UNIT I

Small Signal High Frequency Transistor Amplifier models

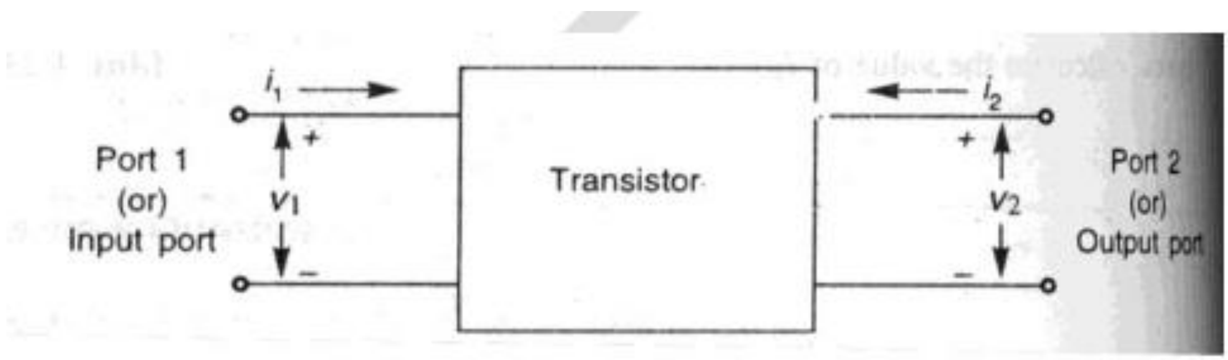
BJT: Transistor at high frequencies, Hybrid- π common emitter transistor model, Hybrid π conductances, Hybrid π capacitances, validity of hybrid π model, determination of high-frequency parameters in terms of low-frequency parameters , CE short circuit current gain, current gain with resistive load, cut-off frequencies, frequency response and gain bandwidth product. **FET:** Analysis of common Source and common drain Amplifier circuits at high frequencies.

Introduction:

Electronic circuit analysis subject teaches about the basic knowledge required to design an amplifier circuit, oscillators etc .It provides a clear and easily understandable discussion of designing of different types of amplifier circuits and their analysis using hybrid model, to find out their parameters. Fundamental concepts are illustrated by using small examples which are easy to understand. It also covers the concepts of MOS amplifiers, oscillators and large signal amplifiers.

Two port devices & Network Parameters:

A transistor can be treated as a two-part network. The terminal behavior of any two-part network can be specified by the terminal voltages V1& V2at parts 1 & 2 respectively and current i1and i2, entering parts 1 & 2, respectively, as shown in figure.



Of these four variables V1, V2, i1and i2, two can be selected as independent variables and the remaining two can be expressed in terms of these independent variables. This leads to various two part parameters out of which the following three are more important.

1. Z –Parameters (or) Impedance parameters
2. Y –Parameters (or) Admittance parameters
3. H –Parameters (or) Hybrid parameters

Hybrid parameters (or) h –parameters:

The equivalent circuit of a transistor can be dram using simple approximation by retaining its essential features. These equivalent circuits will aid in analyzing transistor circuits easily and rapidly.

If the input current i_1 and output Voltage V_2 are takes as independent variables, the input voltage V_1 and output current i_2 can be written as

$$\mathbf{V_1 = h_{11}\, i_1 + h_{12}\, V_2}$$
$$\mathbf{i_2 = h_{21}\, i_1 + h_{22}\, V_2}$$

The four hybrid parameters h11, h12, h21and h22 are defined as follows:

h11= [V1/ i1] with V2= 0 Input Impedance with output part short circuited.
h22= [i2/ V2] with i1= 0 Output admittance with input part open circuited.
h12= [V1/ V2] with i1= 0 reverse voltage transfer ratio with input part open circuited.
h21= [i2/ i1] with V2= 0 Forward current gain with output part short circuited

The dimensions of h–parameters are as follows:

h11-Ω

h22–mhos

h12, h21 –dimension less.

As the dimensions are not alike, (i.e.) they are hybrid in nature, and these parameters are called as hybrid parameters.

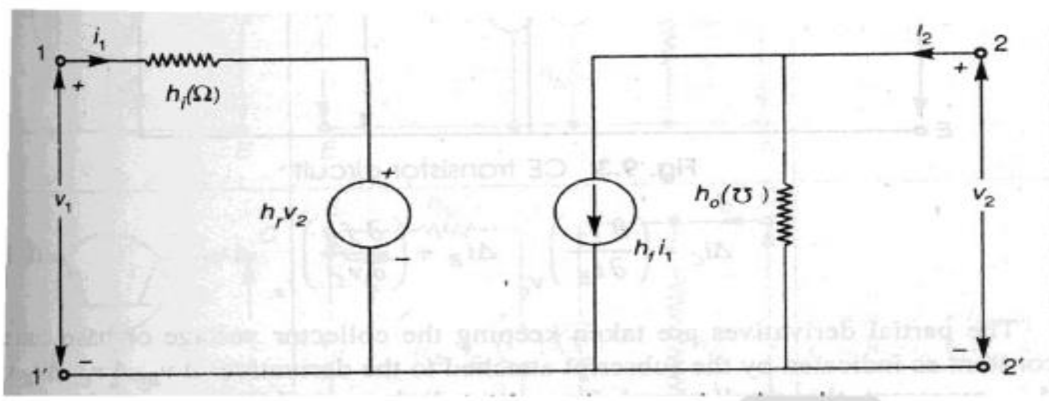
h11 = input; h 22 = output;
h21= forward transfer; h22 = Reverse transfer.

Notations used in transistor circuits:

h_{ie}= h11e= Short circuit input impedance
h_{oe}= h22e= Open circuit output admittance
h_{re} = h12e= Open circuit reverse voltage transfer ratio
h_{fe}= h21e= Short circuit forward current Gain.

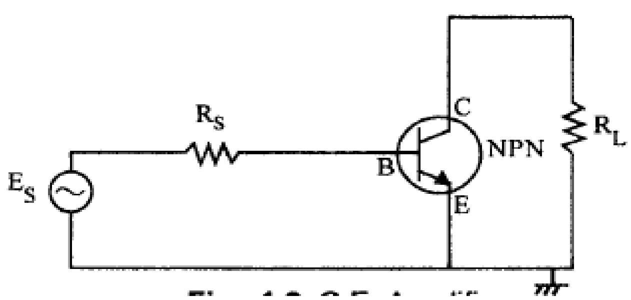
The Hybrid Model for Two-port Network:

$$V1= h11\ i1+ h12V2$$
$$I2= h21i1+ h22V2$$
$$V1= h_1 i1+ h_r V2$$
$$I2 = h_f i1+ h_0 V2$$

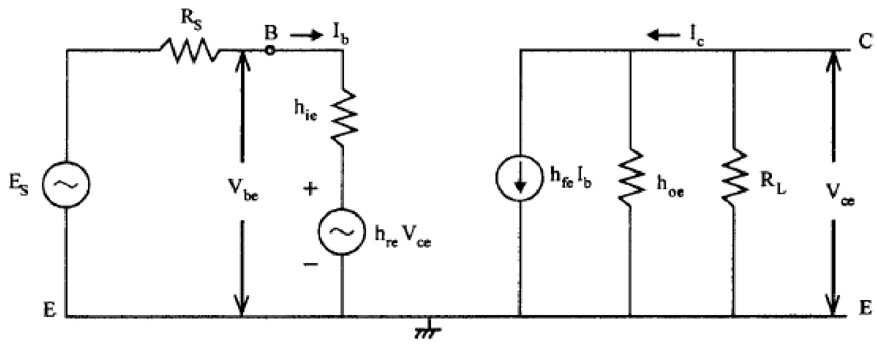


Common Emitter Amplifier

Common Emitter Circuit is as shown in the Fig. The DC supply, biasing resistors and coupling capacitors are not shown since we are performing an AC analysis.



E_s is the input signal source and R_s is its resistance. The h-parameter equivalent for the above circuit is as shown in Fig.



$$h_{ie} = \left.\frac{V_{be}}{I_b}\right|_{V_{ce}=0} \qquad h_{re} = \left.\frac{V_{be}}{V_{ce}}\right|_{I_b=0}$$

$$h_{oe} = \left.\frac{I_c}{V_{ce}}\right|_{I_b=0} \qquad h_{fe} = \left.\frac{I_c}{I_b}\right|_{V_{ce}=0}$$

The typical values of the h-parameter for a transistor in Common Emitter configuration are,

$$h_{ie} = \frac{V_{be}}{I_b}$$

$$h_{ie} = \frac{0.2V}{50\times10^{-6}} = 4K\Omega$$

$$h_{fe} = I_c/I_b \simeq 100\,.$$

$$h_{fe} >> 1 \simeq \beta$$

$$h_{re} = \frac{V_{be}}{V_{ce}}$$

$$h_{re} = 0.2 \times 10^{-3}.$$

$$h_{oe} = \frac{I_c}{V_{ce}}.$$

$$h_{oe} = 8\ \mu\mho$$

$$R_i = h_{ie} - \frac{h_{fe}h_{re}}{h_{oe} + \frac{1}{R_L}}$$

$$R_i = h_{ie} - \frac{h_{fe}h_{re}}{h_{oe} + \frac{1}{R_L}}$$

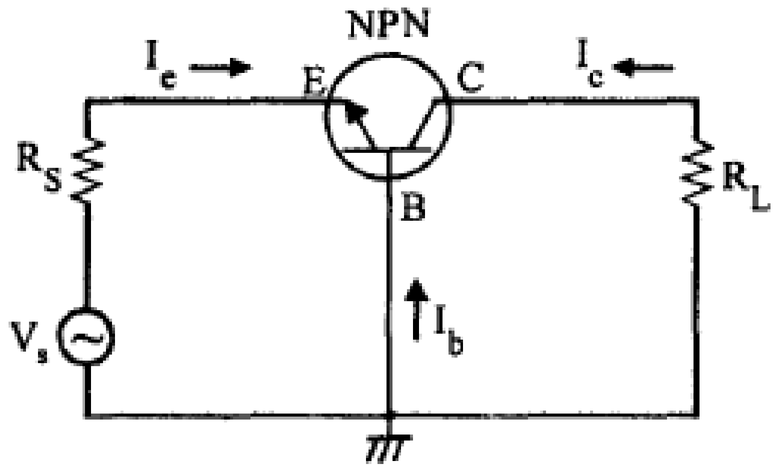
$$R_o = \frac{1}{h_{oe} - \left(\frac{h_{re}h_{fe}}{h_{ie} + R_s}\right)}$$

$$A_i = \frac{-h_{fe}}{1 + h_{oe}R_L}$$

$$A_V = \frac{-h_{fe}R_L}{h_{ie} + R_L(h_{ie}h_{oe} - h_{fe}h_{re})}$$

Common Base Amplifier

Common Base Circuit is as shown in the Fig. The DC supply, biasing resistors and coupling capacitors are not shown since we are performing an AC analysis.



$$h_{ib} = \left.\frac{V_{eb}}{I_e}\right|_{V_{cb}=0}$$

$$h_{fb} = \left.\frac{I_c}{I_e}\right|_{V_{cb}=0} = -0.99 \text{ (Typical Value)}$$

$$I_c < I_e \qquad \therefore \qquad h_{fb} < 1$$

$$h_{ob} = \left.\frac{I_c}{V_{cb}}\right|_{I_e=0} = 7.7 \times 10^{-8} \text{ mhos (Typical Value)}$$

$$h_{rb} = \left.\frac{V_{eb}}{V_{cb}}\right|_{I_e = 0} = 37 \times 10^{-6} \quad \text{(Typical Value)}$$

$$R_i = h_{ib} - \frac{h_{fb} . h_{rb}}{h_{ob} + \frac{1}{R_L}}$$

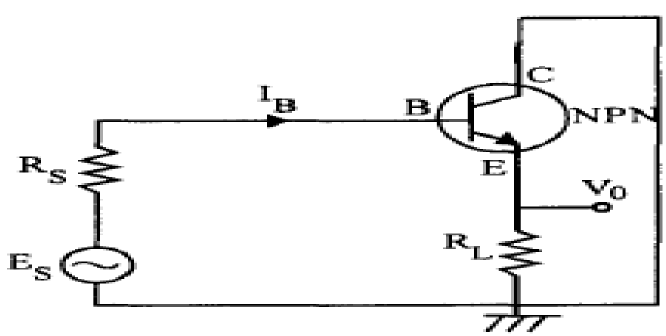
$$R_o = \frac{1}{h_{ob} - \frac{h_{rb} h_{fb}}{h_{ib} + R_S}}$$

$$A_I = \frac{-h_{fb}}{1 + h_{ob} R_L}$$

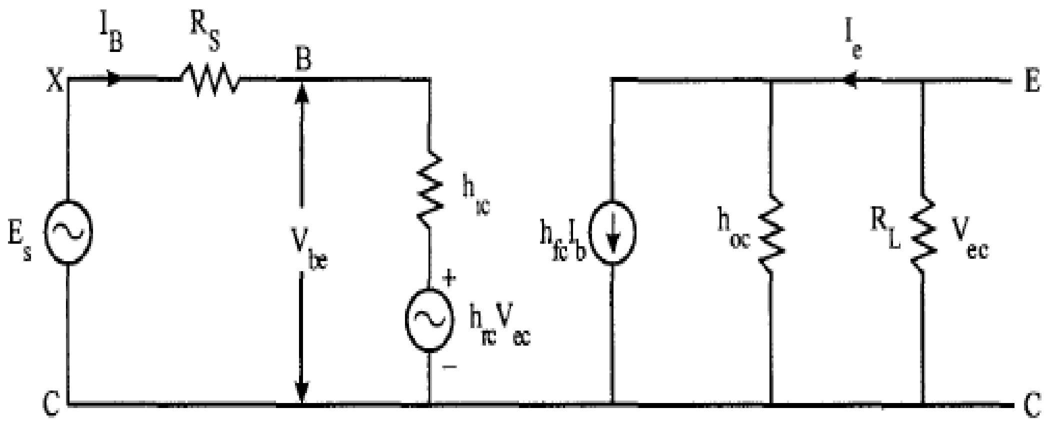
$$A_v = \frac{-h_{fb} R_L}{h_{ib} + R_L (h_{ib} h_{ob} - h_{fb} h_{rb})}$$

Common Collector Amplifier

Common Collector Circuit is as shown in the Fig. The DC supply, biasing resistors and coupling capacitors are not shown since we are performing an AC analysis



The h-parameter model is shown below



$$h_{ic} = \left.\frac{V_{bc}}{I_b}\right|_{V_{ec}=0} = 2{,}780\ \Omega \text{ (Typical Value)}$$

$$h_{oc} = \left.\frac{I_e}{V_{ec}}\right|_{I_b=0} = 7.7 \times 10^{-6} \text{ mhos (Typical Value)}$$

$$h_{fc} = -\left.\frac{I_e}{I_b}\right|_{V_{ce}=0} = 100 \text{ (Typical value)}$$

$\because I_e >> I_b.$

$$h_{rc} = \left.\frac{V_{bc}}{V_{ec}}\right|_{I_b=0};$$

Transistors at High Frequencies

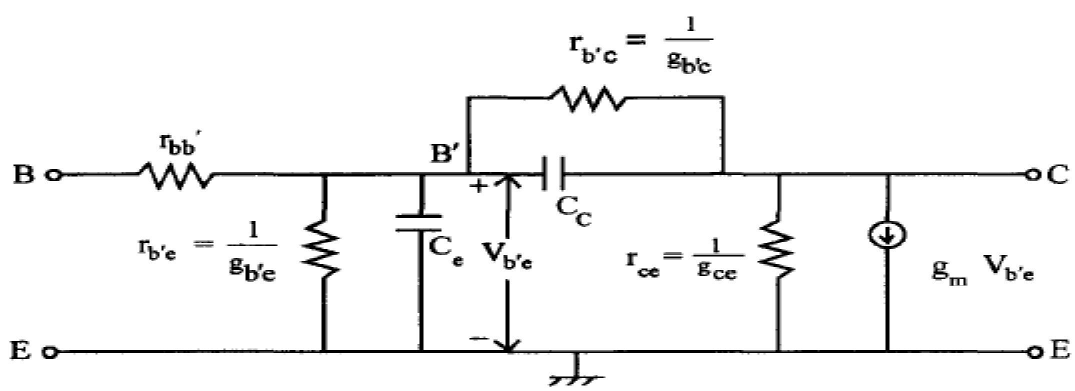
At low frequencies it is assumed that transistor responds instantaneously to changes in the input voltage or current i.e., if you give AC signal between the base and emitter of a Transistor amplifier in Common Emitter configuraii6n and if the input signal frequency is low, the output at the collector will exactly follow the change in the input (amplitude etc.,). If '1' of the input is high (MHz) and the amplitude of the input signal is changing the Transistor amplifier will not be able to respond.

It is because; the carriers from the emitter side will have to be injected into the collector side. These take definite amount of time to travel from Emitter to Base, however small it may be. But if the input signal is varying at much higher speed than the actual time taken by the carries to respond, then the Transistor amplifier will not respond instantaneously. Thus, the junction capacitances of the transistor, puts a limit to the highest frequency signal which the transistor can handle. Thus depending upon doping area of the junction etc, we have transistors which can respond in AF range and also RF range.

To study and analyze the behavior of the transistor to high frequency signals an equivalent model based upon transmission line equations will be accurate. But this model will be very complicated to analyze. So some approximations are made and the equivalent circuit is simplified. If the circuit is simplified to a great extent, it will be easy to analyze, but the results will not be accurate. If no approximations are made, the results will be accurate, but it will be difficult to analyze. The desirable features of an equivalent circuit for analysis are simplicity and accuracy. Such a circuit which is fairly simple and reasonably accurate is the Hybrid-pi or Hybrid-π model, so called because the circuit is in the form of π.

Hybrid - π Common Emitter Transconductance Model

For Transconductance amplifier circuits Common Emitter configuration is preferred. Why? Because for Common Collector (hrc< 1). For Common Collector Configuration, voltage gain Av < 1. So even by cascading you can't increase voltage gain. For Common Base, current gain is hib< 1. Overall voltage gain is less than 1. For Common Emitter, hre>>1. Therefore Voltage gain can be increased by cascading Common Emitter stage. So Common Emitter configuration is widely used. The Hybrid-x or Giacoletto Model for the Common Emitter amplifier circuit (single stage) is as shown below.



Analysis of this circuit gives satisfactory results at all frequencies not only at high frequencies but also at low frequencies. All the parameters are assumed to be independent of frequency.

Where B' = internal node in base
$r_{bb'}$ = Base spreading resistance
$r_{b'e}$ = Internal base node to emitter resistance
r_{ce} = collector to emitter resistance
C_e = Diffusion capacitance of emitter base junction
$r_{b'c}$ = Feedback resistance from internal base node to collector node
g_m = Transconductance
C_C= transition or space charge capacitance of base collector junction

Circuit Components

B' is the internal node of base of the Transconductance amplifier. It is not physically accessible. The base spreading resistance rbb is represented as a lumped parameter between base B and internal node B'. $g_{mVb'e}$ is a current generator. Vb'e is the input voltage across the emitter junction. If Vb'e increases, more carriers are injected into the base of the transistor. So the increase in the number of carriers is proportional to Vb'e. This results in small signal current since we are taking into account changes in Vb'e. This effect is represented by the current generator g_mVb'e. This represents the current that results because of the changes in Vb'e' when C is shorted to E.

When the number of carriers injected into the base increase, base recombination also increases. So this effect is taken care of by gb'e. As recombination increases, base current increases. Minority carrier storage in the base is represented by C_e the diffusion capacitance.

According to Early Effect, the change in voltage between Collector and Emitter changes the base width. Base width will be modulated according to the voltage variations between Collector and Emitter. When base width changes, the minority carrier concentration in base changes. Hence the current which is proportional to carrier concentration also changes. I_E changes and I_C changes. This feedback effect [IE on input side, I_C on output side] is taken into account by connecting gb'e between B', and C. The conductance between Collector and Base is g_{ce}.C_c represents the collector junction barrier capacitance.

Hybrid - *n* Parameter Values

Typical values of the hybrid-n parameter at I_C = 1.3 rnA are as follows:

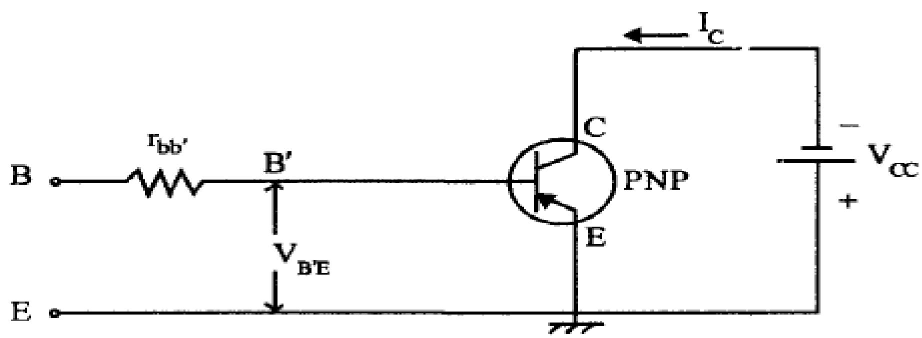
$$g_m = 50 \text{ mA/v}$$

rbb' = 100 Ω
rb'e = 1 kΩ
ree = 80 kΩ
Cc = 3 pf
Ce = 100 pf
rb'c = 4 MΩ

These values depend upon:
1. Temperature 2. Value of I_C

Determination of Hybrid-x Conductances

1. Trans conductance or Mutual Conductance (g_m)



The above figure shows PNP transistor amplifier in Common Emitter configuration for AC purpose, Collector is shorted to Emitter.

$$I_C = I_{C0} - \alpha_0 \cdot I_E$$

I_{CO} opposes I_E. I_E is negative. Hence $I_C = I_{CO} - \alpha_0 IE$ α_0 is the normal value of α at room temperature.

In the hybrid - π equivalent circuit, the short circuit current = $g_m Vb'$ e

Here only transistor is considered, and other circuit elements like resistors, capacitors etc are not considered.

$$g_m = \left.\frac{\partial I_C}{\partial V_{b'e}}\right|_{V_{CE} = K}$$

Differentiate (l) with respect to Vb'e partially. I_{CO} is constant

$$g_m = 0 - \alpha_0 \frac{\partial I_E}{\partial V_{b'e}}$$

For a PNP transistor, Vb'e = $-V_E$ Since, for PNP transistor, base is n-type. So negative voltage is given

$$g_m = \alpha_0 \frac{\partial I_E}{\partial V_E}$$

If the emitter diode resistance is r_e then

$$r_e = \frac{\partial V_E}{\partial I_E}$$

$$g_m = \frac{\alpha_0}{r_e}$$

$r = \frac{\eta . V_T}{I}$ $\eta = 1,$ $I = I_E$ $r = \frac{V_T}{I_E}$

$g_m = \frac{\alpha_0 . I_E}{V_T}$ $\alpha_0 \simeq 1,$ $I_E \simeq I_C$

$I_E = I_{C0} - I_C$

$g_m = \frac{I_{C0} - I_C}{V_T}$

Neglect I_{C0}

$$g_m = \frac{|I_C|}{V_T}$$

g_m is directly proportional to I_C is also inversely proportional to T. For PNP transistor, I_C is negative

$$\boxed{g_m = \frac{-I_C}{V_T}}$$

At room temperature i.e. $T=300^0K$

$$g_m = \frac{|I_C|}{26}, \; I_C \text{ is in mA.}$$

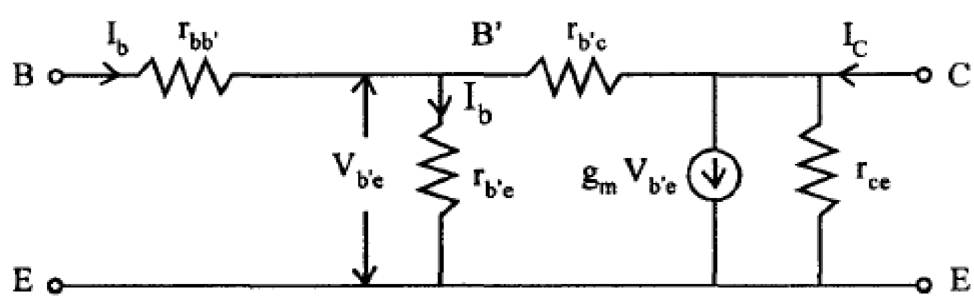
If $I_C = 1.3$ mA, $g_m = 0.05$ A/V

If $I_C = 10$ mA, $g_m = 400$ mA/V

Input Conductance (gb'e):

At low frequencies, capacitive reactance will be very large and can be considered as Open circuit. So in the hybrid-π equivalent circuit which is valid at low frequencies, all the capacitances can be neglected.

The equivalent circuit is as shown in Fig.



The value of rb'c» rb'e (Since Collector Base junction is Reverse Biased)So I_b flows into rb'e only. [This is lb' (IE - I_b)will go to collector junction]

$$V_{b'e} \simeq I_b \cdot r_{b'e}$$

The short circuit collector current,

$$I_C = g_m \cdot V_{b'e}; \qquad V_{b'e} = I_b \cdot r_{b'e}$$

$$I_C = g_m \cdot I_b \cdot r_{b'e}$$

$$h_{fe} = \left.\frac{I_C}{I_B}\right|_{V_{CE}} = g_m \cdot r_{b'e}$$

$$\boxed{r_{b'e} = \frac{h_{fe}}{g_m}}$$

$$g_m = \frac{|I_C|}{V_T}$$

$$r_{b'e} = \frac{h_{fe} \cdot V_T}{|I_C|}$$

$$g_{b'e} = \boxed{\frac{|I_C|}{h_{fe} V_T}} \text{ or } \boxed{\frac{g_m}{h_{fe}}}$$

Feedback Conductance ($g_{b'c}$)

hre = reverse voltage gain, with input open or $I_b = 0$
hre =Vb'e/Vce = Input voltage/Output voltage

$$h_{re} = \frac{r_{b'e}}{r_{b'e} + r_{b'c}}$$

[With input open, i.e., I_b – 0, V_{ce} is output. So it will get divided between $r_{b'e}$ and $r_{b'c}$ only]

or $$h_{re}(r_{b'e} + r_{b'c}) = r_{b'e}$$

$$r_{b'e}[1 - h_{re}] = h_{re} r_{b'c}$$

But $$h_{re} \ll 1$$

∴ $$r_{b'e} = h_{re} \; r_{b'c}; \quad r_{b'c} = \frac{r_{b'e}}{h_{re}}$$

or $$\boxed{g_{b'c} = h_{re} \; g_{b'e}} \quad \frac{1}{r_{b'c}} = g_{b'c} = \frac{h_{re}}{r_{b'e}}$$

$$h_{re} = 10^{-4}$$

∴ $$r_{b'c} \gg r_{b'e}$$

Base Spreading Resistance (r bb')

The input resistance with the output shorted is hie. If output is shorted, i.e., Collector and Emitter arejoined; $r_{b'e}$ is in parallel with $r_{b'c}$.

$$h_{ie} = r_{bb'} + r_{b'e}$$

$$\boxed{r_{bb'} = h_{ie} - r_{b'e}}$$

$$h_{ie} = r_{bb'} + r_{b'e}$$

$$r_{b'e} = \frac{h_{fe}.V_T}{|I_C|}$$

$$h_{ie} = r_{bb'} + \frac{h_{fe}.V_T}{|I_C|}$$

Output Conductance (g_{ce})

This is the conductance with input open circuited. In h-parameters it is represented as hoe. For I_b= 0, we have,

$$I_C = \frac{V_{ce}}{r_{ce}} + \frac{V_{ce}}{r_{b'c} + r_{b'e}} + g_m \ V_{b'e}$$

$$h_{re} = \frac{V_{b'e}}{V_{ce}} \qquad \therefore \qquad V_{b'e} = h_{re}.\ V_{ce}$$

$$I_C = \frac{V_{ce}}{r_{ce}} + \frac{V_{ce}}{r_{b'c} + r_{b'e}} + g_m.\ h_{re}.V_{ce}$$

$$h_{oe} = \frac{1}{r_{ce}} + \frac{1}{r_{b'c}} + g_m \cdot h_{re}$$

$$= g_{ce} + g_{b'c} + g_m \, h_{re}$$

$$g_{b'e} = \frac{g_m}{h_{fe}}$$

$$g_m = g_{b'e} \cdot h_{fe}$$

$$h_{re} = \frac{r_{b'e}}{r_{b'e} + r_{b'c}} \approx \frac{r_{b'e}}{r_{b'c}} = \frac{g_{b'c}}{g_{b'e}}$$

$$h_{oe} = g_{ce} + g_{b'c} + g_{b'e} \, h_{fe} \cdot \frac{g_{b'c}}{g_{b'e}}$$

$$g_{ce} = h_{oe} - (1 + h_{fe}) \cdot g_{b'c}$$

$$h_{fe} >> 1, \; 1 + h_{fe} \approx h_{fe}$$

$$\boxed{\mathbf{g_{ce} = h_{oe} - h_{fe} \cdot g_{b'c}}}$$

$$g_{b'c} = h_{re} \cdot g_{b'e}$$

$$g_{ce} = h_{oe} - h_{fe} \cdot h_{re} \cdot g_{b'e}$$

Hybrid - π Capacitances

In the hybrid - **π** equivalent circuit, there are two capacitances, the capacitance between the Collector Base junction is the C_c or $C_{b'e'}$. This is measured with input open i.e., $I_E = 0$, and is specified by the manufacturers as C_{Ob}. 0 indicates that input is open. Collector junction is reverse biased.

$$C_C \; \alpha \; \frac{1}{(V_{CE})^n}$$

$$n = \frac{1}{2} \quad \text{for abrupt junction}$$

$$= 1/3 \quad \text{for graded junction.}$$

C_e = Emitter diffusion capacitance C_{De} + Emitter junction capacitance C_{Te}

C_T = Transition capacitance.

C_D = Diffusion capacitance.

$C_{De} >> C_{Te}$

$C_e \simeq C_{De}$

$C_{De} \; \alpha \; I_E$ and is independent of Temperature T.

Validity of hybrid-π model

The high frequency hybrid Pi or Giacoletto model of BJT is valid for frequencies less than the unit gain frequency.

High frequency model parameters of a BJT in terms of low frequency hybrid parameters

The main advantage of high frequency model is that this model can be simplified to obtain low frequency model of BJT. This is done by eliminating capacitance's from the high frequency model so that the BJT responds without any significant delay (instantaneously) to the input signal. In practice there will be some delay between the input signal and output signal of BJT which will be very small compared to signal period (1/frequency of input signal) and hence can be neglected. The high frequency model of BJT is simplified at low frequencies and redrawn as shown in the figure below along with the small signal low frequency hybrid model of BJT.

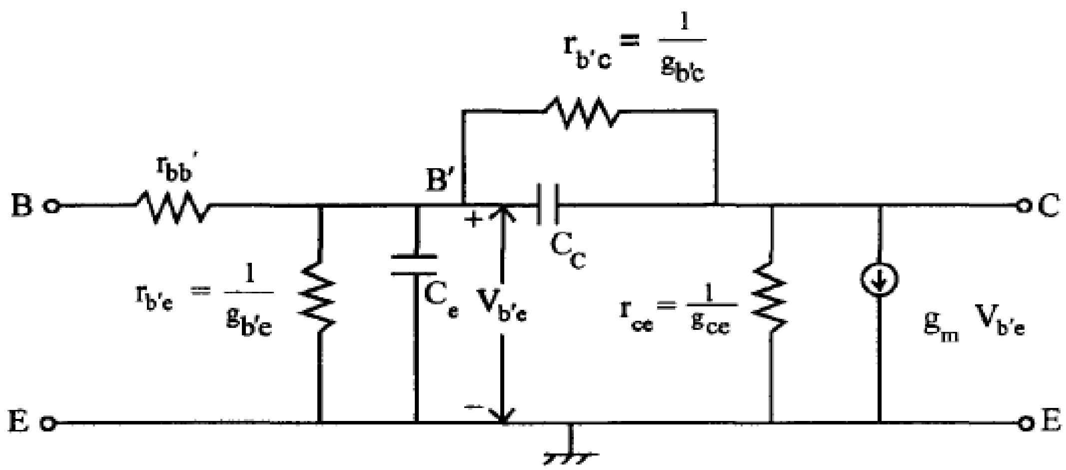


Fig. high frequency model of BJT at low frequencies

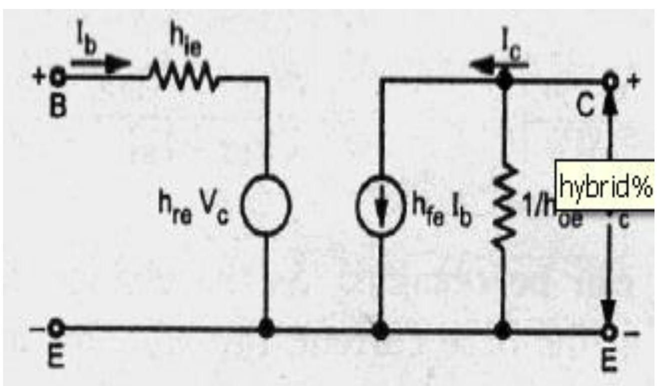


Fig hybrid model of BJT at low frequencies

The High frequency model parameters of a BJT in terms of low frequency hybrid parameters are given below:

Transconductance $g_m = I_c/V_t$
Internal Base node to emitter resistance $r_{b'e} = h_{fe}/ g_m = (h_{fe} * V_t)/ I_c$
Internal Base node to collector resistance $r_{b'e} = (h_{re} * r_{b'c}) / (1- h_{re})$ assuming $h_{re} << 1$ it reduces to $r_{b'e} = (h_{re} * r_{b'c})$
Base spreading resistance $r_{bb'} = h_{ie} - r_{b'e} = h_{ie} - (h_{fe} * V_t)/ I_c$
Collector to emitter resistance $r_{ce} = 1 / (h_{oe} - (1+ h_{fe})/r_{b'c})$

Collector Emitter Short Circuit Current Gain

Consider a single stage Common Emitter transistor amplifier circuit. The hybrid-1t equivalent circuit is as shown:

$$I_L = -g_m V_{b'e}$$

$$V_{b'e} = \frac{I_i}{g_{b'e} + j\omega(C_e + C_c)}$$

A_I under short circuit condition is,

$$A_I = \frac{I_L}{I_i} = \frac{-g_m}{g_{b'e} + j\omega(C_e + C_c)}$$

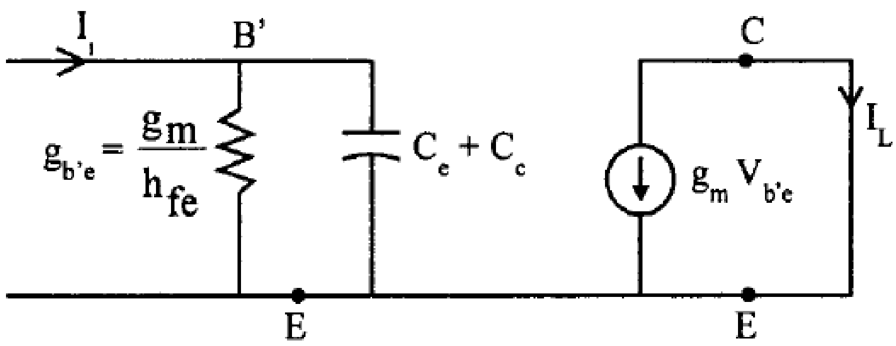
But $$g_{b'e} = \frac{g_m}{h_{fe}}, \quad C_e + C_c \simeq C_e$$

$$C_e = \frac{g_m}{2\pi f_T}$$

$$= \frac{-g_m}{\frac{g_m}{h_{fe}} + \frac{j\,2\pi . g_m . f}{2\pi f_T}}$$

$\therefore$ $$A_I = \frac{-1}{\frac{1}{h_{fe}} + j\left(\frac{f}{f_T}\right)}$$

If the output is shorted i.e. $R_L = 0$, what will be the flow response of this circuit? When$R_L = 0$, $V_o = 0$. Hence $A_v = 0$. So the gain that we consider here is the current gain I_L/Ic. The simplified equivalent circuit with output shorted is,



A current source gives sinusoidal current Ic. Output current or load current is IL· $g_{b'c}$ isneglected since $g_{b'c} \ll g_{b'e}$, g_{ce} is in shunt with short circuit R = 0. Therefore g_{ce} disappears. The current is delivered to the output directly through Ce and $g_{b'c}$ is also neglected since this will be very small.

$$I_L = -g_m V_{b'e}$$

$$V_{b'e} = \frac{I_i}{g_{b'e} + j\omega(C_e + C_c)}$$

A_I under short circuit condition is,

$$A_I = \frac{I_L}{I_i} = \frac{-g_m}{g_{b'e} + j\omega(C_e + C_c)}$$

But $$g_{b'e} = \frac{g_m}{h_{fe}}, \quad C_e + C_c \simeq C_e$$

$$C_e = \frac{g_m}{2\pi f_T}$$

$$= \frac{-g_m}{\frac{g_m}{h_{fe}} + \frac{j\,2\pi . g_m . f}{2\pi f_T}}$$

$\therefore$ $$A_I = \frac{-1}{\frac{1}{h_{fe}} + j\left(\frac{f}{f_T}\right)}$$

$$= \frac{-h_{fe}}{1 + j\, h_{fe}\left(\frac{f}{f_T}\right)}$$

$$A_i = \frac{-h_{fe}}{1 + j\left(\frac{f}{f_\beta}\right)}$$

$$\frac{f_T}{h_{fe}} = f_\beta$$

$$|A_i| = \frac{h_{fe}}{\sqrt{1 + \left(\frac{f}{f_\beta}\right)^2}}$$

Where $f_\beta = \frac{g_{b'e}}{2\pi(C_e + C_C)}$

$$g_{b'e} = \frac{g_m}{h_{fe}}$$

$\therefore$ $f_\beta = \frac{g_m}{h_{fe}\, 2\pi(Ce + C_c)}$

At $f = f_\beta$, $A_i = \frac{1}{\sqrt{2}}$ $= 0.707$ of h_{fe}.

Current Gain with Resistance Load:

$$f_T = f_\beta \cdot h_{fe} = \frac{g_m}{2\pi(C_e + C_c)}$$

Considering the load resistance R_L
$V_{b'e}$ is the input voltage and is equal to V_1
V_{ce} is the output voltage and is equal to V_2

$$K_2 = \frac{V_{ce}}{V_{b'e}}$$

This circuit is still complicated for analysis. Because, there are two time constants associated with the input and the other associated with the output. The output time constant will be much smaller than the input time constant. So it can be neglected.

K = Voltage gain. It will be >> 1

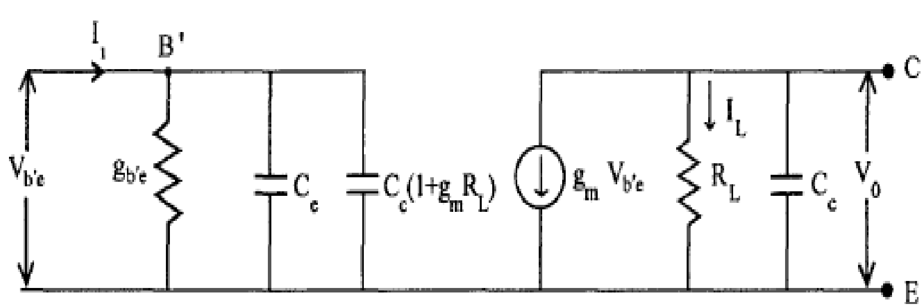
$$g_{b'c}\left(\frac{K-1}{K}\right) \simeq g_{b'c}$$

$$g_{b'c} < g_{ce} \quad \because \quad r_{b'c} \simeq 4\ M\Omega, \qquad r_{ce} = 80\ K \text{ (typical values)}$$

So $g_{b'c}$ can be neglected in the equivalent circuit. In a wide band amplifier RL will not exceed 2KΩ. If R_L is small f_H is large.

$$f_H = \frac{1}{2\pi C_s (R_C \| R_L)}$$

Therefore g_{ce} can be neglected compared with R_L. Therefore the output circuit consists of current generator gm $V_{b'e}$ feeding the load R_L so the Circuit simplifies as shown in Fig.



$$K = \frac{V_{ce}}{V_{b'e}} = -g_m R_L; \quad g_m = 50 \text{ mA\V}, \quad R_L = 2K\Omega \text{ (typical values)}$$

$$K = -100$$

Miller's Theorem

It states that if an impedance Z is connected between the input and output terminals, of a network, between which there is voltage gain, K, the same effect can be had by removing Z and connecting an impedance Z_i at the input =Z/(1-K) and Zo across the output = ZK/(K-1)

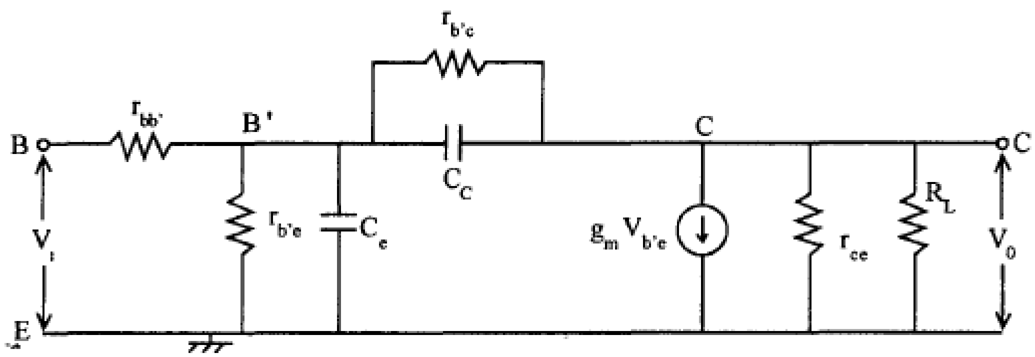


Fig. High frequency equivalent circuit with resistive load R_L

Therefore high frequency equivalent circuit using Miller's theorem reduces to

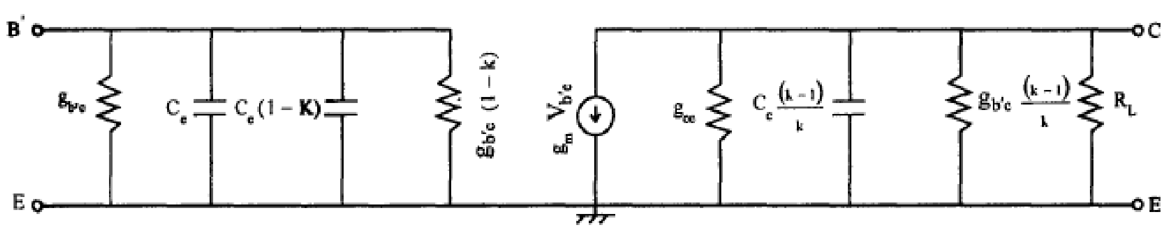


Fig. Circuit after applying Millers' Theorem

$$K = \frac{V_{ce}}{V_{b'e}}$$

$$V_{ce} = -I_c \cdot R_L$$

$$K = \frac{-I_C . R_L}{V_{b'e}}$$

$$\frac{I_C}{V_{b'e}} = g_m$$

$$K = -g_m . R_L$$

The Parameters f_T

f_T is the frequency at which the short circuit Common Emitter current gain becomes unity.

The Parameters f_β

$$A_i = 1, \quad \text{or} \quad \frac{h_{fe}}{\sqrt{1 + \left(\frac{f_T}{f_\beta}\right)^2}} = 1$$

$$f = f_T, \quad A_i = 1$$

$$h_{fe} = \sqrt{1 + \left(\frac{f_T}{f_\beta}\right)^2}$$

$$(h_{fe})^2 = 1 + \left(\frac{f_T}{f_\beta}\right)^2 \cong \left(\frac{f_T}{f_\beta}\right)^2$$

$$h_{fe} \simeq \frac{f_T}{f_\beta} \text{ when } Ai = 1$$

$$\boxed{f_T \simeq h_{fe} \cdot f_\beta}$$

$$f_\beta = \frac{g_m}{h_{fe}\{C_e + C_c\}}$$

$$f_T = f_\beta \cdot h_{fe} = \frac{g_m}{2\pi(C_e + C_c)}$$

$$C_e >> C_c$$

$$\boxed{f_T \simeq \frac{g_m}{2\pi C_e}}$$

$$A_i = \frac{-g_m}{g_{b'e} + j\omega(C_e + C_c)}$$

Dividing by $g_{b'e}$, Numerator and Denominator,

$$A_i = \frac{-g_m / g_{b'e}}{1 + \frac{j2\pi f (C_e + C_c)}{g_{b'e}}}$$

we know that $g_{b'e} = \frac{g_m}{h_{fe}}$

$\therefore$ $\frac{g_m}{g_{b'e}} = h_{fe}$

$$A_i = \frac{-h_{fe}}{1 + jf\left[\frac{2\pi(C_e + C_C)}{g_{b'e}}\right]}$$

But we know that $A_i = \frac{-h_{fe}}{1 + j\frac{f}{f_\beta}}$

Comparing, $f_\beta = \frac{g_{b'e}}{2\pi(C_e + C_c)} = \frac{g_m}{h_{fe}.2\pi(C_e + C_c)}$ $\because$ $g_{b'e} = \frac{g_m}{h_{fe}}$

$\therefore$

$$\boxed{f_\beta = \frac{g_m}{h_{fe}.2\pi(C_e + C_c)}}$$

$$\boxed{f_T = \frac{g_m}{2\pi(C_e + C_c)}}$$

Gain - Bandwidth (B.W) Product

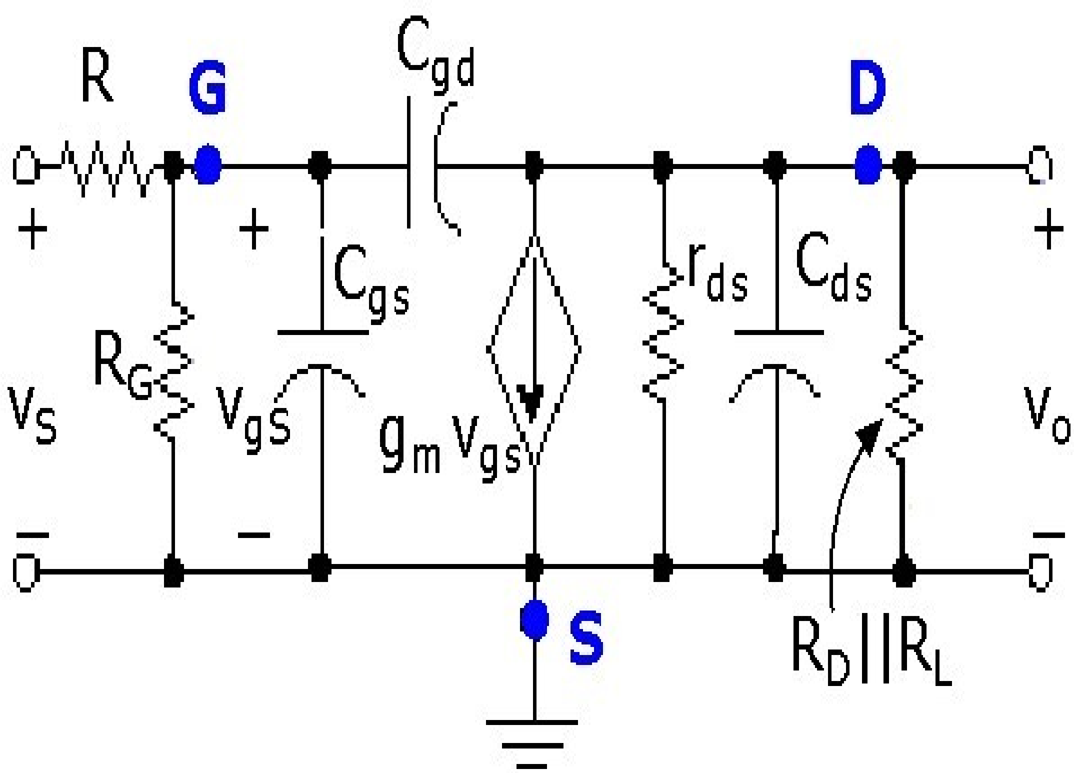
This is a measure to denote the performance of an amplifier circuit. Gain - B. W product is also referred as Figure of Merit of an amplifier. Any amplifier circuit must have large gain and large bandwidth. For certain amplifier circuits, the mid band gain Am maybe large, but not Band width or Vice - Versa. Different amplifier circuits can be compared with thus parameter.

FET: Analysis of common Source and common drain Amplifier circuits at high frequencies.

Just like for the BJT, we could use the original small signal model for low frequency analysis–the only difference was that external capacitances had to be kept in the circuit. Also just like the BJT, for high frequency operation, the internal capacitances between each of the device's terminals can no longer be ignored and the small signal model must be modified. Recall that for high frequency operation, we're stating that external capacitances are so large (in relation to the internal capacitances) that they may be considered short circuits.

High frequency response of Common source amplifier

The JFET implementation of the common-source amplifier is given to the left below, and the small signal circuit in corporating the high frequency FET model is given to the right below. As stated above, the external coupling and bypass capacitors are large enough that we can model them as short circuits for high frequencies.



We may simplify the small signal circuit by making the following approximations and observations:

1. R_{ds} is usually larger than $R_D||R_L$, so that the parallel combination is dominated by $R_D||R_L$ and r_{ds} may be neglected. If this is not the case, a single equivalent resistance, $r_{ds}||R_D||R_L$ may be defined.

2. The Miller effect transforms C_{gd} into separate capacitances seen in the input and output circuits as

$$C_{M1} = C_{gd}(1 - A_v) \quad (input\ circuit)$$
$$C_{M2} = C_{gd}\left(1 - \frac{1}{A_v}\right) (output\ circuit).$$

3. C_{ds} is very small, so the impedance contribution of this capacitance may be considered to be an open circuit and may be ignored.

$$C_{in} = C_{gs} + C_{M1} = C_{gs} + C_{gd}(1 - A_v).$$

4. The parallel capacitances in the input circuit, C_{gs} and C_{M1}, may be combined to a single equivalent capacitance of value

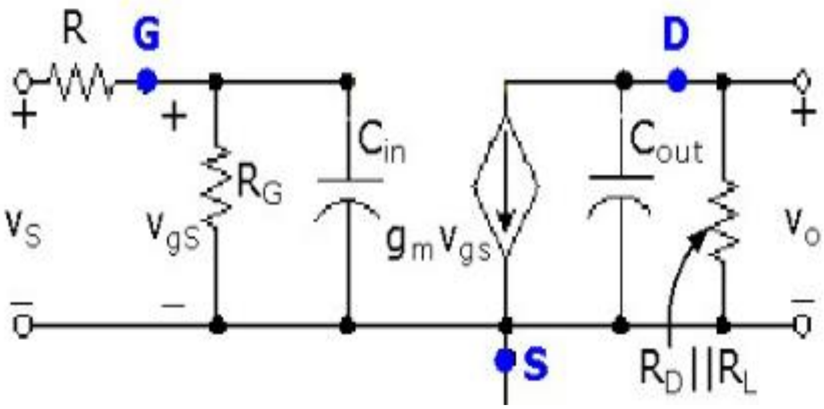
$$C_{in} = C_{gs} + C_{M1} = C_{gs} + C_{gd}(1 - A_v).$$

5. Similarly, the parallel capacitances in the output circuit, C_{ds} and C_{M2}, may be combined to a single equivalent capacitance of value

$$C_{out} = C_{ds} + C_{M2} = C_{ds} + C_{gd}\left(1 - \frac{1}{A_v}\right),$$

Where $A_v = -g_m(R_D||R_L)$ for a common-source amplifier.

Setting the input source, v_S, equal to zero allows us to define the equivalent resistances seen by C_{in} and C_{out}(the Method of Open Circuit Time Constants).Note that, with v_S=0, the dependent current source also goes to zero (opens) and the input and output circuits are separated.



$$R_{Cin} = R \,||\, R_G \,.$$

$$R_{Cout} = R_D \,||\, R_L \,.$$

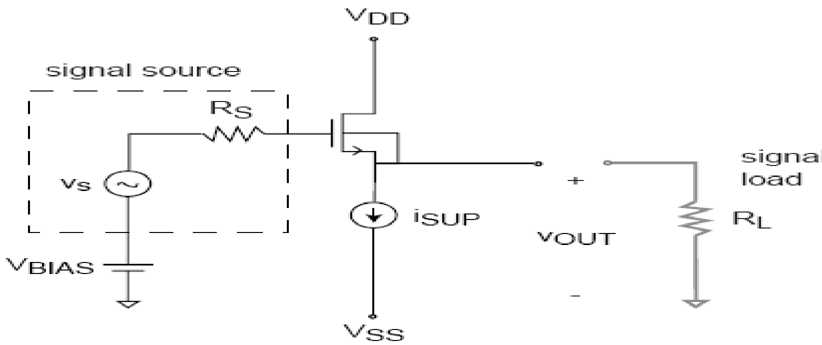
$$\tau_{Cin} = C_{in} R_{Cin} \,; \qquad \tau_{Cout} = C_{out} R_{Cout} \,,$$

$$\omega_H = \frac{1}{\frac{1}{\omega_{Cin}} + \frac{1}{\omega_{Cout}}} = \frac{1}{\tau_{Cin} + \tau_{Cout}} = \frac{1}{C_{in} R_{Cin} + C_{out} R_{Cout}} = \frac{1}{C_{in}(R \,||\, R_G) + C_{out}(R_D \,||\, R_L)}$$

Generally, the input is going to provide the dominant pole, so the high frequency cut off is given by

$$\omega_H = \frac{1}{C_{in}(R \,||\, R_G)} \,; \qquad f_H = \frac{\omega_H}{2\pi} = \frac{1}{2\pi C_{in}(R \,||\, R_G)} \,.$$

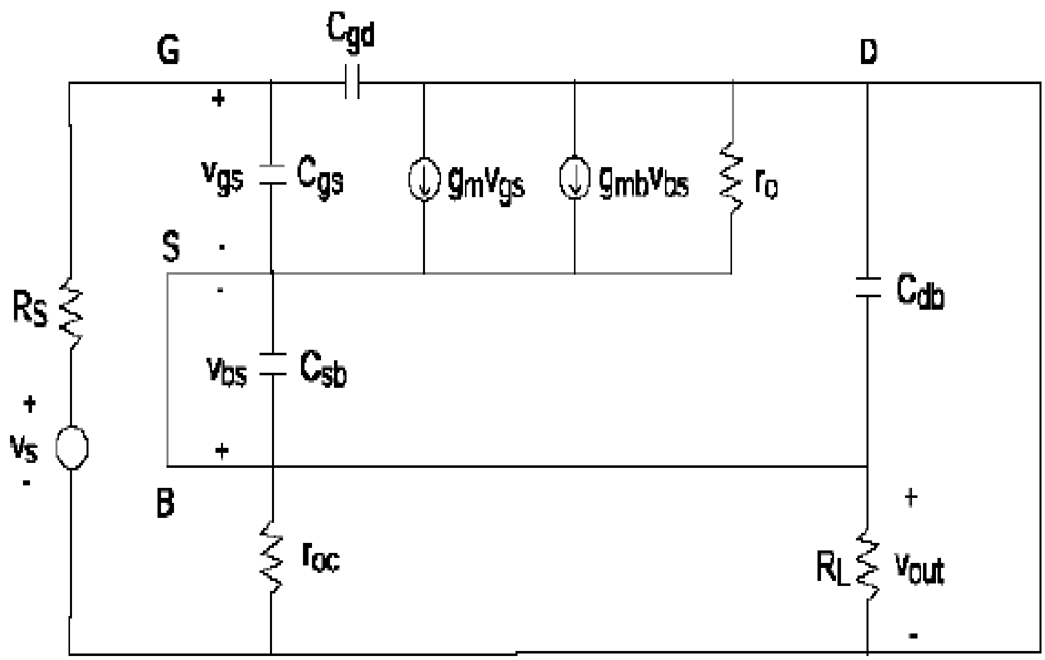
High frequency response of Common source amplifier

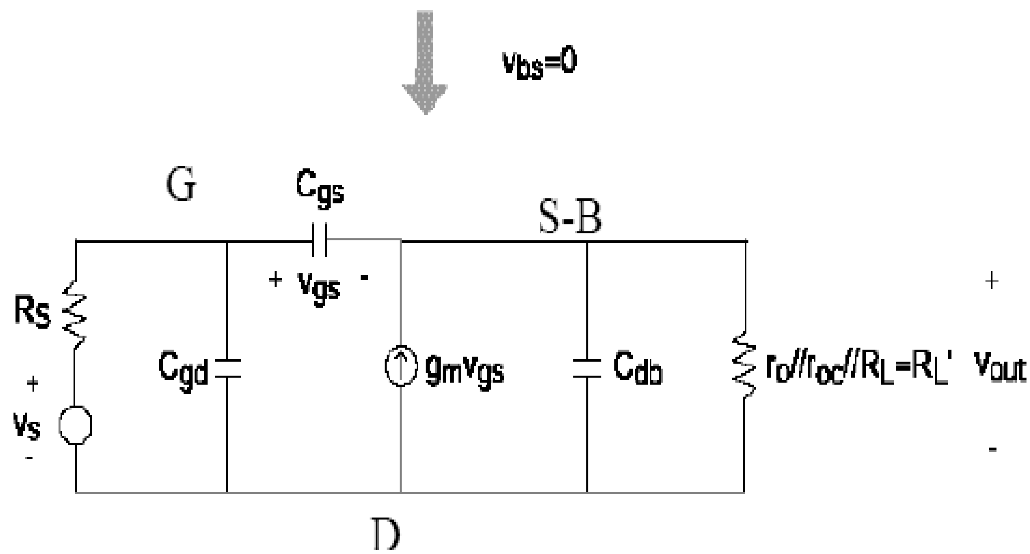


Characteristics ofCDAmplifier:

- Voltagegain ≈1
- Highinputresistance
- Lowoutputresistance
- Goodvoltage buffer

High frequency small signal model





$$A_{vC_{gs}} = \frac{R_L}{R_{out} + R_L} = \frac{g_m R_L}{1 + g_m R_L}$$

$$C_M = C_{gs}\left(1 - A_{vC_{gs}}\right) = C_{gs}\left(1 - \frac{g_m R_L}{1 + g_m R_L}\right) = \frac{C_{gs}}{1 + g_m R_L}$$

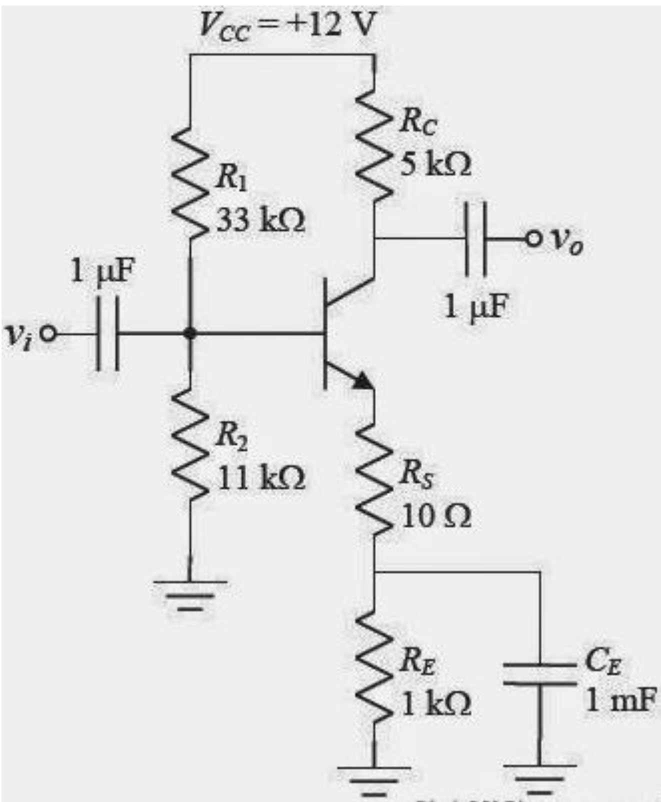
$$C_T = C_{gs}\left(1 - \frac{g_m R_L}{1 + g_m R_L}\right) + C_{gd}$$

$$R_T = R_S \parallel R_{in} = R_S$$

$$R_{C_{db}} = R_{out} \parallel R_L = \frac{R_{out} R_L}{R_{out} + R_L} = \frac{R_L}{1 + g_m R_L}$$

$$\omega_{3dB} \approx \frac{1}{R_S\left(\frac{C_{gs}}{1+g_m R_L} + C_{gd}\right) + C_{db}\frac{R_L}{1+g_m R_L}}$$

If R_S is not too high, bandwidth can be rather high and approach ω_T



UNIT-II

Multistage Amplifiers : Classification of amplifiers, methods of coupling**, c**ascaded transistor amplifier and its analysis, analysis of two stage RC coupled amplifier, high input resistance transistor amplifier circuits and their analysis-Darlington pair amplifier, Cascode amplifier, Boot-strap emitter follower, Analysis of multi stage amplifiers using FET, Differential amplifier using BJT.

Classification of amplifiers

Depending upon the type of coupling, the multistage amplifiers are classified as :

1. Resistance and Capacitance Coupled Amplifiers (RC Coupled)
2. Transformer Coupled Amplifiers
3. Direct Coupled DC Amplifiers
4. Tuned Circuit Amplifiers.

Based upon the B. W. of the amplifiers, they can be classified as :

1. Narrow hand amplifiers
2. Untuned amplifiers

Narrow hand amplifiers: Amplification is restricted to a narrow band of frequencies around a centre frequency. There are essentially tuned amplifiers.

Untuned amplifiers: These will have large bandwidth. Amplification is desired over a considerable range of frequency spectrum.

Untuned amplifiers are further classified w.r.t bandwidth.

I. DC amplifiers (Direct Coupled) DC to few KHz	
2. Audio frequency amplifiers (AF)	20 Hz to 20 KHz
3. Broad band amplifier	DC to few MHz
4. Video amplifier	100 Hz to few MHz

The gain provided by an amplifier circuit is not the same for all frequencies because the reactance of the elements connected in the circuit and the device reactance value depend upon the frequency. Bandwidth of an amplifier is the frequency range over which the amplifier stage gain is reasonably constant within ± 3 db, or O. 707 of A_V Max Value.

Resistance and Capacitance Coupled Amplifiers (RC Coupled)

This type of amplifier is very widely used. It is least expensive and has good frequency response. In the multistage resistive capacitor coupled amplifiers, the output of the first stage is coupled to the next through coupling capacitor and R_L. In two stages Resistor Capacitor coupled amplifiers, there is no separate R_L between collector and ground, but Reo the resistance between collector and V cc (R_C) itself acts as R_L in the AC equivalent circuit.

Transformer Coupled Amplifiers
`

Here the output of the amplifier is coupled to the next stage or to the load through a transformer. With this overall circuit gain will be increased and also impedance matching can be achieved. But such transformer coupled amplifiers will not have broad frequency response i.e., (f_2-f_1) is small since inductance of the transformer windings will be large. So Transformer coupling is done for power amplifier circuits, where impedance matching is critical criterion for maximum power to be delivered to the load.

Direct Coupled (DC) Amplifiers

Here DC stands for direct coupled and not (direct current). In this type, there is no reactive element. L or C used to couple the output of one stage to the other. The AC output from the collector of one stage is directly given to the base of the second stage transistor directly. So type of amplifiers is used for large amplification of DC and using low frequency signals. Resistor Capacitor coupled amplifiers cannot be used for amplifications of DC or low frequency signals since Xc the capacitive reactance of the coupling capacitor will be very large or open circuit for DC

Tuned Circuit Amplifiers

In this type there will be one RC or LC tuned circuit between collector and V_{CC} in the place of Re. These amplifiers will amplify signals of only fixed frequency.fo which is equal to the resonance frequency of the tuned circuit LC. These are also used to amplify signals of a narrow band of frequencies centered on the tuned frequency f_0.

Distortion in Amplifiers

If the input signal is a sine wave the output should also be a true sine wave. But in all the cases it may not be so, which we characterize as distortion. Distortion can be due to the nonlinear characteristic of the device, due to operating point not being chosen properly, due to large signal swing of the input from the operating point or due to the reactive elements Land C in the circuit. Distortion is classified as:

(**a**)**Amplitude distortion**:

This is also called non linear distortion or harmonic distortion. This type of distortion occurs in large signal amplifiers or power amplifiers. It is due to then on linearity of the characteristic of the device. This is due to the presence of new frequency signals which are not present in the input. If the input signal is of 10 KHz the output signal should also be 10 KHz signal. But some harmonic terms will also be present. Hence the amplitude of the signal (rms value) will be different Vo = Ay Vi.

(b) **Frequency distortion**:

The amplification will not be the same for all frequencies. This is due to reactive component in the circuit.

(c) **Phase - shift delay distortion**:

There will be phase shift between the input and the output and this phase shift will not be the same for all frequency signals. It also varies with the frequency of the input signal. In the output signal, all these distortions may be present or anyone may be present because of which the amplifier response will not be good.

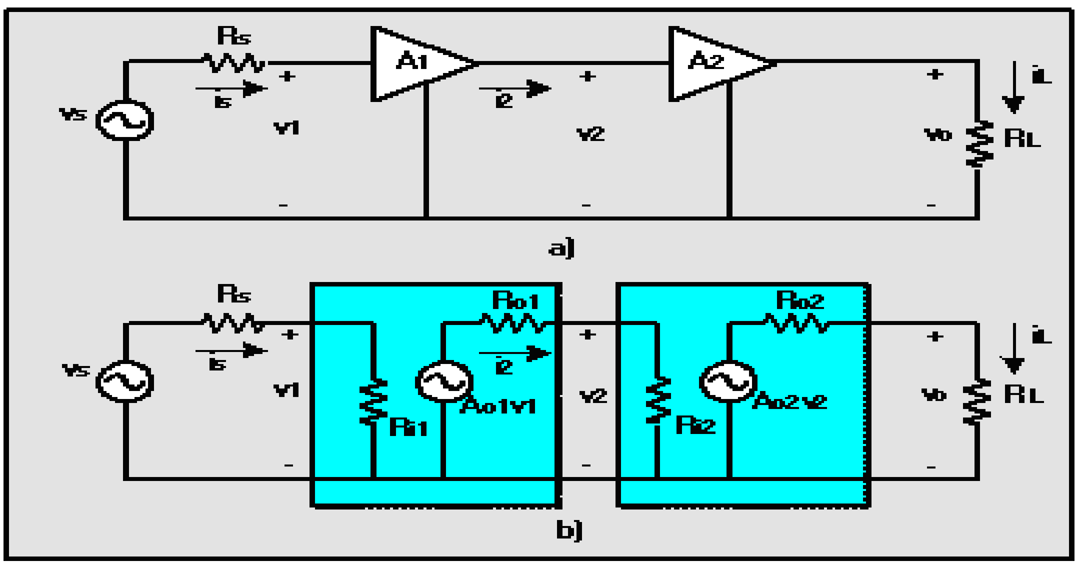
The performance obtainable from a single stage amplifier is often insufficient for many applications; hence several stages may be combined forming a multistage amplifier. These stages may be combined forming a multistage amplifier. These stages are connected in cascade, i.e. output of the first stage is connected to form input of second stage, whose output becomes input of third stage, and so on. The overall gain of a multistage amplifier is the product of the gains of the individual stage (ignoring potential loading effects):

Gain (A) = A1 * A2 * A3 * A4 * …… *An.

Alternately, if the gain of each amplifier stage is expressed in decibels (dB), the total gain is the sum of the gains of the individual stages

Gain in dB (A) = A1 + A2 + A3 + A4 + …… + An.

When we want to achieve higher amplification than a single stage amplifier can offer, it is a common practice to cascade various stages of amplifiers, as it is shown in Fig.1.a. In such a structure the input performance of the resulted multistage amplifier is the input performance of the first amplifier while the output performance is that of the last amplifier. It is understood that combining amplifiers of various types we can create those characteristics that are necessary to fulfill the specifications of a specific application. In addition, using feedback techniques in properly chosen multistage amplifiers can further increase this freedom of the design.



According to the small signal equivalent circuit of a two stage amplifier shown in Fig., we can calculate the ac performance of the circuit.

Voltage amplification

$$A_v = \frac{V_o}{V_1} = \frac{V_o}{V_2} \cdot \frac{V_o}{V_1} = A_{v2} \cdot A_{v1}$$

Current amplification

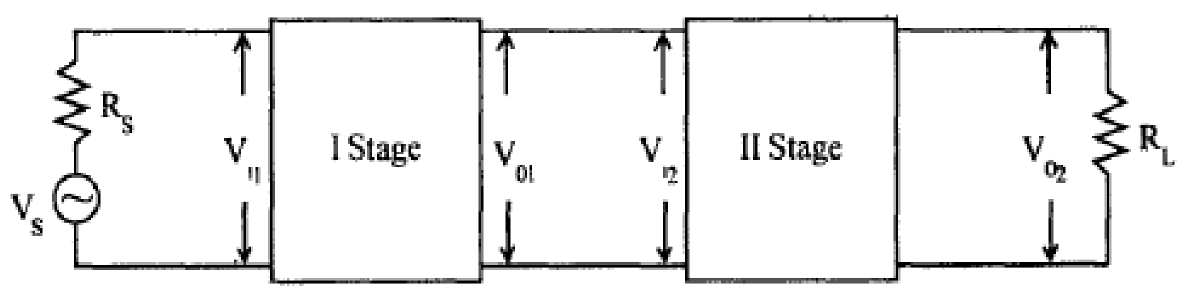
$$A_i = \frac{i_L}{i_s} = \frac{i_L}{i_2} \cdot \frac{i_2}{i_s} = A_{i2} \cdot A_{i1}$$

Power amplification

$$A_p = A_v \cdot A_i = (A_{v2} \cdot A_{i2}) \cdot (A_{v1} \cdot A_{i1}) = A_{p2} \cdot A_{p1}$$

Cascading Transistor Amplifiers

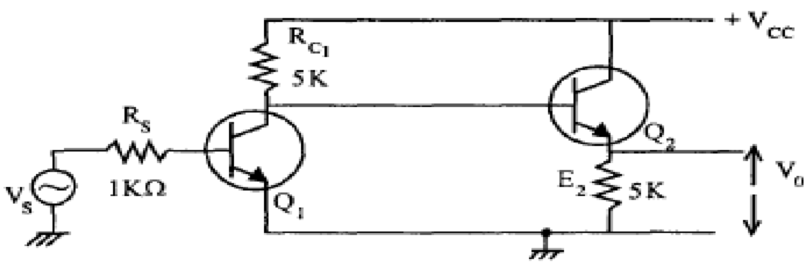
When the amplification of a single transistor is not sufficient for a particular purpose (say to deliver output to the speaker or to drive a transducer etc) or when the input or output impedance is not of the correct magnitude for the desired application, two or more stages may be connected in cascade. Cascade means in series i.e. the output of first stage is connected to the input of the next stage.



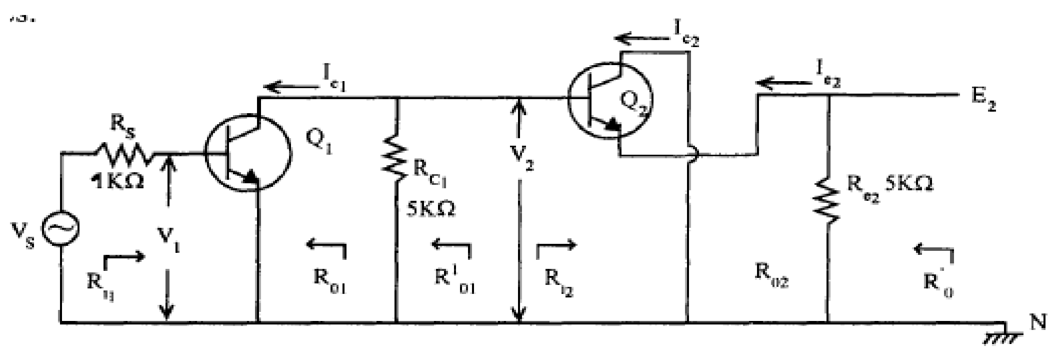
Let us consider two stage cascaded amplifier. Let the first stage is in common emitter configuration. Current gain is high and let the II stage is in common collector configuration to provide high input impedance and low output impedance. So what are the expressions for the total current gain A_I of the entire circuit (i.e. the two stages), Z_i, A_v and Y_o? To get these expressions, we must take the h-par ammeters of these transistors in that particular configuration. Generally manufactures specify the h-parameters for a given transistor in common emitter configuration. It is widely used circuit and also A_I is high. To get the transistor h-parameters in other configurations, converts ion formulae are used.

The Two Stage Cascaded Amplifier Circuit

The Transistor Q_1 is in Common Emitter configuration. The second Transistor Q_2 is in Common Collector (CC) configuration. Output is taken across 5K, the emitter resistance. Collector is at ground potential in the A.C. equivalent circuit. Biasing resistors are not shown since their purpose in only to provide the proper operating point and they do not affect the response of the amplifier. In the low frequency equivalent circuit, since the capacitors have large value, and so is X_c low, and can be neglected. So the capacitive reactance is not considered, and capacitive reactance X_c is low when C is large and taken as short circuit.



The small signal Common Emitter configuration circuit reduces as shown in Fig. In this circuit Q2 collector is at ground potential, in AC equivalent circuit. It is in Common Collector configuration and the output is taken between emitter point E2 and ground. So the circuit is redrawn as shown in Figure indicating voltages at different stages and input and output resistances.



Choice of Transistor in a Cascaded Amplifier Configuration

By connecting transistor in cascade, voltage gain gets multiplied. But what type of configuration should be used? Common Collector(CC) or Common Base(CB) or Common

Emitter(CE)? To get voltage amplification and current amplification, only Common Emitter (CE) configuration is used. Since it is Common Collector amplifier, the voltage gain is less than one for each stage. So the overall amplification is less than 1.

Common Base Configuration is also not used since A_I is less than 1.

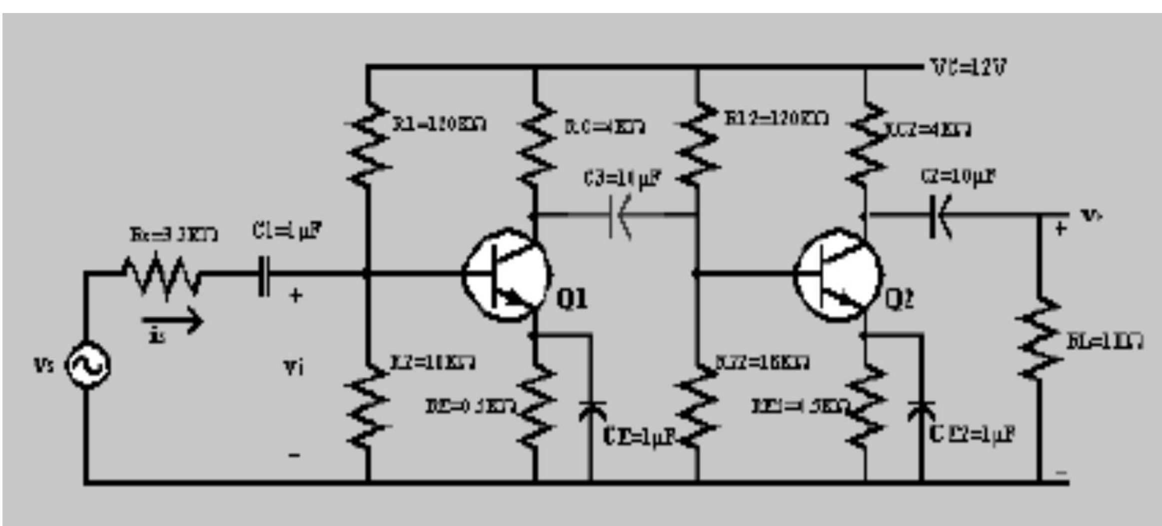
$$A_V = A_I \times \frac{R_L}{R_i}$$

Effective load resistance R_L is parallel combination of R_c and R_i of the following stage, (next stage) (since in multi stage connection, the output of one stage is the input to the other stage). This parallel combination is less than R_i. Therefore $R_L/R_i < 1$.

The current gain A_I in common base configurations is hib< 1 or =1.Therefore overall voltage gain = 1. Therefore Common Base configuration is not used for cascading. So only Common Emitter configuration is used (hfe>> I).Therefore overall voltage gain and current gains are> 1 in Common Emitter configuration.

Two stage RC coupled amplifier

One way to connect various stages of a multistage amplifier is via capacitors, as indicated in the two-stage amplifier in Figure. Where two stages of common emitter amplifier are coupled to each other by the capacitor C_3.



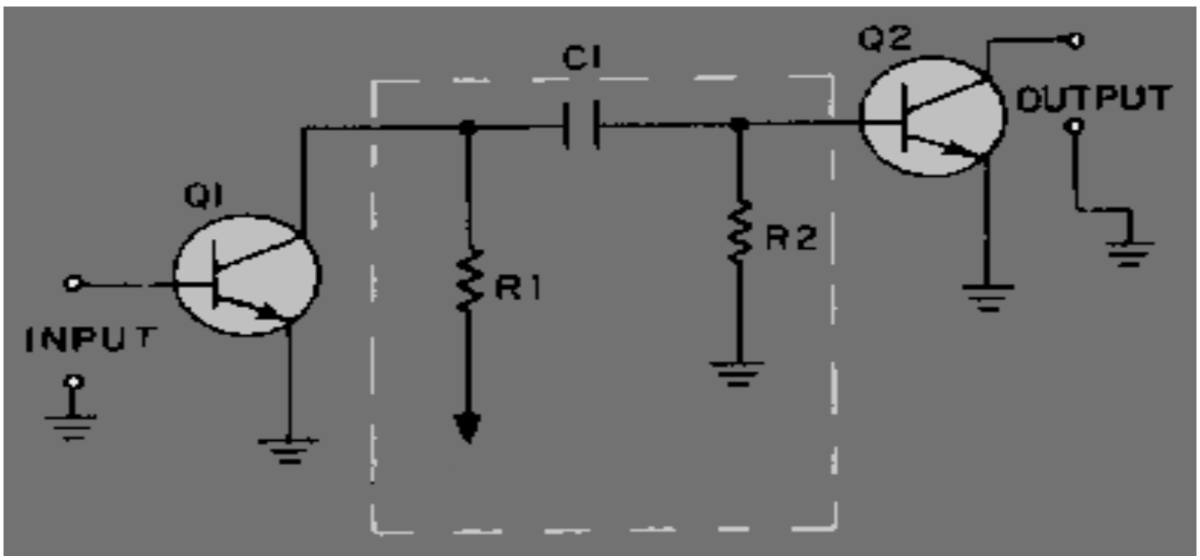
In RC-coupled amplifiers:

1. The various stages are DC isolated. This feature facilitates the biasing of individual stages.

2. The various stages can be similar. Hence the design of the amplifier is simplified.

3. The coupling capacitors influence the responses of the amplifier.

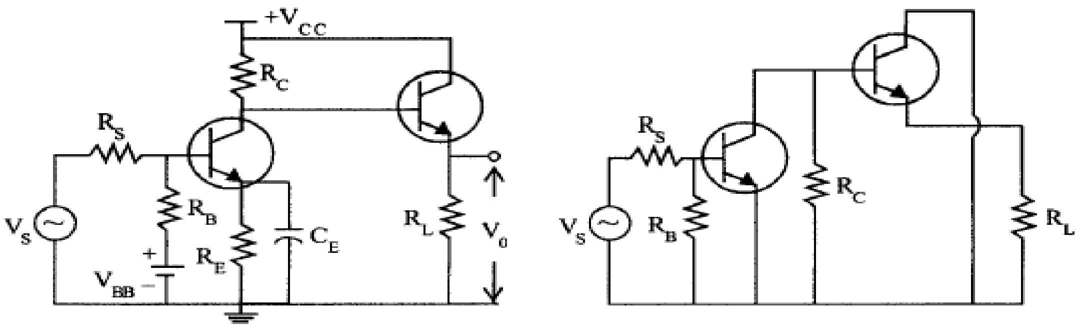
4. A great number of biasing resistors is necessary.



The most commonly used coupling in amplifiers is RC coupling. An RC-coupling network is shown in the illustration above. The network of R1, R2, and C1 enclosed in the dashed lines of the figure is the coupling network. You may notice that the circuitry for Q1 and Q2 is incomplete. That is intentional so that you can concentrate on the coupling network. R1 acts as a load resistor for Q1 (the first stage) and develops the output signal of that stage. Do you remember how a capacitor reacts to ac and dc? The capacitor, C1, "blocks" the dc of Q1's collector, but "passes" the ac output signal. R2 develops this passed, or coupled, signal as the input signal to Q2 (the second stage). This arrangement allows the coupling of the signal while it isolates the biasing of each stage. This solves many of the problems associated with direct coupling.

CE - CC Amplifiers

This is another type of two-stage BJT amplifier. The first stage in Common Emitter (CE) configuration provides voltage and current gains. The second stage in Common-Collector (CC) configuration provides impedance matching. This circuit is used in audio frequency amplifiers. The circuit is shown in Fig.



$$R_{L_2} \simeq R_L$$

$$h_{oc} \quad R_{L_1} \leq 0.1$$

$$A_{I2} = A_{I'2} = (1 + h_{fe})$$

$$R_{i_2} = (1 + h_{fe})\ R_{L_2}$$

$$A_{V2} = A_{V2'} = \frac{A_{I_2} . R_{L_2}}{R_{i_2}}$$

$$= \frac{\left(1 + h_{fe}\ R_{L_2}\right)}{h_{ie} + \left(1 + h_{fe}\right) R_{L_2}}$$

$$= 1 - \frac{h_{ie}}{R_{i_2}}\ ;$$

$$A_{V_2} < 1$$

$$A_{I_1} = A_{I_1}' = -h_{fe}$$

$$R_{i_1} = R_{i_1}' = h_{ie}$$

$$A_{V_1} = A_{V_1}' = (A_{I_1} \ . \ R_{L_1} / R_i)$$

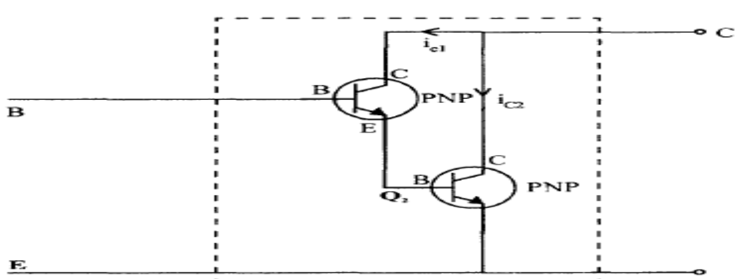
$$A_V = A_{V_1} \cdot A'_{V_2}$$
$$R_i = R_{i_1}$$
$$R_o = R_{o_1}$$
$$A_I = \frac{A_V . R_{i_1}}{R_{L_2}}$$

High Input Resistance Transistor Circuits

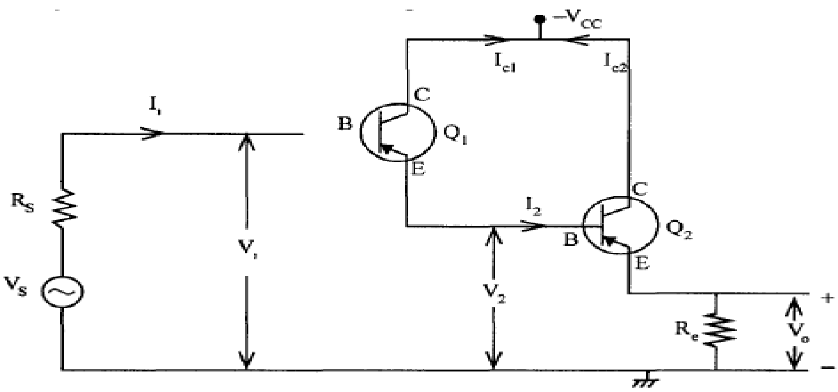
In some applications the amplifier circuit will have to have very high input impedance. Common Collector Amplifier circuit has high input impedance and low output impedance. But it's A_v<1.If the input impedance of the amplifier circuit is to be only 500 KO or less the Common Collector Configuration can be used. But if still higher input impedance is required a circuit. This circuit is known as the Darlington Connection (named after Darlington) or Darlington Pair Circuit.

The Darlington Pair

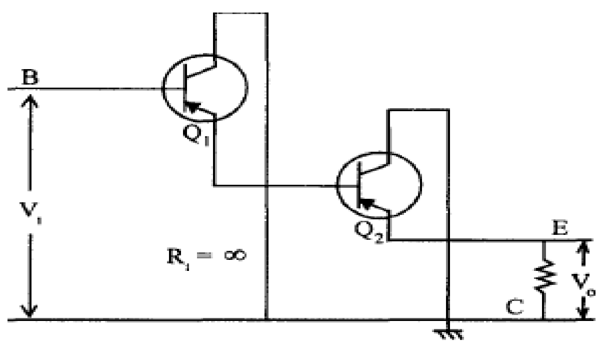
This is two transistors connected together so that the amplified current from the first is amplified further by the second transistor. This gives the Darlington pair a very high current gain such as 10000. Darlington pairs are sold as complete packages containing the two transistors. They have three leads (B, C and E) which are equivalent to the leads of a standard individual transistor.



In this circuit, the two transistors are in Common Collector Configuration. The output of the first transistor Q_1 (taken from the emitter of the Q_1) is the input to the second transistor Q_2 at the base. The input resistance of the second transistor constitutes the emitter load of the first transistor. So, Darlington Circuit is nothing but two transistors in Common Collector Configuration connected in series. The same circuit can be redrawn as AC equivalent circuit. So, DC is taken as ground shown in below Fig. Hence 'C' at ground potential, Collectors of transistors $Q_1$and Q_2 is at ground potential.



There is no resistor connected between the emitter of Q_1 and ground i.e., Collector Point. So, we can assume that infinite resistance is connected between emitter and collector.



The overall current gain is equal to the two individual gains multiplied together:

Darlington pair current gain, $h_{FE} = h_{FE1} \times h_{FE2}$

Here h_{FE1} and h_{FE2} are the gains of the individual transistors

If both the transistors are identical then

Current gain

$$A_I = \frac{I_c}{I_{b_1}} \cong (h_{fe})^2$$

Input resistance

$$R_i \simeq \frac{(1+h_{fe})^2 R_e}{1+h_{oe}h_{fe}R_e}$$

Voltage gain

$$A_v \cong \left(1 - \frac{h_{ie}}{Ri_2}\right)$$

Output resistance

$$R_{o2} = \frac{R_s + h_{ie}}{(1+h_{fe})^2} + \frac{h_{ie}}{1+h_{fe}}$$

Therefore, the characteristic of Darlington Circuit are

1. Very High Input Resistance
2. Very Large Current Gain
3. Very Low Output Resistance
4. Voltage Gain, $A_V < 1$.

This gives the Darlington pair a very high current gain, such as 10000, so that only a tiny base current is required to make the pair switch on.

A Darlington pair behaves like a single transistor with a very high current gain. It has three leads (B, C and E) which are equivalent to the leads of a standard individual transistor. To turn on there must be 0.7V across both the base-emitter junctions which are connected in series inside the Darlington pair, therefore it requires 1.4V to turn on.

Darlington pairs are available as complete packages but you can make up your own from two transistors; TR1 can be a low power type, but normally TR2 will need to be high power. The maximum collector current Ic(max) for the pair is the same as Ic(max) for TR2.

A Darlington pair is sufficiently sensitive to respond to the small current passed by your skin and it can be used to make a touch-switch as shown in the diagram. For this circuit which just lights an LED the two transistors can be any general purpose low power transistors. The 100k resistor protects the transistors if the contacts are linked with a piece of wire. Two transistors may be combined to form a configuration known as the Darlington pair which behaves like a single transistor with a current gain equivalent to the product of the current gain of the two transistors. This is especially useful where very high currents need to be controlled as in a power amplifier or power-regulator circuit. Darlington transistors are available whereby two transistors are combined in one single package. The base-emitter volt-drop is twice that of a small transistor.

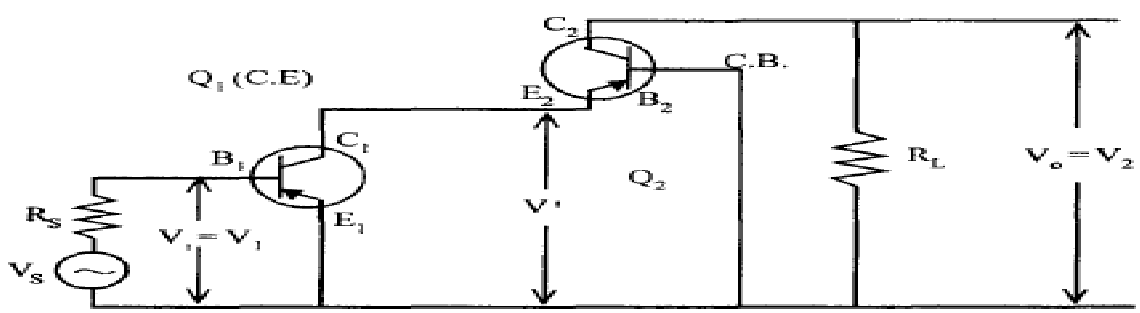
Disadvantages

1. The h-parameters for both the transistors will not be the same.
2. Leakage Current is more

The CASCODE Transistor Configuration

The circuit is shown in Figure. This transistor configuration consists of a Common Emitter Stage in cascade with a Common Base Stage. The collector current of transistor Q) equals the emitter current of Q_2.

The transistor Q_1 is in Common Emitter Configuration and transistor Q_2 is in Common Base Configuration. Let us consider the input impedance (h_{11}) etc., output admittance (h22) i.e. the h - parameters of the entire circuit in terms of the h- parameters of the two transistors



Input impedance

$$h_{11} = \text{Input } Z = \left.\frac{V_1}{I_1}\right|_{V_2 = 0}$$

$$h_{11} \cong h_{ie}$$

Short circuit current gain

$$h_{21} = \left.\frac{I_2}{I_1}\right|_{V_2 = 0}$$

$$h_{21} = \frac{I_2}{I_1} = \frac{I}{I_1} \times \left.\frac{I_2}{I_0}\right|_{V_2 = 0}$$

$$\frac{I'}{I_1} = h_{fe} \qquad \text{since, } I = I_{C_1}.\ I_1 = I_{B_1}$$

$$\frac{I_2}{I'} = -h_{fb} \qquad \text{since, } I = I_{E_2}.\ I_2 = I_{C_2}.$$

$$h_{21} = -h_{fe}.\ h_{fb}.$$

$$h_{fe} >> 1\ .\ -h_{fb} \simeq 1, \qquad \text{since } h_{fb} = \frac{I_C}{I_E}$$

$$\boxed{h_{21} \cong h_{fe}}$$

Output conductance

$$h_{22} = \left.\frac{I_2}{V_2}\right|_{I_1 = 0}$$

Reverse voltage gain

$$h_{12} = \left.\frac{V_1}{V_2}\right|_{I_1 = 0}$$

$$= \left.\frac{V_1}{V'} \times \frac{V^1}{V_2}\right|_{I_1 = 0}$$

$$\left.\frac{V_1}{V_0'}\right|_{I_i = 0} = h_{re} \cdot \left.\frac{V'}{V_2}\right| = h_{rb}$$

$h_{12} \cong h_{re}\, h_{rb}.$

$h_{re} \cong 10^{-4}$ $\quad h_{rb} = 10^{-4}.$ $\because$ h_{12} is very small

$h_i = h_{11} \cong h_{ie}.$ Typical value = 1.1KΩ

$h_f = h_{21} \cong h_{fe}.$ Typical value = 50

$h_o = h_{22} \cong h_{ob}.$ Typical value = 0.49 μ A/V

$h_r = h_{12} \cong h_{re}\, h_{rb}.$ Typical value = 7×10^{-8}.

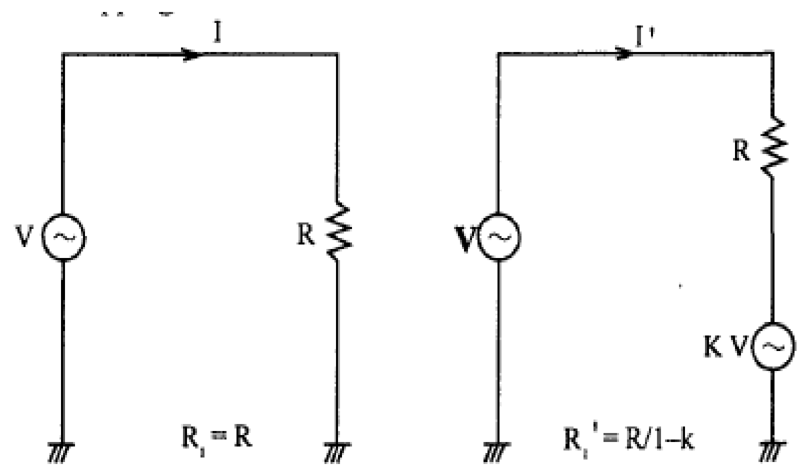
Therefore, for a CASCODE Transistor Configuration, its input Z is equal to that of a single Common Emitter Transistor (hie)' Its Current Gain is equal to that of a single Common Base Transistor (h_{fe}). Its output resistance is equal to that of a single Common Base Transistor (hob)' the reverse voltage gain is very small, i.e., there is no link between V_1 (input voltage) and V_2 (output voltage). In other words, there is negligible internal feedback in the case of, a CASCODE Transistor Circuit, acts like a single stage C.E. Transistor (Since hie and hfe are same) with negligible internal feedback (:.hre is very small) and very small output conductance, (= hob) or large output resistance (=2MΩ equal to that of a Common Base Stage). The above values are correct, if we make the assumption that hob R_L< 0.1 or RL is <200K.

CASCODE Amplifier will have

1. Very Large Voltage Gain.
2. Large Current Gain
3. Very High Output Resistance.

Boot-strap emitter follower

The maximum input resistance of a practical Darlington Circuit is only 2 MΩ. Higher input resistance cannot be achieved because of the biasing resistors R_1, R_2 etc. They come in parallel with R_i of the transistors and thus reduce the value of Ri. The maximum value of Ri is only 1/hob since, hob is resistance between base and collector. The input resistance can be increased greatly by boot strapping, the Darlington Circuit through the addition of Co between the first collector C_1 and emitter B_2.

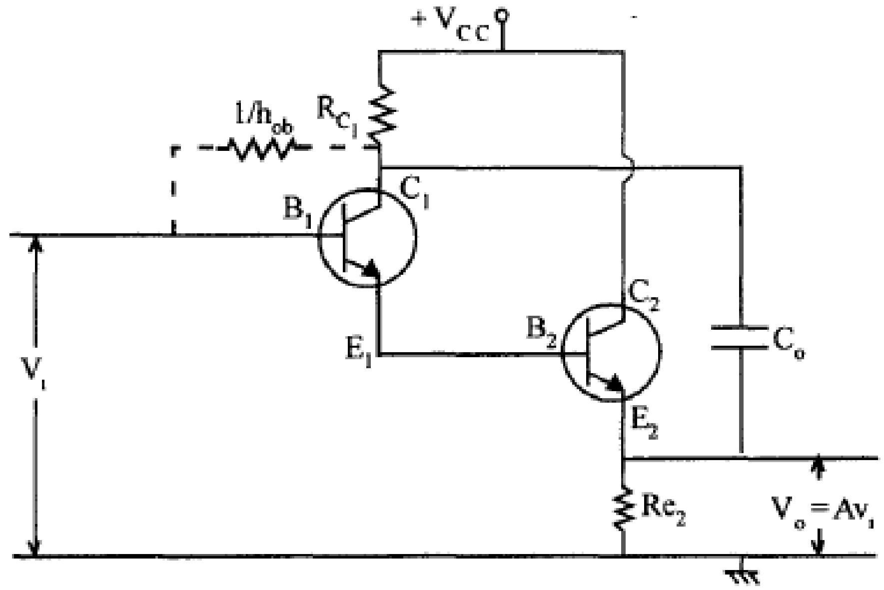


Boot strapping

In Fig, V is an AC signal generator, supplying current I to R. Therefore, the input resistance of V seen by the generator is $R_1 = V/I = R$ itself. Now suppose, the bottom end of R is not at ground potential but at higher potential i.e. another voltage source of KV (K < I) is connected between the bottom end of R and ground. Now the input resistance of the circuit is

$$R_i' = \frac{V}{I'} \qquad I' = \frac{(V - KV)}{R}$$

$$R_i' = \frac{VR}{V(1-K)} = \frac{R}{1-K}$$

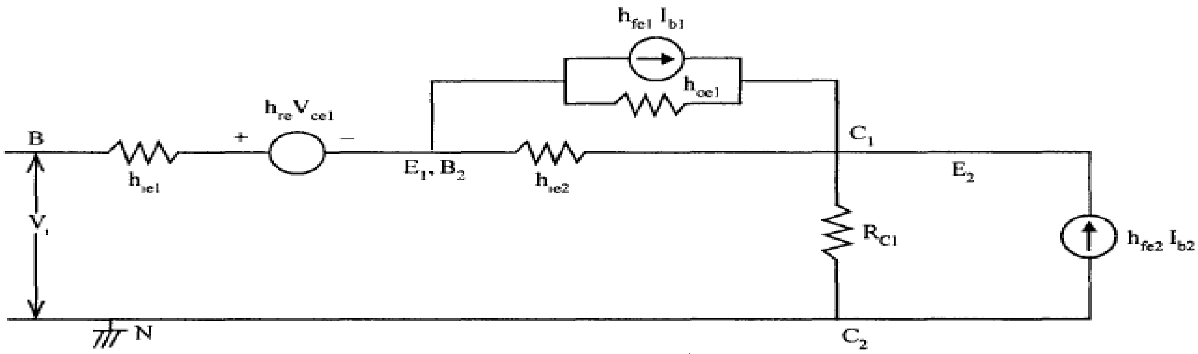


I' can be increased by increasing V. When V increases KV also increases. K is constant.Therefore the potential at the two ends of R will increase by the same amount, K is less than 1, therefore R_i> R. Now if K = 1, there is no current flowing through R (So V = KV there is nopotential difference). So the input resistance R\ = 00. Both the top and bottom of the resistorterminals are at the same potential. This is called as the Boots Strapping method which increases theinput resistance of a circuit. If the potential at one end of the resistance changes, the other end of Ralso moves through the same potential difference. It is as if R is pulling itself up by its boot straps.For CC amplifiers A_v< 1 = 0.095. So R_i can be made very large by this technique. K = Av = 1.Ifwe pull the boot with both the edges of the strap (wire) the boot lifts up. Here also, if the potentialat one end ofR is changed, the voltage at the other end also changes or the potential level ofR_3 rises,as if it is being pulled up from both the ends.

$$R_i = \frac{h_{ie}}{1 - A_V}; \quad A_V \cong 1.$$

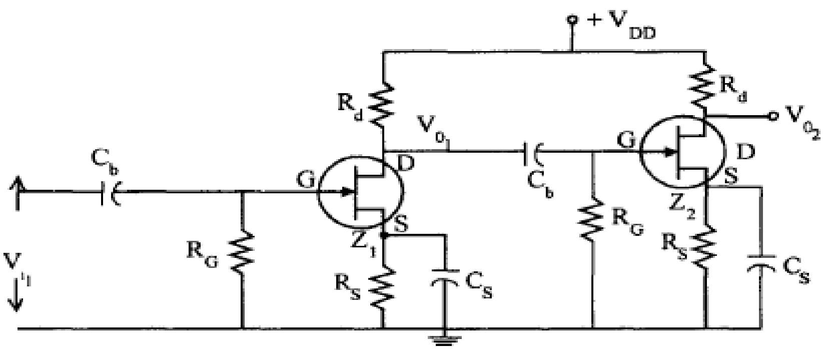
$$R_i = \frac{R}{1-K}$$

AC Equivalent circuit



Two Stage RC Coupled JFET amplifier (in Common Source (CS) configuration)

The circuit for two stages of RC coupled amplifier in CS configuration is as shown in fig.



The output Vo of I Stage is coupled to the input Vi of II Stage through a blocking capacitor C_b. It blocks the DC components present in the output or I Stage from reaching the input of the I stage which will alter the biasing already fixed for the active device. Resistor Rg is connected between gate and ground resistor Ro is connected between drain and V_{DD} supply. C_S is the bypass capacitor used to prevent loss of gain due to negative feedback. The active device is assumed to operate in the linear region. So the small signal model of the device is valid. Frequency Roll-off is the term used for the decrease in gain with frequency in the uppercut-off region. It is expressed as db/octave on db/decade.

The purpose of multistage amplifiers is to get large .gain. So with BJTs, Common Emitter Configuration is used. If JFETs are employed, common source configuration is used.

Difference Amplifier

This is also known as differential amplifier. The function of this is to amplify the difference between the signals. The advantage with this amplifier is, we can eliminate the noise in the input signals which is common to both the inputs. Thus SIN ratio can be improved. The difference amplifier can be represented as a black box with two inputs V_1 and V_2 and output V_0 where $V_0 = A_d (V_1 - V_2)$'.

Where A_d is the gain of the differential amplifier. But the above equation will not correctly describe the characteristic of a differential amplifier. The output Vo depends not only on the difference of the two signals (V_1- V 2) = V d but also on the average level called common mode signal V c = (V_1 + V 2)/ 2.

$$V_d = V_1 - V_2$$

$$V_c = \frac{1}{2}(V_1 + V_2)$$

$$V_1 = V_c + \frac{1}{2} V_d$$

$$V_2 = V_c - \frac{1}{2} V_d$$

$$V_0 = A V + A_2 V_2$$

$$V_0 = A_1 [V_c + \frac{1}{2} v_d] + A_2 [V_c - \frac{1}{2} V_d]$$

$$= A_1 V_c + \frac{A_1}{2} V_d + A_2 V_c - \frac{A_2}{2} V_d$$

$$V_0 = V_c (A_1 + A_2) + V_d \left[\frac{A_1 - A_2}{2}\right]$$

$$V_0 = V_c A_c + V_d A_d$$

$$A_d = \frac{A_1 - A_2}{2} \text{ and } A_c = A_1 + A_2.$$

A_l and A_2 are the voltage gains of the two amplifier circuits separately.

The voltage gain from the difference signal is Ad'

The voltage gain from the common mode signal is Ac'

$$V_0 = A_d V_d + A_c V_c$$

To measure Ad' directly set $V_I = - V_2 = 0.5V$ so that

$$V_d = 0.5 - (-0.5) = 1V.$$

$$V_c = \frac{(0.5 - 0.5)}{2} = 0$$

$$V_0 = A_d \,.\, 1 = A_d \text{ it self}$$

Output voltage directly gives the value of Ad'
Similarly if we set $V_I = V_2 = I$ V. then

$$V_a = 0, \quad V_c = \frac{V_1 + V_2}{2} = \frac{2}{2} = 1V.$$

$$V_0 = 0 + A_c \,.\, 1 = A_c.$$

The measured output voltage directly gives Ac. We want Ad to be large and Ac to be very small because only the difference of the two signals should be amplified and the average of the signals should not be amplified. :. The ratio of the two gains p $=A_d/A_c$ is called the common mode rejection ratio. This should be large for a good difference amplifier.

$$V_O = A_d V_d + A_c V_c$$

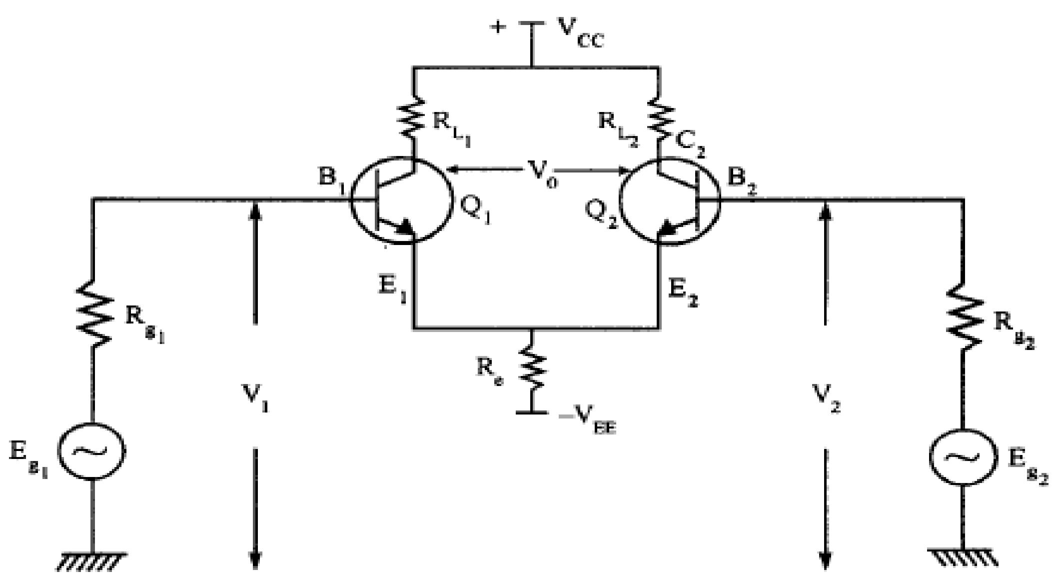
$$\rho = \frac{A_d}{A_c} \qquad \therefore \quad A_c = \frac{A_d}{\rho}$$

$$V_O = A_d V_d + \frac{A_d}{\rho} \cdot V_c$$

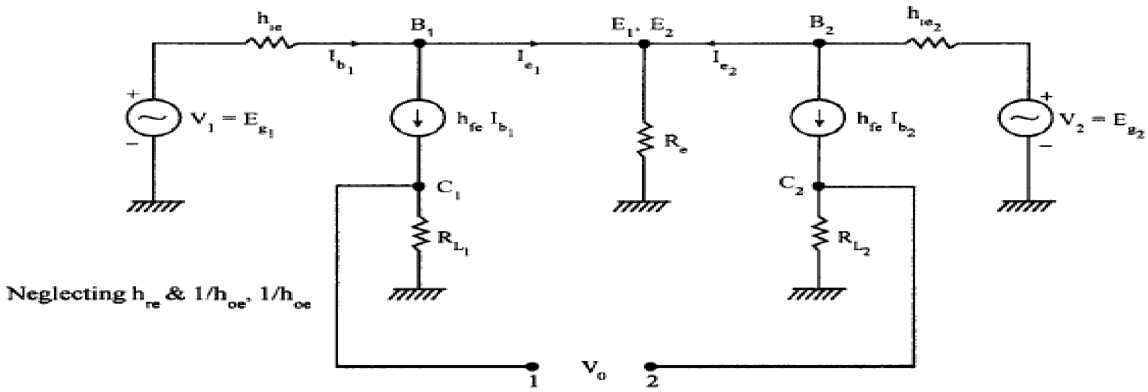
$$V_O = A_d V_d \left(1 + \frac{1}{\rho} \cdot \frac{V_c}{V_d}\right)$$

Circuit for Differential Amplifier

In the previous D.C amplifier viz., C.B, C.C and C.E, the output is measured with respect to ground. But in difference amplifier, the output is proportional to the difference of the inputs. So Vo is not measured w.r.t ground but w.r.t to the output of one transistor Q_1 or output of the other transistor Q_2'.



Equivalent Circuit



The advantage with this type of amplifiers is the drift problem is eliminated. Drift means, even when there is no input, Vi there can be some output Vo which is due to the internal thermal noise of the circuit getting amplified and coming at the output. Drift is reduced in this type of circuit, because, the two points should be exactly identical. Hence, I_E ' h_{FE}' V_{BE} will be the same for the two transistors. Now if I_E rises across R_L (IeR_L) increases with increase in I_{cs}. So the voltage at collector of Q_1decreases. If Q2is also identical Q_1 its collector voltage also I drops by the same amount. Hence Vo which is the difference of these voltages remains' ~he same thus the drift of these transistors gets cancelled.

The input to a differential amplifier is of two types.

1. Differential mode
2. Common mode.

If V_1 and V_2 are the inputs, the differential mode input = $V_2–V_1$

Here two different a.c. signals are being applied V_1& V_2' So these will be interference of these signals and so both the signals will be present simultaneously at both input points i.e., if V_1 is applied at point I, It also prices up the signal V_2 and so the net input is (V I + V 2)' This is due to interference.

Common node voltage = $(V_1+V_2)/2$

An ideal differential amplifier must provide large gain to the differential mode inputs and zero gain to command input.

$$V_0 = A_2 V_2 - A_1 V_1$$

A2 = voltage gain of the transistor Q_2
A1 = voltage gain of the transistor Q_1

We can also express the output in term of the common mode gain Ac and differential gain Ad'

$$V_0 = A_d (V_2 - V_1) + A_c \left(\frac{V_1 + V_2}{2} \right)$$

$$= A_d V_2 - A_d \cdot V_1 + A_c \cdot \frac{V_1}{2} + A_c \cdot \frac{V_2}{2}$$

$$V_0 = V_2 \left(A_d + \frac{A_c}{2} \right) - V_1 \left(A_d - \frac{A_c}{2} \right)$$

$$A_2 = A_d + \frac{A_c}{2}$$

$$A_i = A_d + \frac{A_c}{2}$$

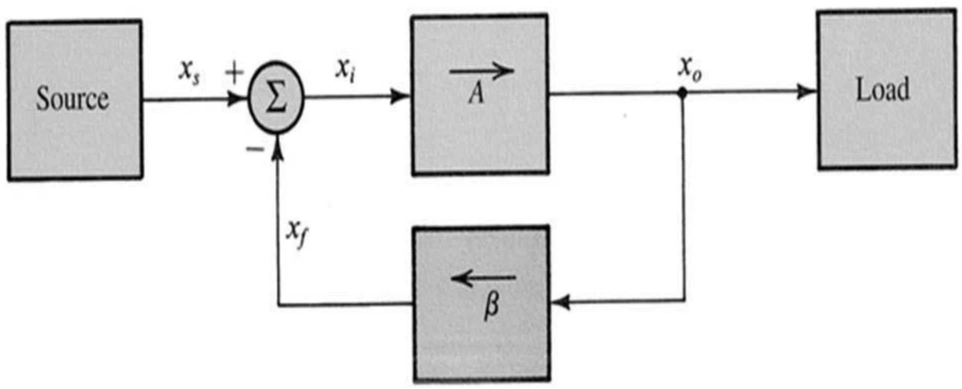
$$\therefore A_d = \frac{A_1 + A_2}{2}$$

$$A_c = A_2 - A_1/2$$

UNIT –III

Feedback Amplifiers : Feedback principle and concept, types of feedback, classification of amplifiers, feedback topologies, Characteristics of negative feedback amplifiers, Generalized analysis of feedback amplifiers, Performance comparison of feedback amplifiers, Method of analysis of feedback amplifiers.

FEEDBACK AMPLIFIER:



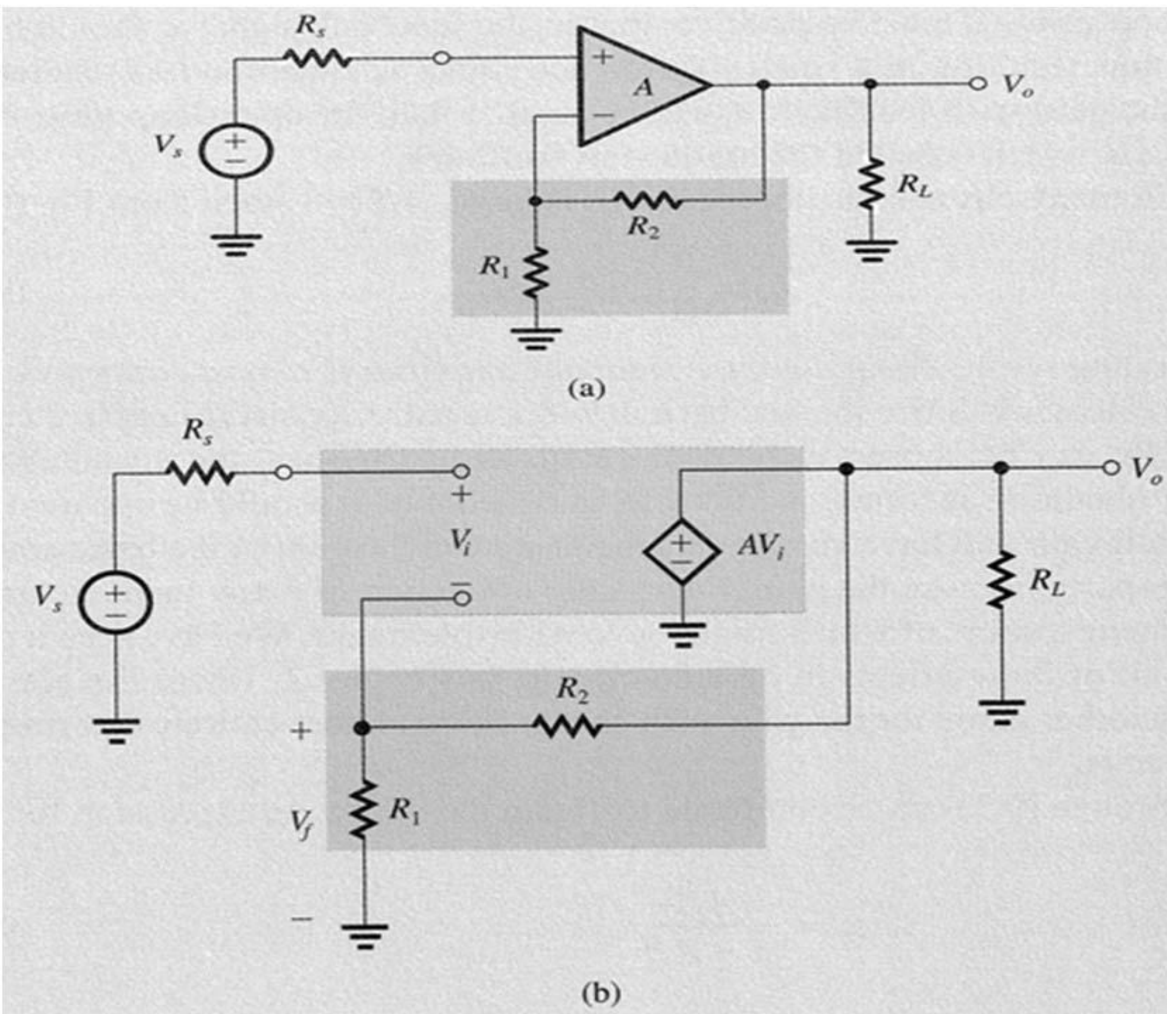
- Signal-flow diagram of a feedback amplifier
- Open-loop gain: A
- Feedback factor:
- Loop gain: A
- Amount of feedback: 1 + A
- Gain of the feedback amplifier (closed-loop gain):

Negative feedback:

- The feedback signal xf is subtracted from the source signal xs
- Negative feedback reduces the signal that appears at the input of the basic amplifier
- The gain of the feedback amplifier Af is smaller than open-loop gain A by a factor of (1+A)
- The loop gain A is typically large (A >>1):
- The gain of the feedback amplifier (closed-loop gain)
- The closed-loop gain is almost entirely determined by the feedback network □ better accuracy of Af.
- xf = xs(A)/(1+A) xs□ error signal xi = xs – xf

For Example, The feedback amplifier is based on an op amp with infinite input resistance and zero output resistance.

- Find an expression for the feedback factor.
- Find the condition under which the closed-loop gain Af is almost entirely determined by the feedback network.
- If the open-loop gain A = 10000 V/V, find R2/R1 to obtain a closed-loop gain Af of 10 V/V.
- What is the amount of feedback in decibel?
- If Vs = 1 V, find Vo, Vf and Vi.
- If A decreases by 20%, what is the corresponding decrease in Af ?



Some Properties of Negative Feedback

Gain de sensitivity:

- The negative reduces the change in the closed-loop gain due to open-loop gain variation

$$dA_f = \frac{dA}{(1+A\beta)^2} \rightarrow \frac{dA_f}{A_f} = \frac{1}{1+A\beta}\frac{dA}{A}$$

- Desensitivity factor: $1+A\beta$

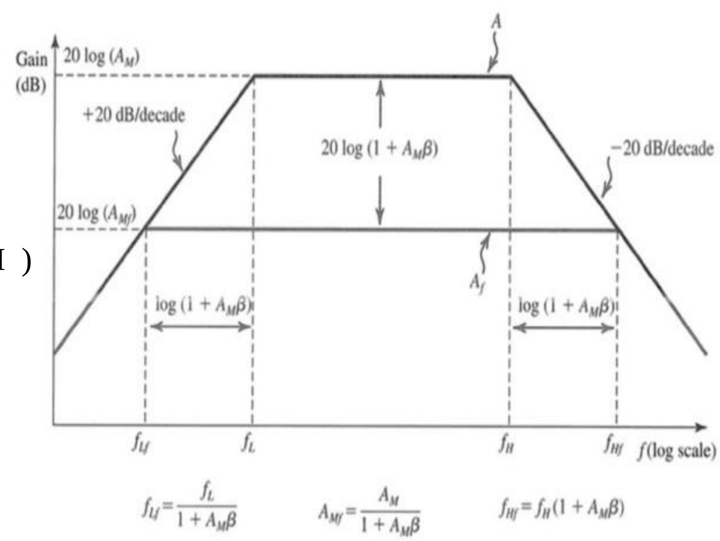
Bandwidth extension

High-frequency response of a single-pole amplifier:

$$A(s) = \frac{A_M}{1+s/\omega_H} \rightarrow A_f(s) = \frac{A_M/(1+A_M\beta)}{1+s/\omega_H(1+A_M\beta)}$$

Low-frequency response of an amplifier with a dominant low-frequency pole:

$$A(s) = \frac{sA_M}{s+\omega_L} \rightarrow A_f(s) = \frac{sA_M/(1+A_M\beta)}{s+\omega_L/(1+A_M\beta)}$$



Negative feedback:

Reduces the gain by a factor of (1+AM)

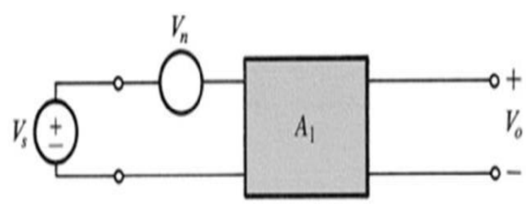
Extends the bandwidth by a factor of (1+AM)

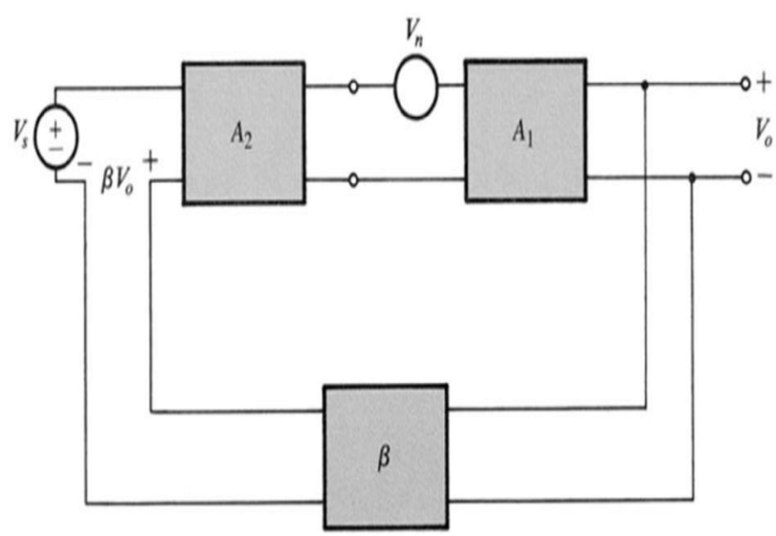
Interference reduction

- ➢ The signal-to-noise ratio:
  - o The amplifier suffers from interference introduced at the input of the amplifier
  - o Signal-to-noise ratio: S/I = Vs/Vn

- ➢ Enhancement of the signal-to-noise ratio:
  - o Precede the original amplifier A1 by a clean amplifier A2
  - o Use negative feedback to keep the overall gain constant.

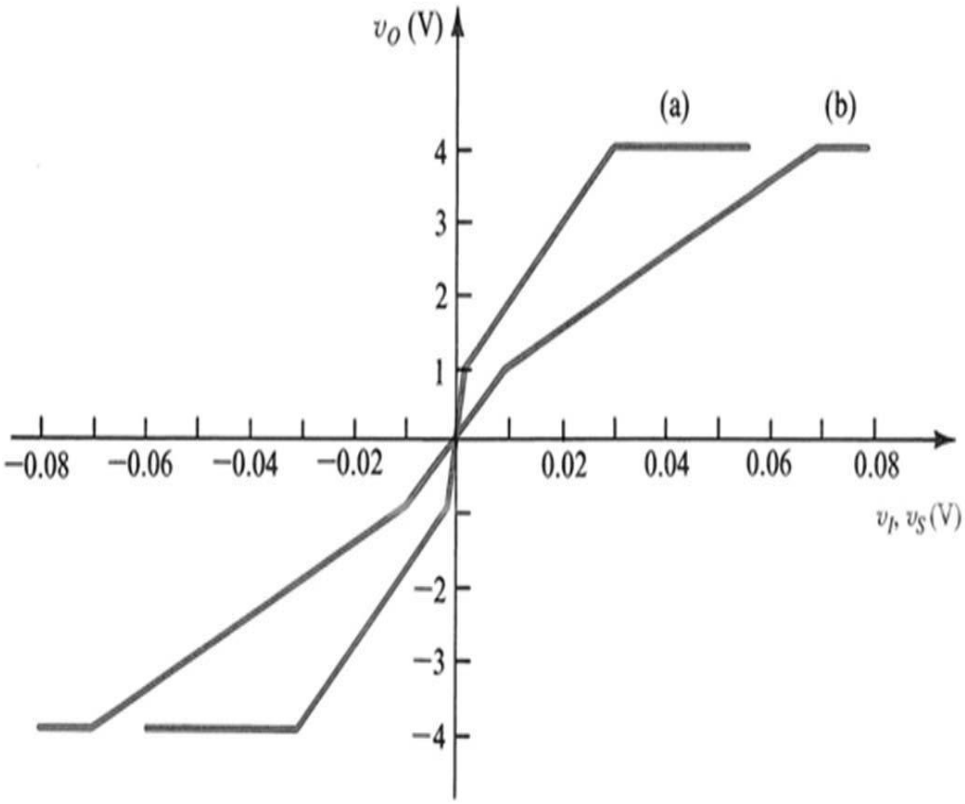
$$V_0 = V_s \frac{A_1 A_2}{1 + A_1 A_2 \beta} + V_n \frac{A_1}{1 + A_1 A_2 \beta} \rightarrow \frac{S}{I} = \frac{V_s}{V_n} A_2$$





Reduction in nonlinear distortion:

The amplifier transfer characteristic is linearised through the application of negative feedback.



□□= 0.01

□*A*□changes from 1000 to 100

$$A_{f1} = \frac{1000}{1+1000\times 0.01} = 90.9$$

$$A_{f2} = \frac{100}{1+100\times 0.01} = 50$$

The Four Basic Feedback Topologies:

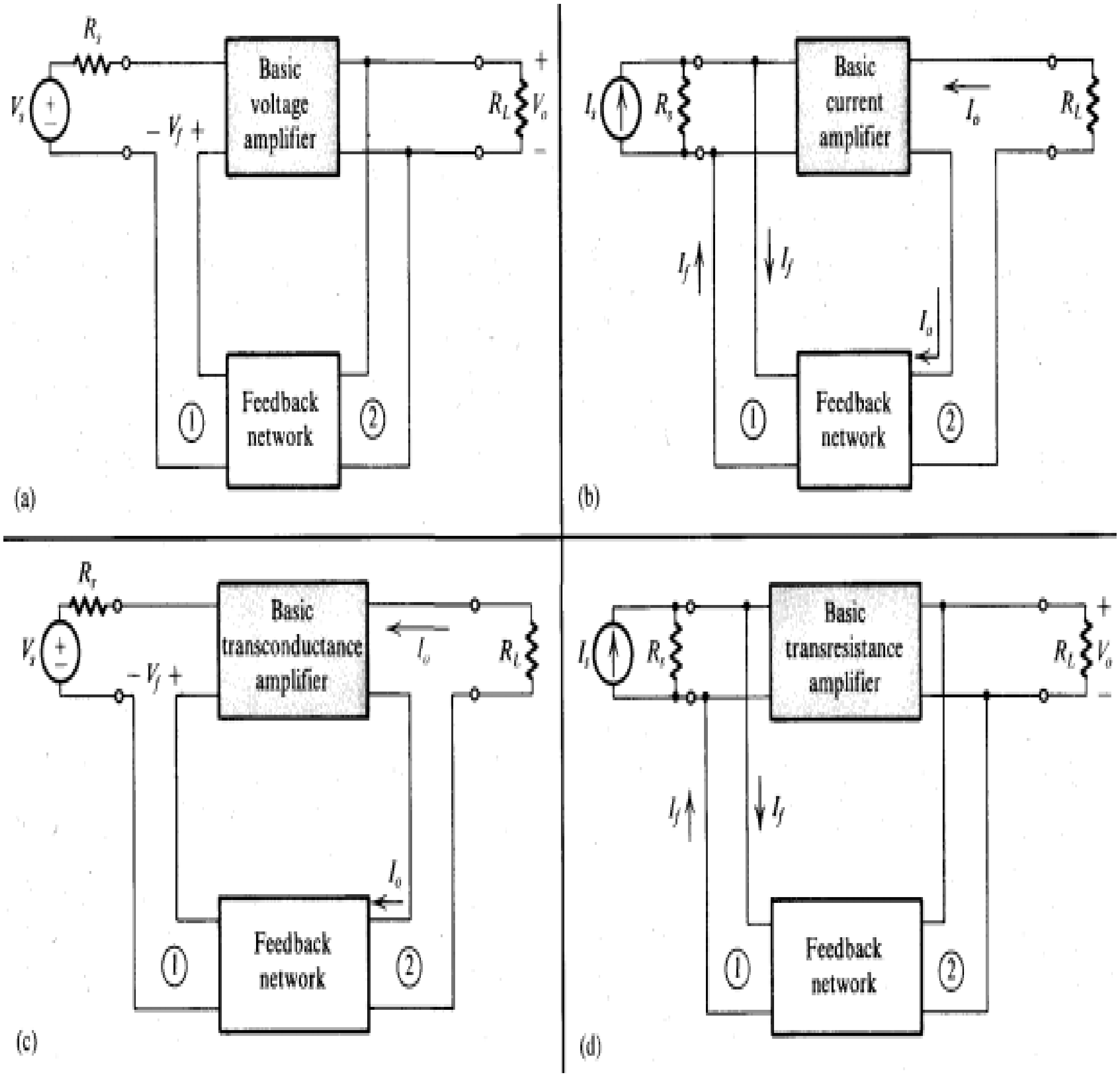
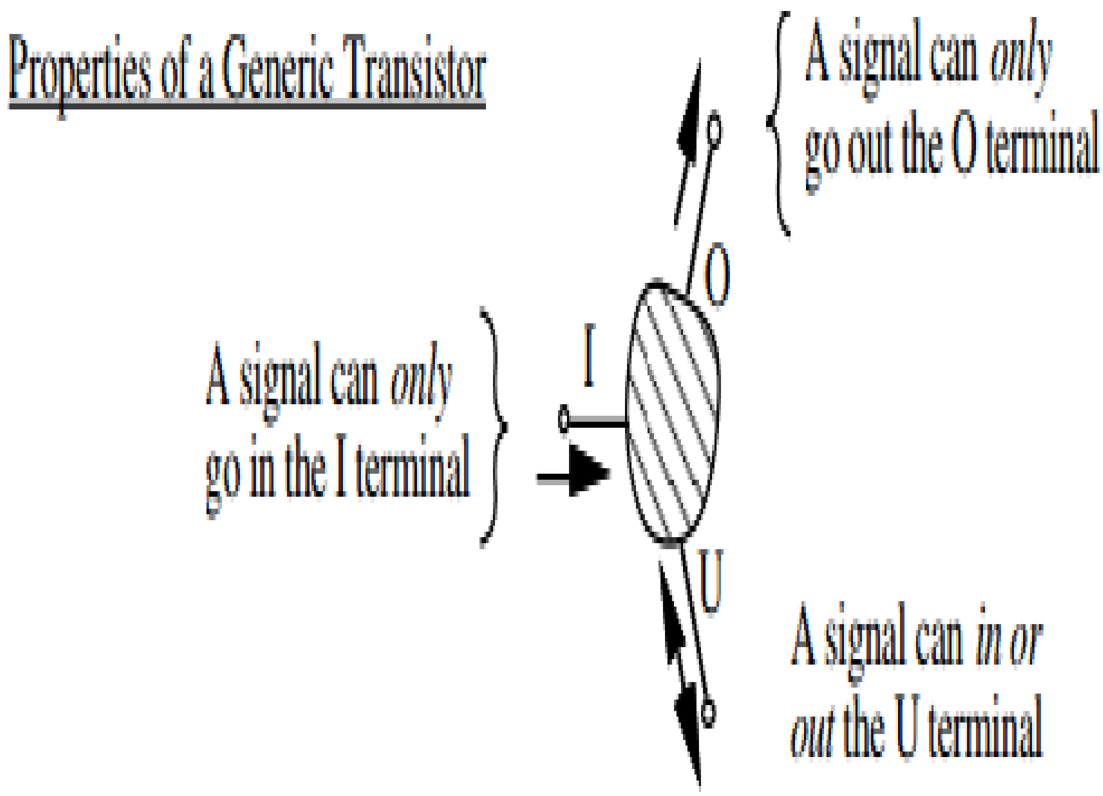


Fig. The four basic feedback topologies: (a) voltage-sampling series-mixing (series-shunt) topology; (b) current-sampling shunt-mixing (shunt-series) topology; (c) current-sampling series-mixing (series-series) topology; (d) voltage-sampling shunt-mixing (shunt-shunt) topology.

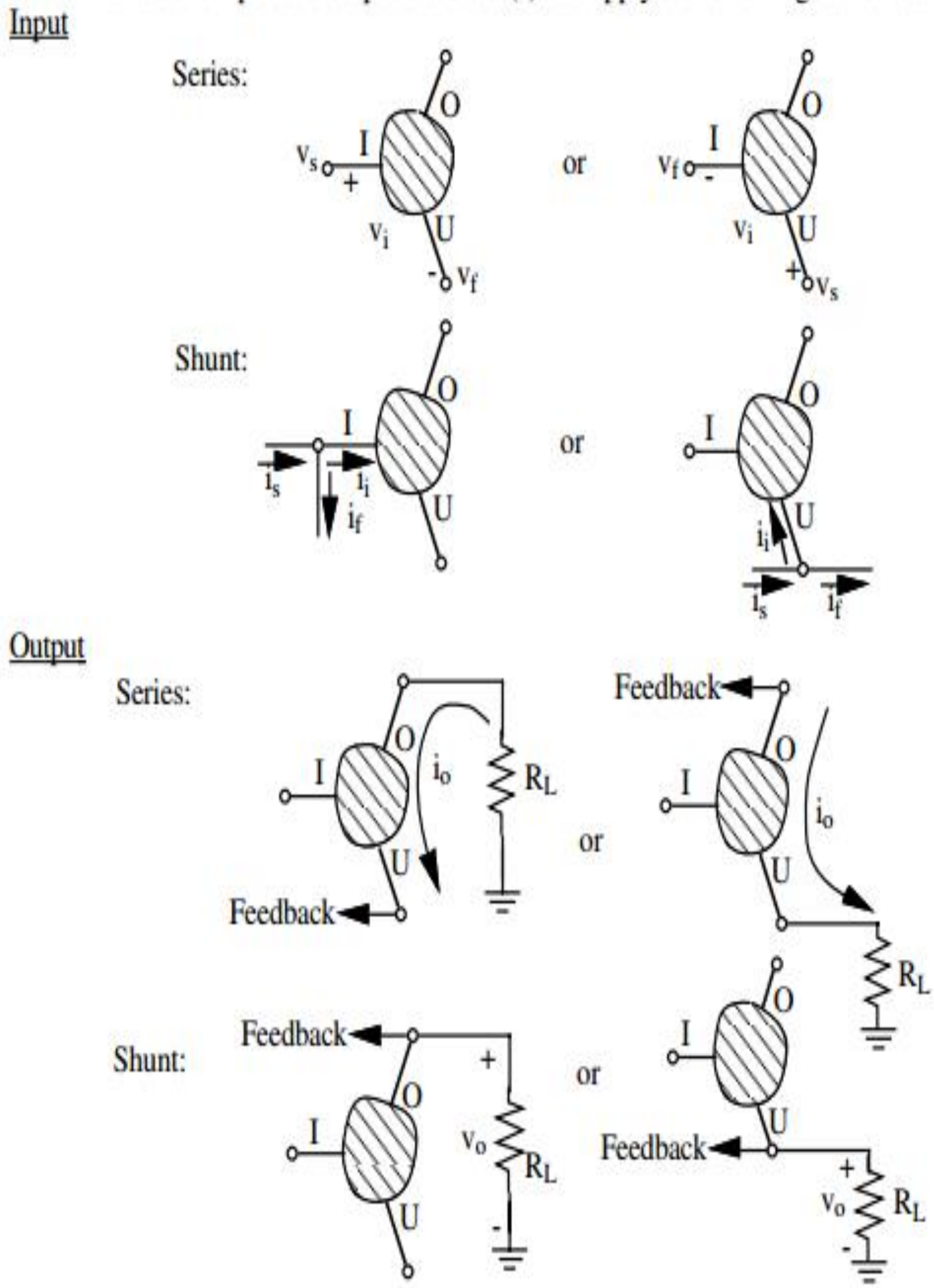
Method of analysis of Feedback Amplifiers:

1. Identify the topology.
2. Determine whether the feedback is positive or negative.
3. Open the loop and calculate A, ß, Ri, and Ro.
4. Use the Table to find Af, Rif and Rof or AF, RiF, and RoF.
5. Use the information in 4 to find whatever is required (v_{out}/v_{in}, R_{in}, R_{out}, etc.)



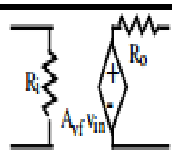
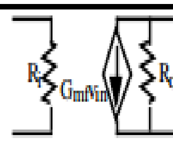
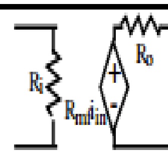
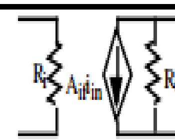
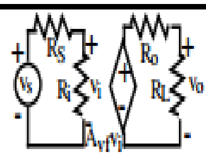
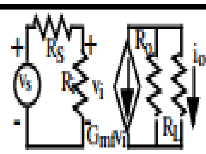
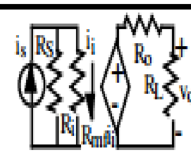
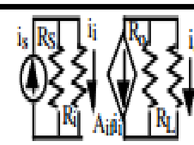
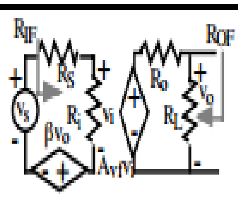
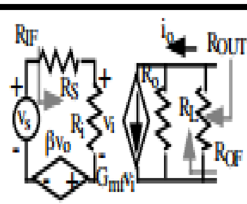
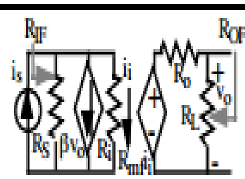
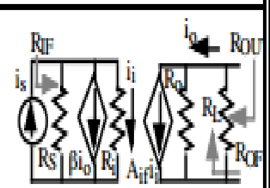
Identification of the Feedback Topology

Isolate the input and output transistor(s) and apply the following identification.



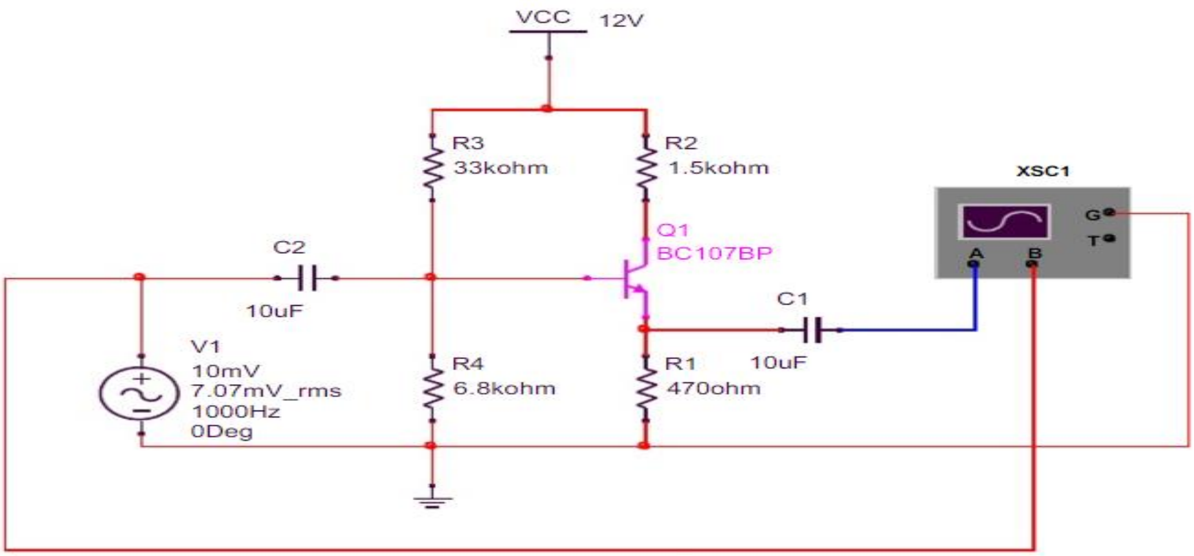
Performance comparison of feedback amplifiers:

Summary of the Important Relationships of Open-loop and Closed-loop Feedback Amplifiers.

Quantity	Voltage Amplifier	Transconductance Amplifier	Transresistance Amplifier	Current Amplifier
Input-output variable	Voltage-voltage	Voltage-current	Current-voltage	Current-current
Small Signal Model				
Small Signal Amplifier with Source & Load				
Ideal R_S	$R_S = 0$ or $R_S << R_i$	$R_S = 0$ or $R_S << R_i$	$R_S = \infty$ or $R_S >> R_i$	$R_S = \infty$ or $R_S >> R_i$
Ideal R_L	$R_L = \infty$ or $R_L >> R_o$	$R_L = 0$ or $R_L << R_o$	$R_L = \infty$ or $R_L >> R_o$	$R_L = 0$ or $R_L << R_o$
Overall Forward Gain	$A_V = \frac{R_i R_L A_{vf}}{(R_S+R_i)(R_L+R_o)}$	$G_M = \frac{R_i R_o G_{mf}}{(R_S+Ri)(R_L+R_o)}$	$R_M = \frac{R_S R_L R_{mf}}{(R_S+Ri)(R_L+R_o)}$	$A_I = \frac{R_S R_o A_{if}}{(R_S+Ri)(R_L+R_o)}$
Feedback Topology	Series-shunt	Series-series	Shunt-shunt	Shunt-series
Ideal ß, finite R_S and R_L Feedback Small Signal Models				
Closed-Loop Gain (Ideal R_S and R_L)	$A_{VF} = \frac{A_{vf}}{(1+A_{vf}\beta_v)}$	$G_{mF} = \frac{G_{mf}}{(1+G_{mf}\beta_g)}$	$R_{mF} = \frac{R_{mf}}{(1+R_{mf}\beta_r)}$	$A_{iF} = \frac{A_{if}}{(1+A_{if}\beta_i)}$

Closed-Loop Input Resistance (Ideal R_S and R_L)	$R_{iF} = R_i(1 + A_{vf}\beta_v)$	$R_{iF} = R_i(1 + G_{mf}\beta_g)$	$R_{iF} = \frac{R_i}{1 + R_{mf}\beta_r}$	$R_{iF} = \frac{R_i}{1 + A_{if}\beta_i}$
Closed-Loop Output Resistance (Ideal R_S and R_L)	$R_{oF} = \frac{R_o}{1 + A_{vf}\beta_v}$	$R_{oF} = R_o(1 + R_{mf}\beta_g)$	$R_{oF} = \frac{R_o}{1 + R_{mf}\beta_r}$	$R_{oF} = R_o(1 + A_{if}\beta_i)$
Closed-Loop Gain	$A_{VF} = \frac{A_V}{(1+A_V\beta_v)}$	$G_{MF} = \frac{G_M}{(1+G_M\beta_g)}$	$R_{MF} = \frac{R_M}{(1+R_M\beta_r)}$	$A_{IF} = \frac{A_I}{(1+A_I\beta_i)}$
Closed-Loop Input Resistance	$R_{IF} = (R_i+R_S)(1+A_V\beta_v)$	$R_{IF} = (R_i+R_S)(1 + G_M\beta_g)$	$R_{IF} = \frac{\frac{R_iR_S}{R_i+R_S}}{1 + R_M\beta_r}$	$R_{IF} = \frac{\frac{R_iR_S}{R_i+R_S}}{1 + A_I\beta_i}$
Closed-Loop Output Resistance	$R_{OF} = \frac{\frac{R_oR_L}{R_o+R_L}}{1 + A_V\beta_v}$	$R_{OF} = (R_o+R_L)(1+G_M\beta_g)$	$R_{OF} = \frac{\frac{R_oR_L}{R_o+R_L}}{1 + R_M\beta_r}$	$R_{OF} = (R_o+R_L)(1+A_I\beta_i)$
Output Resistance of Series Output Fb. Ckt	$R_{OUT} = R_{OF}$	$R_{OUT} = \frac{R_L}{R_{OF}}(R_{OF}-R_L)$	$R_{OUT} = R_{OF}$	$R_{OUT} = \frac{R_L}{R_{OF}}(R_{OF}-R_L)$

In voltage series feedback amplifier, sampling is voltage and series mixing indicates voltage mixing. As both input and output are voltage signals and is said to be voltage amplifier with gain A_{vf}.



Band width is defined as the range frequencies over which gain is greater than or equal to 0.707 times the maximum gain or up to 3 dB down from the maximum gain

Bandwidth (BW) = f_h- f_l

Where f_h= Upper cutoff frequency

And f_l= Lower cutoff frequency.

Cutoff frequency is the frequency at which the gain is 0.707 times the maximum gain or 3dB down from the maximum gain. In all feedback amplifiers we use negative feedback, so gain is reduced and bandwidth is increased

$$\mathbf{A_{vf}= Av/[1+A_v\beta]}$$

$$\mathbf{And\ BW_f.= BW\ [1+ A_v\beta]}$$

Where A_{vf} = Gain with feedback

Av= Gain without feedback

β=feedback gain

BW_f = Bandwidth with feedback and

BW = Bandwidth without feedback Output resistance will decrease due to shunt connection at output and input resistance will increase due to series connection at input.

So $\mathbf{R_{0f} = R_0/[1+A_v\beta]}$ **and**

$$\mathbf{R_{if}=R_i[1+A_v\beta].}$$

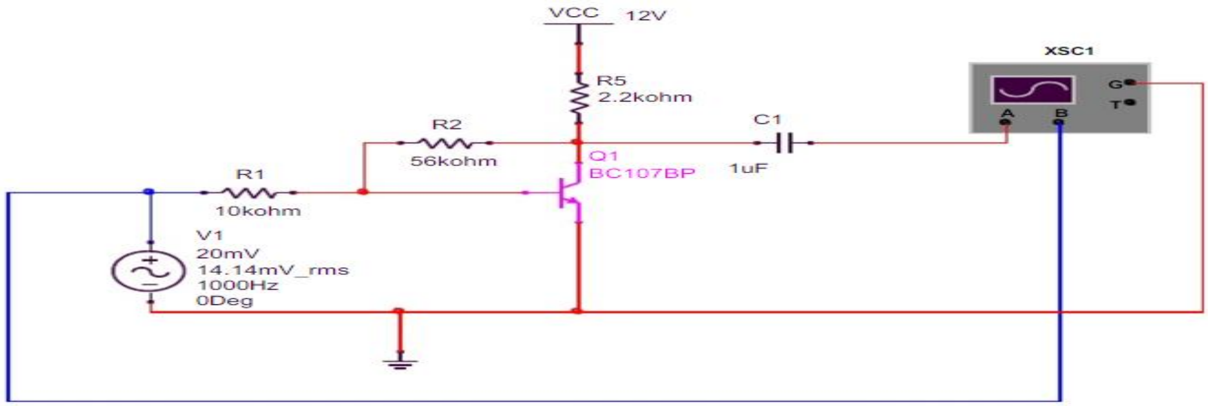
Where R 0f = Output resistance with feedback

R_0= Output resistance without feedback.

R_{if}= Input resistance with feedback

R_i= Input resistance without feedback

In voltage shunt feedback amplifier, sampling is voltage and shunt mixing indicates current mixing. As input is current signal and output is voltage signal, so it is said to be trans-resistance amplifier with gain R_{mf}.



Band width is defined as the range frequencies over which gain is greater than or equal to 0.707 times the maximum gain or up to 3 dB down from the maximum gain.

Bandwidth (BW) = f_h- f_l

Where f_h= Upper cutoff frequency

And f_l= Lower cutoff frequency.

Cutoff frequency is the frequency at which the gain is 0.707 times the maximum gain or 3dB down from the maximum gain. In all feedback amplifiers we use negative feedback, so gain is reduced and bandwidth is increased.

$$R_{mf} = Rm/[1+R_m\beta]$$

$$\text{And } BW_f = BW\,[1+R_m\beta]$$

Where

R_{mf} = Gain with feedback

R_m= Gain without feedback

β=feedback gain

BW = Bandwidth without feedback

Output resistance and input resistance both will decrease due to shunt connections at input and output. So

$$R_{0f} = R_0/[1+R_m\,\beta] \text{ and}$$

$$R_{if} = R_i/[1+R_m\,\beta].$$

Where R 0f = Output resistance with feedback

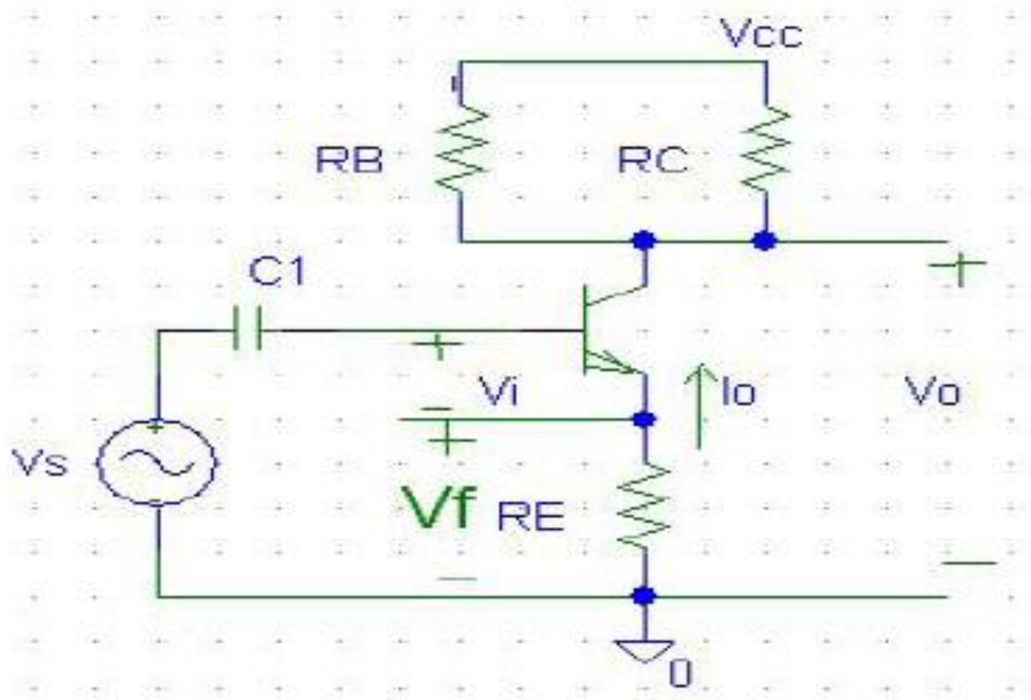
R_0= Output resistance without feedback.

R_{if}= Input resistance with feedback

R_i= Input resistance without feedback

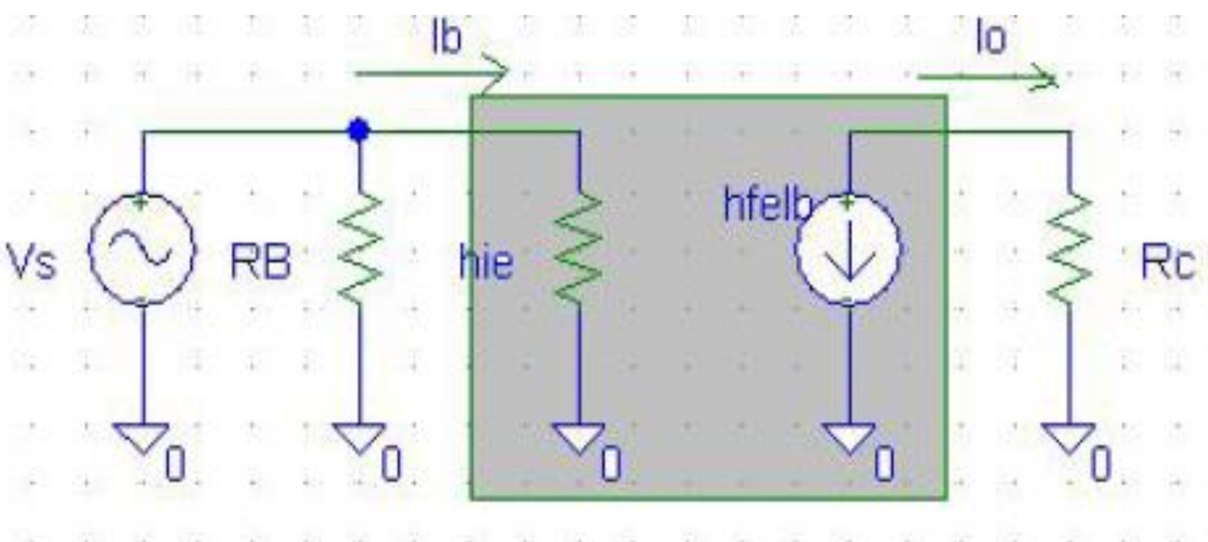
Current series feedback

- Feedback technique is to sample the output current (Io) and return a proportional voltage in series.
- It stabilizes the amplifier gain, the current series feedback connection increases the input resistance.
- In this circuit, emitter of this stage has an un bypassed emitter, it effectively has current-series feedback.
- The current through RE results in feedback voltage that opposes the source signal applied so that the output voltage Vo is reduced.



• To remove the current-series feedback, the emitter resistor must be either removed or bypassed by a capacitor (as is done in most of the amplifiers)

The fig below shows the equivalent circuit for current series feedback



Gain, input and output impedance for this condition is,

$$A_f = \frac{I_o}{V_s} = \frac{A}{1+A\beta} = \frac{-h_{fe}/h_{ie}}{1+(-R_E)\left(\frac{-h_{fe}}{h_{ie}+R_E}\right)}$$

$$Z_{if} = Z_i(1+A\beta) \cong h_{ie}\left(1+\frac{h_{fe}R_E}{h_{ie}}\right)$$

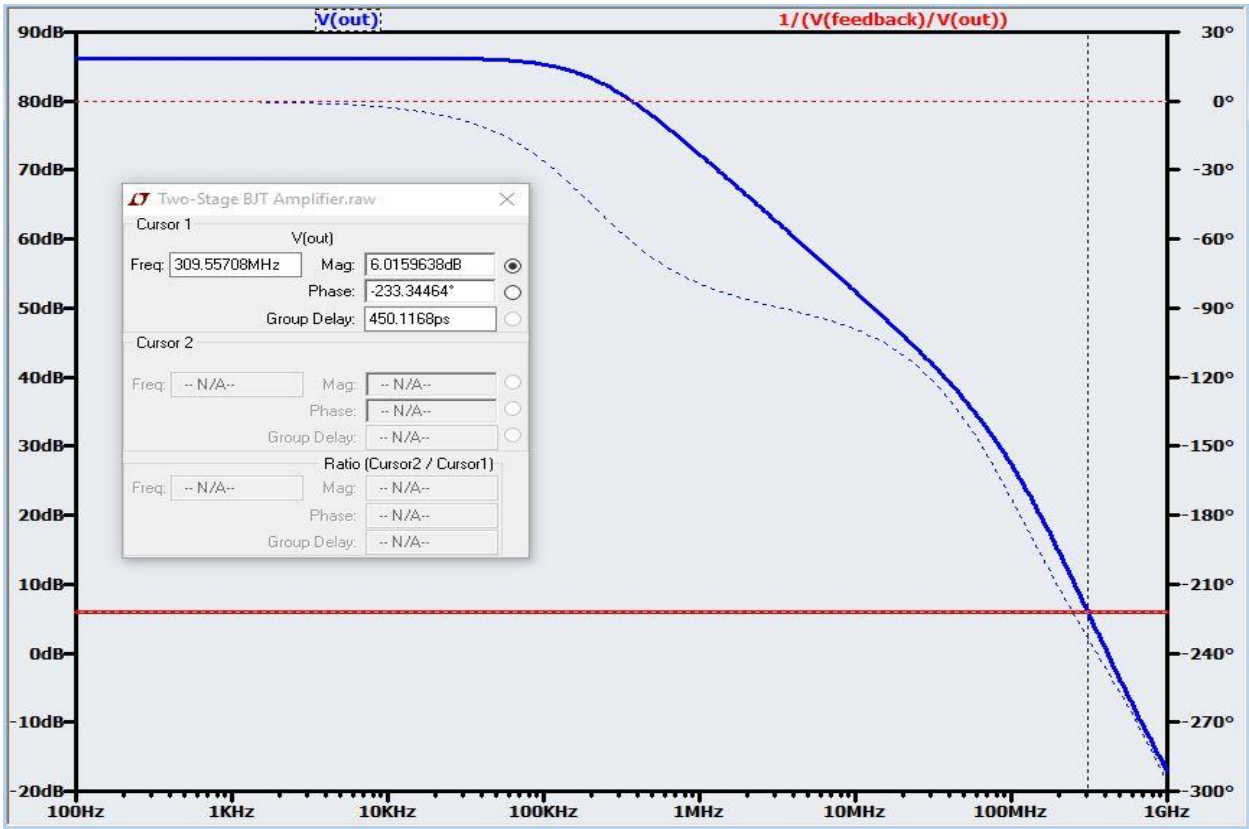
$$Z_{of} = Z_o(1+A\beta) \cong R_c\left(1+\frac{h_{fe}R_E}{h_{ie}}\right)$$

$$with-feedback..A;$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{I_oR_c}{V_s} = \left(\frac{I_o}{V_s}\right)R_c = A_fR_C \cong \frac{-h_{fe}R_C}{h_{ie}+h_{fe}R_E}$$

We now know that by plotting the gain and phase shift of a negative feedback amplifier's loop gain—denoted by $A\beta$, where A is always a function of frequency and β can be considered a function of frequency if necessary—we can determine two things: 1) whether the amplifier is stable, and 2) whether the amplifier is *sufficiently* stable (rather than *marginally* stable). The first determination is based on the stability criterion, which states that the magnitude of the loop gain must be less than unity at the frequency where the phase shift of the loop gain is 180°. The second is based on the amount of gain margin or phase margin; a rule of thumb is that the phase margin should be at least 45°.

It turns out that we can effectively analyze stability using an alternative and somewhat simplified approach in which open-loop gain A and feedback factor β are depicted as separate curves on the same axes. Consider the following plot for the discrete BJT amplifier with a frequency-independent (i.e., resistor-only) feedback network configured for $\beta = 0.5$:



Here you see V(out), which corresponds to the open-loop gain, and 1/(V(feedback)/V(out)). If you recall that *β* is the percentage (expressed as a decimal) of the output fed back and subtracted from the input, you will surely recognize that this second trace is simply 1/*β*. So why did we plot 1/*β*? Well, we know that loop gain is *A* multiplied by *β*, but in this plot the y-axis is in decibels and is thus logarithmic. Our high school math teachers taught us that multiplication of ordinary numbers corresponds to addition with logarithmic values, and likewise numerical division corresponds to logarithmic subtraction. Thus, a logarithmic plot of *A* multiplied by *β* can be represented as the logarithmic plot of *A* **plus** the logarithmic plot of *β*. Remember, though, that the above plot includes not *β* but rather 1/*β*, which is the equivalent of **negative** *β* on a logarithmic scale. Let's use some numbers to clarify this:

$$\beta=0.5 \Rightarrow 20\log(\beta)\approx -6\text{ dB}$$

$$\frac{1}{\beta}=2 \Rightarrow 20\log\left(\frac{1}{\beta}\right)\approx 6\text{ dB}$$

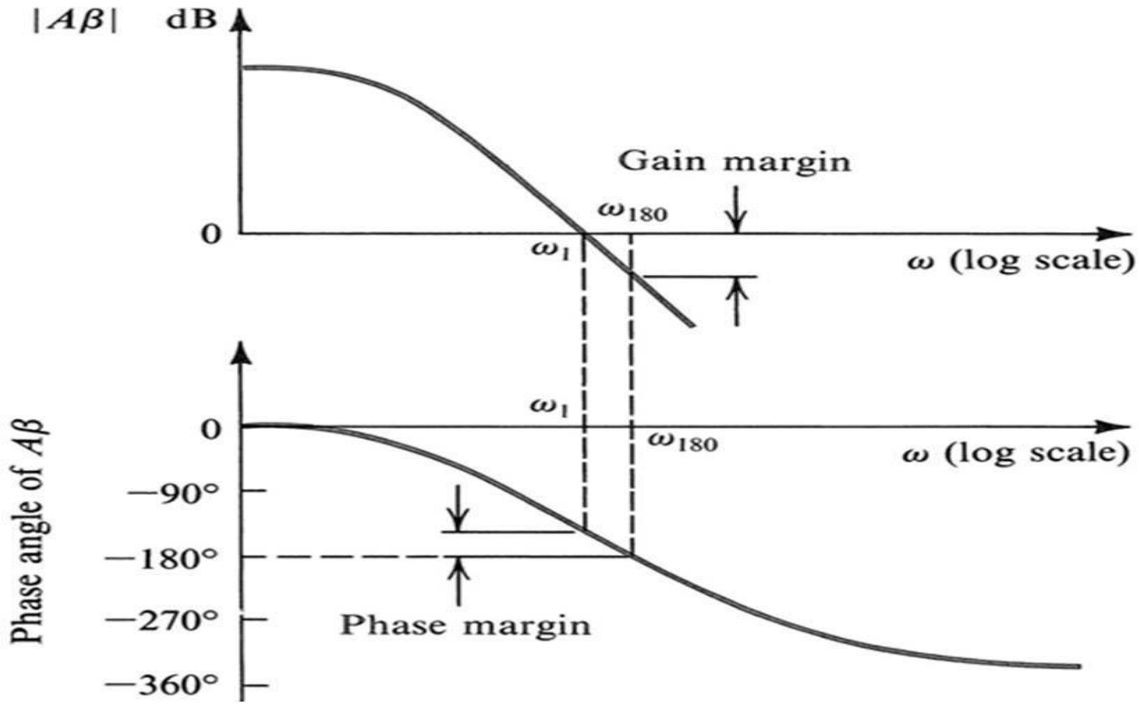
Thus, in this logarithmic plot, we have 20log(A) and -20log(β), which means that to reconstruct 20log($A\beta$) we need to **subtract the 1/β curve from the A curve**:

20log(Aβ)=20log(A)+20log(β) ⇒ 20log(Aβ)=20log(A)−(−20log(β))20log(Aβ)=20log(A)+20 log(β) ⇒ 20log(Aβ)=20log(A)−(−20log(β))

⇒ 20log(Aβ)=20log(A)−20log(1β).

Gain and phase margin

- The stability of a feedback amplifier is determined by examining its loop gain as a function of frequency.
- One of the simplest means is through the use of Bode plot for $A\beta$.
- Stability is ensured if the magnitude of the loop gain is less than unity at a frequency shift of 180 .
- Gain margin:
  - The difference between the value $|A\beta|$ of at $_{180}$ and unity.
  - Gain margin represents the amount by which the loop gain can be increased while maintaining stability.
- Phase margin:
  - A feedback amplifier is stable if the phase is less than 180 at a frequency for which $|A\beta|$ =1.
  - A feedback amplifier is unstable if the phase is in excess of 180 at a frequency for which $|A\beta|$ =1.
  - The difference between the a frequency for which $|A\beta|$ =1 and 180 .



Effect of phase margin on closed-loop response:

- Consider a feedback amplifier with a large low-frequency loop gain ($A_0 >> 1$).
- The closed-loop gain at low frequencies is approximately $1/\beta$.
- Denoting the frequency at which | $A\beta$ | =1 by ω_1: $A(j\omega_1)\beta = 1\times e^{-j\theta}$ and $\theta = 180^\circ -$ phase margin
- The closed-loop gain at $_1$ peaks by a factor of 1.3 above the low-frequency gain for a phase margin of 45 .

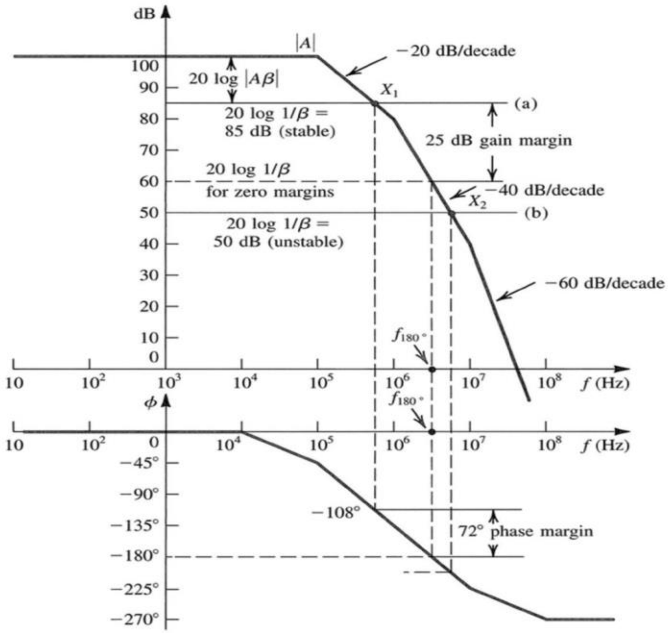
$$A_f(j\omega_1) = \frac{A(j\omega_1)}{1+A(j\omega_1)\beta} = \frac{1}{\beta}\frac{e^{-j\theta}}{1+e^{-j\theta}}$$

$$|A_f(j\omega_1)| = \frac{1/\beta}{|1+e^{-j\theta}|}$$

- This peaking increase as the phase margin is reduced, eventually reaching infinite when the phase margin is zero (sustained oscillations).

An alternative approach for investigating stability

- In a Bode plot, the difference between 20 log$|A(j\omega)|$ and 20 log$(1/\beta)$ is 20 log$|A\beta|$.



$$A = \frac{10^5}{(1 + jf/10^5)(1 + jf/10^6)(1 + jf/10^7)}$$

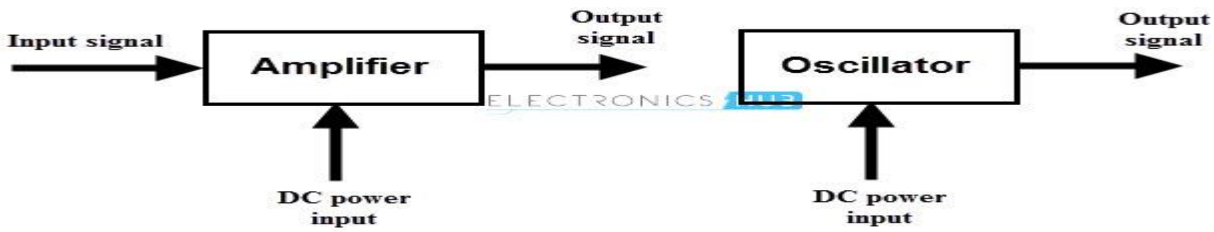
$$\phi = -\left[\tan^{-1}(f/10^5) + \tan^{-1}(f/10^6) + \tan^{-1}(f/10^7)\right]$$

Unit-IV

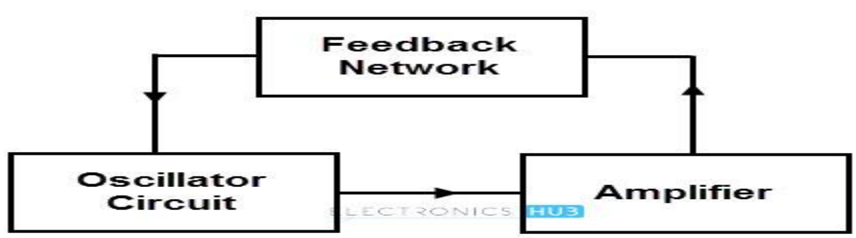
Oscillators: Oscillator principle, condition for oscillations, types of oscillators, RC-phase shift and Wein bridge oscillators with BJT and FET and their analysis, Generalized analysis of LC Oscillators, Hartley and Colpitt's oscillators with BJT and FET and their analysis, Frequency and amplitude stability of oscillators.

Oscillators

An electronic circuit used to generate the output signal with constant amplitude and constant desired frequency is called as an oscillator. It is also called as a waveform generator which incorporates both active and passive elements. The primary function of an oscillator is to convert DC power into a periodic signal or AC signal at a very high frequency. An oscillator does not require any external input signal to produce sinusoidal or other repetitive waveforms of desired magnitude and frequency at the output and even without use of any mechanical moving parts.



In case of amplifiers, the energy conversion starts as long as the input signal is present at the input, i.e., amplifier produces an output signal whose frequency or waveform is similar to the input signal but magnitude or power level is generally high. The output signal will be absent if there is no input signal at the input.In contrast, to start or maintain the conversion process an oscillator does not require any input signal as shown figure. As long as the DC power is connected to the oscillator circuit, it keeps on producing an output signal with frequency decided by components in it.

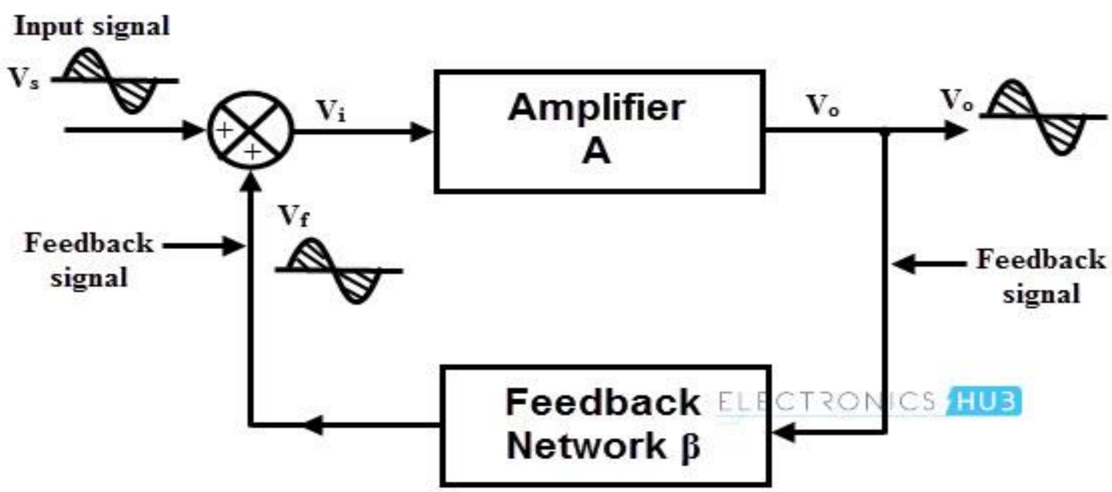


The above figure shows the block diagram of an oscillator. An oscillator circuit uses a vacuum tube or a transistor to generate an AC output. The output oscillations are produced by the tank circuit components either as R and C or L and C. For continuously generating output without the requirement of any input from preceding stage, a feedback circuit is used.

From the above block diagram, oscillator circuit produces oscillations that are further amplified by the amplifier. A feedback network gets a portion of the amplifier output and feeds it the oscillator circuit in correct phase and magnitude. Therefore, un damped electrical oscillations are produced , by continuously supplying losses that occur in the tank circuit.

Oscillators Theory

The main statement of the oscillator is that the oscillation is achieved through positive feedback which generates the output signal without input signal. Also, the voltage gain of the amplifier increases with the increase in the amount of positive feedback. In order to understand this concept, let us consider a non-inverting amplifier with a voltage gain 'A' and a positive feedback network with feedback gain of β as shown in figure.



Let us assume that a sinusoidal input signal Vs is applied at the input. Since the amplifier is non-inverting, the output signal Vo is in phase with Vs. A feedback network feeds the part of Vo to the input and the amount Vo fed back depends on the feedback network gain β. No phase shift is introduced by this feedback network and hence the feedback voltage or signal Vf is in phase with Vs. A feedback is said to be positive when the phase of the feedback signal is same as that of the input signal. The open loop gain 'A' of the amplifier is the ratio of output voltage to the input voltage, i.e.,

$$A = Vo/Vi$$

By considering the effect of feedback, the ratio of net output voltage Vo and input supply Vs called as a closed loop gain Af (gain with feedback).

$$Af = Vo/Vs$$

Since the feedback is positive, the input to the amplifier is generated by adding Vf to the Vs,

$$Vi = Vs + Vf$$

Depends on the feedback gain β, the value of the feedback voltage is varied, i.e.,

$$Vf = \beta\ Vo$$

Substituting in the above equation,

$$Vi = Vs + \beta\ Vo$$

$$Vs = Vi - \beta\ Vo$$

Then the gain becomes

$$Af = Vo/\ (Vi - \beta\ Vo)$$

By dividing both numerator and denominator by Vi, we get

$$Af = (Vo\ /\ Vi)/\ (1 - \beta)\ (Vo\ /\ Vi)$$

$$Af = A/\ (1- A\ \beta)\ \text{since } A = Vo/Vi$$

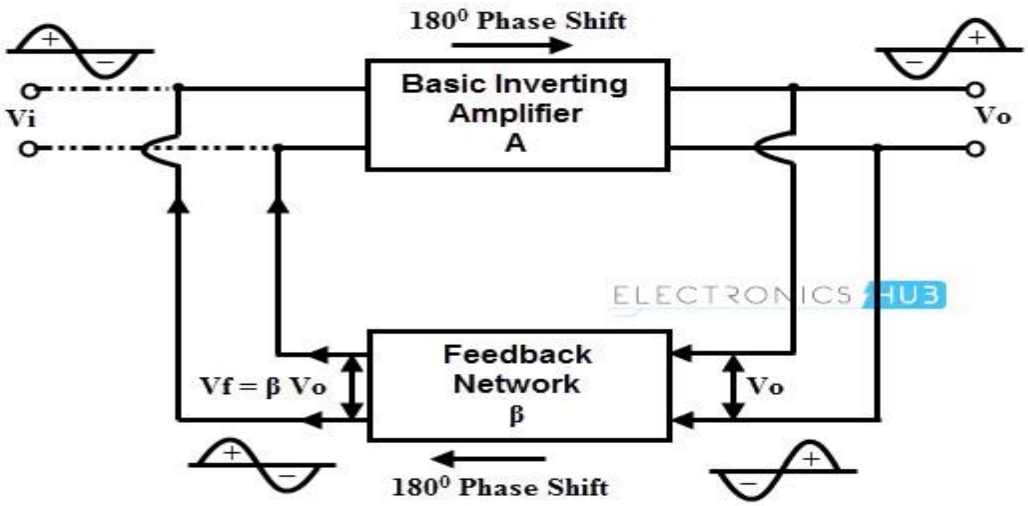
Where Aβ is the loop gain and if Aβ = 1, then Af becomes infinity. From the above expression, it is clear that even without external input (Vs = 0), the circuit can generate the output just by feeding a part of the output as its own input. And also closed loop gain increases with increase in amount of positive feedback gain. The oscillation rate or frequency depends on amplifier or feedback network or both.

Barkhausen Criterion or Conditions for Oscillation

The circuit will oscillate when two conditions, called as Barkhausen's criteria are met. These two conditions are

1. **The loop gain must be unity or greater**
2. **The feedback signal feeding back at the input must be phase shifted by 360 degrees** (which is same as zero degrees). In most of the circuits, an inverting amplifier is used to produce 180 degrees phase shift and additional 180 degrees phase shift is provided by the feedback network. At only one particular frequency, a tuned inductor-capacitor (LC circuit) circuit provides this 180 degrees phase shift.

Let us know how these conditions can be achieved.



Consider the same circuit which we have taken in oscillator theory. The amplifier is a basic inverting amplifier and it produces a phase shift of 180 degrees between input and output.
The input to be applied to the amplifier is derived from the output Vo by the feedback network. Since the output is out of phase with Vi. So the feedback network must ensure a phase shift of 180 degrees while feeding the output to the input. This is nothing but ensuring positive feedback.

Let us consider that a fictitious voltage, Vi is applied at the input of amplifier, then

$$Vo = A\ Vi$$

The amount of feedback voltage is decided by the feedback network gain, then

$$Vf = -\beta\ Vo$$

This negative sign indicates 180 degrees phase shift.
Substituting Vo in above equation, we get

$$Vf = -A\ \beta\ Vi$$

In oscillator, the feedback output must drive the amplifier, hence Vf must act as Vi. For achieving this term – A β in the above expression should be 1, i.e.,

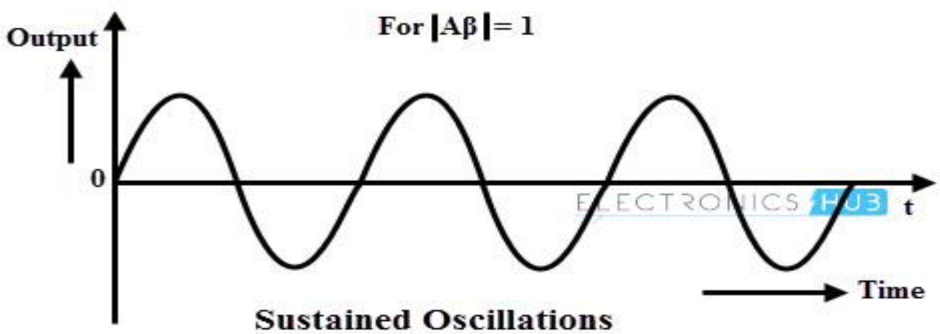
$$Vf = Vs \text{ when } -A\ \beta = 1.$$

This condition is called as Barkhausen criterion for oscillation.
Therefore, A β = -1 + j0. This means that the magnitude of A β (modulus of A β) is equal to 1. In addition to the magnitude, the phase of the Vs must be same as Vi. In order to perform this, feedback network should introduce a phase shift of 180 degrees in addition to phase shift (180 degrees) introduced by the amplifier.

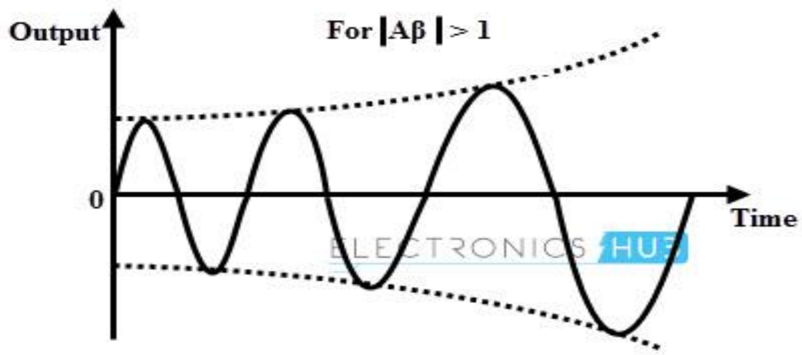
So the total phase shift around the loop is 360 degrees. Thus, under these conditions the oscillator can oscillate or produce the waveform without applying any input (that's why we have considered as fictitious voltage). It is important to know that how the oscillator starts to oscillate even without input signal in practice? The oscillator starts generating oscillations by amplifying the noise voltage which is always present. This noise voltage is result of the movement of free electrons under the influence of room temperature. This noise voltage is not exactly in sinusoidal due to saturation conditions of practical circuit. However, this nose signal will be sinusoidal when A β value is close to one. In practice modulus of A β is made greater than 1 initially, to amplify the small noise voltage. Later the circuit itself adjust to get modulus of A β is equal to one and with a phase shift of 360 degrees.

Nature of Oscillations

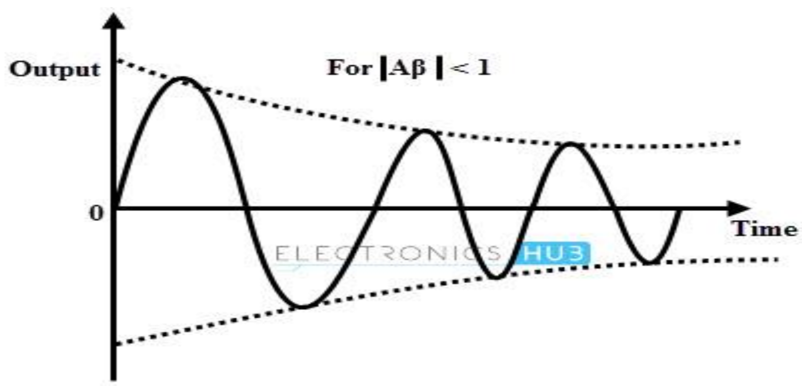
Sustained Oscillations: Sustained oscillations are nothing but oscillations which oscillate with constant amplitude and frequency. Based on the Barkhausen criterion sustained oscillations are produced when the magnitude of loop gain or modulus of A β is equal to one and total phase shift around the loop is 0 degrees or 360 ensuring positive feedback.



Growing Type of Oscillations: If modulus of A β or the magnitude of loop gain is greater than unity and total phase shift around the loop is 0 or 360 degrees, then the oscillations produced by the oscillator are of growing type. The below figure shows the oscillator output with increasing amplitude of oscillations.



Exponentially Decaying Oscillations: If modulus of A β or the magnitude of loop gain is less than unity and total phase shift around the loop is 0 or 360 degrees, then the amplitude of the oscillations decreases exponentially and finally these oscillations will cease.



Classification of oscillators

The oscillators are classified into several types based on various factors like nature of waveform, range of frequency, the parameters used, etc. The following is a broad classification of oscillators.

According to the Waveform Generated

Based on the output waveform, oscillators are classified as sinusoidal oscillators and non-sinusoidal oscillators.

Sinusoidal Oscillators: This type of oscillator generates sinusoidal current or voltages.

Non-sinusoidal Oscillators: This type of oscillators generates output, which has triangular, square, rectangle, saw tooth waveform or is of pulse shape.

According to the Circuit Components: Depends on the usage of components in the circuit, oscillators are classified into LC, RC and crystal oscillators. The oscillator using inductor and capacitor components is called as LC oscillator while the oscillator using resistance and capacitor components is called as RC oscillators. Also, crystal is used in some oscillators which are called as crystal oscillators.

According to the Frequency Generated: Oscillators can be used to produce the waveforms at frequencies ranging from low to very high levels. Low frequency or audio frequency oscillators are used to generate the oscillations at a range of 20 Hz to 100-200 KHz which is an audio frequency range.

High frequency or radio frequency oscillators are used at the frequencies more than 200-300 KHz up to gigahertz. LC oscillators are used at high frequency range, whereas RC oscillators are used at low frequency range.

Based on the Usage of Feedback

The oscillators consisting of feedback network to satisfy the required conditions of the oscillations are called as feedback oscillators. Whereas the oscillators with absence of feedback network are called as non-feedback type of oscillators. The UJT relaxation oscillator is the example of non-feedback oscillator which uses a negative resistance region of the characteristics of the device.

Some of the sinusoidal oscillators under above categories are

- Tuned-circuits or LC feedback oscillators such as Hartley, Colpitts and Clapp etc.
- RC phase-shift oscillators such as Wein-bridge oscillator.
- Negative-resistance oscillators such as tunnel diode oscillator.
- Crystal oscillators such as Pierce oscillator.
- Heterodyne or beat-frequency oscillator (BFO).

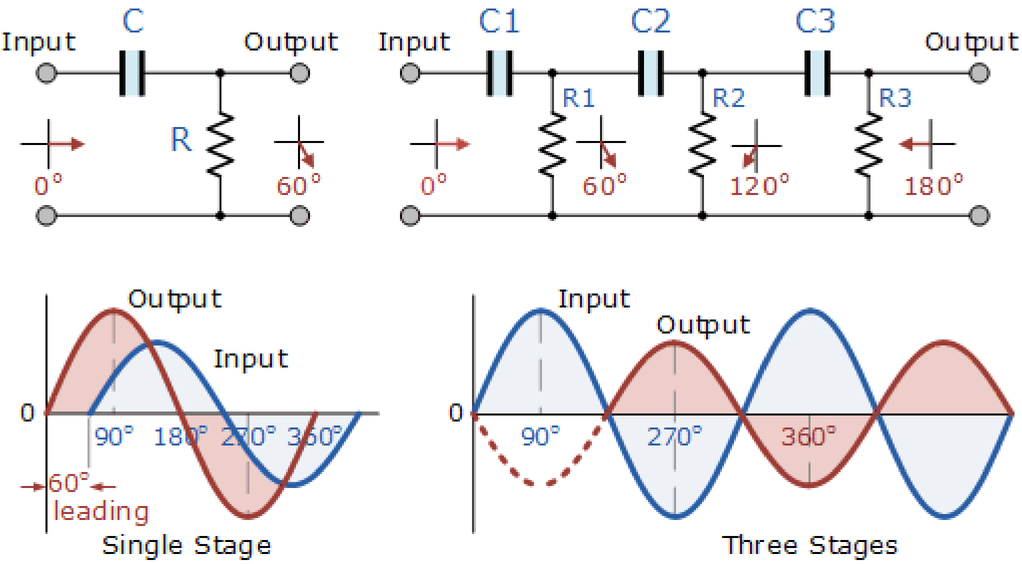
A single stage amplifier will produce 180^o of phase shift between its output and input signals when connected in a class-A type configuration.

For an oscillator to sustain oscillations indefinitely, sufficient feedback of the correct phase, that is “Positive Feedback” must be provided along with the transistor amplifier being used acting as an inverting stage to achieve this.

In an **RC Oscillator** circuit the input is shifted 180^o through the amplifier stage and 180^oagain through a second inverting stage giving us “$180^o + 180^o = 360^o$” of phase shift which is effectively the same as 0^o thereby giving us the required positive feedback. In other words, the phase shift of the feedback loop should be “0”.

In a **Resistance-Capacitance Oscillator** or simply an **RC Oscillator**, we make use of the fact that a phase shift occurs between the input to a RC network and the output from the same network by using RC elements in the feedback branch, for example.

RC Phase-Shift Network



The circuit on the left shows a single resistor-capacitor network whose output voltage "leads" the input voltage by some angle less than 90°. An ideal single-pole RC circuit would produce a phase shift of exactly 90°, and because 180° of phase shift is required for oscillation, at least two single-poles must be used in an RC oscillator design.

However in reality it is difficult to obtain exactly 90° of phase shift so more stages are used. The amount of actual phase shift in the circuit depends upon the values of the resistor and the capacitor, and the chosen frequency of oscillations with the phase angle (Φ) being given as:

RC Phase Angle

$$X_C = \frac{1}{2\pi f C} \qquad R = R,$$

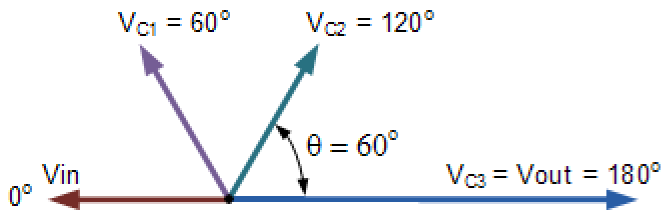
$$Z = \sqrt{R^2 + (X_C)^2}$$

$$\therefore \quad \phi = \tan^{-1}\frac{X_C}{R}$$

Where: X_C is the Capacitive Reactance of the capacitor, R is the Resistance of the resistor, and *f* is the Frequency.

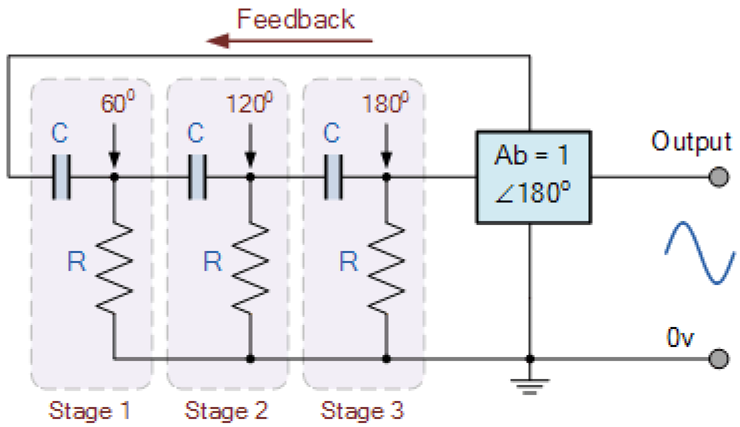
In our simple example above, the values of R and C have been chosen so that at the required frequency the output voltage leads the input voltage by an angle of about 60°. Then the phase angle between each successive RC section increases by another 60°giving a phase difference between the input and output of 180° (3 x 60°) as shown by the following vector diagram.

Vector Diagram



Then by connecting together three such RC networks in series we can produce a total phase shift in the circuit of 180° at the chosen frequency and this forms the bases of a "phase shift oscillator" otherwise known as a **RC Oscillator** circuit.

We know that in an amplifier circuit either using a Bipolar Transistor or an Operational Amplifier, it will produce a phase-shift of 180° between its input and output. If a three-stage RC phase-shift network is connected between this input and output of the amplifier, the total phase shift necessary for regenerative feedback will become 3 x 60°+ 180° = 360° as shown.



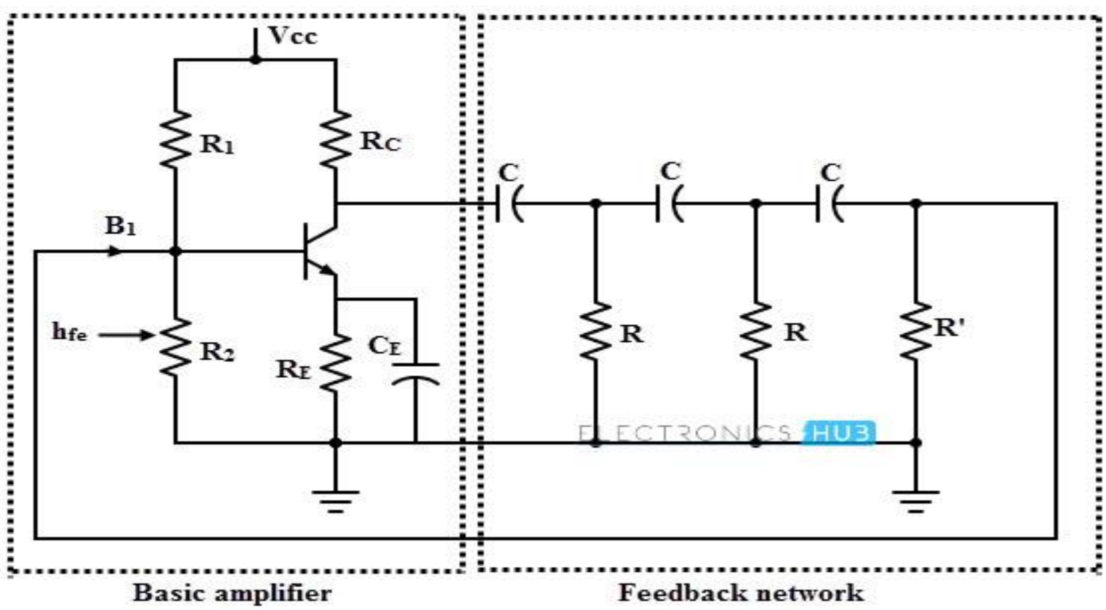
The three RC stages are cascaded together to get the required slope for a stable oscillation frequency. The feedback loop phase shift is -180° when the phase shift of each stage is -60°. This occurs when $\omega = 2\pi f = 1.732/RC$ as ($\tan 60° = 1.732$). Then to achieve the required phase shift in an RC oscillator circuit is to use multiple RC phase-shifting networks such as the circuit below.

RC Phase Shift Oscillator Using BJT

In this transistorized oscillator, a transistor is used as active element of the amplifier stage. The figure below shows the RC oscillator circuit with transistor as active element. The DC operating point in active region of the transistor is established by the resistors R1, R2, RC and RE and the supply voltage Vcc.

The capacitor CE is a bypass capacitor. The three RC sections are taken to be identical and the resistance in the last section is R' = R – hie. The input resistance hie of the transistor is added to R', thus the net resistance given by the circuit is R.

The biasing resistors R1 and R2 are larger and hence no effect on AC operation of the circuit. Also due to negligible impedance offered by the RE – CE combination, it is also no effect on AC operation.



When the power is given to the circuit, noise voltage (which is generated by the electrical components) starts the oscillations in the circuit. A small base current at the transistor amplifier produces a current which is phase shifted by 180 degrees. When this signal is feedback to the input of the amplifier, it will be again phase shifted by 180 degrees. If the loop gain is equal to unity then sustained oscillations will be produced.

By simplifying the circuit with equivalent AC circuit, we get the frequency of oscillations,

$$\mathbf{f = 1/ (2 \pi R C \sqrt{((4Rc / R) + 6)})}$$

If Rc/R << 1, then

$$f = 1/(2 \pi R C \sqrt{6})$$

The condition of sustained oscillations,

$$h_{fe}(min) = (4 R_c/R) + 23 + (29 R/R_c)$$

For a phase shift oscillator with R = Rc, hfe should be 56 for sustained oscillations.

From the above equations it is clear that, for changing the frequency of oscillations, R and C values have to be changed.

But for satisfying oscillating conditions, these values of the three sections must be changed simultaneously. So this is not possible in practice, therefore a phase shift oscillator is used as a fixed frequency oscillator for all practical purposes.

Advantages of Phase Shift Oscillators:

- Due to the absence of expensive and bulky high-value inductors, circuit is simple to design and well suited for frequencies below 10 KHz.
- These can produce pure sinusoidal waveform since only one frequency can fulfill the Barkhausen phase shift requirement.
- It is fixed to one frequency.

Disadvantages of Phase Shift Oscillators:

For a variable frequency usage, phase shift oscillators are not suited because the capacitor values will have to be varied. And also, for frequency change in every time requires gain adjustment for satisfying the condition of oscillations.

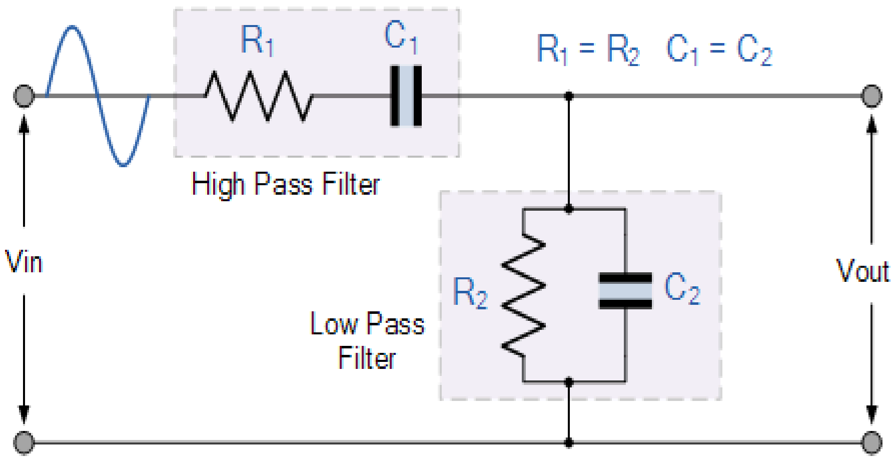
- These oscillators produce 5% of distortion level in the output.
- This oscillator gives only a small output due to smaller feedback
- These oscillator circuits require a high gain which is practically impossible.
- The frequency stability is poor due to the effect of temperature, aging, etc. of various circuit components.

One of the simplest sine wave oscillators which uses a RC network in place of the conventional LC tuned tank circuit to produce a sinusoidal output waveform, is called a **Wien Bridge Oscillator**.

The **Wien Bridge Oscillator** is so called because the circuit is based on a frequency-selective form of the Wheatstone bridge circuit. The Wien Bridge oscillator is a two-stage RC coupled amplifier circuit that has good stability at its resonant frequency, low distortion and is very easy to tune making it a popular circuit as an audio frequency oscillator but the phase shift of the output signal is considerably different from the previous phase shift **RC Oscillator**.

The **Wien Bridge Oscillator** uses a feedback circuit consisting of a series RC circuit connected with a parallel RC of the same component values producing a phase delay or phase advance circuit depending upon the frequency. At the resonant frequency fr the phase shift is 0°. Consider the circuit below.

RC Phase Shift Network

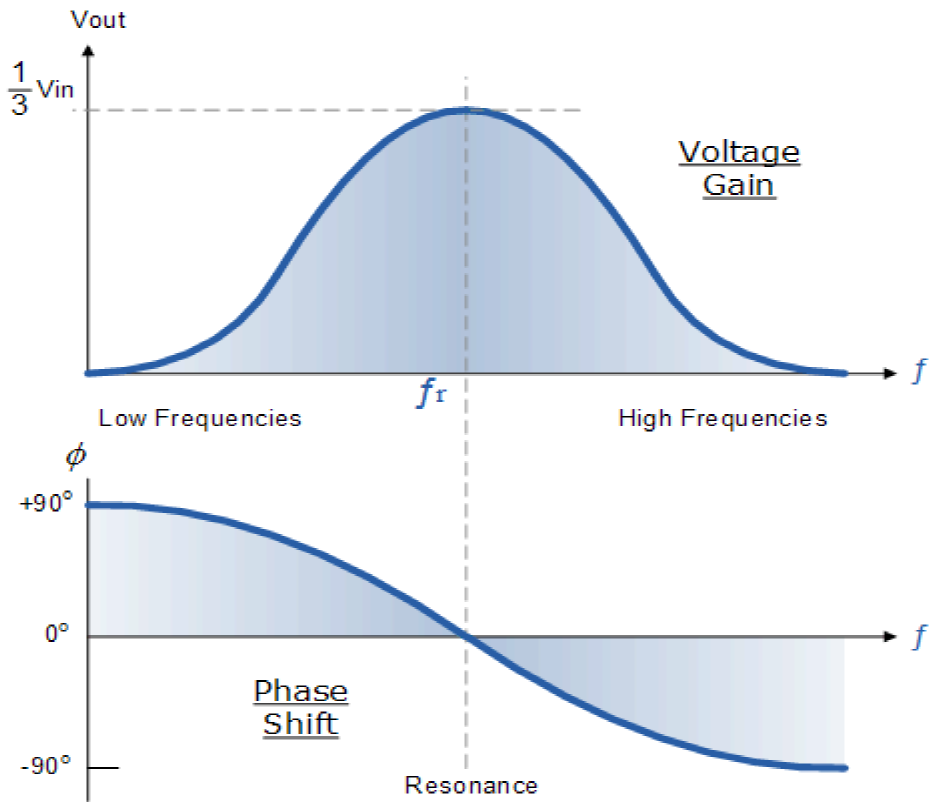


The above RC network consists of a series RC circuit connected to a parallel RC forming basically a **High Pass Filter** connected to a **Low Pass Filter** producing a very selective second-order frequency dependant **Band Pass Filter** with a high Q factor at the selected frequency, fr.

At low frequencies the reactance of the series capacitor (C1) is very high so acts like an open circuit and blocks any input signal at Vin. Therefore there is no output signal, Vout. At high frequencies, the reactance of the parallel capacitor, (C2) is very low so this parallel connected capacitor acts like a short circuit on the output so again there is no output signal. However, between these two extremes the output voltage reaches a maximum value with the frequency at which this happens being called the *Resonant Frequency*, (fr).

At this resonant frequency, the circuits reactance equals its resistance as Xc = R so the phase shift between the input and output equals zero degrees. The magnitude of the output voltage is therefore at its maximum and is equal to one third (1/3) of the input voltage as shown.

Oscillator Output Gain and Phase Shift



It can be seen that at very low frequencies the phase angle between the input and output signals is "Positive" (Phase Advanced), while at very high frequencies the phase angle becomes "Negative" (Phase Delay). In the middle of these two points the circuit is at its resonant frequency, (fr) with the two signals being "in-phase" or 0°. We can therefore define this resonant frequency point with the following expression.

Wien Bridge Oscillator Frequency

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

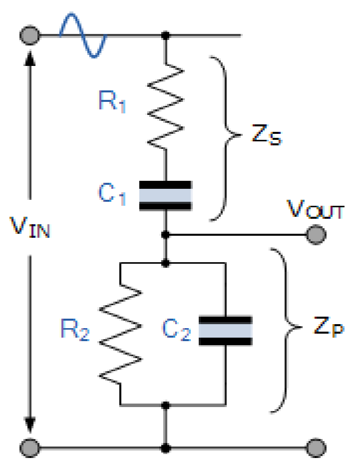
Where:

- f_r is the Resonant Frequency in Hertz
- R is the Resistance in Ohms
- C is the Capacitance in Farads

We said previously that the magnitude of the output voltage, Vout from the RC network is at its maximum value and equal to one third (1/3) of the input voltage, Vin to allow for oscillations to occur. But why one third and not some other value. In order to understand why the output from the RC circuit above needs to be one-third, that is 0.333xVin, we have to consider the complex impedance ($Z = R \pm jX$) of the two connected RC circuits.

We know from our **AC Theory tutorials** that the real part of the complex impedance is the resistance, R while the imaginary part is the reactance, X. As we are dealing with capacitors here, the reactance part will be capacitive reactance, Xc.

The RC Network



If we redraw the above RC network as shown, we can clearly see that it consists of two RC circuits connected together with the output taken from their junction. Resistor R_1 and capacitor C_1 form the top series network, while resistor R_2 and capacitor C_2 form the bottom parallel network.

Therefore the total impedance of the series combination (R_1C_1) we can call, Z_S and the total impedance of the parallel combination (R_2C_2) we can call, Z_P. As Z_S and Z_P are effectively connected together in series across the input, V_{IN}, they form a voltage divider network with the output taken from across $Z_P$as shown.

Let's assume then that the component values of R_1 and R_2 are the same at: 12kΩ, capacitors C_1 and C_2 are the same at: 3.9nF and the supply frequency, f is 3.4kHz.

Series Circuit

The total impedance of the series combination with resistor, R_1 and capacitor, C_1 is simply:

$$R = 12k\Omega, \quad \text{but} \quad X_C = \frac{1}{2\pi f C}$$

$$\therefore X_C = \frac{1}{2\pi \times 3.4kHz \times 3.9nF} = 12k\Omega$$

$$Z_S = \sqrt{R^2 + X_C^2} = \sqrt{12000^2 + 12000^2}$$

$$\therefore Z_S = 16{,}970\Omega \text{ or } 17k\Omega$$

We now know that with a supply frequency of 3.4kHz, the reactance of the capacitor is the same as the resistance of the resistor at 12kΩ. This then gives us an upper series impedance Z_S of 17kΩ.

For the lower parallel impedance Z_P, as the two components are in parallel, we have to treat this differently because the impedance of the parallel circuit is influenced by this parallel combination.

Parallel Circuit

The total impedance of the lower parallel combination with resistor, R_2 and capacitor, C_2 is given as:

$$R = 12k\Omega, \quad \text{and} \quad X_C = 12k\Omega$$

$$\frac{1}{Z} = \frac{1}{R} + \frac{1}{XC} = \frac{1}{12000} + \frac{1}{12000}$$

$$\therefore Z = 6000\Omega \text{ or } 6k\Omega$$

At the supply frequency of 3400Hz, or 3.4KHz, the combined resistance and reactance of the RC parallel circuit becomes 6kΩ (R||Xc) and their parallel impedance is therefore calculated as:

$$R = 6k\Omega, \text{ and } X_C = 6k\Omega \text{ (Parallel)}$$

$$Z_P = \sqrt{R^2 + X_C^2} = \sqrt{6000^2 + 6000^2}$$

$$\therefore Z_P = 8485\Omega \text{ or } 8.5k\Omega$$

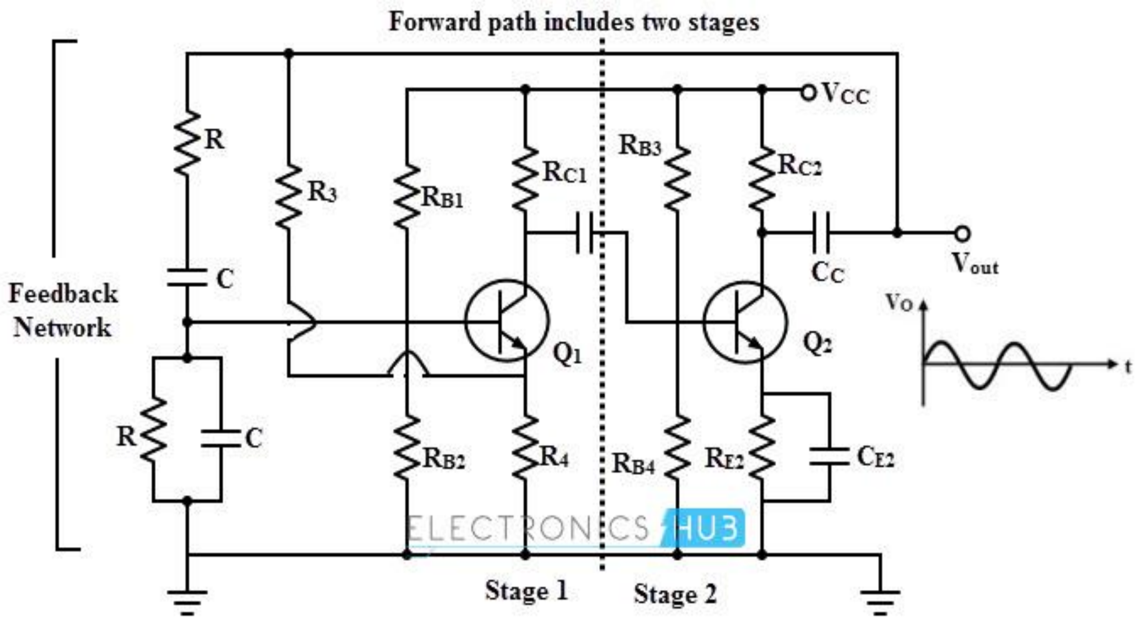
So we now have the value for the series impedance of: 17kΩ's, (Z_S = 17kΩ) and for the parallel impedance of: 8.5kΩ's, (Z_S = 8.5kΩ). Therefore the output impedance, Zout of the voltage divider network at the given frequency is:

$$Z_{OUT} = \frac{Z_P}{Z_P + Z_S} = \frac{8.5k\Omega}{8.5k\Omega + 17k\Omega} = 0.333 \text{ or } \frac{1}{3}$$

Then at the oscillation frequency, the magnitude of the output voltage, Vout will be equal to Zout x Vin which as shown is equal to one third (1/3) of the input voltage, Vin and it is this frequency selective RC network which forms the basis of the **Wien Bridge Oscillator** circuit.
If we now place this RC network across a non-inverting amplifier which has a gain of 1+R1/R2 the following basic wien bridge oscillator circuit is produced.

Transistorized Wien Bridge Oscillator:
The figure below shows the transistorized Wien bridge oscillator which uses two stage common emitter transistor amplifier. Each amplifier stage introduces a phase shift of 180 degrees and hence a total 360 degrees phase shift is introduced which is nothing but a zero phase shift condition.

The feedback bridge consists of RC series elements, RC parallel elements, R3 and R4 resistances. The input to the bridge circuit is applied from the collector of transistor T2 through a coupling capacitor.



When the DC source is applied to the circuit, a noise signal is at the base of the transistor T1 is generated due to the movement of charge carriers through transistor and other circuit components. This voltage is amplified with gain A and produce output voltage 180 degrees out of phase with input voltage. This output voltage is applied as input to second transistor at base terminal of T2. This voltage is multiplied with gain of the T2. The amplified output of the transistor T2 is 180 degrees out of phase with the output of the T1. This output is feedback to the transistor T1 through the coupling capacitor C. So the oscillations are produced at wide range of frequencies by this positive feedback when Barkhausen conditions are satisfied. Generally, the Wien bridge in the feedback network incorporates the oscillations at single desired frequency. The bridge is get balanced at the frequency at which total phase shift is zero.

The output of the two stage transistor acts as an input to the feedback network which is applied between the base and ground.

Feedback voltage,

$$V_f = (V_o \times R4) / (R3 + R4)$$

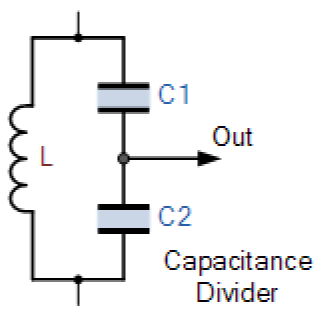
Advantages:

- Because of the usage of two stage amplifier, the overall gain of this oscillator is high.
- By varying the values of C1 and C2 or with use of variable resistors, the frequency of oscillations can be varied.
- It produces a very good sine wave with less distortion
- The frequency stability is good.
- Due to the absence of inductors, no interference occurs from external magnetic fields.

Disadvantages:

- More number of components is needed for two stage amplifier type of Wien bridge oscillators.
- Very high frequencies cannot be generated.

In many ways, the Colpitts oscillator is the exact opposite of the **Hartley Oscillator** we looked at in the previous tutorial. Just like the Hartley oscillator, the tuned tank circuit consists of an LC resonance sub-circuit connected between the collector and the base of a single stage transistor amplifier producing a sinusoidal output waveform. The basic configuration of the **Colpitts Oscillator** resembles that of the *Hartley Oscillator* but the difference this time is that the centre tapping of the tank sub-circuit is now made at the junction of a "capacitive voltage divider" network instead of a tapped autotransformer type inductor as in the Hartley oscillator. Related Products: Oscillators and Crystals | Controlled Oscillator | MEMS Oscillators | Oscillator Misc | Silicon Oscillators

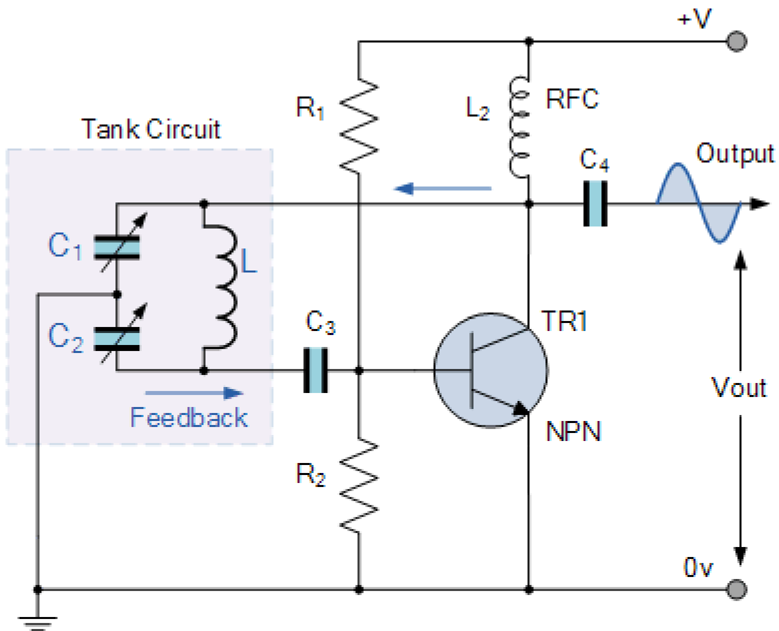


Colpitts Oscillator:

Tank Circuit:

The Colpitts oscillator uses a capacitive voltage divider network as its feedback source. The two capacitors, C1 and C2 are placed across a single common inductor, L as shown. Then C1, C2 and L form the tuned tank circuit with the condition for oscillations being: $X_{C1} + X_{C2} = X_L$, the same as for the Hartley oscillator circuit. The advantage of this type of capacitive circuit configuration is that with less self and mutual inductance within the tank circuit, frequency stability of the oscillator is improved along with a more simple design. As with the Hartley oscillator, the Colpitts oscillator uses a single stage bipolar transistor amplifier as the gain element which produces a sinusoidal output. Consider the circuit below.

Basic Colpitts Oscillator Circuit



The emitter terminal of the transistor is effectively connected to the junction of the two capacitors, C1 and C2 which are connected in series and act as a simple voltage divider. When the power supply is firstly applied, capacitors C1 and C2 charge up and then discharge through the coil L. The oscillations across the capacitors are applied to the base-emitter junction and appear in the amplified at the collector output.

Resistors, R1 and R2 provide the usual stabilizing DC bias for the transistor in the normal manner while the additional capacitors act as a DC-blocking bypass capacitors. A radio-frequency choke (RFC) is used in the collector circuit to provide a high reactance (ideally open circuit) at the frequency of oscillation, (fr) and a low resistance at DC to help start the oscillations.

The required external phase shift is obtained in a similar manner to that in the Hartley oscillator circuit with the required positive feedback obtained for sustained undamped oscillations. The amount of feedback is determined by the ratio of C1 and C2. These two capacitances are generally “ganged” together to provide a constant amount of feedback so that as one is adjusted the other automatically follows.

The frequency of oscillations for a Colpitts oscillator is determined by the resonant frequency of the LC tank circuit and is given as:

$$f_r = \frac{1}{2\pi\sqrt{L C_T}}$$

Where C_T is the capacitance of C1 and C2 connected in series and is given as:

$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} \quad \text{or} \quad C_T = \frac{C_1 \times C_2}{C_1 + C_2}$$

The configuration of the transistor amplifier is of a **Common Emitter Amplifier** with the output signal 180° out of phase with regards to the input signal. The additional 180°phase shift require for oscillation is achieved by the fact that the two capacitors are connected together in series but in parallel with the inductive coil resulting in overall phase shift of the circuit being zero or 360°. The amount of feedback depends on the values of C1 and C2. We can see that the voltage across C1 is the the same as the oscillators output voltage, Vout and that the voltage across C2 is the oscillators feedback voltage. Then the voltage across C1 will be much greater than that across C2.

Therefore, by changing the values of capacitors, C1 and C2 we can adjust the amount of feedback voltage returned to the tank circuit. However, large amounts of feedback may cause the output sine wave to become distorted, while small amounts of feedback may not allow the circuit to oscillate.

Then the amount of feedback developed by the Colpitts oscillator is based on the capacitance ratio of C1 and C2 and is what governs the the excitation of the oscillator. This ratio is called the "feedback fraction" and is given simply as:

$$\text{Feedback Fraction} = \frac{C_1}{C_2}\ \%$$

Colpitts Oscillator Example :

A Colpitts Oscillator circuit having two capacitors of 24nF and 240nF respectively are connected in parallel with an inductor of 10mH. Determine the frequency of oscillations of the circuit, the feedback fraction and draw the circuit.

The oscillation frequency for a Colpitts Oscillator is given as:

$$f_r = \frac{1}{2\pi\sqrt{L\,C_T}}$$

As the colpitts circuit consists of two capacitors in series, the total capacitance is therefore:

$$C_T = \frac{24nF \times 240nF}{24nF + 240nF} = 21.82nF$$

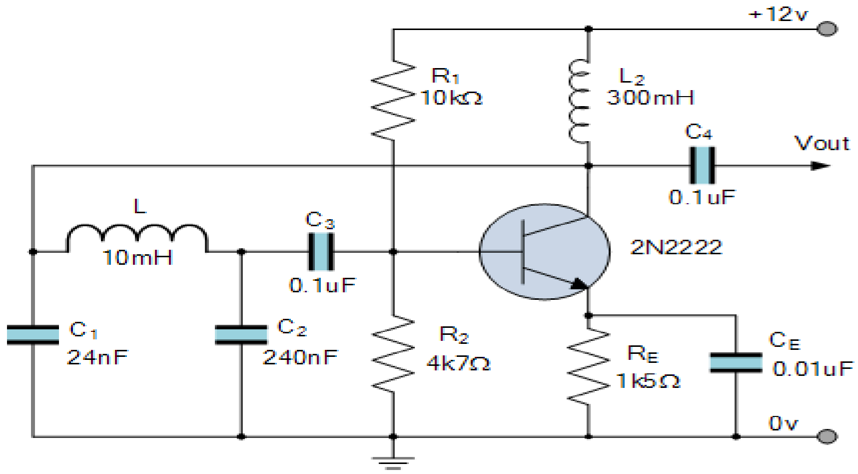
The inductance of the inductor is given as 10mH, then the frequency of oscillation is:

$$f_r = \frac{1}{2\pi\sqrt{LC_T}} = \frac{1}{6.283\sqrt{0.01 \times 21.82 \times 10^{-9}}} = 10.8kHz$$

The frequency of oscillations for the Colpitts Oscillator is therefore 10.8kHz with the feedback fraction given as:

$$F_F = \frac{C_1}{C_2} = \frac{24nF}{240nF} = 10\%$$

Colpitts Oscillator Circuit:



Colpitts Oscillator Summary:

Then to summaries, the Colpitts Oscillator consists of a parallel LC resonator tank circuit whose feedback is achieved by way of a capacitive divider. Like most oscillator circuits, the Colpitts oscillator exists in several forms, with the most common form being the transistor circuit above.

The centre tapping of the tank sub-circuit is made at the junction of a "capacitive voltage divider" network to feed a fraction of the output signal back to the emitter of the transistor. The two capacitors in series produce a 180^o phase shift which is inverted by another 180^o to produce the required positive feedback. The oscillating frequency which is a purer sine-wave voltage is determined by the resonance frequency of the tank circuit.

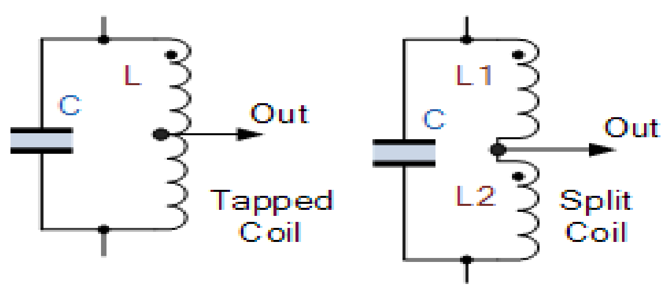
Applications of Colpitts oscillator

- Colpitts oscillators are used for high frequency range and high frequency stability
- A surface acoustical wave (SAW) resonator
- Microwave applications
- Mobile and communication systems
- These are used in chaotic circuits which are capable to generate oscillations from audio frequency range to the optical band. These application areas include broadband communications, spectrum spreading, signal masking, etc.

One of the main disadvantages of the basic LC Oscillator circuit we looked at in the previous tutorial is that they have no means of controlling the amplitude of the oscillations and also, it is difficult to tune the oscillator to the required frequency. If the cumulative electromagnetic coupling between L_1 and L_2 is too small there would be insufficient feedback and the oscillations would eventually die away to zero.

Likewise if the feedback was too strong the oscillations would continue to increase in amplitude until they were limited by the circuit conditions producing signal distortion. So it becomes very difficult to "tune" the oscillator. However, it is possible to feed back exactly the right amount of voltage for constant amplitude oscillations. If we feed back more than is necessary the amplitude of the oscillations can be controlled by biasing the amplifier in such a way that if the oscillations increase in amplitude, the bias is increased and the gain of the amplifier is reduced. If the amplitude of the oscillations decreases the bias decreases and the gain of the amplifier increases, thus increasing the feedback. In this way the amplitude of the oscillations are kept constant using a process known as Automatic Base Bias**.**

One big advantage of automatic base bias in a voltage controlled oscillator, is that the oscillator can be made more efficient by providing a Class-B bias or even a Class-C bias condition of the transistor. This has the advantage that the collector current only flows during part of the oscillation cycle so the quiescent collector current is very small. Then this "self-tuning" base oscillator circuit forms one of the most common types of LC parallel resonant feedback oscillator configurations called the **Hartley Oscillator** circuit.



Hartley Oscillator Tank Circuit:

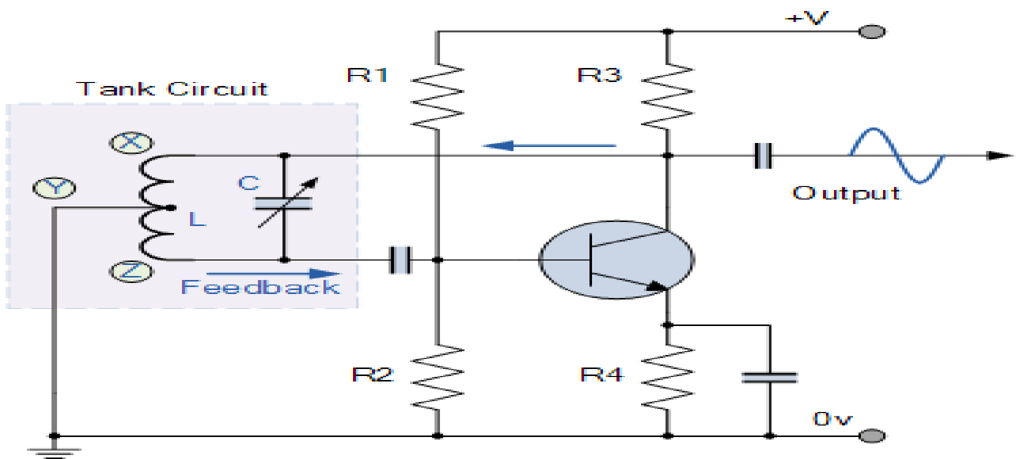
In the **Hartley Oscillator** the tuned LC circuit is connected between the collector and the base of a transistor amplifier. As far as the oscillatory voltage is concerned, the emitter is connected to a tapping point on the tuned circuit coil.

The feedback part of the tuned LC tank circuit is taken from the centre tap of the inductor coil or even two separate coils in series which are in parallel with a variable capacitor, C as shown.

The Hartley circuit is often referred to as a split-inductance oscillator because coil L is centre-tapped. In effect, inductance L acts like two separate coils in very close proximity with the current flowing through coil section XY induces a signal into coil section YZ below.

An Hartley Oscillator circuit can be made from any configuration that uses either a single tapped coil (similar to an autotransformer) or a pair of series connected coils in parallel with a single capacitor as shown below.

Basic Hartley Oscillator Design



When the circuit is oscillating, the voltage at point X (collector), relative to point Y (emitter), is 180^o out-of-phase with the voltage at point Z (base) relative to point Y. At the frequency of oscillation, the impedance of the Collector load is resistive and an increase in Base voltage causes a decrease in the Collector voltage.

Then there is a 180^o phase change in the voltage between the Base and Collector and this along with the original 180^o phase shift in the feedback loop provides the correct phase relationship of positive feedback for oscillations to be maintained.

The amount of feedback depends upon the position of the “tapping point” of the inductor. If this is moved nearer to the collector the amount of feedback is increased, but the output taken between the Collector and earth is reduced and vice versa. Resistors, R1 and R2 provide the usual stabilizing DC bias for the transistor in the normal manner while the capacitors act as DC-blocking capacitors.

In this **Hartley Oscillator** circuit, the DC Collector current flows through part of the coil and for this reason the circuit is said to be "Series-fed" with the frequency of oscillation of the Hartley Oscillator being given as.

$$f = \frac{1}{2\pi\sqrt{L_T C}}$$

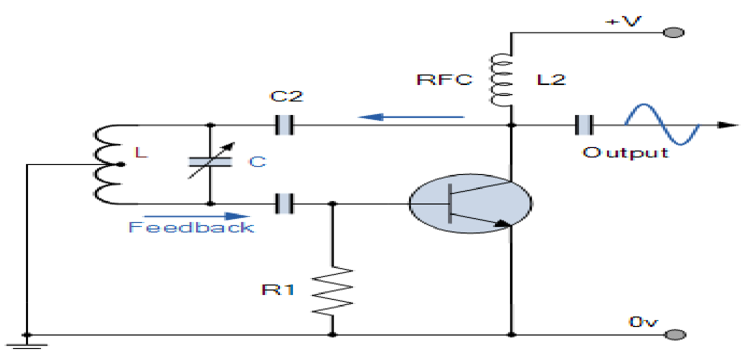
$$\text{where:} \quad L_T = L_1 + L_2 + 2M$$

Note: L_T is the total cumulatively coupled inductance if two separate coils are used including their mutual inductance, M.

The frequency of oscillations can be adjusted by varying the "tuning" capacitor, C or by varying the position of the iron-dust core inside the coil (inductive tuning) giving an output over a wide range of frequencies making it very easy to tune. Also the **Hartley Oscillator** produces an output amplitude which is constant over the entire frequency range.

As well as the Series-fed Hartley Oscillator above, it is also possible to connect the tuned tank circuit across the amplifier as a shunt-fed oscillator as shown below.

Shunt-fed Hartley Oscillator Circuit:



In the shunt-fed Hartley oscillator circuit, both the AC and DC components of the Collector current have separate paths around the circuit. Since the DC component is blocked by the capacitor, C2 no DC flows through the inductive coil, L and less power is wasted in the tuned circuit.

The Radio Frequency Coil (RFC), L2 is an RF choke which has a high reactance at the frequency of oscillations so that most of the RF current is applied to the LC tuning tank circuit via capacitor, C2 as the DC component passes through L2 to the power supply. A resistor could be used in place of the RFC coil, L2 but the efficiency would be less.

Hartley Oscillator Example

A Hartley Oscillator circuit having two individual inductors of 0.5mH each, are designed to resonate in parallel with a variable capacitor that can be adjusted between 100pF and 500pF. Determine the upper and lower frequencies of oscillation and also the Hartley oscillators bandwidth.

From above we can calculate the frequency of oscillations for a Hartley Oscillator as:

$$f_r = \frac{1}{2\pi\sqrt{L_T C}}$$

The circuit consists of two inductive coils in series, so the total inductance is given as:

$$L_T = L_1 + L_2 = 0.5\text{mH} + 0.5\text{mH} = 1.0\text{mH}$$

Hartley Oscillator Upper Frequency

$$f_H = \frac{1}{2\pi\sqrt{1\text{mH} \times 100\text{pF}}} = \frac{1}{6.283\sqrt{1 \times 10^{-13}}} = 503228\text{Hz}$$

$$\therefore f_H = 503\text{kHz}$$

Hartley Oscillator Lower Frequency

$$f_L = \frac{1}{2\pi\sqrt{1mH \times 500pF}} = \frac{1}{6.283\sqrt{5\times10^{-13}}} = 225050Hz$$

$$\therefore f_L = 225kHz$$

Hartley Oscillator Bandwidth

$$Bandwidth = f_H - f_L$$

$$= 503 - 225 = 278kHz$$

Mutual Inductance in Hartley Oscillator:

The change in current through coil induces the current in other vicinity coil by the magnetic field is called as mutual inductance. It is an additional amount of inductance caused in one inductor due to the magnetic flux of other inductor. By considering the effect of mutual inductance, the total inductance of the coils can be calculated by the formula given below.

$$Leq = L1 + L2 + 2M$$

Where M is the mutual inductance and its value depends on the effective coupling between the inductors, spacing between them, dimensions of each coil, number of turns in each coil and type of material used for the common core. In radio frequency oscillators, depending on the North and south polarities of the fields generated by the closely coupled inductors, the total inductance of the circuit is determined. If the fields generated by the individual coils are in the same direction, then the mutual inductance will add to the total inductance, hence the total inductance is increased. If the fields are in opposite direction, then the mutual inductance will reduce the total inductance. Therefore, the oscillator working frequency will be increased.

The design of the Hartley oscillator considers this mutual effect of the two inductors. In practical, a common core is used for both inductors, however depending on the coefficient of coupling the mutual inductance effect can be much greater. This coefficient value is unity when there is hundred percent magnetic coupling between the inductors and its value is zero if there is no magnetic coupling between the inductors.

The Hartley Oscillator Summary:

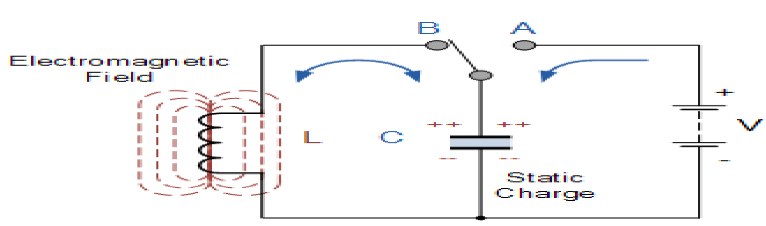
Then to summarize, the Hartley Oscillator consists of a parallel LC resonator tank circuit whose feedback is achieved by way of an inductive divider. Like most oscillator circuits, the Hartley oscillator exists in several forms, with the most common form being the transistor circuit above.

This *Hartley Oscillator* configuration has a tuned tank circuit with its resonant coil tapped to feed a fraction of the output signal back to the emitter of the transistor. Since the output of the transistors emitter is always "in-phase" with the output at the collector, this feedback signal is positive. The oscillating frequency which is a sine-wave voltage is determined by the resonance frequency of the tank circuit.

In the next tutorial about Oscillators, we will look at another type of LC oscillator circuit that is the opposite to the Hartley oscillator called the Colpitts oscillator. The Colpitts oscillator uses two capacitors in series to form a centre tapped capacitance in parallel with a single inductance within its resonant tank circuit.

When a constant voltage but of varying frequency is applied to a circuit consisting of an inductor, capacitor and resistor the reactance of both the Capacitor/Resistor and Inductor/Resistor circuits is to change both the amplitude and the phase of the output signal as compared to the input signal due to the reactance of the components used. At high frequencies the reactance of a capacitor is very low acting as a short circuit while the reactance of the inductor is high acting as an open circuit. At low frequencies the reverse is true, the reactance of the capacitor acts as an open circuit and the reactance of the inductor acts as a short circuit. Between these two extremes the combination of the inductor and capacitor produces a "Tuned" or "Resonant" circuit that has a **Resonant Frequency**, (fr) in which the capacitive and inductive reactance's are equal and cancel out each other, leaving only the resistance of the circuit to oppose the flow of current. This means that there is no phase shift as the current is in phase with the voltage. Consider the circuit below.

Basic LC Oscillator Tank Circuit:



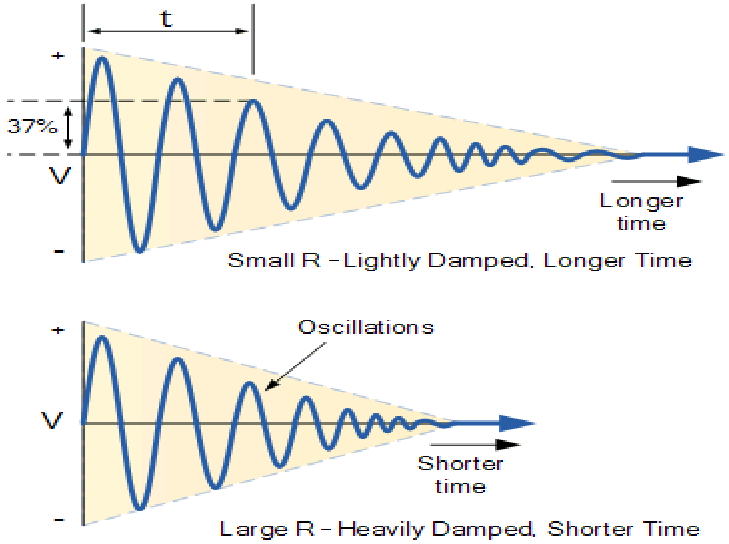
The circuit consists of an inductive coil, L and a capacitor, C. The capacitor stores energy in the form of an electrostatic field and which produces a potential (*static voltage*) across its plates, while the inductive coil stores its energy in the form of an electromagnetic field. The capacitor is charged up to the DC supply voltage, V by putting the switch in position A.

When the capacitor is fully charged the switch changes to position B. The charged capacitor is now connected in parallel across the inductive coil so the capacitor begins to discharge itself through the coil. The voltage across C starts falling as the current through the coil begins to rise. This rising current sets up an electromagnetic field around the coil which resists this flow of current. When the capacitor, C is completely discharged the energy that was originally stored in the capacitor, C as an electrostatic field is now stored in the inductive coil, L as an electromagnetic field around the coils windings. As there is now no external voltage in the circuit to maintain the current within the coil, it starts to fall as the electromagnetic field begins to collapse. A back emf is induced in the coil ($e = -Ldi/dt$) keeping the current flowing in the original direction. These current charges up capacitor, C with the opposite polarity to its original charge. C continues to charge up until the current reduces to zero and the electromagnetic field of the coil has collapsed completely.

The energy originally introduced into the circuit through the switch, has been returned to the capacitor which again has an electrostatic voltage potential across it, although it is now of the opposite polarity. The capacitor now starts to discharge again back through the coil and the whole process is repeated. The polarity of the voltage changes as the energy is passed back and forth between the capacitor and inductor producing an AC type sinusoidal voltage and current waveform. This process then forms the basis of an LC oscillators tank circuit and theoretically this cycling back and forth will continue indefinitely.

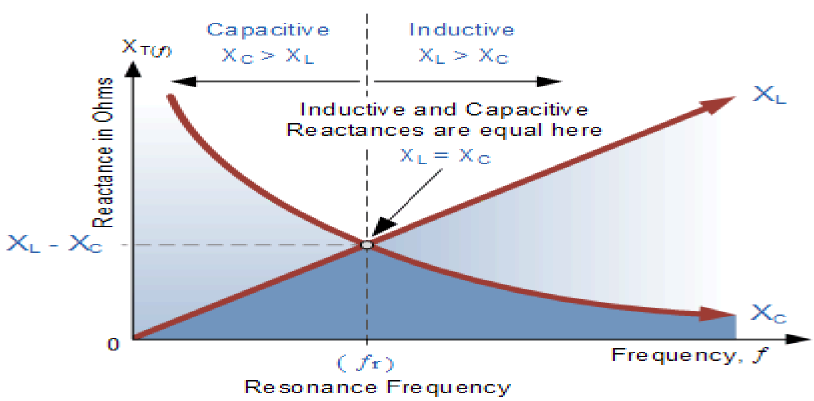
However, things are not perfect and every time energy is transferred from the capacitor, C to inductor, L and back from L to C some energy losses occur which decay the oscillations to zero over time. This oscillatory action of passing energy back and forth between the capacitor, C to the inductor, L would continue indefinitely if it was not for energy losses within the circuit. Electrical energy is lost in the DC or real resistance of the inductors coil, in the dielectric of the capacitor, and in radiation from the circuit so the oscillation steadily decreases until they die away completely and the process stops. Then in a practical LC circuit the amplitude of the oscillatory voltage decreases at each half cycle of oscillation and will eventually die away to zero. The oscillations are then said to be “damped” with the amount of damping being determined by the quality or Q-factor of the circuit.

Damped Oscillations



The frequency of the oscillatory voltage depends upon the value of the inductance and capacitance in the LC tank circuit. We now know that for *resonance* to occur in the tank circuit, there must be a frequency point were the value of X_C, the capacitive reactance is the same as the value of X_L, the inductive reactance ($X_L = X_C$) and which will therefore cancel out each other out leaving only the DC resistance in the circuit to oppose the flow of current. If we now place the curve for inductive reactance of the inductor on top of the curve for capacitive reactance of the capacitor so that both curves are on the same frequency axes, the point of intersection will give us the resonance frequency point, (fr or ωr) as shown below.

Resonance Frequency



Where: f_r is in Hertz, L is in Henries and C is in Farads. Then the frequency at which this will happen is given as:

$$X_L = 2\pi f L \quad \text{and} \quad X_C = \frac{1}{2\pi f C}$$

$$\text{at resonance:} \quad X_L = X_C$$

$$\therefore \quad 2\pi f L = \frac{1}{2\pi f C}$$

$$2\pi f^2 L = \frac{1}{2\pi C}$$

$$\therefore \quad f^2 = \frac{1}{(2\pi)^2 LC}$$

$$f = \frac{\sqrt{1}}{\sqrt{(2\pi)^2 LC}}$$

Then by simplifying the above equation we get the final equation for **Resonant Frequency**, fr in a tuned LC circuit as:

Resonant Frequency of a LC Oscillator:

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

Where:

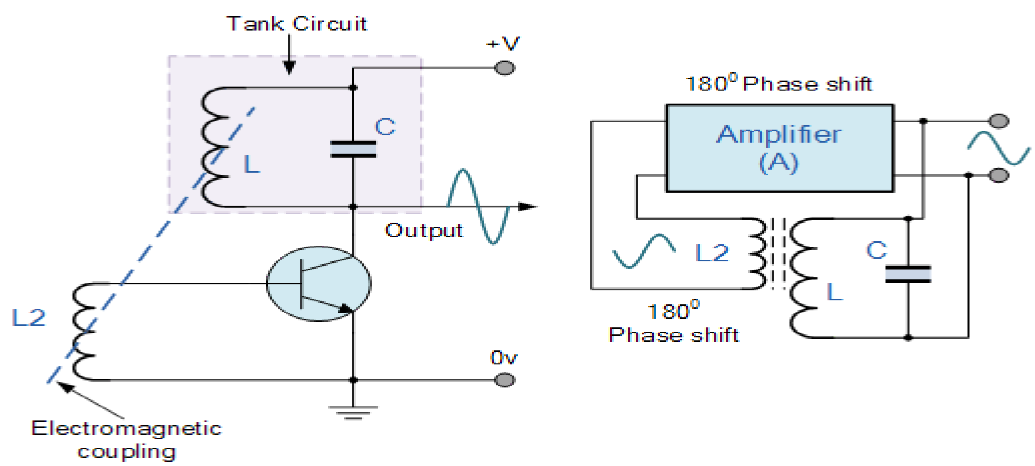
- L is the Inductance in Henries
- C is the Capacitance in Farads
- fr is the Output Frequency in Hertz

This equation shows that if either L or C are decreased, the frequency increases. This output frequency is commonly given the abbreviation of (fr) to identify it as the “resonant frequency”.To keep the oscillations going in an LC tank circuit, we have to replace all the energy lost in each oscillation and also maintain the amplitude of these oscillations at a constant level. The amount of energy replaced must therefore be equal to the energy lost during each cycle.

If the energy replaced is too large the amplitude would increase until clipping of the supply rails occurs. Alternatively, if the amount of energy replaced is too small the amplitude would eventually decrease to zero over time and the oscillations would stop. The simplest way of replacing this lost energy is to take part of the output from the LC tank circuit, amplify it and then feed it back into the LC circuit again. This process can be achieved using a voltage amplifier using an op-amp, FET or bipolar transistor as its active device. However, if the loop gain of the feedback amplifier is too small, the desired oscillation decays to zero and if it is too large, the waveform becomes distorted.

To produce a constant oscillation, the level of the energy fed back to the LC network must be accurately controlled. Then there must be some form of automatic amplitude or gain control when the amplitude tries to vary from a reference voltage either up or down.To maintain a stable oscillation the overall gain of the circuit must be equal to one or unity. Any less and the oscillations will not start or die away to zero, any more the oscillations will occur but the amplitude will become clipped by the supply rails causing distortion. Consider the circuit below.

Basic Transistor LC Oscillator Circuit



A **Bipolar Transistor** is used as the LC oscillator’s amplifier with the tuned LC tank circuit acts as the collector load. Another coil L2 is connected between the base and the emitter of the transistor whose electromagnetic field is “mutually” coupled with that of coil L.

“Mutual inductance” exists between the two circuits and the changing current flowing in one coil circuit induces, by electromagnetic induction, a potential voltage in the other (transformer effect) so as the oscillations occur in the tuned circuit, electromagnetic energy is transferred from coil L to coil L2 and a voltage of the same frequency as that in the tuned circuit is applied between the base and emitter of the transistor. In this way the necessary automatic feedback voltage is applied to the amplifying transistor.

The amount of feedback can be increased or decreased by altering the coupling between the two coils L and L2. When the circuit is oscillating its impedance is resistive and the collector and base voltages are 180^o out of phase. In order to maintain oscillations (called frequency stability) the voltage applied to the tuned circuit must be “in-phase” with the oscillations occurring in the tuned circuit.

Therefore, we must introduce an additional 180^o phase shift into the feedback path between the collector and the base. This is achieved by winding the coil of L2 in the correct direction relative to coil L giving us the correct amplitude and phase relationships for the **Oscillators** circuit or by connecting a phase shift network between the output and input of the amplifier. The **LC Oscillator** is therefore a “Sinusoidal Oscillator” or a “Harmonic Oscillator” as it is more commonly called. LC oscillators can generate high frequency sine waves for use in radio frequency (RF) type applications with the transistor amplifier being of a Bipolar Transistor or FET.

Harmonic Oscillators come in many different forms because there are many different ways to construct an LC filter network and amplifier with the most common being the **Hartley LC Oscillator**, **Colpitts LC Oscillator**, **Armstrong Oscillator** and **Clapp Oscillator** to name a few.

LC Oscillator Example:
An inductance of 200mH and a capacitor of 10pF are connected together in parallel to create an LC oscillator tank circuit. Calculate the frequency of oscillation.

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{200mH \times 10pF}} = 112.5\ kHz$$

Then we can see from the above example that by decreasing the value of either the capacitance, C or the inductance, L will have the effect of increasing the frequency of oscillation of the LC tank circuit.

LC Oscillators Summary:

The basic conditions required for an **LC oscillator** resonant tank circuit are given as follows.

- For oscillations to exist an oscillator circuit MUST contain a reactive (frequency-dependant) component either an “Inductor”, (L) or a “Capacitor”, (C) as well as a DC power source.
- In a simple inductor-capacitor, LC circuit, oscillations become damped over time due to component and circuit losses.
- Voltage amplification is required to overcome these circuit losses and provide positive gain.
- The overall gain of the amplifier must be greater than one, unity.
- Oscillations can be maintained by feeding back some of the output voltage to the tuned circuit that is of the correct amplitude and in-phase, (0^o).
- Oscillations can only occur when the feedback is “Positive” (self-regeneration).
- The overall phase shift of the circuit must be zero or 360^o so that the output signal from the feedback network will be “in-phase” with the input signal.

UNIT-V

Power Amplifiers:

Classification of amplifiers, Class A power Amplifiers and their analysis, Harmonic Distortions, Class B Push-pull amplifiers and their analysis, Complementary symmetry push pull amplifier, Class AB power amplifier, Class-C power amplifier, Thermal stability and Heat sinks, Distortion in amplifiers.

Introduction:

When the output to be delivered is large, much greater than mW range and is of the order of few watts or more watts, conventional transistor (BJT) amplifiers cannot be used. Such electronic amplifier circuits, delivering significant output power to the load (in watts range) are termed as *Power Amplifiers.* Since the input to this type of amplifier circuits is also *large*, they are termed as *Large Signal Amplifiers.*

In order to improve the circuit efficiency, which is the ratio of output power delivered to the load Po to input power, the device is operated in varying conduction angles of 360^0,180^0 less than 180^0 etc. Based on the variation of conduction angle, the amplifier circuits are classified as Class A, Class B, Class C, Class AB, Class D, and Class S.

Power Amplifier:

Large input signals are used to obtain appreciable power output from amplifiers. But if the input signal is large in magnitude, the operating point is driven over a considerable portion of the output characteristic of the transistor (BJT).

The transfer characteristic of a transistor which is a plot between the output current Ie and input voltage V BE is not linear. The transfer characteristic indicates the change in ic when Vb or IB is changed. For equal increments of VBE, increase in I_e will not be uniform since output characteristics are not linear (for equal increments of VBE, I_e will not increase by the same current). So the transfer characteristic is not linear.Hence because of this, when the magnitude of the input signal is very large, distortion is introduced in the output in large signal power amplifiers. To eliminate distortion in the output, push pull connection and negative feedback are employed.

Class A Operation:

If the *Q point is placed near the centre a/the linear region a/the* dynamic curve, class A operation results. Because the transistor will conduct for the complete 360°, distortion is low for small signals and conversion efficiency is low.

Class B Operation:

class B operation the Q point is set near cutoff. So output power will be more and conversion efficiency (ll) is more. Conduction is only for 180°, from 1t - 21t. Since the transistor Q point is beyond cutoff, the output is zero or the transistor will not conduct.

Output power is more because the complete linear region is available for an operating signal excursion, resulting from one half of the input wave. The other half of input wave gives no output, because it drives the transistor below cutoff.

Class C Operation:

Here Q point is set well beyond cutoff and the device conducts for less than *1800.* The conversion efficiency (η) can theoretically reach 100%. Distortion is very high. These are used in radio frequency circuits where resonant circuit may be used to filter the output waveform.

Class A and class B amplifiers are used in the audio frequency range. Class B and class C are used in Radio Frequency range where conversion efficiency is important.

Large Signal Amplifiers:

With respect to the input signal, the amplifier circuits are classified as

(i) Small signal amplifiers
(ii) (ii) Large signal amplifiers

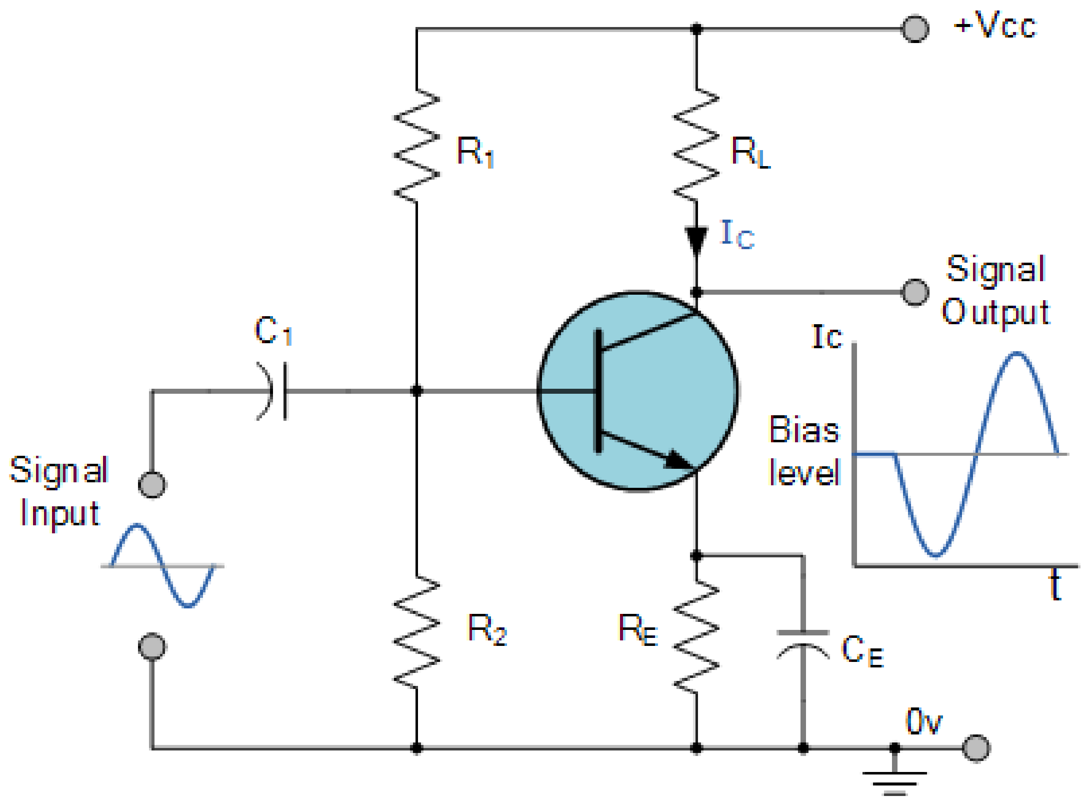
Class A Amplifier

The most commonly used type of power amplifier configuration is the **Class A Amplifier**. The Class A amplifier is the most common and simplest form of power amplifier that uses the switching transistor in the standard common emitter circuit configuration as seen previously.

The transistor is always biased "ON" so that it conducts during one complete cycle of the input signal waveform producing minimum distortion and maximum amplitude to the output. This means then that the **Class A Amplifier** configuration is the ideal operating mode, because there can be no crossover or switch-off distortion to the output waveform even during the negative half of the cycle.

Class A power amplifier output stages may use a single power transistor or pairs of transistors connected together to share the high load current. Consider the **Class A amplifier** circuit below.

Single Stage Amplifier Circuit

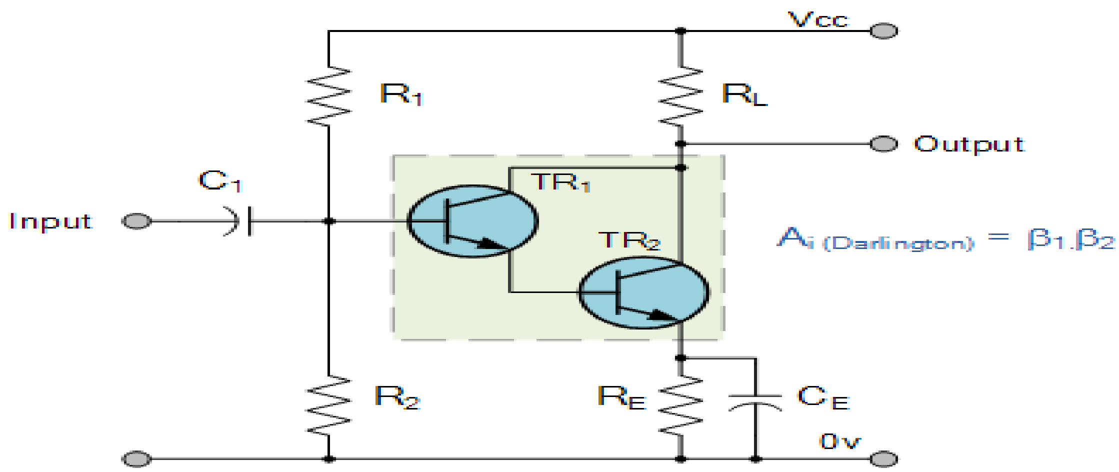


This is the simplest type of Class A power amplifier circuit. It uses a single-ended transistor for its output stage with the resistive load connected directly to the Collector terminal. When the transistor switches "ON" it sinks the output current through the Collector resulting in an inevitable voltage drop across the Emitter resistance thereby limiting the negative output capability.

The efficiency of this type of circuit is very low (less than 30%) and delivers small power outputs for a large drain on the DC power supply. A Class A amplifier stage passes the same load current even when no input signal is applied so large heatsinks are needed for the output transistors.

However, another simple way to increase the current handling capacity of the circuit while at the same time obtain a greater power gain is to replace the single output transistor with a **Darlington Transistor**. These types of devices are basically two transistors within a single package, one small "pilot" transistor and another larger "switching" transistor. The big advantage of these devices are that the input impedance is suitably large while the output impedance is relatively low, thereby reducing the power loss and therefore the heat within the switching device.

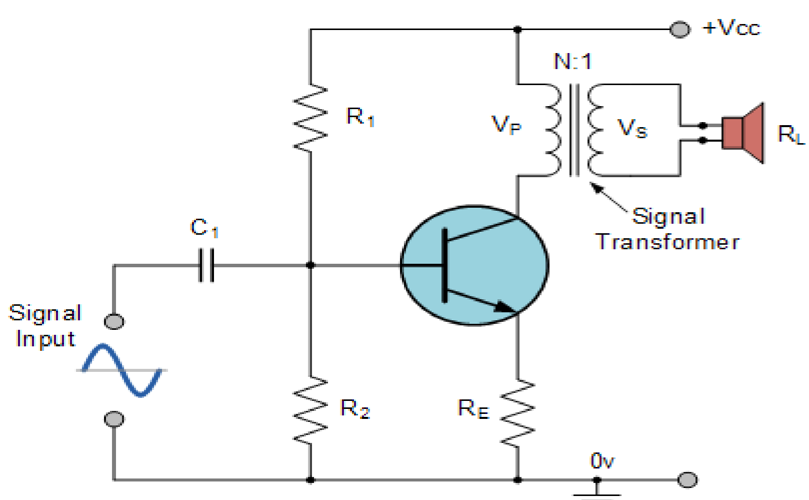
Darlington Transistor Configurations



The overall current gain Beta (β) or hfe value of a Darlington device is the product of the two individual gains of the transistors multiplied together and very high β values along with high Collector currents are possible compared to a single transistor circuit.

To improve the full power efficiency of the **Class A amplifier** it is possible to design the circuit with a transformer connected directly in the Collector circuit to form a circuit called a **Transformer Coupled Amplifier**. The transformer improves the efficiency of the amplifier by matching the impedance of the load with that of the amplifiers output using the turns ratio (n) of the transformer and an example of this is given below.

Transformer-coupled Amplifier Circuit



As the Collector current, Ic is reduced to below the quiescent Q-point set up by the base bias voltage, due to variations in the base current, the magnetic flux in the transformer core collapses causing an induced emf in the transformer primary windings. This causes an instantaneous collector voltage to rise to a value of twice the supply voltage 2Vcc giving a maximum collector current of twice Ic when the Collector voltage is at its minimum. Then the efficiency of this type of Class A amplifier configuration can be calculated as follows.

The r.m.s. Collector voltage is given as:

$$V_{CE} = \frac{V_{C(max)} - V_{C(min)}}{2\sqrt{2}} = \frac{2V_{CC} - 0}{2\sqrt{2}}$$

The r.m.s. Collector current is given as:

$$I_{CE} = \frac{I_{C(max)} - I_{C(min)}}{2\sqrt{2}} = \frac{2I_C - 0}{2\sqrt{2}}$$

The r.m.s. Power delivered to the load (Pac) is therefore given as:

$$P_{ac} = V_{CE} \times I_{CE} = \frac{2V_{CC}}{2\sqrt{2}} \times \frac{2I_C}{2\sqrt{2}} = \frac{2V_{CC}\,2I_C}{8}$$

The average power drawn from the supply (Pdc) is given by:

$$P_{dc} = V_{CC} \times I_C$$

and therefore the efficiency of a Transformer-coupled Class A amplifier is given as:

$$\eta_{(max)} = \frac{P_{ac}}{P_{dc}} = \frac{2V_{CC}\,2I_C}{8V_{CC}I_C} \times 100\%$$

An output transformer improves the efficiency of the amplifier by matching the impedance of the load with that of the amplifiers output impedance. By using an output or signal transformer with a suitable turns ratio, class-A amplifier efficiencies reaching 40% are possible with most commercially available Class-A type power amplifiers being of this type of configuration.

However, the transformer is an inductive device due to its windings and core so the use of inductive components in amplifier switching circuits is best avoided as any back emf’s generated may damage the transistor without adequate protection. Also another big disadvantage of this type of transformer coupled class A amplifier circuit is the additional cost and size of the audio transformer required.

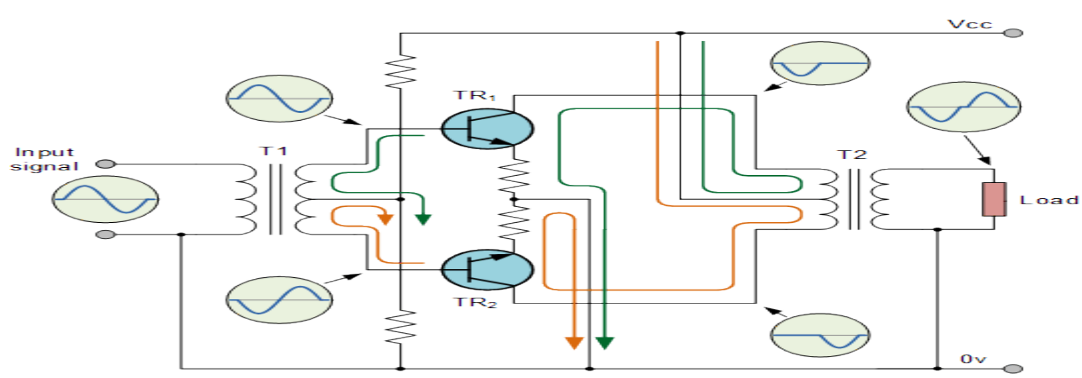
The type of "Class" or classification that an amplifier is given really depends upon the conduction angle, the portion of the 360° of the input waveform cycle, in which the transistor is conducting. In the Class A amplifier the conduction angle is a full 360° or 100% of the input signal while in other amplifier classes the transistor conducts during a lesser conduction angle.

It is possible to obtain greater power output and efficiency than that of the **Class A amplifier** by using two complementary transistors in the output stage with one transistor being an NPN or N-channel type while the other transistor is a PNP or P-channel (the complement) type connected in what is called a "push-pull" configuration. This type of configuration is generally called a **Class B Amplifier** and is another type of audio amplifier circuit that we will look at in the next tutorial.

Push-pull amplifiers use two "complementary" or matching transistors, one being an NPN-type and the other being a PNP-type with both power transistors receiving the same input signal together that is equal in magnitude, but in opposite phase to each other. This results in one transistor only amplifying one half or 180° of the input waveform cycle while the other transistor amplifies the other half or remaining 180° of the input waveform cycle with the resulting "two-halves" being put back together again at the output terminal.

Then the conduction angle for this type of amplifier circuit is only 180° or 50% of the input signal. This pushing and pulling effect of the alternating half cycles by the transistors gives this type of circuit its amusing "push-pull" name, but are more generally known as the **Class B Amplifier** as shown below.

Class B Push-pull Transformer Amplifier Circuit



The circuit above shows a standard **Class B Amplifier** circuit that uses a balanced center-tapped input transformer, which splits the incoming waveform signal into two equal halves and which are 180° out of phase with each other.

Another center-tapped transformer on the output is used to recombined the two signals providing the increased power to the load. The transistors used for this type of transformer push-pull amplifier circuit are both NPN transistors with their emitter terminals connected together.

Here, the load current is shared between the two power transistor devices as it decreases in one device and increases in the other throughout the signal cycle reducing the output voltage and current to zero.

The result is that both halves of the output waveform now swings from zero to twice the quiescent current thereby reducing dissipation. This has the effect of almost doubling the efficiency of the amplifier to around 70%.

Assuming that no input signal is present, then each transistor carries the normal quiescent collector current, the value of which is determined by the base bias which is at the cut-off point.

If the transformer is accurately center tapped, then the two collector currents will flow in opposite directions (ideal condition) and there will be no magnetization of the transformer core, thus minimizing the possibility of distortion.

When an input signal is present across the secondary of the driver transformer T1, the transistor base inputs are in “anti-phase” to each other as shown, thus if TR1 base goes positive driving the transistor into heavy conduction, its collector current will increase but at the same time the base current of TR2 will go negative further into cut-off and the collector current of this transistor decreases by an equal amount and vice versa.

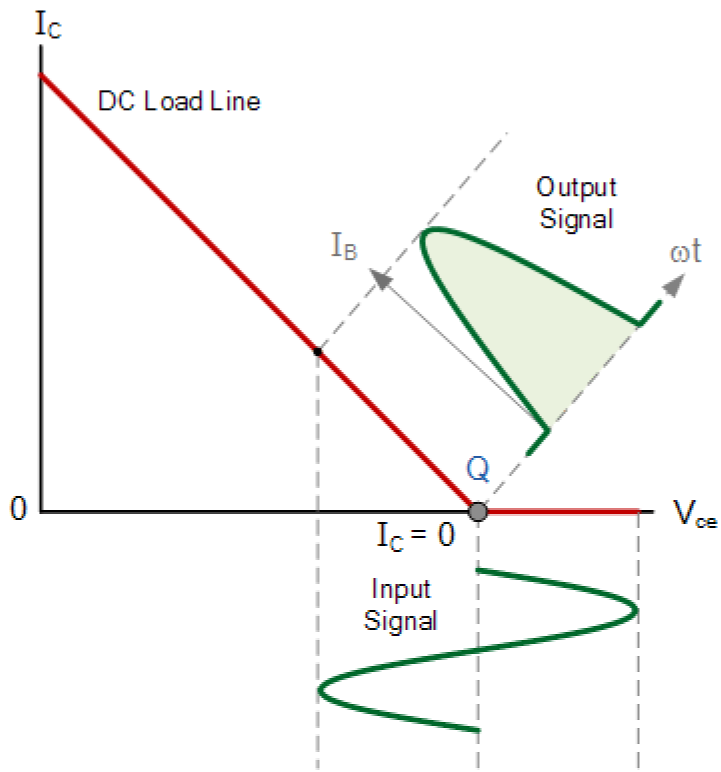
Hence negative halves are amplified by one transistor and positive halves by the other transistor giving this push-pull effect.

Unlike the DC condition, these alternating currents are **ADDITIVE** resulting in the two output half-cycles being combined to reform the sine-wave in the output transformers primary winding which then appears across the load.

Class B Amplifier operation has zero DC bias as the transistors are biased at the cut-off, so each transistor only conducts when the input signal is greater than the base-emitter voltage.

Therefore, at zero input there is zero output and no power is being consumed. This then means that the actual Q-point of a Class B amplifier is on the Vce part of the load line as shown below.

Class B Output Characteristics Curves



The **Class B Amplifier** has the big advantage over their Class A amplifier cousins in that no current flows through the transistors when they are in their quiescent state (ie, with no input signal), therefore no power is dissipated in the output transistors or transformer when there is no signal present unlike Class A amplifier stages that require significant base bias thereby dissipating lots of heat – even with no input signal present.

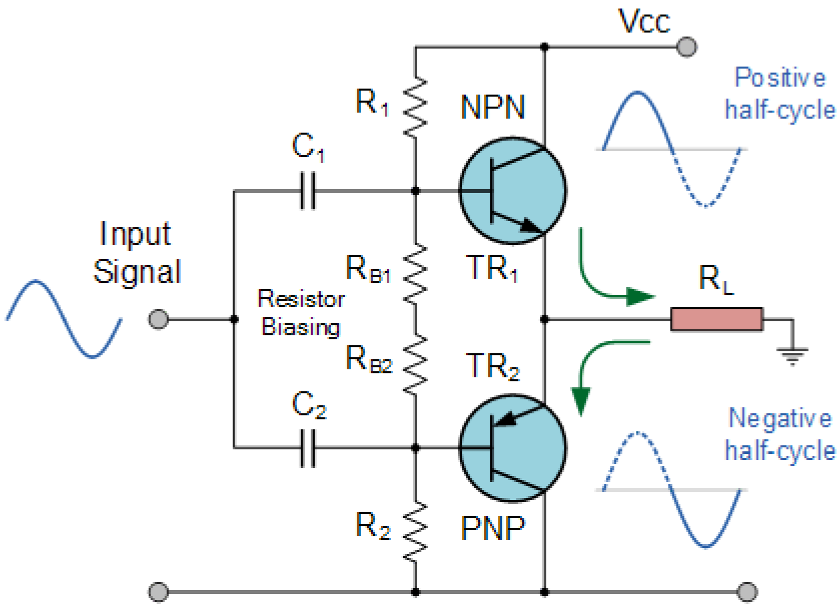
So the overall conversion efficiency (η) of the amplifier is greater than that of the equivalent Class A with efficiencies reaching as high as 70% possible resulting in nearly all modern types of push-pull amplifiers operated in this Class B mode.

Complementary symmetry push pull amplifier

One of the main disadvantages of the Class B amplifier circuit above is that it uses balanced center-tapped transformers in its design, making it expensive to construct. However, there is another type of Class B amplifier called a **Complementary-Symmetry Class B Amplifier** that does not use transformers in its design therefore, it is transformer less using instead complementary or matching pairs of power transistors.

As transformers are not needed this makes the amplifier circuit much smaller for the same amount of output, also there are no stray magnetic effects or transformer distortion to effect the quality of the output signal. An example of a "transformer less" Class B amplifier circuit is given below.

Class B Transformer less Output Stage



The Class B amplifier circuit above uses complimentary transistors for each half of the waveform and while Class B amplifiers have a much high gain than the Class A types, one of the main disadvantages of class B type push-pull amplifiers is that they suffer from an effect known commonly as Crossover Distortion.

Hopefully we remember from our tutorials about Transistors that it takes approximately 0.7 volts (measured from base to emitter) to get a bipolar transistor to start conducting. In a pure class B amplifier, the output transistors are not "pre-biased" to an "ON" state of operation.

This means that the part of the output waveform which falls below this 0.7 volt window will not be reproduced accurately as the transition between the two transistors (when they are switching over from one transistor to the other), the transistors do not stop or start conducting exactly at the zero crossover point even if they are specially matched pairs.

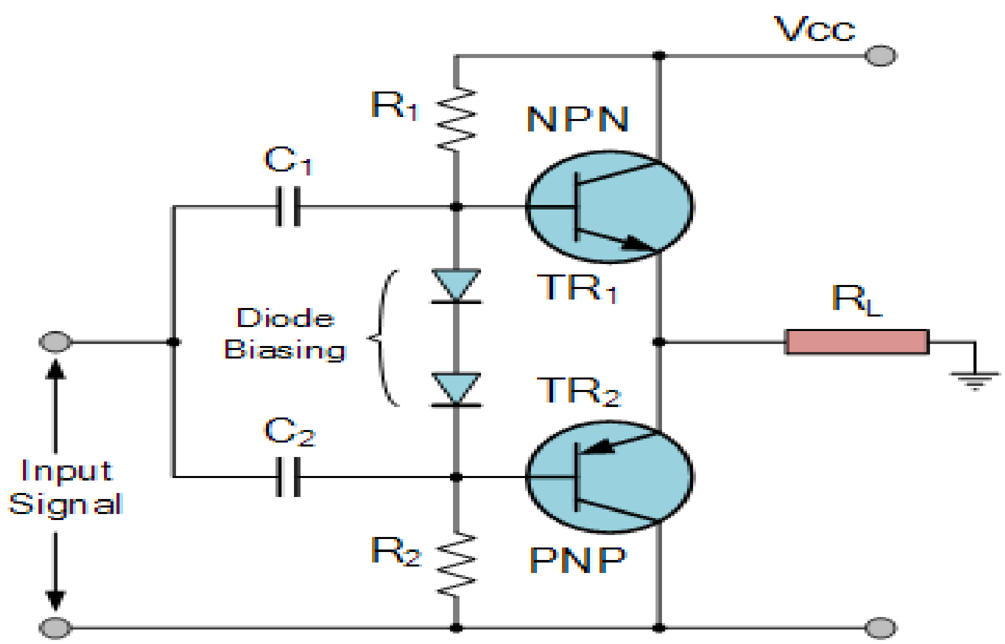
The output transistors for each half of the waveform (positive and negative) will each have a 0.7 volt area in which they are not conducting. The result is that both transistors are turned "OFF" at exactly the same time.

A simple way to eliminate crossover distortion in a Class B amplifier is to add two small voltage sources to the circuit to bias both the transistors at a point slightly above their cut-off point. This then would give us what is commonly called an**Class AB Amplifier** circuit. However, it is impractical to add additional voltage sources to the amplifier circuit so PN-junctions are used to provide the additional bias in the form of silicon diodes**.**

The Class AB Amplifier

We know that we need the base-emitter voltage to be greater than 0.7v for a silicon bipolar transistor to start conducting, so if we were to replace the two voltage divider biasing resistors connected to the base terminals of the transistors with two silicon Diodes, the biasing voltage applied to the transistors would now be equal to the forward voltage drop of the diode. These two diodes are generally called **Biasing Diodes** or **Compensating Diodes** and are chosen to match the characteristics of the matching transistors. The circuit below shows diode biasing.

Class AB Amplifier



The **Class AB Amplifier** circuit is a compromise between the Class A and the Class B configurations. This very small diode biasing voltage causes both transistors to slightly conduct even when no input signal is present. An input signal waveform will cause the transistors to operate as normal in their active region thereby eliminating any crossover distortion present in pure Class B amplifier designs.

A small collector current will flow when there is no input signal but it is much less than that for the Class A amplifier configuration. This means then that the transistor will be “ON” for more than half a cycle of the waveform but much less than a full cycle giving a conduction angle of between 180 to 360^o or 50 to 100% of the input signal depending upon the amount of additional biasing used. The amount of diode biasing voltage present at the base terminal of the transistor can be increased in multiples by adding additional diodes in series.

Class B amplifiers are greatly preferred over Class A designs for high-power applications such as audio power amplifiers and PA systems. Like the class-A amplifier circuit, one way to greatly boost the current gain (A_i) of a Class B push-pull amplifier is to use Darlington transistors pairs instead of single transistors in its output circuitry.

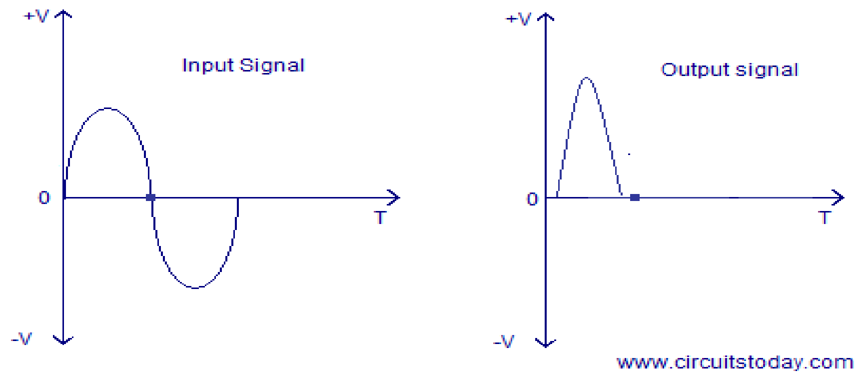
In the next tutorial about Amplifiers we will look more closely at the effects of Crossover Distortion in Class B amplifier circuits and ways to reduce its effect.

But we also know that we can improve the amplifier and almost double its efficiency simply by changing the output stage of the amplifier to a Class B push-pull type configuration. However, this is great from an efficiency point of view, but most modern Class B amplifiers are transformerless or complementary types with two transistors in their output stage.

Class C power amplifier.

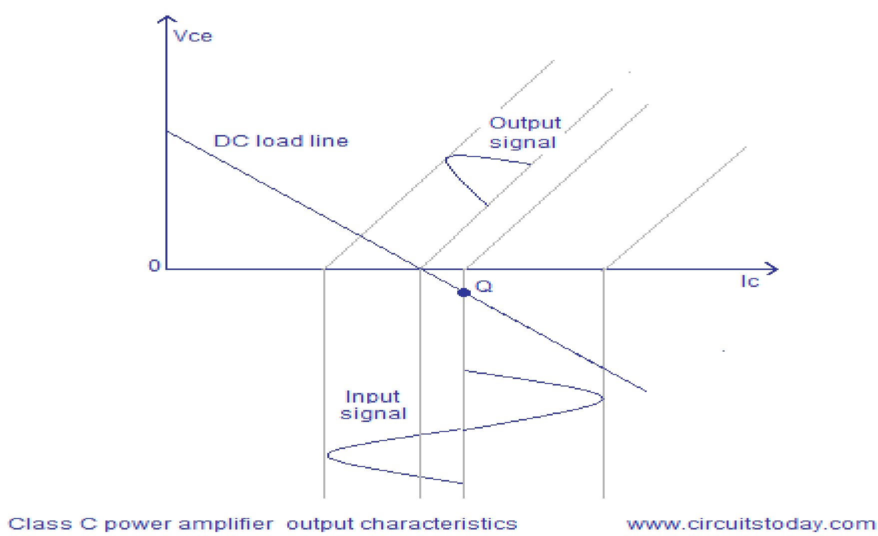
Class C power amplifier is a type of amplifier where the active element (transistor) conduct for less than one half cycle of the input signal. Less than one half cycle means the conduction angle is less than 180° and its typical value is 80° to 120°. The reduced conduction angle improves the efficiency to a great extend but causes a lot of distortion. Theoretical maximum efficiency of a Class C amplifier is around 90%.

Due to the huge amounts of distortion, the Class C configurations are not used in audio applications. The most common application of the Class C amplifier is the RF (radio frequency) circuits like RF oscillator, RF amplifier etc where there are additional tuned circuits for retrieving the original input signal from the pulsed output of the Class C amplifier and so the distortion caused by the amplifier has little effect on the final output. Input and output waveforms of a typical Class C power amplifier are shown in the figure below.



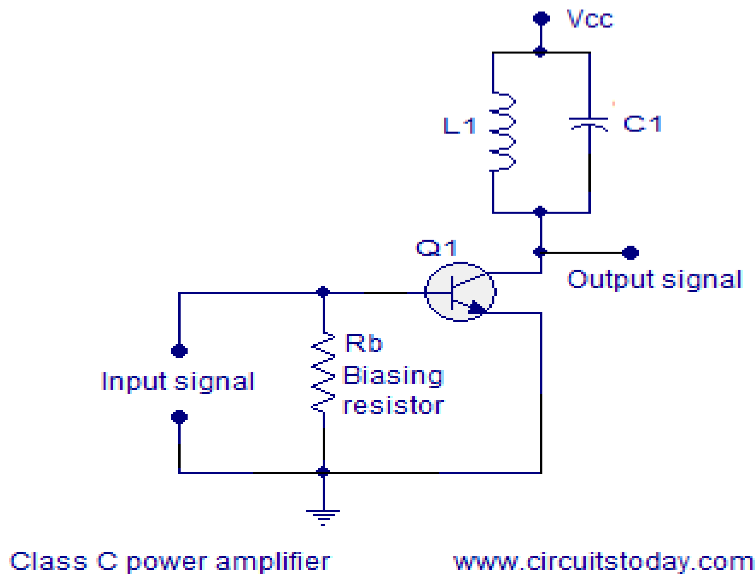
From the above figure it is clear that more than half of the input signal is missing in the output and the output is in the form of some sort of a pulse.

Output characteristics of Class C power amplifier



In the above figure you can see that the operating point is placed some way below the cut-off point in the DC load-line and so only a fraction of the input waveform is available at the output.

Class C power amplifier circuit diagram.



Biasing resistor Rb pulls the base of Q1 further downwards and the Q-point will be set some way below the cut-off point in the DC load line. As a result the transistor will start conducting only after the input signal amplitude has risen above the base emitter voltage (Vbe~0.7V) plus the downward bias voltage caused by Rb. That is the reason why the major portion of the input signal is absent in the output signal.

Inductor L1 and capacitor C1 forms a tank circuit which aids in the extraction of the required signal from the pulsed output of the transistor. Actual job of the active element (transistor) here is to produce a series of current pulses according to the input and make it flow through the resonant circuit. Values of L1 and C1 are so selected that the resonant circuit oscillates in the frequency of the input signal. Since the resonant circuit oscillates in one frequency (generally the carrier frequency) all other frequencies are attenuated and the required frequency can be squeezed out using a suitably tuned load. Harmonics or noise present in the output signal can be eliminated using additional filters. A coupling transformer can be used for transferring the power to the load.

Advantages of Class C power amplifier.

- High efficiency.
- Excellent in RF applications.
- Lowest physical size for a given power output.

Disadvantages of Class C power amplifier.

- Lowest linearity.
- Not suitable in audio applications.
- Creates a lot of RF interference.
- It is difficult to obtain ideal inductors and coupling transformers.
- Reduced dynamic range.

Applications of Class C power amplifier.

- RF oscillators.
- RF amplifier.
- FM transmitters.
- Booster amplifiers.
- High frequency repeaters.
- Tuned amplifiers etc.

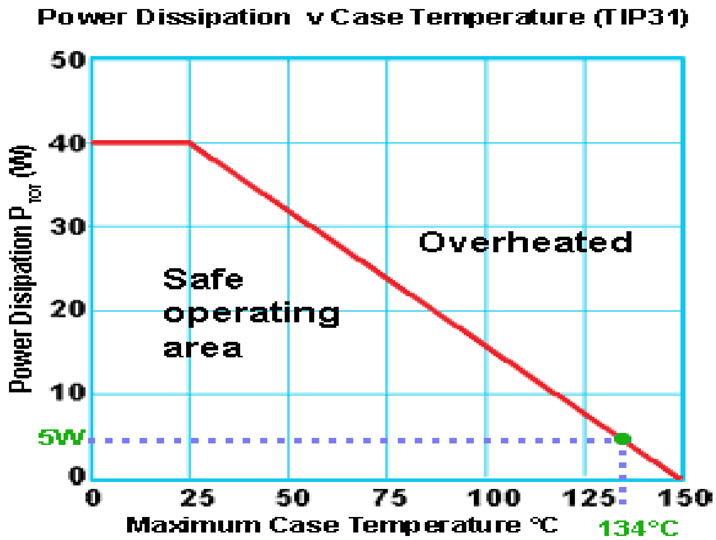
There is not a clear cut difference between ‘ordinary’ transistors used in voltage amplifiers and power transistors, but generally Power transistors can be categorized as those than can handle more than 1 Ampere of collector (or Drain in the case of FETs) current. Because power transistors, such as those voltages, they have a different construction to small signal devices. They must have low output resistances so that they can deliver large currents to the load, and good junction insulation to withstand high voltages. They must also be able to dissipate heat very quickly so they do not overheat. As most heat is generated at the collector/base junction, the area of this junction is made as large as possible.

Power and Temperature

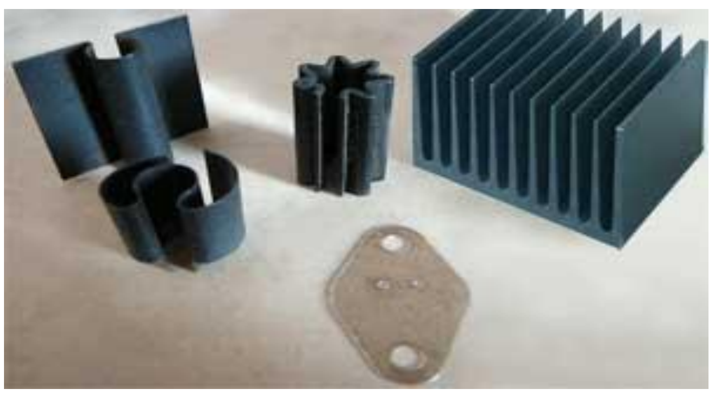
The maximum power rating of a transistor is largely governed by the temperature of the collector/base junction as can be seen from the power de-rating graph in power is dissipated, this junction gets too hot and the transistor will be destroyed, a typical maximum temperature is between 100°C and 150°C, although some devices can withstand higher maximum junction temperatures. The maximum power output available from a power transistor is closely linked to temperature, and above 25°C falls in a linear manner to zero power output as the maximum permissible temperature is reached.

Power De-rating

For example, a transistor such as the TIP31 having a quoted maximum power output PTOT of 40W can only handle 40W of power IF the case temperature (slightly less than the junction temperature) is kept below 25°C. The performance of a power transistor is closely dependent on its ability to dissipate the heat generated at the collector base junction. Minimizing the problem of heat is approached in two main ways:



1. By operating the transistor in the most efficient way possible, that is by choosing a class of biasing that gives high efficiency and is least wasteful of power.



2. By ensuring that the heat produced by the transistor can be removed and effectively transferred to the surrounding air as quickly as possible. Method 2 above highlights the importance of the relationship between a power transistor and its heat sink, a device attached to the transistor for the purpose of removing heat. The physical construction of power transistors is therefore designed to maximize the transfer of heat to the heat sink. In addition to the usual collector lead-out wire, the collector of a power transistor, which has a much larger area than that of a small signal transistor, is normally in direct contact with the metal case of the transistor, or a metal mounting pad, which may then be bolted or clipped directly on to a heat-sink. Typical metal cased and metal body power transistors are Because power amplifiers generate substantial

amounts of heat, which is wasted power, they are made to be as efficient as possible. With voltage amplifiers, low distortion is of greater importance than efficiency, but with power amplifiers, although distortion cannot be ignored, efficiency is vital.

Calculating the Required Thermal Resistance R_{th} for a Heat-sink

The heat-sink chosen must be able to dissipate heat from the transistor to the surrounding air, quickly enough to prevent the junction temperature of the transistor exceeding its maximum permitted value (usually quoted on the transistor's data sheet), typically 100 to 150°C.

Each heat-sink has a parameter called its Thermal Resistance (R_{th}) measured in °C/Watt and the lower the value of R_{th} the faster heat is dissipated. Other factors affecting heat dissipation include the power (in Watts) being dissipated by the transistor, the efficiency of heat transfer between the internal transistor junction and the transistor case, and the case to the heat-sink.

The difference between the temperature of the heat-sink and the air temperature surrounding the heat-sink (the ambient temperature) must also be taken into account. The main criteria is that the heat-sink should be efficient enough, too efficient is not a problem.

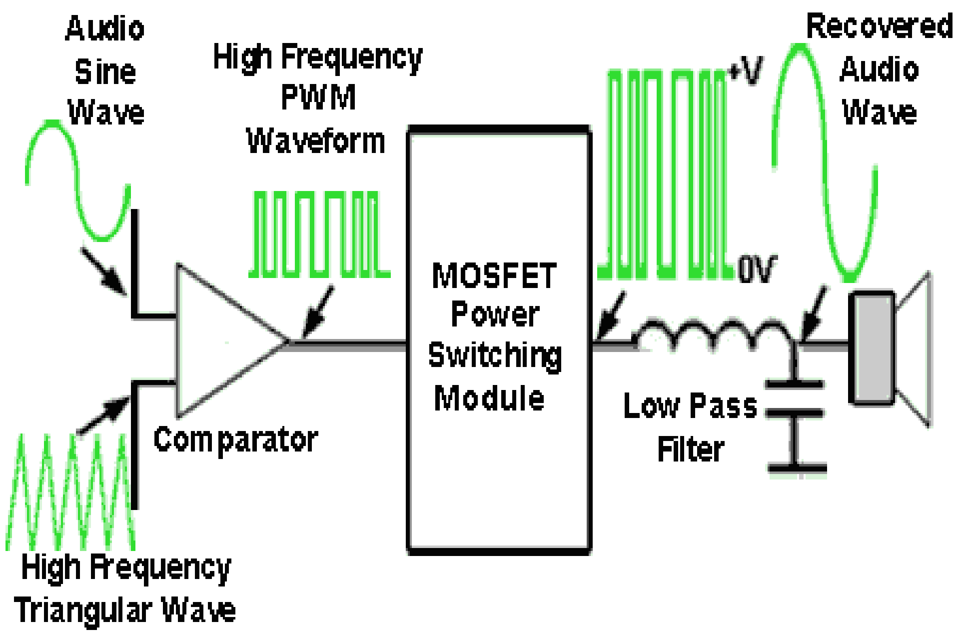
Therefore, any heat-sink with a thermal resistance lower or equal to the calculated value should be OK, but to avoid continually running the transistor at, or close to the maximum permitted temperature, which is almost guaranteed to shorten the life of the transistor, it is advisable to use a heat-sink with a lower thermal resistance where possible.

The power de-rating graph for a TIP31 transistor shown in Fig illustrates the relationship between the power dissipated by the transistor and the case temperature. When the transistor is dissipating 5W, it can be estimated from the graph that the maximum safe case temperature, for a junction temperature of 150°C would be about 134 to 135°C, confirming the above calculation of max. case temperature.

The TIP31 transistor has maximum power dissipation P_{TOT} of 40W but it can be seen from the graph in Fig that this is only attainable if the case temperature of the transistor can be held at 25°C. The case temperature can only be allowed to rise to 150°C (the same as the maximum junction temperature) if the power dissipation is zero.

The class A Common Emitter Voltage Amplifier described in Amplifier Module 1, Module 2 and Module 3 has some excellent properties that make it useful for many amplification tasks; however it is not suitable for every purpose. Class A biasing is good at preserving the original wave shape as the transistor is biased using the most linear part of the transistor's characteristics. However the big problem with class A is its poor efficiency. Amplifiers Module explains how push-pull class B power amplifiers improve efficiency at the expense of added crossover distortion.

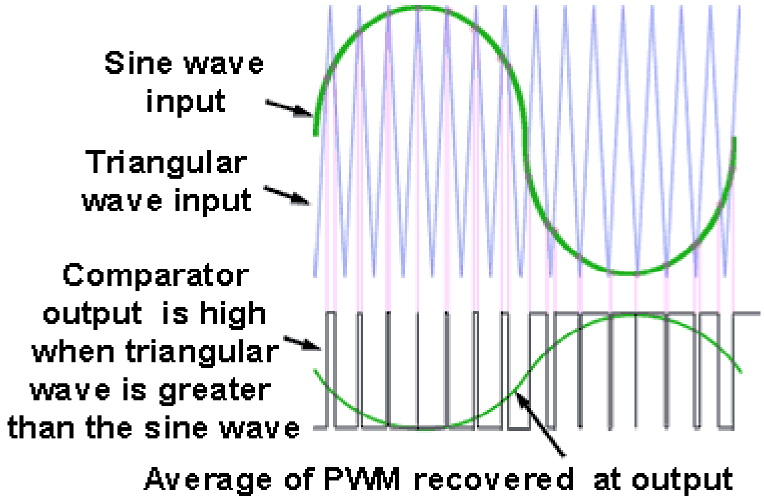
Class D operation makes the output circuit extremely efficient (around 90%) allowing high power output without the need for such high power transistors and elaborate heat-sinks. However this big increase in efficiency is only achieved at the expense of some increase in distortion and especially of noise, in the form of electromagnetic interference (EMI).



Nevertheless, class D is a very efficient class of amplifier suited to both high power audio and RF amplifiers and low power portable amplifiers, where battery life can be considerably extended because of the amplifier's high efficiency. The increased interest in class D amplifiers has led to a number of class D integrated circuits becoming available.

Class E and F Power Amplifiers:

Amplifier classes such as E and F are basically enhancements of class D, offering more complex and improved output filtering, including some additional wave shaping of the PWM signal to prevent audio distortion.



Class G and H Power Amplifiers:

Classes G and H offer enhancements to the basic class AB design.

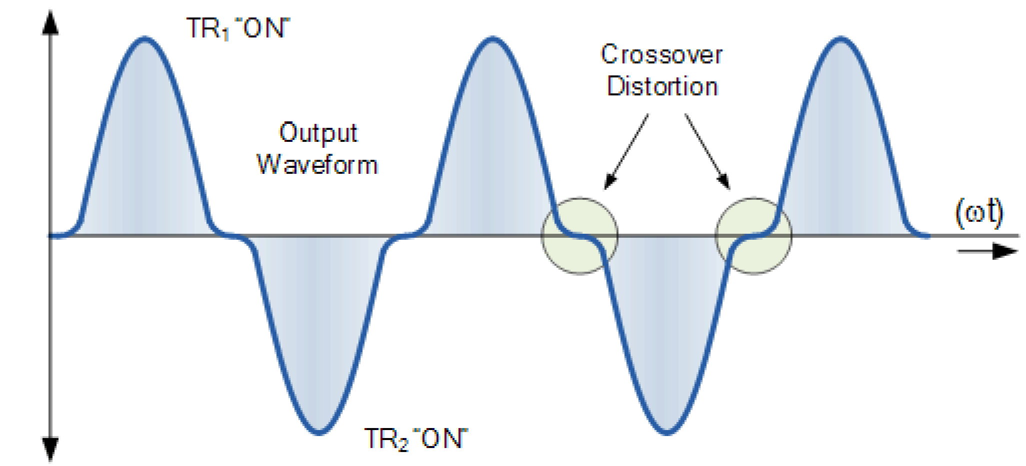
Class G uses multiple power supply rails of various voltages, rapidly switching to a higher voltage when the audio signal wave has a peak value that is a higher voltage than the level of supply voltage, and switching back to a lower supply voltage when the peak value of the audio signal reduces. By switching the supply voltage to a higher level only when the largest output signals are present and then switching back to a lower level, average power consumption, and therefore heat caused by wasted power is reduced.

Class H improves on class G by continually varying the supply voltage at any time where the audio signal exceeds a particular threshold level. The power supply voltage tracks the peak level of the signal to be only slightly higher than the instantaneous value of the audio wave, returning to its lower level once the signal peak value falls below the threshold level again. Both classes G and H therefore require considerably more complex power supplies, which add to the cost of implementing these features.

Crossover Distortion

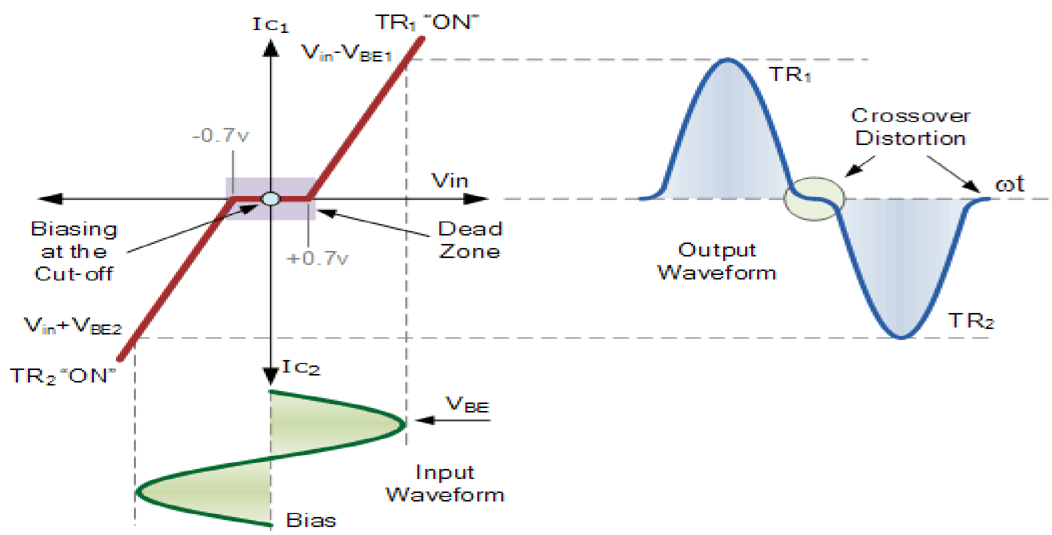
It produces a zero voltage "flat spot" or "deadband" on the output wave shape as it crosses over from one half of the waveform to the other. The reason for this is that the transition period when the transistors are switching over from one to the other, does not stop or start exactly at the zero crossover point thus causing a small delay between the first transistor turning "OFF" and the second transistor turning "ON". This delay results in both transistors being switched "OFF" at the same instant in time producing an output wave shape as shown below.

Crossover Distortion Waveform:



In order that there should be no distortion of the output waveform we must assume that each transistor starts conducting when its base to emitter voltage rises just above zero, but we know that this is not true because for silicon bipolar transistors the base voltage must reach at least 0.7v before the transistor starts to conduct thereby producing this flat spot. This crossover distortion effect also reduces the overall peak to peak value of the output waveform causing the maximum power output to be reduced as shown below.

Non-Linear Transfer Characteristics

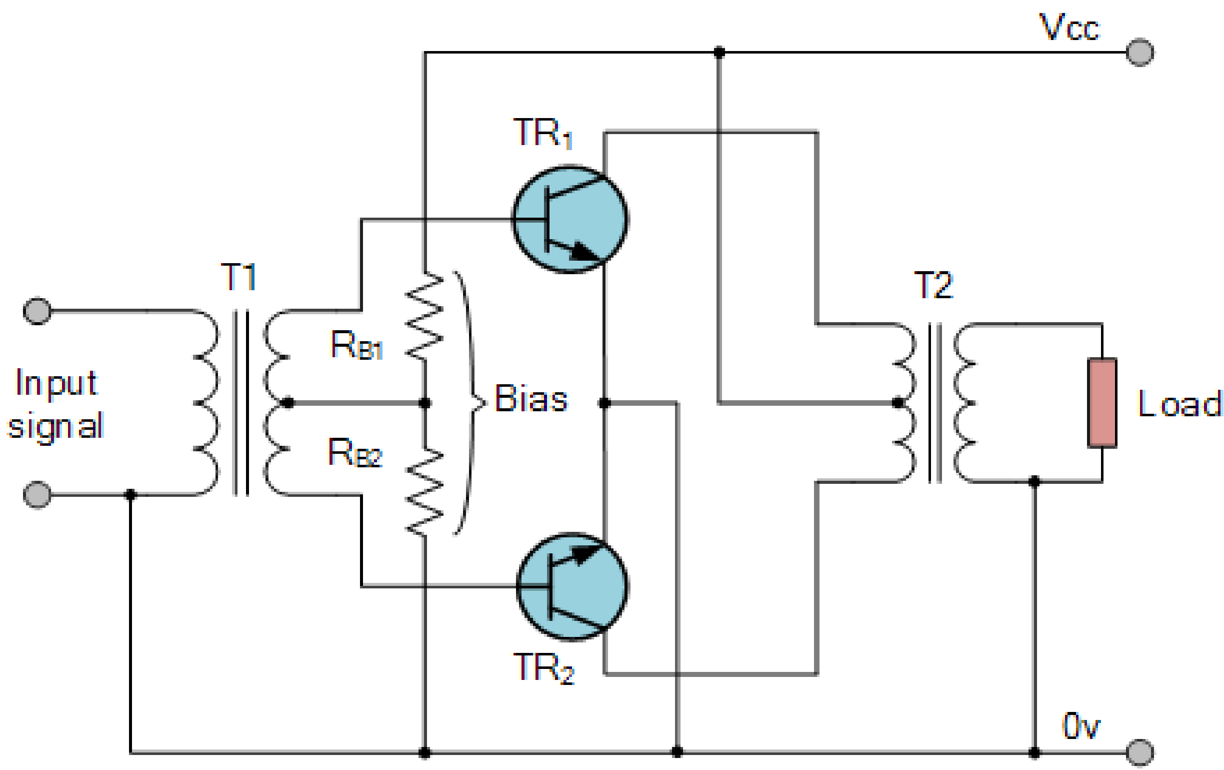


This effect is less pronounced for large input signals as the input voltage is usually quite large but for smaller input signals it can be more severe causing audio distortion to the amplifier.

Pre-biasing the Output

The problem of **Crossover Distortion** can be reduced considerably by applying a slight forward base bias voltage (same idea as seen in the Transistor tutorial) to the bases of the two transistors via the center-tap of the input transformer, thus the transistors are no longer biased at the zero cut-off point but instead are "Pre-biased" at a level determined by this new biasing voltage.

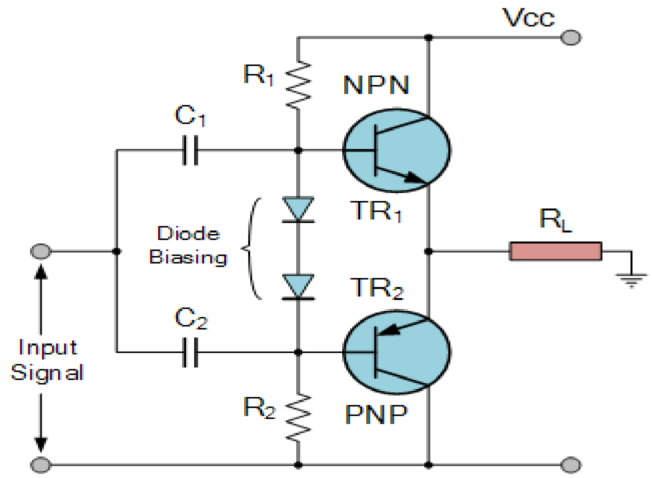
Push-pull Amplifier with Pre-biasing



This type of resistor pre-biasing causes one transistor to turn "ON" exactly at the same time as the other transistor turns "OFF" as both transistors are now biased slightly above their original cut-off point. However, to achieve this the bias voltage must be at least twice that of the normal base to emitter voltage to turn "ON" the transistors.

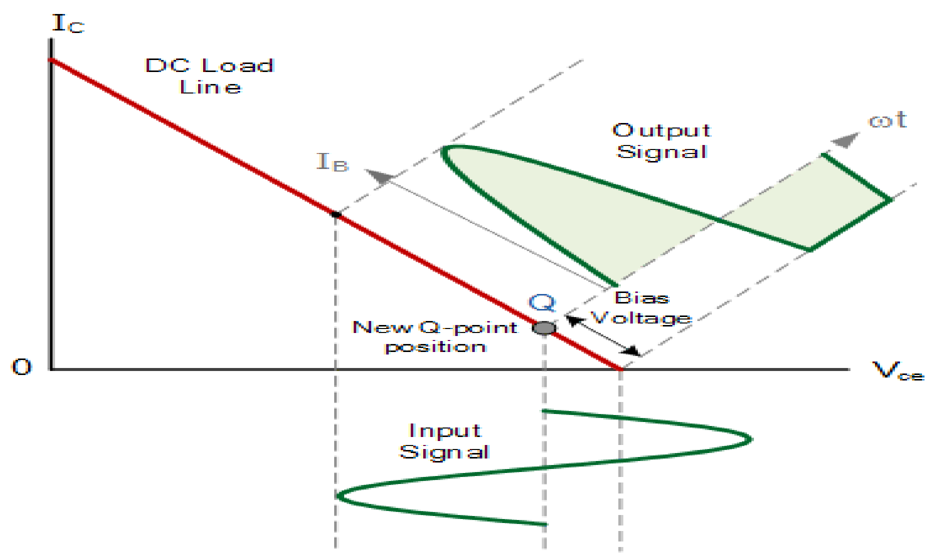
This pre-biasing can also be implemented in transformerless amplifiers that use complementary transistors by simply replacing the two potential divider resistors with **Biasing Diodes** as shown below.

Pre-biasing with Diodes



This pre-biasing voltage either for a transformer or transformer less amplifier circuit, has the effect of moving the amplifiers Q-point past the original cut-off point thus allowing each transistor to operate within its active region for slightly more than half or 180° of each half cycle. In other words 180° + Bias. The amount of diode biasing voltage present at the base terminal of the transistor can be increased in multiples by adding additional diodes in series. This then produces an amplifier circuit commonly called a **Class AB Amplifier** and its biasing arrangement is given below.

Class AB Output Characteristics



Crossover Distortion Summary

Then to summarize, **Crossover Distortion** occurs in Class B amplifiers because the amplifier is biased at its cut-off point. This then results in BOTH transistors being switched “OFF” at the same instant in time as the waveform crosses the zero axis. By applying a small base bias voltage either by using a resistive potential divider circuit or diode biasing this crossover distortion can be greatly reduced or even eliminated completely by bringing the transistors to the point of being just switched “ON”.

The application of a biasing voltage produces another type or class of amplifier circuit commonly called a **Class AB Amplifier**. Then the difference between a pure Class B amplifier and an improved Class AB amplifier is in the biasing level applied to the output transistors. One major advantage of using diodes over resistors is that the PN-junctions compensate for variations in the temperature of the transistors. Therefore, we can say the a Class AB amplifier is a Class B amplifier with “Bias” and we can therefore summaries as:

- Class A Amplifiers – No Crossover Distortion as they are biased in the center of the load line.
- Class B Amplifiers – Large amounts of Crossover Distortion due to biasing at the cut-off point.
- Class AB Amplifiers – Some Crossover Distortion if the biasing level is set too low.

As well as the three amplifier classes above, there are a number of high efficiency Amplifier Classes relating to switching amplifier designs that use different switching techniques to reduce power loss and increase efficiency. Some of these amplifier designs use RLC resonators or multiple power-supply voltages to help reduce power loss and distortion.

UNIT-VI

Tuned Amplifiers : Introduction, Q-Factor, small signal tuned amplifier, capacitance single tuned amplifier, double tuned amplifiers, effect of cascading single tuned amplifiers on band width, effect of cascading double tuned amplifiers on band width, staggered tuned amplifiers, stability of tuned amplifiers, wideband amplifiers.

Introduction:

Most of the audio amplifiers we have discussed in the earlier chapters will also work at radio frequencies *i.e*. above 50 kHz. However, they suffer from two major drawbacks. First, they become less efficient at radio frequency. Secondly, such amplifiers have mostly resistive loads and consequently their gain is independent of signal frequency over a large bandwidth.

In other words, an audio amplifier amplifies a wide band of frequencies equally well and does not permit the selection of a particular desired frequency while rejecting all other frequencies. However, sometimes it is desired that an amplifier should be selective *i.e*. it should select a desired frequency or narrow band of frequencies for amplification.

For instance, radio and television transmission are carried on a specific radio frequency assigned to the broadcasting station. The radio receiver is required to pick up and amplify the radio frequency desired while discriminating all others. To achieve this, the simple resistive load is replaced by a parallel tuned circuit whose impedance strongly depends upon frequency. Such a tuned circuit becomes very selective and amplifies very strongly signals of resonant frequency and narrow band on either side.

Therefore, the use of tuned circuits in conjunction with a transistor makes possible the selection and efficient amplification of a particular desired radio frequency. Such an amplifier is called a tuned amplifier. In this chapter, we shall focus our attention on transistor tuned amplifiers and their increasing applications in high frequency electronic circuits.

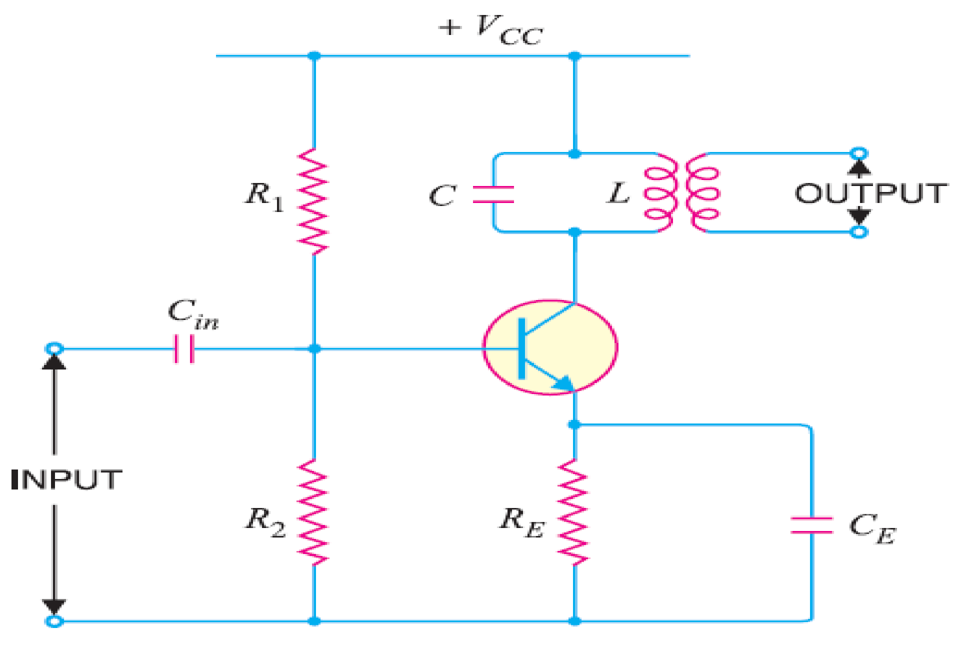
Amplifiers which amplify a specific frequency or narrow band of frequencies are called **tuned amplifiers.**

Tuned amplifiers are mostly used for the amplification of high or radio frequencies. It is because radio frequencies are generally single and the tuned circuit permits their selection and efficient amplification.

However, such amplifiers are not suitable for the amplification of audio frequencies as they are mixture of frequencies from 20 Hz to 20 kHz and not single. Tuned amplifiers are widely used in radio and television circuits where they are called upon to handle radio frequencies. Figure shows the circuit of a simple transistor tuned amplifier. Here, instead of load resistor, we have a parallel tuned circuit in the collector.

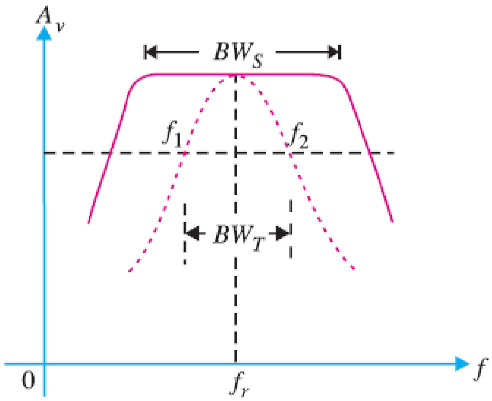
The impedance of this tuned circuit strongly depends upon frequency. It offers a very high impedance at *resonant frequency* and very small impedance at all other frequencies.

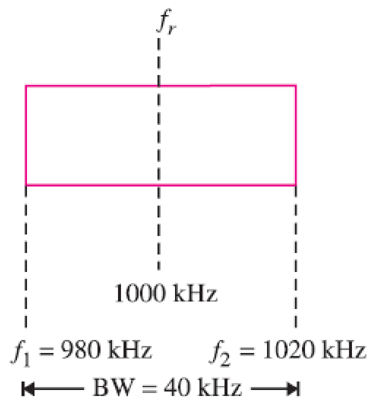
If the signal has the same frequency as the resonant frequency of *LC* circuit, large amplification will result due to high impedance of *LC* circuit at this frequency. When signals of many frequencies are present at the input of tuned amplifier, it will select and strongly amplify the signals of resonant frequency while rejecting all others. Therefore, such amplifiers are very useful in radio receivers to select the signal from one particular broadcasting station when signals of many other frequencies are present at the receiving aerial.



Distinction between Tuned Amplifiers and other Amplifiers:

We have seen that amplifiers (*e.g.*, voltage amplifier, power amplifier *etc.*) provide the constant gain over a limited band of frequencies *i.e.*, from lower cut-off frequency *f*1 to upper cut-off frequency *f*2. Now bandwidth of the amplifier, *BW* = *f*2 −□*f*1. The reader may wonder, then, what distinguishes a tuned amplifier from other mplifiers? The difference is that tuned amplifiers are designed to have specific, usually narrow bandwidth. This point is illustrated in in Fig. 15.2. Note that *BWS* is the bandwidth of standard frequency response while *BWT* is the bandwidth of the tuned amplifier. In many applications, the narrower the bandwidth of a tuned amplifier, the better it is.





Consider a tuned amplifier that is designed to amplify only those frequencies that are within ± 20 kHz of the central frequency of 1000 kHz (*i.e.*, *fr* = 1000 kHz). Here *f*1 = 980 kHz, *fr* = 1000 kHz, *f*2 = 1020 kHz, *BW* = 40 kHz This means that so long as the input signal is within the range of 980 – 1020 kHz, it will be amplified. If the frequency of input signal goes out of this range, amplification will be drastically reduced.

A parallel tuned circuit consists of a capacitor *C* and inductor *L* in parallel as shown in Fig In practice, some resistance *R* is always present with the coil. If an alternating voltage is applied across this parallel circuit, the frequency of oscillations will be that of the applied voltage. However, if the frequency of applied voltage is equal to the natural or resonant frequency of *LC* circuit, then *electrical resonance* will occur. Under such conditions, the impedance of the tuned circuit becomes maximum and the line current is minimum. The circuit then draws just enough energy from a.c. supply necessary to overcome the losses in the resistance *R*.

Parallel resonance: A parallel circuit containing reactive elements (*L* and *C*) is *resonant when the circuit power factor is unity *i.e*. applied voltage and the supply current are in phase. The phasor diagram of the parallel circuit is shown in Fig. The coil current *IL* has two rectangular components *viz* active component *IL* cosφ $_L$ and reactive component *IL* sin φ $_L$. This parallel circuit will resonate when the circuit power factor is unity. This is possible only when the net reactive component of the circuit current is zero *i.e*.

IC □*IL* sin φ $_L$ = 0

or *IC* = *IL* sin φ $_L$

Resonance in parallel circuit can be obtained by changing the supply frequency. At some frequency *fr* (called resonant frequency), *IC* = *IL* sin φ $_L$ and resonance occurs.

Resonant frequency. The frequency at which parallel resonance occurs (*i.e.* reactive component of circuit current becomes zero) is called the *resonant frequency fr*.

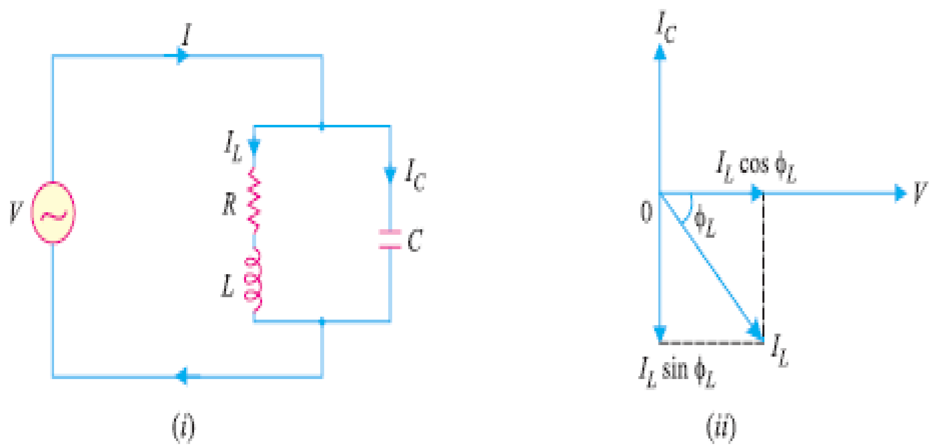


Fig. 15.4

At parallel resonance, we have, $I_C = I_L \sin \phi_L$

Now $\quad I_L = V/Z_L; \quad \sin \phi_L = X_L/Z_L \quad \text{and} \quad I_C = V/X_C$

$$\therefore \quad \frac{V}{X_C} = \frac{V}{Z_L} \times \frac{X_L}{Z_L}$$

$$\text{or} \quad X_L X_C = Z_L^2$$

$$\text{or} \quad \frac{\omega L}{\omega C} = Z_L^2 = R^2 + X_L^2 \qquad ...(i)$$

$$\text{or} \quad \frac{L}{C} = R^2 + (2\pi f_r L)^2$$

$$\text{or} \quad (2\pi f_r L)^2 = \frac{L}{C} - R^2$$

$$\text{or} \quad 2\pi f_r L = \sqrt{\frac{L}{C} - R^2}$$

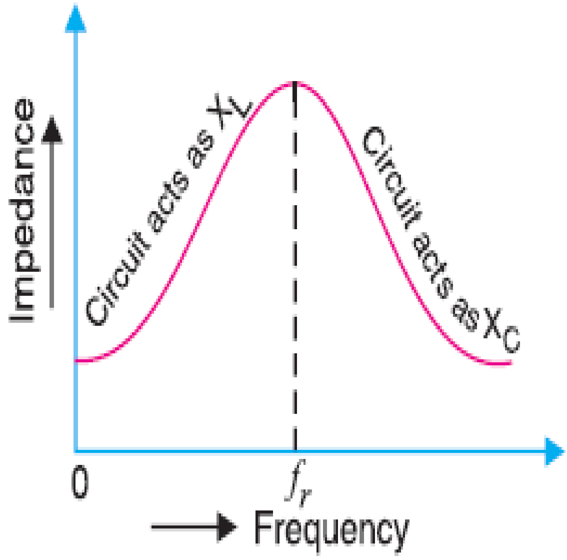
$$\text{or} \quad f_r = \frac{1}{2\pi L}\sqrt{\frac{L}{C} - R^2}$$

$$\therefore \quad \text{Resonant frequency, } f_r = \frac{1}{2\pi}\sqrt{\frac{1}{LC} - \frac{R^2}{L^2}} \qquad ...(ii)$$

If coil resistance R is small (as is generally the case), then,

$$f_r = \frac{1}{2\pi\sqrt{LC}} \qquad ...(iii)$$

The resonant frequency will be in Hz if *R*, *L* and *C* are in ohms, henry and farad respectively.



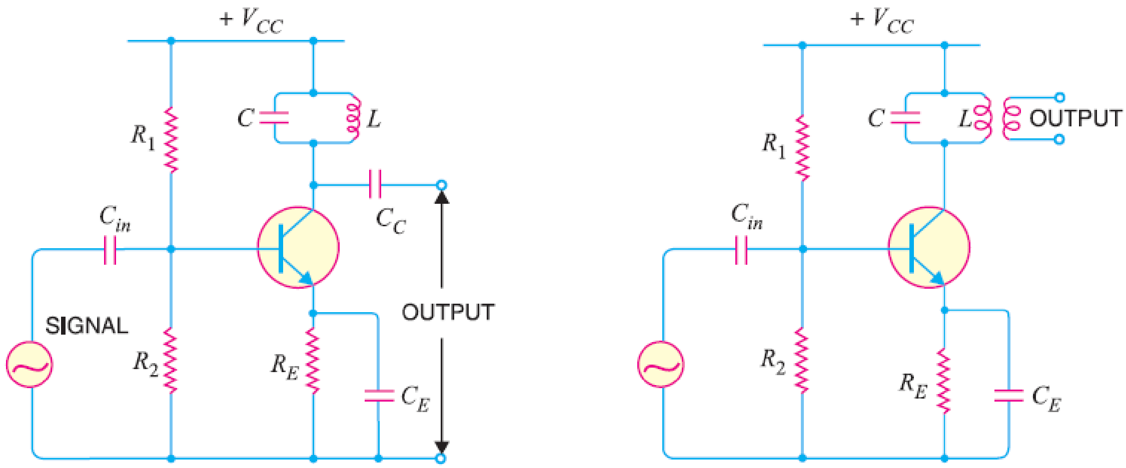
Quality factor Q: It is desired that resonance curve of a parallel tuned circuit should be as sharp as possible in order to provide selectivity. The sharp resonance curve means that impedance falls rapidly as the frequency is varied from the resonant frequency. The smaller the resistance of coil, the more sharp is the resonance curve. This is due to the fact that a small resistance consumes less power and draws a relatively small line current. The ratio of inductive reactance and resistance of the coil at resonance, therefore, becomes a measure of the quality of the tuned circuit. This is called ***quality factor*** and may be defined as under: *The ratio of inductive reactance of the coil at resonance to its resistance is known as* **quality factor Q** *i.e.*,

$$Q = X_L / R = \frac{2\pi f_r L}{R}$$

The quality factor *Q* of a parallel tuned circuit is very important because the sharpness of resonance curve and hence selectivity of the circuit depends upon it. The higher the value of *Q*, the more selective is the tuned circuit. Figure shows the effect of resistance *R* of the coil.

Single Tuned Amplifier

A single tuned amplifier consists of a transistor amplifier containing a parallel tuned circuit as the collector load. The values of capacitance and inductance of the tuned circuit are so selected that its resonant frequency is equal to the frequency to be amplified. The output from a single tuned amplifier can be obtained either (*a*) by a coupling capacitor *CC* as shown in Fig. (*i*) or (*b*) by a secondary coil as shown in Fig. (*ii*).



Operation: The high frequency signal to be amplified is given to the input of the amplifier. The resonant frequency of parallel tuned circuit is made equal to the frequency of the signal by changing the value of *C*. Under such conditions, the tuned circuit will offer very high impedance to the signal frequency. Hence a large output appears across the tuned circuit. In case the input signal is complex containing many frequencies, only that frequency which corresponds to the resonant frequency of the tuned circuit will be amplified. All other frequencies will be rejected by the tuned circuit. In this way, a tuned amplifier selects and amplifies the desired frequency.

Analysis of Tuned Amplifier

Fig. (*i*) Shows a single tuned amplifier. Note the presence of the parallel *LC* circuit in the collector circuit of the transistor. When the circuit has a high *Q*, the parallel resonance occurs at a frequency *fr* given by.

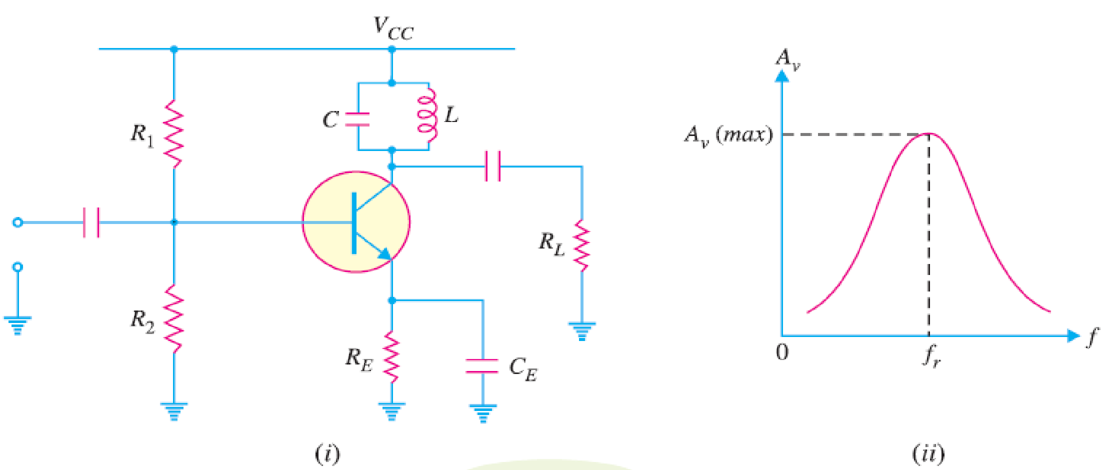


Fig. 15.10

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

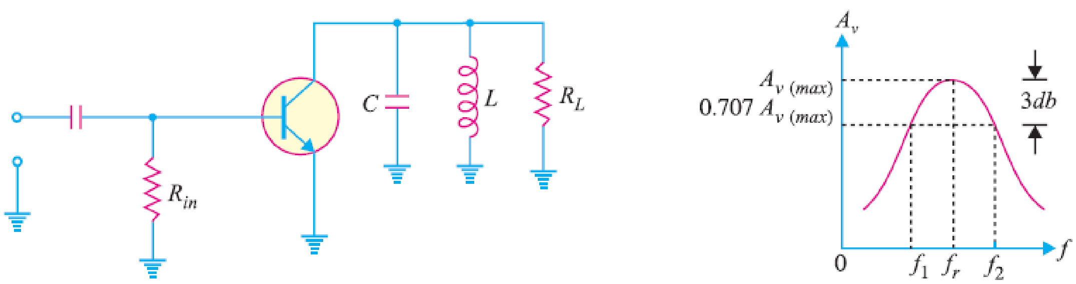
At the resonant frequency, the impedance of the parallel resonant circuit is very high and is purely resistive. Therefore, when the circuit is tuned to resonant frequency, the voltage across *RL* is maximum. In other words, the voltage gain is maximum at *fr*. However, above and below the resonant frequency, the voltage gain decreases rapidly. The higher the *Q* of the circuit, the faster the gain drops off on either side of resonance.

A.C. Equivalent Circuit of Tuned Amplifier

Fig. (*i*) shows the *ac* equivalent circuit of the tuned amplifier. Note the tank circuit components are not shorted. In order to completely understand the operation of this circuit, we shall see its behaviour at three frequency conditions *viz*.,

(*i*) *fin* = *fr* **(*ii*)** *fin* < *fr* **(*iii*)** *fin* > *fr*

(*i*) When input frequency equals *fr* (*i.e.*, *fin* = *fr*). When the frequency of the input signal is equal to *fr*, the parallel *LC* circuit offers a very high impedance *i.e.*, it acts as an open. Since *RL* represents the only path to ground in the collector circuit, all the ac collector current flows through *RL*. Therefore, voltage across *RL* is maximum *i.e.*, the voltage gain is maximum as shown in Fig.ii



(*i*) (*ii*)

***(ii)* When input frequency is less than** *fr* (*i.e.*, *fin* < *fr*). When the input signal frequency is less than *fr*, the circuit is effectively* inductive. As the frequency decreases from *fr*, a point is reached when *XC* −□*XL* = *RL*. When this happens, the voltage gain of the amplifier falls by 3 *db*. In other words, the lower cut-off frequency *f*1 for the circuit occurs when *XC* □*XL* = *RL*.

(*iii*) When input frequency is greater than *fr* (*i.e.*, *fin* > *fr*). When the input signal frequency is greater than *fr*, the circuit is effectively capacitive. As *fin* is increased beyond *fr*, a point is reached when *XL* −□*XC* = *RL*. When this happens, the voltage gain of the amplifier will again fall by 3*db*. In other words, the upper cut-off frequency for the circuit will occur when *XL* □*XC* = *RL*.

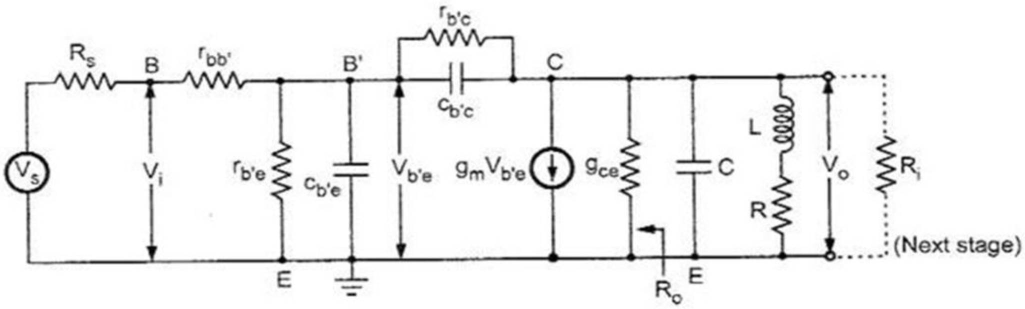


Fig. 3.14 Equivalent circuit of single tuned amplifier

The Fig. 3.14 shows the equivalent circuit for single tuned amplifier using hybrid π 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 $g_m V_{b'e}$. 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 single tuned amplifier.

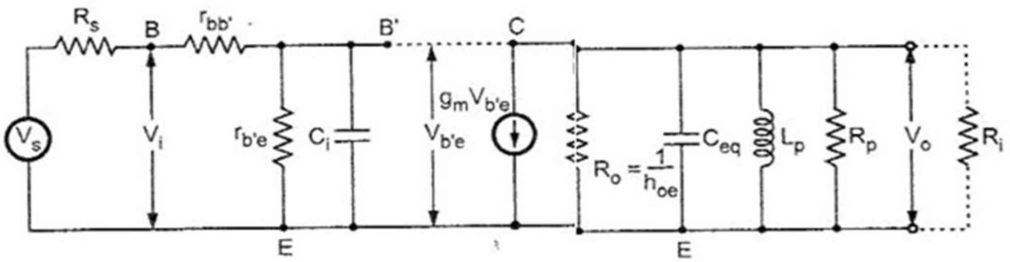


Fig. 3.15 Simplified equivalent circuit for single tuned amplifier

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

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

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

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

$$g_{ce} = \frac{1}{r_{ce}} = h_{oe} - g_m h_{re} \approx h_{oe} = \frac{1}{R_o} \qquad \text{... (3)}$$

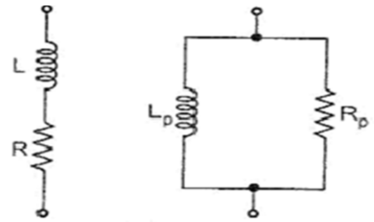


Fig. 3.16

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 $$R_p = \frac{R^2 + \omega^2 L^2}{R} \quad \text{... (4)}$$

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

Centre frequency

The centre frequency or resonant frequency is given as,

$$f_r = \frac{1}{2\pi\sqrt{L_p C_{eq}}} \quad \text{...(6)}$$

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

and $$C_{eq} = C_{b'c}\left(\frac{A-1}{A}\right) + C \quad \text{...(7)}$$

$$= C_o + C$$

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,

$$Q_r = \frac{\omega_r L}{R} \quad \text{...(8)}$$

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 $\omega L >> R$ 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 $\frac{\omega^2 L^2}{R} >> 1$, $R_p \approx \frac{\omega^2 L^2}{R}$ ...(9)

From equation (5) we have,

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

$$\approx L \qquad \because \omega L >> R \qquad \text{... (10)}$$

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

$$R_p = \frac{\omega_r^2 L^2}{R}$$

$$= \omega_r Q_r L \quad \because Q_r = \frac{\omega_r L}{R} \qquad \text{... (11)}$$

Therefore, Q_r can be expressed interms of R_p as,

$$Q_r = \frac{R_p}{\omega_r L} \qquad \text{...(12)}$$

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

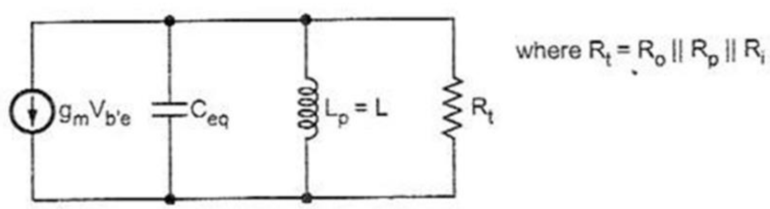


Fig. 3.17 Simplified output circuit for single tuned amplifier

$$\text{Effective quality factor } Q_{eff} = \frac{\text{Susceptance of inductance L or capacitance C}}{\text{Conductance of shunt resistance } R_t}$$

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

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

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

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

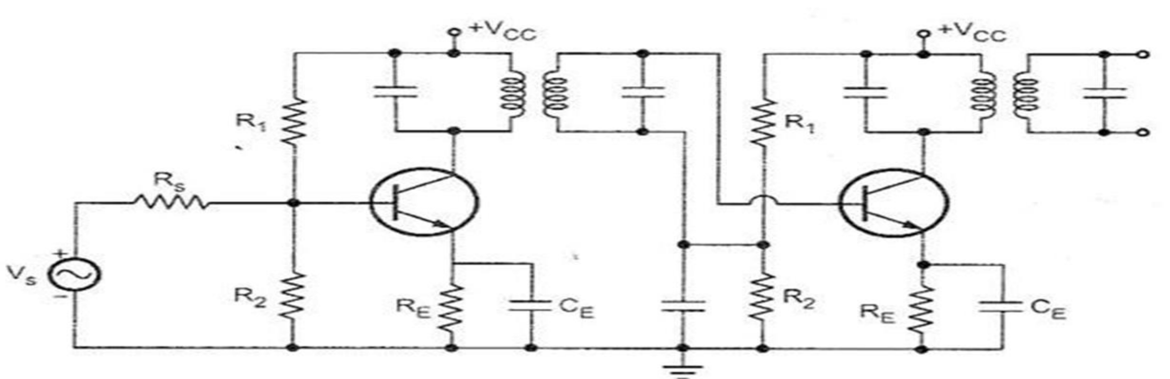
3 dB bandwidth

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

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

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

Below figure shows the 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.



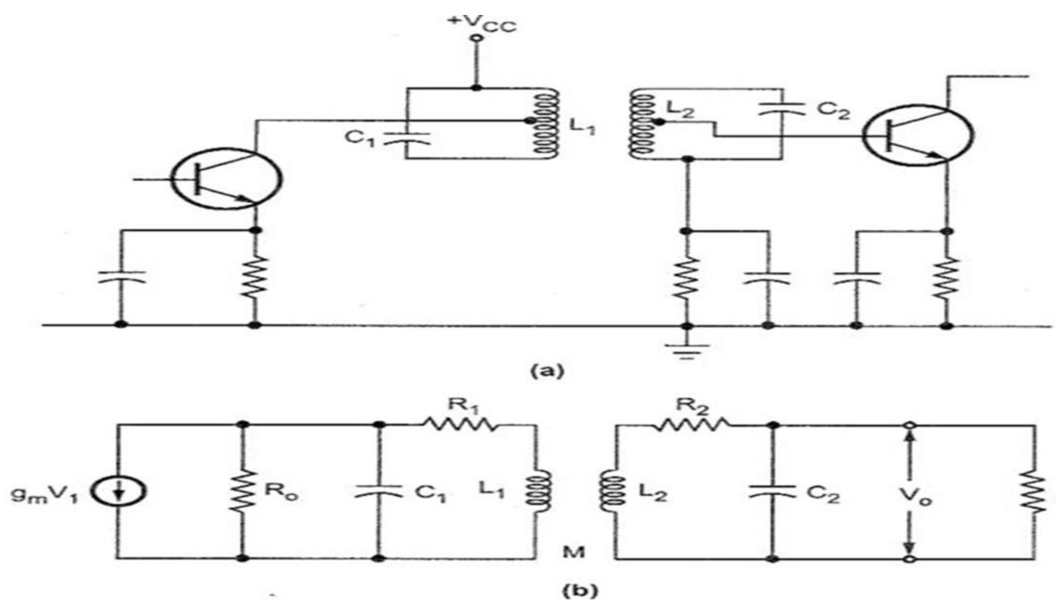
The double tuned circuit can provide a bandwidth of several percent of the resonant frequency and gives steep sides to the response curve.

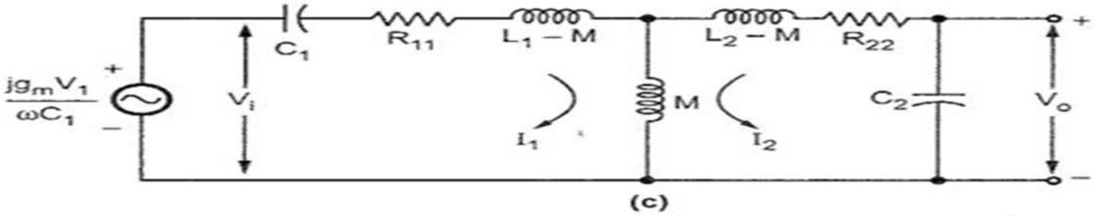
The Fig. 3.19 (a) shows the coupling section of a transformer coupled double tuned amplifier. The Fig. 3.19 (b) shows the equivalent circuit for it. In which transistor is replaced by the current source with its output resistance (R_o). The C_1 and L_1 are the tank circuit components of the primary side.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

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

where R represents series resistance and R_p represents parallel resistance.





(c)

Fig. 3.19 Equivalent circuits for double tuned amplifier

Therefore we can write,

$$R_{11} = \frac{\omega_o^2 L_1^2}{R_o} + R_1$$

$$R_{12} = \frac{\omega_o^2 L_2^2}{R_i} + R_2$$

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, $Q = \frac{\omega_r L}{R}$

Therefore, the Q factors of the individual tank circuits are

$$Q_1 = \frac{\omega_r L_1}{R_{11}} \text{ and } Q_2 = \frac{\omega_r L_2}{R_{22}} \qquad \ldots(1)$$

Usually, the Q factors for both circuits are kept same. Therefore, $Q_1 = Q_2 = Q$ and the resonant frequency $\omega_r^2 = 1/L_1 C_1 = 1/L_2C_2$.

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

$$V_o = -\frac{j}{\omega_r C_2} I_2 \qquad \ldots (2)$$

To calculate V_o/V_1 it is necessary to represent I_2 interms 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,

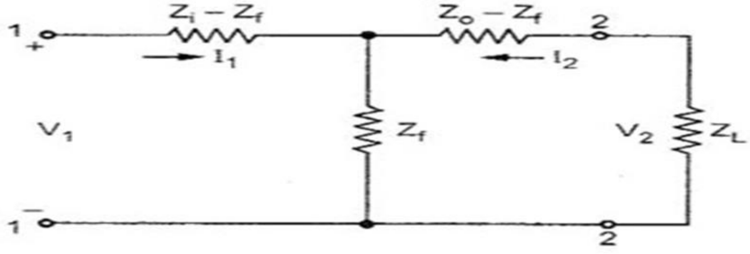


Fig. 3.20

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

$$= \frac{Z_f}{Z_f^2 - Z_i\ (Z_o + Z_L)}$$

where $$Z_{11} = \frac{V_1}{I_1} = Z_i - \frac{Z_f^2}{Z_o + Z_L} \text{ and}$$

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

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

$$Z_f = j\,\omega_r M$$

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

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

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

$$Z_f = j\omega_r M = j\omega_r k\sqrt{L_1 L_2}$$

where, k is the coefficient of coupling.

Multiplying numerator and denominator by $\omega_r L_1$ for Z_i we get,

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

$$= \frac{\omega_r\,L_1}{Q} + j\omega_r\,L_1\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right) \qquad \because Q = \frac{\omega_r\,L}{R_{11}} \text{ and } \frac{1}{\omega_r\,L} = \omega_r\,C$$

$$= \frac{\omega_r\,L_1}{Q} + j\omega_r\,L_1\,(2\delta) \qquad \because \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega} = 1 + \delta - (1 - \delta) = 2\,\delta$$

$$= \frac{\omega_r\,L_1}{Q} + (1 + j2Q\delta)$$

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

By doing similar analysis as for Z_i we can write,

$$Z_o + Z_L = \frac{\omega_r\,L_2}{Q} + (1 + j2Q\delta)$$

Then

$$Y_T = \frac{Z_f}{Z_f^2 - Z_i\,(Z_o + Z_L)} = \frac{1}{Z_f - Z_i\,(Z_o + Z_L)\,/\,Z_f}$$

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

$$\dot{Y}_T = \frac{kQ^2}{\omega_r \sqrt{L_1 L_2}\left[4Q\delta - j(1 + k^2Q^2 - 4Q^2\delta^2)\right]} \quad \dots (3)$$

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

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

$$\because V_i = \frac{j g_m V_1}{\omega C_1}$$

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

$$\because \frac{1}{\omega_r C} = \omega_r L$$

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

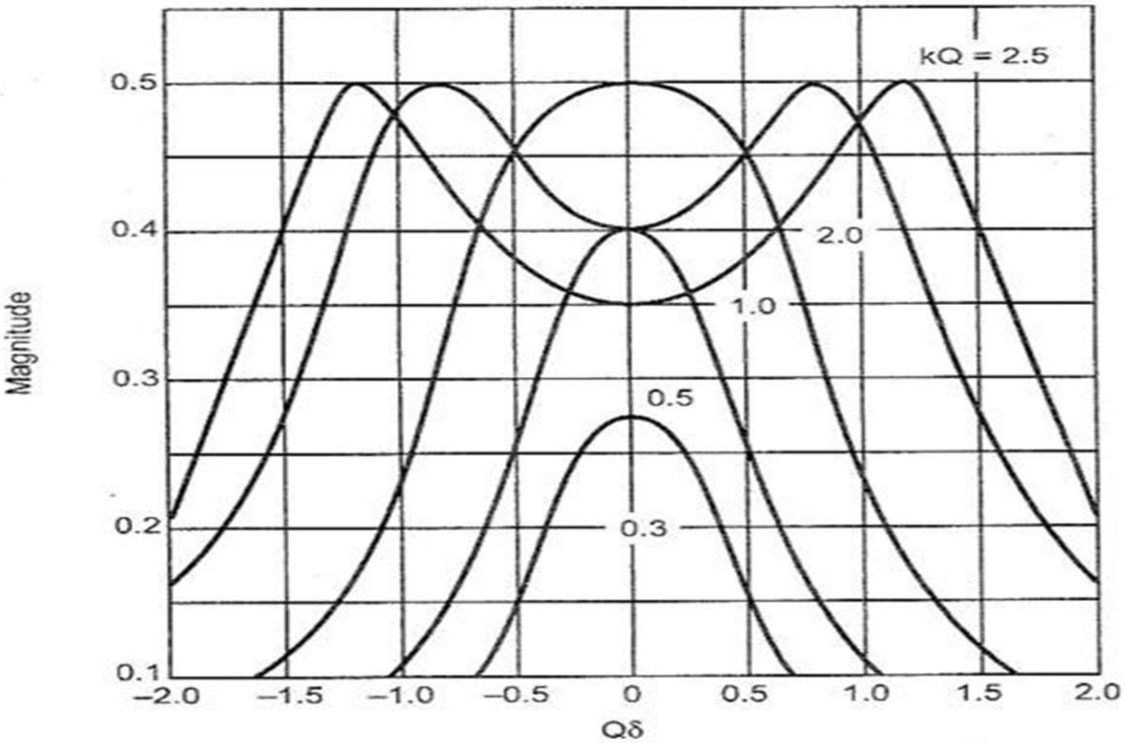
Taking the magnitude of equation (4) we have,

$$|A_v| = g_m \omega_r \sqrt{L_1 L_2}\, Q \frac{kQ}{\sqrt{1 + k^2Q^2 - 4Q^2\delta^2 + 16Q^2\delta^2}} \quad \dots (5)$$

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

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

$$4Q\delta - j(1 + k^2Q^2 - 4Q^2\delta^2) = 0 \quad \dots (6)$$



kQ = 2.5
2.0
1.0
0.5
0.3
0.5
0.4
0.3
0.2
0.1
Magnitude
−2.0
−1.5
−1.0
−0.5
0
0.5
1.0
1.5
2.0
Qδ

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,

$$|A_p| = \frac{g_m \omega_o \sqrt{L_1 L_2}\, kQ}{2} \qquad \text{... (8)}$$

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

$$|A_d| = |A_p| \frac{2\,kQ}{1+k^2Q^2} \qquad \text{... (9)}$$

The ratio of peak gain and dip gain is denoted as γ and it represents the magnitude of the ripple in the gain curve.

$$\gamma = \left|\frac{A_p}{A_d}\right| = \frac{1+k^2Q^2}{2\,kQ} \qquad \text{... (10)}$$

$$\gamma = \left|\frac{A_p}{A_d}\right| = \frac{1+k^2Q^2}{2\,kQ} \qquad \text{... (10)}$$

Using quadratic simplification and choosing positive sign we get,

$$kQ = \gamma + \sqrt{\gamma^2 - 1} \qquad \text{... (11)}$$

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,

$$BW = 2\,\delta' = \sqrt{2}\,(f_2 - f_1) \qquad \dots (12)$$

At 3 dB bandwidth,

$$\gamma = \sqrt{2}$$

$$\therefore \quad kQ = \gamma + \sqrt{\gamma^2 + 1} = \sqrt{2} + \sqrt{\sqrt{2}^2 + 1} = 2.414$$

$$\therefore \quad 3\text{ dB BW} = 2\,\delta' = \sqrt{2}\,(f_2 - f_1)$$

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

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

$$= \sqrt{2}\left[\frac{f_r}{Q}\sqrt{(2.414)^2 - 1}\right] = \frac{3.1\,f_r}{Q}$$

We know that, the 3 dB bandwidth for single tuned amplifier is $2\,f_r/Q$. 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.

Effect of cascading single tuned amplifier on bandwidth:

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 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 is given from equation (14) of section 3.4.

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

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

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

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

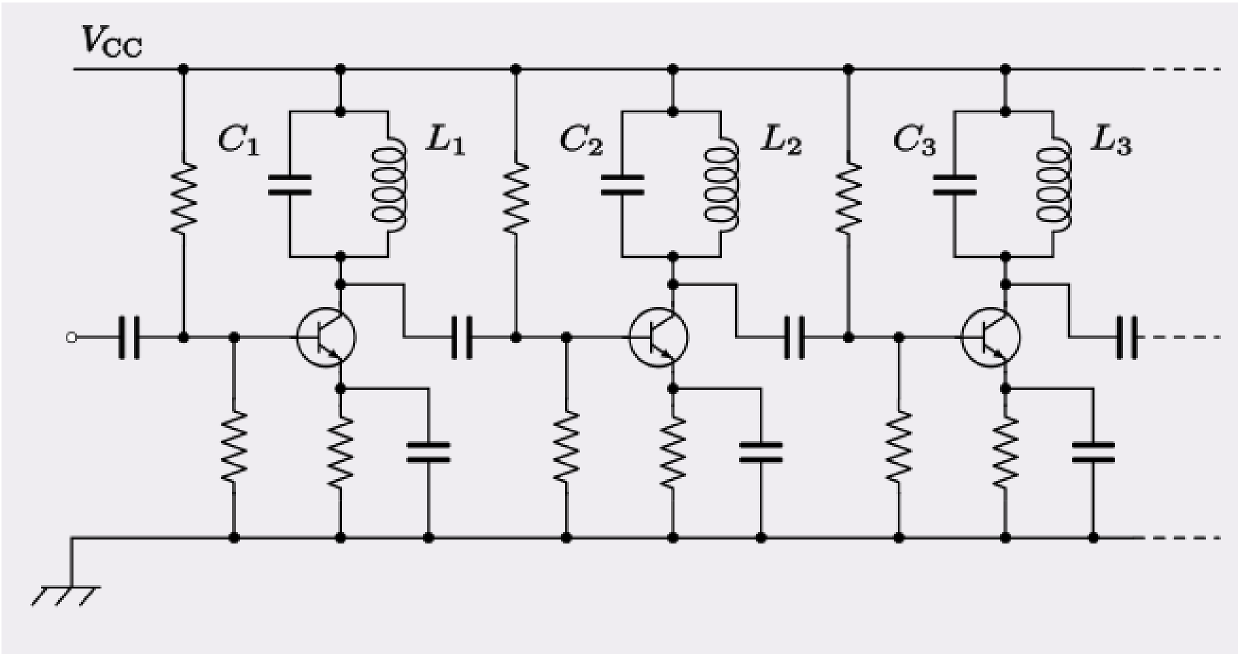


Fig. n-stage single tuned amplifier

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

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

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

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

Substituting for δ, the fractional frequency variation, i.e.

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

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

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

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

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

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

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

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

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

The bandwidth of n stage identical amplifier is given as,

$$\begin{aligned} BW_n &= f_2 - f_1 = (f_2 - f_r) + (f_r - f_1) \\ &= \frac{f_r}{2Q_{eff}} \sqrt{2^{\frac{1}{n}} - 1} + \frac{f_r}{2Q_{eff}} \sqrt{2^{1/n} - 1} \\ &= \frac{f_r}{Q_{eff}} \sqrt{2^{\frac{1}{n}} - 1} \\ &= BW_1 \sqrt{2^{\frac{1}{n}} - 1} \qquad \text{... (1)} \end{aligned}$$

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

Effect of cascading double tuned amplifier on bandwidth:

When a number of identical double tuned amplifier stages are cascaded 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 Δ_2 of such a system can be shown to be 3 dB bandwidth for

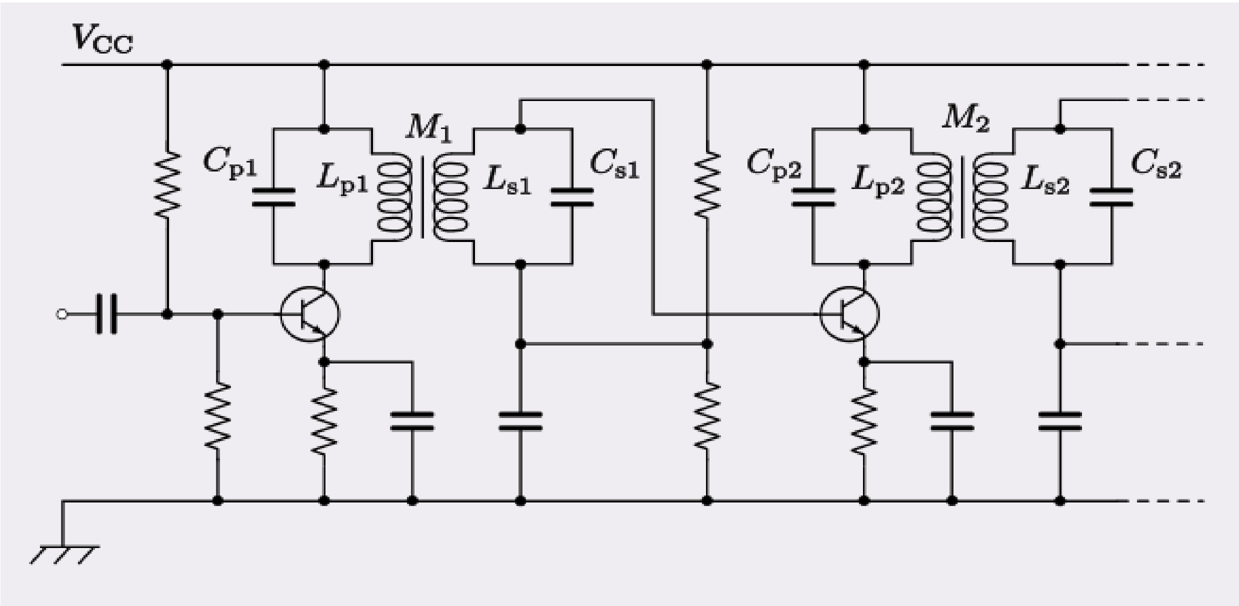


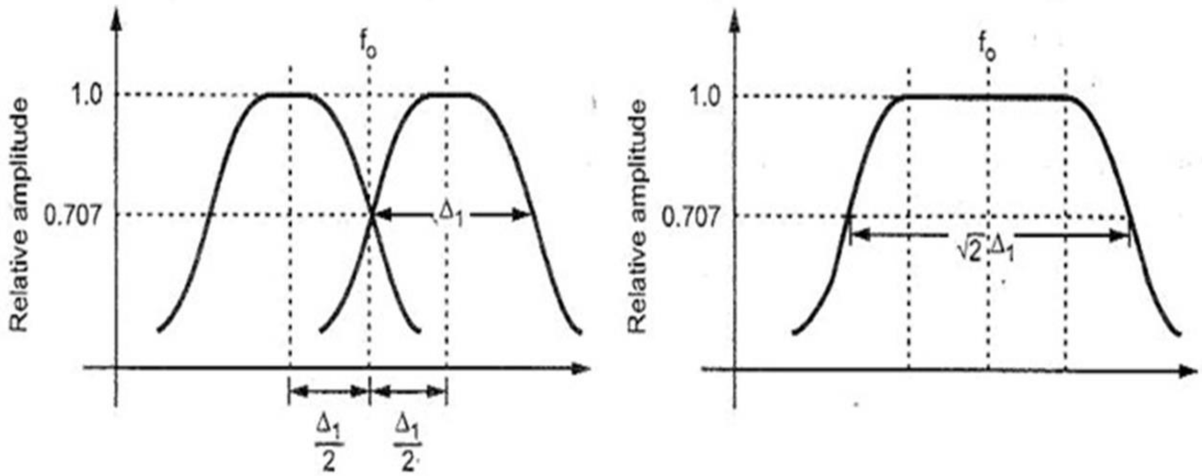
Fig. 2 stage double tuned amplifier

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

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

STAGGER TUNED AMPLIFIER:

The double tuned amplifier gives greater 3dB 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 adjusted that they are separated by an amount equal to the bandwidth of each stage. Since 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 very sharp, rejective, narrow band characteristics of synchronously tuned circuits (tuned to same resonant frequencies). Fig. 3.23 shows the relationship of amplification characteristics of individual stages in a staggered pair to the overall amplification of the two stages.



(a) Response of individual stages (b) Overall response of staggered pair

Fig. 3.23

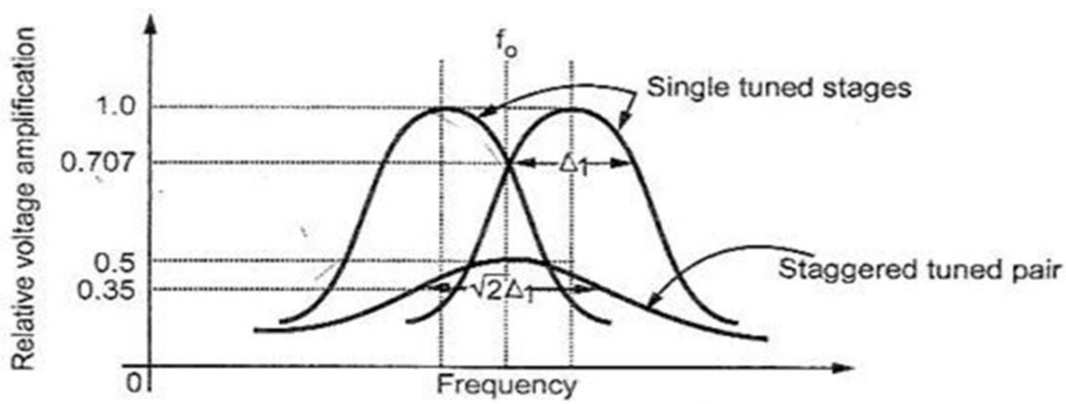


Fig. 3.24 Response of individually tuned and staggered tuned pair

The overall response of the two stage stagger tuned pair is compared in 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 an 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 an 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 easily be extended to more stages. In case of three stage 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.

Analysis of stagger tuned amplifier:

Analysis

From equation (14) of section 3.4 we can write the gain of the single tuned amplifier as,

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

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

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,

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

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

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

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

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

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

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

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

Substituting the value of X we get,

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

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

Wide Band amplifiers/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 efficiency is prime concern, class B and class C amplifiers are used at the output stages in transmitter. 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 single 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 an narrow bandwidth is desired, a resonant circuit is employed in class B and class C tuned RF power amplifiers to eliminate the harmonics.

Class B tuned amplifier

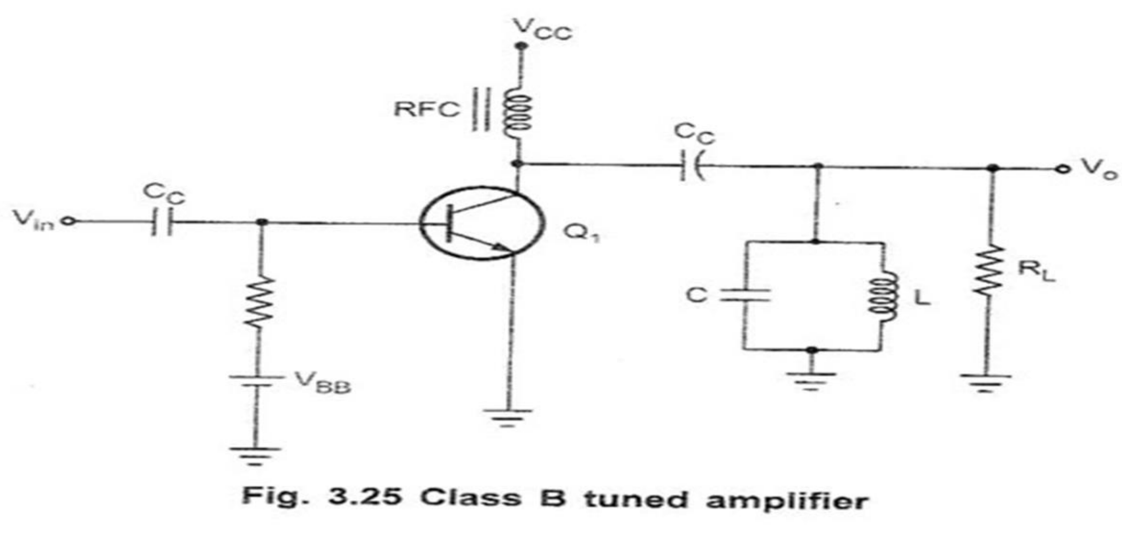


Fig. 3.25 Class B tuned amplifier

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 on a swing. We only need to push in one direction, and the reactive energy stored will swing the person back in the reverse direction.

Class C tuned amplifier:

The amplifier is said to be class C amplifier, if the Q point and the input signal are selected such that the output signal is obtained for less than a half cycle, for a full input cycle. Due to such a selection of the Q point, transistor remains active, for less than a half cycle. Only that much 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.

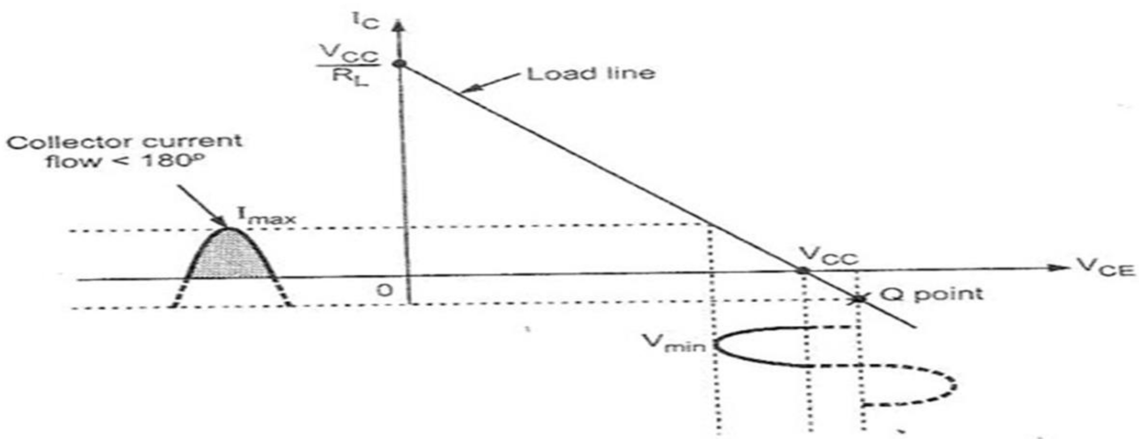
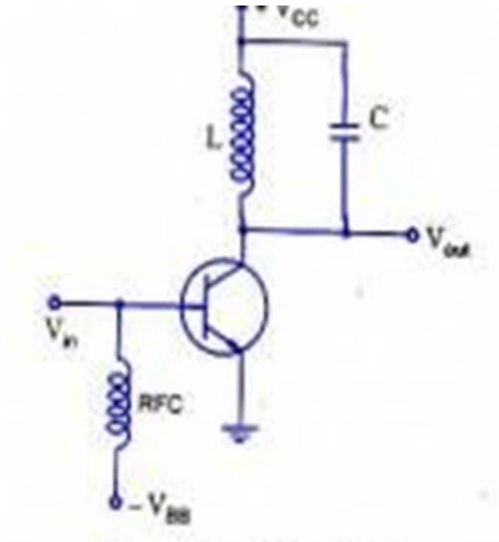


Fig. 3.26 Waveform representing class C operation

From the figure, it is apparent that the total angle during which current flows is less than 180^{o}. this angle is called the conduction angle, θ_c.



The above shows the class C tuned amplifier. Here a parallel resonant circuit acts 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 a produce a sine wave output voltage consisting of fundamental component of the input signal.

Fast track material for QUICK REFERENCE:

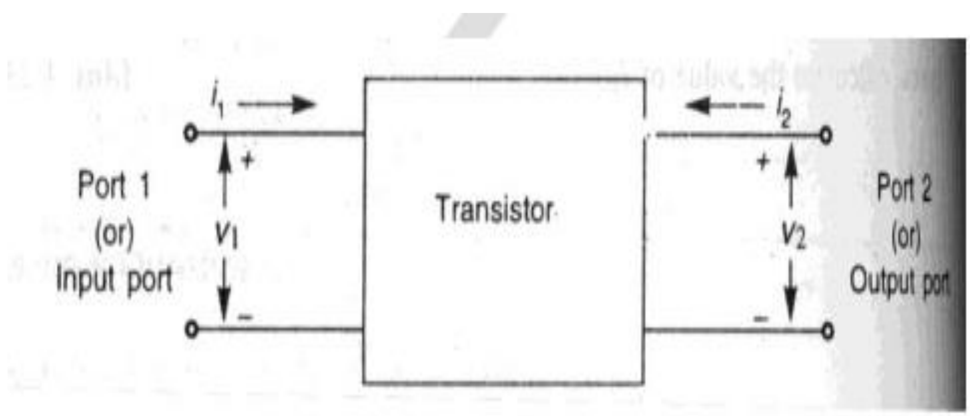
Small signal high frequency transistor amplifier

Introduction:

Electronic circuit analysis subject teaches about the basic knowledge required to design an amplifier circuit, oscillators etc .It provides a clear and easily understandable discussion of designing of different types of amplifier circuits and their analysis using hybrid model, to find out their parameters. Fundamental concepts are illustrated by using small examples which are easy to understand. It also covers the concepts of MOS amplifiers, oscillators and large signal amplifiers.

Two port devices & Network Parameters: -

A transistor can be treated as a two-part network. The terminal behavior of any two-part network can be specified by the terminal voltages V1& V2at parts 1 & 2 respectively and current i1and i2, entering parts 1 & 2, respectively, as shown in figure.

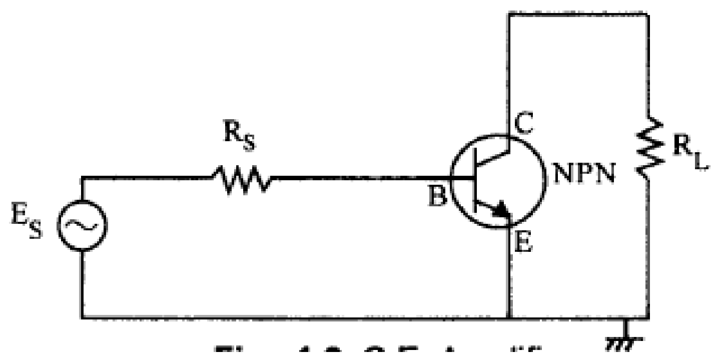


Of these four variables V1, V2, i1and i2, two can be selected as independent variables and the remaining two can be expressed in terms of these independent variables. This leads to various two part parameters out of which the following three are more important.

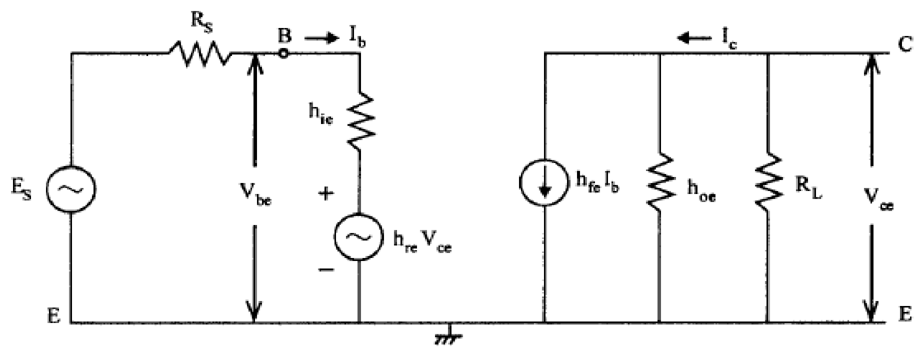
1. Z –Parameters (or) Impedance parameters
2. Y –Parameters (or) Admittance parameters
3. H –Parameters (or) Hybrid parameters

Common Emitter Amplifier:

Common Emitter Circuit is as shown in the Fig. The DC supply, biasing resistors and coupling capacitors are not shown since we are performing an AC analysis.



E_S is the input signal source and R_S is its resistance. The h-parameter equivalent for the above circuit is as shown in Fig.



$$h_{ie} = \left.\frac{V_{be}}{I_b}\right|_{V_{ce}=0} \qquad h_{re} = \left.\frac{V_{be}}{V_{ce}}\right|_{I_b=0}$$

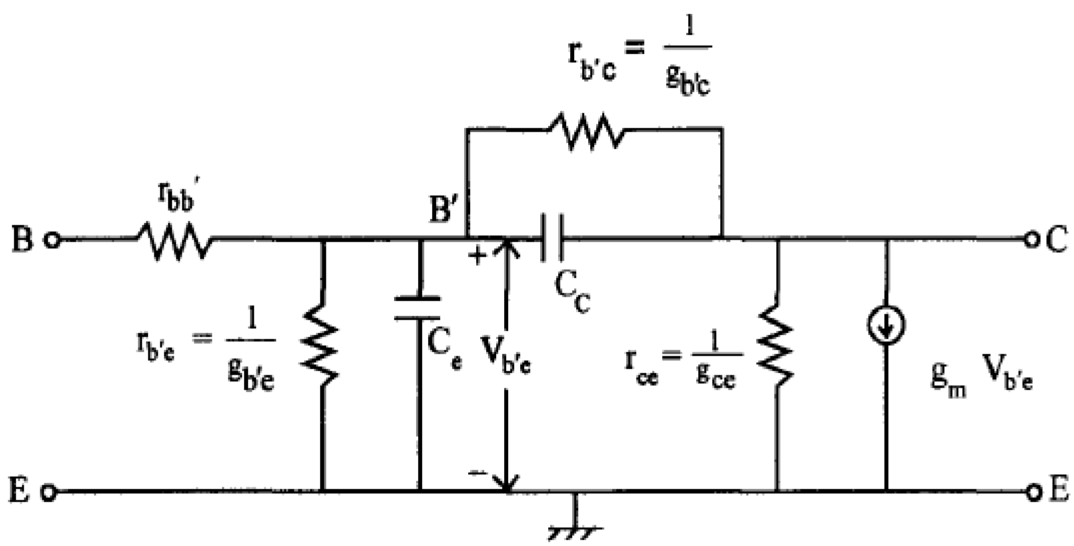
$$h_{oe} = \left.\frac{I_c}{V_{ce}}\right|_{I_b=0} \qquad h_{fe} = \left.\frac{I_c}{I_b}\right|_{V_{ce}=0}$$

The typical values of the h-parameter for a transistor in Common Emitter configuration are,

$$h_{ie} = \frac{V_{be}}{I_b}$$

Hybrid - π Common Emitter Tran conductance Model:

For Tran conductance amplifier circuits Common Emitter configuration is preferred. Why? Because for Common Collector (hrc< 1). For Common Collector Configuration, voltage gain Av < 1. So even by cascading you can't increase voltage gain. For Common Base, current gain is hib< 1. Overall voltage gain is less than 1. For Common Emitter, hre>>1. Therefore Voltage gain can be increased by cascading Common Emitter stage. So Common Emitter configuration is widely used. The Hybrid-x or Giacoletto Model for the Common Emitter amplifier circuit (single stage) is as shown below



Analysis of this circuit gives satisfactory results at all frequencies not only at high frequencies but also at low frequencies. All the parameters are assumed to be independent of frequency.

Where

B’ = internal node in base

$r_{bb'}$ = Base spreading resistance

$r_{b'e}$ = Internal base node to emitter resistance

r_{ce} = collector to emitter resistance

C_e = Diffusion capacitance of emitter base junction
$r_{b'c}$ = Feedback resistance from internal base node to collector node
g_m = Transconductance
C_C= transition or space charge capacitance of base collector junction

Hybrid - π Capacitances:

In the hybrid - **π** equivalent circuit, there are two capacitances, the capacitance between the Collector Base junction is the C_c or $C_{b'e}$'. This is measured with input open i.e., $I_E = 0$, and is specified by the manufacturers as C_{Ob}. 0 indicates that input is open. Collector junction is reverse biased.

$$C_C \; \alpha \; \frac{1}{(V_{CE})^n}$$

$$n = \frac{1}{2} \quad \text{for abrupt junction}$$

$$= 1/3 \quad \text{for graded junction.}$$

C_e = Emitter diffusion capacitance C_{De} + Emitter junction capacitance C_{Te}

C_T = Transition capacitance.

C_D = Diffusion capacitance.

$C_{De} >> C_{Te}$

$C_e \simeq C_{De}$

$C_{De} \; \alpha \; I_E$ and is independent of Temperature T.

Validity of hybrid-π model:

The high frequency hybrid Pi or Giacoletto model of BJT is valid for frequencies less than the unit gain frequency.

Current Gain with Resistance Load:

$$f_T = f_\beta \cdot h_{fe} = \frac{g_m}{2\pi(C_e + C_c)}$$

The Parameters f_T

f_T is the frequency at which the short circuit Common Emitter current gain becomes unity.

The Parameters f_β

$$A_i = 1, \quad \text{or} \quad \frac{h_{fe}}{\sqrt{1 + \left(\frac{f_T}{f_\beta}\right)^2}} = 1$$

$$f = f_T, \quad A_i = 1$$

$$h_{fe} = \sqrt{1 + \left(\frac{f_T}{f_\beta}\right)^2}$$

$$(h_{fe})^2 = 1 + \left(\frac{f_T}{f_\beta}\right)^2 \cong \left(\frac{f_T}{f_\beta}\right)^2$$

$$h_{fe} \simeq \frac{f_T}{f_\beta} \text{ when } Ai = 1$$

$$\boxed{f_T \simeq h_{fe} \cdot f_\beta}$$

$$f_\beta = \frac{g_m}{h_{fe}\{C_e + C_c\}}$$

$$f_T = f_\beta \cdot h_{fe} = \frac{g_m}{2\pi(C_e + C_c)}$$

$$C_e >> C_c$$

$$\boxed{f_T \simeq \frac{g_m}{2\pi C_e}}$$

Gain - Bandwidth (B.W) Product

This is a measure to denote the performance of an amplifier circuit. Gain - B. W product is also referred as Figure of Merit of an amplifier. Any amplifier circuit must have large gain and large bandwidth. For certain amplifier circuits, the midband gain Am maybe large, but not Band width or Vice - Versa. Different amplifier circuits can be compared with thus parameter.

Multistage Amplifiers:

Classification of amplifiers

Depending upon the type of coupling, the multistage amplifiers are classified as:

1. Resistance and Capacitance Coupled Amplifiers (RC Coupled)
2. Transformer Coupled Amplifiers
3. Direct Coupled DC Amplifiers
4. Tuned Circuit Amplifiers.

Based upon the B. W. of the amplifiers, they can be classified as:

1. Narrow hand amplifiers
2. Untuned amplifiers

Narrow hand amplifiers: Amplification is restricted to a narrow band offrequencies arounda centre frequency. There are essentially tuned amplifiers.

Untuned amplifiers: These will have large bandwidth. Amplification is desired over a Considerable range of frequency spectrum.

Untuned amplifiers are further classified w.r.t bandwidth.

I. DC amplifiers (Direct Coupled) DC to few KHz	
2. Audio frequency amplifiers (AF)	20 Hz to 20 KHz
3. Broad band amplifier	DC to few MHz
4. Video amplifier	100 Hz to few MHz

The gain provided by an amplifier circuit is not the same for all frequencies because the reactance of the elements connected in the circuit and the device reactance value depend upon the frequency. Bandwidth of an amplifier is the frequency range over which the amplifier stage gain is reasonably constant within ± 3 db, or O. 707 of A_V Max Value.

Resistance and Capacitance Coupled Amplifiers (RC Coupled)

This type of amplifier is very widely used. It is least expensive and has good frequency response.In the multistage resistive capacitor coupled amplifiers, the output of the first stage is coupled to the next through coupling capacitor and R_L. In two stages Resistor Capacitor coupled amplifiers, there is no separate R_L between collector and ground, but Reo the resistance between collector and V cc (R_C) itself acts as R_L in the AC equivalent circuit.

Transformer Coupled Amplifiers

` Here the output of the amplifier is coupled to the next stage or to the load through a transformer. With this overall circuit gain will be increased and also impedance matching can be achieved. But such transformer coupled amplifiers will not have broad frequency response i.e., (f_2-f_1)is small since inductance of the transformer windings will be large. So Transformer coupling is done for power amplifier circuits, where impedance matching is critical criterion for maximum power to be delivered to the load.

Direct Coupled (DC) Amplifiers

Here DC stands for direct coupled and not (direct current). In this type, there is no reactive element. L or C used to couple the output of one stage to the other. The AC output from the collector of one stage is directly given to the base of the second stage transistor directly. So type of amplifiers is used for large amplification of DC and using low frequency signals. Resistor Capacitor coupled amplifiers cannot be used for amplifications of DC or low frequency signals since Xc the capacitive reactance of the coupling capacitor will be very large or open circuit for DC

Tuned Circuit Amplifiers

In this type there will be one RC or LC tuned circuit between collector and V_{CC} in the place of Re. These amplifiers will amplify signals of only fixed frequency.fo which is equal to the resonance frequency of the tuned circuit LC. These are also used to amplify signals of a narrow band of frequencies centered on the tuned frequency f_0.

Distortion in Amplifiers

If the input signal is a sine wave the output should also be a true sine wave. But in all the cases it may not be so, which we characterize as distortion. Distortion can be due to the nonlinear characteristic of the device, due to operating point not being chosen properly, due to large signal swing of the input from the operating point or due to the reactive elements Land C in the circuit. Distortion is classified as:

(a)**Amplitude distortion**: This is also called non linear distortion or harmonic distortion. This type of distortion occurs in large signal amplifiers or power amplifiers. It is due to then on linearity of the characteristic of the device. This is due to the presence of new frequency signals which are not present in the input. If the input signal is of 10 KHz the output signal should also be 10 KHz signal. But some harmonic terms will also be present. Hence the amplitude of the signal (rms value) will be different Vo = Ay Vi.

(b) **Frequency distortion**: The amplification will not be the same for all frequencies. This is due to reactive component in the circuit.

(c) **Phase - shift delay distortion**: There will be phase shift between the input and the output and this phase shift will not be the same for all frequency signals. It also varies with the frequency of the input signal. In the output signal, all these distortions may be present or anyone may be present because of which the amplifier response will not be good.

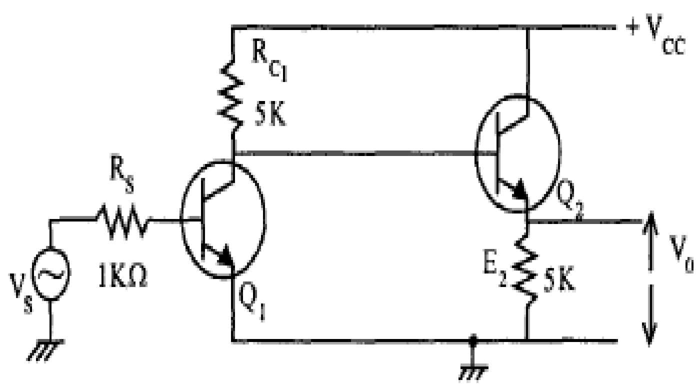
The overall gain of a multistage amplifier is the product of the gains of the individual stage (ignoring potential loading effects):

Gain (A) = A1 * A2 * A3 * A4 * …… *An.

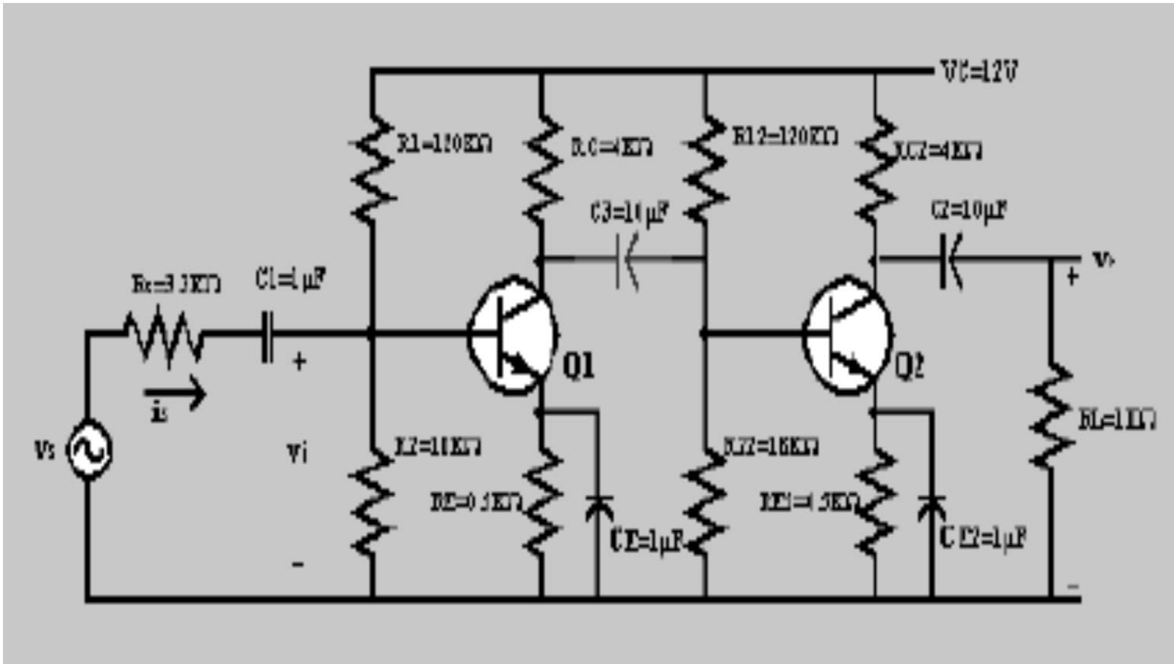
Alternately, if the gain of each amplifier stage is expressed in decibels (dB), the total gain is the sum of the gains of the individual stages

Gain in dB (A) = A1 + A2 + A3 + A4 + …… + An.

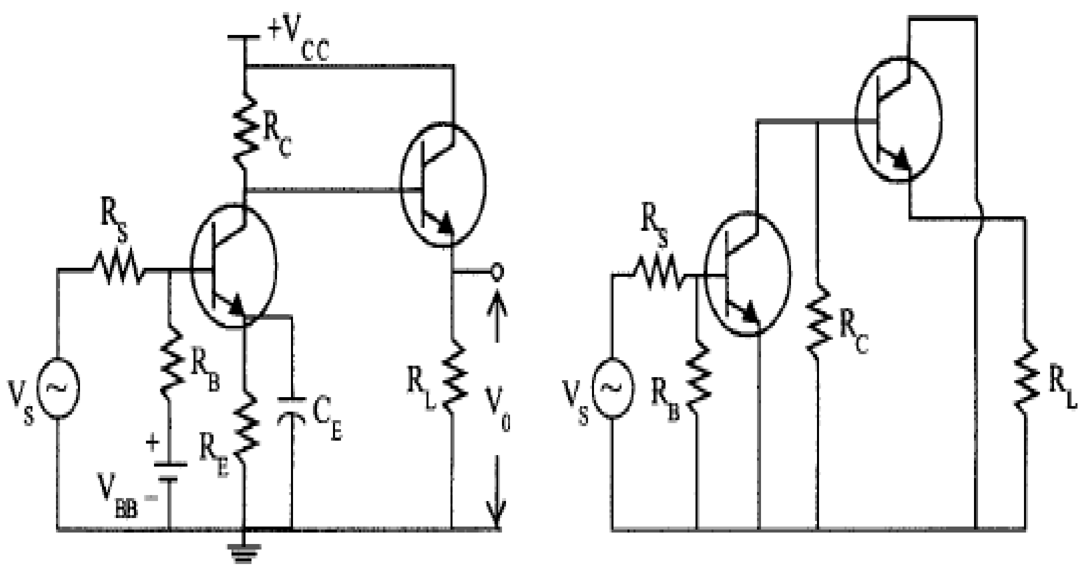
The Two Stage Cascaded Amplifier Circuit:



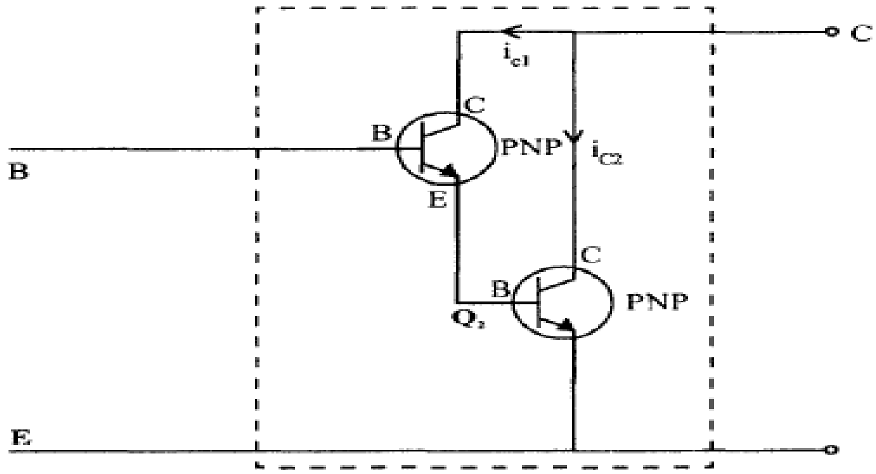
Two stage RC coupled amplifier:



CE - CC Amplifiers:



The Darlington Pair:



Current gain

$$A_I = \frac{I_c}{I_{b_1}} \cong (h_{fe})^2$$

Input resistance

$$R_i \simeq \frac{(1+h_{fe})^2 R_e}{1 + h_{oe} h_{fe} R_e}$$

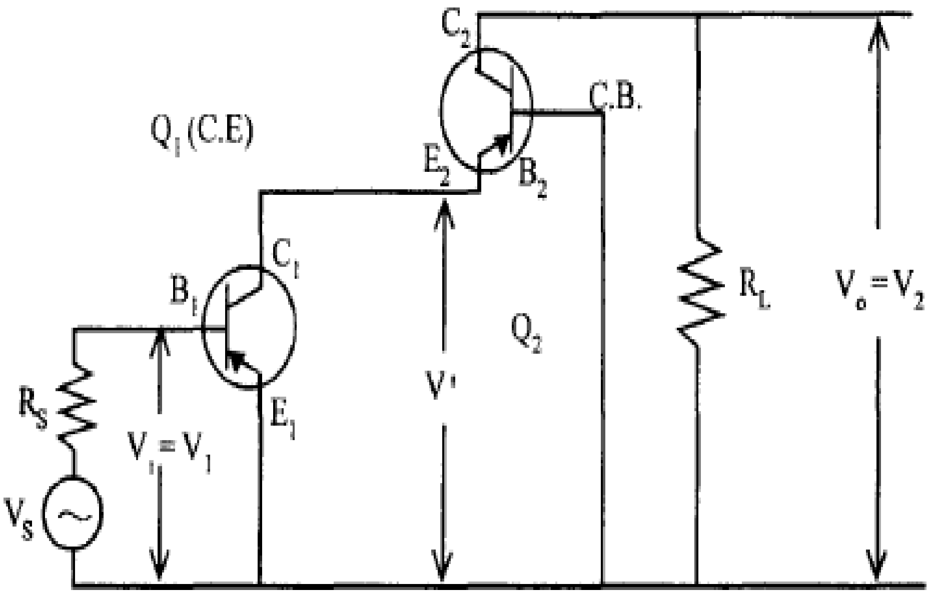
Voltage gain

$$A_v \cong \left(1 - \frac{h_{ie}}{Ri_2}\right)$$

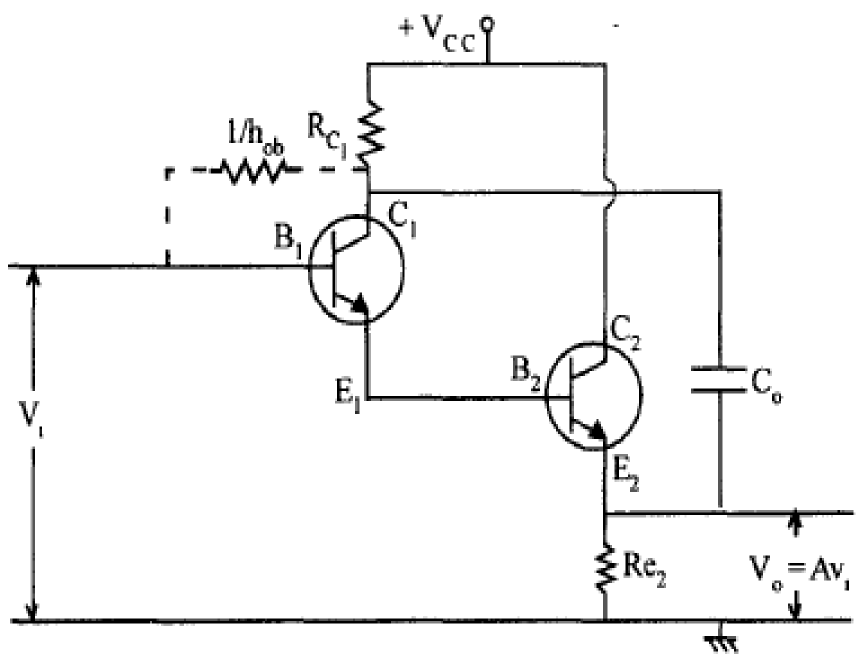
Output resistance

$$R_{o2} = \frac{R_s + h_{ie}}{(1+h_{fe})^2} + \frac{h_{ie}}{1+h_{fe}}$$

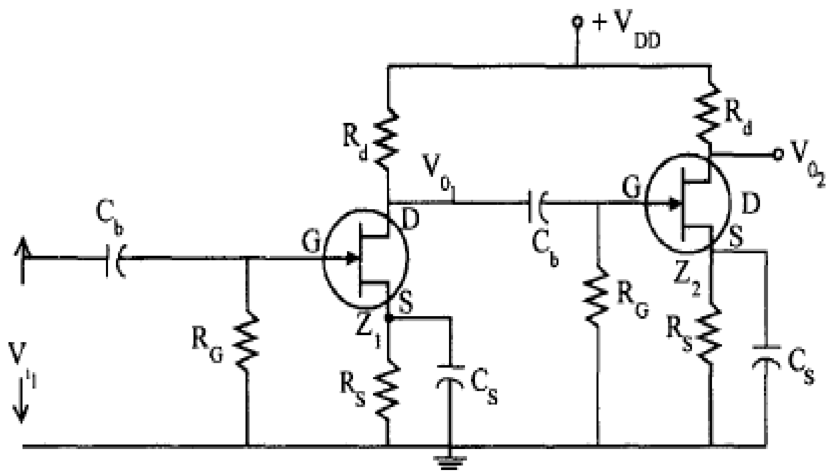
The CASCODE Transistor Configuration:



Boot-strap emitter follower:

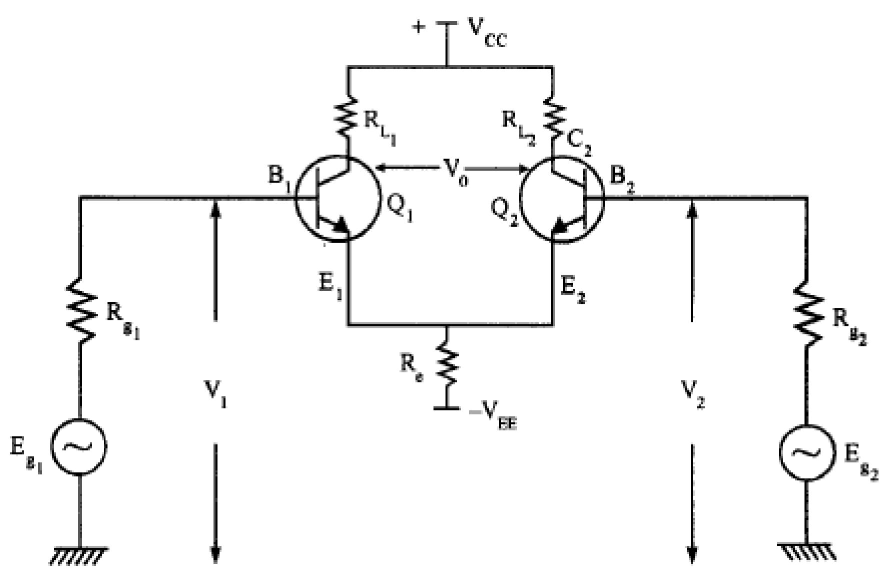


Two Stage RC Coupled JFET amplifier (in Common Source (CS) configuration):



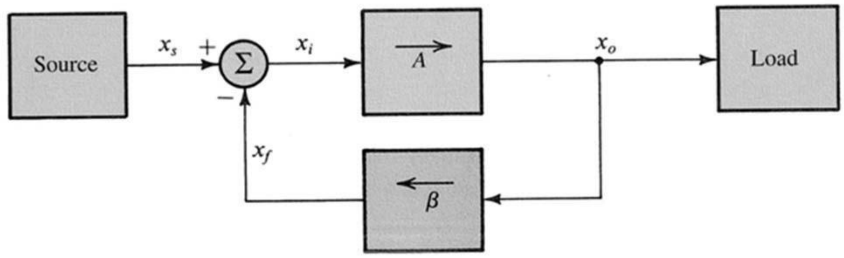
Circuit for Differential Amplifier

In the previous D.C amplifier viz., C.B, C.C and C.E, the output is measured with respect to ground. But in difference amplifier, the output is proportional to the difference of the inputs. So Vo is not measured w.r.t ground but w.r.t to the output of one transistor Q_1 or output of the other transistor Q_2'.



Feedback Amplifier

FEEDBACK AMPLIFIER:



- ➢ Signal-flow diagram of a feedback amplifier
- ➢ Open-loop gain: A
- ➢ Feedback factor:
- ➢ Loop gain: A
- ➢ Amount of feedback: 1 + A
- ➢ Gain of the feedback amplifier (closed-loop gain): □

Negative feedback:

- ➢ The feedback signal xf is subtracted from the source signal xs
- ➢ Negative feedback reduces the signal that appears at the input of the basic amplifier
- ➢ The gain of the feedback amplifier Af is smaller than open-loop gain A by a factor of (1+A)
- ➢ The loop gain A is typically large (A >>1):
- ➢ The gain of the feedback amplifier (closed-loop gain)
- ➢ The closed-loop gain is almost entirely determined by the feedback network □ better accuracy of Af.
- ➢ xf = xs(A)/(1+A) xs□ error signal xi = xs – xf

For Example, The feedback amplifier is based on an opamp with infinite input resistance and zero output resistance

- ➢ Find an expression for the feedback factor.
- ➢ Find the condition under which the closed-loop gain Af is almost entirely determined by the feedback network.
- ➢ If the open-loop gain A = 10000 V/V, find R2/R1 to obtain a closed-loop gain Af of 10 V/V.
- ➢ What is the amount of feedback in decibel?
- ➢ If Vs = 1 V, find Vo, Vf and Vi.

- If A decreases by 20%, what is the corresponding decrease in Af ?

Some Properties of Negative Feedback

Gain desensitivity:

- The negative reduces the change in the closed-loop gain due to open-loop gain variation

$$dA_f = \frac{dA}{(1+A\beta)^2} \rightarrow \frac{dA_f}{A_f} = \frac{1}{1+A\beta}\frac{dA}{A}$$

- Desensitivity factor: $1+A\beta$

The Four Basic Feedback Topologies:

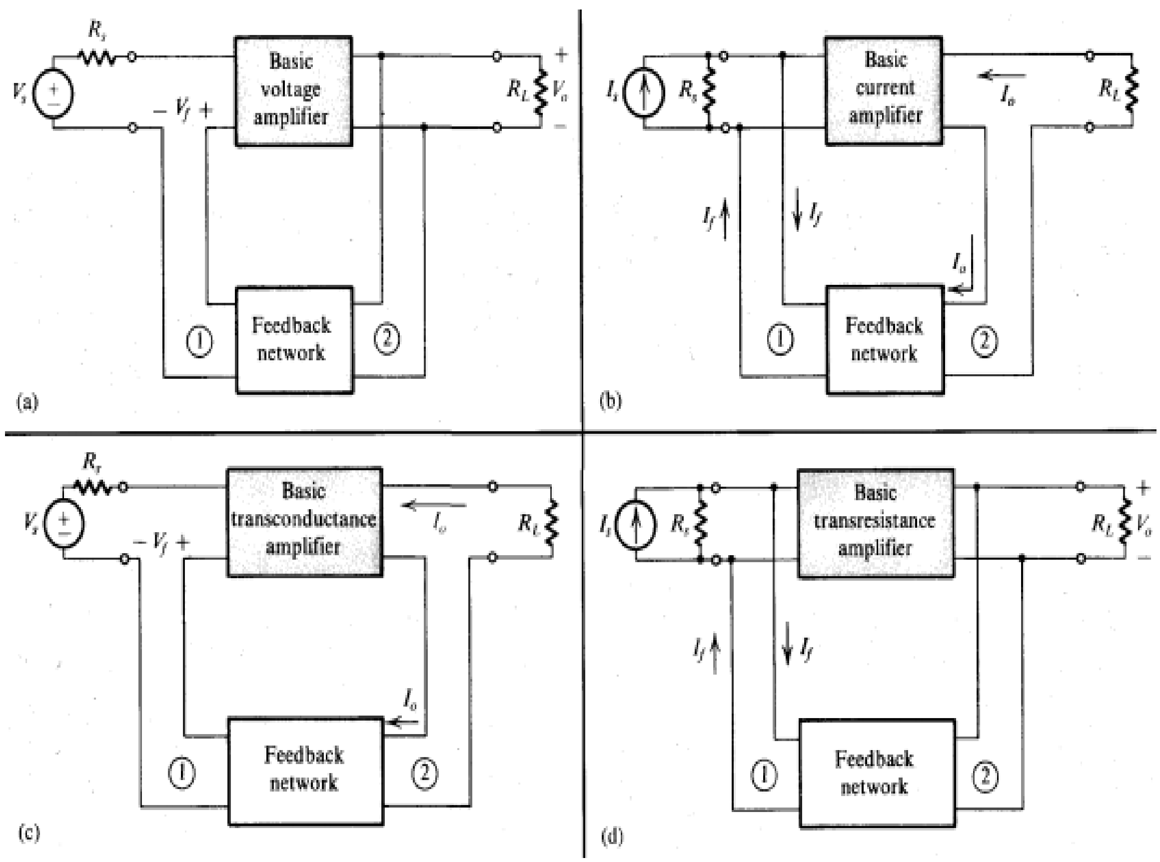
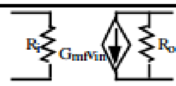
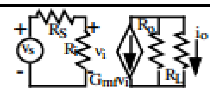
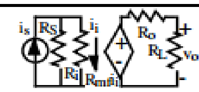
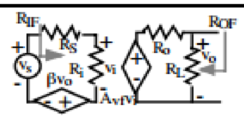
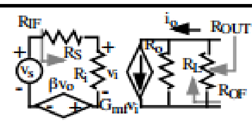
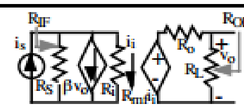


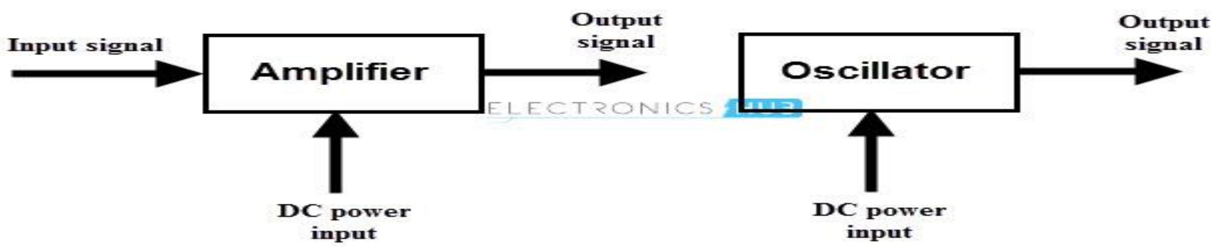
Fig. The four basic feedback topologies: (a) voltage-sampling series-mixing (series-shunt) topology; (b) current-sampling shunt-mixing (shunt-series) topology; (c) current-sampling series-mixing (series-series) topology; (d) voltage-sampling shunt-mixing (shunt-shunt) topology.

Summary of the Important Relationships of Open-loop and Closed-loop Feedback Amplifiers.

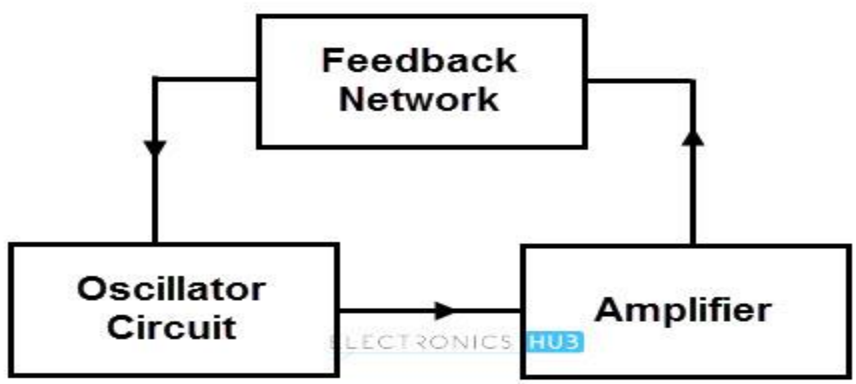
Quantity	Voltage Amplifier	Transconductance Amplifier	Transresistance Amplifier	Current Amplifier
Input-output variable	Voltage-voltage	Voltage-current	Current-voltage	Current-current
Small Signal Model				
Small Signal Amplifier with Source & Load				
Ideal R_S	$R_S = 0$ or $R_S << R_i$	$R_S = 0$ or $R_S << R_i$	$R_S = \infty$ or $R_S >> R_i$	$R_S = \infty$ or $R_S >> R_i$
Ideal R_L	$R_L = \infty$ or $R_L >> R_o$	$R_L = 0$ or $R_L << R_o$	$R_L = \infty$ or $R_L >> R_o$	$R_L = 0$ or $R_L << R_o$
Overall Forward Gain	$A_V = \frac{R_i R_L A_{vf}}{(R_S+R_i)(R_L+R_o)}$	$G_M = \frac{R_i R_o G_{mf}}{(R_S+Ri)(R_L+R_o)}$	$R_M = \frac{R_S R_L R_{mf}}{(R_S+Ri)(R_L+R_o)}$	$A_I = \frac{R_S R_o A_{if}}{(R_S+Ri)(R_L+R_o)}$
Feedback Topology	Series-shunt	Series-series	Shunt-shunt	Shunt-series
Ideal ß, finite R_S and R_L Feedback Small Signal Models				
Closed-Loop Gain (Ideal R_S and R_L)	$A_{vF} = \frac{A_{vf}}{(1+A_{vf}\beta_v)}$	$G_{mF} = \frac{G_{mf}}{(1+G_{mf}\beta_g)}$	$R_{mF} = \frac{R_{mf}}{(1+R_{mf}\beta_r)}$	$A_{iF} = \frac{A_{if}}{(1+A_{if}\beta_i)}$
Closed-Loop Input Resistance (Ideal R_S and R_L)	$R_{iF} = R_i(1 + A_{vf}\beta_v)$	$R_{iF} = R_i(1 + G_{mf}\beta_g)$	$R_{iF} = \frac{R_i}{1 + R_{mf}\beta_r}$	$R_{iF} = \frac{R_i}{1 + A_{if}\beta_i}$
Closed-Loop Output Resistance (Ideal R_S and R_L)	$R_{oF} = \frac{R_o}{1 + A_{vf}\beta_v}$	$R_{oF} = R_o(1 + R_{mf}\beta_g)$	$R_{oF} = \frac{R_o}{1 + R_{mf}\beta_r}$	$R_{oF} = R_o(1 + A_{if}\beta_i)$
Closed-Loop Gain	$A_{VF} = \frac{A_V}{(1+A_V\beta_v)}$	$G_{MF} = \frac{G_M}{(1+G_M\beta_g)}$	$R_{MF} = \frac{R_M}{(1+R_M\beta_r)}$	$A_{IF} = \frac{A_I}{(1+A_I\beta_i)}$
Closed-Loop Input Resistance	$R_{IF} = (R_i+R_S)(1+A_V\beta_v)$	$R_{IF} = (R_i+R_S)(1 + G_M\beta_g)$	$R_{IF} = \frac{\frac{R_i R_S}{R_i+R_S}}{1 + R_M\beta_r}$	$R_{IF} = \frac{\frac{R_i R_S}{R_i+R_S}}{1 + A_I\beta_i}$
Closed-Loop Output Resistance	$R_{OF} = \frac{\frac{R_o R_L}{R_o+R_L}}{1 + A_V\beta_v}$	$R_{OF} = (R_o+R_L)(1+G_M\beta_g)$	$R_{OF} = \frac{\frac{R_o R_L}{R_o+R_L}}{1 + R_M\beta_r}$	$R_{OF} = (R_o+R_L)(1+A_I\beta_i)$
Output Resistance of Series Output Fb. Ckt	$R_{OUT} = R_{OF}$	$R_{OUT} = \frac{R_L}{R_{OF}}(R_{OF}-R_L)$	$R_{OUT} = R_{OF}$	$R_{OUT} = \frac{R_L}{R_{OF}}(R_{OF}-R_L)$

Oscillators

An electronic circuit used to generate the output signal with constant amplitude and constant desired frequency is called as an oscillator. It is also called as a waveform generator which incorporates both active and passive elements. The primary function of an oscillator is to convert DC power into a periodic signal or AC signal at a very high frequency. An oscillator does not require any external input signal to produce sinusoidal or other repetitive waveforms of desired magnitude and frequency at the output and even without use of any mechanical moving parts.



In case of amplifiers, the energy conversion starts as long as the input signal is present at the input, i.e., amplifier produces an output signal whose frequency or waveform is similar to the input signal but magnitude or power level is generally high. The output signal will be absent if there is no input signal at the input. In contrast, to start or maintain the conversion process an oscillator does not require any input signal as shown figure. As long as the DC power is connected to the oscillator circuit, it keeps on producing an output signal with frequency decided by components in it.

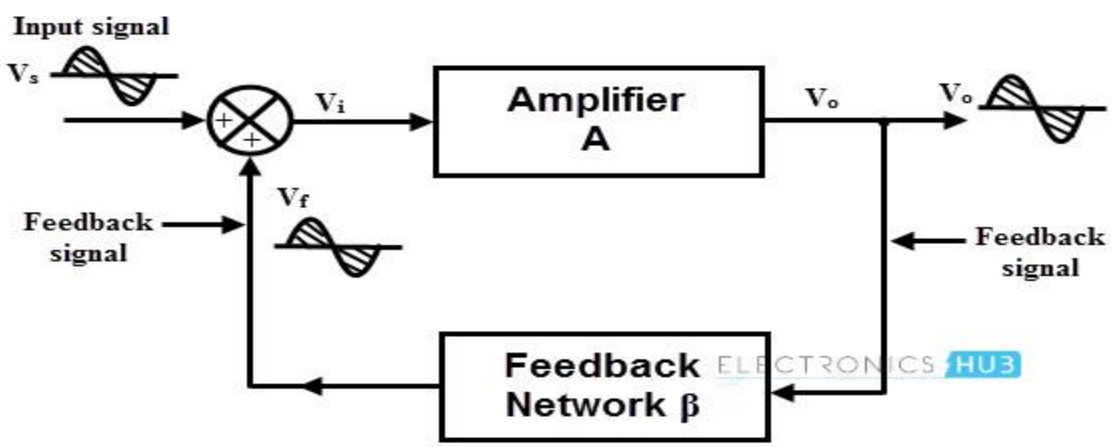


The above figure shows the block diagram of an oscillator. An oscillator circuit uses a vacuum tube or a transistor to generate an AC output. The output oscillations are produced by the tank circuit components either as R and C or L and C. For continuously generating output without the requirement of any input from preceding stage, a feedback circuit is used.

From the above block diagram, oscillator circuit produces oscillations that are further amplified by the amplifier. A feedback network gets a portion of the amplifier output and feeds it the oscillator circuit in correct phase and magnitude. Therefore, un damped electrical oscillations are produced , by continuously supplying losses that occur in the tank circuit.

Oscillators Theory

The main statement of the oscillator is that the oscillation is achieved through positive feedback which generates the output signal without input signal. Also, the voltage gain of the amplifier increases with the increase in the amount of positive feedback. In order to understand this concept, let us consider a non-inverting amplifier with a voltage gain ‘A’ and a positive feedback network with feedback gain of β as shown in figure.



Let us assume that a sinusoidal input signal Vs is applied at the input. Since the amplifier is non-inverting, the output signal Vo is in phase with Vs. A feedback network feeds the part of Vo to the input and the amount Vo fed back depends on the feedback network gain β. No phase shift is introduced by this feedback network and hence the feedback voltage or signal Vf is in phase with Vs. A feedback is said to be positive when the phase of the feedback signal is same as that of the input signal. The open loop gain ‘A’ of the amplifier is the ratio of output voltage to the input voltage, i.e.,

$$A = Vo/Vi$$

By considering the effect of feedback, the ratio of net output voltage Vo and input supply Vs called as a closed loop gain Af (gain with feedback).

$$Af = Vo/Vs$$

Since the feedback is positive, the input to the amplifier is generated by adding Vf to the Vs,

$$Vi = Vs + Vf$$

Depends on the feedback gain β, the value of the feedback voltage is varied, i.e.,

$$Vf = \beta\ Vo$$

Substituting in the above equation,

$$Vi = Vs + \beta\ Vo$$

$$Vs = Vi - \beta\ Vo$$

Then the gain becomes

$$Af = Vo/\ (Vi - \beta\ Vo)$$

By dividing both numerator and denominator by Vi, we get

$$Af = (Vo\ /\ Vi)/\ (1 - \beta)\ (Vo\ /\ Vi)$$

$$Af = A/\ (1-\ A\ \beta)\ \text{since}\ A = Vo/Vi$$

Where Aβ is the loop gain and if Aβ = 1, then Af becomes infinity. From the above expression, it is clear that even without external input (Vs = 0), the circuit can generate the output just by feeding a part of the output as its own input. And also closed loop gain increases with increase in amount of positive feedback gain. The oscillation rate or frequency depends on amplifier or feedback network or both.

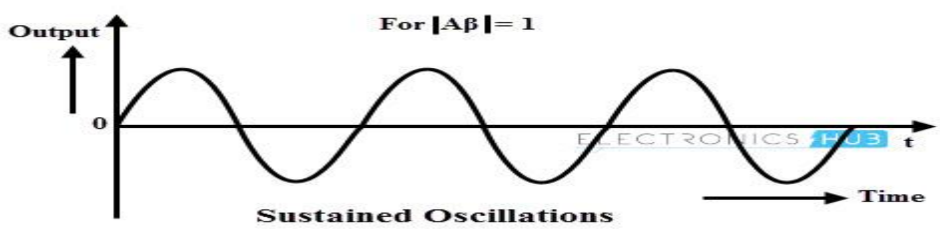
Barkhausen Criterion or Conditions for Oscillation:

The circuit will oscillate when two conditions, called as Barkhausen's criteria are met. These two conditions are

1. The loop gain must be unity or greater
2. The feedback signal feeding back at the input must be phase shifted by 360 degrees (which is same as zero degrees).

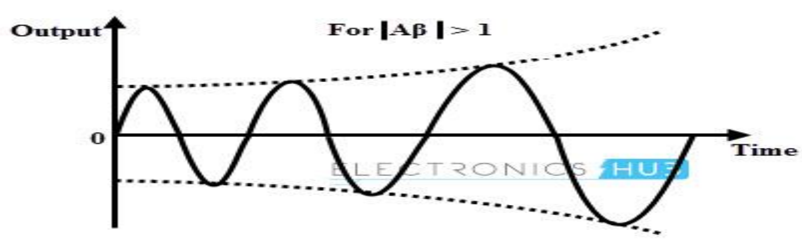
Nature of Oscillations

Sustained Oscillations: Sustained oscillations are nothing but oscillations which oscillate with constant amplitude and frequency. Based on the Barkhausen criterion sustained oscillations are produced when the magnitude of loop gain or modulus of A β is equal to one and total phase shift around the loop is 0 degrees or 360 ensuring positive feedback.



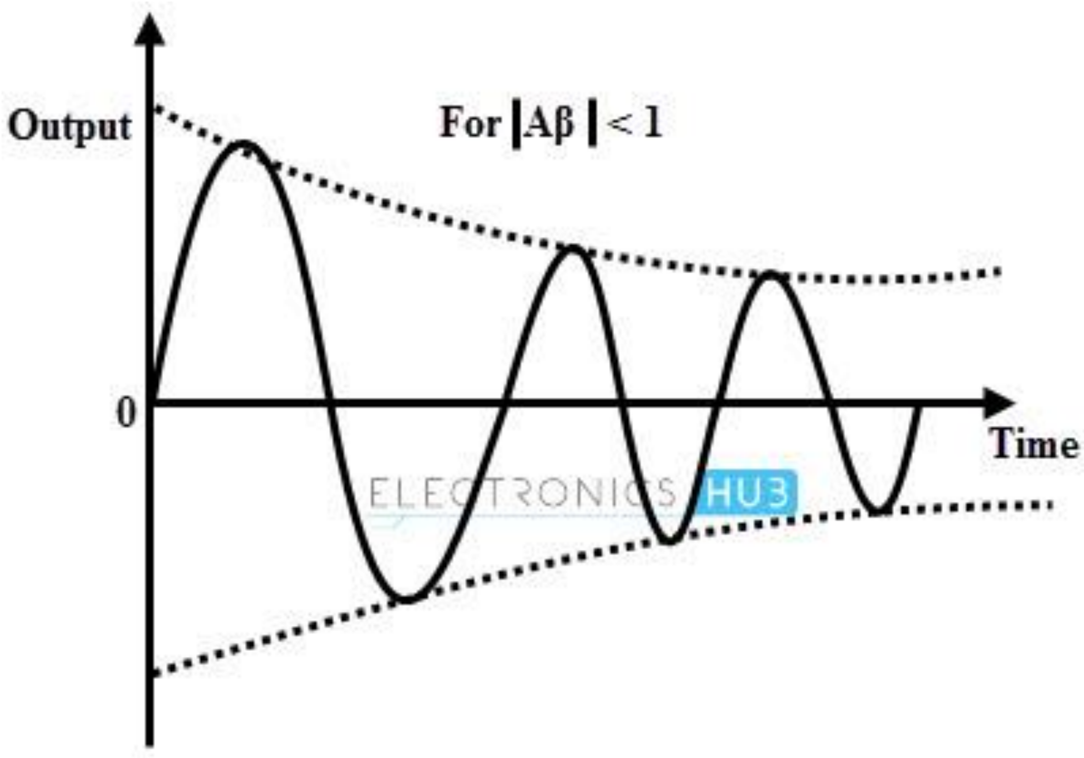
Growing Type of Oscillations:

If modulus of A β or the magnitude of loop gain is greater than unity and total phase shift around the loop is 0 or 360 degrees, then the oscillations produced by the oscillator are of growing type. The below figure shows the oscillator output with increasing amplitude of oscillations.



Exponentially Decaying Oscillations:

If modulus of A β or the magnitude of loop gain is less than unity and total phase shift around the loop is 0 or 360 degrees, then the amplitude of the oscillations decreases exponentially and finally these oscillations will cease.



Classification of oscillators

The oscillators are classified into several types based on various factors like nature of waveform, range of frequency, the parameters used, etc. The following is a broad classification of oscillators.

According to the Waveform Generated

Based on the output waveform, oscillators are classified as sinusoidal oscillators and non-sinusoidal oscillators.

Sinusoidal Oscillators: This type of oscillator generates sinusoidal current or voltages.

Non-sinusoidal Oscillators: This type of oscillators generates output, which has triangular, square, rectangle, saw tooth waveform or is of pulse shape.

According to the Circuit Components: Depends on the usage of components in the circuit, oscillators are classified into LC, RC and crystal oscillators. The oscillator using inductor and capacitor components is called as LC oscillator while the oscillator using resistance and capacitor components is called as RC oscillators. Also, crystal is used in some oscillators which are called as crystal oscillators.

According to the Frequency Generated: Oscillators can be used to produce the waveforms at frequencies ranging from low to very high levels. Low frequency or audio frequency oscillators are used to generate the oscillations at a range of 20 Hz to 100-200 KHz which is an audio frequency range.

High frequency or radio frequency oscillators are used at the frequencies more than 200-300 KHz up to gigahertz. LC oscillators are used at high frequency range, whereas RC oscillators are used at low frequency range.

Based on the Usage of Feedback

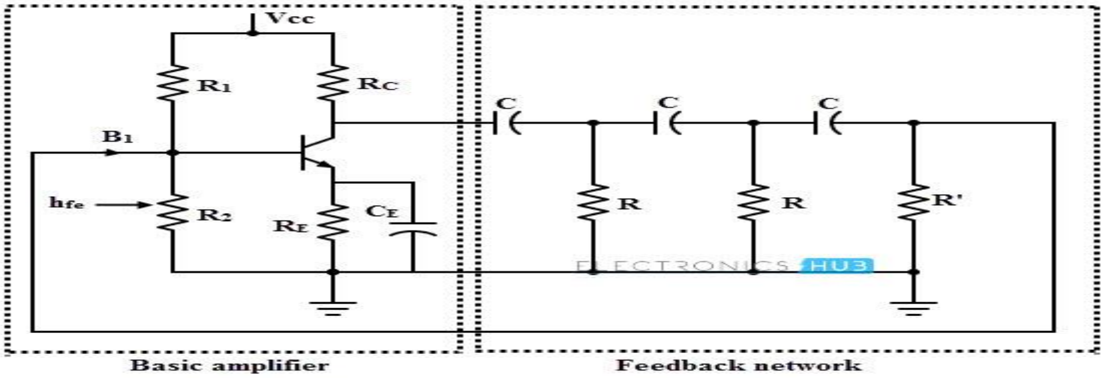
The oscillators consisting of feedback network to satisfy the required conditions of the oscillations are called as feedback oscillators. Whereas the oscillators with absence of feedback network are called as non-feedback type of oscillators.

The UJT relaxation oscillator is the example of non-feedback oscillator which uses a negative resistance region of the characteristics of the device.

Some of the sinusoidal oscillators under above categories are

- Tuned-circuits or LC feedback oscillators such as Hartley, Colpitts and Clapp etc.
- RC phase-shift oscillators such as Wein-bridge oscillator.
- Negative-resistance oscillators such as tunnel diode oscillator.
- Crystal oscillators such as Pierce oscillator.
- Heterodyne or beat-frequency oscillator (BFO).

RC Phase-shift Oscillator:



$$f = 1/(2\pi R C \sqrt{((4Rc/R) + 6)})$$

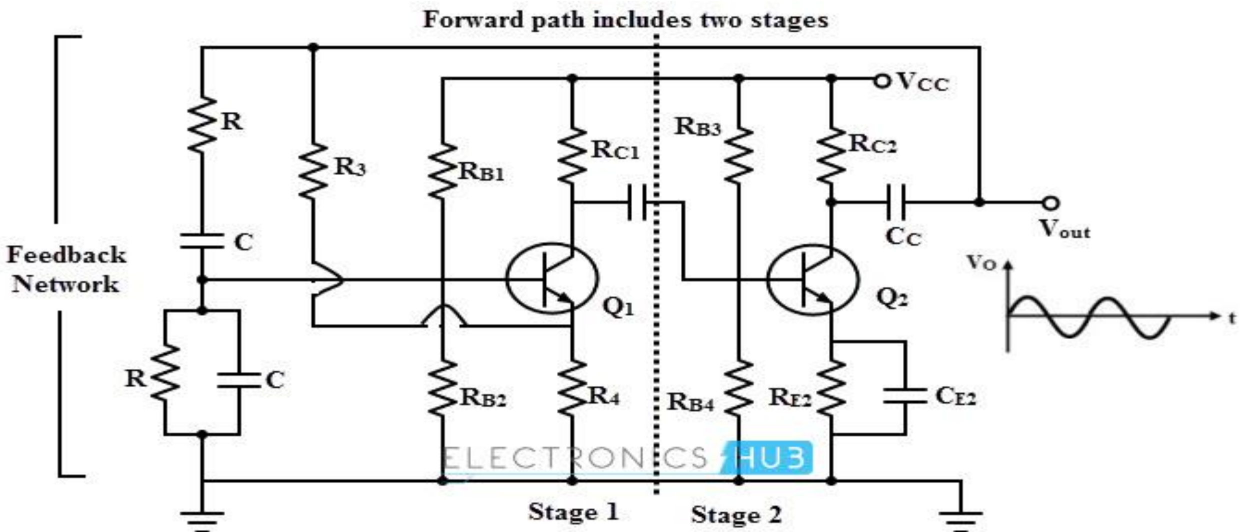
If Rc/R << 1, then

$$f = 1/(2\pi R C \sqrt{6})$$

The condition of sustained oscillations,

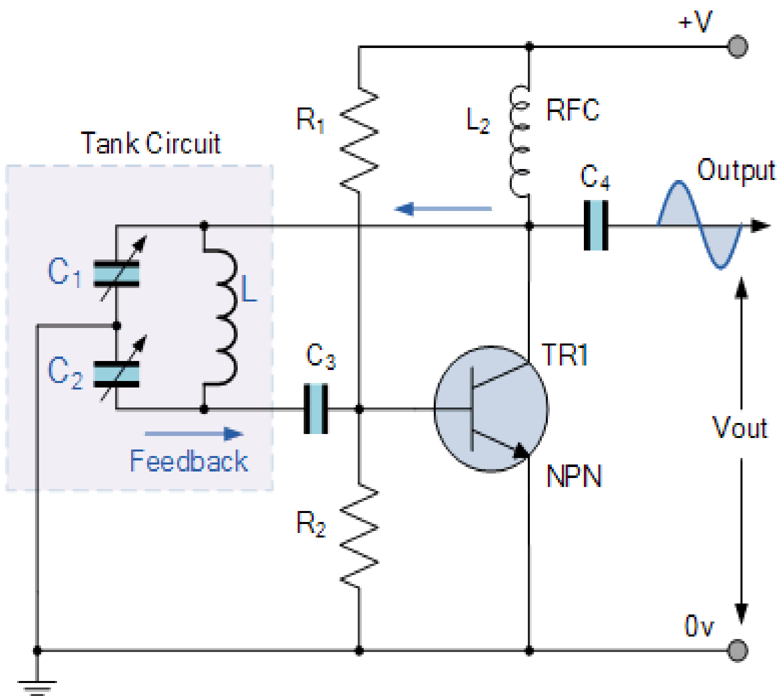
$$hfe\,(min) = (4\,Rc/R) + 23 + (29\,R/Rc)$$

Wien Bridge Oscillator:



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

Colpitts Oscillator:



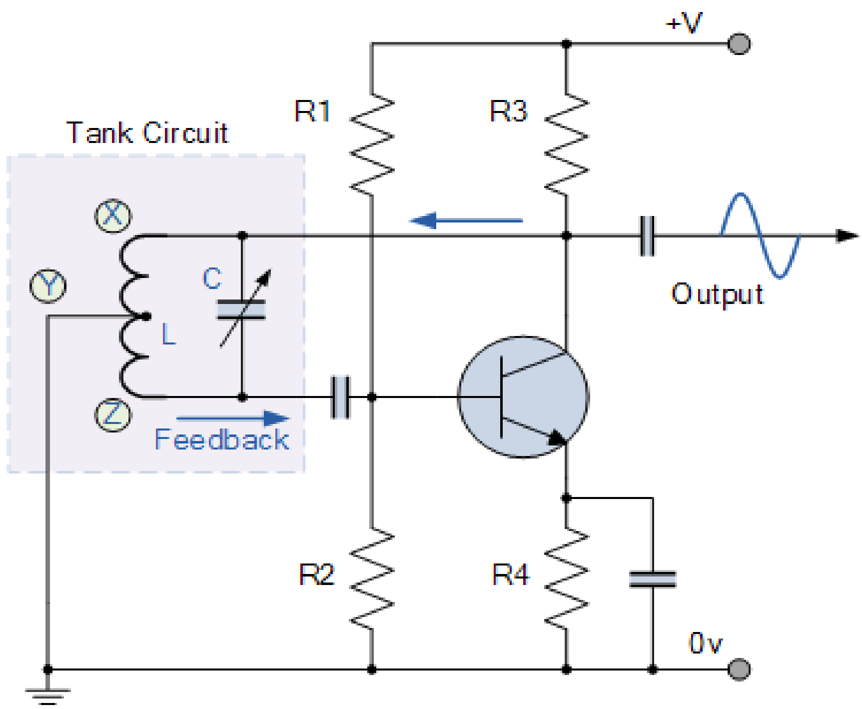
The frequency of oscillations for a Colpitts oscillator is determined by the resonant frequency of the LC tank circuit and is given as:

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

where C_T is the capacitance of C1 and C2 connected in series and is given as:

$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} \quad \text{or} \quad C_T = \frac{C_1 \times C_2}{C_1 + C_2}$$

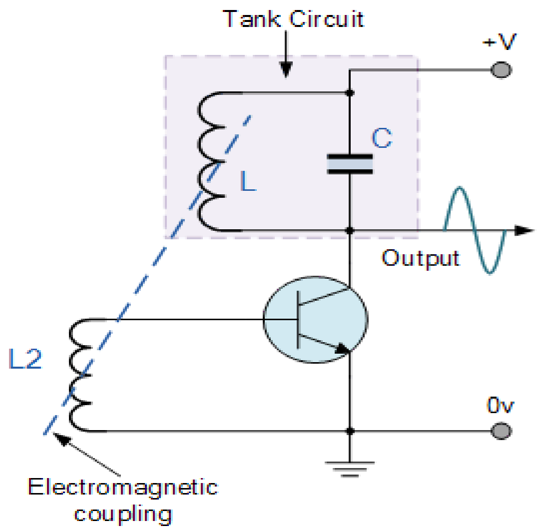
Hartley Oscillator:

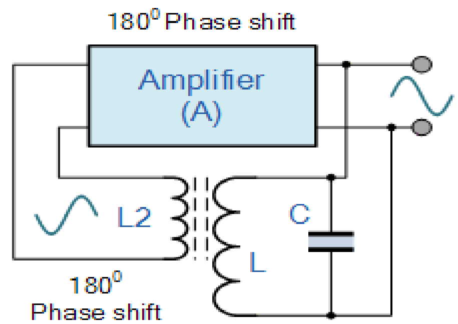


$$f = \frac{1}{2\pi\sqrt{L_T C}}$$

where: $L_T = L_1 + L_2 + 2M$

Basic Transistor LC Oscillator Circuit:





An inductance of 200mH and a capacitor of 10pF are connected together in parallel to create an LC oscillator tank circuit. Calculate the frequency of oscillation.

$$f = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{200mH \times 10pF}} = 112.5\ kHz$$

POWER AMPLIFIERS

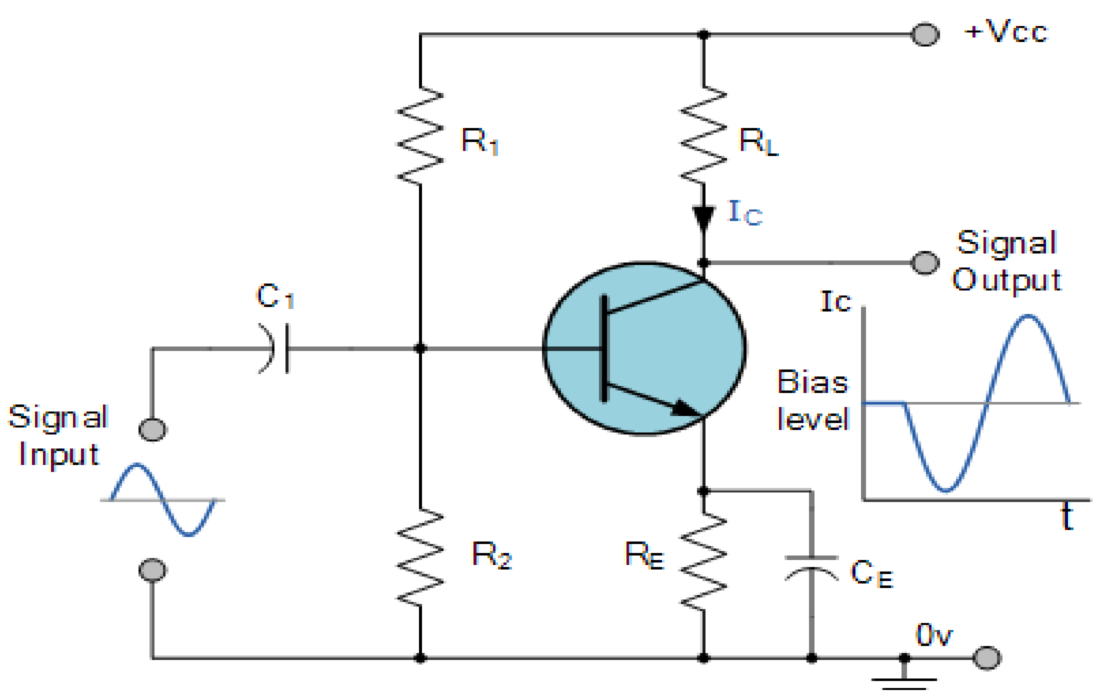
Power Amplifier:

Large input signals are used to obtain appreciable power output from amplifiers. But if the input signal is large in magnitude, the operating point is driven over a considerable portion of the output characteristic of the transistor (BJT). The transfer characteristic of a transistor which is a plot between the output current Ie and input voltage V BE is not linear. The transfer characteristic indicates the change in ic when Vb or IB is changed. For equal increments of VBE, increase in I_e will not be uniform since output characteristics are not linear (for equal increments of VBE, I_e will not increase by the same current). So the transfer characteristic is not linear. Hence because of this, when the magnitude of the input signal is very large, distortion is introduced in the output in large signal power amplifiers. To eliminate distortion in the output, push pull connection and negative feedback are employed.

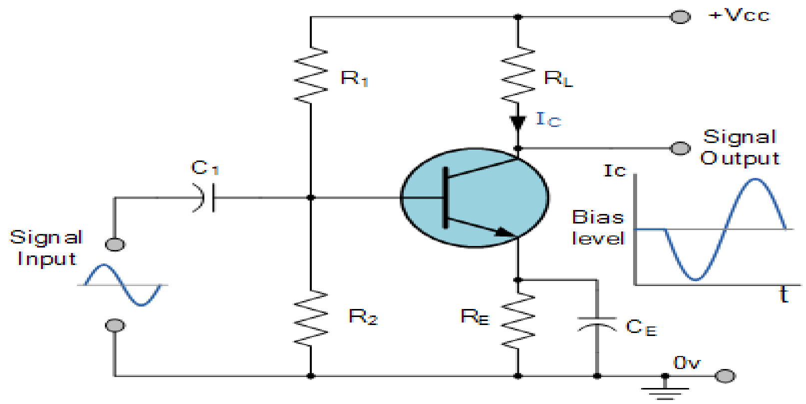
Class A Operation:

If the *Q point is placed near the centre a/the linear region a/the* dynamic curve, class A operation results. Because the transistor will conduct for the complete 360°, distortion is low for small signals and conversion efficiency is low.

Single Stage Amplifier Circuit



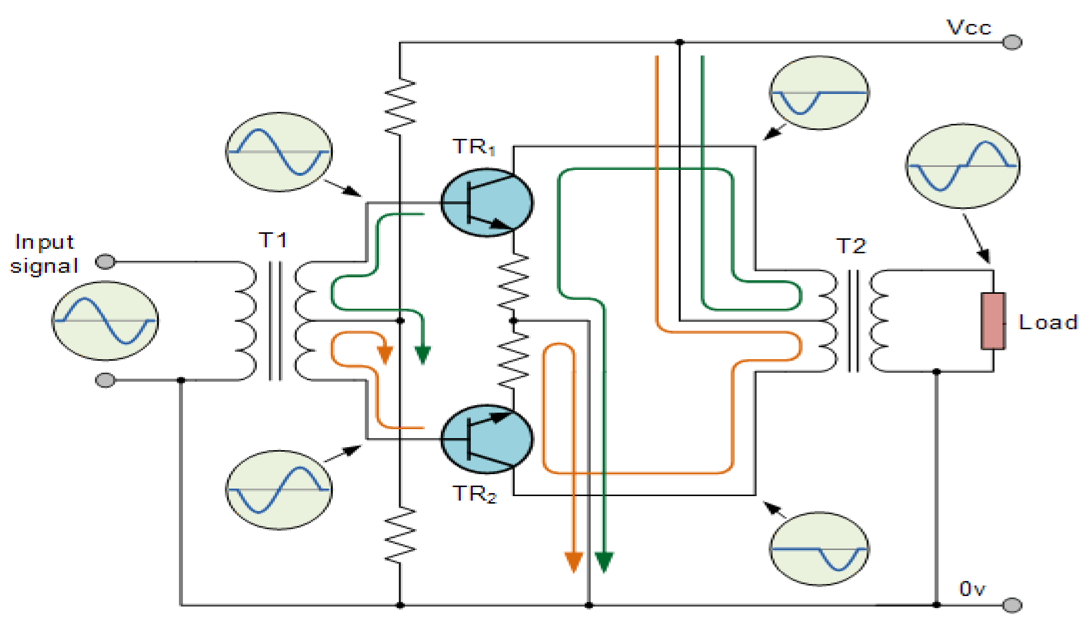
Single Stage Amplifier Circuit



Class B Operation:

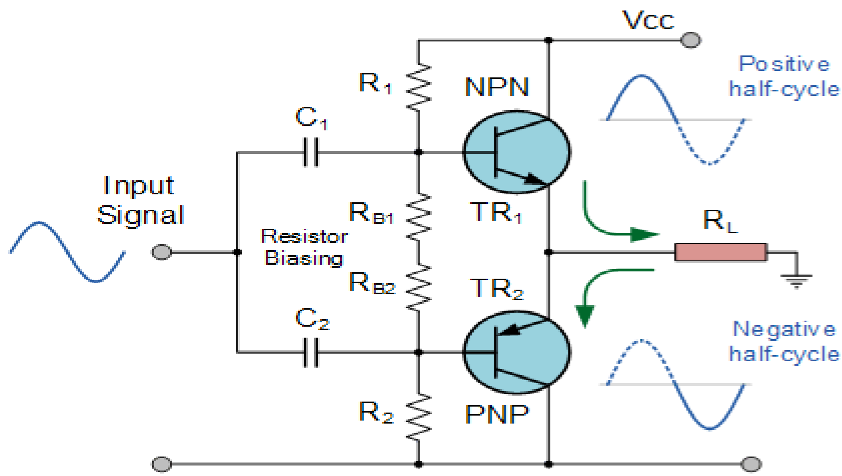
class B operation the Q point is set near cutoff. So output power will be more and conversion efficiency (ll) is more. Conduction is only for 180°, from 1t - 21t. Since the transistor Q point is beyond cutoff, the output is zero or the transistor will not conduct. Output power is more because the complete linear region is available for an operating signal excursion, resulting from one half of the input wave. The other half of input wave gives no output, because it drives the transistor below cutoff.

Class B Push-pull Transformer Amplifier Circuit

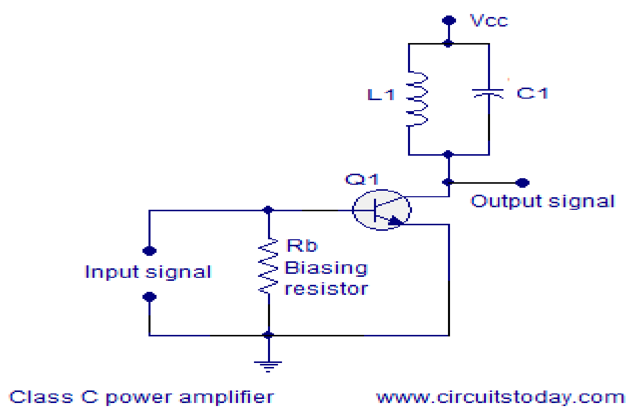


The circuit above shows a standard **Class B Amplifier** circuit

Complementary symmetry push pull amplifier



Class C Operation:



Here Q point is set well beyond cutoff and the device conducts for less than *1800.* The conversion efficiency (ll) can theoretically reach 100%. Distortion is very high. These are used in radio frequency circuits where resonant circuit may be used to filter the output waveform.

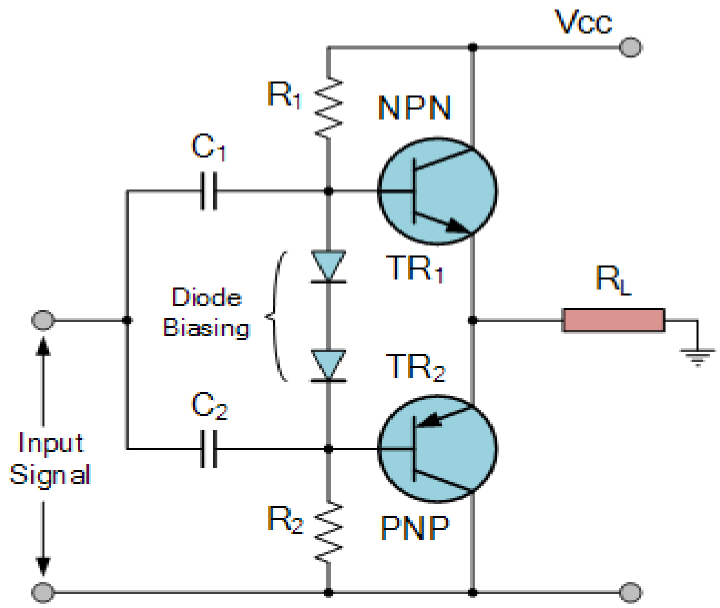
Class A and class B amplifiers are used in the audio frequency range. Class B and class C are used in Radio Frequency range where conversion efficiency is important.

Large Signal Amplifiers:

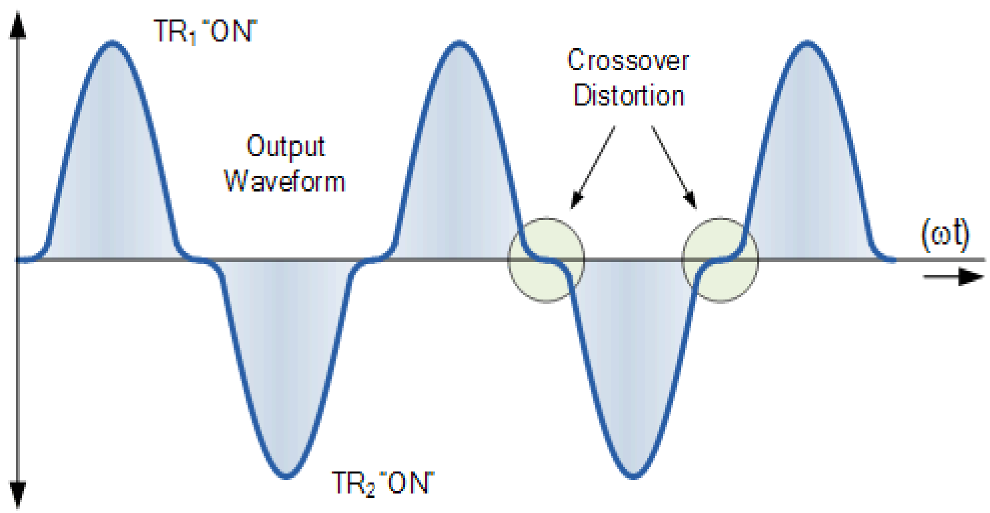
With respect to the input signal, the amplifier circuits are classified as

(i) Small signal amplifiers (ii) Large signal amplifiers

The Class AB Amplifier



Crossover Distortion Waveform



In order that there should be no distortion of the output waveform we must assume that each transistor starts conducting when its base to emitter voltage rises just above zero, but we know that this is not true because for silicon bipolar transistors the base voltage must reach at least 0.7v before the transistor starts to conduct thereby producing this flat spot. This crossover distortion effect also reduces the overall peak to peak value of the output waveform causing the maximum power output.

Tuned Amplifiers

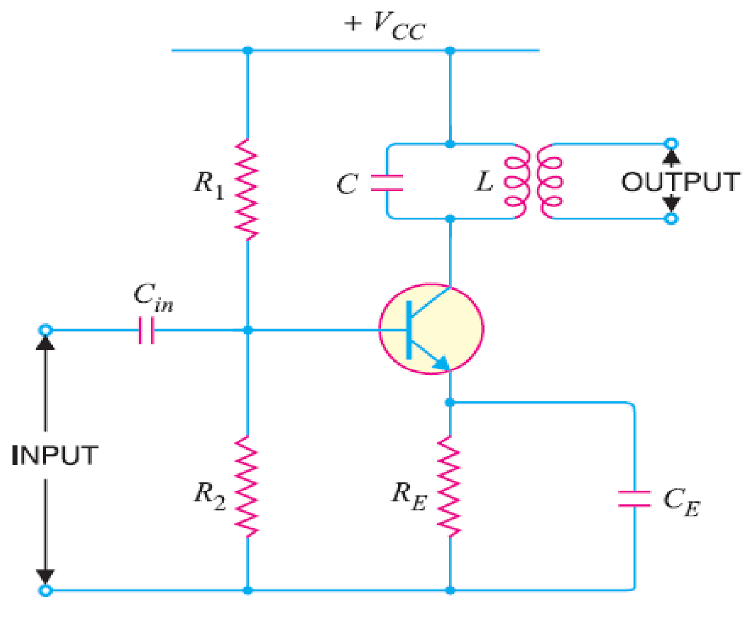
Most of the audio amplifiers we have discussed in the earlier chapters will also work at radio frequencies *i.e*. above 50 kHz. However, they suffer from two major drawbacks. First, they become less efficient at radio frequency. Secondly, such amplifiers have mostly resistive loads and consequently their gain is independent of signal frequency over a large bandwidth. In other words, an audio amplifier amplifies a wide band of frequencies equally well and does not permit the selection of a particular desired frequency while rejecting all other frequencies. However, sometimes it is desired that an amplifier should be selective *i.e*. it should select a desired frequency or narrow band of frequencies for amplification. For instance, radio and television transmission are carried on a specific radio frequency assigned to the broadcasting station. The radio receiver is required to pick up and amplify the radio frequency desired while discriminating all others. To achieve this, the simple resistive load is replaced by a parallel tuned circuit whose impedance strongly depends upon frequency. Such a tuned circuit becomes very selective and amplifies very strongly signals of resonant frequency and narrow band on either side. Therefore, the use of tuned circuits in conjunction with a transistor makes possible the selection and efficient amplification of a particular desired radio frequency. Such an amplifier is called a tuned amplifier. In this chapter, we shall focus our attention on transistor tuned amplifiers and their increasing applications in high frequency electronic circuits.

Amplifiers which amplify a specific frequency or narrow band of frequencies are called **tuned amplifiers.**

Tuned amplifiers are mostly used for the amplification of high or radio frequencies. It is because radio frequencies are generally single and the tuned circuit permits their selection and efficient amplification.

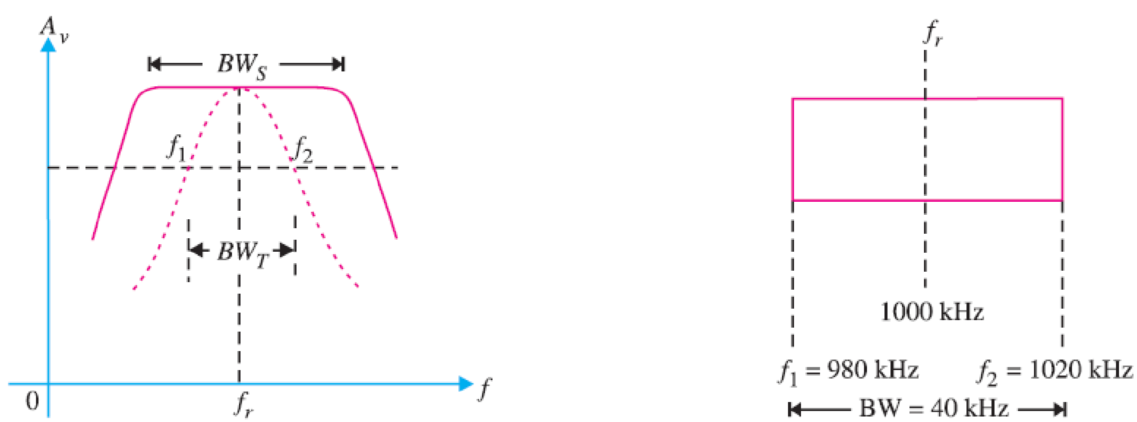
However, such amplifiers are not suitable for the amplification of audio frequencies as they are mixture of frequencies from 20 Hz to 20 kHz and not single. Tuned amplifiers are widely used in radio and television circuits where they are called upon to handle radio frequencies. Figure shows the circuit of a simple transistor tuned amplifier. Here, instead of load resistor, we have a parallel tuned circuit in the collector. The impedance of this tuned circuit strongly depends upon frequency. It offers a very high impedance at *resonant frequency* and very small impedance at all other frequencies. If the signal has the same frequency as the resonant frequency of

LC circuit, large amplification will result due to high impedance of *LC* circuit at this frequency. When signals of many frequencies are present at the input of tuned amplifier, it will select and strongly amplify the signals of resonant frequency while *rejecting all others. Therefore, such amplifiers are very useful in radio receivers to select the signal from one particular broadcasting station when signals of many other frequencies are present at the receiving aerial.



Distinction between Tuned Amplifiers and other Amplifiers:

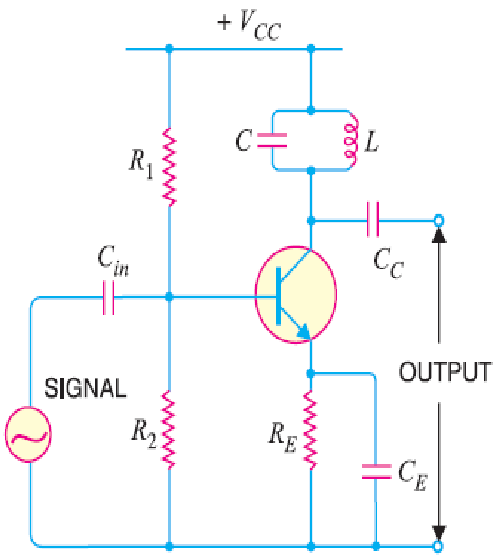
We have seen that amplifiers (*e.g.*, voltage amplifier, power amplifier *etc.*) provide the constant gain over a limited band of frequencies *i.e.*, from lower cut-off frequency *f*1 to upper cut-off frequency *f*2. Now bandwidth of the amplifier, *BW* = *f*2 −□*f*1. The reader may wonder, then, what distinguishes a tuned amplifier from other mplifiers? The difference is that tuned amplifiers are designed to have specific, usually narrow bandwidth. This point is illustrated in in Fig. 15.2. Note that *BWS* is the bandwidth of standard frequency response while *BWT* is the bandwidth of the tuned amplifier. In many applications, the narrower the bandwidth of a tuned amplifier, the better it is.

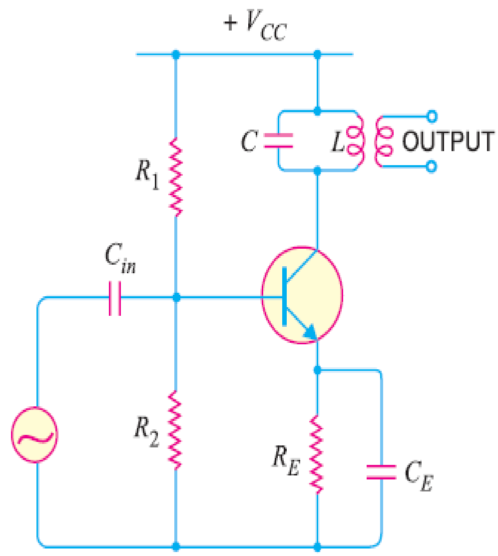


Consider a tuned amplifier that is designed to amplify only those frequencies that are within ± 20 kHz of the central frequency of 1000 kHz (*i.e.*, *fr* = 1000 kHz). Here *f*1 = 980 kHz, *fr* = 1000 kHz, *f*2 = 1020 kHz, *BW* = 40 kHz This means that so long as the input signal is within the range of 980 – 1020 kHz, it will be amplified. If the frequency of input signal goes out of this range, amplification will be drastically reduced.

Single Tuned Amplifier

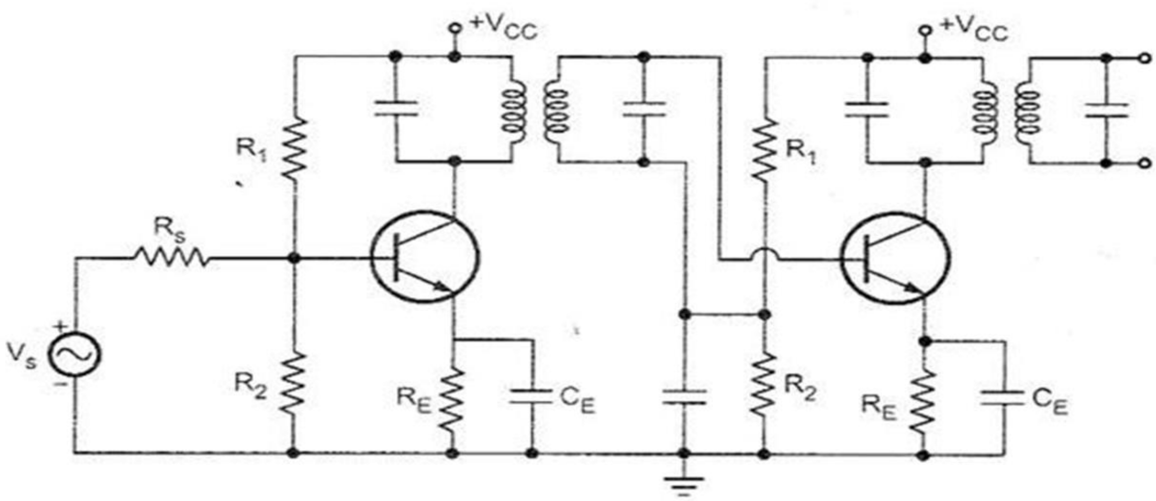
A single tuned amplifier consists of a transistor amplifier containing a parallel tuned circuit as the collector load. The values of capacitance and inductance of the tuned circuit are so selected that its resonant frequency is equal to the frequency to be amplified. The output from a single tuned amplifier can be obtained either (*a*) by a coupling capacitor *CC* as shown in Fig. (*i*) or (*b*) by a secondary coil as shown in Fig. (*ii*).





DOUBLE TUNED AMPLIFIER:

Below figure shows the 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.

STAGGER TUNED AMPLIFIER:

The double tuned amplifier gives greater 3dB 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 adjusted that they are separated by an amount equal to the bandwidth of each stage. Since 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 very sharp, projective, narrow band characteristics of synchronously tuned circuits (tuned to same resonant frequencies). Fig. 3.23 shows the relationship of amplification characteristics of individual stages in a staggered pair to the overall amplification of the two stages.

Wide Band amplifiers/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 efficiency is prime concern, class B and class C amplifiers are used at the output stages in transmitter. 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 single 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 an narrow bandwidth is desired, a resonant circuit is employed in class B and class C tuned RF power amplifiers to eliminate the harmonics.

Class B tuned amplifier

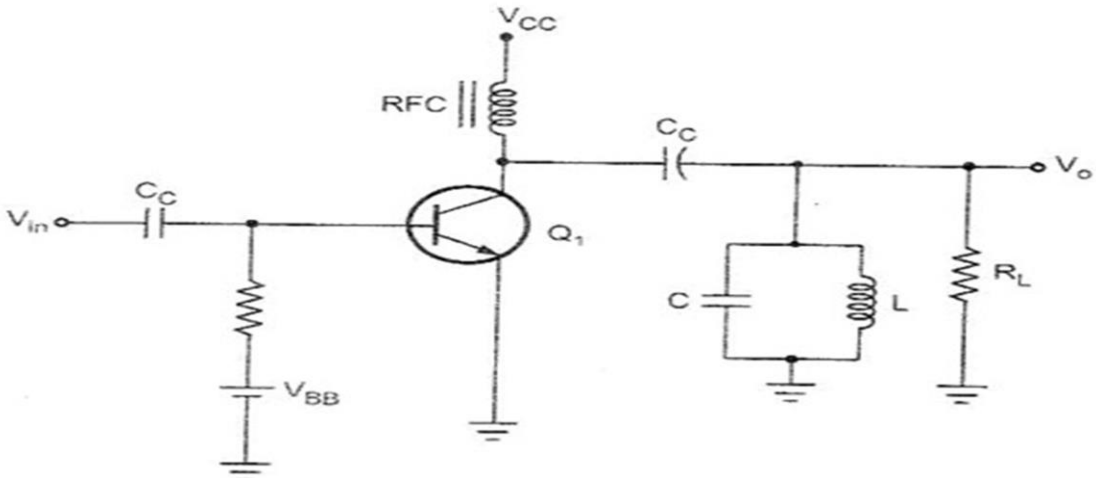


Fig. 3.25 Class B tuned amplifier