1) Amplifier Definition

2) h-Parameter Vs hybrid-π model.

3) Comparison of transistor configurations

4) Classification of Amplifiers, Applications of Amplifiers

5) Need for cascading

6) Types of coupling

7) What is the use of Darlington pair Amp ?

8) What is the use of Bootstrapped Emitter follower ?

9) What is Bandwidth of an amplifier - Definition.

10) Effect of capacitors on freq. response of amplifier.

11). What is Gain Bandwidth product ?

12) Feedback - Definition

13) Types of feedback.

14) Negative F/b Vs Postive F/b

15) Feed back Topologies.

16) characteristics of negative f/b

17) Oscillator - Definition

18) Types of oscillators

19). conditions for a circuit to be oscillator (Barkhausen criteria)

20). Applications of oscillators.

~~[illegible]~~

21) What are RC and Wien Bridge oscillators.

22) What is tank circuit ?

23) LC oscillators.

24) crystal oscillators ? Advantages

25) Voltage Amp Vs Power Amp

26) Types of power Amplifiers

27) comparison of power Amplifiers

28) Need for push-pull Power Amp

29) What is the use of complementary symmetry power Amplifiers

30) Need for Tuned Amplifiers

31) classification of Tuned Amplifiers, Applications.

32) Q-factor - Definition

33) What is synchronous tuning & stagger tuning.

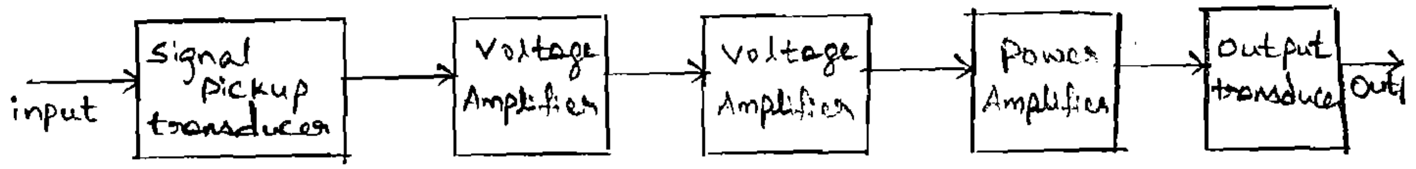
34) Effect of cascading tuned Amp on Band width.

UNIT-V

POWER AMPLIFIERS

Introduction

Consider an amplifier system shown in the below fig(a)



Fig(a)

* A transducer is used to convert one form of energy into another type. Ex:- A microphone converts acoustic signal into electrical signal.
* The voltage amplifier is used to provide high resistance to the input transducer to minimize the loading effects and to provide a large voltage gain
* The power amplifiers are the amplifiers which raise the power levels of the signal. The power amplifier may also be defined as a device which converts the dc power to ac power and whose action is controlled by the ac input signal. power amplifiers does not amplify the power, but it takes power from d.c power supply connected to the output circuit and converts it into useful ac signal power. The type of a.c power available at the output terminals is controlled by the input signal
* power amplifiers are also known as large signal amplifiers
* The transistors used in the power amplifiers are called power transistors. They differ from other transistor in the following aspects.
  (i) Base is made thicker to handle large currents in power amplifiers, i.e, β is small.
  (ii) The area of collector region of power transistor is made large inorder to dissipate the heat developed in the transistor during operation.

(iii) The emitter and base layers are heavily doped.

Differences between voltage and power Amplifiers

Voltage Amplifiers	Power Amplifiers
1) Voltage is amplified	1) power (or) current is amplified
2) h-parameter analysis is applicable.	2) Graphical analysis is required.
3) Harmonics are not present for sinusoidal signals	3) Harmonics are present
4) Normal transistors (BJT's & FET's) are used	4) power transistors are used.
5) Transistors used in voltage amplifier has a large current gain of order 100db	5) power transistors has a small current gain of order 20-50 db.
6) Heat dissipation is not considered	6) Heat dissipation is considered, so heat sinks are used
7) Power handling capacity is small.	7) power handling capacity is large.
8) The output voltage and current swings are small.	8) The output voltage and current swings are large
9) R-C coupling is used normally.	9) Transformer coupling is used.

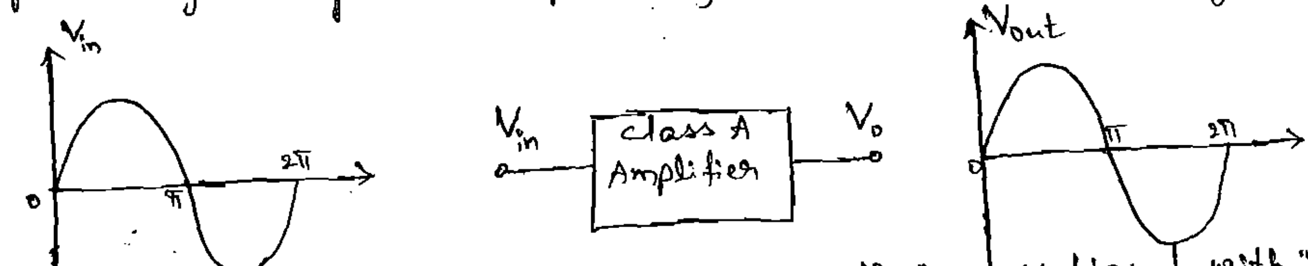
Classification of power Amplifiers

Based on the amount of transistor bias and amplitude of the input signal, amplifiers can be classified as

(i) class A amplifier (ii) class B amplifier
(iii) class C amplifier (iv) class AB amplifier.
(v) class D amplifier (vi) class S amplifier

(i) class A Amplifier

It is an amplifier in which the transistor is biased in such a way that the output current flows for the complete cycle of the input signal as shown in fig(a)

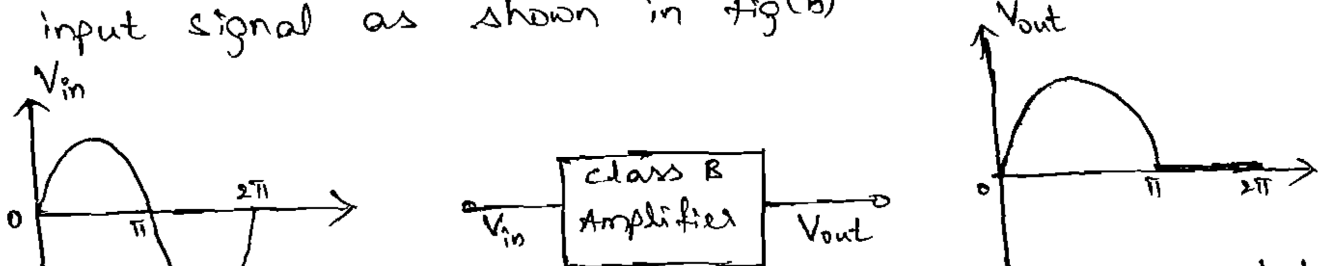


Transistor is well biased with in the active region.

Fig(a)

(ii) class B Amplifier

It is an amplifier in which the transistor is biased, such that the output current flows only for one half cycle of input signal as shown in fig(b)



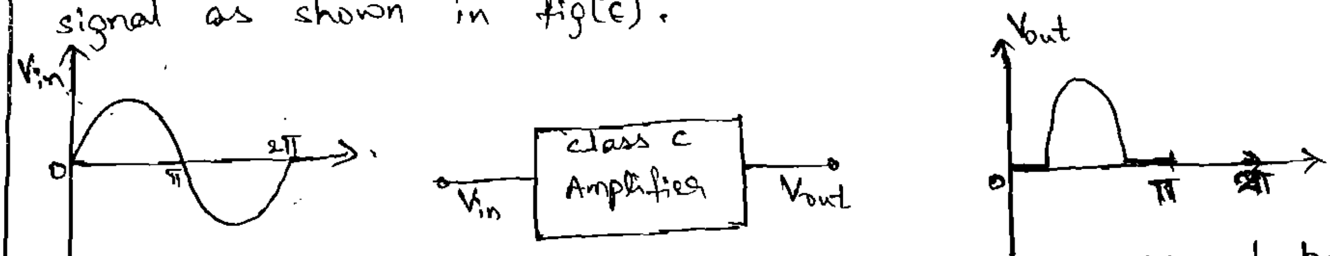
180° conduction.

Fig(b).

Transistor is biased at cut-off region.

(iii) class C Amplifier

It is an amplifier in which the transistor biased such that output current flows for less than half cycle of the input signal as shown in fig(c).

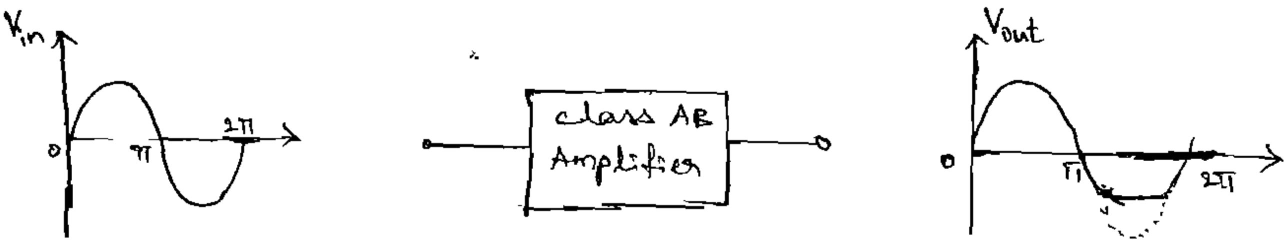


Fig(c)

Transistor biased beyond the cut-off region.

(iv) class AB Amplifier

The characteristics of class AB amplifier lies between class A and class B amplifier. Thus in a class AB amplifier, the output current flows for more than half cycle but less than full cycle as shown in fig(c).

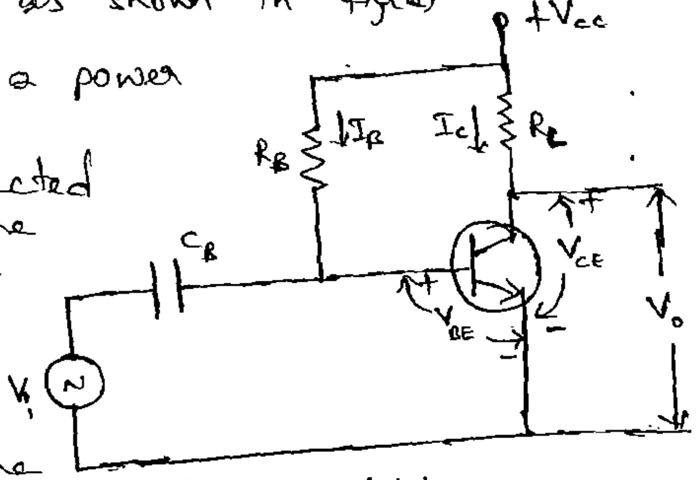


Direct coupled class-A large signal Amplifier

A simple fixed-bias circuit can be used as a large signal class A amplifier as shown in fig(a)

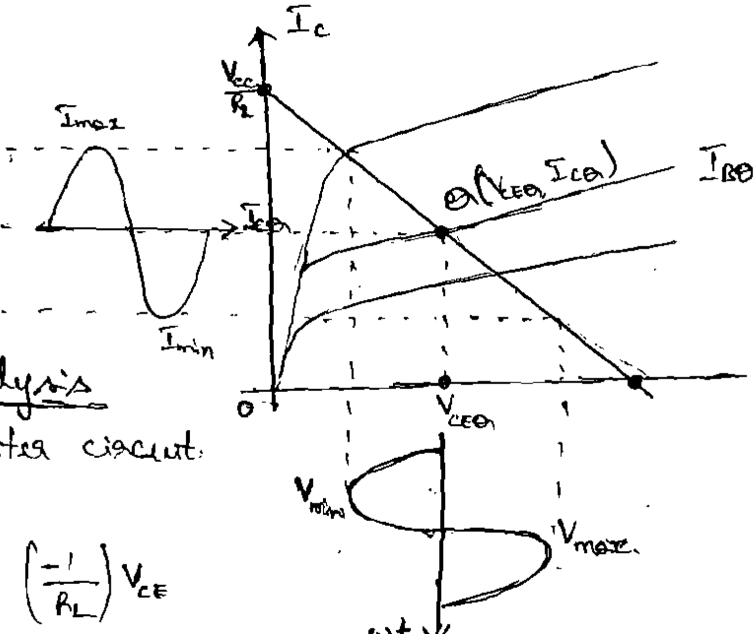
* The transistor used is a power transistor.
* The value of R_B is selected in such a way that the Q point lies ~~between~~ at the centre of the d.c load line.
* The load resistor is directly connected in the collector circuit, so it is called as directly coupled class A amplifier.

Mostly the load ~~resistance~~ is a loud speaker of impedance of ~~···~~ 4 Ω to 16 Ω. The beta(β) of the transistor is less than 100.



Fig(a)

* circuit handles large power in the range of tens of watts. without providing much voltage gain



D.C Analysis

~~Vcc~~ Apply KVL to collector emitter circuit.

$$V_{cc} = I_C R_L + V_{CE}$$

$$\therefore\ I_C = \frac{V_{cc} - V_{CE}}{R_L} = \frac{V_{cc}}{R_L} + \left(\frac{-1}{R_L}\right) V_{CE}$$

A.c Analysis

(i) D.c input power

$$\boxed{P_{in(D.c)} = V_{cc} \times I_{CQ}}$$

(ii) A.c output power

$$P_{o(A.c)} = V_{rms} \cdot I_{rms}$$
$$= \frac{V_m}{\sqrt{2}} \cdot \frac{I_m}{\sqrt{2}}$$
$$= \frac{V_m \cdot I_m}{2}.$$

$$\therefore \boxed{P_{o(A.c)} = \frac{V_m I_m}{2}}$$ using rms values (or) peak values.

(iii) A.c output power

$$\boxed{P_{o(A.c)} = \frac{(V_{max} - V_{min})(I_{max} \cdot I_{min})}{8}}$$ using peak-to-peak values.

(iv) Maximum efficiency

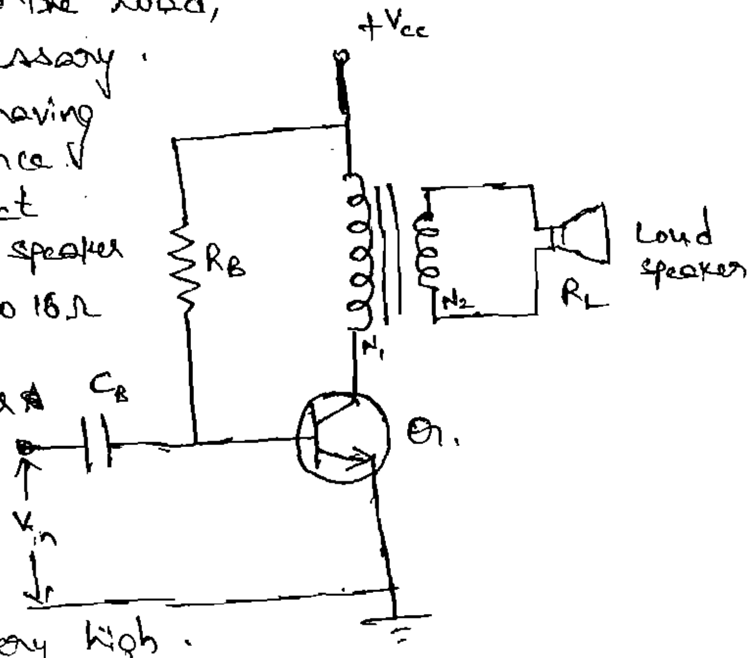
$$\% \eta = \frac{P_{ac}}{P_{dc}} \times 100$$
$$= \frac{(V_{max}-V_{min})(I_{max}-I_{min})}{8 V_{cc} I_{CQ}} \times 100$$
$$= \frac{(V_{cc}-0)(2I_{CQ}-0)}{8 V_{cc} I_{CQ}} \times 100$$

$$\therefore \boxed{\eta = 25\%} \rightarrow$$ very low efficiency

Transformer coupled class A Power Amplifier

For maximum power transfer to the load, the impedance matching is necessary. For loads like loud speaker, having low impedance values, impedance matching is difficult using direct coupling. This is because loud speaker resistance is in the range of 3 to 16 Ω while the output impedance of series fed directly coupled amplifiers is very much high.

To overcome this problem, we must replace R_c by a component whose D.c resistance is zero but A.c resistance is very high. So, a choke (or) inductor can be used.



Fig(a)

The circuit of a transformer coupled class A power amplifier is shown in fig(a).

Transformer Impedance matching

let

n_1 - turns in the Primary

n_2 - turns in the secondary

V_1 - Primary voltage

V_2 - secondary voltage

I_1 - Primary current

I_2 - secondary current

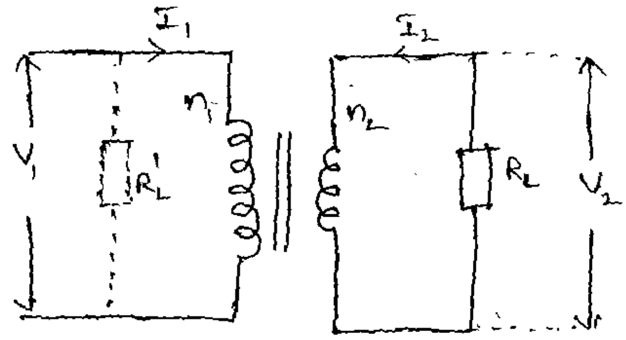


Fig (b)

W.k.t. turns ratio is related to.

$$\frac{V_1}{V_2} = \frac{n_1}{n_2} \quad \text{and} \quad \frac{I_1}{I_2} = \frac{n_2}{n_1}.$$

$$\Rightarrow V_1 = \frac{n_1}{n_2} V_2 \quad \text{and} \quad I_1 = \frac{n_2}{n_1} I_2.$$

$$\Rightarrow \frac{V_1}{I_1} = \left(\frac{n_1}{n_2}\right)^2 \cdot \frac{V_2}{V_1}.$$

$$\Rightarrow \boxed{R_L' = \left(\frac{n_1}{n_2}\right)^2 R_L}$$

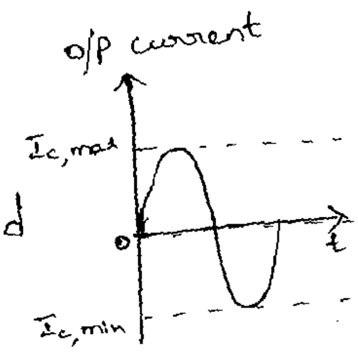
where,

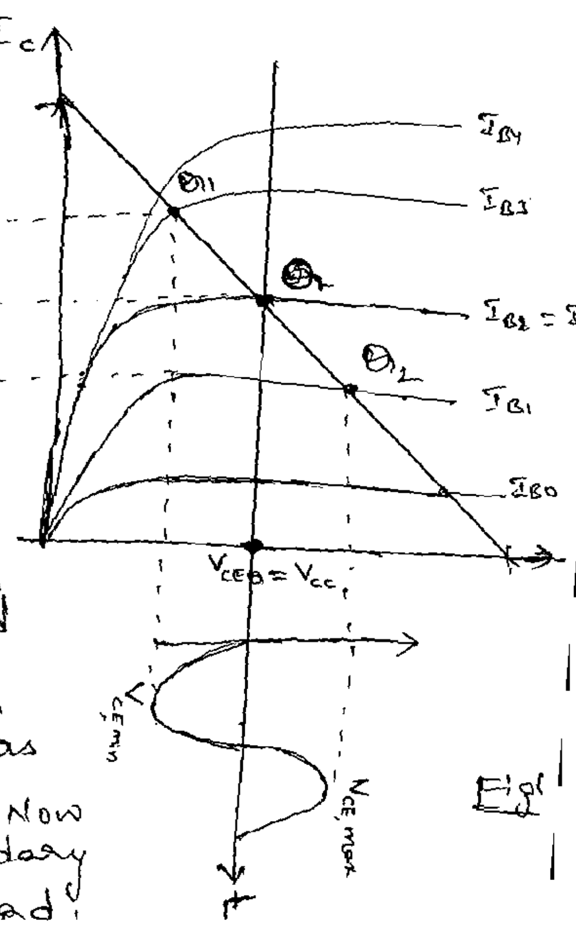
$\frac{V_1}{I_1} = R_L'$ = effective input resistance.

$\frac{V_2}{I_2} = R_L$ = effective output resistance.

Analysis

In order to get maximum a.c power output, ~~the peak value of the collector current~~ the Q should be located at the centre of load line as shown in fig(c).





Fig(c)

When a.c signal is applied, collector current fluctuates. Now the operating point Q moves up and down the load line. The collector voltage varies in the opposite to the collector current. The variation of collector voltage appears across the primary of transformer. Now a.c voltage is induced in the secondary which inturn develops a.c power in load.

Power calculations

(i) D.c input power

$$\boxed{P_{in(D.c)} = V_{cc} \cdot I_{CQ}}$$

(ii) A.c output power

$$P_{o(A.c)} = \frac{\left(V_{CE(max)} - V_{CE(min)}\right)\left(I_{c(max)} - I_{c(min)}\right)}{8}$$

For maximum A.c output power, the Q-point is selected such that it gives the maximum output ~~power~~ voltage and current swings.

$P_{o(A.c)}$ becomes maximum when $V_{CE(min)} = 0$ & $I_{c(min)} = 0$.

$$\Rightarrow P_{o(A.c)} = \frac{V_{CE(max)} \cdot I_{c(max)}}{8}$$

$$= \frac{V_{CE(max)} \cdot \left(\frac{V_{CE(max)}}{R_L'}\right)}{8} \qquad \left[\because I_{cmax} = \frac{V_{CE(max)}}{R_L'}\right]$$

$$= \frac{(2V_{cc})(2V_{cc})}{8R_L'} \qquad \because V_{CE(max)} = 2V_{cc}$$

$$\therefore \boxed{P_{o(A.c)} = \frac{V_{cc}^2}{2R_L'}}$$

(iii) ~~collector~~ Efficiency

$$\%\eta = \frac{(P_o)_{ac}}{(P_i)_{d.c}}$$

$$= \frac{V_{cc}^2/2R_L'}{V_{cc} \cdot I_{CQ}} \times 100\%$$

$$= \frac{V_{cc}^2/2R_L'}{V_{cc} \cdot \frac{V_{cc}}{R_L'}} \times 100\%$$

$$= 50\%$$

Advantages

1.) The efficiency of the amplifier is increased to 50%.
2.) Due to transformer coupling, at the collector, impedan matching can be achieved.
3.) The quiescent collector current does not flow throug the load resistance and hence there is no wastage of power.

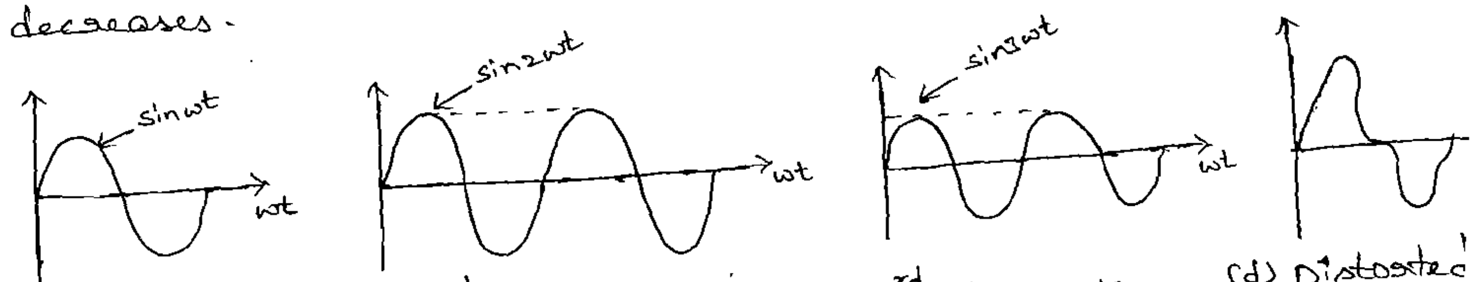
Disadvantages

1.) Due to the transformer, the circuit is bulky, and costly.
2.) Frequency response is poor due to transformer coupling

Harmonic Distortion

Harmonic distortion means the presence of the frequency components in the output waveform, which are not present in the input signal. The component with frequency same as the input signal is called fundamental frequency component.

As the order of the harmonic increases, its amplitude decreases.



(a) Fundamental (b) 2^{nd} Harmonic (c) 3^{rd} Harmonic (d) Distorted waveform

As the second harmonic waveform is largest, so the second harmonic distortion is more important in the analysis of power amplifiers.

The fig(d) shows the distorted waveform which is obtained by adding the fundamental and the harmonic components.

2^{nd} Harmonic Distortion

$$\% D_2 = \frac{|B_2|}{|B_1|},$$

where,

B_2 = amplitude of 2^{nd} harmonic component

B_1 = fundamental frequency component amplitude

$$\% D_3 = \frac{|B_3|}{|B_1|}$$ and so o

Where

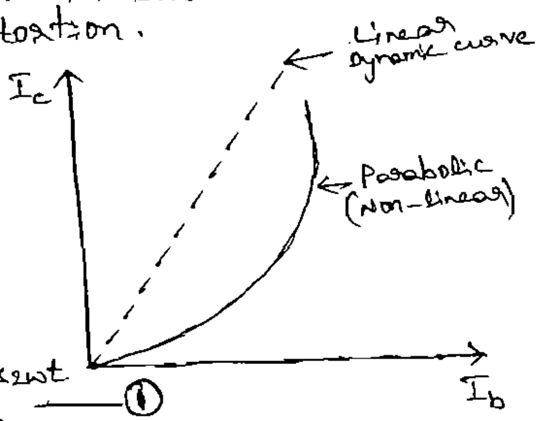
B_3 = amplitude of 3^{rd} harmonic compone

Total harmonic Distortion

$$\% D = \sqrt{D_2^2 + D_3^2 + D_4^2 + \cdots} \times 100$$

Second Harmonic Distortion (Three point Method)

In the analysis of series fed class-A power amplifier it is assumed that the active device is linear. But in practice it is not true since the output characteristics of the transistor are not equidistant straight line for equal increments in base current. This results in non-linear (or) amplitude distortion.

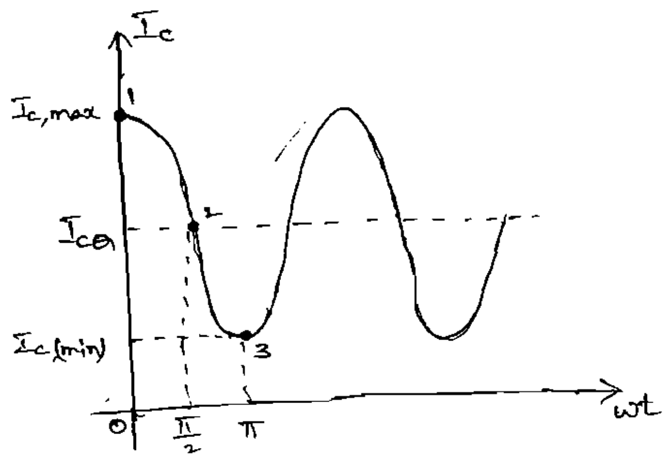


The total collector current I_c which swings about its quiescent value I_{CQ} is the sum of I_{CQ} and I_c.

$$\therefore I_{c,total} = I_{CQ} + I_c$$

$$\Rightarrow I_{c,total} = I_{CQ} + B_0 + B_1 \cos\omega t + B_2 \cos 2\omega t \quad \text{—— (1)}$$

where, B_0, B_1 and B_2 are coefficients of harmonic components.



Fig(b)

At point 1

$\omega t = 0$.

By substituting $\omega t = 0$ in eq (1), we get

$$I_c = I_{CQ} + B_0 + B_1 + B_2 \quad \text{—— (2)}$$

At point 2

$\omega t = \pi/2$, sub in eq (1).

$$I_c = I_{CQ} + B_0 - B_2 \quad \text{—— (3)}$$

At point 3

$\omega t = \pi$, sub in eq (1),

$$I_c = I_{CQ} + B_0 - B_1 + B_2 \quad \text{—— (4)}$$

By solving (2), (3) & (4), we get % second harmonic distortion as

$$\boxed{\% D_2 = \frac{|I_{c,max} + I_{c,min} - 2 I_{CQ}|}{|I_{c,max} - I_{c,min}|} \times 50\%}$$

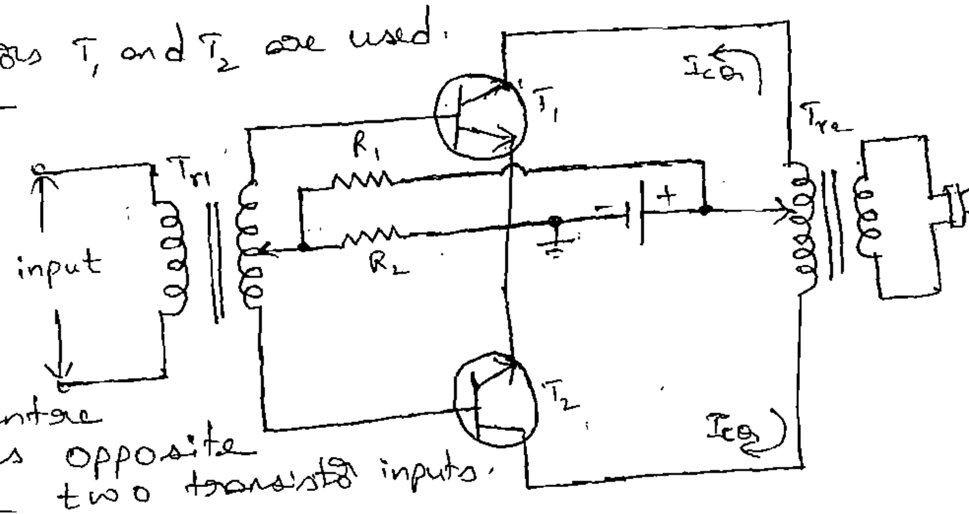
III^{ly} interms of collector emitter Voltage

$$\boxed{\% D_2 = \frac{|V_{CE,max} + V_{CE,min} - 2 V_{CEQ}|}{|V_{CE,max} - V_{CE,min}|} \times 50\%}$$

Class A Push-Pull Power Amplifier

The distortion introduced by non-linearity of the dynamic transfer characteristics using a single transistor as amplifier can be minimised by using push-pull arrangement shown in fig(a).

In this, two transistors T_1 and T_2 are used. The two emitters are connected together. The input is applied to the two inputs through centre tapped transformer. The transformer is centre tapped. so, it provides opposite polarity signals to the two transistor inputs.



Fig(a)

As shown in fig(a), the two transistors Q_1 and Q_2 carry D.C components of collector currents I_{CQ}. These currents are equal in magnitude and flow in opposite direction through the primary of transformer T_{r2}. so, there is practically no net D.C power output which is obtained by a single transistor.

When A.C signal is applied to the input,
when the input signal is positive, the base of Q_1 is more positive than base of Q_2. Hence i_{c1} of transistor Q_1 increases while i_{c2} of transistor Q_2 decreases. These currents flow in opposite directions in two halves of the primary of output transformer. The flux produced ~~also~~ by these currents will also be in opposite directions. As a result voltage across the load will be induced whose magnitude will be proportional to the difference of collector currents. i.e, $(i_{c1} - i_{c2})$

$\|^{ly}$, when the input signal is negative, the ~~base gets~~ voltage across load induced will be proportional to $(i_{c2} - i_{c1})$. The overall operation results in an A.C voltage induced in the secondary of output transformer and hence A.C power is delivered to the load.

Hence, during any half cycle of input signal, one transistor is being driven driven deep into conduction, while the other being non-conducting. Hence the name push-pull amplifier.

Harmonic Distortion in class-A push-pull Amplifier

The load ~~to~~ current, is the difference of two collector currents.

i.e, $i_L = i_{c_1} - i_{c_2}$

$$\therefore \quad i_L = 2B_1 \sin \omega t + 2B_3 \sin 3\omega t + \dots \quad \text{— (1)}$$

$\therefore$ From (1), it is clear that, in push-pull arrangement, even harmonic components $2^{nd}, 4^{th}, 6^{th}, \dots$ gets eliminated. Hence the total distortion is less and contains only odd harmonics.

$$\therefore \quad \% D_3 = \frac{|B_3|}{|B_1|} \times 100\%, \quad \% D_5 = \frac{|B_5|}{|B_1|} \times 100 \quad \dots$$

Hence, the total Distortion is,

$$\boxed{\% D = \sqrt{D_3^2 + D_5^2 + D_7^2 + \dots}}$$

Advantages

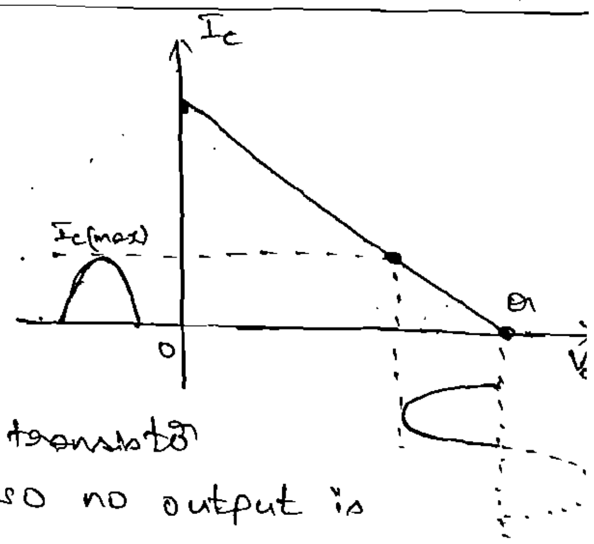
1.) Even harmonics are absent in the output
2.) High A.C output power is obtained.
~~3.)~~

Disadvantages

1) Two identical transistors are required
2) centre tapping is required in transformer
3) Transformers used are bulky and expensive.

~~Cross-over Distortion~~

Class - B Amplifier

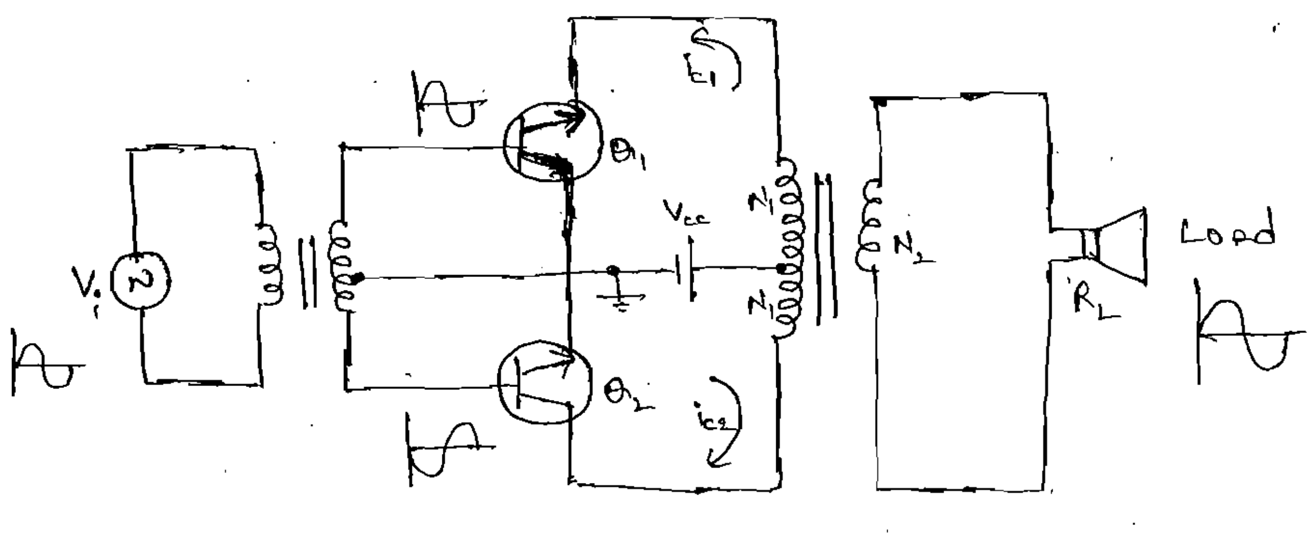


* Q-point is on the x-axis
* The transistor remains in the active region only for positive half cycle of the input signal. Hence this positive half cycle is reproduced at the output
* For negative half input cycle the transistor enters into a cut-off region and so no output is produced
* so, the collector current flows only for 180° of the input signal
* As only a half cycle is obtained at the output for full input cycle, the output signal is distorted.
* The efficiency of class B operation is much higher than the class A operation.

Push-Pull class B Amplifier

To get full cycle across the load, a pair of transistors are used in class-B operation. The both transistors are of same type. i.e, either n-p-n or p-n-p then the circuit is called Push-Pull class-B power Amplifier.

The circuit arrangement of class-B push-pull amplifier is shown in fig(a)



Fig(a)

D.c input power

$$P_{in(D.C)} = V_{cc} \times I_{dc}$$
$$= V_{cc} \times \left(\frac{2I_m}{\pi}\right)$$

$$\therefore \boxed{P_{in(D.C)} = \frac{2I_m}{\pi} \cdot V_{cc}}$$

A.c output power

$$P_{o(A.C)} = V_{rms} \cdot I_{rms}$$
$$= (I_{rms} \cdot R_L') \cdot I_{rms}$$
$$= I_{rms}^2 \cdot R_L' = \frac{V_{rms}^2}{R_L'}$$

$$\Rightarrow \boxed{P_{o(A.C)} = \frac{I_m^2 \cdot R_L'}{2} = \frac{V_m^2}{2R_L'}}$$

Efficiency

$$\% \eta = \frac{P_{ac}}{P_{DC}} \times 100.$$

$$= \frac{\left(\frac{V_m I_m}{2}\right)}{\frac{2}{\pi} V_{cc} \cdot I_m} \times 100.$$

$$\therefore \boxed{\% \eta = \frac{\pi}{4} \frac{V_m}{V_{cc}} \times 100}$$

$$\eta_{max} = \frac{\pi}{4} \times \frac{V_{cc}}{V_{cc}} \times 100$$
$$= 78.5\%.$$

Total Harmonic Distortion

$$\% D = \sqrt{D_3^2 + D_5^2 + D_7^2 + \cdots}$$

Cross over Distortion

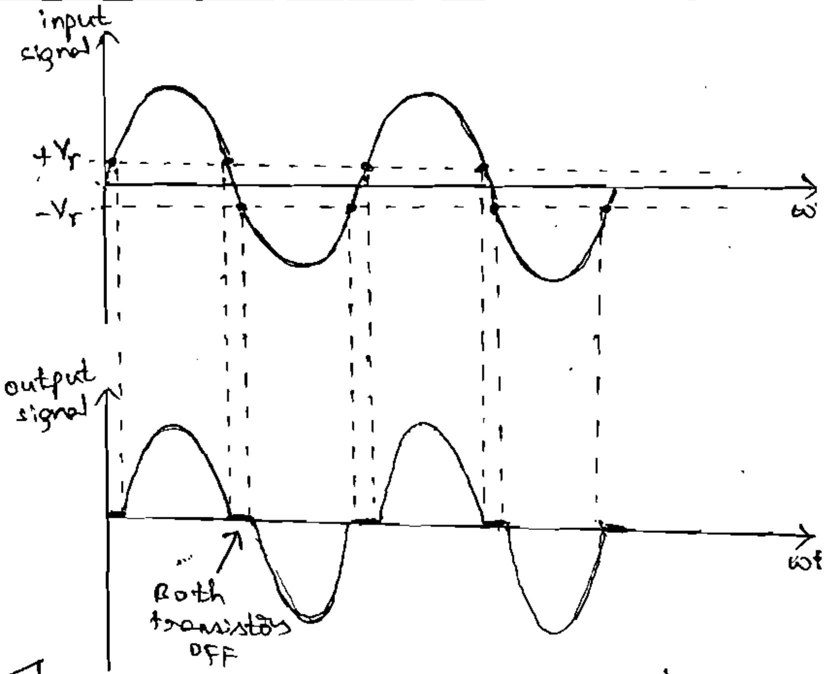
For a transistor to be in active region, the base-emitter junction must be forward biased. The junction will be forward biased when $V_{BE} > V_\gamma$, where V_γ = cut-in voltage and V_γ = 0.2V for Ge and 0.7V for Si.

For $V_{BE} < V_\gamma$ the ~~~~ transistor remains in cut-off and hence collector current will be zero.

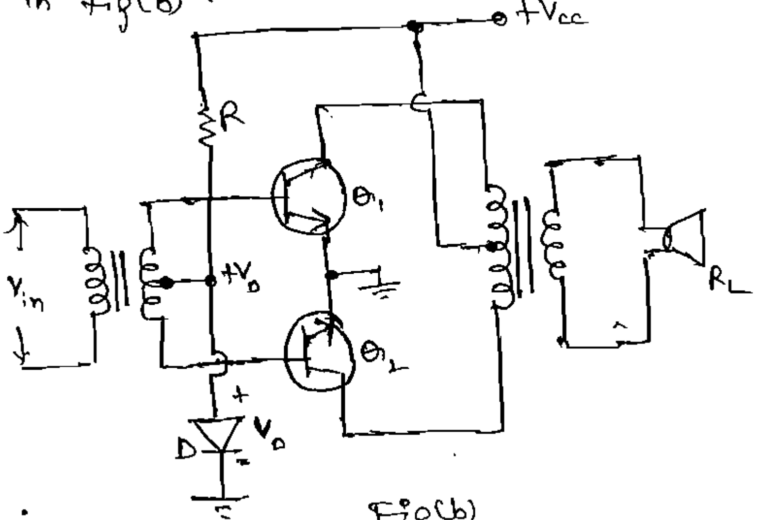
Hence, there is a period between the crossing of half cycles of input signal for which none of the transistors is active and output is zero. Hence the nature of the output signal is distorted and no longer remains same as the input. Such a distortion in the output signal is called a cross-over distortion. Due to cross-over distortion ~~the~~ each transistor conducts for less than a half cycle rather than the complete half cycle.

The basic reason for cross over distortion is the cut-in voltage of the transistor junction.

To overcome this cut-in voltage, a small forward bias is applied to the transistors as shown in fig(b).



fig(a) cross-over Distortion



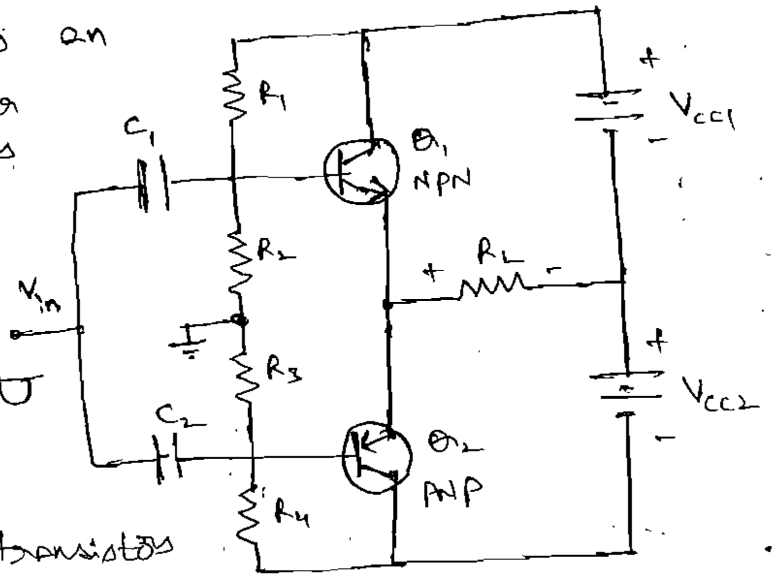
Fig(b)

Complementary Symmetry Push-Pull class B Power Amplifier

The use of transformers at input and as well as at the output in the push-pull amplifier makes it bulky and expensive.

Another drawback of the circuit is that it needs two out-of phase signals which requires an centre tapped transformer and thus the circuit gets complicated.

The above two drawbacks of an ordinary push-pull can be overcome in the complementary symmetry push-pull amplifier.



Fig(a)

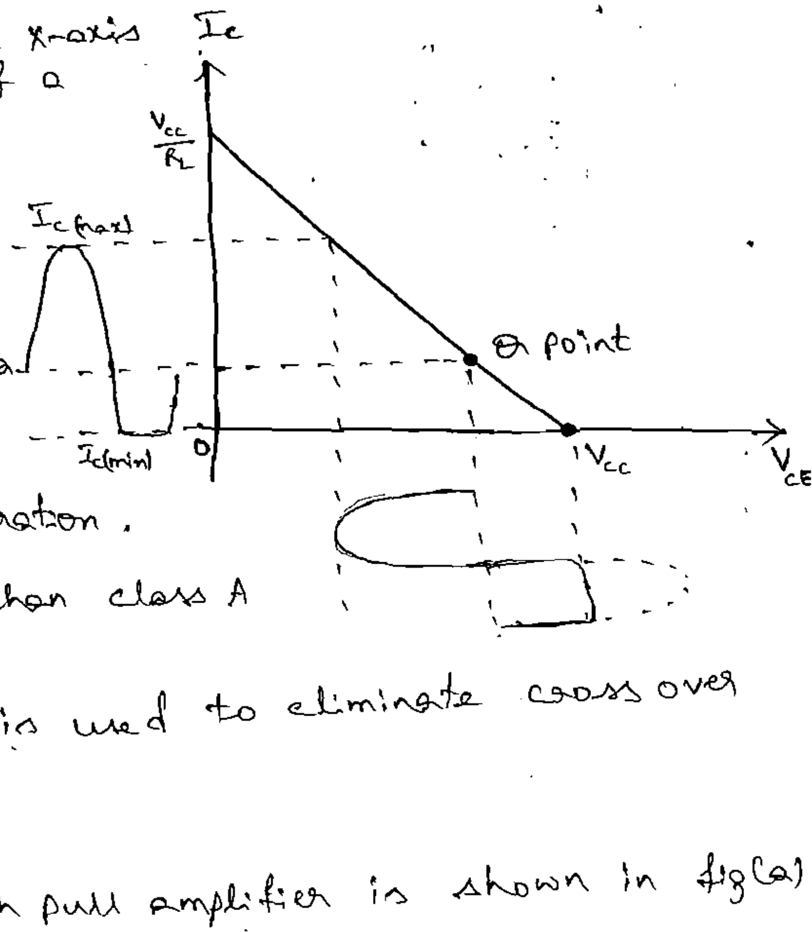
This arrangement uses two transistors having complementary & symmetry (one NPN and other PNP). The term complementary means uses two identical transistors one NPN and the other PNP. The term "symmetry" means the biasing resistors connected in both transistors are equal. As a result of this emitter base

junction of each transistors is biased with the same voltage.
The resistors R_1 and R_2 provide the voltage divider bias that forward bias the emitter base junction of transistor Q_1 and similarly the resistors R_3 and R_4 forward bias the emitter base junction of transistor Q_2.

During positive half cycle of the input, the NPN transistor Q_1 is forward biased and conducts while the PNP transistor Q_2 is reverse biased and do not conducts. similarly in the negative half cycle of the input Q_2 conducts and Q_1 do not conducts. Thus for a complete cycle of input, a complete cycle of the output signal is developed.

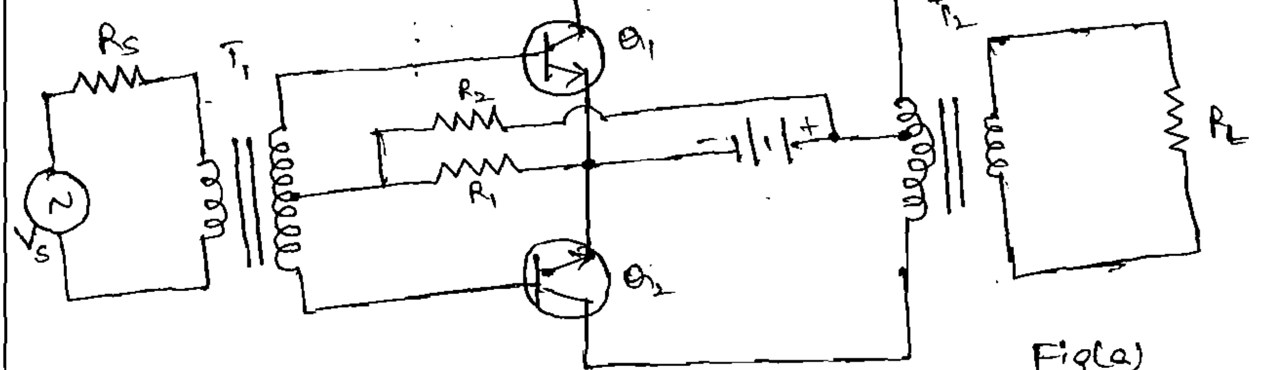
Class AB Amplifiers

* The Q-point is above the x-axis but below the mid point of a load line
* The output signal is obtained for more than 180° but less than 360° for a full input cycle.
* The output signal is distorted in class AB operation.
* The efficiency is more than class A but less than class B
* The class AB operation is used to eliminate cross over distortion.



class-AB Push Pull Amplifier

The circuit for class-AB push pull amplifier is shown in fig(a)



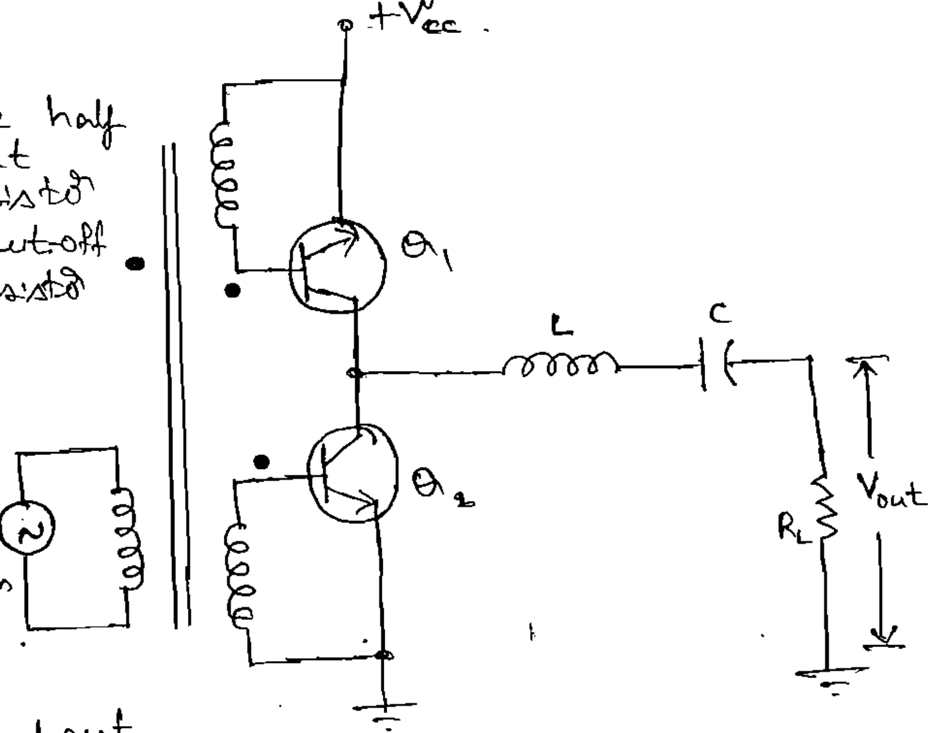
Fig(a)

- The cross over distortion caused by the non-linear transistor characteristics can be eliminated by using class-AB push pull amplifier.

To minimize the cross over distortion, a small stand by current flows at zero excitation. The voltage drop across R_1 is adjusted to be approximately equal to V_r. Thus class-AB operation results in less distortion than class B.

Class D Amplifier

Class-D Amplifier is an electronic amplifier where all power devices (usually MOSFET's) are operated as binary switches. The name class-D arises from the fact that the circuit is designed to operate with digital or pulse-type signals. Fig(a) shows the basic circuit of class-D amplifier. Here Push-pull connection with complementary symmetry transistor is used.



Fig(a)

* During the positive half cycle of the input voltage, the transistor Q_1 is driven into cut-off state and the transistor Q_2 is driven into saturation state.

* During the negative half cycle of the input voltage, the transistor Q_1 is ON and Q_2 is OFF.

As a result, the output voltage is a square wave output alternating between two values 0 to 1 volts. This square wave is given into high Q tank circuit which converts the pulse-type signal back to sinusoidal signal.

D.C input power

$$P_{in(D.C)} = V_{cc}(2I_c)$$

$$\therefore \boxed{P_{in(D.C)} = \frac{8V_{cc}^2}{\pi^2 R_L}} \qquad \left[\because I_c = \frac{4V_{cc}}{\pi^2 R_L}\right]$$

A.C output power

$P_{o(A.C)} = I_{rms}^2 \cdot R_L = \left(\frac{I_m}{\sqrt{2}}\right)^2 R_L$

$$\boxed{P_{o(A.C)} = \frac{8V_{cc}^2}{\pi^2 R_L}} \qquad \left[\because I_m = \frac{4V_{cc}}{\pi R_L}\right]$$

Efficiency

$$\% \eta = \frac{P_{o(A.C)}}{P_{in(D.C)}} \times 100$$

$$= \left(\frac{8V_{cc}^2 / \pi^2 R_L}{8V_{cc}^2 / \pi^2 R_L}\right) \times 100$$

$$\approx 100\%$$

Advantages

1) Very high power conversion efficiency.
2) Reduction in cost due to compact circuitry.
3) Reduction of power wastage.
4) Reduction in size and weight of the amplifier.

Class-S Amplifier

The class-D amplifier can be used to amplify an FM signal or any constant amplitude sinusoidal signal but the output is insensitive to change in the amplitude of a signal. The class-S power amplifier can be used to amplifies either constant amplitude or varying amplitude signal (such as FM or AM).

The figure shows the block diagram of class-S amplifier.

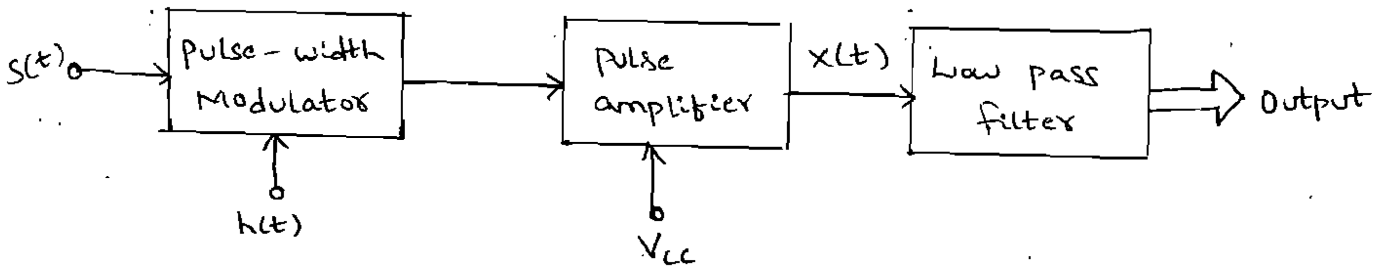
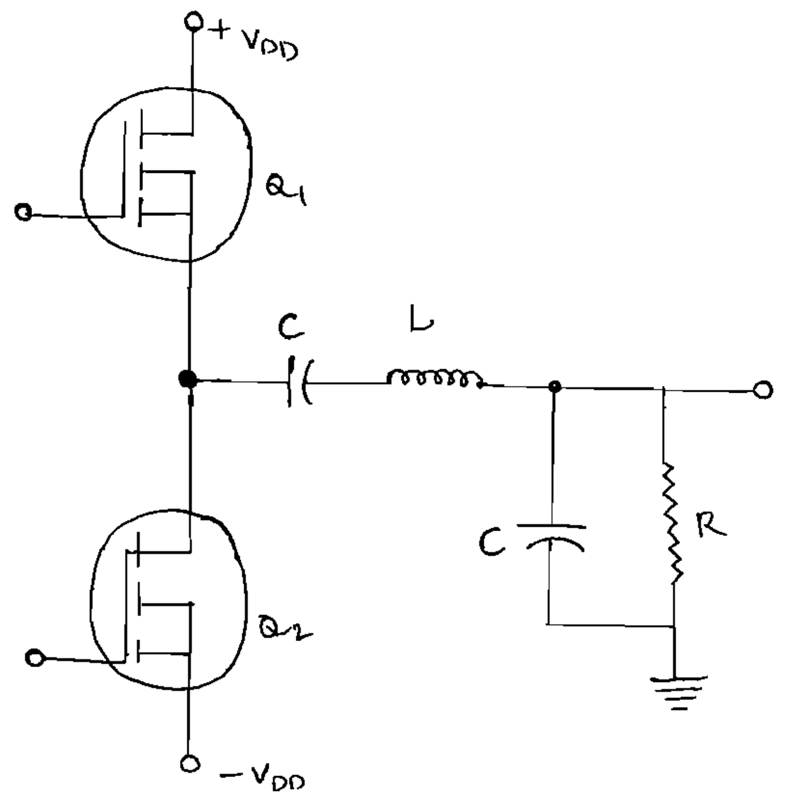


fig (a): Block diagram of class-S amplifier.

The input signal is pulse width modulated which generates constant amplitude pulses. This PWM wave is modified by pulse amplifier which is highly efficient switching amplifier like class-D amplifier. The low pass filter eliminates the frequency components, higher than modulating frequency.

A simplified class-S amplifier is shown in figure (b).



figure(b): Simplified class – S Amplifier.

From the figure (b) it can be noticed that the basic difference b/w class – D and class – S amplifiers is that the output stage of the class – D uses tuned circuit which is tuned to the fundamental frequency of the input while the output stage of the class – S uses the LPF which recovers the input voltage signal.

Heat Sinks

The transistors handling small signals, the heat developed at the collector junction is small, therefore, such transistors have little chances of thermal runway. However in power transistors, heat produced at the collector junction is large which may lead to transistor destruction.

Thus, heat sink is a sheet of metal used to dissipate heat developed at the collector junction of a power transistor. A fin – type heat sink is used for low power transistor and rectangular heat sink is used for large power transistor.

Heat sink increases the dissipation area from which the heat is to be dissipated to the radiation. The heat produced at the collector junction spreads over the metal sheet and is dissipate by convection radiation.

Thermal Resistance

It is experimentally found that rise in temperature at collector junction is directly proportional to the power dissipated at the junction that is

$T_J - T_A \propto P_D$

$\Rightarrow \boxed{T_J - T_A = \Theta P_D}$ —— ①

where $T_J \rightarrow$ Collector base junction temperatu

$T_A \rightarrow$ Ambient temperature of air around the transistor

$P_D \rightarrow$ power dissipated.

In eq ① Θ is called as thermal resistance which resists the heat flow between the two temperature points.

Figure (a) shows the power - temperature derating curves for a germanium transistor. From figure it can be noticed that maximum collector current allowed for safe operation is at 25°C. For the ambient temperatures above this value of P_D reduces to zero.

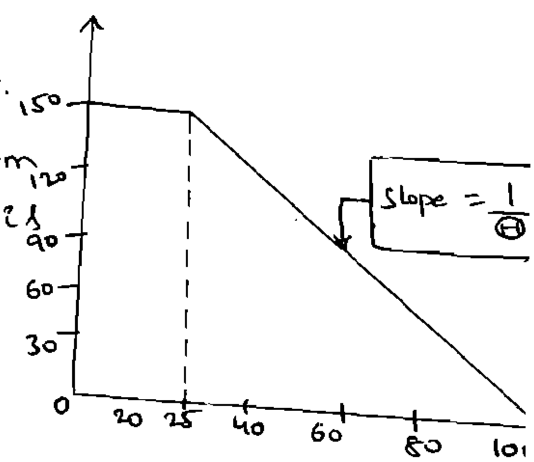


figure (a): Power - temperature Derating curve (thermal

Thermal stability

To avoid thermal runaway the required condition is that, the rate at which heat is released at the collector junction must not exceed the rate at which the heat can be dissipated under steady state condition. i.e.,

$$\boxed{\frac{\partial P_C}{\partial T_J} < \frac{\partial P_D}{\partial T_J}}$$

UNIT-VI

Tuned Amplifiers

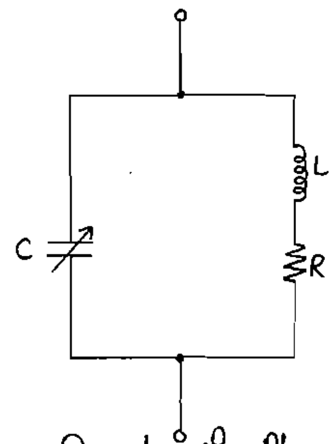
* Introduction :-

→ 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 which such a tuned circuit as a load are known as tuned circuit. Since tuned amplifiers amplify narrow band of frequencies they are also known as narrow band amplifiers.

→ The resonance frequency and the impedance of tuned circuit is given as,

$$f_r = \frac{1}{2\pi\sqrt{LC}}$$

and $Z_r = \frac{L}{CR}$.



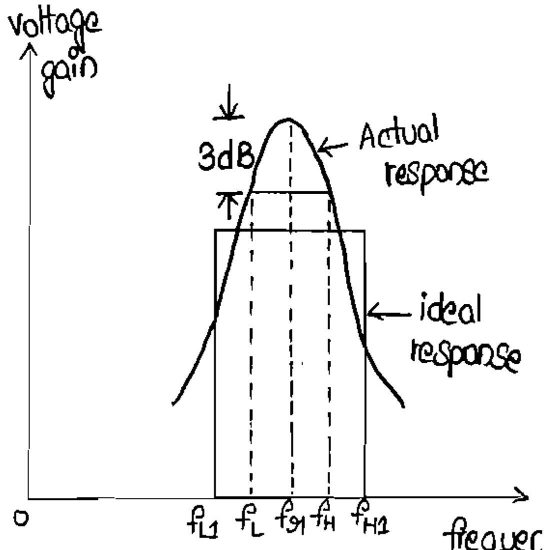
Tuned circuit

→ The response of tuned amplifiers is maximum at resonant frequency at it falls sharply for frequencies below and above the resonant frequency, as shown in the below figure. It is designed to reject all frequencies below a lower cut-off frequency, f_L and above a upper cut-off frequency f_H.

As shown in the figure, 3 dB band width is denoted as B and 30 dB band width is denoted as S. The ratio of 30dB band width (S) to the 3 dB bandwidth (B) is known as skirt selectivity.

f_L and f_H : Actual response

f_{L1} and f_{H1} : Ideal response.



Ideal & Actual response of tuned circuit

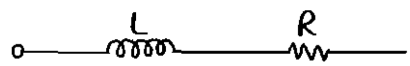
→ At resonance, inductive and capacitive effect of tuned circuit cancel each other. As a result, circuit is like resistive and $\cos\phi = 1$ i.e., voltage and current are in phase. For frequencies, above resonance circuit is like capacitive and for frequencies below resonance it is like inductive. Since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

* Coil losses :-

→ As shown in the first figure, 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 in series with the Inductor. The total loss of the coil is comprised of copper loss, eddy current loss and hysterisis loss.

→ The copper loss at low frequencies is equivalent to the d.c resistance of the coil. Copper loss is inversely proportional to frequency. There, fore, the frequency increases, the copper loss decreases.

→ Eddy current loss in iron and copper loss coil are due to current flowing within the copper (or) core cased by induction. The result of eddy current is a loss due to the heating within the inductors copper (or) core.



Inductor with leakage resistance

→ Eddy current losses are directly proportional to frequency. Hysterisis loss is proportional to the area enclosed by the hysterisis loop and to the rate at which the loop is transversed. It is a function of signal level and increases with frequency. Hysterisis loss is however independent of. frequency.

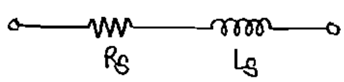
→ As mentioned earlier, the total losses in the coil or inductor i represented by inductance in series with leakage resistance in the coil. As shown in the above figure.

* Q-factor :-

→ Quality factor (Q) is important characteristics of an 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 contains only reactance). The higher the Q of an inductor the few 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.

$$Q = 2\pi \frac{\text{Maximum energy stored per cycle}}{\text{Energy dissipate per cycle}}.$$

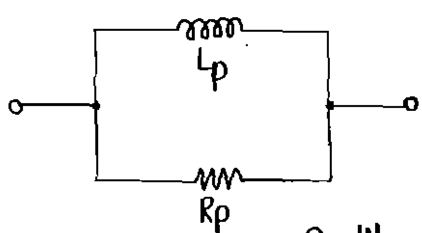
Series circuit



Inductive impedance.

$$\frac{\omega L_S}{R_S}$$

parallel circuit



Inductive admittance

$$\frac{R_P}{\omega L_P}.$$

Quality factor equations

→ The dissipation factor (D) that can be refered as the total loss within a component is defined as 1/Q. The above fig. shows the quality factor equations for series and parallel circuits and its equations relations with dissipation factor.

⇒ Quality factor equation $Q = \frac{1}{D} = \frac{\omega L_S}{R_S}$

$$= \frac{R_P}{\omega L_P}.$$

* Derive the expression for quality factor, Q of an inductor.

→ consider the circuit shown in the below fig. Here, a sinusoidal voltage $V_m \sin\omega t$ is applied to the inductor with an internal resistor R_S.

The maximum energy stored per cycle $= \frac{1}{2} I_m^2 L_S$ and → ①

Average power dissipated per cycle $= \left[\frac{I_m}{\sqrt{2}}\right]^2 \cdot R_S$

Energy = power × time.

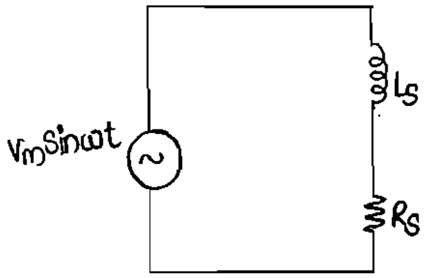
∴ Average energy dissipated in the inductor per cycle = power × periodic time for one cycle.

$= \text{power} \times T$

$= \left[\frac{I_m}{\sqrt{2}}\right]^2 R_S \times T$

$= \left[\frac{I_m}{\sqrt{2}}\right]^2 \times R_S \times \frac{1}{f} \qquad \left[\because T = \frac{1}{f}\right]$

$= \frac{I_m^2 R_S}{2f} \rightarrow$ ②



Circuit for obtaining equation of Q

Substituting values of equation ① and equation ② in the equation of Q. we have,

$$Q = 2\pi \frac{\frac{1}{2} I_m^2 L_S}{I_m^2 \frac{R_S}{2f}}$$

$$= \frac{2\pi f L_S}{R_S}$$

$$= \frac{\omega L_S}{R_S}.$$

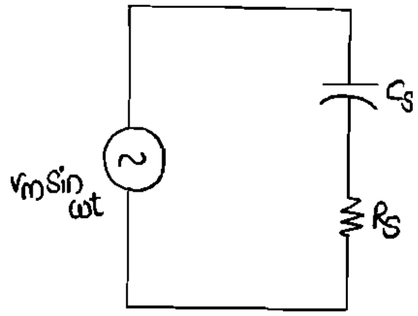
$$\therefore Q = \frac{\omega L_S}{R_S}.$$

* Derive the expression for Q-factor of a capacitor.

1. Capacitor with a small resistor in series:-

→ Consider a circuit shown in the fig.

Here, a sinusoidal voltage $V_m \sin\omega t$ is applied to the capacitor with a small resistor.



Circuit for obtaining equation for Q

Maximum energy stored in the capacitor per cycle

$$= \frac{1}{2} C_S V_{max}^2$$

where

$$V_{max} = \frac{I_m}{\omega C_S} \quad \text{when } R_S << \frac{1}{\omega C_S}.$$

∴ Maximum energy stored in the capacitor per cycle $= \frac{1}{2} CV_{max}^2 = \frac{I_m^2}{2\omega^2 C_S} \rightarrow (1)$

Energy dissipated per cycle $= \frac{I_m^2 R_S}{2f} \rightarrow (2)$

Substituting equations (1) and (2) in the equation Q we have,

$$Q = 2\pi \times \frac{\frac{I_m^2}{2\omega^2 C_S}}{\frac{I_m^2 R_S}{2f}} = 2\pi \times \frac{I_m^2}{2\omega^2 C_S} \times \frac{2f}{I_m^2 R_S}.$$

$$= \frac{2\pi \times I_m^2 \times 2f}{2\times(2\pi f)^2 C_S \times I_m^2 R_S}$$

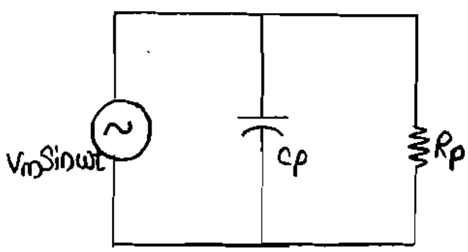
$$= \frac{1}{2\pi f \times C_S \times R_S} = \frac{1}{\omega C_S R_S}.$$

$$\therefore Q = \frac{1}{\omega C_S R_S}.$$

2. Capacitor with a high resistor in parallel:-

→ Consider a circuit shown in the fig.

Here, a sinusoidal voltage $V_m \sin\omega t$ is applied to the capacitor with high resistor in parallel.



Circuit for obtaining equation for Q

Maximum energy stored in the capacitor

$$= \frac{1}{2} CV_{max}^2 \rightarrow (1).$$

Average power dissipated per cycle in $R_p = \left[\frac{V_{max}}{\sqrt{2}}\right]^2 \times \frac{1}{R_p}$

$$= \frac{V_{max}^2}{2R_p}$$

Energy dissipated per cycle $= \frac{V_{max}^2}{2R_p} \times T = \frac{V_{max}^2}{2R_pf} \quad \left[\because T = \frac{1}{f}\right] \rightarrow (2)$

Substituting equations (1) and (2) in the equation of Q, we have

$$Q = 2\pi \times \frac{\frac{1}{2} C_p V_{max}^2}{\frac{V_{max}^2}{2R_pf}}$$

$$= 2\pi f C_p R_p = \omega C_p R_p.$$

$$\therefore Q = \omega C_p R_p.$$

* **Unloaded and loaded Q :-**

→ When the tank circuit is not connected to any external circuit (or) load, Q accounts for the internal losses and it is known as unloaded quality factor, Q_U. It is defined as,

$$Q_U = 2\pi \times \frac{\text{Maximum energy stored per cycle}}{\text{Energy dissipated per cycle in tank circuit}}$$

→ In practise, the tank circuit is connected to the load. Hence, the energy dissipation takes place in the tank circuit as well as in the external load. The loaded quality factor, Q_L is defined as

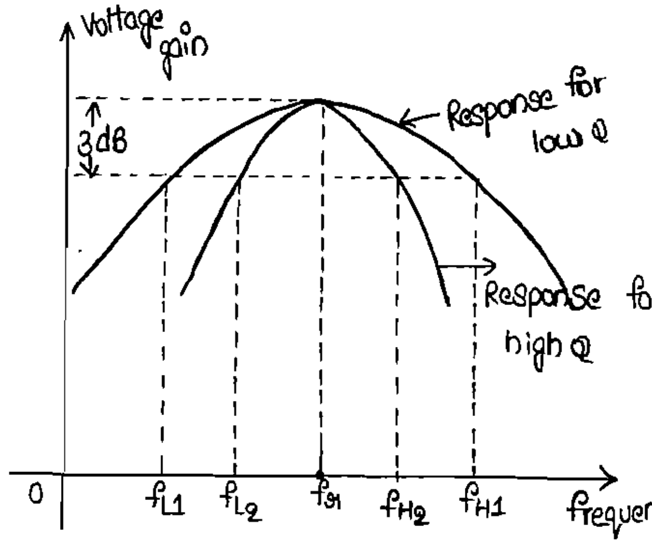
$$Q_L = 2\pi \times \frac{\text{Maximum energy stored per cycle}}{(\text{Energy dissipated per cycle in tank circuit} + \text{Energy dissipated per cycle due to presence of external load})}$$

The quality factor Q_L determines the 3dB bandwidth for the resonant circuit. The 3 dB band width for resonant circuit is given as,

$$B.W = \frac{f_{r}}{Q_L}.$$

where f_r represents the centre frequency of a resonator and BW represents the band width.

If Q is large, band width 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.



Variation of 3dB bandwidth with variation in quality factor

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 select band of frequencies.

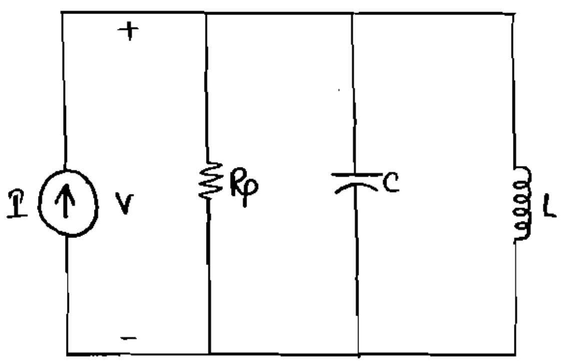
* **parallel Resonant circuit :-**

→ The fig. shows the tuned ||el LC circuit which resonates at a particular frequency.

The total admittance of the ||el tuned circuit is given by

$$Y_T = \frac{1}{R_p} + \frac{1}{1/j\omega C} + \frac{1}{j\omega L}.$$

$$= \frac{1}{R_p} + j\left[\omega C - \frac{1}{\omega L}\right] \rightarrow (1).$$



parallel resonant LC circuit

At resonance imaginary part is zero, thus equating it to zero, we get

$$\omega_0 C = \frac{1}{\omega_0 L} \quad \therefore \omega_0^2 LC = 1 \rightarrow ②$$

$$f_0 = \frac{1}{2\pi\sqrt{LC}} \rightarrow ③$$

In parallel resonant circuit, the voltage v is common to the three circuit elements, and we can write the maximum energy of the circuit interms of capacitance as $C\frac{V_m^2}{2}$. The energy loss per cycle is $\frac{\left[\frac{V_m^2}{2R_p}\right]}{f}$. Then Q is

$$Q = \frac{2\pi V_m^2 C/2}{V_m^2/2R_p f}$$

$$= \omega_0 R_p C$$

$$= \frac{R_p}{\omega_0 L} = R_p\sqrt{\frac{C}{L}} \rightarrow ④$$

Once the determined resonant condition by $\omega_0^2 LC = 1$, the value of Q of a resonant circuit is determined by R_p, or by the ratio of C to L.

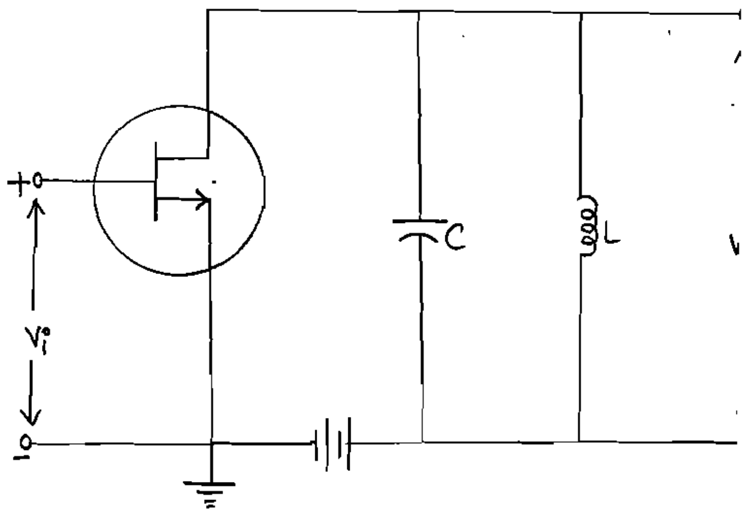
At resonance, reactive term is equal to zero,

$$\therefore Y_T = \frac{1}{R_p}.$$

$\therefore$ Impedance at resonance, $Z_0 = R_p \rightarrow ⑤$

using eq ④ and ⑤, we can write

$$Z_0 = Q_0 \omega_0 L = \frac{Q}{\omega_0 C}. \rightarrow ⑥$$



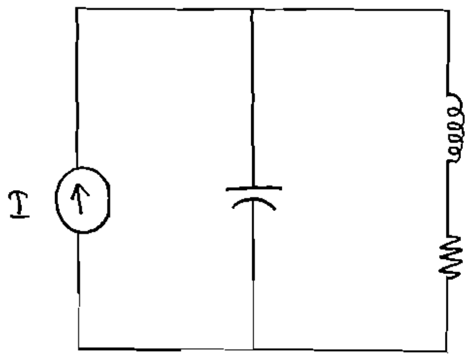
LC resonant circuit

The impedance of the resonant circuit is required in determining circuit gain. The gain of the circuit shown in the figure.

$$\therefore A_V = -g_m \cdot R_L = -g_m \cdot Q\omega_0 L \rightarrow ⑦$$

* Series Resonant circuit :-

→ The figure shows the series resonant circuit. Here, the loss element R_S is in series with L. The admittance of the R_SL series branch is



R in Series

$$Y = \frac{1}{R_S + j\omega L} = \frac{R_S - j\omega L}{R_S^2 + \omega^2 L^2} \rightarrow ①$$

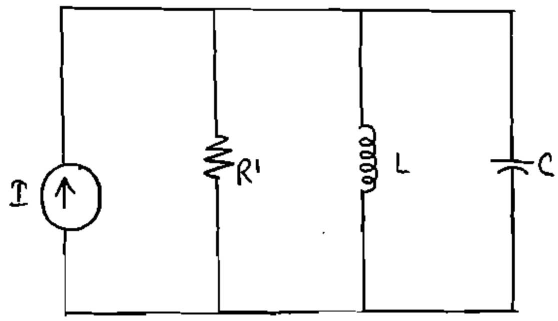
By multiplying numerator and denominator $R_S - j\omega L$.

usually, at high Q conditions, $\omega^2 L^2 >> R_S^2$.

Therefore, we can drop term R_S^2 in the denominator, we get

$$Y = \frac{R_S}{\omega^2 L^2} + \frac{1}{j\omega L}$$

$$= \frac{1}{R'} + \frac{1}{j\omega L} \rightarrow ②$$



R in parallel

The equation gives us the parallel arrangement as shown in fig. where R' is given by,

$$R' = \frac{\omega^2 L^2}{R_S} \rightarrow ③$$

$$\therefore R_S = \frac{\omega^2 L^2}{R'} \rightarrow ④$$

The equations ③ and ④ represent transformations for passing from the series form of circuit to the ||el form, or vice versa. The inductance L does not change in the transformations but a small series R_S transforms to a large R' in parallel with L.

In the previous section, we have seen that for parallel circuit is

$$Q = \omega_0 C R_p.$$

Here R_p is represented by R', $\therefore Q = \omega_0 C R' = \omega_0 C \frac{\omega_0^2 L^2}{R_S} \rightarrow ⑤$.

Since, $\omega_0^2 LC = 1$, we have

$$\therefore Q = \frac{\omega_0 L}{R_S} \rightarrow ⑥$$

* Requirements of Tuned amplifier :-

The basic requirements of tuned amplifiers are :-

1. The amplifier should provide selectivity of resonant frequency over a 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 are 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 amplifi components to the desired frequency over a considerable range or band of frequencies.

* Tuned circuits :-

→ At radio frequencies, iron core transformers are not used as eddy currents losses and hysteresis losses increase with frequency. Thus at radio frequency air core transformers are frequently used. As there is air path between windings, the leakage flux increases and coefficient of the coupling decreases.

In RF circuit design, tuned circuits are generally employed either for obtaining maximum power transfer to the load connected to secondary or for obtaining maximum possible value of secondary voltage.

They are two types of tuned circuits, namely:-

1. Single tuned circuit
2. Double tuned circuit.

Similarly let Q_2 be the quality factor of secondary circuit which is same as quality factor of secondary coil if R_2 is the coil resistance,

$$\therefore Q_2 = \frac{\omega_r L_2}{R_2} \rightarrow (15)$$

The coefficient of coupling is given by,

$$K = \frac{M}{\sqrt{L_1 L_2}}$$

$$L_1 L_2 \; K^2 = M^2$$

Substituting values of L_1 and L_2 from equation (14) and (15)

$$K^2 \left(\frac{Q_1 R_1}{\omega_r} \right) \left(\frac{Q_2 R_2}{\omega_r} \right) = M^2$$

$$K^2 (Q_1 Q_2) = \frac{\omega_r M^2}{R_1 R_2} \rightarrow (16)$$

Hence expression for maximum voltage is given by,

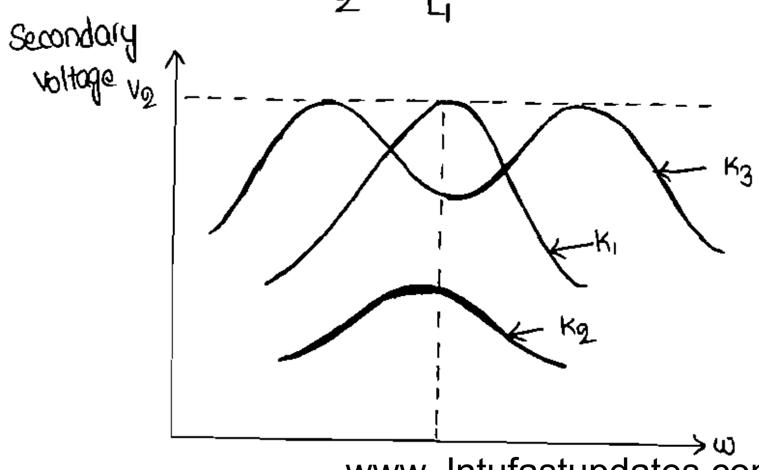
$$V_2 = \frac{K Q_1 Q_2 \sqrt{L_2/L_1} \; V_1}{K^2 Q_1 Q_2 + 1} \rightarrow (17)$$

In the above expression is maximised with respect to K, we get the condition for critical coupling.

$$K_C = \frac{1}{\sqrt{Q_1 Q_2}} \rightarrow (18)$$

Thus the maximum voltage under critical coupling is given by,

$$V_2 = \frac{Q_1 Q_2}{2} \frac{L_2}{L_1} V_1 \rightarrow (19)$$



$K_1 = K_C$ = Critical coupling

$K_2 < K_1$ = Sufficient coupling

$K_3 = K_1$ = over coupling

Hence, to get maximum secondary voltage with kritical coupling high values Q_1, Q_2 and L_2 are selected.

Under critical coupling secondary current is maximum, hence secondary voltage is also maximum. Frequency response of secondary voltage for different coupling coefficients as shown in the above fig.

* classification of tuned amplifiers :-

→ Multistage amplifiers are used to obtain large overall gain, the cascaded stage of multistage tuned amplifiers can be categorised as given below.

1. Single tuned amplifiers.
2. Double tuned amplifiers.
3. Sttagger tuned amplifiers.

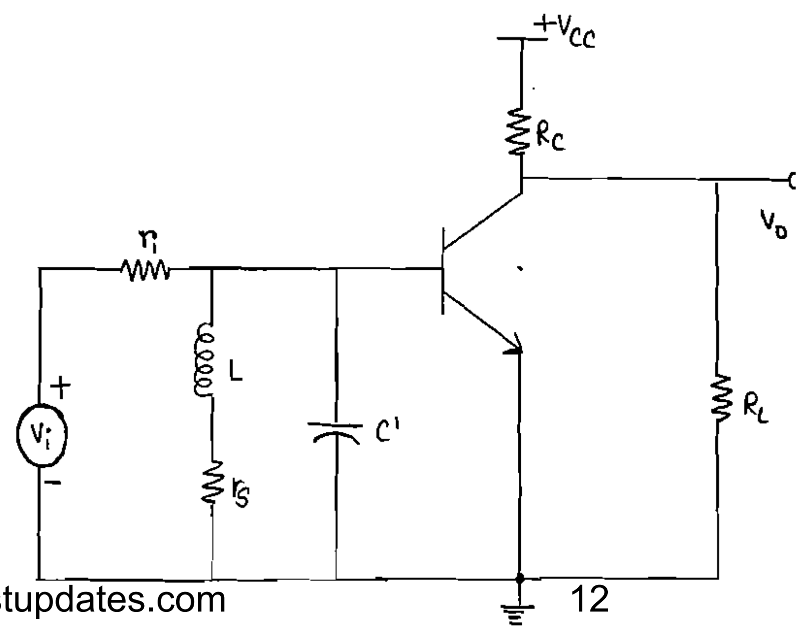
These amplifiers are further classified according to coupling used to cascade the stage of multistage amplifier.

1. capacitive coupled.
2. Inductive coupled.
3. Transformer coupled.

* Small signal tuned amplifiers :-

A common emitter amplifier can be converted into a single tuned amplifier by including a parallel tuned amplifier as shown in fig.

The biasing components are not shown for simplicity.

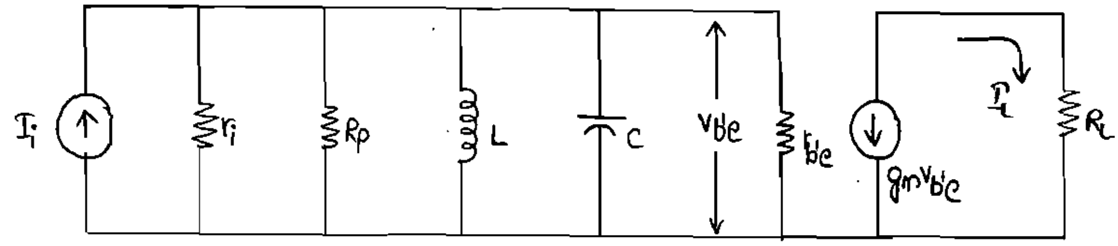


Assumptions :-

1. $R_L << R_C$

2. $r_{bb'} = 0$

with these assumptions, the simplified equivalent circuit for a single tuned amplifier as shown in the fig.



Equivalent circuit for single tuned amplifier

where $C = C' + C_{b'e} + (1+g_m R_L) C_{b'c}$.

C' = external capacitance used to tune the circuit

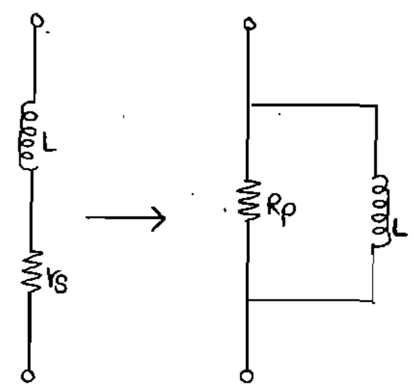
$(1+g_m R_L) C_{b'c}$:- The miller capacitance

r_s = represents the losses in the coil.

The series RL circuit in the 1st fig is replaced by the equivalent RL circuit in above fig assuming coil losses are low over frequency band of interest, i.e., the coil Q is high.

$$Q_c = \frac{\omega L}{r_s} >> 1 \longrightarrow (1)$$

The conditions for equivalence are most easily established by equating the admittances of the two circuits as shown in figure.



equivalent circuits

$$Y_1 = \frac{1}{r_s + j\omega L} = \frac{r_s - j\omega L}{r_s^2 + (j\omega L)^2} = \frac{r_s - j\omega L}{r_s^2 + \omega^2 L^2}$$

$$= \frac{r_s}{r_s^2 + \omega^2 L^2} - \frac{j\omega L}{r_s^2 + \omega^2 L^2}$$

$\therefore \omega L >> r_s$ from eq (1).

$$Y_1 = \frac{r_s}{\omega^2 L^2} + \frac{1}{j\omega L}.$$

$$y_2 = \frac{1}{R_p} + \frac{1}{j\omega L}.$$

$\therefore$ By equating y_1 and y_2, we get

$$\frac{r_s}{\omega^2 L^2} + \frac{1}{j\omega L} = \frac{1}{R_p} + \frac{1}{j\omega L}.$$

$$\therefore \frac{1}{R_p} = \frac{r_s}{\omega^2 L^2} = \frac{r_s^2}{r_s \omega^2 L^2} = \frac{1}{r_s Q_s^2}.$$

$$\therefore R_p = r_s Q_c^2 = \omega L Q_c \longrightarrow (2) \qquad \because \omega L = Q_c r_s \text{ from eq (1)}$$

looking at equivalent circuit, we have

$$\therefore R = r_i \| R_p \| r_{b'e} \longrightarrow (3)$$

The current gain of the amplifier is then

$$A_i = \frac{-g_m R}{1+j(\omega R C - R/\omega L)} = \frac{-g_m R_k}{1+j\omega_0 R_C(\omega/\omega_0 - \omega_0/\omega)} \longrightarrow (4) \quad \text{where } \omega_0^2 = \frac{1}{LC}.$$

we define the Q of the tuned circuit at the resonant frequency ω_0 to be

$$Q_i = \frac{R}{\omega_0 L} = \omega_0 RC \rightarrow (5)$$

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

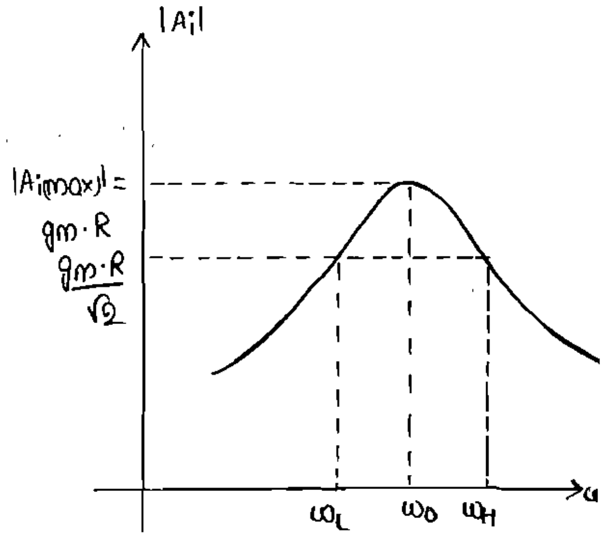
At $\omega = \omega_0$, gain is maximum and it is given by

$$\therefore A_{i(max)} = -g_m \cdot R \longrightarrow (6)$$

These fig shows the gain vs frequency plot for single tuned amplifier. It shows the variation of the magnitude of gain as a function of frequency.

At 3 dB frequency,

$$|A_i| = \frac{g_m \cdot R}{\sqrt{2}} \rightarrow (7)$$



gain vs frequency for single tuned amplifier

$\therefore$ At 3dB frequency,

$$1+jQ_i\left[(\omega/\omega_0)-(\omega_0/\omega)\right]=\sqrt{2}$$

$$\therefore \quad 1+Q_i^2\left[\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega}\right]^2=2 \quad \rightarrow (8)$$

This equation is quadratic in ω^2 and has two positive solution, ω_H and ω_L. As After solving equation (8), we get 3dB bandwidth as given below,

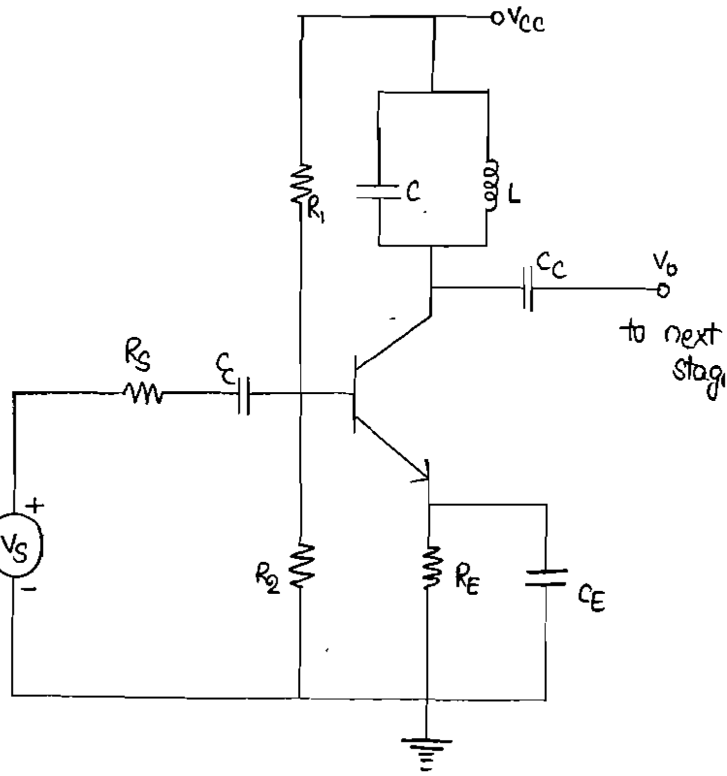
$$B.W = f_H - f_L = \frac{\omega_0}{2\pi Q_i} = \frac{1}{2\pi RC} \rightarrow (9)$$

$$\therefore B.W = \frac{1}{2\pi RC}.$$

* 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 the same frequency in fig shows a typical single tuned amplifier in CE configuration.

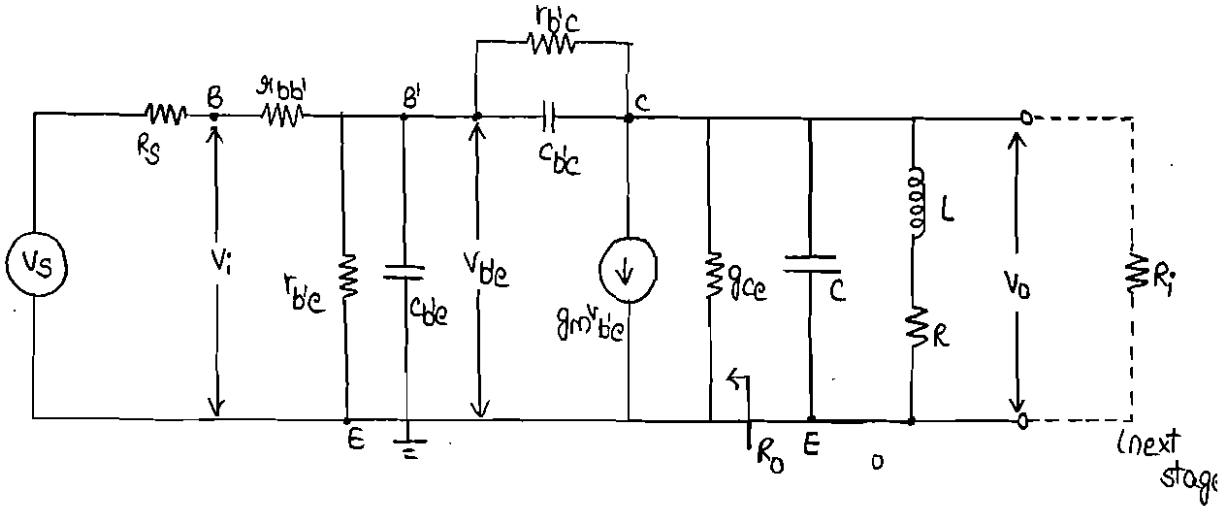
As these circuit, tuned circuit formed by L and C acts as collector load and resonates at frequency of operation. Resistors R_1, R_2 and R_E along with capacitor C_E provides self bias for the circuit.



Single tuned capacitive coupled transformer amplifier

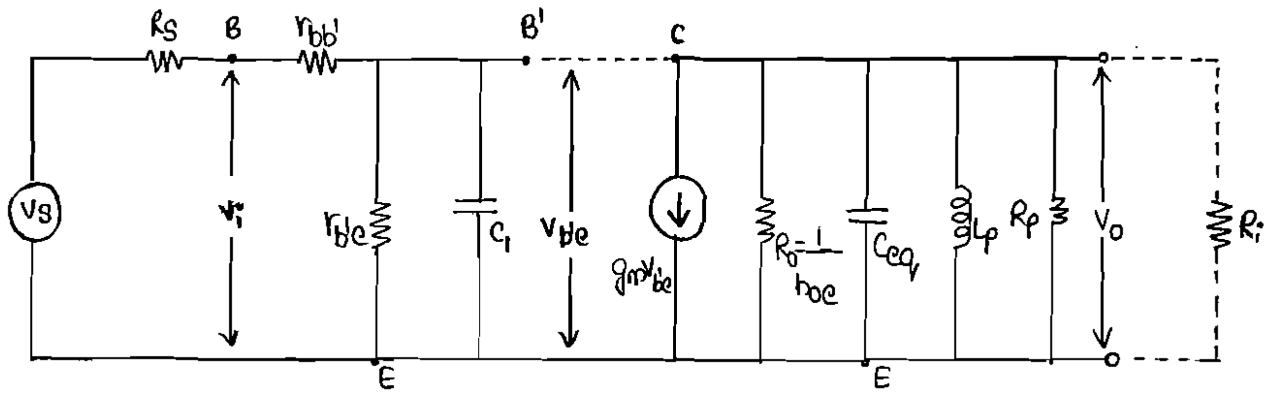
The below figure shows the equivalent circuit for single tuned amplifier using hybrid π parameters.

As shown in the below fig. R_i is the input resistance of the next stage and R_o is the output resistance of the current generator $g_m V_{b'e}$. The reactances of the bypass capacitor C_E and the coupling capacitors C_C are negligible small at the operating frequency at hence these elements are neglected in the equivalent circuit in the below fig.



Equivalent circuit of single tuned amplifier

The equivalent circuit shown in the above fig. can be simplified by applying Miller's theorem. Below fig shows the simplified equivalent circuit for single tuned amplifier.



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,

$$C_i = C_{b'e} + C_{b'c}(1-A) \longrightarrow ①$$

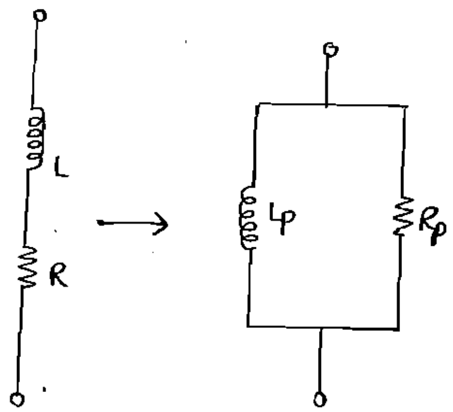
where A is the voltage gain of the amplifier

$$C_{eq} = C_{b'c}\left(\frac{A-1}{A}\right)+C$$ where C is the tuned circuit capacitance $\rightarrow$ (2)

The g_{ce} is represented as the output resistance of current generator $g_m v_{b'e}$.

$$g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} = h_{oe} = \frac{1}{R_o} \rightarrow (3)$$

The Series RL circuit is represented by its equivalent parallel circuit. The condition for equivalence are most easily established by equating the admittance of the two circuits as shown in fig.



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

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

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

$$= \frac{1}{R_p} + \frac{1}{j\omega L_p} \rightarrow (4)$$ where $R_p = \frac{R^2+\omega^2L^2}{R}$

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

centre frequency :-

The centre frequency or resonant frequency is given as,

$$f_r = \frac{1}{2\pi\sqrt{L_pC_{eq}}} \rightarrow (6)$$

where $$L_p = \frac{R^2+\omega^2L^2}{\omega^2 L} \rightarrow (4)$$

and $C_{eq} = C_{b'c}\left[\frac{A-1}{A}\right] + C \longrightarrow (7)$
$= C_0 + C$

$\therefore C_{eq}$ 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

$$Q_r = \frac{\omega_r L}{R} \longrightarrow (8)$$

Here ω_r = centre frequency or resonant frequency.

This quality factor is also called unloaded Q. But in practise, 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 $\omega L >> R$ in the frequency range of operation.

From eq (4), we have

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

As $\frac{\omega^2 L^2}{R} >> 1$, $R_p = \frac{\omega^2 L^2}{R} \longrightarrow (9)$

From eq (5), we have

$$L_p = \frac{R^2 + \omega^2 L^2}{\omega^2 L} = \frac{R^2}{\omega^2 L} + L$$
$$\approx L \qquad \because \omega L >> R \longrightarrow (10)$$

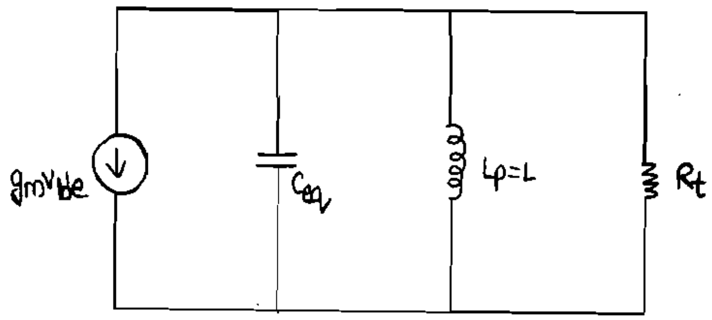
From eq (9), we can express R_p at resonance as,

$$R_p = \frac{\omega_r^2 L^2}{R}$$
$$= \omega_r Q_r L \qquad \because Q_r = \frac{\omega_r L}{R} \longrightarrow (11)$$

$\therefore Q_r$ can be expressed interms of R_p as,

$$Q_r = \frac{R_p}{\omega_r L} \longrightarrow (12)$$

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



where $R_t = R_o \| R_p \| R_i$

Simplified output circuit for single tuned amplifier

$$\text{effective quality factor } Q_{eff} = \frac{\text{Susceptance of inductance } L \text{ (or) capacitance } C}{\text{conductance of shunt resistance } R_t}$$

$$= \frac{R_t}{\omega_r L} \text{ (or) } \omega_r C_{eq} R_t \longrightarrow (13)$$

Voltage gain (A_v):-

The voltage gain for single tuned amplifier is given by,

$$A_v = -g_m \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \times \frac{R_t}{1 + 2jQ_{eff}\delta}$$

where $R_t = R_o \| R_p \| R_i$

δ = fraction variation in the resonant frequency

$$A_v(\text{at resonance}) = -g_m \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \times R_t.$$

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

3 dB bandwidth :-

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

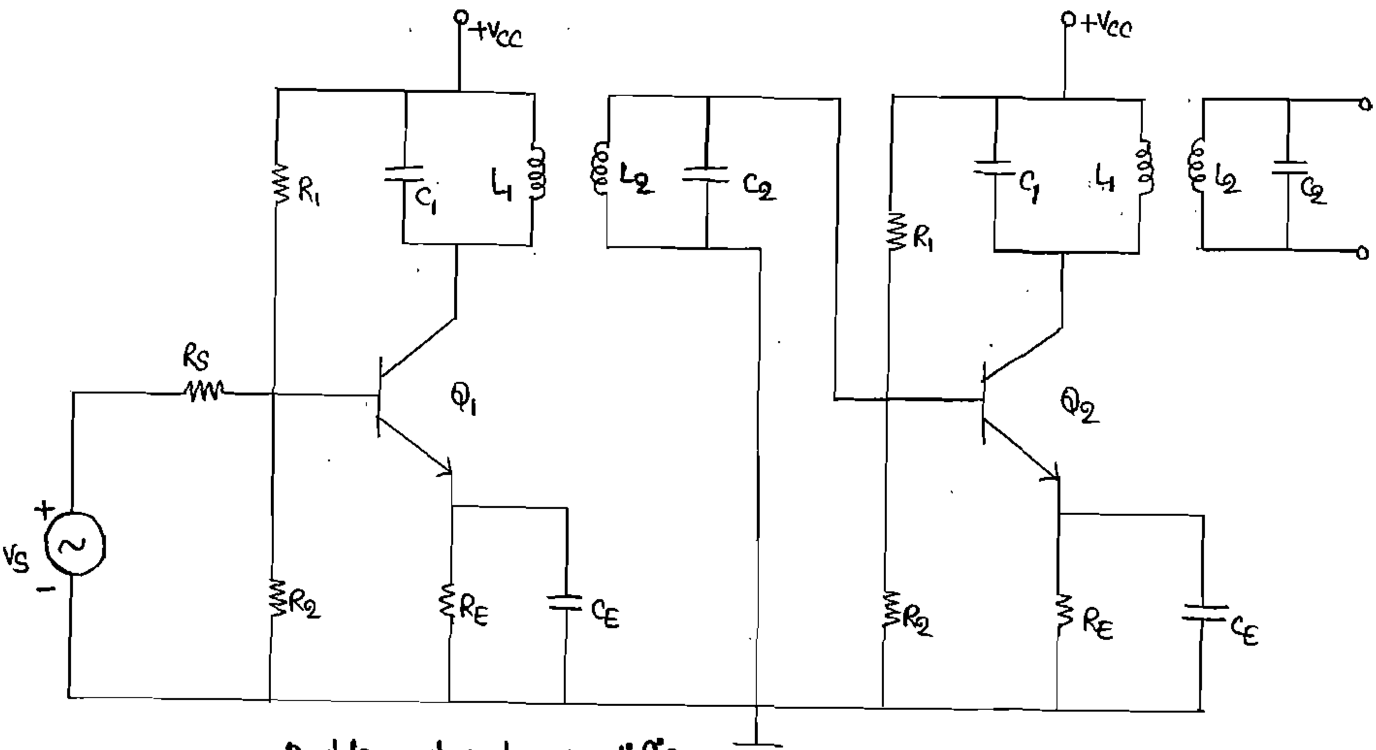
$$\Delta f = \frac{1}{2\pi R_t C_{eq}}$$

$$= \frac{\omega_r}{2\pi Q_{eff}} \longrightarrow (15) \quad \because Q_{eff} = \omega_r R_t C_{eq}.$$

$$= \frac{f_r}{Q_{eff}} \longrightarrow (16) \quad \because \omega_r = 2\pi f_r.$$

* **Double tuned amplifier :-**

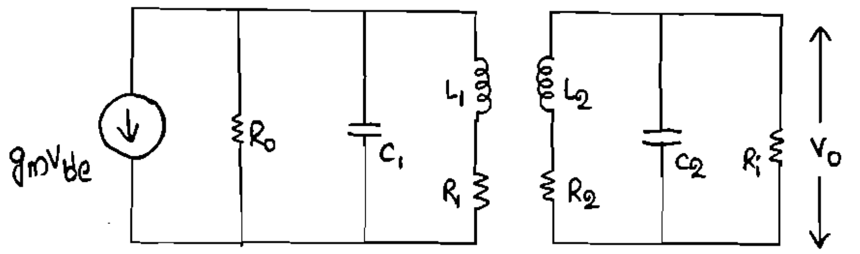
From the below fig. shows the double tuned amplifier in CE configuration. Here, voltage developed across tuned circuit is coupled inductively to another tuned circuit. Both tuned circuits are tuned to the same frequency.

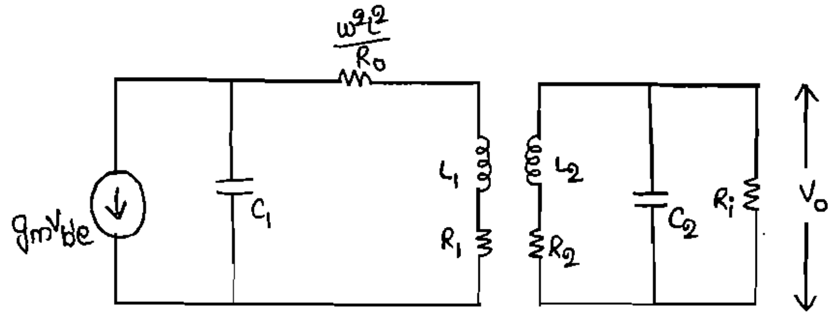


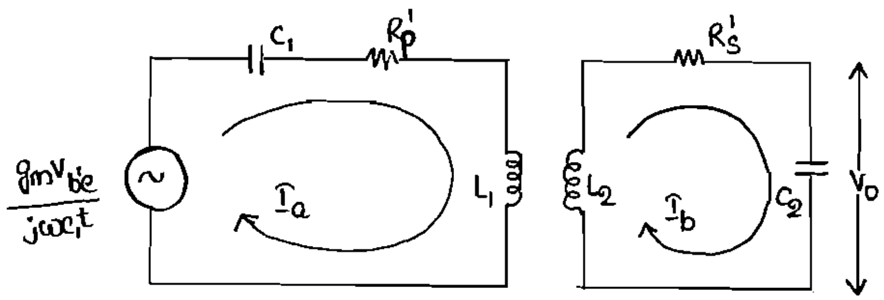
Double tuned amplifier

The double tuned circuit can provide a bandwidth of several percent of the resonant frequency and gives step sides to the response curve. Let us analyze the double tuned circuit.

Equivalent circuit for the o/p stage.







where

$$R_p' = \frac{\omega^2 L^2}{R_o} + R_1 \quad , \quad R_s' = R_2 \| R_i$$

Apply KVL to the input and output loops.

$$\frac{-g_m V_{b'e}}{j\omega c_1} = Z_p \cdot I_a + j\omega m I_b \rightarrow (1)$$

$$0 = j\omega m I_a + Z_s I_b \longrightarrow (2)$$

where $Z_p = R_p' + j\omega L_1 + \frac{1}{j\omega C_1}$

$$= R_p' + j\left[\omega L_1 - \frac{1}{\omega C_1}\right]$$

$$= R_p'\left[1 + j\left[\frac{\omega L_1}{R_p'} - \frac{1}{\omega C_1 R_p'}\right]\right]$$

$$= R_p'\left[1 + j\left(\frac{\omega_0}{\omega_0}\frac{\omega L_1}{R_p'} - \frac{\omega_0}{\omega_0}\frac{1}{\omega C_1 R_p'}\right)\right]$$

$$= R_p'\left[1 + j\left(\frac{\omega}{\omega_0}Q_1 - \frac{\omega_0}{\omega}Q_1\right)\right]$$

where $Q_1 = \frac{\omega_0 L_1}{R_p'} = \frac{1}{\omega_0 C_1 R_p'}$

$$Z_p = R_p'\left[1+jQ_1\left(\frac{\omega}{\omega_0}-\frac{\omega_0}{\omega}\right)\right]$$

$$= R_p'\left[1+jQ_1 \frac{\delta(\delta+2)}{\delta+1}\right].$$

where δ is the fractional change at resonant frequency is given by,

$$\delta = \frac{\omega-\omega_0}{\omega_0} = \frac{f-f_0}{f_0}$$

In eq ②, $Z_S = R_S' + j\omega L_2 + \frac{1}{j\omega C_2}$

Ily $Z_S = R_S'\left[1+jQ_2 \frac{\delta(\delta+2)}{\delta+1}\right]$

usually $\delta = \frac{f-f_0}{f_0} << 1$ then $\frac{\delta+2}{\delta+1} = 2$.

$$Z_p = R_p'\left[1+j2\delta Q_1\right] \rightarrow ③$$

$$Z_S = R_S'\left[1+j2\delta Q_2\right] \rightarrow ④$$

Solving ① and ② for I_b

From ②,

$$0 = j\omega M I_a + Z_S I_b$$

$$-j\omega M I_a = Z_S I_b$$

$$I_a = \frac{-Z_S I_b}{j\omega M}.$$

Substituting I_a value in eq ①

$$\frac{-g_m V_{b'e}}{j\omega C_1} = \frac{-Z_p Z_S I_b}{j\omega M} + j\omega M I_b$$

$$= I_b\left[\frac{-Z_p Z_S}{j\omega M} + j\omega M\right]$$

$$I_b = \frac{\frac{-g_m V_{b'e}}{j\omega C_1}}{j\omega M - \frac{Z_p Z_S}{j\omega M}} = \frac{-g_m V_{b'e}/j\omega C_1}{\frac{-\omega^2 M^2 - Z_p Z_S}{j\omega M}} = \frac{g_m V_{b'e} M}{C_1\left[\omega^2 M^2 + Z_p Z_S\right]}$$

from eq ③ and ④,

$$I_b = \frac{g_m V_{b'e} M}{C_1[R_p' R_s (1+j2\delta Q_1)(1+j2\delta Q_2) + \omega^2 M^2]}$$

$$I_b = \frac{g_m V_{b'e} M}{C_1[R_p' R_s \{1+2j\delta(Q_1+Q_2) - 4\delta^2 Q_1 Q_2\} + \omega^2 M^2]} \rightarrow ⑤$$

from the output loop, we have

$$V_0 = I_b \times C_2$$

$$= I_b \times \frac{1}{j\omega C_2}.$$

$$V_0 = \frac{g_m V_{b'e} M}{j\omega C_1 C_2[R_p' R_s \{1+2j\delta(Q_1+Q_2) - 4\delta^2 Q_1 Q_2\} + \omega^2 M^2]}$$

The voltage gain $A_V = V_0/V_i = V_0/V_{b'e}$

when $\delta = 0$ at resonance

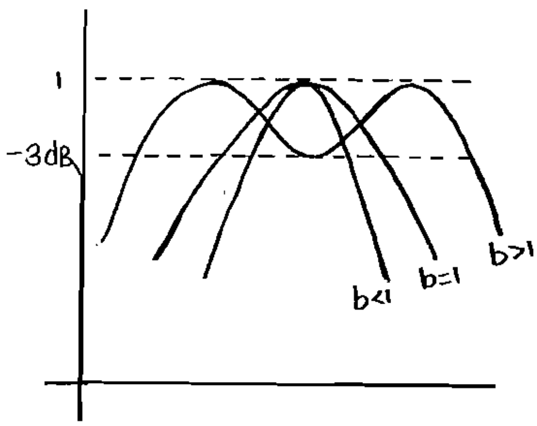
$\omega = \omega_0$ & $A_V = A_{V(max)}$

$A_{V_{max}} = A_{V_{resonance}}$.

$$\frac{A_V}{A_{V(resonance)}} = \frac{R_p' R_s' + \omega^2 M^2}{R_p' R_s' [1+2j\delta(Q_1+Q_2) - 4\delta^2 Q_1 Q_2] + \omega^2 M^2}$$

$$= \frac{1 + \frac{\omega^2 M^2}{R_p' R_s'}}{1 + 2j\delta(Q_1+Q_2) - 4\delta^2 Q_1 Q_2 + \frac{\omega^2 M^2}{R_p' R_s'}}$$

$$= \frac{1}{\frac{1+2j(Q_1+Q_2)\delta - 4\delta^2 Q_1 Q_2 + b^2}{1+b^2}}$$



where $b = \frac{\omega M}{\sqrt{R_p' R_s'}}$.

Gain band width product :- $G.B.W = A_{V_{resonance}} + B.W$

$$= \frac{g_m W}{j\omega C_1 C_2 [R_p' + R_s' + \omega^2 M^2]} \times \frac{1}{2\pi R_C}.$$

Advantages :-

Compared to the single tuned amplifier, the double tuned amplifier has the following advantages.

1. possesses a flatter response having steeper sides.
2. provides larger 3 dB bandwidth.
3. provides large gain bandwidth product.

* Effect of cascading single tuned amplifier on band width :-

→ In order to obtain a high overall gain, several identical stages of tuned 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 circuit direct coupled amplifiers connected in cascade. we know that the relative gain of a single tuned amplifier with respect to the gain at resonant frequency f_r in the equation from the

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

∴ 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

$$\left|\frac{A_v}{A_v(\text{at resonance})}\right|^n = \frac{1}{\sqrt{2}}.$$

$$\therefore \left|\frac{A_V}{A_V(\text{at resonance})}\right|^n = \frac{1}{[1+(2\delta Q_{eff})^2]^{n/2}} = \frac{1}{\sqrt{2}}$$

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

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

$$\therefore 1+(2\delta Q_{eff})^2 = 2^{1/n}.$$

$$\therefore 2\delta Q_{eff} = \pm\sqrt{2^{1/n}-1}.$$

Substituting for δ, the fractional frequency variation.

i.e., $\delta = \frac{\omega-\omega_r}{\omega_r} = \frac{f-f_r}{f_r}$

$$\therefore 2\left[\frac{f-f_r}{f_r}\right]Q_{eff} = \pm\sqrt{2^{\frac{1}{n}}-1}$$

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

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

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

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

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

The bandwidth of n stage amplifiers is given as,

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

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

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

$$= BW_1\sqrt{2^{1/n}-1}.$$

where BW_1 is the bandwidth of single stage.

BW_n is the bandwidth of n stages.

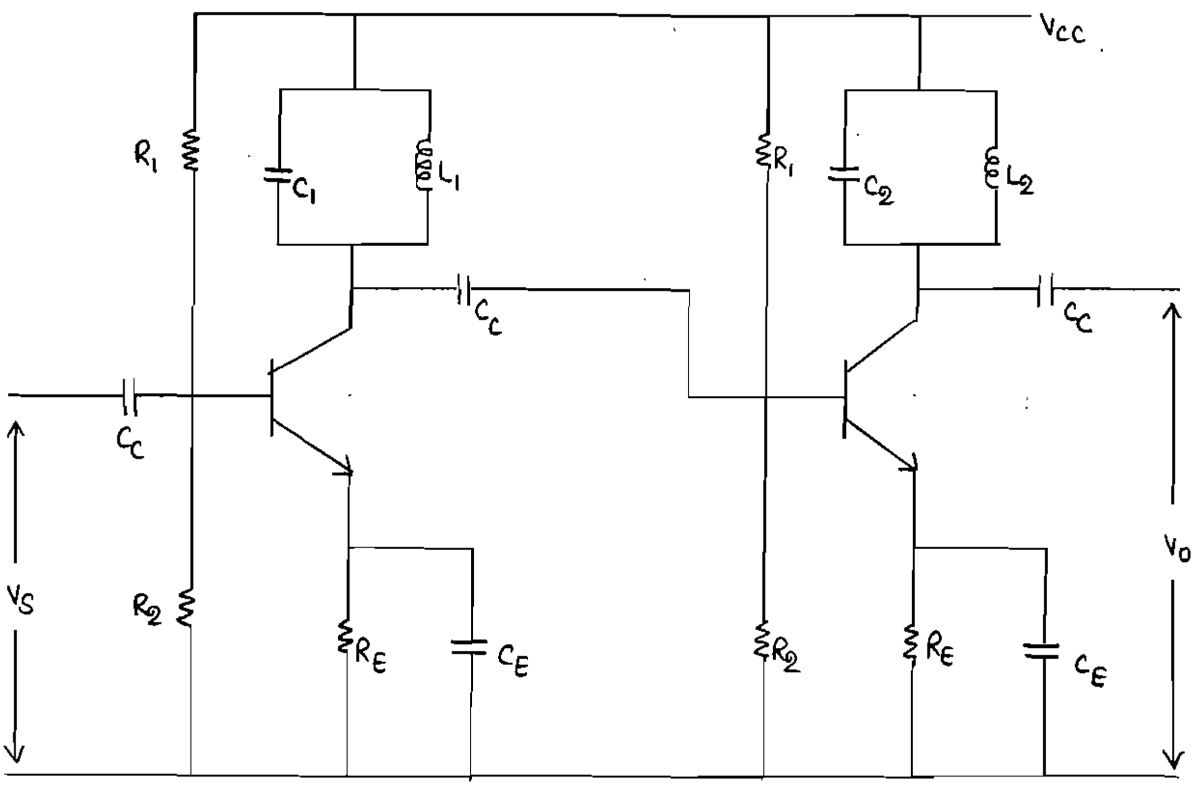
* Effect of cascading Double tuned Amplifiers on Bandwidth :-

→ When a number of identical double tuned amplifier stages are connected in cascade, the overall bandwidth of the system is there by 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 3dB bandwidth of n identical double tuned critically coupled stages compared with the bandwidth Δ_2 of such a system can be shown to be 3 dB bandwidth for

$$N \text{ identical stages double tuned amplifiers} = \Delta_2 \times \left[2^{\frac{1}{n}} - 1\right]^{1/4} \rightarrow (1)$$

where Δ_2 = 3 dB bandwidth of single stage double tuned amplifier.

* Staggered tuned amplifiers :-



The gain of single tuned amplifier as,

$$\frac{A_V}{A_V(\text{at resonance})} = \frac{1}{1+2j\delta Q_{eff}} = \frac{1}{1+jX} \quad \text{where } X = 2Q_{eff}\,\delta.$$

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

$$f_{r_1}=f_r+\delta,\quad f_{r_2}=f_r-\delta.$$

~~The overall gain~~

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

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

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

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

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

$$=\frac{1}{1+j(X+1)}\times\frac{1}{1+j(X-1)}=\frac{1}{2+2jX-X^2}=\frac{1}{(2-X^2)+(2jX)}$$

$$\therefore \left|\frac{A_v}{A_v(\text{at resonance})}\right|_{\text{cascaded}}=\frac{1}{\sqrt{(2-X^2)^2+(2X)^2}}$$

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

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

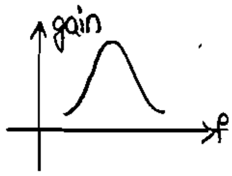
Substituting the value of X, we get

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

$$=\frac{1}{\sqrt{4+16Q_{eff}^4\delta^4}}$$

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

* Comparision between tuned circuits:-

S.No	parameter	Single tuned circuit	Double tuned circuit	Stagger tuned circuit
1.	No. of tuned circuits	one	Two	More than two
2.	Q-factor	High	High	moderate low
3.	Selectivity	very high	moderate	low
4.	Band width	Small	Moderate	high
5.	frequency response vs gain			
6.	Application	RF amplifier stage in radio receivers	IF amplifier stage in radio receiver	Band pass filter

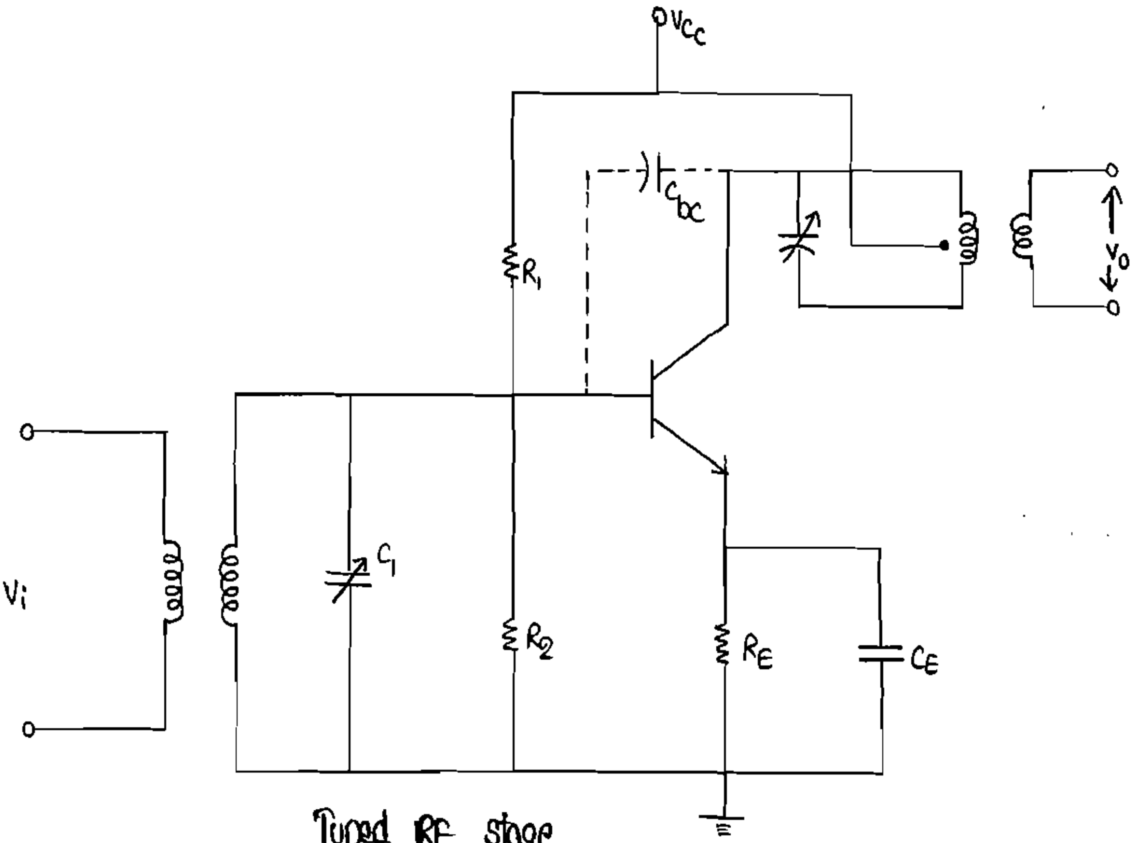
* Comparision between Synchronously and stagger tuned amplifier:-

Synchronously tuned amplifier	Stagger tuned amplifier
1. Several identical stages of single tuned amplifier are connected in cascade & tuned to same frequency.	1. Single tuned cascaded amplifiers with their resonant frequencies are seperated by an amount equal to bandwidth of each stage.
2. The overall bandwidth is lower than that of the single tuned circuit alone.	2. The overall bandwidth is larger than that of the single tuned circuit alone.
3. The magnitude of the gain is more as compared to stagger tuned amplifier.	3. The magnitude of the gain is less than that of the synchronously tuned circuit.
4. The passband is narrow.	4. The passband is wider and flatter than passband of synchronously tuned amplifier and thus provides a better approximation to the ideal band pass.

This is called Synchronous tuning and amplifier is Synchronously tuned amplifier.

The overall bandwidth of the Synchronously tuned amplifier is lower than that of the sigle tuned circuit alone.

* Stability of tuned amplifiers :-



Tuned RF stage

* Hazeltine Neutralisation :-

→ The below fig shows one variation of 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.

There fore, 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}.

* Advantages and disadvantages of tuned circuits :-

Advantages :-

→ They amplify defined frequencies.

→ Signal to noise ratio at output is good.

→ They are well suited for radio transmitters and receivers.

→ 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.

* Applications :-

The important applications of tuned amplifiers are as follows :-

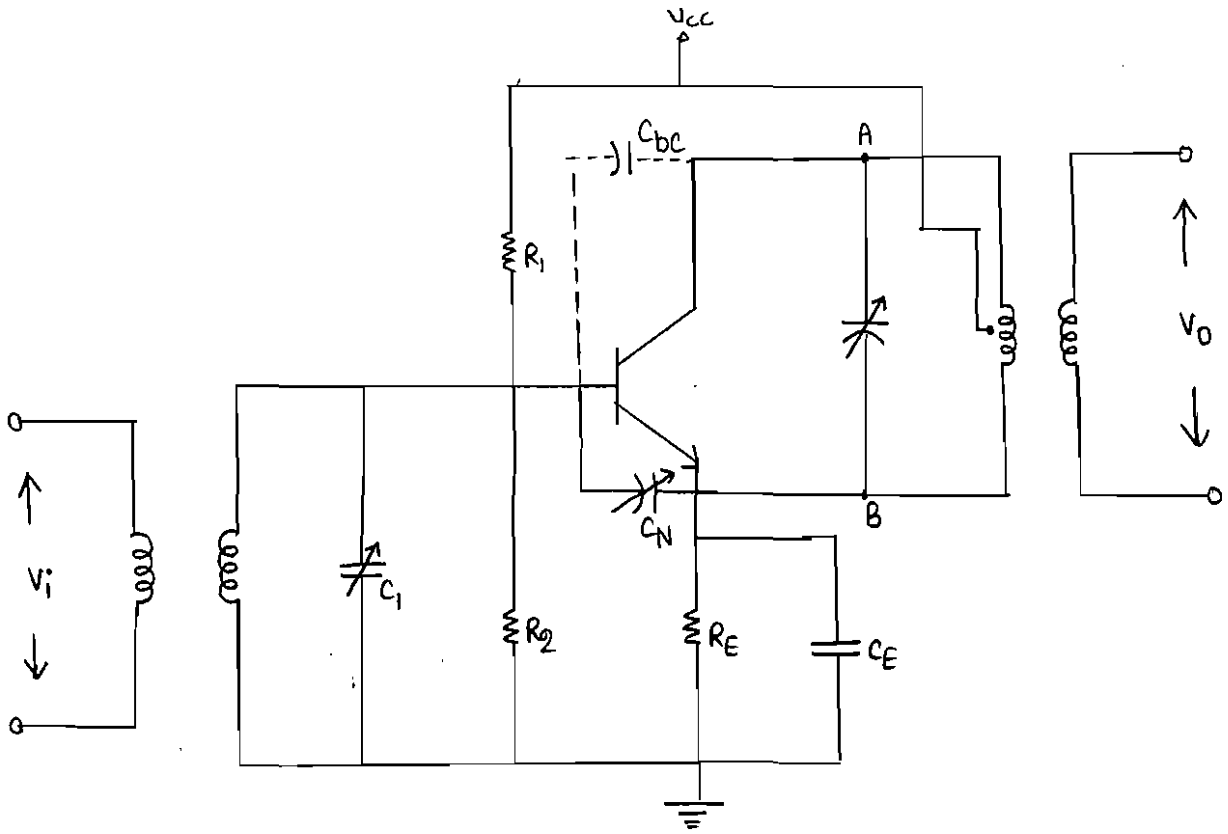
→ Tuned amplifiers are used in radio receivers to simplify a particular band of frequencies for which the radio receiver is tuned.

→ Tuned class B and class C amplifiers are used as an output of RF amplifiers in radio transmitters to increase the output efficiency and to reduce the harmonics.

→ 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.

* Synchronously tuned amplifier :-

→ In order to obtain a high overall gain, several identical stages of tuned amplifiers can be used in cascade. The overall gain is the product of the voltage gain of the individual stages. All amplifier stages are assumed to be identitical to be tuned to the same frequency ω_0

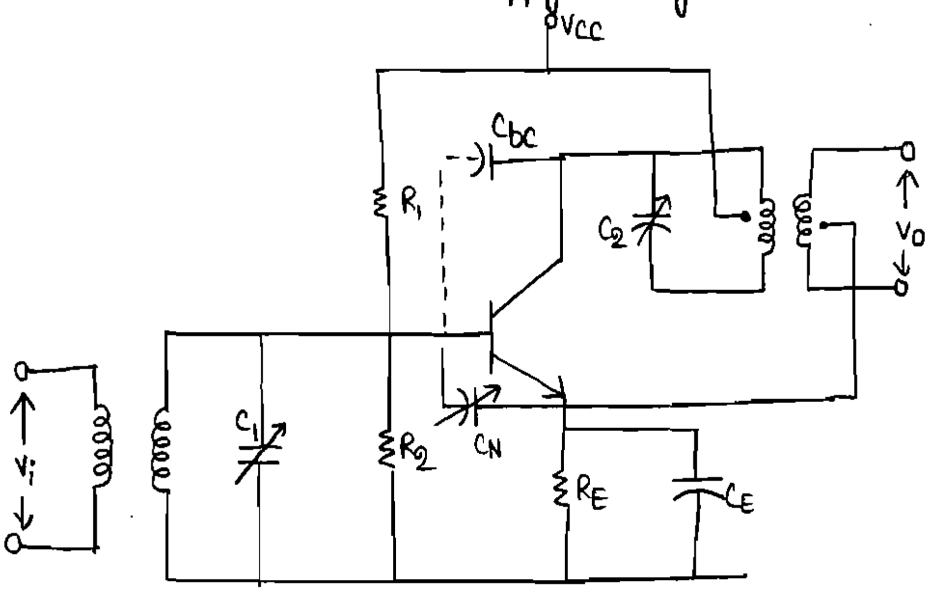


Tuned RF amplifier with Hazeltine neutralization

* Neutrodyne Neutralization :-

→ The below fig 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.

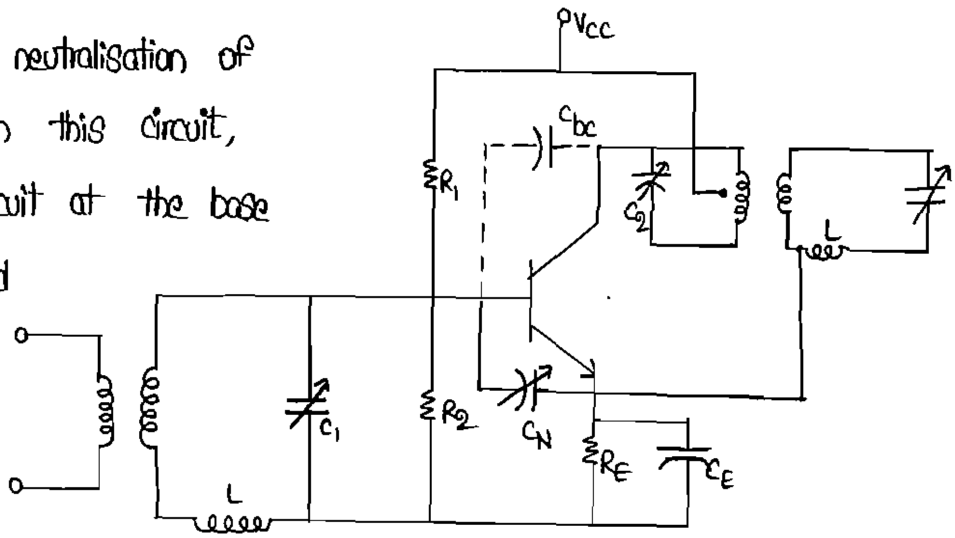
In principle, the circuit functions in the same manner as the Hazeltine neutralization circuit with the advantage that the neutralizing capacitor does not have the supply voltage across it.



Tuned RF amplifier with Neutrodyne neutralisation

* ## Neutralization using coil:-

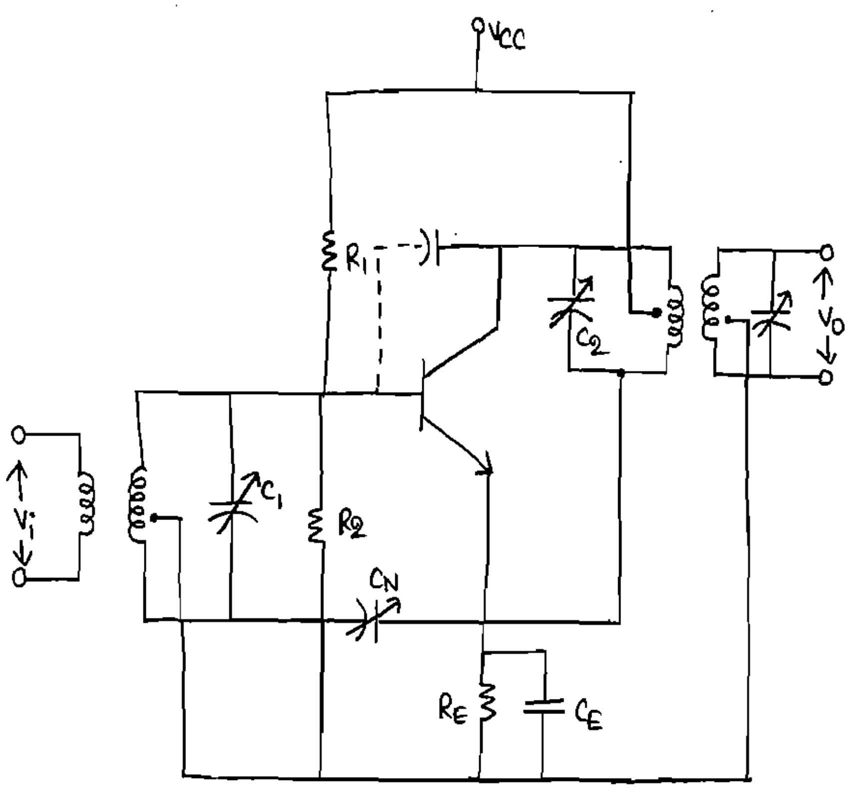
These fig shows the neutralisation of RF amplifier using coil. In this circuit, L part of the tuned circuit at the base of next stage is oriented for minimum coupling to the other windings. It is wound on the seperate from and mounted at right angles to the coupled windings. If the windings are properly 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.



Tuned RF amplifier using coil

* ## Rice Neutralization:-

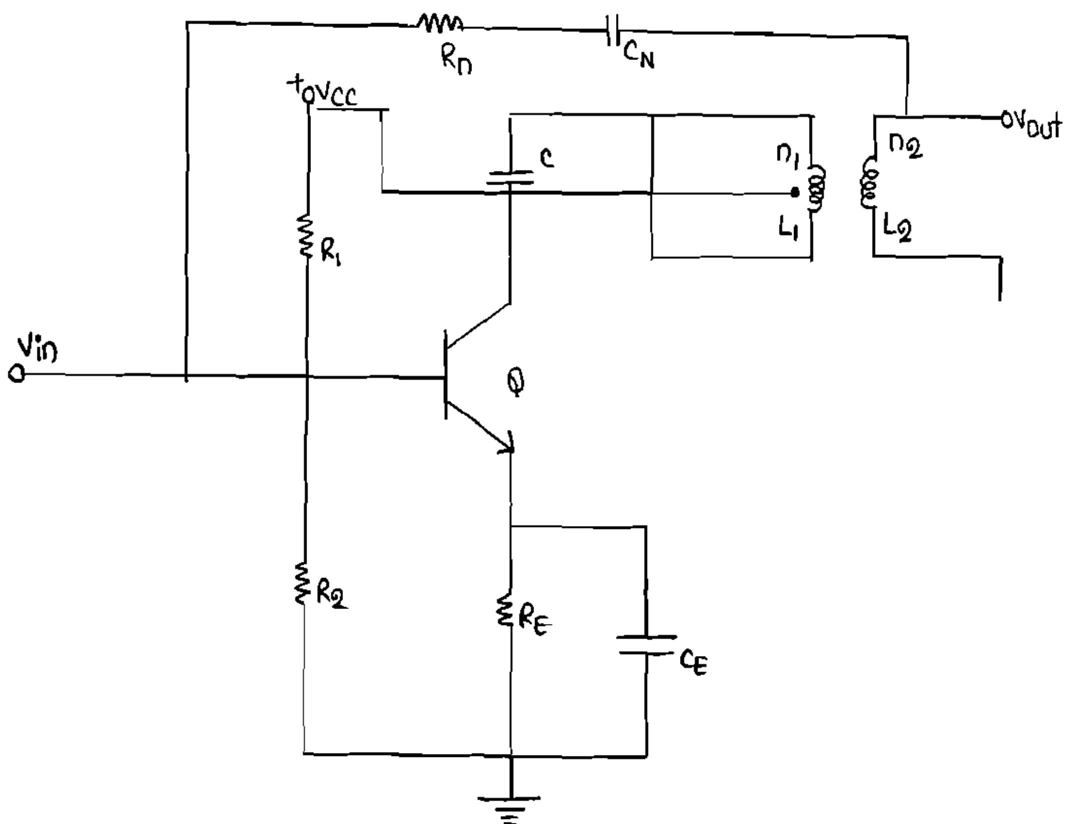
This figure shows the Rice circuit of neutralization. It uses a center of tapped coil in the base circuit. with this arrangement the signal voltages at the ends of the tuned base coil are equal and out of phase.



Tuned RF using Rice Neutralisation

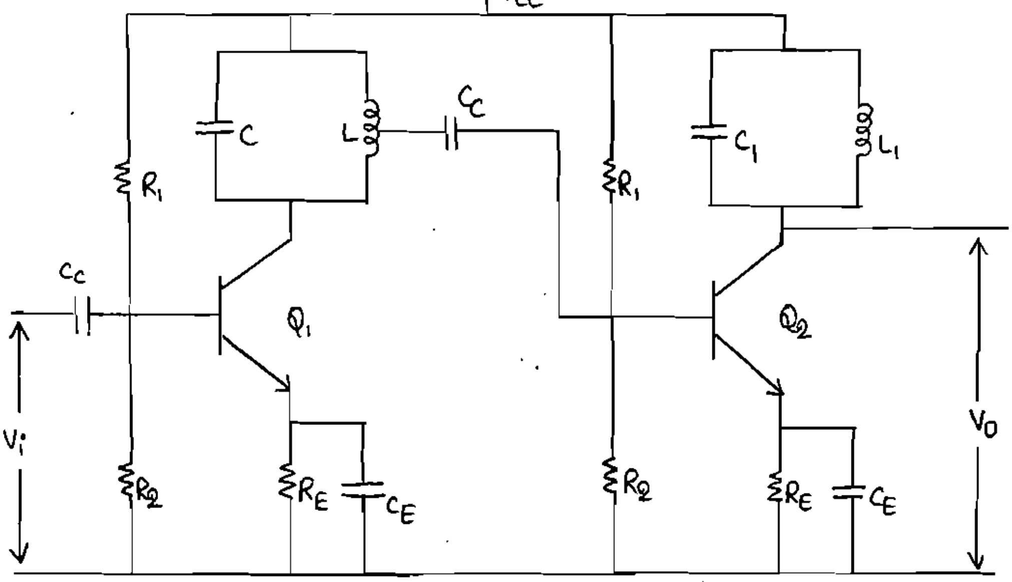
* Mismatch technique :-

The load impedance of a tuned amplifier is usually very small and hence a large current flows through this impedance. This results in a very little feedback to the input through the neutralising capacitance C_n. However, we require high impedance to be presented to the tank ckt in order to achieve a high Q value. Thus, there is a mismatch between the output of the transistor and tuned circuit. The step up transformer is used to solved this problem as shown in the figure.



This mismatch technique improves stability. The advantage is that stability is achieved at all the frequencies unlike neutralising techniques were the stability is achievable over a very small band of frequencies. This technique also decreases the alignment problem of various stages. Each tank circuit can be tuned without must effect on adjacent stages when the combination of mismatch and neutralisation techniques are employed.

* Tapped Single tuned capacitance Coupled Amplifier :-

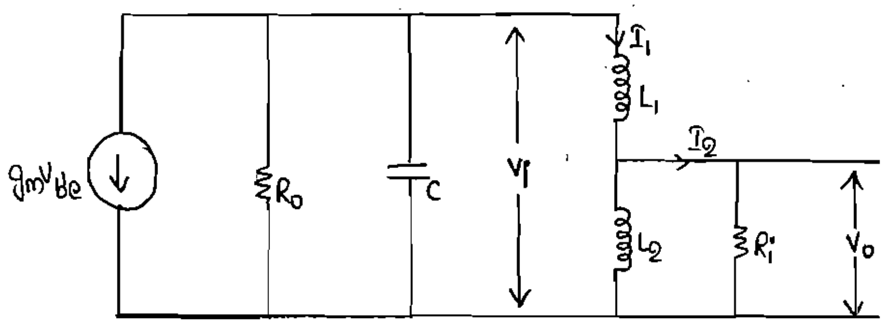


Tapped single tuned capacitive coupled amplifier.

Equivalent circuit on the output side of the stage 1.

R_i is the input resistance of the II stage.

R_0 is the output resistance of the 1st stage.



Equivalent circuit.

Expression for Inductance for maximum power transfer :-

let the tapping point divide the impedance into two parts L_1 and L_2.

Let $L_1 = nL$ so that $L_2 = (1-n)L$

Apply Kirchoff's voltage law (KVL)

$$V_1 = j\omega L_1 \cdot I_1 - j\omega (L_2 + M) I_2 \rightarrow (1)$$

$$0 = -j\omega (L_1 + M) I_1 + (R_i + j\omega L_2) I_2 \longrightarrow (2)$$

where M is the mutual Inductance between L_1 and L_2. By Solving (1)&

$$I_1 = \frac{V_1 (R_i + j\omega L_2)}{j\omega L_1 (R_i + j\omega L_2) + \omega^2 (L_2 + M)^2} \longrightarrow (3)$$

Hence |Z| offered by the coil replaced along with input resistance R_i of the next stage is

$$Z_1 = \frac{V_1}{I_1} = \frac{j\omega L_1(R_i+j\omega L_2)+\omega^2(L_2+M)^2}{(R_i+j\omega L_2)} \longrightarrow (4)$$

$$= j\omega L_1 + \frac{\omega^2(L_2+M)^2}{R_i+j\omega L_2} \rightarrow (5)$$

But ωL_2 much less than R_i.

As R_i, the input resistance of transistor circuit II stage is $K\Omega$ and much greater than ωL_2.

$$Z_1 = j\omega L_1 + \frac{\omega^2(L_2+M)^2}{R_i} \longrightarrow (6)$$

$M = K\sqrt{L_1L_2}$, where M = mutual Inductance.

where k is the coefficient of coupling, Since $L_1 = nL$, $L_2 = (1-n)L$.

$$= K\sqrt{nL(1-n)L}$$

$$= KL\sqrt{(n-n^2)} \longrightarrow (7)$$

putting K=1, we get $M \simeq L\sqrt{n-n^2} \longrightarrow (8)$

Substituting thus value of M in eq (6)

$$Z_1 = j\omega Ln \pm \frac{\omega^2[(1-n)L+L\sqrt{n-n^2}]^2}{R_i} \longrightarrow (9)$$

$$= j\omega nL + \frac{\omega^2L^2[(1-n)+\sqrt{n-n^2}]^2}{R_i} \longrightarrow (10)$$

$$= j\omega nL + R_{is}. \longrightarrow (10)$$

The resistance effectively reflected in series with the coil due to the resistance R_i is given by

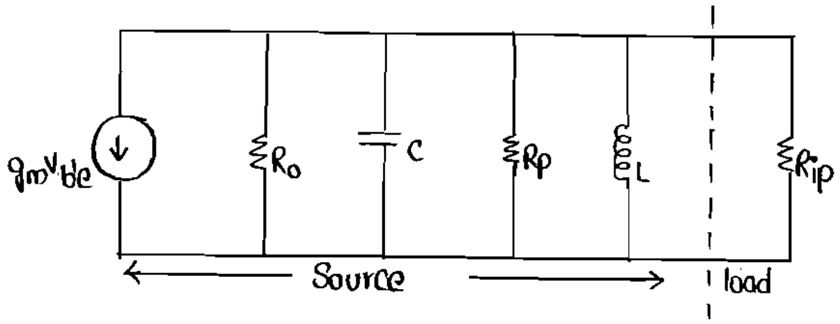
$$R_{is} = \frac{\omega^2L^2\left[(1-n)+\sqrt{n-n^2}\right]^2}{R_i}$$

This is the resistance component, S:- Series i:- input.

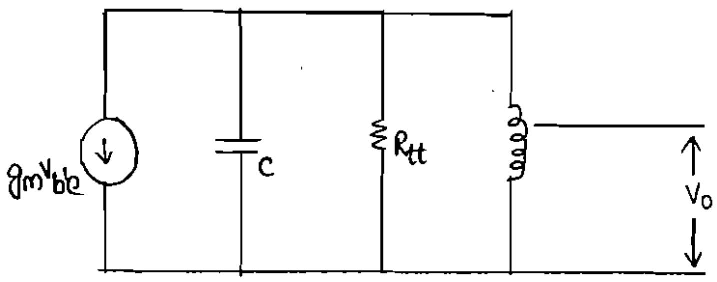
The resistance R_{is} in Series with the coil L may be equated to a resistance R_{ip} in shunt with the coil R_{ip} is given by

$$R_{ip} = (\omega L)^2/R_{is}.$$

So the equivalent circuit is



Simplifying



$$\frac{1}{R_{tt}} = \frac{1}{R_o} + \frac{1}{R_p} + \frac{1}{R_{ip}} \quad , \; Q_e = \frac{R_{tt}}{\omega_0 L}.$$ Here tt :- tapped tuned circuit

$$\omega_0 = \frac{1}{\sqrt{LC}}.$$

under the conditions of maximum power transfer theorem, the total resistance appearing in shunt with the coil is $= R_{op}$.

Since it is resonant circuit, at resonance, the $|z|$ is purely resistance. For maximum power transfer $|z| = R/2$.

$$\therefore Q_e = \frac{R_{op}/2}{\omega_0 L} \; ; \; R_{tt} = \frac{R_{op}}{2}$$

$$R_{op} = 2Q_e \cdot \omega_0 L.$$

But $$R_{op} = \frac{R_o R_p}{R_o + R_p}$$

$$2Q_e \omega_0 L = \frac{R_o \, \omega_0 Q_o L}{R_o + \omega_0 Q_o L}.$$

Solving for L, we get $$L = \frac{R_o(Q_o - 2Q_e)}{2\omega_0 Q_o Q_e}$$

Expression for L for maximum power transfer $L = \frac{R_o}{\omega_0}\left[\frac{1}{2Q_e} - \frac{1}{Q_o}\right].$

This is the value of L for maximum power transfer.

Expression for voltage gain and Bandwidth are determined in the same ways as done for a single tuned circuit. we have,

1. R_{tt} instead of R_t.

2. output voltage equals $(1-n)$ times the voltage developed across the complete coil. $|Z|$ at any frequency close to ω_0 is given by

$$Z = \frac{R_{tt}}{1+j2\delta Q_e}, \quad R_{tt} = \text{resistance of tuned tapped circuit.}$$

output voltage $V_0 = \dfrac{-g_m V_i r_{b'e}}{r_{b'e}+r_{bb'}} \cdot Z(1-n)$.

$$\therefore \text{voltage gain } A = \frac{V_0}{V_i} = -g_m(1-n)\frac{r_{b'e}}{r_{b'e}+r_{bb'}} \cdot Z.$$

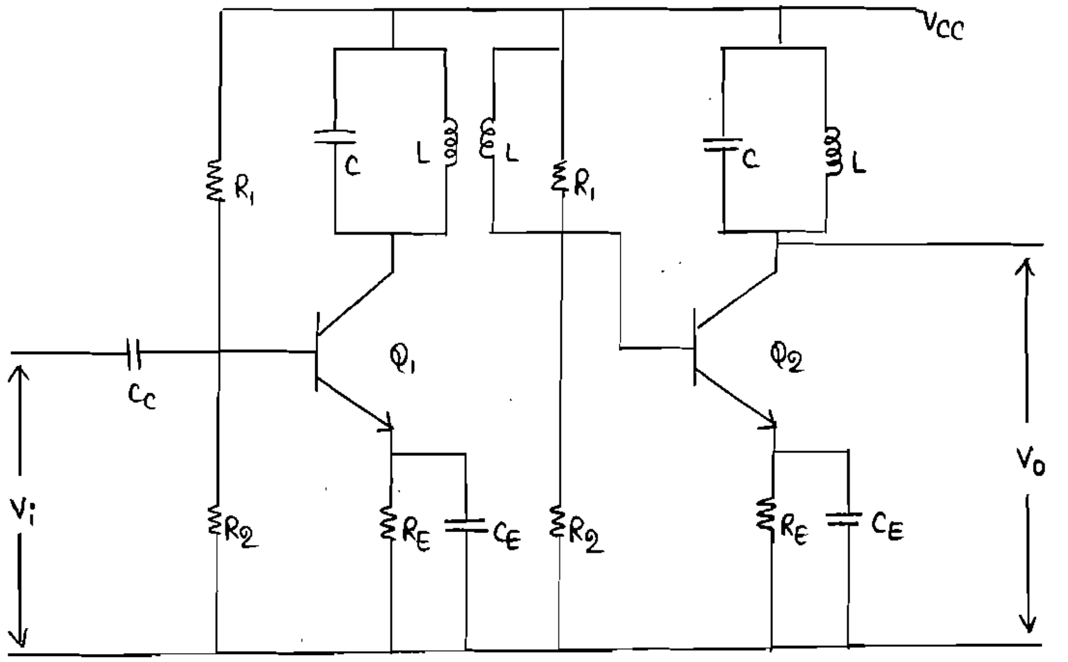
$$= -g_m(1-n) \cdot \frac{r_{b'e}}{r_{b'e}+r_{bb'}} \cdot \frac{R_{tt}}{1+2j\delta Q_e}.$$

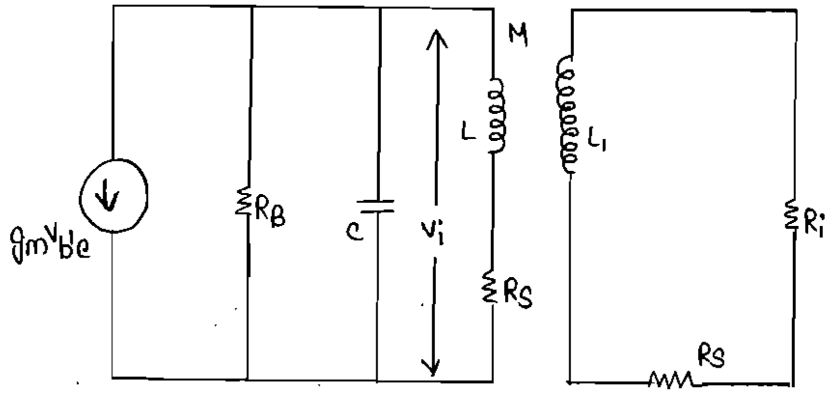
At resonance, voltage gain is

$$A_{res} = -g_m(1-n) \cdot \frac{r_{b'e}}{r_{bb'}+r_{b'e}} \cdot R_{tt}.$$

$$\therefore \frac{A}{A_{res}} = \frac{1}{1+j2\delta Q_e}.$$

* Single tuned transformer coupled or Inductively coupled amplifier:-





Expression for L for maximum power transfer :-

Apply KVL to the primary and secondary coils, we get.

$$V_1 = I_1 \cdot z_{11} + I_2 \cdot z_{12} \rightarrow (1)$$

$$0 = I_1 \cdot z_{21} + I_2 \cdot z_{22} \rightarrow (2)$$

where $z_{11} = R_1 + j\omega L$, $z_{12} = z_{21} = j\omega M$, $z_{22} = R_i + R_1 + j\omega L_1$

By solving (1) and (2), we get

$$I_1 = \frac{V_1 \cdot z_{22}}{z_{11} z_{22} \mp z_{12}^2}.$$

$$z_{in} = \frac{V_1}{I_1} = \frac{z_{11} z_{22} - z_{12}^2}{z_{22}}$$

$$z_{in} = z_{11} - \frac{z_{12}^2}{z_{22}}.$$

Substituting z_{11}, z_{12} and z_{22} in the above equation,

$$z_{in} = R + j\omega L - \frac{(j\omega M)^2}{R_i + R_1 + j\omega L_1}$$

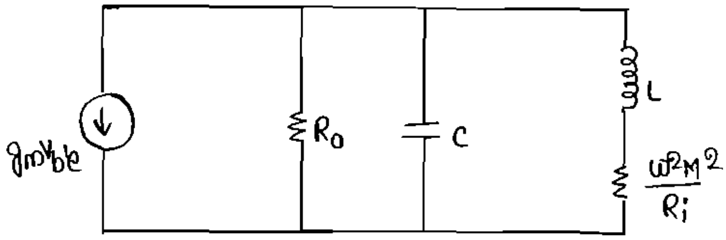
$$z_{in} = R + j\omega L + \frac{\omega^2 M^2}{R_i + R_1 + j\omega L_1}.$$

AS $R_i >> R_1 + j\omega L_1 \Rightarrow z_{in} = R + j\omega L + \frac{\omega^2 M^2}{R_i}$

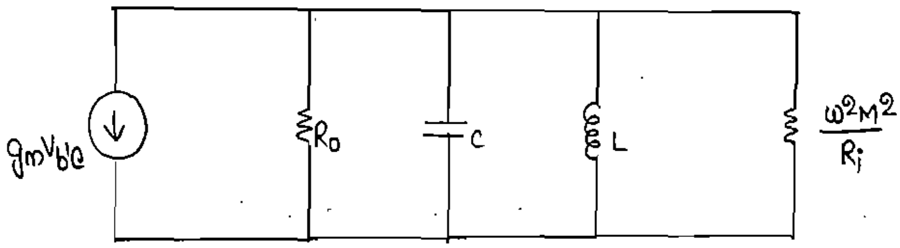
AS if the mutual inductance M, between the coils is larger than

i.e., $R << M$

$$z_{in} = j\omega L + \frac{\omega^2 M^2}{R_i}.$$



where $\frac{\omega^2 M^2}{R_i}$ is the impedance of secondary side reflected to the primar converting inductance L with series resistance $\frac{\omega^2 M^2}{R_i}$ may be represented as L in shunt with R_{ip}.

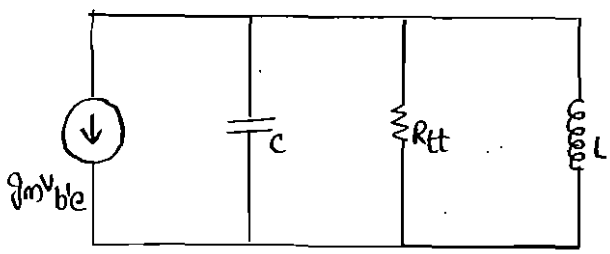


$$R_{ip} = \frac{\omega^2 L^2}{R_{is}} = \frac{\omega^2 L^2}{\frac{\omega^2 M^2}{R_i}} = \left[\frac{L}{M}\right]^2 R_i.$$

$$\therefore R_{ip} = \left[\frac{L}{M}\right]^2 \cdot R_i.$$

For maximum power transfer :-

$$R_{ip} = R_o, \quad R_o = \left[L/M\right]^2 R_i \rightarrow (3)$$



where $\frac{1}{R_{tt}} = \frac{1}{R_o} + \frac{1}{R_{ip}} \rightarrow (3)$

As the L and L_1 are primary and secondary windings of the coils,

$$\therefore M = K\sqrt{LL_1} \rightarrow (4)$$

substituting (4) in eq (3), $R_o = \left[\frac{L}{K\sqrt{LL_1}}\right]^2 R_i$

$$= \left[\frac{\cancel{L}}{K^2 \cancel{L} L_i}\right] R_i$$

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$$= \frac{R_i L}{K^2 L_i}$$

$$\therefore R_o = \frac{L}{K^2 L_i} \cdot R_i$$

from the above expression, for a given value of the R_0, K and R_i we can determine L_2 for the maximum power transfer.

The effective quality factor is

$$Q_e = \frac{R_{tt}}{\omega_0 L}.$$

For maximum power transfer, $R_{tt} = \frac{R_0}{2}$.

$$Q_e = \frac{R_0/2}{\omega_0 L} = \frac{R_0 \omega_0 L}{2}$$

$$\therefore R_0 = \frac{2Q_e}{\omega_0 L}.$$

The gain of single tuned transformer coupled amplifier is given as

$$A_V = -g_m \frac{g_{b'e}}{g_{b'e} + g_{bb'}} \cdot \frac{R_{tt}}{1+2j\delta Q_e} \cdot \frac{R_i z_{21}}{z_{11}z_{22} - z_{12}^2}$$

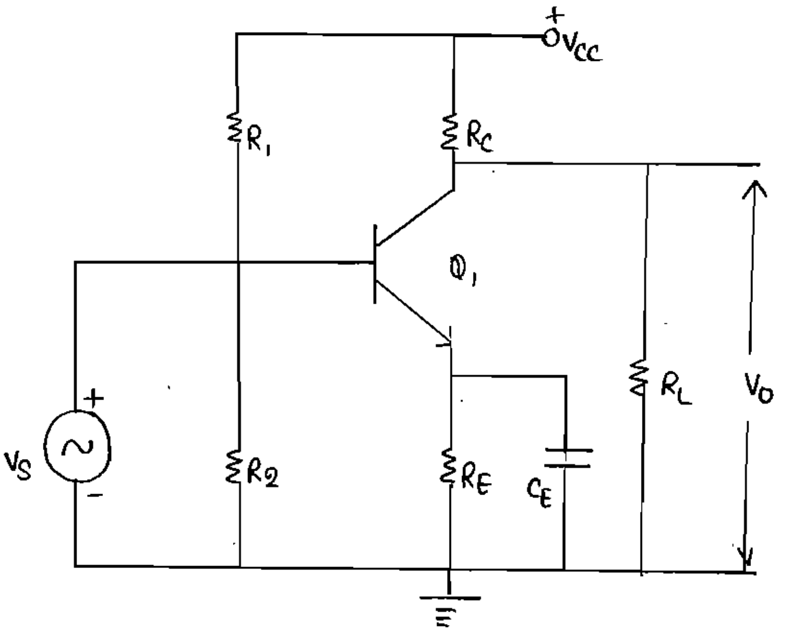
A_V at resonance:-

$$= -g_m \cdot \frac{g_{b'e}}{g_{bb'}+g_{b'e}} \cdot \frac{R_{tt} \cdot R_i \cdot z_{21}}{z_{11}z_{22} - z_{12}^2}.$$

$$\left|\frac{A_V}{A_V(\text{at resonance})}\right| = \left|\frac{1}{1+2j\delta Q_e}\right|.$$

$$\therefore \left|\frac{A_V}{A_V(\text{at resonance})}\right| = \frac{1}{\sqrt{1+(2\delta Q_e)^2}}$$

* **Wideband Amplifiers:-**



* Wide band amplifiers:-

Tuned voltage amplifier having its frequency response with uniform gain for signals covering a frequency range from few Hertz to tens of Mega hertzs.

→ Wide band amplifiers are also called as video amplifiers whose frequencies lies between 15 KHz to 5 MHz.

→ Wide band amplifiers can be achieved or designed by following any one of the compensation techniques.

1. low frequency compensation technique:-

To improve the low frequency range.

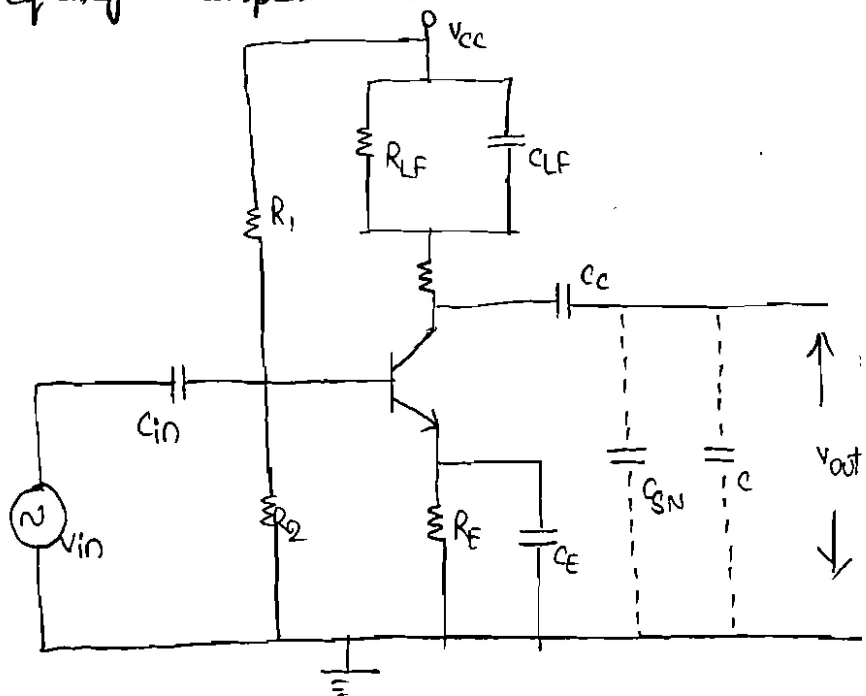
2. High frequency compensation technique:-

To improve the high frequency range.

3. Both low and high frequency compensation technique:-

To improve overall band width.

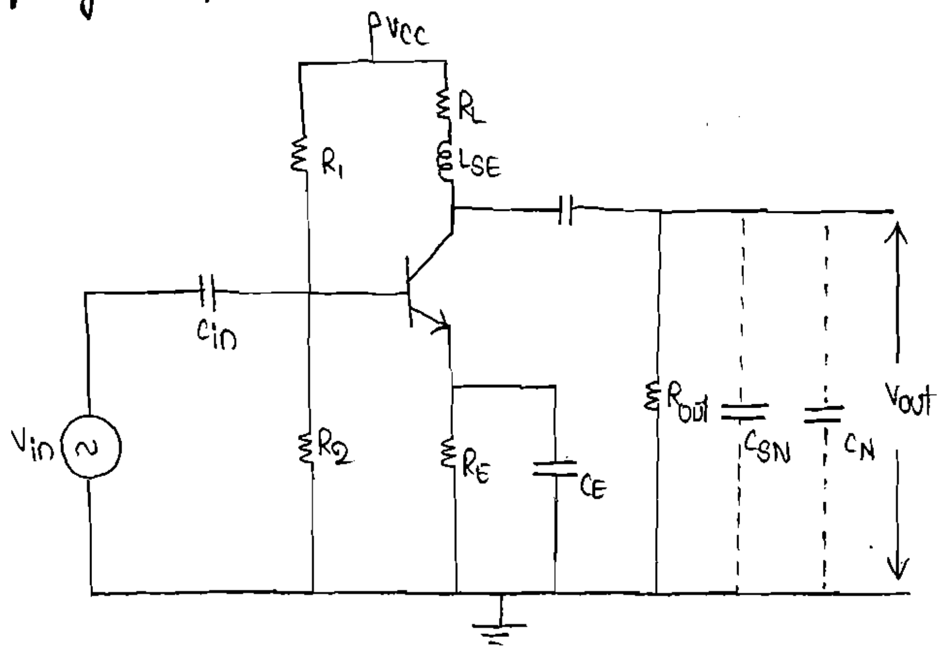
low frequency compensation:-



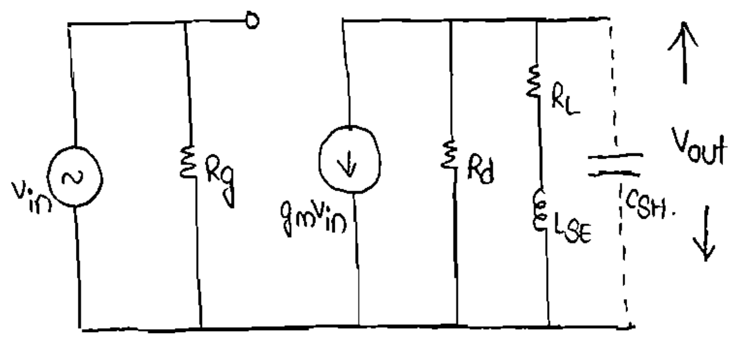
The main reason for the downfall of voltage gain in amplifiers is due to the presence of coupling capacitor (C_C). To improve voltage gain, we add a parallel combination of resistor and capacitor (R_{LF} & C_{LF}) b/w V_{CC} and load (R_L).

At mid frequencies, and high frequencies capacitor C_{LF} acts as a short circuit and total load resistance is equal to R_L. For low frequencies, C_{LF} acts as o.c and total load resistance is equal to $R_{LF}+R_L$.

High frequency compensation :-



Equivalent ckt for FET wide band Amplifiers :-



To compensate the loss at high frequencies, we add small capacitance inductance (L_{SE}) in series with load resistance (R_L) in high frequency compensation technique.

Gain at mid frequency :- $A_{mid} = -g_m \cdot R_{eq}$.

Gain of CS amplifier is $A = -g_m \cdot R_{eq}$

at mid frequency range L_{SE} is negligible . So $R_{eq} = R_L$

$$A_{mid} = -g_m \cdot R_L$$

At high frequencies, $R_{eq} = \dfrac{1}{Y_L + Y_{CSH}}$

where $Y_L = \dfrac{1}{R_L + j\omega L_{SE}}$

$$y_{CSH} = j\omega C_{SH}$$

$$A_{High} = -g_m \cdot \frac{1}{\frac{1}{R_L + j\omega L_{SE}} + j\omega C_{SH}}$$

$$\frac{A_{High}}{A_{mid}} = \frac{\dfrac{\cancel{-g_m}}{\frac{1}{R_L + j\omega L_{SE}} + j\omega C_{SH}}}{-g_m \cdot R_L} = \frac{1}{R_L\left[\frac{1}{R_L + j\omega L_{SE}} + j\omega C_{SH}\right]}$$

$$= \frac{R_L + j\omega L_{SE}}{R_L\left[1 + j\omega R_L C_{SH} + j^2\omega^2 L_{SE} C_{SH}\right]}$$

$$= \frac{R_L + \frac{j\omega L_{SE}}{R_L}}{1 + j\omega R_L C_{SH} + (-1)\omega^2 L_{SE} C_{SH}}$$

$$\alpha \left(= 1 + \frac{j\omega L_{SE}}{R_L}\left[\frac{\omega_r}{\omega_r}\right]\right) \times$$

$$= \frac{1 + \frac{j\omega L_{SE}}{R_L}\left[\frac{\omega_r}{\omega_r}\right]}{\frac{1 + j\omega}{1/R_L C_{SH}} - \omega^2 L_{SE} C_{SH}\left[\frac{\omega_r R_L}{\omega_r R_L}\right]}$$

we know that

$$Q_{eff} = \frac{\omega_r L_{SE}}{R_L}$$

where $\omega_r = \frac{1}{R_L C_{SH}}$.

$$\frac{A_{high}}{A_{mid}} = \frac{1 + j\left(\frac{\omega}{\omega_r}\right) Q_{eff}}{1 + j\left(\frac{\omega}{\omega_r}\right) - \frac{\omega^2}{\omega_r}\frac{Q_{eff}}{\omega_r}}.$$

$$\therefore \frac{A_{high}}{A_{mid}} = \frac{1 + j\left(\frac{\omega}{\omega_r}\right) Q_{eff}}{1 + j\left(\frac{\omega}{\omega_r}\right) - \left(\frac{\omega}{\omega_r}\right)^2 Q_{eff}}$$