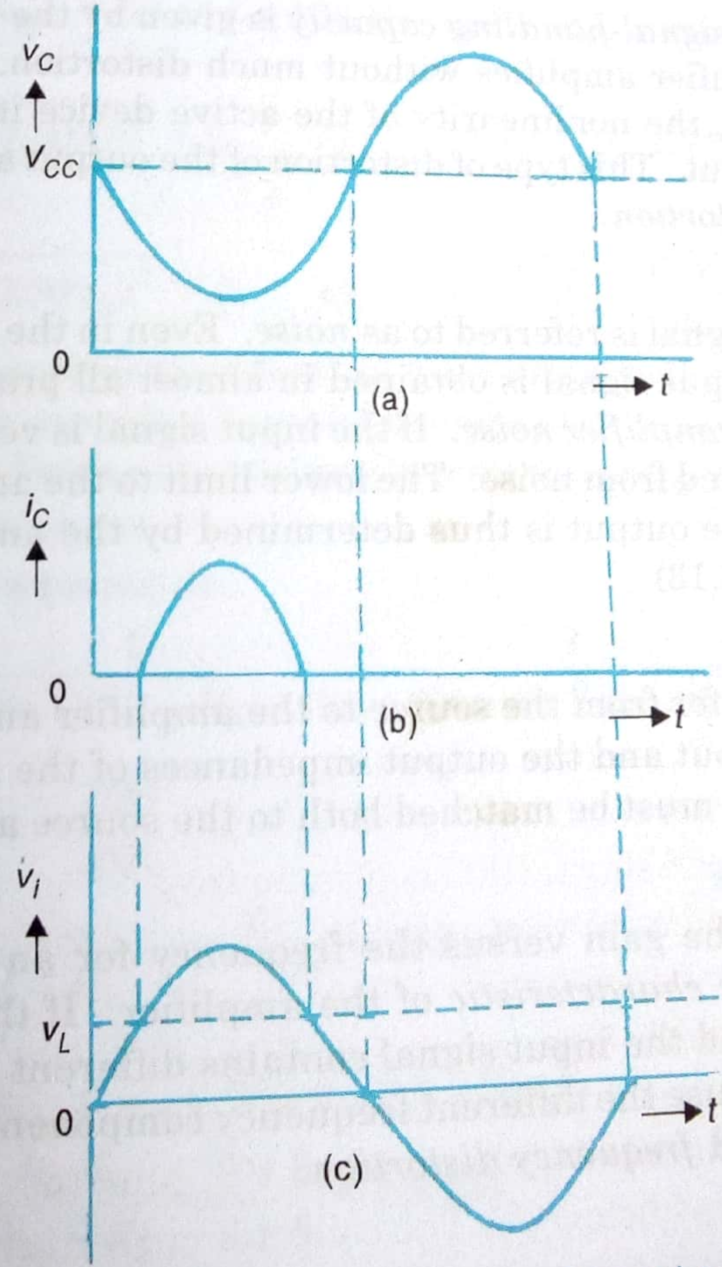

Fig. 9.24 Typical waveforms in class C tuned amplifiers: (a) collector voltage v_c against time t, (b) collector current i_c against t, and (c) input ac voltage v_i as a function of t.

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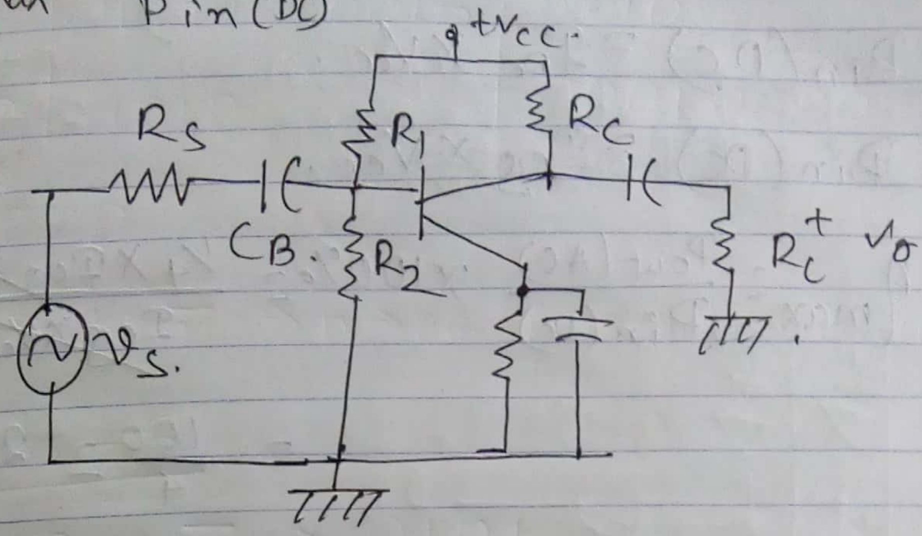
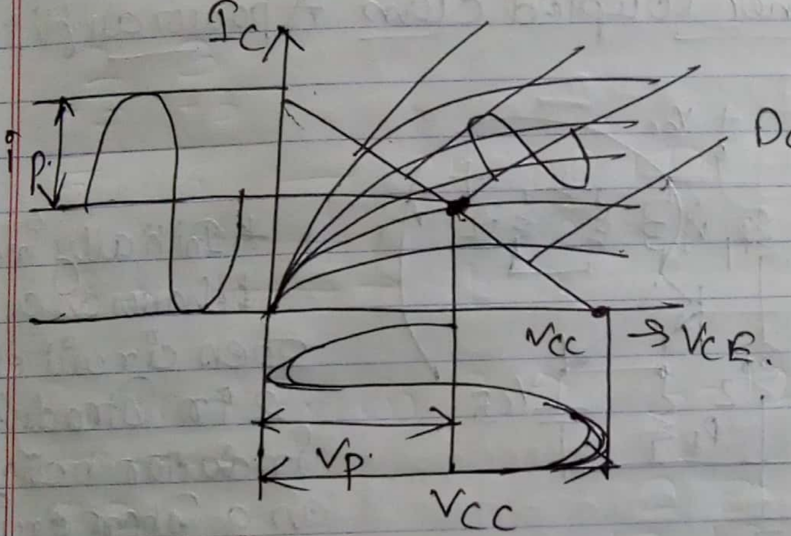
PB(Analog)

$$\eta_{max} = \frac{P_{out}(AC)}{P_{in}(DC)} \times 100\%.$$





$$V_p = \frac{V_{cc}}{2}.$$

$$i_p = I_{CQ}.$$

$P_{out}(AC) = i_c/rms \times V_{CE}/rms.$

$$I_C = \underset{(DC)}{I_{CQ}} + \underset{(AC)}{i_p \cos\omega_0 t}$$

$$V_{CE} = \underset{(DC)}{V_{CEQ}} + \underset{(AC)}{V_p \cos(\omega_0 t + \pi)}.$$

Again,

$$P_{out}(AC) = i_c/rms \times V_{CE}/rms$$
$$= i_p/\sqrt{2} \times V_p/\sqrt{2} = \frac{1}{2} \times I_{CQ} \times \frac{V_{cc}}{2}.$$

$P_{out}(AC) = \frac{1}{4} \times I_{CQ} \times V_{CC}$.

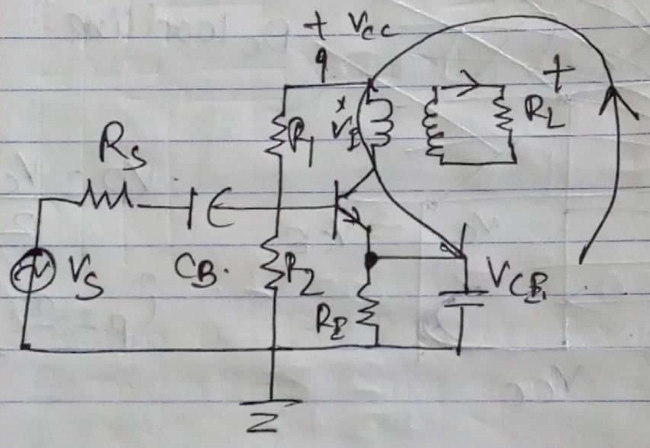
$P_{in}(DC) = I_{DC} \times V_{CC}$.

$P_{in}(DC) = I_{CQ} \times V_{CC}$.

$\eta_{max} = \frac{P_{out}(AC)}{P_{in}(DC)} \times 100\% = \frac{1/4 \times I_{CQ} \cancel{\times V_{CC}}}{I_{CQ} \cancel{\times V_{CC}}} \times 100$

$= \frac{100}{4} = 25\%$.

Transformer coupled class A power amplifier:

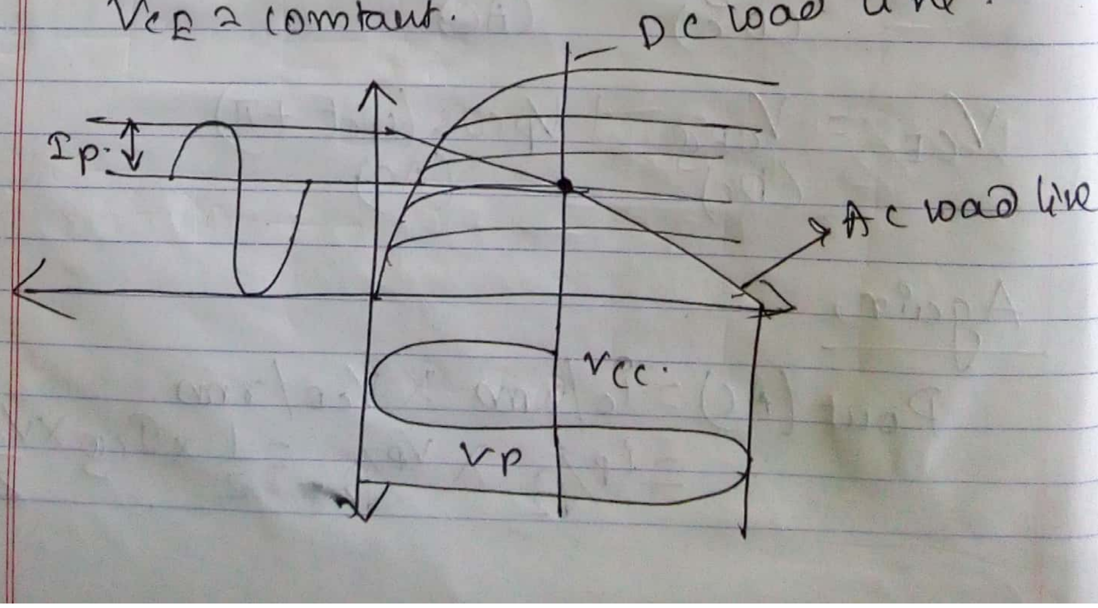


* Initially a inductor behaves as a open circuit element & in steady state inductor behaves as a short circuit element.

$-V_{CC} + V_T + V_{CE} = 0$.

$V_{CE} = V_{CC} - V_T$.

V_{CE} = constant.



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$$\boxed{V_p = V_{cc}}$$

$$\boxed{i_p = I_{cq}}$$

$$P_{out}(AC) = i_c/rms \times V_{ce}/rms$$

$$= i_p/\sqrt{2} \times \frac{V_p}{\sqrt{2}}$$

$$= \frac{1}{2} \times i_p \times V_p.$$

$$P_{out}(AC) = \frac{1}{2} \times i_{cq} \times V_{cc}.$$

$$P_{in}(DC) = I_{DC} \times V_{cc}.$$

$$\boxed{P_{in}(DC) = I_{cq} \times V_{cc}}.$$

$$\eta_{max} = \frac{P_{out}(AC)}{P_{in}(DC)} \times 100.$$

$$= \frac{1/2 \times \cancel{i_{cq} \times V_{cc}} \times 100}{\cancel{I_{cq} \times V_{cc}}}$$

$$= \frac{100}{2} = 50\%.$$

$\boxed{\text{2ND MID SEM}}$

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Feedback & Oscillator

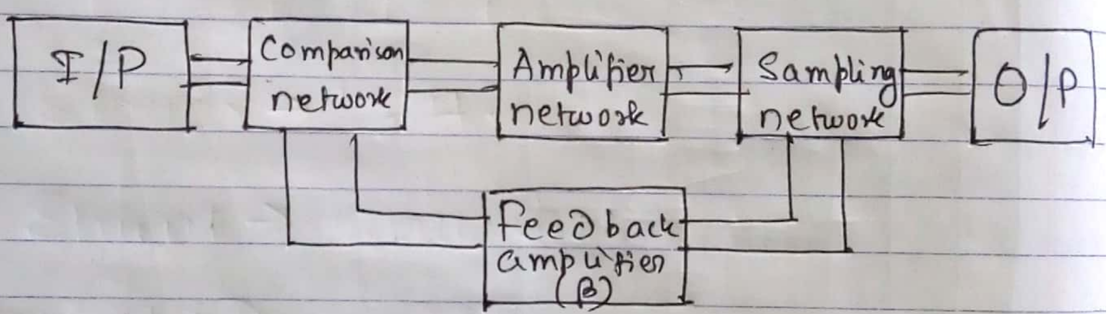
The term 'Feedback' implies transfer of energy from the output of a system to the input. It is the position of the input signal of a amplifier is feed and super imposed on the input signal, the performance of the amplifier changes significantly. The amplifier is then said to be a feedback amplier.

Types of feedback:—

i) +ve feedback → It's unstable & gain is infinite.

ii) -ve " → It's stable & gain is low

Block Diagram of Feedback Amplifier:—



Feedback Topologies:—

1) Voltage Series feedback.

2) Voltage Shunt "

3) Current Series "

4) Current Shunt "

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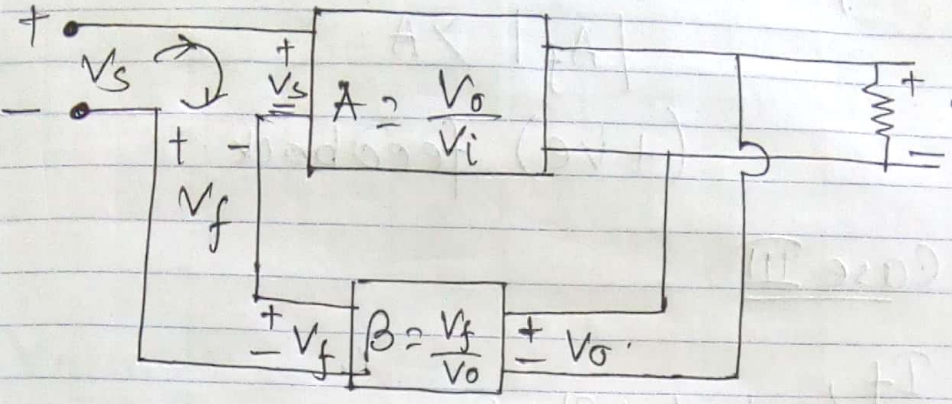
(*) General expression for obtaining gain of feedback amplifier.



Overall Transfer Gain $(A_f) = \frac{V_o}{V_s}$.

$A = \frac{V_o}{V_i}$.

$\Rightarrow \cancel{A/7}\ V_o = A \cdot V_i$

$\Rightarrow V_o = A(V_s - V_f)$

$\Rightarrow V_o = AV_s - AV_f$.

$\Rightarrow V_o = A \cdot V_s - AV_f$.

$\Rightarrow V_o + A\beta V_o = AV_s$.

$\Rightarrow V_o[1 + A\beta] = A \cdot V_s$.

$\boxed{V_i = V_s - V_f}$

$\beta = \frac{V_f}{V_o}$

$\Rightarrow \boxed{V_f = \beta V_o}$

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

Case - 1

If $|1 + A\beta| > 1$

$|A_f| < A$, then -ve feedback.

Case - II

If $|1+A\beta| < 1$.

$|A_f| > A$.

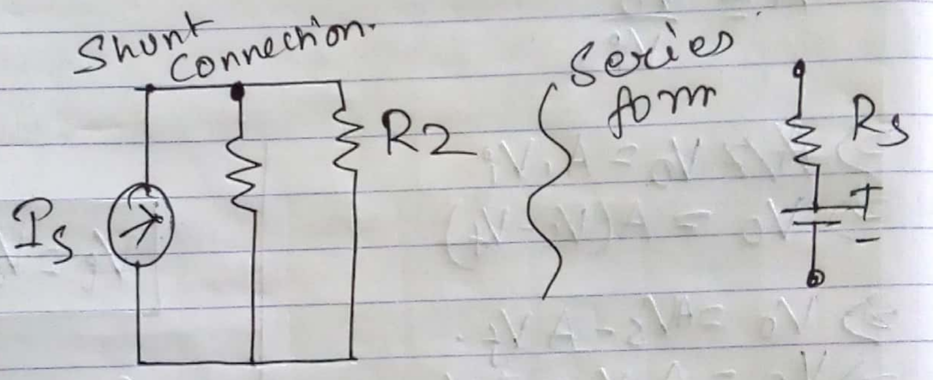
(+ve) feedback.

Case III

If, $|1+A\beta| = 0$

$|A_f| = \infty$

Then it treated as a Oscillator



Series → V_o H.

Shunt → Current.

Volt → Parallel

Current → Series.

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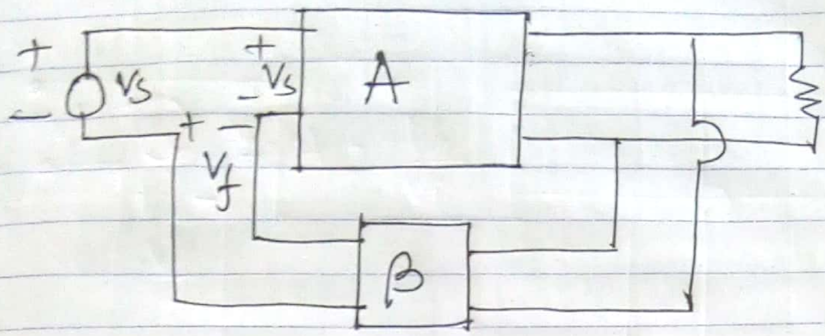
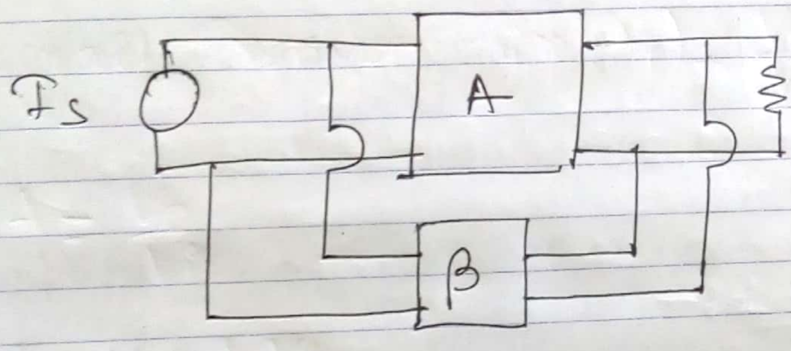
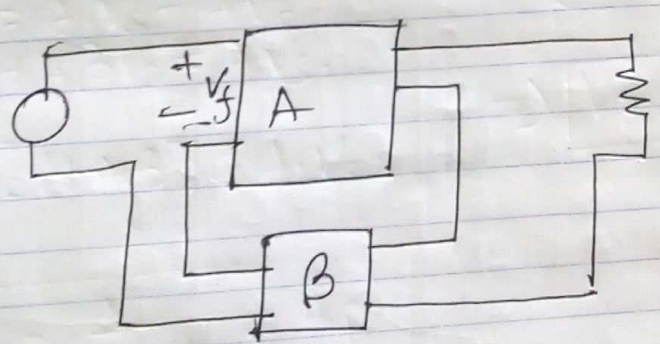
VOLTAGE SERIES



Voltage Shunt:-



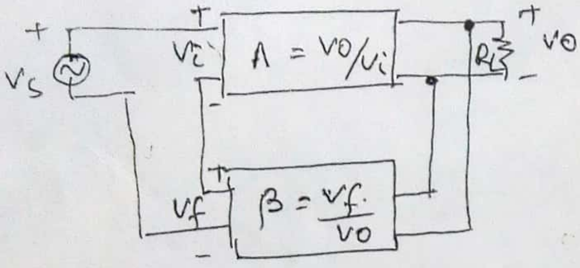
Current Series:-



Topologies of feedback

1) voltage series feedback
2) voltage shunt feedback
3) Current series feedback
4) Current shunt feedback

voltage series feedback



$V_f = \beta V_o$ as $\beta = \frac{V_f}{V_o}$

$V_i = V_s - V_f = V_s - \beta V_o$

$A = \frac{V_o}{V_i}$

$V_o = AV_i = A(V_s - \beta V_o)$

$A_f = \frac{V_o}{V_s}$

$V_o = A(V_s - \beta V_o)$

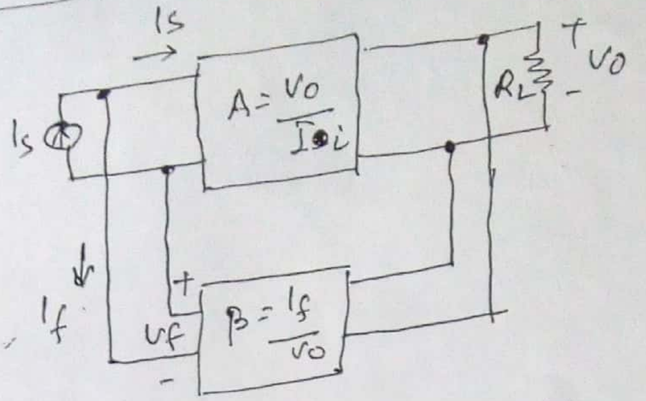
$= AV_s - A\beta V_o$

or $V_o + A\beta V_o = AV_s \Rightarrow V_o(1+A\beta) = AV_s$.

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

∴ the feedback reduces the gain by a factor $(1+A\beta)$ [negative feedback].

Voltage Shunt



$I_i \rightarrow$ input current

$I_f \rightarrow$ feedback current

$I_s \rightarrow$ Source current

$V_o \rightarrow$ O/P voltage

$$A = \frac{V_o}{I_i} \quad \text{or} \quad V_o = A I_i$$

$$\beta = \frac{I_f}{V_o} \quad \text{or} \quad I_f = \beta V_o$$

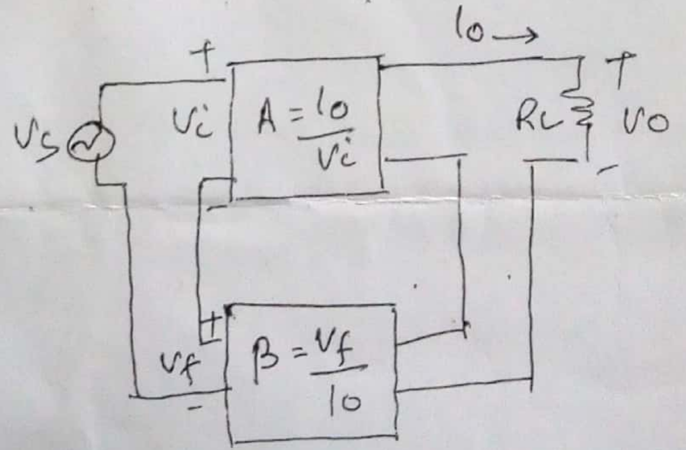
$$\therefore I_s = I_i + I_f = I_i + \beta V_o$$

$$A_f = \frac{V_o}{I_s} = \frac{V_o}{I_i + \beta V_o}$$

$$= \frac{A I_i}{I_i + \beta V_o} = \frac{A}{1 + A\beta}$$

- amplifier gain is reduced by a factor $(1 + A\beta)$ due to feedback.

Current Series feedback.



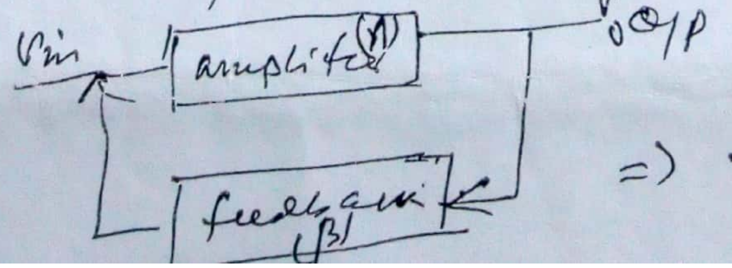
$$A_f = \frac{I_o}{I_s} = \frac{I_o}{I_i + I_f} = \frac{I_o}{I_i + \beta I_o}$$

$$= \frac{A I_i}{I_i + \beta A I_i} = \frac{A}{1 + A\beta}$$

the gain is reduced by a factor $(1 + A\beta)$ when feedback is introduced.

Oscillator

An oscillator is a circuit which produces a continuous, repeated, alternating waveform without any input. This circuit converts unidirectional current flow from a DC source into an alternating waveform which is of the desired frequency, as decided by circuit components.



=) Typical oscillator.

Oscillator :-

A system that consist of active and passive circuit element to produce a sinusoidal or other repetative wave form.

Types of oscillator :-

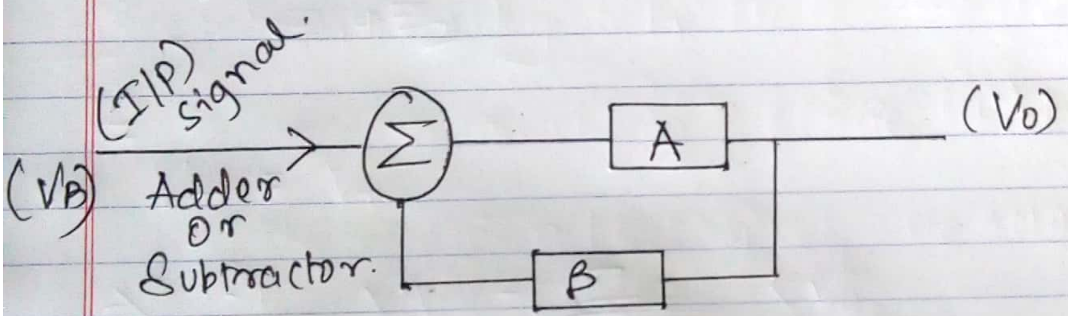
1) AF (0-20 kHz)

(a) ~~Wein~~ Wien Bridge
(b) phase-shift.

(ii) RF (20 kHz - 30 kHz)

(a) Hartley
(b) Colpitts
(c) Tuned

Block diagram of oscillator :-



$$\boxed{\frac{V_o}{V_s} = \frac{A}{1 - A\beta}}$$

Condition of oscillator :-

$1 - A\beta = 0 \Rightarrow A\beta = 1$.

$A\beta$ → loop gain.

$(0^\circ \text{ or } \pm 360^\circ)$.

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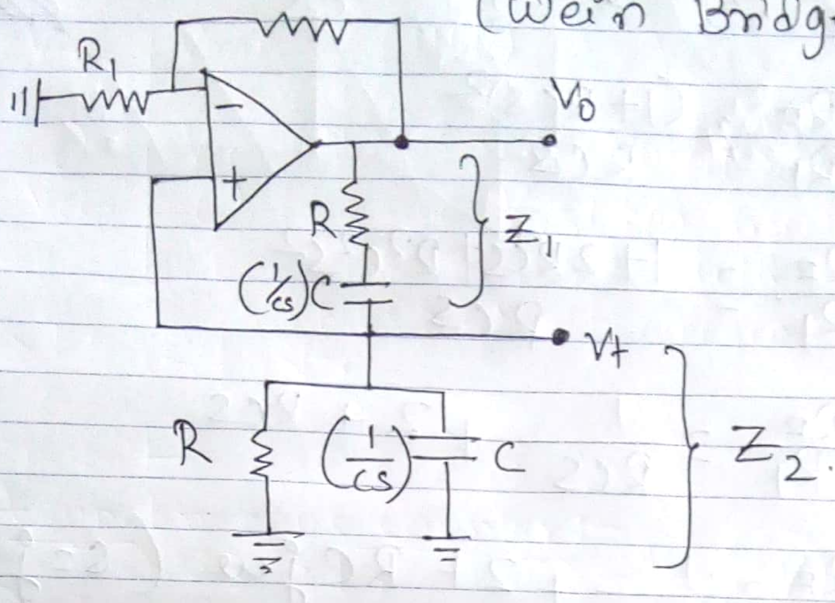
PB (Analog). (Wein Bridge Oscillator).



$$V_- = \frac{R_1}{R_1+R_F} \times V_0$$

$$V_+ = \left(\frac{Z_2}{Z_1+Z_2}\right) \times V_0.$$

$$V_- = V_+$$

$$\Rightarrow \frac{R_1}{R_1+R_F} \times \cancel{V_0} = \frac{Z_2}{Z_1+Z_2} \times \cancel{V_0}$$

$$\Rightarrow \frac{R_1+R_F}{R_1} = \frac{Z_1+Z_2}{Z_2}$$

$$\Rightarrow \frac{R_F}{R_1}+1 = \frac{Z_1}{Z_2}+1$$

$$\Rightarrow Z_1 = R + 1/cs.$$

$$\Rightarrow Z_2 = \frac{R \times 1/cs}{R + 1/cs} = \frac{R}{cs} \times \frac{cs}{RCs+1} = \frac{R}{RCs+1}.$$

$$\frac{R_F}{R_1} = \frac{Z_1}{Z_2} \Rightarrow \frac{R_F}{R_1} = \frac{(1+RCS)}{CS} \times \frac{(1+RCS)}{R}$$

$$\Rightarrow \frac{R_F}{R_1} = \frac{(1+RCS)^2}{RCS}$$

$$\Rightarrow \frac{R_F}{R_1} = \frac{1+2RCS+R^2C^2S^2}{RCS}$$

$$\Rightarrow \frac{R_F}{R_1} = \frac{1}{RCS} + 2 + RCS$$

$$\Rightarrow \frac{R_F}{R_1} = \frac{1}{RC\cancel{S}j\omega} + 2 + RCj\omega \quad (s = j\omega)$$

$$\Rightarrow \frac{R_0}{R_1} + 0 \cdot j = 2 + j\left(RC\omega - \frac{1}{RC\omega}\right)$$

$$\frac{R_F}{R_1} \geq 2$$

$$RC\omega - \frac{1}{RC\omega} = 0$$

$$\Rightarrow RC\omega = \frac{1}{RC\omega} \Rightarrow R^2C^2\omega^2 = 1$$

$$\Rightarrow \omega = \frac{1}{RC} \Rightarrow 2\pi f = \frac{1}{RC}$$

$$\Rightarrow \boxed{f = \frac{1}{2\pi RC}}$$

11.7 PHASE-SHIFT OSCILLATOR

Fig. 11.6 shows the circuit of a phase-shift oscillator. The dc operating point of the transistor in the active region of its characteristics is established by the resistors R_1, R_2, R_L and R_E, and the supply voltage $-V_{cc}$. The capacitor C_E is a bypass capacitor.

As the transistor is in the CE configuration, it introduces a phase difference of 180° between its input and output voltages. The three sections of RC network give an extra phase difference of 180° so that the net phase shift around the loop is 0° or 360°.

For convenience, the three RC sections are taken to be identical. The resistance in the last

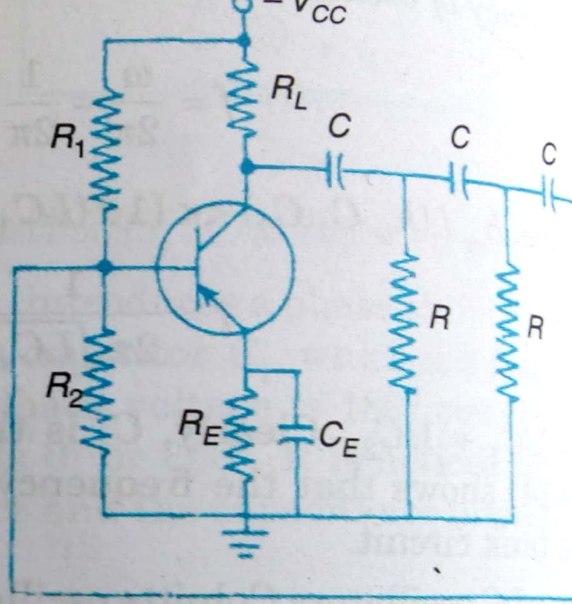


Fig. 11.6 Circuit of a phase-shift

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section is $R' = R - h_{ie}$. The input resistance h_{ie} of the transistor is added to R', thus giving a net resistance R. The RC phase-shift networks constitute the frequency determining circuit.

Analysis

The resistances R_1 and R_2 are large and therefore have no effect on the AC operation of the circuit. The $R_E - C_E$ parallel combination is also absent from the AC equivalent circuit of Fig. 11.7(*a*) due to the negligible impedance offered by this combination to ac. As $1/h_{oe} >> R_L$, one can neglect $1/h_{oe}$ which is in parallel with R_L. Also, h_{re} being small, $h_{re} V_2$ can be omitted. The equivalent circuit of Fig. 11.7(*a*) then simplified to the circuit of Fig. 11.7(*b*) where the current source is replaced by its equivalent voltage source.

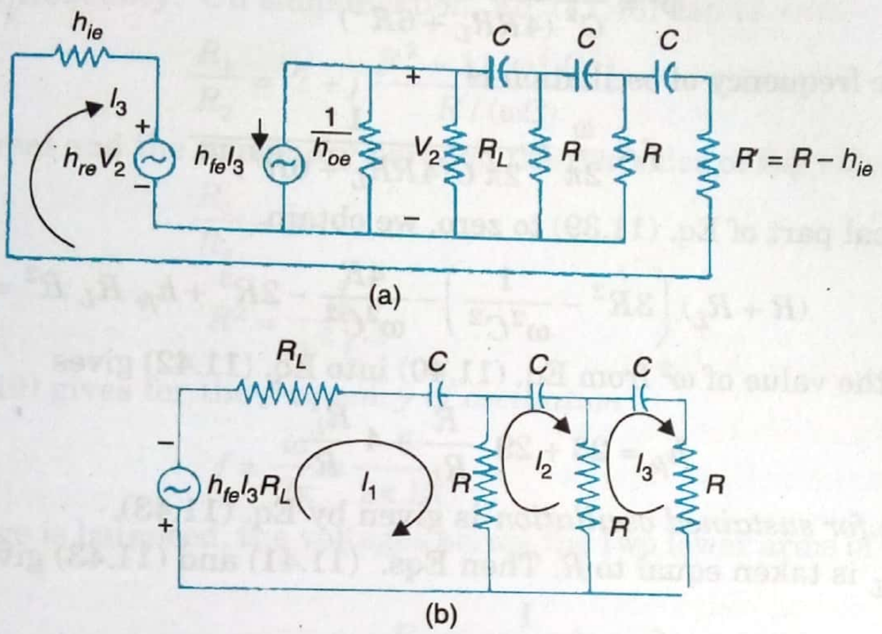

Fig. 11.7 (a) AC equivalent circuit of Fig. 11.6, (b) Simplified AC equivalent circuit with h_{re} V_2 and $1/h_{oe}$ neglected.

Kirchhoff's voltage law equations for the three loops in Fig. 11.7(*b*) are respectively written as

$$\left(R_L + R - \frac{j}{\omega C}\right) I_1 - RI_2 + h_{fe} R_L I_3 = 0 \tag{11.35}$$

$$-RI_1 + \left(2R - \frac{j}{\omega C}\right) I_2 - RI_3 = 0 \tag{11.36}$$

and

$$0.I_1 - RI_2 + \left(2R - \frac{j}{\omega C}\right) I_3 = 0 \tag{11.37}$$

where ω is the angular frequency of oscillation. Since the currents I_1, I_2, and I_3 are nonzero, the determinant of the coefficients of I_1, I_2, and I_3 in Eqs. (11.35) through (11.37) must vanish. Thus

$$\begin{vmatrix} R_L + R - \frac{j}{\omega C} & -R & h_{fe} R_L \\ -R & 2R - \frac{j}{\omega C} & -R \\ 0 & -R & 2R - \frac{j}{\omega C} \end{vmatrix} = 0. \tag{11.38}$$

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or,
$$\left(R_L + R - \frac{j}{\omega C}\right)\left(3R^2 - j\frac{4R}{\omega C} - \frac{1}{\omega^2 C^2}\right)$$
$$- R^2\left(2R - \frac{j}{\omega C}\right) + h_{fe} R_L R^2 = 0 \tag{11.39}$$

Equating the imaginary part of Eq. (11.39) to zero, we get
$$- 4R\frac{(R + R_L)}{\omega C} - \frac{1}{\omega C}\left(3R^2 - \frac{1}{\omega^2 C^2}\right) + \frac{R^2}{\omega C} = 0$$

or,
$$\omega^2 = \frac{1}{C^2(4RR_L + 6R^2)} \tag{11.40}$$

Therefore, the frequency of oscillation is
$$f = \frac{\omega}{2\pi} = \frac{1}{2\pi C(4RR_L + 6R^2)^{1/2}} \tag{11.41}$$

Putting the real part of Eq. (11.39) to zero, we obtain
$$(R + R_L)\left(3R^2 - \frac{1}{\omega^2 C^2}\right) - \frac{4R}{\omega^2 C^2} - 2R^3 + h_{fe} R_L R^2 = 0 \tag{11.42}$$

Substituting the value of ω^2 from Eq. (11.40) into Eq. (11.42) gives
$$h_{fe} = 23 + 29\frac{R}{R_L} + 4\frac{R_L}{R} \tag{11.43}$$

The *condition for sustained oscillation* is given by Eq. (11.43).

Ordinarily, R_L is taken equal to R. Then Eqs. (11.41) and (11.43) give
$$f = \frac{1}{2\sqrt{10}\,\pi CR}, \tag{11.44}$$

and
$$h_{fe} = 56. \tag{11.45}$$

Hence, for sustained oscillations, the transistor should have an h_{fe} of 56 when $R_L = R$.

Phase-shift oscillators are commonly employed in the AF range. The frequency of oscillation here can be changed by using a ganged variable capacitor with three sections. To vary the frequency, the capacitances of the three sections are varied simultaneously. As conventional variable capacitors have capacitances in the range 50 pF to 500 pF, the frequency of oscillation can be altered in the ratio 10 : 1. For a greater variation of frequency, different resistors of resistances differing by a factor of 10 are employed.

11.4 TUNED-COLLECTOR OSCILLATOR

Fig. 11.2 (*a*) depicts the circuit diagram of a tuned-collector oscillator. The dc operating point of the transistor in the active region of its characteristics is determined by the resistances R_1, R_2 and R_E and the collector supply voltage $-V_{cc}$. The emitter by-pass capacitor C_E shunts the alternating components so that R_E does not appear in the ac equivalent circuit of Fig. 11.2(*b*).

As the transistor is in the CE configuration, it gives a phase shift of 180° between its input and output voltages. The transformer introduces another 180° phase shift needed for oscillation. The frequency-determining circuit is made up by the capacitor C together with the transformer primary inductance L. The LC tuned circuit connected to the collector accounts for the name 'tuned-collector oscillator'. The LC tuned circuit is called the *tank circuit* because this circuit determines or *stores* the frequency of oscillation. Furthermore, it is a reservoir of current, and stores electric and magnetic energies.

The resistance R_B is effectively in parallel with the resistance R_2. Since R_B is very large, its effect on the dc operating point is small. The main purposes served by R_B are to (*i*) reduce the loading of the collector circuit by the low input resistance of the transistor, (*ii*) introduce regenerative feedback just required to sustain oscillations, and (*iii*) decrease the input nonlinear distortion.

Analysis

The resistance R_B is chosen to be much larger than h_{ie} of the transistor and ωL_s, where ω is the angular frequency of oscillation and L_s is the transformer secondary inductance. As a result, the current I_1 in the transformer secondary is so small that it induces a negligible voltage in the transformer primary. In other words, the large value of R_B ensures that the LC tuned circuit is not significantly loaded. Therefore, the frequency of oscillation f [$= \omega/(2\pi)$] is approximately given by the natural resonant frequency of the LC tank circuit. Thus,

$$f \approx \frac{1}{2\pi\sqrt{LC}} \tag{11.2}$$

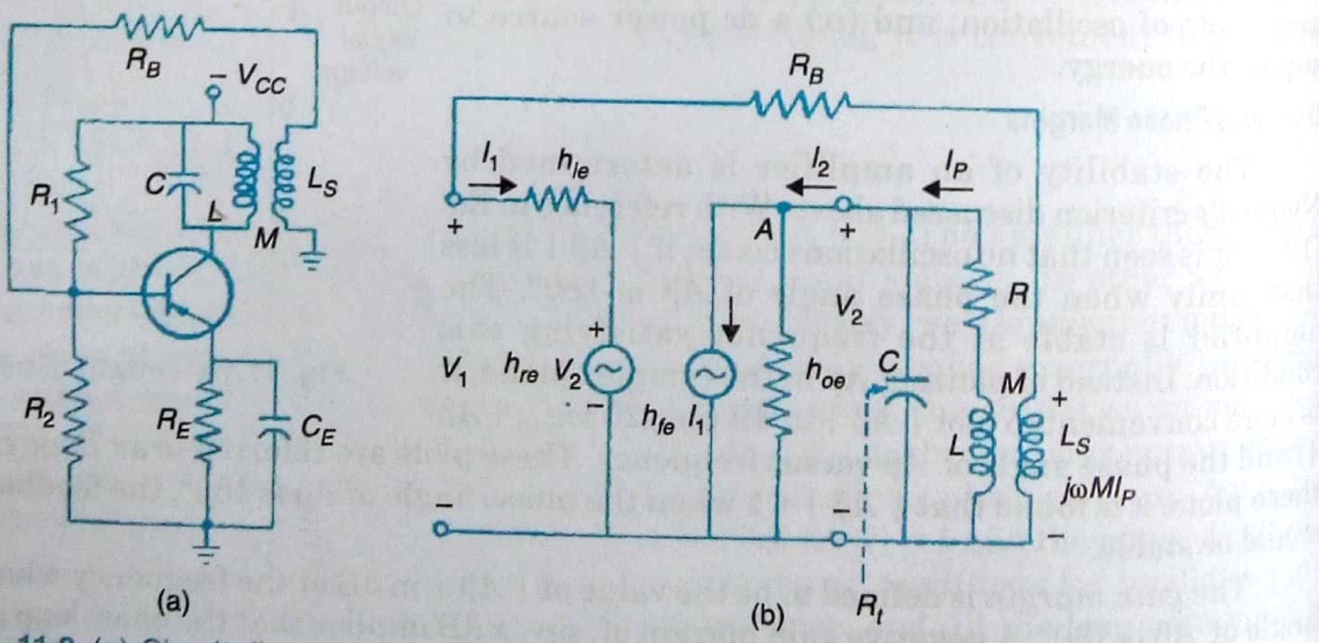

Fig. 11.2 (a) Circuit diagram of a tuned-collector oscillator (b) Its ac equivalent circuit.

The frequency of oscillation can be changed by varying either L or C, or both. Usually, the capacitor C is changed for practical convenience. With available values of L and C, tuned-collector oscillators generate oscillations in the RF range.

The resistance R of the transformer primary is small so that the tank circuit has a high Q. At resonance, the resistance offered by this circuit is*

$$R_t = \frac{L}{CR} \tag{11.3}$$

From the h-parameter AC equivalent circuit of Fig. 11.2(b), we obtain

$$V_2 = -I_2 R_t = -\frac{I_2 L}{CR} \tag{11.4}$$

Application of Kirchhoff's current law at the point A gives

$$I_2 = h_{oe} V_2 + h_{fe} I_1 \tag{11.5}$$

Substituting for V_2 from Eq. (11.4) into Eq. (11.5), we have

$$I_2 \left(1 + h_{oe} \frac{L}{CR}\right) = h_{fe} I_1 \tag{11.6}$$

The voltage induced in the transformer secondary is $j\omega MI_p$, where M is the mutual inductance between the primary and the secondary, and I_p is the primary current.

Since $R_B > |\omega L_s|$, the effective impedance in the secondary circuit is $R_B + h_{ie}$. Applying Kirchhoff's voltage law in the secondary circuit, we obtain

$$I_1 = \frac{j\omega MI_p - h_{re} V_2}{R_B + h_{ie}} \tag{11.7}$$

From Fig. 11.2(b), we also get

$$I_p = \frac{1/(j\omega C)}{R + j\omega L + \dfrac{1}{j\omega C}} I_2 = \frac{1}{j\omega CR} I_2 \qquad \left[\text{since } \omega = \frac{1}{\sqrt{LC}}\right]$$

$$= -j\omega \frac{L}{R} I_2 \tag{11.8}$$

Substituting for I_p and V_2 from Eqs. (11.8) and (11.4), respectively, in Eq. (11.7) gives

$$I_1 = \left(\frac{\omega^2 ML}{R} + \frac{h_{re} L}{CR}\right) I_2 /(R_B + h_{ie}). \tag{11.9}$$

Putting this value of I_1 in Eq. (11.6), we get

$$1 + h_{oe}\frac{L}{CR} = \frac{h_{fe}}{R_B + h_{ie}}\left(\omega^2 M + \frac{h_{re}}{C}\right)\frac{L}{R} \tag{11.10}$$

Substituting $\omega = 1/\sqrt{LC}$ in Eq. (11.10) gives

$$M = \frac{R_B}{h_{fe}}(CR + h_{oe} L) + CR\frac{h_{ie}}{h_{fe}} + L\frac{\Delta_{he}}{h_{fe}}, \tag{11.11}$$

where $\Delta_{he} = h_{ie} h_{oe} - h_{fe} h_{re}$. (11.12)

Equation (11.11) gives the value of M required for sustained oscillations.

11.5 HARTLEY OSCILLATOR

Fig. 11.3 shows the circuit diagram of a Hartley oscillator. The dc operating point in the active region of the characteristics is established by the resistors R_1, R_2 and R_E and the collector supply voltage $-V_{cc}$. The capacitor C_B is the blocking capacitor, while C_E is the bypass capacitor. Since the transistor is in the CE configuration, a phase shift of 180° exists between the input and output voltages. The output voltage appears across the tank circuit connected to the collector. The feedback voltage is a part of the output voltage, namely, V_1 appearing across the inductance L_1. The phase shift between the feedback voltage and the output voltage is 180°. Therefore, the total phase shift around the loop is 0° or 360°. The frequency-determining circuit is constituted by the capacitor C, and the inductors L_1 and L_2.

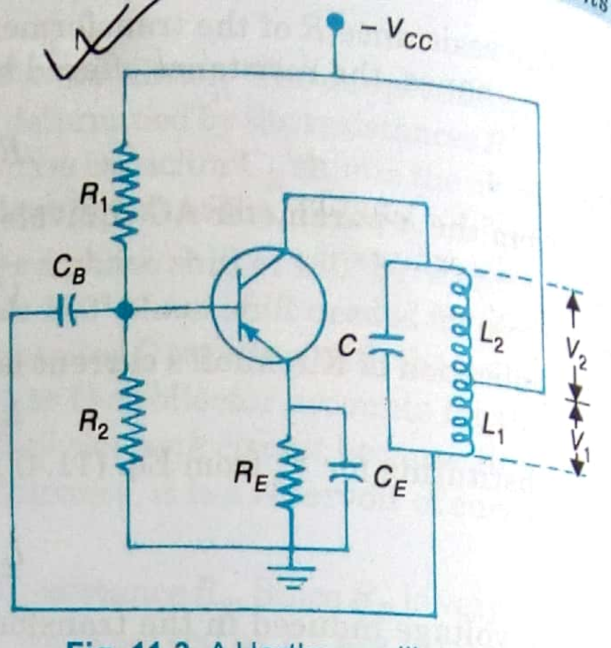

Fig. 11.3 A Hartley oscillator circuit.

Analysis

The ac operation of the circuit is not affected by the resistances R_1 and R_2 which are large. The resistance R_E is also ineffective in the ac behaviour due to the bypass capacitor C_E. The hybrid model ac equivalent circuit of the Hartley oscillator is depicted in Fig. 11.4(a).

We apply Thevenin's theorem looking towards the left at the terminals X, Y in the circuit of Fig. 11.4(a). The current source $h_{fe}\,I_1$ in shunt with the resistance $(1/h_{oe})$ is replaced by an equivalent Thevenin voltage source of generated voltage $(h_{fe}\,I_1/h_{oe})$ and internal resistance $(1/h_{oe})$. The equivalent circuit with this Thevenin representation is given in Fig 11.4(b). We neglect here, for simplicity, the mutual inductance between L_1 and L_2.

From Fig. 11.4(b), we obtain for the voltage across the terminals X, Y

$$V_2 = \frac{1}{h_{oe}}\,I_2 - \frac{h_{fe}}{h_{oe}}\,I_1 \tag{11.13}$$

The Kirchhoff voltage law equations for the loops (1), (2), and (3) in Fig. 11.4(b) are, respectively,

$$\left(h_{ie} + j\omega L_1 - \frac{h_{fe}\,h_{re}}{h_{oe}}\right) I_1 + \frac{h_{re}}{h_{oe}}\,I_2 - j\omega L_1\,I_3 = 0, \tag{11.14}$$

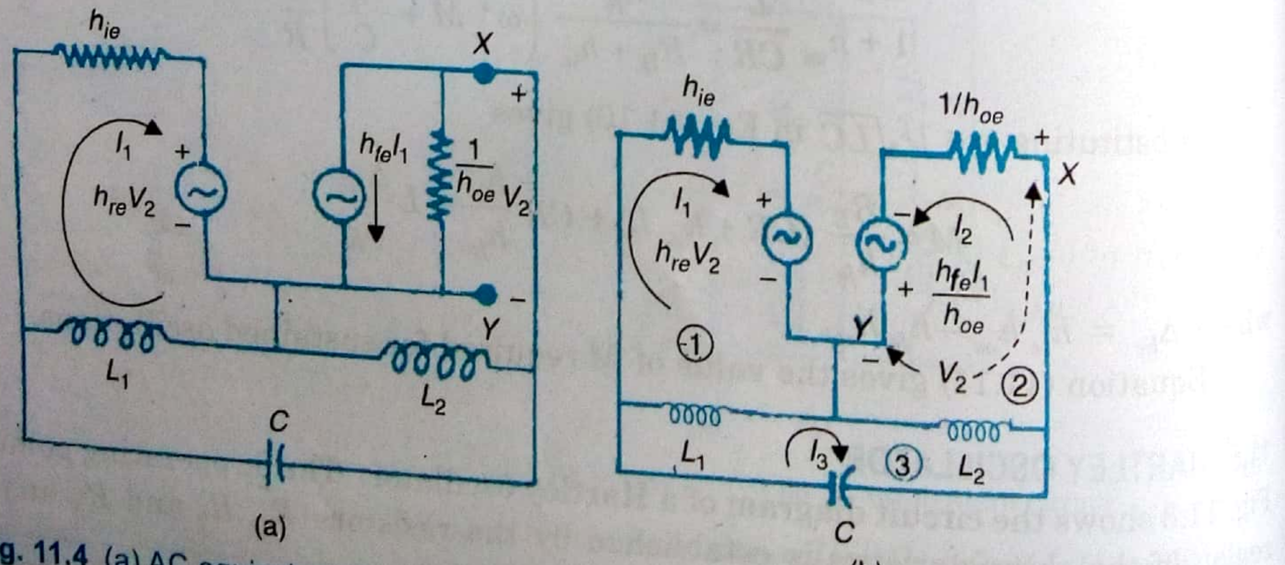

Fig. 11.4 (a) AC equivalent circuit of the Hartley oscillator. (b) AC equivalent circuit with the current source replaced by an equivalent voltage source.

$$-\frac{h_{fe}}{h_{oe}} I_1 + \left(\frac{1}{h_{oe}} + j\omega L_2\right) I_2 + j\omega L_2 I_3 = 0, \tag{11.15}$$

$$-j\omega L_1 I_1 + j\omega L_2 I_2 + [j\omega L_1 + j\omega L_2 - j/(\omega C)] I_3 = 0, \tag{11.16}$$

and

where ω is the angular frequency of oscillation. Since the currents I_1, I_2 and I_3 are nonvanishing, the determinant of the coefficients of I_1, I_2 and I_3 in Eqs. (11.14) through (11.16) must be zero. That is,

$$\begin{vmatrix} h_{ie} + j\omega L_1 - \dfrac{h_{fe}\, h_{re}}{h_{oe}} & \dfrac{h_{re}}{h_{oe}} & -j\omega L_1 \\ -\dfrac{h_{fe}}{h_{oe}} & \dfrac{1}{h_{oe}} + j\omega L_2 & j\omega L_2 \\ -j\omega L_1 & j\omega L_2 & j\omega L_1 + j\omega L_2 - \dfrac{j}{\omega C} \end{vmatrix} = 0 \tag{11.17}$$

At the frequency of oscillation, we approximately have

$$\omega (L_1 + L_2) - \frac{1}{\omega C} \approx 0 \tag{11.18}$$

Therefore, Eq. (11.17) yields

$$\left(h_{ie} + j\omega L_1 - \frac{h_{fe}\, h_{re}}{h_{oe}}\right) L_2^2 + \frac{h_{re}}{h_{oe}} L_1 L_2 - \frac{h_{fe}}{h_{oe}} L_1 L_2 + \left(\frac{1}{h_{oe}} + j\omega L_2\right) L_1^2 = 0 \tag{11.19}$$

Equating the real part of Eq. (11.19) to zero, we obtain

$$\Delta_{he}\, L_2^2 - (h_{fe} - h_{re}) L_1 L_2 + L_1^2 = 0 \tag{11.20}$$

here $\Delta_{he} = h_{ie}\, h_{oe} - h_{fe}\, h_{re}$.

As $h_{re} << 1$, Eq. (11.20) reduces to

$$\Delta_{he}\, L_2^2 - h_{fe}\, L_1 L_2 + L_1^2 = 0 \tag{11.21}$$

Solving Eq. (11.21) for L_2, we get

$$L_2 = \frac{h_{fe}\, L_1 \pm \sqrt{h_{fe}^2\, L_1^2 - 4\Delta_{he}\, L_1^2}}{2\Delta_{he}} \tag{11.22}$$

Again, $h_{fe}^2 >> 4\ \Delta_{he}$. Therefore, Eq. (11.22) further simplifies to

$$L_2 \approx \frac{h_{fe}\, L_1}{\Delta_{he}} \tag{11.23}$$

The *condition for sustaining the oscillations* is given by Eq. (11.23). Putting the imagi-
y part of Eq. (11.17) to zero, we obtain

$$\frac{\omega}{C} L_1 L_2 + \frac{h_{ie}}{h_{oe}}\left(\omega L_1 + \omega L_2 - \frac{1}{\omega C}\right) = 0$$

$$\omega^2 = \frac{1}{(h_{oe}\ L_1 L_2 / h_{ie}) + C(L_1 + L_2)} \tag{11.24}$$

The *frequency of oscillation* is

$$f = \frac{\omega}{2\pi} = \frac{1}{2\pi\ [(h_{oe}\ L_1 L_2 / h_{ie}) + C(L_1 + L_2)]^{1/2}} \tag{11.25}$$

In practice, $h_{oe} L_1 L_2/h_{ie} << C(L_1 + L_2)$, so that Eq. (11.25) gives

$$f \approx \frac{1}{2\pi\sqrt{C(L_1 + L_2)}} = \frac{1}{2\pi\sqrt{LC}}, \tag{11.26}$$

where $L (= L_1 + L_2)$ is the total inductance of the tank-circuit coil. *Thus the circuit gives oscillations at nearly the resonant frequency of the tank circuit.*

As mentioned earlier, we have neglected in the above analysis the mutual inductance M between L_1 and L_2. It can be shown that, when M is incorporated, L_1 and L_2 in Eqs. (11.23) and (11.26) are respectively replaced by $(L_1 + M)$ and $(L_2 + M)$.

Hartley oscillators usually generate oscillations in the RF range since the required values of L and C are convenient from practical considerations. The frequency of oscillation can be altered by varying L or C or both. The capacitance C can be easily changed, but smooth variations of L over a large range are inconvenient. So, for a smooth variation of frequency, capacitive tuning is preferred. Since variable capacitors have capacitances in the range 50 pF to 500 pF, Eq. (11.26) shows that the frequency of oscillation can be changed in the ratio 3 : 1.

11.6 COLPITTS OSCILLATOR

A Colpitts oscillator circuit is shown in Fig. 11.5(*a*). The dc operating point of the transistor in the active region of its characteristics is established by the resistors R_1, R_2, R_L, and R_E, and the supply voltage $-V_{cc}$. The capacitor C_B blocks the dc current flow from the collector to the base of the transistor through the coil of inductance L. The capacitor C_E is a bypass capacitor. The reactances of C_E and C_B are negligible at the frequency of oscillation. The inductance L and the capacitances C_1 and C_2 constitute the frequency-determining network.

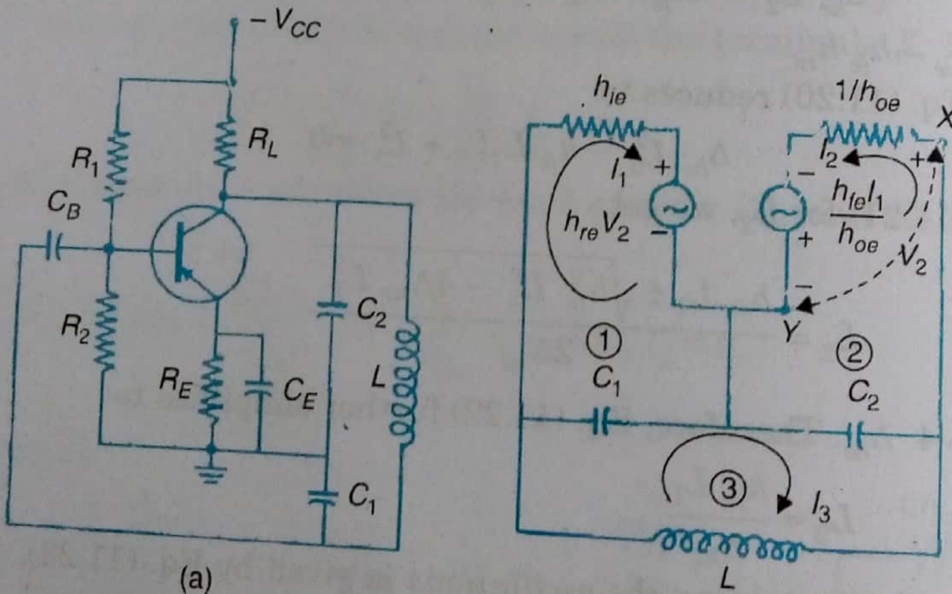

Fig. 11.5 (a) The circuit of a Colpitts oscillator, (b) Its AC equivalent circuit.

The transistor being in the CE configuration, introduces a phase shift of 180° between its input and output voltages. The voltage across the capacitor C_1, which is a fraction of the output voltage, is the feedback voltage. As the feedback voltage is 180° out of phase with the output voltage, the phase shift around the loop is 0° or 360°. It is noticed that Hartley and Colpitts oscillators are similar with the inductance and the capacitance interchanged.

Analysis

Since R_1 and R_2 are large resistances, they do not affect the ac operation of the circuit. Also, R_E, being shunted by C_E which bypasses the ac, is excluded from the AC equivalent circuit, shown in Fig. 11.5 (b). Also, R_L is omitted since it is much larger than $1/h_{oe}$, and the current source is transformed into a voltage source in the AC equivalent circuit to facilitate the analysis.

The potential difference between the points X, Y in Fig. 11.5(b) is

$$V_2 = \frac{1}{h_{oe}} I_2 - \frac{h_{fe}}{h_{oe}} I_1 \tag{11.27}$$

Applying Kirchhoff's voltage law to loops (1), (2) and (3) in Fig. 11.5 (b), we obtain respectively

$$\left(h_{ie} - \frac{h_{fe}\, h_{re}}{h_{oe}} - \frac{j}{\omega C_1} \right) I_1 + \frac{h_{re}}{h_{oe}} I_2 + \frac{j}{\omega C_1} I_3 = 0 \tag{11.28}$$

$$-\frac{h_{fe}}{h_{oe}} I_1 + \left(\frac{1}{h_{oe}} - \frac{j}{\omega C_2} \right) I_2 - \frac{j}{\omega C_2} I_3 = 0 \tag{11.29}$$

and
$$\frac{j}{\omega C_1} I_1 - \frac{j}{\omega C_2} I_2 + j\left(\omega L - \frac{1}{\omega C_1} - \frac{1}{\omega C_2} \right) I_3 = 0. \tag{11.30}$$

where Eq. (11.27) has been used. At the angular frequency ω of oscillation, the tuned circuit is nearly resonant so that $\omega L - \frac{1}{\omega C_1} - \frac{1}{\omega C_2} = 0$, Assuming $h_{re} << 1$, and $h^2_{fe} >> 4\Delta_{he}$, the condition for the sustained oscillations and the frequency of oscillation can be derived in the same manner as for a Hartley oscillator. Thus, equating the real part of the determinant of the coefficients of I_1, I_2 and I_3 in Eqs. (11.28) through (11.30) to zero, we get

$$\frac{C_1}{C_2} \approx \frac{h_{fe}}{\Delta_{he}}, \tag{11.31}$$

where $\Delta_{he} = h_{ie}\, h_{oe} - h_{fe}\, h_{re}$. The *condition for sustained oscillations* in the Colpitts oscillator is given by Eq. (11.31). As $h_{fe} \approx 50$ and $\Delta_{he} \approx 0.5$, Eq. (11.31) shows that $C_1/C_2 \approx 100$.

Putting the imaginary part of the determinant of the coefficients of I_1, I_2 and I_3 in Eqs. (11.28) through (11.30) to zero, we get

$$\omega^2 = \frac{h_{oe}}{h_{ie}\, C_1 C_2} + \frac{1}{LC_1} + \frac{1}{LC_2} \tag{11.32}$$

The *frequency of oscillation* is

$$f = \frac{\omega}{2\pi} = \frac{1}{2\pi} \left(\frac{h_{oe}}{h_{ie}\, C_1 C_2} + \frac{1}{LC_1} + \frac{1}{LC_2} \right)^{1/2} \tag{11.33}$$

In practice, $h_{oe} / (h_{ie}\, C_1\, C_2) << [1 / (LC_1) + 1 / (LC_2)]$. Hence, Eq. (11.33) reduces to

$$f \approx \frac{1}{2\pi\sqrt{LC_s}} \tag{11.34}$$

where $1/C_s = 1/C_1 + 1/C_2$. Clearly, C_s is the equivalent capacitance of C_1 and C_2 in series. Equation (10.34) shows that the frequency of oscillation is approximately the resonant frequency of the tank circuit.

Like other LC oscillators Colpitts oscillators also produce RF signals. In principle, Colpitts oscillators can be tuned by changing the inductance or the capacitance of the tank circuit.

11.10 MULTIVIBRATORS

The multivibrator is a relaxation oscillator generating nonsinusoidal waveforms. Basically, the multivibrator is a two-stage amplifier or oscillator operating in two *modes* or *states*. Each amplifier stage feeds back the other such that the active element of one stage is driven to saturation and the other to cut off. A new set of actions, producing the opposite effects, then follows. Thus the saturated stage becomes cutoff and the cutoff stage saturates. The operation of the multivibrator is based on the fact that *no two active elements have exactly identical characteristics*.

There are three classes of multivibrators: *astable, bistable*, and *monostable*. The *astable* or the *free-running* multivibrator alternates automatically and continuously between two states at a rate dictated by the circuit components. The *bistable* multivibrator, also referred to as an *Eccles-Jordan,* or a *flip-flop* multivibrator, operates in two states but requires an external trigger pulse to change from one state of operation to the other. The circuit continues in the existing state till a trigger pulse is applied. The *monostable* multivibrator, also known as a *single-shot*, a *single-swing* or a *one-shot* multivibrator, has a *normal* or a *quiescent state* and a *transient state*. When an external trigger pulse is applied, the circuit goes over to the transient state from the quiescent state, and automatically returns to its initial quiescent state after a time interval determined by the circuit components.

(1) Astable Multivibrator

An astable multivibrator using two similar transistors Q_1 and Q_2 is shown in Fig. 11.11. The circuit is basically two symmetrical CE amplifier stages, each supplying a feedback signal to the other. The transistor Q_1 is forward-biased by the supply voltage V_{CC} and the resistance R_3 while Q_2 is forward-biased by V_{CC} and the resistance R_2. The collector-emitter voltages of Q_1 and Q_2 are determined respectively by the load resistances R_1 and R_4 together with V_{CC}. The capacitor C_1 couples the output of Q_1 to the input of Q_2, whereas C_2 couples the output of Q_2 to the input of Q_1. In practice, $R_1 = R_4$, $R_2 = R_3$, and $C_1 = C_2$.

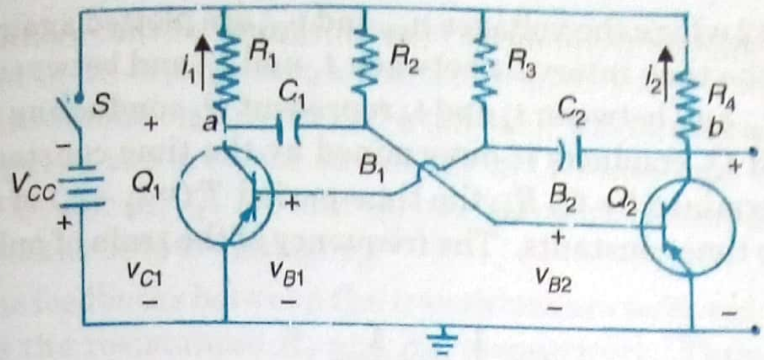

Fig. 11.11 An astable multivibrator.

Circuit Operation

When the switch S is closed, Q_1 and Q_2 are forward biased and so collector currents begin to flow in Q_1 and Q_2. As no two transistors have exactly the same characteristics, one of the transistors, say Q_1, will start conducting before the other and will carry a larger collector current. The currents i_1 and i_2 passing through R_1 and R_4, respectively, will produce a changing voltage (or signal) at the points a and b. With $i_1 > i_2$, an increase of i_1 enhances the voltage drop $i_1 R_1$, thus raising the potential of the point a. Due to the coupling capacitor C_1, the forward bias of Q_2 consequently decreases, so that i_2 falls. The decrease of i_2 reduces the voltage $i_2 R_4$ and drives the point b to a more negative potential. The capacitor C_2 couples this potential to the base of Q_1, so that the forward bias of Q_1 increases. This results in a further increase of i_1. This series of action continues, driving quickly Q_1 to saturation and Q_2 to cutoff.

When Q_1 is at saturation and Q_2 is cutoff, the collector voltage of Q_1 is $v_{C1} = V_{CE\,(sat)}$ and the collector voltage of Q_2 is $v_{C2} = -V_{CC}$. Now the capacitor C_1 charges through R_2, and so the base voltage of Q_2, i.e. v_{B2} falls exponentially towards $-V_{CC}$ with a time constant $R_2 C_1$. When v_{B2} attains the cutin voltage, Q_2 starts conducting and quickly goes into saturation. If I_s is the current through R_4 at saturation, v_{C2} increases by $I_s R_4$ to $V_{CE\,(sat)}$, as shown in Fig. 11.12. This increase of v_{C2} raises the voltage v_{B1} by $I_s R_4$ because of the capacitive coupling. The increase of v_{B1} cuts off Q_1 and its collector voltage drops towards $-V_{CC}$. The time interval in which Q_2 remains at saturation and Q_1 is cutoff, is determined by the time constant $C_2 R_3$, since C_2 now charges through R_3. The series of events continues as before, making Q_2 cutoff and Q_1 on again.

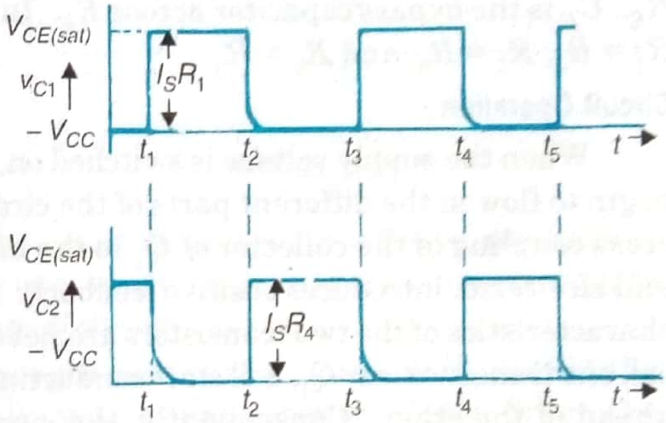

Fig. 11.12 Collector voltage waveforms for the astable multivibrator.

When Q_1 goes from saturation to cut off and Q_2 gets forward biased, the capacitor C_1 starts charging in the reverse direction by the battery voltage V_{CC} through the forward biased emitter-base junction of Q_2 and the resistance R_1. The charging time constant is $(R_1 + r_b)C_1$, where r_b is the base spreading resistance of Q_2. Usually, $R_2 >> R_1$ and so this time constant is much less than the time constant $R_2 C_1$. Because of the transient characterized by the time constant $(R_1 + r_b)\,C_1$, there is a slight rounding of the edge of the curve for v_{c1} in Fig. 11.12 as it jumps from $V_{CE\,(sat)}$ to $-V_{CC}$. A simlar rounding is found in the curve for v_{C2} due to the charging of the capacitor C_2. However, these effects are quite small.

Referring to Fig. 11.12 where the voltages v_{C1} and v_{C2} are plotted against time, Q_1 conducts and Q_2 remains cutoff in the time intervals between t_1 and t_2, and between t_3 and t_4. The time intervals between t_2 and t_3, and between t_4 and t_5 represent Q_2 conducting and Q_1 cutoff. Since the time interval in which Q_1 conducts is determined by the time constant $C_1 R_2$ and that in which Q_2 conducts is determined by $C_2 R_3$, the time period $T (= t_3 - t_1)$ of the collector voltage signal is fixed by these two time constants. The frequency of the train of pulses representing v_{C1} or v_{C2} is

$$f = \frac{1}{T} = \frac{1}{t_3 - t_1}.$$

The frequency can be varied from a few Hz to a few MHz by varying R_2, C_1, R_3, and C_2.

All the waveforms are inverted if *npn* transistors are used in place of *pnp* transistors in Fig. 11.11, because voltages and currents are negative of those for a circuit with *pnp* devices.

Astable multivibrators are widely used in pulse circuits. They find applications as generators of square waves. As no triggering signal is required, these multivibrators serve as sources of fast waveforms.

(2) Bistable Multivibrator

Fig. 11.13 shows the circuit diagram of a bistable multivibrator. Q_1 and Q_2 are two similar transistors having the collector resistances R_1 and R_4, respectively. The base-emitter bias of Q_1 is established by the voltage source V_{CC} and the voltage divider circuit formed by R_4, R_3, R_5 and R_E. Similarly, the base-emitter bias of Q_2 is determined by V_{CC}, R_1, R_2, R_6 and R_E. C_E is the bypass capacitor across R_E. In practice, $R_1 = R_4$, $R_5 = R_6$, and $R_2 = R_3$.

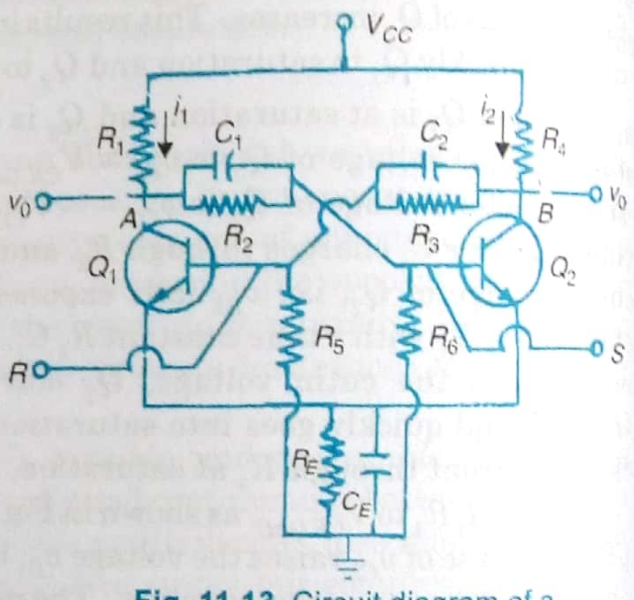

Fig. 11.13 Circuit diagram of a bistable multivibrator.

Circuit Operation

When the supply voltage is switched on, currents begin to flow in the different parts of the circuit. The cross coupling of the collector of Q_1 to the base of Q_2, and *vice-versa*, introduces positive feedback. Since the characteristics of the two transistors are never identical, one transistor, say Q_1, will start conducting slightly ahead of the other. Consequently, the current i_1 is larger than the current i_2. An increase of i_1 enhances the voltage drop across R_1, thus decreasing the voltage at the point A. So, the collector current of Q_2, i.e. i_2 will diminish, thereby increasing the voltage at B. This will raise the base voltage of Q_1. Consequently, the collector current of Q_1, i.e., i_1 will increase. The process cumulates rapidly, driving Q_1 to saturation and Q_2 to cutoff. If a positive trigger pulse of sufficient magnitude, as shown in Fig. 11.14 (*a*), is applied to the input S, Q_2 starts conducting and goes into saturation, whereas Q_1 is driven to cutoff. A negative trigger pulse [Fig. 11.14(*b*)] applied at R can bring about the same change.

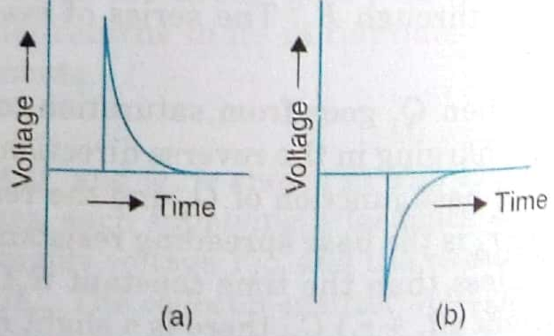

Fig. 11.14 (a) A positive trigger pulse (b) A negative trigger pulse.

Q_2 remains on and Q_1 cutoff indefinitely even when the trigger pulse is removed. Q_1 can go over to saturation and Q_2 to cutoff when a positive trigger pulse is applied to R or a negative trigger pulse to S. This condition will prevail even after the removal of the trigger. Thus, the application of trigger pulses can drive the circuit of Fig. 11.13 into any of the following two stable states: (i) Q_1 *on* and Q_2 *off*, (ii) Q_1 *off* and Q_2 *on*. Clearly, the spacing of the trigger pulses determines the shape of the output voltage v_0.

In some cases, the feedbacks between the transistors are achieved by connecting the capacitors C_1 and C_2 across the resistances R_2 and R_3, respectively. These capacitors assist in the abrupt transitions, thereby speeding up the switch-over of the flip-flop from one stable state to another. Hence, these capacitors are called *speed-up capacitors* or *commutating capacitors*.

Bistable multivibrators are used in digital circuits for counting and storing of binary information. They are extensively employed for the generation and processing of pulse type waveforms.

(3) Monostable Multivibrator

The circuit of a basic monostable multivibrator is illustrated in Fig. 11.15. In the absence of any trigger pulse v_i, the transistor Q_1 remains cutoff as it is reverse-biased by V_{BB} and R_5. However, V_{CC} and R_2 forward bias Q_2 and drives it into saturation. This is the *quiescent state* of the circuit.

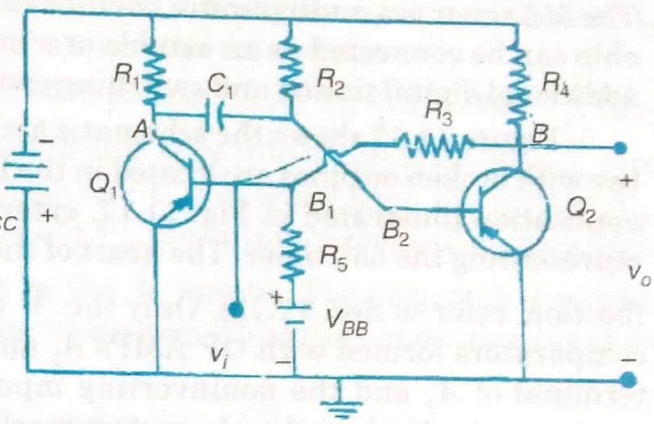

Fig. 11.15 The circuit of a monostable multivibrator.

Let a negative trigger pulse of sufficient magnitude be applied to the base of Q_1. Then Q_1 starts conducting and voltage of the point A rises. Due to the capacitive coupling of C_1, the base voltage of Q_2 also increases, so that the collector current of Q_2 falls. Consequently, the potential of the point B decreases. As the collector is connected to the base of Q_1 through R_3, the decrease of the potential at B reduces the base voltage of Q_1. So, the collector current of Q_1 increases. These events are repeated rapidly to drive Q_1 to saturation and Q_2 to cutoff. This is the *transient state* of the circuit.

When Q_1 is in saturation, the capacitor C_1 is charged towards V_{CC} through R_2, so that the potential of the base B_2 of Q_2 decreases exponentially with a time constant $C_1 R_2$. When the emitter-base voltage of Q_2 attains the cutin value, Q_2 conducts. The resulting voltage drop across R_4 increases the potential of the point B and hence of the point B_1. The collector current of Q_1 thus decreases, which, in turn, decreases the potential of A. The base voltage of Q_2 consequently falls due to the capacitive coupling. The collector current of Q_2, therefore, increases further. The rapid cumulative process drives Q_1 to cutoff and Q_2 to saturation. The circuit thus automatically returns to its quiescent state after a time interval determined by the time constant $C_1 R_2$. As Q_2 is driven to saturation, the capacitor C_1 is charged in the reverse direction by the battery voltage V_{CC} through the forward, biased base-emitter junction of Q_2 and the resistance R_1. The corresponding time constant is $C_1 (R_1 + r_b)$ where r_b is the base spreading resistance of Q_2. Usually, $R_2 >> R_1$, so that this time constant is much smaller than the time constant $C_1 R_2$. Hence the transients as Q_2 is driven to saturation die out quickly.

The trigger voltage v_i and the output voltage v_o are plotted as a function of time t in Fig. 11.16. The first trigger is applied at time t_1 so that v_o drops from $V_{CE\,(sat)}$ to $-V_{CC}$ at t_1. In the time interval from t_1 to t_2, Q_2 remains off and Q_1, on. At time t_2, Q_2 returns to its on state and Q_1 to its off state. Thus, at t_2, v_o rises from $-V_{CC}$ to $V_{CE\ (sat)}$. Another trigger at t_3 repeats the waveform. The time constant C_1R_2 determines the time interval $(t_2 - t_1)$ or $(t_4 - t_3)$.

If *npn* transistors are used instead of the *pnp* devices in Fig. 11.15, all waveforms are inverted.

Monostable multivibrators are extensively used in pulse circuits. Such multivibrators give a fixed time interval, the starting and the end of which are characterized by an abrupt change of a voltage waveform. These circuits are used to gate other circuits, and are then referred to as *gating circuits.*

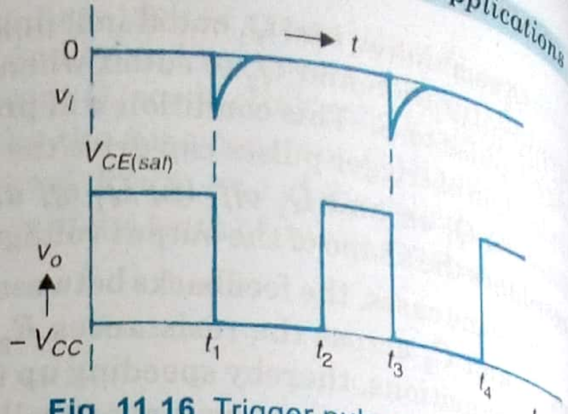

Fig. 11.16 Trigger pulses and the output voltage waveform for a monostable multivibrator.

11.11 The 555 IC Timer

The 555 timer is a multivibrator circuit available in integrated circuit (IC) forms. This versatile IC chip can be connected as an astable or a monostable multivibrator, and can be applied in many additional digital timing and switching operations.

Figure 11.17 shows the schematic architecture of the 555 timer. The components inside the box with broken outlines are located in the IC chip. For specific functions, such as for a monostable application illustrated in Fig. 11.17, external components are to be connected outside the box representing the 555 timer. The heart of the timer is a simple *RS* flip-flop. (For a discussion of RS flip-flop, refer to Sec. 17.13). Only the $\overline{Y}$ output is shown in Fig. 11.17. The open-loop analog comparators formed with OP AMPs A_1 and A_2 supply the S and R inputs. The inverting input terminal of A_1 and the noninverting input terminal of A_2 are connected to a voltage divider constructed with three equal resistors marked R_1. Each of the three resistances R_1 is 5 kΩ; hence the numbers 555 in the name of the IC. The voltages V_1 and V_2 are given by

$$V_1 = \frac{2V_{CC}}{3} \quad \text{and} \quad V_2 = \frac{V_{CC}}{3}.$$

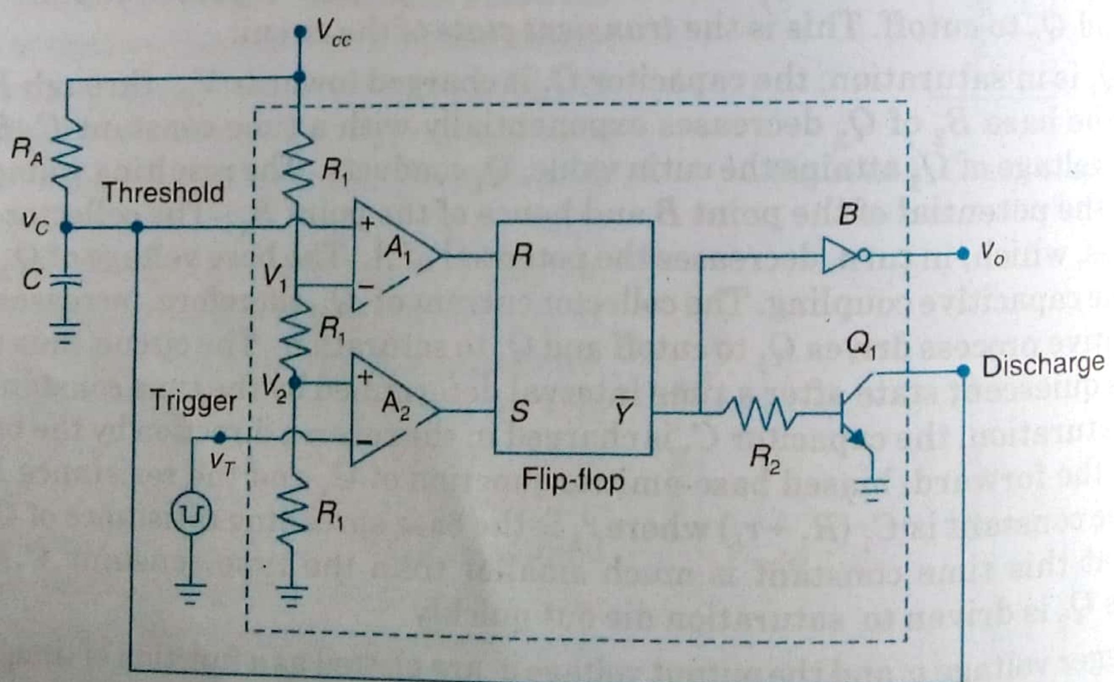

Fig. 11.17 Architecture of the 555 timer and its connections for monostable operation.

Astable Operation

The circuit connections to the 555 timer for the astable mode, shown in Fig. 11.19, give a periodic train of square waves indefinitely. Here the noninverting input of A_1 and the inverting input of A_2 are both connected to the capacitor C. Consider that the circuit is in a state with $\overline{Y}$ high, the transistor Q_1 in saturation, and the capacitor voltage v_C decaying towards V_{sat}. As v_C drops just below $V_2 = V_{CC}/3$, the output voltage of A_2 goes high, but the output voltage of A_1 remains low. Since R is low and S is high now, the flip-flop is driven into its *set* state with $\overline{Y}$ low. Hence, the transistor Q_1 goes to cutoff. Consequently, the capacitor starts to charge towards the supply voltage V_{CC} through the resistors R_A and R_B in series. When v_C exceeds V_2, the output voltage of A_2 is driven low. Since both S and R are now low, the set state of the flip-flop is retained.

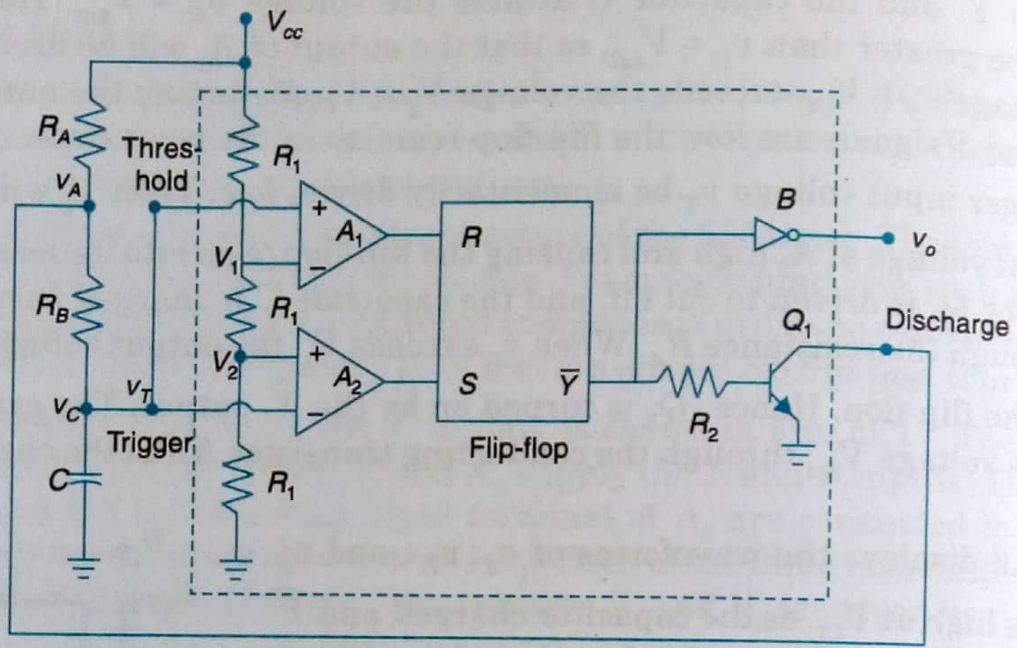

Fig. 11.19 The 555 timer and its connections for astable operation.

As charge continues to build up on the capacitor, v_C at some instant of time will reach the value $V_1 = 2V_{CC}/3$. Then the output voltage of A_1 goes high, and the flip-flop will be *reset*. Thus $\overline{Y}$ goes high, thereby turning Q_1 on. The capacitor now discharges through R_B and Q_1 towards the saturation voltage V_{sat} with a time constant R_BC approximately.

At some instant of time, v_C drops below $V_2 = V_{CC}/3$. Then the output voltage of A_2 becomes high, and the cycle of events repeats, generating a square-wave output voltage v_o.

Figure 11.20 depicts the waveforms of v_o and v_C. Let at time $t = 0$, $v_C = V_2 = V_{CC}/3$, forcing the input S to go high. The capacitor voltage v_C now starts to rise towards V_{CC} obeying the equation

$$v_C = \frac{V_{CC}}{3}\exp\left[-\frac{t}{C(R_A+R_B)}\right] + V_{CC}\left(1-\exp\left[-\frac{t}{C(R_A+R_B)}\right]\right)$$

Suppose that at $t = t_1$, $v_C = V_1 = \dfrac{2V_{CC}}{3}$. Hence

$$\frac{2V_{CC}}{3} = \frac{V_{CC}}{3}\exp\left[-\frac{t_1}{C(R_A+R_B)}\right] + V_{CC}\left(1-\exp\left[-\frac{t_1}{C(R_A+R_B)}\right]\right)$$

or,

$$t_1 = C(R_A+R_B)\ln 2 = 0.69\, C\,(R_A+R_B)$$

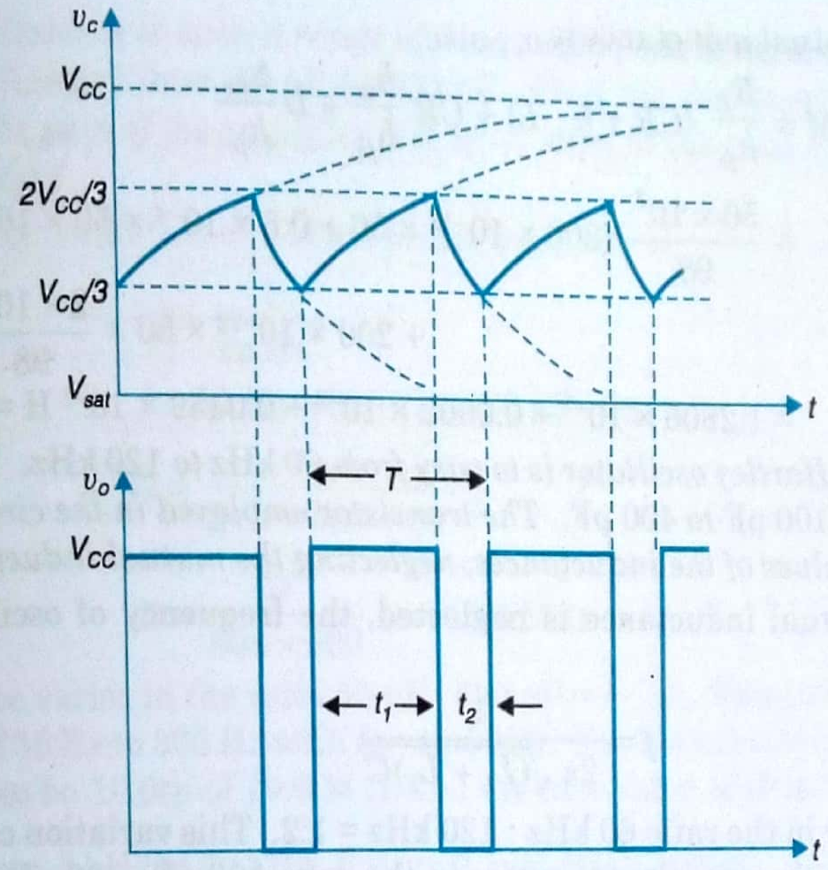

Fig. 11.20 Variations of v_0 and v_c with time, t.

At $t = t_1$, the flop-flop is reset, and C discharges through R_B and conducting Q_1 towards V_{sat}, obeying the following equation approximately :

$$v_C = \frac{2V_{CC}}{3}\exp\left(-\frac{t}{CR_B}\right) + V_{sat}\left[1-\exp\left(-\frac{t}{CR_B}\right)\right]$$

In this equation, the former time t_1 is designated as $t = 0$. The equation is approximate since it is assumed that the transistor voltage stays at a constant value V_{sat} during the entire discharging interval. If $v_C = V_2 = V_{CC}/3$ at $t = t_2$ (with the instant t_1 taken as zero), we have

$$\frac{V_{CC}}{3} = \frac{2V_{CC}}{3}\exp\left(-\frac{t_2}{CR_B}\right) + V_{sat}\left[1-\exp\left(-\frac{t_2}{CR_B}\right)\right]$$

or,

$$t_2 = CR_B \ln\frac{2V_{CC}/3 - V_{sat}}{V_{CC}/3 - V_{sat}}$$

If $V_{sat} << V_{CC}$, we obtain $t_2 \simeq CR_B \ln 2 = 0.69\, CR_B$.

The time period of the square waves for v_0 is

$$T = t_1 + t_2 = 0.69\, C\,(R_A + R_B) + 0.69\, CR_B = 0.69\, C\,(R_A + 2R_B).$$