5.1 Introduction:
To amplify the selective range of frequencies, the resistive load $R_C$ is replaced by a tuned circuit. The tuned circuit is capable of amplifying a signal over a narrow band of frequencies centered at $f_r$. The amplifiers with such a tuned circuit as a load are known as tuned amplifiers.

![Fig. 3.1 Tuned circuit](image)

The Fig. 3.1 shows the tuned parallel LC circuit which resonates at a particular frequency. The resonance frequency and impedance of tuned circuit is given as,

$$f_r = \frac{1}{2\pi\sqrt{LC}} \quad \ldots (1)$$

$$Z_r = \frac{L}{CR} \quad \ldots (2)$$

The response of tuned amplifiers is maximum at resonant frequency and it falls sharply for frequencies below and above the resonant frequency, as shown in the Fig. 3.2. In the Fig. 3.2, 3 dB bandwidth is denoted as $B$ and 30dB bandwidth is denoted as $S$. The ratio of the 30dB bandwidth ($S$) to the 3 dB bandwidth ($B$) is known as Skirt selectivity.

At resonance, inductive and capacitive effects of tuned circuit cancel each other. As a result, circuit is like resistive and $\cos \varphi = 1$ i.e. voltage and current are in phase. For frequencies above resonance, circuit is capacitive and for frequencies below resonance, circuit is inductive. Since tuned circuit is purely resistive at resonance it can be used as a load for amplifier.

5.1.1 Coil Losses:
As shown in the Fig. 3.1, the tuned circuit consists of a coil. Practically, coil is not purely inductive. It consists of few losses and they are represented in the form of leakage resistance $R_L$ with the inductor. The total loss of the coil is comprised of copper loss, eddy current loss and hysteresis loss. The copper loss at low frequencies is equivalent to the d.c. resistance of the coil. Copper loss is inversely proportional to frequency. Therefore, as frequency increases, the
copper loss decreases. Eddy current loss in iron and copper coil are due to currents flowing within the copper or core based by induction. The result of eddy currents is a loss due to heating within the inductors copper or core. Eddy current losses are directly proportional to frequency. Hysteresis loss is proportional to the area enclosed by the hysteresis loop and to the rate at which this loop is transversed. It is a function of signal level and increase with frequency. Hysteresis loss is however independent of frequency.

As mentioned earlier, the total losses in the coil or inductor is represented by inductance in series with leakage resistance of the coil. It is shown in the Fig.3.3

5.1.2 Q Factor:

Quality Factor (Q) is important characteristics of inductor. The Q is the ratio of reactance to resistance and therefore it is unitless. It is the measure of how 'pure' or 'real' an inductor is (i.e. the inductor consists only reactance). The higher the Q of an inductor, the fewer losses there are in the inductor. The Q factor also can be defined as the measure of efficiency with which inductor can store the energy. The dissipation factor (D) that can be referred to as the total loss within a component is defined as 1/Q. The fig. 3.4 shows the quality factor equations for series and parallel circuits and its relation with dissipation factor.

![Fig. 3.4 Quality factor equations](image)

**Quality factor equation**

\[ Q = \frac{1}{D} = \frac{\omega L_s}{R_s} = \frac{R_p}{\omega L_p} \]

5.1.3 Unloaded and Loaded Q:

Unloaded Q is the ratio of stored energy to dissipated energy in a reactor or resonator. The unloaded Q or Q_U of an inductor or capacitor is X/R_s, where X represents the reactance and R_s represents the series resistance. The loaded Q or Q_L of a resonator is determined by how tightly the resonator is coupled to its terminations.

![Fig. 3.5 Tuned load circuit](image)

Let us consider the tuned load circuit as shown in the Fig. 3.5. Here, L and C represent tank circuit. The internal circuit losses of inductor are represented by R_o and R_c represents the coupled in load.
For this circuit we can write,
\[ R_o = \frac{\omega_0 L}{Q_U} \] and \[ R_C = \frac{\omega_0 L}{Q_L} \]

where \( Q_U \) is unloaded \( Q \) and \( Q_L \) is loaded \( Q \).
The circuit efficiency for the above tank circuit is given as,
\[ \eta = \frac{I^2 R_C}{I^2 (R_C + R_o)} = \frac{Q_U}{Q_U + Q_L} \times 100\% \]

From above equation, it can be easily realized that for high overall power efficiency, the coupled-in load \( R_C \) should be large in comparison to the internal circuit losses represented by \( R_o \) of the inductor.
The quality factor \( Q_L \) determines the 3 dB bandwidth for the resonant circuit. The 3 dB bandwidth for the resonant circuit is given as,
\[ BW = \frac{f_r}{Q_L} \]

Where \( f_r \) represents the centre frequency of a resonator and \( BW \) represents the bandwidth.
If \( Q \) is large, bandwidth is small and circuit will be highly selective. For small \( Q \) values, bandwidth is high and selectivity of the circuit is lost, as shown in the Fig. 3.6.

![Fig. 3.6 Variation of 3dB bandwidth with variation in quality factor](image)

Thus in tuned amplifier \( Q \) is kept as high as possible to get the better selectivity. Such tuned amplifiers are used in communication or broadcast receivers where it is necessary to amplify only selected band of frequencies.

5.1.4 Requirements of Tuned Amplifier:
The basic requirements of Tuned Amplifiers are:
1. The amplifier should provide selectivity of resonant frequency over a very narrow band.
2. The signal should be amplified equally well at all frequencies in the selected narrow band.
3. The tuned circuit should be so mounted that it can be easily tuned. If there is more than one circuit to be tuned, there should be an arrangement to tune all circuit simultaneously.
4. The amplifier must provide the simplicity in tuning of the amplifier components to the desired frequency over a considerable range of band of frequencies.

5.1.5 Classification of Tuned Amplifier:
We know that, multistage amplifiers are used to obtain large overall gain. The cascaded stages of multistage tuned amplifiers can be categorized as given below:
1. Single Tuned Amplifiers
2. Double Tuned Amplifiers
3. Stagger Tuned Amplifiers
These amplifiers are further classified according to coupling used to cascade the stages of multistage amplifier. They are,

1. Capacitive Coupled
2. Inductive Coupled
3. Transformer Coupled

5.2 Small Signal Tuned Amplifier:

A common emitter amplifier can be converted into a single tuned amplifier by including a parallel tuned circuit as shown in the Fig. 3.7. The biasing components are not shown for simplicity.

Before going to study the analysis of this amplifier, the practical assumptions to simplify the analysis are:

Assumptions:
1. \( R_L << R_C \)
2. \( r_{bb'} = 0 \)

With these assumptions, the simplified equivalent circuit for a single tuned amplifier is shown in Fig 3.8.

Fig. 3.7 Single tuned transistor amplifier

Fig. 3.8 Equivalent circuit of single tuned amplifier

Where \( C_{eq} = C' + C_{b'e} + (1 + g_m R_L) C_{b'c} \) and \( C' \) is the external capacitance used to tune the circuit, \( (1 + g_m R_L) C_{b'c} \) is the Miller Capacitance. \( r_s \) represents the losses in the coil. The series RL circuit in Fig. 3.7 is replaced by the equivalent RL circuit in Fig 3.8 assuming coil losses are low over the frequency band of interest, i.e. the coil Q is high.
The conditions for equivalence are most easily established by equating the admittances of the two circuits shown in the Fig. 3.9.

\[
Q_c = \frac{\omega L}{r_c} \gg 1
\]

\[Y_1 = \frac{1}{r_c + j\omega L} = \frac{r_c - j\omega L}{r_c^2 + \omega^2 L^2}
\]

\[= \frac{r_c}{r_c^2 + \omega^2 L^2} - \frac{j\omega L}{r_c^2 + \omega^2 L^2}
\]

\[Y_1 = \frac{r_c}{\omega^2 L^2} + \frac{1}{j\omega L}
\]

\[\therefore \omega L, r_c \text{ from equation (1)}
\]

\[Y_2 = \frac{1}{R_p} + \frac{1}{j\omega L}
\]

\[\therefore \text{Therefore, equating } Y_1 \text{ and } Y_2 \text{ we get,}
\]

\[\frac{r_c}{\omega^2 L^2} + \frac{1}{j\omega L} = \frac{1}{R_p} + \frac{2}{j\omega L}
\]

\[\Rightarrow \frac{1}{R_p} = \frac{r_c}{\omega^2 L^2}
\]

\[= \frac{r_c^2}{r_c\omega^2 L^2} = \frac{1}{r_c Q_c^2}
\]

\[\therefore R_p = r_c Q_c^2 = \omega L Q_c
\]

\[\therefore \omega L = Q_c r_c \text{ from equation (1)} \quad \ldots (2)
\]

Looking at Fig. 3.8 we have,
\[R = r_i \parallel R_p \parallel r_{v_e}
\]

\[\therefore \quad \ldots (3)
\]

The current gain of the amplifier is then
\[A_1 = \frac{-g_m R}{1 + j(\omega RC - R/\omega L)} \quad \frac{-g_m R}{1 + j \omega_c RC(\omega_c - \omega_c / \omega)}
\]

where \[\omega_c^2 = \frac{1}{LC}
\]

\[\ldots (4)
\]
We define the $Q$ of the tuned circuit at the resonant frequency $w_0$ to be

$$Q_i = \frac{R}{\omega_0 L} = \omega_o RC$$  \hspace{1cm} \ldots \text{(5)}$$

$$A_i = \frac{-g_m R}{1+jQ_i(\omega/\omega_0-\omega_o/\omega)}$$

At $\omega = \omega_0$, gain is maximum and it is given as,

$$A_{i(\text{max})} = -g_m R$$ \hspace{1cm} \ldots \text{(5)}$$

The Fig. 3.10 shows the gain versus frequency plot for the single tuned amplifier. It shows the variation of the magnitude of the gain as a function of frequency.

![Gain versus frequency plot](image)

**Fig. 3.10 Gain versus frequency for single tuned amplifier**

At 3 dB frequency,

$$|A_i| = \frac{g_m R}{\sqrt{2}} \hspace{1cm} \ldots \text{(7)}$$

At 3 dB frequency

$$1 + jQ_i[(\omega/\omega_o)-(\omega_o/\omega)] = \sqrt{2}$$

$$1 + Q_i^2\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)^2 = 2 \hspace{1cm} \ldots \text{(8)}$$

This equation is quadratic in $w^2$ and has two positive solutions, $w_H$ and $w_L$. After solving equation (8), 3 dB bandwidth is given below:

$$BW = f_H - f_L = w_0 / 2\pi Q_i = 1 / 2\pi RC \quad \text{(since } Q_i = w_0 RC) \hspace{1cm} \ldots \text{ (9)}$$

$$BW = \frac{1}{2\pi RC}$$
Example 3.1: Design a single tuned amplifier for following specifications:

1. Centre frequency = 500 kHz
2. Bandwidth = 10 kHz

Assume transistor parameters: $g_m = 0.04 S$, $h_{fe} = 100$, $C_{bc} = 1000 pF$ and $C_{bc} = 100 pF$. The bias network and the input resistance are adjusted so that $r_i = 4 k\Omega$ and $R_L = 510 \Omega$.

Solution: From equation (9) we have,

$$BW = \frac{1}{2\pi RC}$$

$$\therefore \quad RC = \frac{1}{2\pi BW} = \frac{1}{2\pi \times 10 \times 10^3}$$

$$= 15.912 \times 10^{-6}$$

From equation (3) we have,

$$R = r_i \parallel R_p \parallel r_{bc'}$$

where

$$r_i = 4 k\Omega$$

$$r_{bc'} = \frac{h_{fe}}{g_m} = \frac{100}{0.04} = 2500 \Omega$$

$$R_p = Q_c \omega_L = \frac{Q_c}{\omega_c C}$$

$$\therefore \quad R = 4 \times 10^3 \parallel 2500 \parallel \frac{Q_c}{\omega_c C}$$

$$C = \frac{1}{2\pi \times 10 \times 10^3 \times R}$$

$$\therefore \quad C = \frac{1}{2\pi \times 10 \times 10^3 \times \left[ 4 \times 10^3 \parallel 2500 \parallel \frac{Q_c}{\omega_c \times C} \right]}$$

The typical range for $Q_c$ is 10 to 150. However, we have to assume $Q$ such that value of $C_p$ should be positive. Let us assume $Q = 100$.

$$\therefore \quad C = \frac{1}{2\pi \times 10 \times 10^3 \times \left[ 1538.5 \parallel \frac{1}{2\pi \times 5000 \times C} \right]}$$

$$= \frac{1}{2\pi \times 10 \times 10^3 \times \left[ \frac{1}{1538.5} + \frac{1}{2\pi \times 5000 \times C} \right]}$$
Solving for $C$ we get,

$$C = 0.02 \, \mu F$$

We have,

$$C = C' + C_{V_e} + (1 + g_m R_L) C_{V_e}$$

$$\therefore \quad C' = C - [C_{V_e} + (1 + g_m R_L) C_{V_e}]$$

$$= 0.02 \times 10^{-6} - [1000 \times 10^{-12} + (1 + 0.04 \times 510) \times 100 \times 10^{-12}]$$

$$\therefore \quad C' = 0.01686 \, \mu F$$

We have,

$$\omega_0^2 = \frac{1}{LC}$$

$$\therefore \quad L = \frac{1}{\omega_0^2 C} = \frac{1}{(2\pi \times 500 \times 10^{-3})^2 \times 0.02 \times 10^{-6}}$$

$$= 5 \, \mu H$$

From equation (2) we have,

$$R_p = \omega L Q_c = 2\pi \times 500 \times 10^3 \times 5 \times 10^{-6} \times 100$$

$$= 1570 \, \Omega$$

$$\therefore \quad R = r_i \parallel R_p \parallel r_{V_e}$$

$$= 4 \times 10^3 \parallel 1570 \parallel 2500$$

$$= 777 \, \Omega$$

We have mid frequency gain as,

$$A_{i_{\text{max}}} = -g_m R = (-0.04) (777) = -31$$

### 5.3 Single Tuned FET Amplifier:

The Fig.3.11 shows the single tuned FET amplifier.

![Diagram](image.png)

**Fig. 3.11 Single tuned FET amplifier**

The equivalent circuit for the given amplifier is as shown in the Fig. 3.12.
The voltage gain is given by,
\[ A_v = -a g_m \left( r_{ds} || R_L \right) \left[ (r_i || R_p) / r_i \right] \]  
\[ \text{where} \]
\[ C_i = a^2 \left( C_{gs} + C_{gd} \left[ 1 + g_m \left( r_{ds} || R_i \right) \right] \right) \]  
\[ Q_i = f_0 \left( r_i || R_p \right) \left( C' + C_i \right) \]  
\[ \omega_0^2 = \frac{1}{L \left( C' + C_i \right)} \]  

At centre frequency, i.e., at \( \omega = \omega_0 \) gain is
\[ A_{v_{\text{max}}} = -a g_m \left( r_{ds} || R_L \right) \frac{R_p}{r_i + R_p} \]  

The 3 dB bandwidth is given by,
\[ BW = \frac{1}{2 \pi \left( r_i || R_p \right) \left( C' + C_i \right)} \]  

### 5.4 Single Tuned Capacitive Coupled Amplifier:

Single Tuned multistage amplifier circuit uses one parallel tuned circuit as a load in each stage with tuned circuits in all stages tuned to same frequency. Fig 3.13 shows a typical single tuned amplifier in CE configuration. As shown in Fig.3.13, tuned circuit formed by L and C acts as collector load and resonates at frequency of operation. Resistors \( R_1, R_2 \) and \( R_E \) along with capacitor \( C_E \) provides self bias for the circuit.
Fig. 3.14 Equivalent circuit of single tuned amplifier

The Fig. 3.14 shows the equivalent circuit for a single tuned amplifier using hybrid \( \pi \) parameters.

As shown in the Fig. 3.14, \( R_i \) is the input resistance of the next stage and \( R_o \) is the output resistance of the current generator \( g_m V_{be} \). The reactances of the bypass capacitor \( C_e \) and the coupling capacitors \( C_C \) are negligibly small at the operating frequency and hence these elements are neglected in the equivalent circuit shown in the Fig. 3.14.

The equivalent circuit shown in Fig. 3.14 can be simplified by applying Miller's theorem. Fig. 3.15 shows the simplified equivalent circuit for a single tuned amplifier.

Fig. 3.15 Simplified equivalent circuit for single tuned amplifier

Here \( C_i \) and \( C_{eq} \) represent input and output circuit capacitances, respectively. They can be given as,

\[
C_i = C_{b'e} + C_{b'e} (1 - A) \quad \text{where } A \text{ is the voltage gain of the amplifier.} \quad \ldots(1)
\]

\[
C_{eq} = C_{b'e} \left( \frac{A - 1}{A} \right) + C \quad \text{where } C \text{ is the tuned circuit capacitance.} \quad \ldots(2)
\]

The \( g_{ce} \) is represented as the output resistance of the current generator \( g_m V_{be} \).

\[
g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} = h_{oe} = \frac{1}{R_o} \quad \ldots(3)
\]
The series RL circuit is represented by its equivalent parallel circuit. The conditions for equivalence are most easily established by equating the admittances of the two circuits shown in Fig. 3.16.

Admittance of the series combination of RL is given as,

\[ Y = \frac{1}{R+j\omega L} \]

Multiplying numerator and denominator by \( R - j\omega L \) we get,

\[ Y = \frac{R - j\omega L}{R^2 + \omega^2 L^2} = \frac{R}{R^2 + \omega^2 L^2} - \frac{j\omega L}{R^2 + \omega^2 L^2} \]

\[ = \frac{R}{R^2 + \omega^2 L^2} - \frac{j\omega^2 L}{\omega(R^2 + \omega^2 L^2)} \]

\[ = \frac{1}{R_p} - \frac{1}{j\omega L_p} \]

where \( R_p = \frac{R^2 + \omega^2 L^2}{R} \) \( \ldots \) \( (4) \)

and \( L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L} \) \( \ldots \) \( (5) \)

**Centre frequency**

The centre frequency or resonant frequency is given as,

\[ f_r = \frac{1}{2\pi\sqrt{L_p C_{eq}}} \] \( \ldots \) \( (6) \)

where \( L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L} \)

and \( C_{eq} = C_n + \left( \frac{A_1}{A} \right) + C \)

\[ = C_o + C \] \( \ldots \) \( (7) \)

Therefore, \( C_{eq} \) is the summation of transistor output capacitance and the tuned circuit capacitance.

**Quality factor Q**

The quality factor \( Q \) of the coil at resonance is given by,

\[ Q_r = \frac{\omega L}{R} \] \( \ldots \) \( (8) \)
where \( \omega_c \) is the centre frequency or resonant frequency.

This quality factor is also called unloaded Q. But in practice, transistor output resistance and input resistance of next stage act as a load for the tuned circuit. The quality factor including load is called as loaded Q and it can be given as follows:

The Q of the coil is usually large so that \( \omega L \gg R \) in the frequency range of operation.

From equation (4) we have,

\[ R_p = \frac{R^2 + \omega^2 L^2}{R} = R + \frac{\omega^2 L^2}{R} \]

As \( \frac{\omega^2 L^2}{R} \gg L \), \( R_p = \frac{\omega^2 L}{R} \) \( \cdots (9) \)

From equation (5) we have,

\[ L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + L \]

\[ \approx L \quad \therefore \omega L \gg R \quad \cdots (10) \]

From equation (9), we can express \( R_p \) at resonance as,

\[ R_p = \frac{\omega^2 L^2}{R} \]

\[ = \omega_c Q_r \cdot L \quad \therefore Q_r = \frac{\omega_c L}{R} \quad \cdots (11) \]

Therefore, \( Q_r \) can be expressed in terms of \( R_p \) as,

\[ Q_r = \frac{R_p}{\omega_c L} \quad \cdots (12) \]

The effective quality factor including load can be calculated looking at the simplified equivalent output circuit for single tuned amplifier.

![Fig. 3.17 Simplified output circuit for single tuned amplifier](image)

**Effective quality factor**

\[ Q_{eff} = \frac{\text{Susceptance of inductance } L \text{ or capacitance } C}{\text{Conductance of shunt resistance } R_t} \]

\[ = \frac{R_t}{\omega_c L} \text{ or } \omega_c C_{eq} R_t \quad \cdots (13) \]
Voltage gain ($A_v$)

The voltage gain for a single tuned amplifier is given by,

$$A_v = -g_m \frac{r_{ve}}{r_{be} + r_{ve}} \times \frac{R_t}{1 + 2jQ_{\text{eff}} \delta}$$

where

$$R_t = R_o || R_p || R_i$$

$$\delta = \text{Fraction variation in the resonant frequency}$$

$$A_v \text{ (at resonance)} = -g_m \frac{r_{ve}}{r_{be} + r_{ve} R_t}$$

$$\therefore \left| \frac{A_v}{A_v \text{ (at resonance)}} \right| = \frac{1}{\sqrt{1 + (28Q_{\text{eff}})^2}} \quad \ldots (14)$$

3 dB bandwidth

The 3 dB bandwidth of a single tuned amplifier is given by,

$$\Delta f = \frac{1}{2\pi R_t C_{\text{eq}}}$$

$$= \frac{\omega_r}{2\pi Q_{\text{eff}}} \quad \therefore Q_{\text{eff}} = \omega_r R_t C_{\text{eq}} \quad \ldots (15)$$

$$= \frac{f_r}{Q_{\text{eff}}} \quad \therefore \omega_r = 2\pi f_r \quad \ldots (16)$$

Example 3.2: A single tuned RF amplifier uses a transistor with an output resistance of 50 $\Omega$, output capacitance of 15 pF and input resistance of next stage is 20 k$\Omega$. The tuned circuit consists of 47 pF capacitance in parallel with series combination of 1 $\mu$H inductance and 2 $\Omega$ resistance. Calculate

i) Resonant frequency

ii) Effective quality factor

iii) Bandwidth of the circuit

Solution: i) Resonant frequency $f_r$ is given as,

$$f_r = \frac{1}{2\pi \sqrt{1/C_{\text{eq}}}}$$

$$= \frac{1}{2\pi \sqrt{1/\mu H \times (15 \text{ pF} + 47 \text{ pF})}}$$

$$= 20.2 \text{ MHz}$$

ii) Effective quality factor is given as,

$$Q_{\text{eff}} = \frac{\omega_r C_{\text{eq}}}{R_t}$$

$$= 2\pi f_r C_{\text{eq}} \times (R_o || R_p || R_i)$$
where \[ R_p = \frac{\omega^2 L^2}{R} = \frac{(2\pi \times 20.2 \times 10^6)^2 (1 \times 10^{-6})^2}{2} \]
\[ = 8054 \ \Omega \]

\[ \therefore \quad Q_{eff} = 2 \pi \times 20.2 \times 10^6 \times (15 \text{ pF} + 47 \text{ pF}) \times (50 \text{ K} || 8.054 \text{ K} || 20 \text{ K}) \]
\[ = 40.52 \]

iii) Bandwidth of the circuit is given as,
\[ BW = \frac{f_r}{Q_{eff}} = \frac{20.2 \times 10^6}{40.52} \]
\[ = 498.5 \text{ kHz} \]

**Example 3.3:** A single tuned transistor amplifier is used to amplify modulated RF carrier of 600 kHz and bandwidth of 15 kHz. The circuit has a total output resistance, \( R_t = 20 \text{ k}\Omega \) and output capacitance \( C_o = 50 \text{ pF} \). Calculate values of inductance and capacitance of the tuned circuit.

**Solution:**

\[ f_r = 600 \text{ kHz} \]
\[ BW = 15 \text{ kHz} \]
\[ R_t = 20 \text{ k}\Omega \]
\[ C_o = 50 \text{ pF} \]

\[ \therefore \quad C_{eq} = (50 \text{ pF} + C) \]
\[ Q_{eff} = \frac{f_r}{BW} = \frac{600 \text{ kHz}}{15 \text{ kHz}} \]
\[ = 40 \]

i) We know that,
\[ Q_{eff} = \frac{\omega \cdot C_{eq} \cdot R_t}{\omega \cdot R_t} \]
\[ \therefore \quad C_{eq} = \frac{Q_{eff}}{\omega \cdot R_t} = \frac{40}{2\pi \times 600 \times 10^3 \times 20 \times 10^3} \]
\[ = 530.5 \text{ pF} \]
\[ C_{eq} = (50 \text{ pF} + C) \]
\[ \therefore \quad C = 530.5 \text{ pF} - 50 \text{ pF} \]
\[ = 480.5 \text{ pF} \]

ii) We know that,
\[ f_r = \frac{1}{2\pi \sqrt{L \cdot C_{eq}}} \]
\[ \therefore \quad L = \frac{1}{(2\pi f_r)^2 \cdot C_{eq}} = \frac{1}{(2\pi \times 600 \times 10^3)^2 \times 530.5 \times 10^{-12}} \]
\[ = 132.6 \text{ \mu H} \]
5.5 Double Tuned Amplifier:

Fig 3.18 Double tuned amplifier

Fig 3.18 shows double tuned RF amplifier in CE configuration. Here, voltage developed across tuned circuit is coupled inductively to another tuned circuit. Both tuned circuits are tuned to the same frequency. The double tuned circuit can provide a bandwidth of several percent of the resonant frequency and gives steep sides to the response curve. Let us analyze the double tuned circuit.

Analysis:
The Fig. 3.19 (a) shows the coupling section of a transformer coupled double tuned amplifier. The Fig. 3.19 (b) shows the equivalent circuit for it, in which transistor is replaced by the current source with its output resistance ($R_o$). The $C_1$ and $L_1$ are the tank circuit components of the primary side. The resistance $R_1$ is the series resistance of the inductance $L_1$. Similarly on the second side $L_2$ and $C_2$ represent the tank circuit components of the secondary side and $R_2$ represents resistance of the inductance $L_2$. The resistance $R_i$ represents the input resistance of the next stage.

The Fig. 3.19 (c) shows the simplified equivalent circuit for the Fig. 3.19 (b). In simplified equivalent circuit the series and parallel resistances are combined into series elements. Referring equation (9) we have,

$$R_F = \frac{\omega^2 L^2}{R} \text{ i.e. } R = \frac{\omega^2 L^2}{R_F}$$

where $R$ represents series resistance and $R_p$ represents parallel resistance.

Therefore we can write,

$$R_{i1} = \frac{\omega^2 L_1^2}{R_o} + R_1$$

$$R_{i2} = \frac{\omega^2 L_2^2}{R_i} + R_2$$

In the simplified circuit the current source is replaced by voltage source, which is now in series with $C_1$. It also shows the effect of mutual inductance on primary and secondary sides.

We know that, $Q = \frac{\omega L}{R}$

Therefore, the $Q$ factors of the individual tank circuits are

$$Q_1 = \frac{\omega L_1}{R_{i1}} \text{ and } Q_2 = \frac{\omega L_2}{R_{i2}}$$

... (1)

Usually, the $Q$ factors for both circuits are kept same. Therefore, $Q_1 = Q_2 = Q$ and the resonant frequency $\omega^2 = 1/L_1C_1 = 1/L_2C_2$.

Looking at Fig. 3.19 (c), the output voltage can be given as,

$$V_o = -\frac{j}{\omega C_2}I_2$$

... (2)
To calculate $V_o/V_1$ it is necessary to represent $I_2$ in terms of $V_1$. For this we have to find the transfer admittance $Y_T$. Let us consider the circuit shown in Fig. 3.20. For this circuit, the transfer admittance can be given as,

![Circuit Diagram]

\[ Y_T = \frac{I_2}{V_1} = \frac{I_2}{I_1 Z_{11}} = \frac{A_i}{Z_{11}} \]

\[ = \frac{Z_f}{Z_f^2 - Z_1 (Z_o + Z_L)} \]

where

\[ Z_{11} = \frac{V_1}{I_1} = Z_1 - \frac{Z_f^2}{Z_o + Z_L} \] and

\[ A_i = \frac{I_2}{I_1} = \frac{-Z_f}{Z_o + Z_L} \]

The simplified equivalent circuit for double tuned amplifier is similar to the circuit shown in Fig. 3.20 with

\[ Z_f = j \omega M \]

\[ Z_i = R_{11} + j \left( \omega L_i - \frac{1}{\omega C_i} \right) \]

\[ Z_o + Z_L = R_{22} + j \left( \omega L_2 - \frac{1}{\omega C_2} \right) \]

The equations for $Z_f$, $Z_i$ and $Z_o + Z_L$ can be further simplified as shown below.

\[ Z_i = j \omega M = j \omega k \sqrt{L_1 L_2} \]

where, $k$ is the coefficient of coupling.
Multiplying numerator and denominator by \( \omega, L_1 \) for \( Z_i \) we get,

\[
Z_i = \frac{R_{11} \omega, L_1}{\omega, L_1} + j \omega, L_1 \left( \frac{\omega L_1}{\omega L_1} - \frac{1}{\omega C_1 \omega L_1} \right)
\]

\[
= \frac{\omega L_1}{Q} + j \omega, L_1 \left( \frac{\omega L_1}{\omega L_1} - \frac{\omega L_1}{\omega L_1} \right)
\]

\[
= \frac{\omega L_1}{Q} + j \omega, L_1 \left( \frac{\omega}{\omega} - \frac{\omega}{\omega} \right)
\]

\[
= \frac{\omega L_1}{Q} + (1 + j 2 Q \delta)
\]

\[
Z_o + Z_L = R_{22} + j \left( \omega L_2 - \frac{1}{\omega C_2} \right)
\]

By doing similar analysis as for \( Z_i \) we can write,

\[
Z_o + Z_L = \frac{\omega L_2}{Q} + (1 + j 2 Q \delta)
\]

Then

\[
Y_T = \frac{Z_i}{Z_i^2 - Z_i (Z_o + Z_L)} = \frac{1}{Z_i - Z_i (Z_o + Z_L) / Z_i}
\]

\[
Y_T = \frac{1}{j \omega, k \sqrt{L_1 \frac{L_2}{Q}} \left[ \frac{\omega L_1}{Q} (1 + j 2 Q \delta) \right] \left[ \frac{\omega L_2}{Q} (1 + j 2 Q \delta) \right]}
\]

\[
Y_T = \frac{\frac{k Q^2}{\omega, \sqrt{L_1 \frac{L_2}{Q}} \left[ 4 Q \delta - j \left( 1 + k^2 Q^2 - 4 Q^2 \delta^2 \right) \right]}}{... (3)}
\]

Substituting value of \( I_2 \), i.e. \( V_i \times Y_T \) we get,

\[
V_o = -j \frac{j \beta_m V_i}{\omega C_2 \omega, C_1 \left[ 4 Q \delta - j \left( 1 + k^2 Q^2 - 4 Q^2 \delta^2 \right) \right]} \frac{k Q^2}{\omega, \sqrt{L_1 \frac{L_2}{Q}} \left[ 4 Q \delta - j \left( 1 + k^2 Q^2 - 4 Q^2 \delta^2 \right) \right]}
\]

\[
\therefore V_i = \frac{j \beta_m V_i}{\omega C_1}
\]
\[ A_v = \frac{V_o}{V_i} = g_m \omega^2 L_1 L_2 \frac{kQ^2}{\omega_r \sqrt{L_1 L_2} \left[ \frac{1}{4Q^2 - j (1 + k^2 Q^2 - 4Q^2 \delta^2)} \right]} \]

or \[ \frac{1}{\omega_r C} = \omega_r L \]

\[ = \left[ \frac{g_m \omega_r \sqrt{L_1 L_2} kQ^2}{4Q^2 - j (1 + k^2 Q^2 - 4Q^2 \delta^2)} \right] \]

... (4)

Taking the magnitude of equation (4) we have,

\[ |A_v| = g_m \omega_r \sqrt{L_1 L_2} \frac{Q}{\sqrt{1 + k^2 Q^2 - 4Q^2 \delta^2 + 16Q^2 \delta^2}} \]

... (5)

The Fig. 3.21 shows the universal response curve for double tuned amplifier plotted with \( kQ \) as a parameter.

The frequency deviation \( \delta \) at which the gain peaks occur can be found by maximizing equation (4), i.e.

\[ 4Q^2 - j (1 + k^2 Q^2 - 4Q^2 \delta^2) = 0 \]

... (6)
As shown in the Fig. 3.22, two gain peaks in the frequency response of the double tuned amplifier can be given at frequencies:

\[ f_1 = f_r \left(1 - \frac{1}{2Q} \sqrt{k^2Q^2 - 1}\right) \]

\[ f_2 = f_r \left(1 + \frac{1}{2Q} \sqrt{k^2Q^2 - 1}\right) \] ... (7)

![Fig. 3.22](image)

At \( k^2Q^2 = 1 \), i.e. \( k = \frac{1}{Q} \), \( f_1 = f_2 = f_r \). This condition is known as critical coupling. For values of \( k < 1/Q \), the peak gain is less than maximum gain and the coupling is poor.

At \( k > 1/Q \), the circuit is overcoupled and the response shows the double peak. Such double peak response is useful when more bandwidth is required.

The gain magnitude at peak is given as,

\[ |A_p| = \frac{g_m Q_0 \sqrt{L_1 L_2}}{2} kQ \] ... (8)

And gain at the dip at \( \delta = 0 \) is given as,

\[ |A_d| = |A_p| \frac{2 kQ}{1 + k^2Q^2} \] ... (9)

The ratio of peak gain and dip gain is denoted as \( \gamma \) and it represents the magnitude of the ripple in the gain curve.

\[ \gamma = \frac{|A_p|}{|A_d|} = \frac{1 + k^2Q^2}{2 kQ} \] ... (10)

Using quadratic simplification and choosing positive sign we get,

\[ kQ = \gamma + \sqrt{\gamma^2 - 1} \] ... (11)

The bandwidth between the frequencies at which the gain is \( |A_d| \) is the useful bandwidth of the double tuned amplifier. It is given as,

\[ BW = 2 \delta' = \sqrt{2} (f_2 - f_1) \] ... (12)
We know that, the 3 dB bandwidth for single tuned amplifier is $2f_r / Q$. Therefore, the 3 dB bandwidth provided by double tuned amplifier ($3.1f_r / Q$) is substantially greater than the 3 dB bandwidth of single tuned amplifier.

Compared with a single tuned amplifier, the double tuned amplifier
1. Possesses a flatter response having steeper sides.
2. Provides larger 3 dB bandwidth.
3. Provides large gain-bandwidth product.

5.6 Effect of cascading Single Tuned Amplifier on Bandwidth:
In order to obtain a high overall gain, several identical stages of amplifiers can be used in cascade. The overall gain is the product of the voltage gains of the individual stages. Let us see the effect of cascading of stages on bandwidth.

Consider $n$ stages of single tuned direct coupled amplifiers connected in cascade. We know that the relative gain of a single tuned amplifier w.r.t. the gain at resonant frequency $f_r$ is given from equation (14).

\[
\frac{A_v}{A_v \text{ (at resonance)}} = \frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}}
\]

Therefore, the relative gain of $n$ stage cascaded amplifier becomes

\[
\left(\frac{A_v}{A_v \text{ (at resonance)}}\right)^n = \left(\frac{1}{\sqrt{1 + (2\delta Q_{eff})^2}}\right)^n = \frac{1}{\left[1 + (2\delta Q_{eff})^2\right]^\frac{n}{2}}
\]

The 3 dB frequencies for the $n$ stage cascaded amplifier can be found by equating

\[
\frac{A_v}{A_v \text{ (at resonance)}} = \frac{1}{\sqrt{2}}
\]

\[
\frac{A_v}{A_v \text{ (at resonance)}} = \frac{1}{\left[1 + (2\delta Q_{eff})^2\right]^{\frac{n}{2}}} = \frac{1}{\sqrt{2}}
\]

\[
\left[1 + (2\delta Q_{eff})^2\right]^\frac{n}{2} = 2^\frac{1}{n}
\]

\[
\left[1 + (2\delta Q_{eff})^2\right]^n = 2
\]

\[
1 + (2\delta Q_{eff})^2 = 2^\frac{1}{n}
\]

\[
2\delta Q_{eff} = \pm \sqrt{2^\frac{1}{n} - 1}
\]
Substituting for $\delta$, the fractional frequency variation, i.e.

$$\delta = \frac{\omega - \omega_r}{\omega_r} = \frac{f - f_r}{f_r}$$

$$2 \left( \frac{f - f_r}{f_r} \right) Q_{eff} = \pm \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

$$2 (f - f_r) Q_{eff} = \pm f_r \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

$$f - f_r = \pm \frac{f_r}{2Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

Let us assume $f_1$ and $f_2$ are the lower 3 dB and upper 3 dB frequencies, respectively. Then we have

$$f_2 - f_r = + \frac{f_r}{2Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

$$f_r - f_1 = + \frac{f_r}{2Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

The bandwidth of $n$ stage identical amplifiers is given as,

$$BW_n = f_2 - f_1 = (f_2 - f_r) + (f_r - f_1)$$

$$= \frac{f_r}{2Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1} + \frac{f_r}{2Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

$$= \frac{f_r}{Q_{eff}} \sqrt{\frac{1}{2^\frac{2}{n}} - 1}$$

$$= BW_1 \sqrt{\frac{1}{2^\frac{2}{n}} - 1} \quad \ldots (1)$$

where $BW_1$ is the bandwidth of single stage and $BW_n$ is the bandwidth of $n$ stages.

Example 3.4 : The bandwidth for single tuned amplifier is 20 kHz. Calculate the bandwidth if such three stages are cascaded. Also calculate the bandwidth for four stages.

Solution : i) We know that,

$$BW_n = BW_1 \sqrt{\frac{1}{2^\frac{2}{n}} - 1} = 20 \times 10^3 \times \sqrt{\frac{1}{2^\frac{2}{3}} - 1}$$

$$= 10.196 \text{ kHz}$$

ii) $BW_n = 20 \times 10^3 \times \sqrt{\frac{1}{2^\frac{2}{4}} - 1} = 8.7 \text{ kHz}$

The above example shows that bandwidth decreases as number of stages increase.

5.7 Effect of cascading Double Tuned Amplifiers on Bandwidth:

When a number of identical double tuned amplifier stages are connected in cascade, the overall bandwidth of the system is thereby narrowed and the steepness of the sides of the response is increased, just as when single tuned stages are cascaded. The quantitative relation
between the 3 dB bandwidth of n identical double tuned critically coupled stages compared with the bandwidth $\Delta_2$ of such a system can be shown to be 3 dB bandwidth for n identical stages double tuned amplifiers is,

$$\Delta_2 \times \left(1 - \frac{1}{2^n} \right)^{\frac{1}{4}}$$

where $\Delta_2$ is the 3 dB bandwidth of single stage double tuned amplifier.

The above equation assumes that the bandwidth $\Delta_2$ is small compared with the resonant frequency.

Example 3.5: The bandwidth for double tuned amplifier is 20 kHz. Calculate the bandwidth if such three stages are cascaded.

Solution: We know that for double tuned cascaded stages,

$$BW_n = BW_1 \times \left(\frac{2^{1/n} - 1}{2^{1/n} - 1} \right)^{\frac{1}{4}}$$

$$= 20 \times \left(\frac{2^{1/3} - 1}{2^{1/3} - 1} \right)^{\frac{1}{4}}$$

$$= 14.28 \text{ kHz}$$

Example 3.6: A three stage double tuned amplifier system is to have a half power BW of 20 kHz centred on a centre frequency of 450 kHz. Assuming that all stages are identical, determine the half power bandwidth of single stage. Assume that each stage couple to get maximum flatness.

Solution: We get maximum flat response when each stage is critically coupled. When stages are critically coupled we have

$$BW_n = BW_1 \times (2^{1/n} - 1)^{\frac{1}{4}}$$

$$BW_n = \frac{BW_n}{(2^{1/n} - 1)^{\frac{1}{4}}}$$

For $n = 3$

$$BW_n = \frac{BW_n}{(2^{1/3} - 1)^{\frac{1}{4}}}$$

$$= 20 \times 10^3$$

$$= 28.01 \text{ kHz}$$

5.8 Staggered Tuned Amplifier:

Double tuned amplifiers give greater 3 dB bandwidth having steeper sides and flat top. But alignment of double tuned amplifier is difficult. To overcome this problem, two single tuned cascaded amplifiers having certain bandwidth are taken and their resonant frequencies are so that they are separated by an amount equal to the bandwidth of each stage. Since the resonant frequencies are displaced or staggered, they are known as stagger tuned amplifiers. The advantage of stagger tuned amplifier is to have better flat, wideband characteristics in contrast with a very sharp, narrow band characteristic of synchronously tuned circuits (tuned to same resonant frequencies). Fig 3.23 shows the relation of amplification characteristics of individual stages in a staggered pair to the overall amplification of the two stages.
The overall response of the two stage stagger tuned pair is compared in the Fig. 3.24 with the corresponding individual single tuned stages having same resonant circuits. Looking at Fig 3.24, it can be seen that staggering reduces the total amplification of the centre frequency to 0.5 of the peak amplification of the individual stage and at the centre frequency each stage has amplification that is 0.707 of the peak amplification of the individual stage. Thus the equivalent voltage amplification per stage of the staggered pair is 0.707 times as great as when the same two stages are used without staggering. However, the half power (3 dB) bandwidth of the staggered pair is $\sqrt{2}$ times as great as the half power (3 dB) bandwidth of on individual single tuned stage. Hence the equivalent gain bandwidth product per stage of a stagger tuned pair is $0.707 \times \sqrt{2} = 1.00$ times that of the individual single tuned stages.

The stagger tuned idea can be easily extended to more stages. In case of three stages staggering, the first tuning circuit is tuned to a frequency lower than centre frequency while the third circuit is tuned to higher frequency than centre frequency. The middle tuned circuit is tuned at exact centre frequency.

![Fig. 3.23 Overall response of staggered pair](image1)

![Fig. 3.24 Response of individually tuned and staggered tuned pair](image2)
Analysis:

Form equation (14), the gain of the single stage tuned amplifier can be written as,

$$A_v(\text{at resonance}) = \frac{1}{1 + \frac{2jQ_{eq}X}{\delta}}$$

$$= \frac{1}{1 - jX} \text{ where } X = 2\ Q_{eff} \delta$$

Since in stagger tuned amplifiers, the two single tuned cascaded amplifiers with separate resonant frequencies are used, we can assume that the one stage is tuned to the frequency $f_r + \delta$ and other stage is tuned to the frequency $f_r - \delta$. Therefore, we have,

$$f_{r1} = f_r + \delta$$

and

$$f_{r2} = f_r - \delta$$

According to these tuned frequencies, the selectivity functions can be given as,

$$\frac{A_v}{A_v(\text{at resonance})_1} = \frac{1}{1 + j(X+1)}$$

and

$$\frac{A_v}{A_v(\text{at resonance})_2} = \frac{1}{1 + j(X-1)}$$

The overall gain of these two stages is the product of individual gains of the two stages.

$$\therefore \frac{A_v}{A_v(\text{at resonance})_{\text{cascaded}}} = \frac{A_v}{A_v(\text{at resonance})_1} \times \frac{A_v}{A_v(\text{at resonance})_2}$$

$$= \frac{1}{1 + j(X+1)} \times \frac{1}{1 + j(X-1)}$$

$$= \frac{1}{2 + 2jX} \times \frac{1}{2 - 2jX}$$

$$= \frac{2 - 2j^2X^2}{4 - 4j^2X^2}$$

$$= \frac{1}{4\ +\ (2X)^2}$$

$$= \frac{1}{\sqrt{4\ -\ 4X^2 + X^4 + 4X^2}} = \frac{1}{\sqrt{4 + X^4}}$$

Substituting the value of $X$, we get,

$$\left|\frac{A_v}{A_v(\text{at resonance})}_{\text{cascaded}}\right| = \frac{1}{\sqrt{4 + (2Q_{eff}\delta)^2}} = \frac{1}{\sqrt{4 + 16Q_{eff}^4\delta^2}}$$

$$= \frac{1}{2\sqrt{1 + 4Q_{eff}^4\delta^2}}$$

5.9 Large Signal Tuned 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 operation of class B and class C amplifiers are non-linear since the amplifying elements remain cut-off during a part of the input signal cycle. The non-linearity generates harmonics of the signal frequency at the output of the amplifier. In the push-pull arrangement
where the bandwidth requirement is no limited, these harmonics can be eliminated or reduced. When a narrow bandwidth is desired, a resonant circuit is employed in class B and class C tuned RF power amplifiers to eliminate the harmonics.

5.9.1 Class B Tuned Amplifier:

The Fig 3.25 shows the class B tuned amplifier. It works with a single transistor by sending half sinusoidal current pulses to the load. The transistor is biased at the edge of the conduction. Even though the input is half sinusoidal, the load voltage is sinusoidal because a high Q RLC tank shunts harmonics to ground. The negative half is delivered by the RLC tank. The Q factor of the tank needs to be large enough to do this. This is analogous to pushing someone in swing. We only need to push in one direction, and the reactive energy stored will swing the person back in the reverse direction.

5.9.2 Class C Tuned Amplifier:

The amplifier is said to be class C amplifier, if the Q point and the input signal are selected such the output signal is obtained for less than a half cycle, for a full cycle.

Due to such a selection of the Q point, transistor remains active, for less than a half cycle. Hence only that part is reproduced at the output. For remaining cycle of the input cycle, the transistor remains cut-off and no signal is produced at the output.

The current and voltage waveforms for a class C operation are shown in the Fig. 3.26. Looking at Fig 3.26, it is apparent that the total angle during which current flows is less than 180°. This angle is called the conduction angle, θc.
Fig. 3.26 Waveform representing class C operation

Fig 3.27 shows the class C tuned amplifier. Here a parallel resonant circuit as load impedance. As collector current flows for less than half a cycle, the collector current consists of a series of pulses with the harmonics of the input signal. A parallel tuned circuit acting as load impedance is tuned to the input frequency. Therefore, it filters the harmonic frequencies and produces a sine wave output voltage consisting of fundamental component of the input signal.
A class C tuned amplifier can be used as a frequency multiplier if the resonant circuit is tuned to a harmonic of the input signal.

\[ f_r = \frac{1}{2\pi \sqrt{LC}} \]  

... (1)

**Resonant frequency**

Here, class C amplifier is used with parallel tuned circuit. Therefore, the output voltage is maximum at the resonant frequency. The resonant frequency for parallel tuned circuit is given as.

**Power gain**

Power gain is defined as,

\[ G = \frac{P_{out}}{P_{in}} \]

... (2)

In words, the power gain equals the a.c. output power divided by the a.c. input power.

**Output power**

If we measure the output voltage of Fig. 3.27 in r.m.s. volts, the output power is given by

\[ P_{out} = \frac{V_{rms}^2}{R_L} \]

... (3)

Usually we measure the output voltage in peak to peak volts \( (V_{pp}) \) with an oscilloscope and

\[ V_{pp} = 2\sqrt{2} V_{rms} \]

\[ V_{rms} = \frac{V_{pp}}{2\sqrt{2}} \]

.: 

Substituting value of \( V_{rms} \) we get,

\[ P_{out} = \left(\frac{V_{pp}}{2\sqrt{2}}\right)^2 \times \frac{V_{pp}}{8R_L} \]

... (4)
A.C. collector resistance

Any inductor has a series resistance $R_s$ as indicated in Fig. 3.29. The $Q$ of the inductor is defined as,

$$Q_L = \frac{X_L}{R} = \frac{\omega L}{R} = \frac{\omega L}{R_{eq}}$$  \hspace{1cm} \ldots (5)

where

$Q_L$ = Quality factor of coil

$X_L$ = Inductive reactance

$R$ = Coil resistance

![Fig. 3.29](image)

As shown in the Fig. 3.29, the series resistance of the coil can be replaced by a parallel resistance $R_p$. This equivalent parallel resistance can be given as,

$$R_p = Q_L \omega R_L$$  \hspace{1cm} \ldots (6)

At resonance, $X_L$ cancels $X_C$, leaving only $R_p$ in parallel with $R_L$. Therefore, the a.c. resistance seen by the collector at resonance is:

$$r_C = R_p || R_L$$  \hspace{1cm} \ldots (7)

$\therefore$ $Q$ of the overall circuit is given by,

$$Q = \frac{r_C}{\omega L}$$  \hspace{1cm} \ldots (8)

Transistor power dissipation

Fig. 3.30 shows the ideal collector-emitter voltage in a class C transistor amplifier. In Fig. 3.30, the maximum output is given by,

$$V_{pp\text{ (max)}} = 2 V_{CC}$$  \hspace{1cm} \ldots (9)

Since the maximum voltage is approximately $2 V_{CC}$, the transistor must have $V_{CEO}$ rating greater than $2 V_{CC}$.

![Fig. 3.30](image)
Fig. 3.30 (b) shows the collector current for a class C amplifier. Typically, the conduction angle $\phi$ is much less than 180°. The power dissipation of the transistor depends on the conduction angle. It increases when conduction angle increases as shown in Fig. 3.30 (c). The maximum power dissipation for class C amplifier can be given as,

$$P_{D_{\text{max}}} = \frac{V_{p_{\text{r(max)}}}^2}{40r_c} \quad \ldots (10)$$

where $r_c = \text{A.C. collector resistance}$

Under normal condition, conduction angle will be less than 180° and the transistor power dissipation will be less than $\frac{V_{p_{\text{r(max)}}}^2}{40r_c}$. But considering worst case condition, transistor power rating must be greater than $P_{D_{\text{max}}}$.  

**D.C. input power**

D.C. input power can be given as,

$$P_{d_c} = V_{cc}I_{dc} \quad \ldots (11)$$

**Efficiency**

The efficiency of the amplifier is given as

$$\eta = \frac{P_{\text{out}}}{P_{d_c}} \times 100 \% = \frac{P_{\text{out}}}{V_{cc}I_{dc}} \times 100 \% \quad \ldots (12)$$

The d.c. collector current depends on the conduction angle for a conduction angle of 180° (a half-wave signal), the average or d.c. collector current is $I_{c_{(\text{sat})}}/\pi$. For smaller conduction angles, the d.c. collector current is less than this, as shown in Fig. 3.30 (d).

In a class C amplifier, most of the d.c. input power is converted into a.c. load power because the transistor and coil losses are small. When the conduction angle is 180°, the efficiency is 78.5 %. The efficiency increases when conduction angle decreases. As indicated class C amplifier has maximum efficiency of 100 %, approached at very small conduction angles.

**Bandwidth**

We know that, bandwidth of resonant circuit is defined as

$$BW = f_2 - f_1$$

where $f_1$ = Lower half power (3 dB) frequency

$f_2$ = Upper half power (3 dB) frequency

The half power frequencies are identical to the frequencies at which the voltage gain equal 0.707 times the maximum gain.

The bandwidth of the class C tuned amplifier is given as,

$$BW = \frac{f_1}{Q} \quad \ldots (13)$$

where $Q$ is the quality factor of the circuit,
Example 3.7: A class C tuned amplifier has inductance of 3 µH and capacitance of 470 pF in the tank circuit. Calculate the resonant frequency.

Solution: We know that,

\[
f_r = \frac{1}{2\pi \sqrt{LC}} = \frac{1}{2\pi \sqrt{3 \mu H \times 470 \text{ pF}}} = 4.238 \text{ MHz}
\]

Now we will discuss a few equations that are useful in the analysis of class C amplifier.

Example 3.8: For the circuit shown in Fig. 3.31, calculate resonant frequency, ac collector resistance, quality factor and bandwidth. Assume \( Q_c = 100 \).

![Fig. 3.31]

Solution: i) Resonant frequency

\[
f_r = \frac{1}{2\pi \sqrt{LC}} = \frac{1}{2\pi \sqrt{2 \mu H \times 470 \text{ pF}}} = 5.19 \text{ MHz}
\]

ii) A.C. collector resistance is given as,

\[
r_c = R_p \parallel R_L = Q_c \omega_c L \parallel R_L
\]

\[
= (Q_c 2\pi f_r L) \parallel R_L
\]

\[
= (100 \times 2\pi \times 5.19 \times 10^6 \times 2 \mu H) \parallel 1 \text{k}\Omega
\]

\[
= 652 \text{k}\Omega \parallel 1 \text{k}\Omega = 867 \Omega
\]

iii) Quality factor of the overall circuit is given as,

\[
Q = \frac{r_c}{\omega_c L} = \frac{r_c}{2\pi f_r L}
\]

\[
= \frac{867}{2\pi \times 5.19 \times 10^6 \times 2 \times 10^{-6}} = 13.29
\]

iv) Bandwidth is given as,

\[
BW = \frac{f_r}{Q} = \frac{5.19 \times 10^6}{13.29} = 390.5 \text{ kHz}
\]
Example 3.9: For the circuit shown in example 3.8, what is the worst case power dissipation?

Solution: The maximum peak to peak output is given as,

$$V_{pp(max)} = 2V_{cc} = 2 \times 15 = 30 \text{ V}$$

The worst case power dissipation (maximum power dissipation) of the transistor is given as,

$$P_{D(max)} = \frac{V_{pp}^2}{40r_e} = \frac{30^2}{40 \times 8 \frac{1}{G}} = 26 \text{ mW}$$

Example 3.10: For the circuit shown in Fig. 3.32, calculate

i) Output power if the output voltage is 50 $V_{pp}$.

ii) Maximum ac output power.

iii) D.C. input power if current drain is 0.5 mA.

iv) Efficiency if the current drain is 0.4 mA and the output voltage is 30 $V_{pp}$.

v) Bandwidth of amplifier if $Q = 125$.

vi) Worst case transistor power dissipation.
Solution: i) The output power of amplifier is given as
\[ P_{\text{out}} = \frac{(V_{\text{out}})^2}{8R_L} = \frac{(V_{pp})^2}{8R_L} = \frac{(50)^2}{8 \times 10 \, \text{K}} = 31.25 \, \text{mW} \]

ii) Maximum a.c. output power is given as,
\[ P_{\text{ac(max)}} = \frac{(V_{pp(max)})^2}{8R_L} = \frac{(2 \times V_{CC})^2}{8R_L} = \frac{(60)^2}{8 \times 10 \, \text{K}} = 45 \, \text{mW} \]

iii) D.C. input power is given as,
\[ P_{\text{dc}} = V_{CC} \times I_{dc} = 30 \times 0.5 \, \text{mA} = 15 \, \text{mW} \]

iv) Efficiency is given as,
\[ \eta = \frac{P_{\text{out}}}{P_{\text{dc}}} \times 100 \]

Given: \( I_{dc} = 0.4 \, \text{mA} \), \( V_{out} = V_{pp} = 30 \, \text{V} \)
\[ = \frac{\left( \frac{V_{pp}^2}{8R_L} \right)}{V_{CC} \times I_{dc}} \times 100 = \frac{30^2}{8 \times 10 \, \text{K} \times 30 \times 0.4 \, \text{mA} \times 100} = 93.75 \, \% \]
v) Bandwidth of the amplifier is given as,

\[ BW = \frac{f_r}{Q} \]

where

\[ f_r = \frac{1}{2\pi\sqrt{LC}} \]

\[ = \frac{1}{2\pi\sqrt{1 \mu\text{H} \times 220 \text{ pF}}} = 10.73 \text{ MHz} \]

\[ \therefore BW = \frac{10.73}{125} = 85.84 \text{ kHz} \]

vi) Worst case transistor power dissipation is given as,

\[ P_{D\text{ (max)}} = \frac{\left(V_{pp\text{ (max)}}\right)^2}{40r_c} \]

where

\[ r_c = R_p \parallel R_L \]

\[ = Q_L \omega L \parallel R_L \]

\[ = \frac{(125 \times 2\pi \times 10.73 \times 10^6 \times 1 \times 10^{-6})}{10 \times 10^3} = 8427.3 \parallel 10 \times 10^3 = 4573.27\Omega \]

\[ \therefore P_{D\text{ (max)}} = \frac{(2 \times 30)^2}{40 \times 4573.27} = 19.68 \text{ mW} \]

Example 3.11: If class C tuned amplifier has \( R_L = 6 \text{ k}\Omega \) and required tank circuit \( Q = 80 \). Calculate the values of \( L \) and \( C \) of the tank circuit. Assume \( V_{CC} = 20 \text{ V} \), resonant frequency = 5 MHz and worst case power dissipation = 20 mW.

Solution: We know that,

\[ P_{D\text{ max}} = \frac{\left(V_{pp\text{ max}}\right)^2}{40r_c} = \frac{(2V_{CC})^2}{40r_c} \]

\[ \therefore r_c = \frac{(2V_{CC})^2}{40 \times P_{D\text{ max}}} \]

\[ = \frac{(2 \times 20)^2}{40 \times 20 \text{ mW}} = 2 \text{ k}\Omega \]
We know that

\[ r_c = R_p \parallel R_L \]

\[ \frac{1}{R_p} = \frac{1}{r_c} - \frac{1}{R_L} \]

\[ = \frac{1}{2 \, \Omega} - \frac{1}{6 \, \Omega} \]

\[ = 3.33 \times 10^{-4} \]

\[ R_p = 3 \, k\Omega \]

We know that,

\[ R_p = Q_L \times \omega_r \times L \]

\[ = Q_L \times 2\pi \times f_r \times L \]

\[ L = \frac{R_p}{Q_L \times 2\pi \times f_r} \]

\[ = \frac{3000}{80 \times 2\pi \times 5 \times 10^6} = 1.19 \, \mu H \]

We know that

\[ f_r = \frac{1}{2\pi \sqrt{LC}} \]

\[ C = \frac{1}{(2\pi)^2 \times L \times f_r^2} \]

\[ = \frac{1}{(2\pi)^2 \times 1.19 \times 10^{-6} \times (5 \times 10^6)^2} = 851 \, \text{pF} \]

3.9.3 Application of Class C Tuned Amplifier

As an application of tuned amplifier we see the mixer or frequency converter circuit. Frequency conversion is the process of translating a modulated signal to a higher or lower frequency while still retaining all the originally transmitted information. Although modulation itself is a form of frequency translation, frequency conversion is often used before and after transmission or reception to provide some benefit.

![Fig. 3.33 Block schematic of mixer circuit](image)

Frequency conversion is a form of AM. It is carried out by a mixer circuit. In some applications, the mixer is referred to as a converter. The function performed by the mixer is called heterodyning.

Fig. 3.33 shows the block schematic of mixer circuit. The mixer accepts two inputs: The signal \( f_s \) which is to be translated to another frequency and sine wave from the oscillator, \( f_o \). The mixer, like an amplitude modulator, performs a mathematical multiplication of its two input signals and produces output signals: \( f_s \), \( f_o \), \( f_s + f_o \) and \( f_s - f_o \). Only one of these signals is the desired one. A tuned circuit or filter is normally used at the output of the mixer to select the desired signal.
The Fig. 3.34 shows the mixer circuit using class C tuned amplifier. Here, transistor is biased to operate as a class C amplifier so that the collector current does not vary linearly with variations in the base current. This results in analog multiplication which produces the sum and difference frequencies. As shown in the Fig. 3.34, both the incoming signal and oscillator signal are applied to the base of the transistor. The tuned circuit selects the sum or difference frequency.

![Mixer circuit using class C tuned amplifier](image)

**Fig. 3.34 Mixer circuit using class C tuned amplifier**

**Applications of mixer or converter circuits**

1. One of the most common applications for mixer is in radio receivers. The mixer is used to convert incoming signal to a lower frequency where it is easier to obtain the high gain and selectivity required.

2. Mixer circuits are used to translate signal frequency to some lower frequency or to some higher frequency. When it is used to translate signal to lower frequency it is called down converter. When it is used to translate signal to higher frequency, it is called up converter.

**3.10 Stability of Tuned Amplifiers**

In tuned RF amplifiers, transistors are used at the frequencies nearer to their unity gain bandwidths (i.e. $f_T$), to amplify a narrow band of high frequencies centered around a radio frequency. At this frequency, the inter junction capacitance between base and collector, $C_{bc}$ of the transistor becomes dominant, i.e. its reactance becomes low enough to be considered, which is otherwise infinite to be neglected as open circuit. Being CE configuration capacitance $C_{bc}$, shown in the Fig. 3.35 come across input and output circuits of an amplifier. As reactance of $C_{bc}$ at RF is low enough it provides the feedback path from collector to base. With this circuit condition, if some feedback signal manages to reach the input from output in a positive manner with proper phase shift, then there is possibility of circuit converted to an unstable one, generating its own oscillations and can stop working as an amplifier. This circuit will always oscillate if enough energy is fed back from the collector to the base in the correct phase to overcome circuit losses. Unfortunately, the conditions for best gain and selectivity are also those which promote oscillation. In order to prevent oscillations in tuned RF amplifiers it was necessary to reduce the stage gain to a level that ensured circuit stability. This could be accomplished in several ways such as lowering the Q of tuned circuits; stagger tuning, loose coupling.
between the stages or inserting a 'loser' element into the circuit. While all these methods reduced gain, detuning and Q reduction had detrimental effects on selectivity. Instead of loosing the circuit performance to achieve stability, the professor L.A. Hazeltine introduced a circuit in which the troublesome effect of the collector to base capacitance of the transistor was neutralized by introducing a signal which cancels the signal coupled through the collector to base capacitance. He proved that the neutralization can be achieved by deliberately feeding back a portion of the output signal to the input in such a way that it has the same amplitude as the unwanted feedback but the opposite phase. Later on many neutralizing circuits were introduced. Let us study some of these circuits.

3.10.1 Hazeltine Neutralization

The Fig. 3.36 shows one variation of the Hazeltine circuit. In this circuit a small value of variable capacitance $C_N$ is connected from the bottom of coil, point B, to the base. Therefore, the internal capacitance $C_{bc}$, shown dotted, feeds a signal from the top end of the coil, point A, to the transistor base and the $C_N$ feeds a signal of equal magnitude but opposite polarity from the bottom of coil, point B, to the base. The neutralizing capacitor, $C_N$, can be adjusted correctly to completely nullify the signal fed through the $C_{bc}$.

Fig. 3.36 Tuned RF amplifier with Hazeltine neutralization
3.10.2 Neutrodyne Neutralization

The Fig. 3.37 shows typical neutrodyne circuit. In this circuit the neutralization capacitor is connected from the lower end of the base coil of the next stage to the base of the transistor.

In principle, this circuit functions in the same manner as the Hazeltine neutralization circuit with the advantage that the neutralizing capacitor does not have the supply voltage across it.

![Fig. 3.37 Tuned RF amplifier with Neutrodyne neutralization](image)

3.10.3 Neutralization using Coil

The Fig. 3.38 shows the neutralization of RF amplifier using coil. In this circuit, L part of the tuned circuit at the base of next stage is oriented for minimum coupling to the other windings. It is wound on a separate form and is mounted at right angles to the coupled windings. If the windings are properly polarized, the voltage across L due to the circulating current in the base circuit will have the proper phase to cancel the signal coupled through the base to collector, C_{bc} capacitance.

![Fig. 3.38 Tuned RF amplifier using coil](image)
3.10.4 Rice Neutralization

The Fig. 3.39 shows the Rice circuit of neutralization. It uses a centre tapped coil in the base circuit. With this arrangement the signal voltages at the ends of the tuned base coil are equal and out of phase.

![Fig. 3.39 Tuned RF amplifier using Rice neutralization](image)

3.11 Advantages and Disadvantages of Tuned Amplifiers

**Advantages:**

1) They amplify defined frequencies.
2) Signal to noise ratio at output is good.
3) They are well suited for radio transmitters and receivers.
4) The band of frequencies over which amplification is required can be varied.

**Disadvantages:**

1) Since they use inductors and capacitors as tuning elements, the circuit is bulky and costly.
2) If the band of frequency is increased, design becomes complex.
3) They are not suitable to amplify audio frequencies.

3.12 Applications of Tuned Amplifiers

The important applications of tuned amplifiers are as follows:

1. Tuned amplifiers are used in radio receivers to amplify a particular band of frequencies for which the radio receiver is tuned.
2. Tuned class B and class C amplifiers are used as an output RF amplifiers in radio transmitters to increase the output efficiency and to reduce the harmonics.
3. Tuned amplifiers are used in active filters such as low pass, high pass and band pass to allow amplification of signal only in the desired narrow band.