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Semester-IV

UNIT I & UNIT II

NAME-

REG NO-

MOHAMED SATHAK

A J COLLEGE OF ENGINEERING

SIRUSERI IT PARK, OMR, CHENNAI - 603103

EC8452 - ELECTRONIC CIRCUITS II

UNIT I FEEDBACK AMPLIFIERS AND STABILITY

Feedback Concepts – gain with feedback – effect of feedback on gain stability, distortion, bandwidth, input and output impedances; topologies of feedback amplifiers – analysis of series-series, shunt-shunt andshunt-series feedback amplifiers-stability problem-Gain and Phase-margins-Frequency compensation.

UNIT II OSCILLATORS

Barkhausen criterion for oscillation – phase shift, Wien bridge - Hartley & Colpitt‘s oscillators – Clapp oscillator-Ring oscillators and crystal oscillators – oscillator amplitude stabilization.

UNIT III TUNED AMPLIFIERS

Coil losses, unloaded and loaded Q of tank circuits, small signal tuned amplifiers –Analysis of capacitor coupled single tuned amplifier – double tuned amplifier - effect of cascading single tuned and double tuned amplifiers on bandwidth – Stagger tuned amplifiers - Stability of tuned amplifiers – Neutralization - Hazeltine neutralization method.

UNIT IV WAVE SHAPING AND MULTIVIBRATOR CIRCUITS

Pulse circuits – attenuators – RC integrator and differentiator circuits – diode clampers and clippers – Multivibrators - Schmitt Trigger- UJT Oscillator.

UNIT V POWER AMPLIFIERS AND DC CONVERTERS

Power amplifiers- class A-Class B-Class AB-Class C-Power MOSFET-Temperature Effect- Class AB Power amplifier using MOSFET –DC/DC convertors – Buck, Boost, Buck-Boost analysis and design

TEXT BOOKS:
1. Sedra and Smith, ―Micro Electronic Circuits‖; Sixth Edition, Oxford University Press, 2011. (UNIT I, III,IV,V)
2. Jacob Millman, ‗Microelectronics‘, McGraw Hill, 2nd Edition, Reprinted, 2009. (UNIT I,II,IV,V)
REFERENCES:
1. Robert L. Boylestad and Louis Nasheresky, ―Electronic Devices and Circuit Theory‖, 10th Edition, Pearson Education / PHI, 2008
2. David A. Bell, ―Electronic Devices and Circuits‖, Fifth Edition, Oxford University Press, 2008.
3. Millman J. and Taub H., ―Pulse Digital and Switching Waveforms‖, TMH, 2000. 4. Millman and Halkias. C., Integrated Electronics, TMH, 2007.

UNIT – I FEED BACK AMPLIFIERS

A practical amplifier has a gain of nearly one million i.e. its output is one million times the input. Consequently, even a casual disturbance at the input will appear in the amplified form in the output. There is a strong tendency in amplifiers to introduce hum due to sudden temperature changes or stray electric and magnetic fields. Therefore, every high gain amplifier tends to give noise along with signal in its output. The noise in the output of an amplifier is undesirable and must be kept to as small a level as possible. The noise level in amplifiers can be reduced considerably by the use of negative feedback i.e. by injecting a fraction of output in phase opposition to the input signal. The object of this chapter is to consider the effects and methods of providing negative feedback in transistor amplifiers.

1.1 Feedback

The process of injecting a fraction of output energy of some device back to the input is known as **feedback**. The principle of feedback is probably as old as the invention of first machine but it is only some 50 years ago that feedback has come into use in connection with electronic circuits. It has been found very useful in reducing noise in amplifiers and making amplifier operation stable. Depending upon whether the feedback energy aids or opposes the input signal, there are two basic types of feedback in amplifiers viz positive feedback and negative feedback.

(i) Positive feedback. When the feedback energy (voltage or current) is in phase with the input signal and thus aids it, it is called positive feedback. This is illustrated in Fig. 1.1. Both amplifier and feedback network introduce a phase shift of 180°. The result is a 360° phase shift around the loop, causing the feedback voltage Vf to be in phase with the input signal Vin.

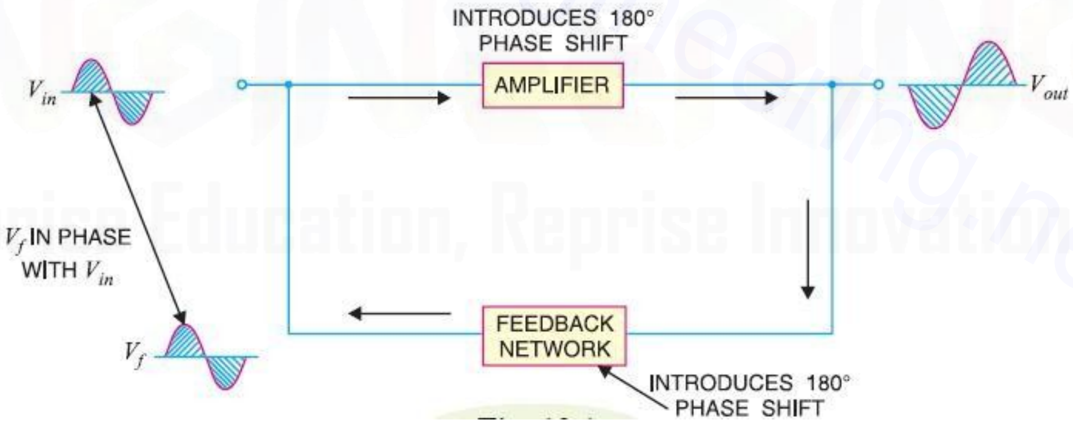


Figure 1.1

The positive feedback increases the gain of the amplifier. However, it has the disadvantages of increased distortion and instability. Therefore, positive feedback is seldom employed in amplifiers. One important use of positive feedback is in oscillators. As we shall see in the next chapter, if positive feedback is sufficiently large, it leads to oscillations. As a matter of fact, an oscillator is a device that converts d.c. power into a.c. power of any desired frequency.

(ii) Negative feedback. When the feedback energy (voltage or current) is out of phase with the input signal and thus opposes it, it is called negative feedback. This is illustrated in Fig. 1.2. As you can see, the amplifier introduces a phase shift of 180° into the circuit while the feedback network is so designed that it introduces no phase shift (i.e., 0° phase shift). The result is that the feedback voltage Vf is 180° out of phase with the input signal Vin.

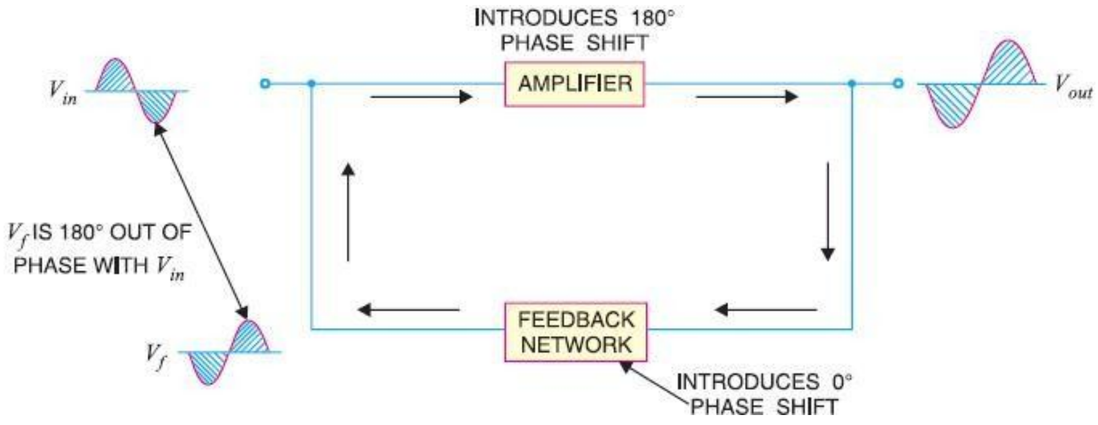


Figure 1.2

Negative feedback reduces the gain of the amplifier. However, the advantages of negative feedback are: reduction in distortion, stability in gain, increased bandwidth and improved input and output impedances. It is due to these advantages that negative feedback is frequently employed in amplifiers.

1.2 Principles of Negative Voltage Feedback In Amplifiers

A feedback amplifier has two parts viz an amplifier and a feedback circuit. The feedback circuit usually consists of resistors and returns a fraction of output energy back to the input. Fig. 1.3 *shows the principles of negative voltage feedback in an amplifier. Typical values have been assumed to make the treatment more illustrative. The output of the amplifier is 10 V. The fraction mv of this output i.e. 100 mV is fedback to the input where it is applied in series with the input signal of 101 mV. As the feedback is negative, therefore, only 1 mV appears at the input terminals of the amplifier. Referring to Fig. 1.3, we have, Gain of amplifier without feedback,

$$A_v = \frac{10 \text{ V}}{1 \text{ mV}} = 10,000$$

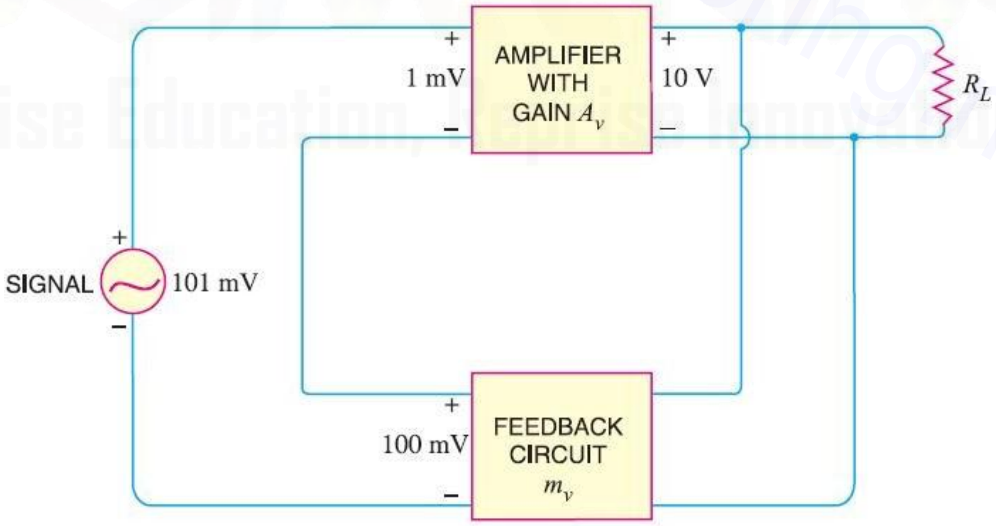


Figure 1.3

$$\text{Fraction of output voltage fedback, } m_v = \frac{100 \text{ mV}}{10 \text{ V}} = 0.01$$

$$\text{Gain of amplifier with negative feedback, } A_{vf} = \frac{10 \text{ V}}{101 \text{ mV}} = 100$$

The following points are worth noting :

- ✓ When negative voltage feedback is applied, the gain of the amplifier is reduced. Thus, the gain of above amplifier without feedback is 10,000 whereas with negative feedback, it is only 100.
- ✓ When negative voltage feedback is employed, the voltage actually applied to the amplifier is extremely small. In this case, the signal voltage is 101 mV and the negative feedback is 100 mV so that voltage applied at the input of the amplifier is only 1 mV.
- ✓ In a negative voltage feedback circuit, the feedback fraction mv is always between 0 and 1.
- ✓ The gain with feedback is sometimes called closed-loop gain while the gain without feedback is called open-loop gain. These terms come from the fact that amplifier and feedback circuits form a “loop”. When the loop is “opened” by disconnecting the feedback circuit from the input, the amplifier's gain is Av, the “open-loop” gain. When the loop is “closed” by connecting the feedback circuit, the gain decreases to Avf , the “closed-loop” gain.

1.3 Gain of Negative Voltage Feedback Amplifier

$$(e_g - m_v e_0) A_v = e_0$$

or $$A_v e_g - A_v m_v e_0 = e_0$$

or $$e_0 (1 + A_v m_v) = A_v e_g$$

or $$\frac{e_0}{e_g} = \frac{A_v}{1 + A_v m_v}$$

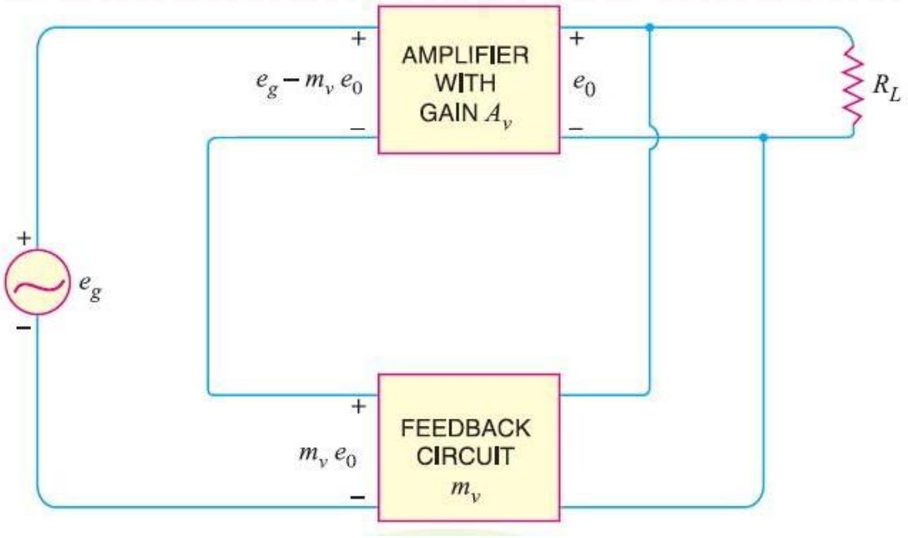


Figure 1.4

But e0/eg is the voltage gain of the amplifier with feedback.

Voltage gain with negative feedback is

$$A_{vf} = \frac{A_v}{1 + A_v\ m_v}$$

It may be seen that the gain of the amplifier without feedback is Av. However, when negative voltage feedback is applied, the gain is reduced by a factor 1 + Av mv. It may be noted that negative voltage feedback does not affect the current gain of the circuit.

1.4 Advantages of Negative Voltage Feedback

The following are the advantages of negative voltage feedback in amplifiers :

(i) Gain stability. An important advantage of negative voltage feedback is that the resultant gain of the amplifier can be made independent of transistor parameters or the supply voltage variations.

$$A_{vf} = \frac{A_v}{1 + A_v\ m_v}$$

For negative voltage feedback in an amplifier to be effective, the designer deliberately makes the product Av mv much greater than unity. Therefore, in the above relation, 1 can be neglected as compared to Av mv and the expression becomes :

$$A_{vf} = \frac{A_v}{A_v\ m_v} = \frac{1}{m_v}$$

It may be seen that the gain now depends only upon feedback fraction mv i.e., on the characteristics of feedback circuit. As feedback circuit is usually a voltage divider (a resistive network), therefore, it is unaffected by changes in temperature, variations in transistor parameters and frequency. Hence, the gain of the amplifier is extremely stable.

(ii) Reduces non-linear distortion. A large signal stage has non-linear distortion because its voltage gain changes at various points in the cycle. The negative voltage feedback reduces the nonlinear distortion in

$$D_{vf} = \frac{D}{1 + A_v\ m_v}$$

where D = distortion in amplifier without feedback

D_{vf} = distortion in amplifier with negative feedback

It is clear that by applying negative voltage feedback to an amplifier, distortion is reduced by a factor 1 + Av mv.

(iii) Improves frequency response. As feedback is usually obtained through a resistive network, therefore, voltage gain of the amplifier is *independent of signal frequency. The result is that voltage gain of the amplifier will be substantially constant over a wide range of signal frequency. The negative voltage feedback, therefore, improves the frequency response of the amplifier.

(iv) Increases circuit stability. The output of an ordinary amplifier is easily changed due to variations in ambient temperature, frequency and signal amplitude. This changes the gain of the amplifier, resulting in distortion. However, by applying negative voltage feedback, voltage gain of the amplifier is stabilised or accurately fixed in value.

This can be easily explained. Suppose the output of a negative voltage feedback amplifier has increased because of temperature change or due to some other reason. This means more negative feedback since feedback is being given from the output. This tends to oppose the increase in amplification and maintains it stable. The same is true should the output voltage decrease. Consequently, the circuit stability is considerably increased.

(v) Increases input impedance and decreases output impedance. The negative voltage feedback increases the input impedance and decreases the output impedance of amplifier. Such a change is profitable in practice as the amplifier can then serve the purpose of impedance matching.

(a) Input impedance. The increase in input impedance with negative voltage feedback can be explained by referring to Fig. 13.5. Suppose the input impedance of the amplifier is Zin without feedback and Z ′in with negative feedback. Let us further assume that input current is i1. Referring to Fig. 13.5, we have,

$$e_g - m_v e_0 = i_1 Z_{in}$$

Now

$$\begin{aligned} e_g &= (e_g - m_v e_0) + m_v e_0 \\ &= (e_g - m_v e_0) + A_v m_v (e_g - m_v e_0) \qquad [\because e_0 = A_v (e_g - m_v e_0)] \\ &= (e_g - m_v e_0)(1 + A_v m_v) \\ &= i_1 Z_{in} (1 + A_v m_v) \qquad [\because e_g - m_v e_0 = i_1 Z_{in}] \end{aligned}$$

or

$$\frac{e_g}{i_1} = Z_{in} (1 + A_v m_v)$$

But eg/i1 = Z ′i n , the input impedance of the amplifier with negative voltage

$$Z'_{in} = Z_{in} (1 + A_v m_v)$$

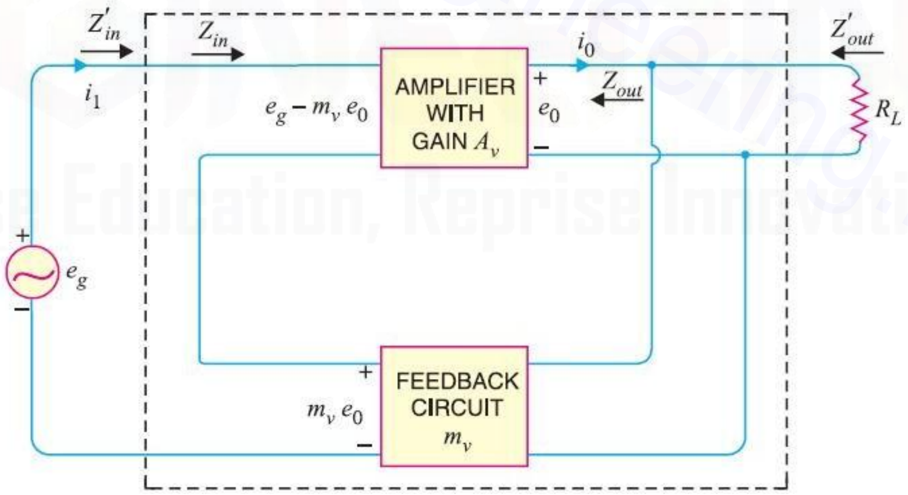


Figure 1.5

It is clear that by applying negative voltage feedback, the input impedance of the amplifier is increased by a factor 1 + Av mv. As Av mv is much greater than unity, therefore, input impedance is increased considerably. This is an advantage, since the amplifier will now present less of a load to its source circuit.

(b) Output impedance. Following similar line, we can show that output impedance with negative voltage feedback is given by :

$$Z'_{out} = \frac{Z_{out}}{1 + A_v\, m_v}$$

where Z'_{out} = output impedance with negative voltage feedback

Z_{out} = output impedance without feedback

It is clear that by applying negative feedback, the output impedance of the amplifier is decreased by a factor 1 + Av mv. This is an added benefit of using negative voltage feedback. With lower value of output impedance, the amplifier is much better suited to drive low impedance loads.

1.5 Feedback Circuit

The function of the feedback circuit is to return a fraction of the output voltage to the input of the amplifier. Fig. 13.6 shows the feedback circuit of negative voltage feedback amplifier. It is essentially a potential divider consisting of resistances R1 and R2. The output voltage of the amplifier is fed to this potential divider which gives the feedback voltage to the input. Referring to Fig. 13.6, it is clear that :

$$\text{Voltage across } R_1 = \left(\frac{R_1}{R_1 + R_2}\right) e_0$$

$$\text{Feedback fraction, } m_v = \frac{\text{Voltage across } R_1}{e_0} = \frac{R_1}{R_1 + R_2}$$

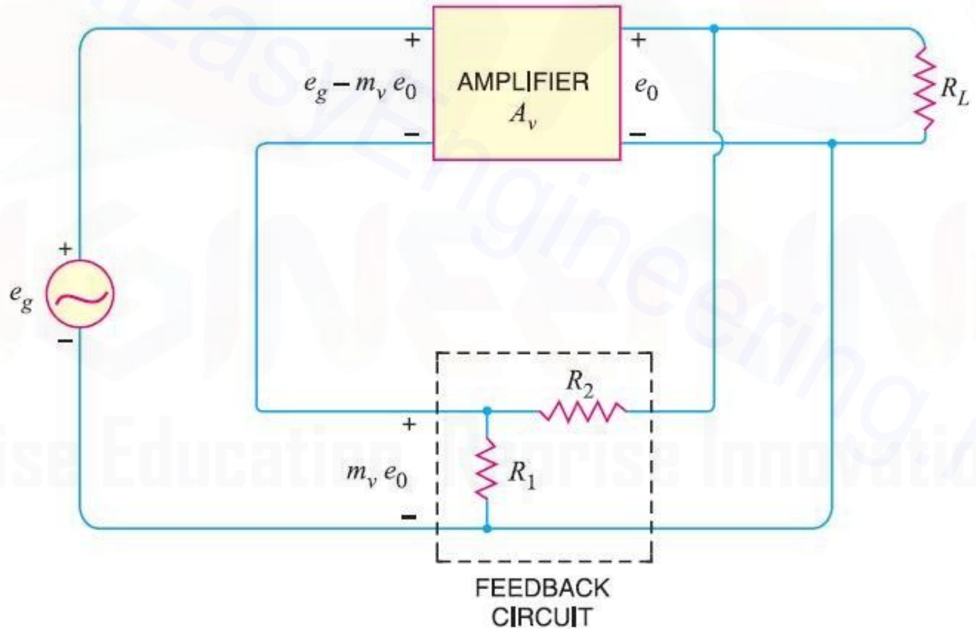


Figure 1.6

1.6 Principles of Negative Current Feedback

In this method, a fraction of output current is fedback to the input of the amplifier. In other words, the feedback current (If) is proportional to the output current (Iout) of the amplifier. Fig. 1.7 shows the principles of negative current feedback. This circuit is called current-shunt feedback circuit. A feedback resistor Rf is connected between input and output of the amplifier. This amplifier has a current gain of Ai without feedback. It means that a current I1 at the input terminals of the amplifier will appear as Ai I1 in the output circuit i.e., Iout = Ai I1.

Now a fraction mi of this output current is fedback to the input through Rf. The fact that arrowhead shows the feed current being fed forward is because it is negative feedback.

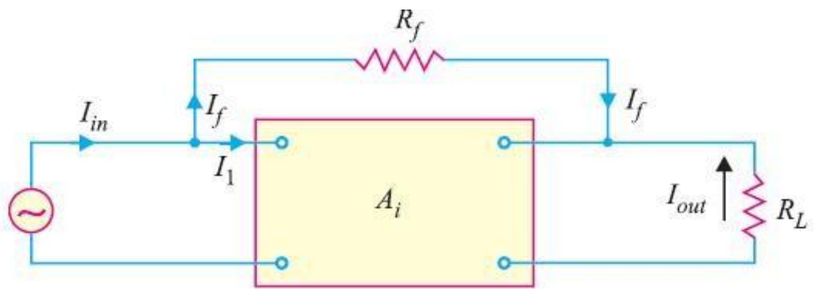


Figure 1.7

Feedback current, If = mi Iout

Note that negative current feedback reduces the input current to the amplifier and hence its current gain.

1.7 Current Gain with Negative Current Feedback

Referring to Fig. 13.6, we have,

Iin = I1 + If = I1 + mi Iout

But Iout = Ai I1, where Ai is the current gain of the amplifier without feedback.

Iin = I1 + mi Ai I1 (ä Iout = Ai I1)

$$A_{if} = \frac{I_{out}}{I_{in}} = \frac{A_i I_1}{I_1 + m_i A_i I_1}$$

$$\text{or} \quad A_{if} = \frac{A_i}{1 + m_i A_i}$$

This equation looks very much like that for the voltage gain of negative voltage feedback amplifier. The only difference is that we are dealing with current gain rather than the voltage gain.

The following points may be noted carefully :

(i) The current gain of the amplifier without feedback is Ai. However, when negative current feedback is applied, the current gain is reduced by a factor (1 + mi Ai).

(ii) The feedback fraction (or current attenuation) mi has a value between 0 and 1.

(iii) The negative current feedback does not affect the voltage gain of the amplifier.

1.8 Effects of Negative Current Feedback

The negative current feedback has the following effects on the performance of amplifiers :

(i) Decreases the input impedance. The negative current feedback decreases the input impedance of most amplifiers.

Let

Zin = Input impedance of the amplifier without feedback

Z ′in = Input impedance of the amplifier with negative current feedback

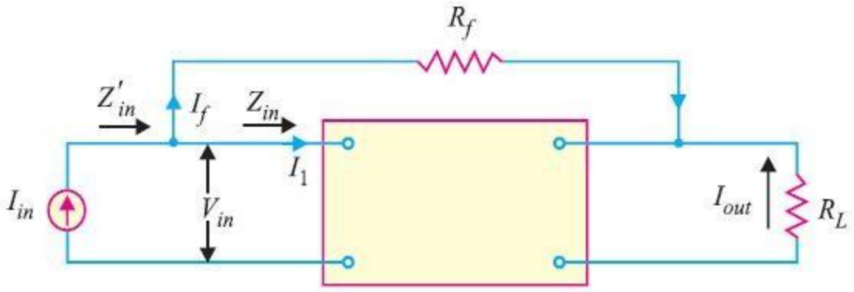


Figure 1.8

Referring to Fig. 1.8, we have,

where

Zout = output impedance of the amplifier without feedback

Z ′out = output impedance of the amplifier with negative current feedback

The reader may recall that with negative voltage feedback, the output impedance of the amplifier is decreased.

Increases bandwidth. It can be shown that with negative current feedback, the bandwidth of the amplifier is increased by the factor (1 + mi Ai).

BW′ = BW (1 + mi Ai)

where

BW = Bandwidth of the amplifier without feedback

BW′ = Bandwidth of the amplifier with negative current feedback

1.9 Emitter Follower

It is a negative current feedback circuit. The emitter follower is a current amplifier that has no voltage gain. Its most important characteristic is that it has high input impedance and low output impedance. This makes it an ideal circuit for impedance matching.

Circuit details. Fig. 1.9 shows the circuit of an emitter follower. As you can see, it differs from the circuitry of a conventional CE amplifier by the absence of collector load and emitter bypass capacitor. The emitter resistance RE itself acts as the load and a.c. output voltage (Vout) is taken across RE. The biasing is generally provided by voltage-divider method or by base resistor method. The following points are worth noting about the emitter follower :

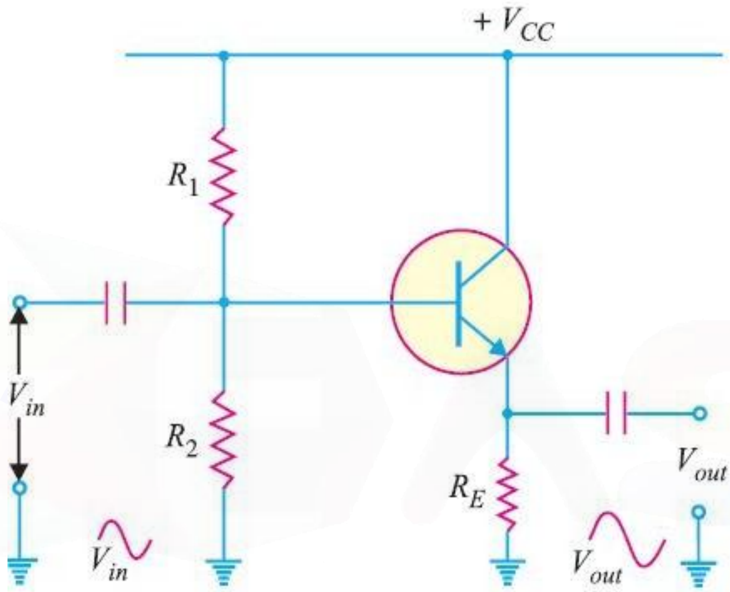


Figure 1.9

(i) There is neither collector resistor in the circuit nor there is emitter bypass capacitor. These are the two circuit recognition features of the emitter follower.

(ii) Since the collector is at ac ground, this circuit is also known as common collector (CC) amplifier.

Operation. The input voltage is applied between base and emitter and the resulting a.c. emitter current produces an output voltage ieRE across the emitter resistance. This voltage opposes the input voltage, thus providing negative feedback. Clearly, it is a negative current feedback circuit since the voltage fedback is proportional to the emitter current i.e., output current. It is called emitter follower because the output voltage follows the input voltage.

Characteristics.

The major characteristics of the emitter follower are :

(i) No voltage gain. In fact, the voltage gain of an emitter follower is close to 1.

(ii) Relatively high current gain and power gain.

(iii) High input impedance and low output impedance.

(iv) Input and output ac voltages are in phase.

1.10 D.C. Analysis of Emitter Follower

The d.c. analysis of an emitter follower is made in the same way as the voltage divider bias circuit of a CE amplifier. Thus referring to Fig. 1.9 above, we have,

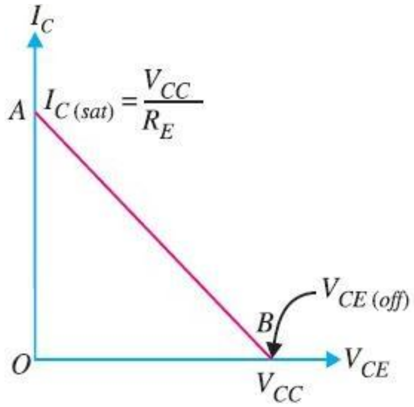


Figure 1.10

$$\text{Voltage across } R_2,\quad V_2 = \frac{V_{CC}}{R_1 + R_2} \times R_2$$

$$\text{Emitter current,}\quad I_E = \frac{V_E}{R_E} = \frac{V_2 - V_{BE}}{R_E}$$

Collector-emitter voltage, VCE = VCC − VE

D.C. Load Line. The d.c. load line of emitter follower can be constructed by locating the two end points viz., IC(sat) and VCE(off).

$$I_{C(sat)} = \frac{V_{CC}}{R_E}$$

This locates the point A (OA = VCC÷RE) of the d.c. load line as shown in Fig. 1.10.

(ii) When the transistor is cut off, IC = 0. Therefore, VCE(off) = VCC. This locates the point B (OB = VCC) of the d.c. load line.

By joining points A and B, d.c. load line AB is constructed.

13.11 Voltage Gain of Emitter Follower

Fig. 1.11 shows the emitter follower circuit. Since the emitter resistor is not bypassed by a capacitor, the a.c. equivalent circuit of emitter follower will be as shown in Fig. 1.12. The ac resistance rE of the emitter circuit is given by

$$r_E = r'_e + R_E \qquad \text{where } r'_e = \frac{25 \text{ mV}}{I_E}$$

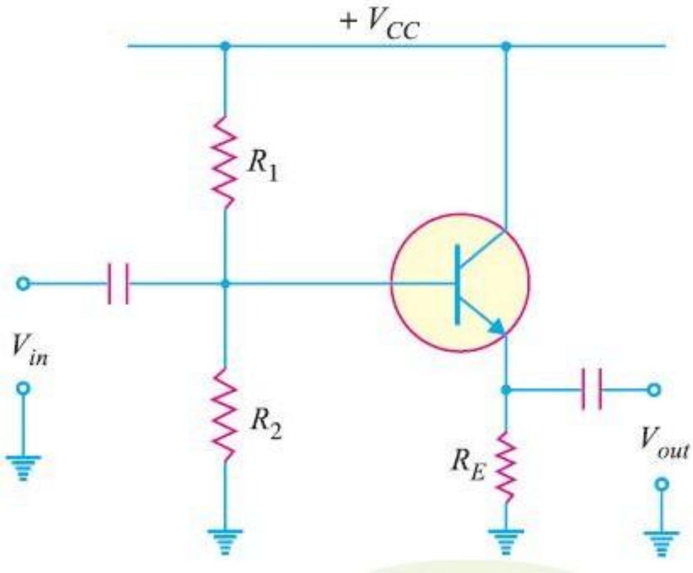


Figure 1.11

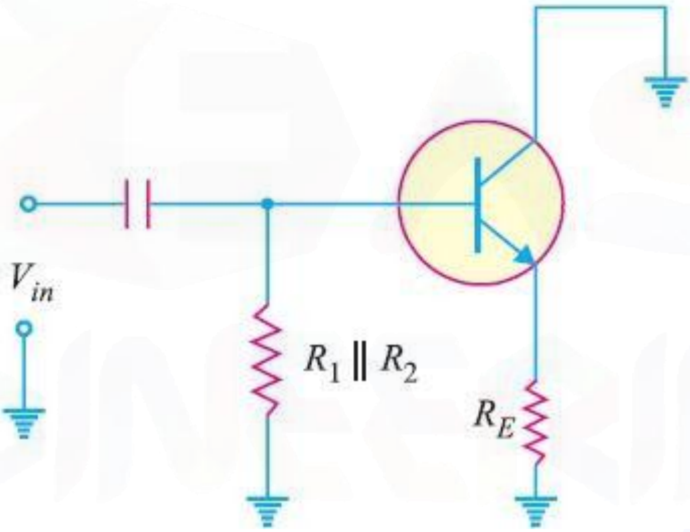


Figure 1.12

In order to find the voltage gain of the emitter follower, let us replace the transistor in Fig. 1.12 by its equivalent circuit. The circuit then becomes as shown in Fig. 1.13. Note that input voltage is applied across the ac resistance of the emitter circuit i.e., (r'e + RE). Assuming the emitter diode to be ideal,

Output voltage, Vout = ie RE

Input voltage, Vin = ie (r'e + RE)

Voltage gain of emitter follower is

$$A_v = \frac{V_{out}}{V_{in}} = \frac{i_e R_E}{i_e (r'_e + R_E)} = \frac{R_E}{r'_e + R_E}$$

$$\text{or} \qquad A_v = \frac{R_E}{r'_e + R_E}$$

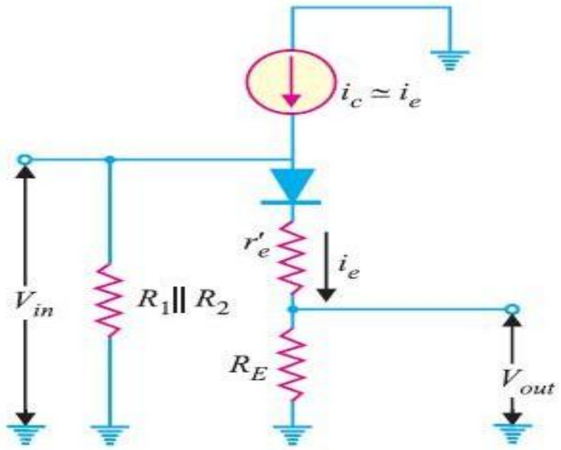


Figure 1.13

In most practical applications, RE >> r'e so that Av =1.

is

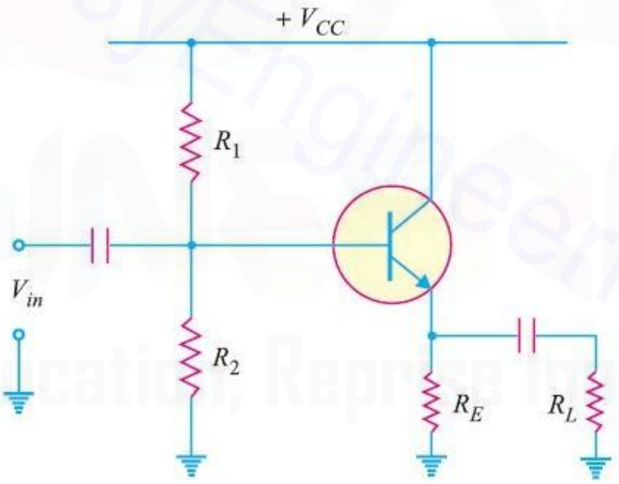


Figure 1.14 (i)

As for CE amplifier, the input impedance of emitter follower is the combined effect of biasing resistors (R1 and R2) and the input impedance of transistor base [Zin (base)]. Since these resistances are in parallel to the ac signal, the input impedance Zin of the emitter follower is given by :

$$Z_{in} = R_1 \,||\, R_2 \,||\, Z_{in(\text{base})}$$

where $$Z_{in\,(base)} = \beta\,(r'_e + R'_E)$$

Now $$r'_e = \frac{25 \text{ mV}}{I_E} \quad \text{and} \quad R'_E = R_E \,||\, R_L$$

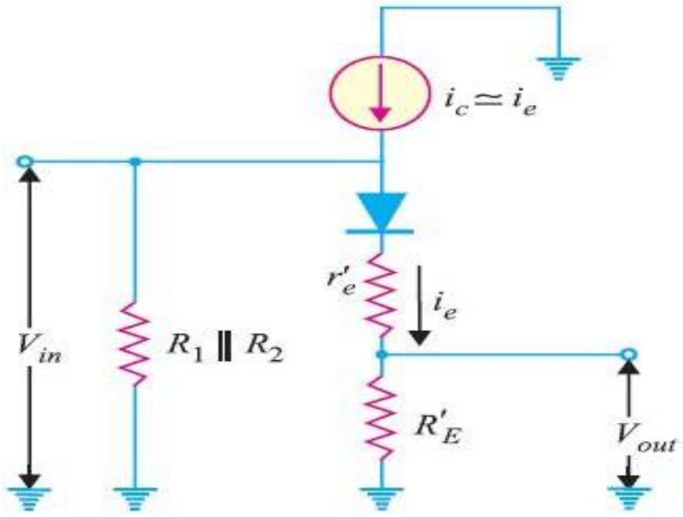


Figure 1.14 (ii)

1.13 Output Impedance of Emitter Follower

The output impedance of a circuit is the impedance that the circuit offers to the load. When load is connected to the circuit, the output impedance acts as the source impedance for the load. Fig.1.15 shows the circuit of emitter follower. Here Rs is the output resistance of amplifier voltage source. It can be proved that the output impedance

$$Z_{out} = R_E \parallel \left(r'_e + \frac{R'_{in}}{\beta} \right)$$

In practical circuits, the value of RE is large enough to be ignored. For this reason, the output impedance of emitter follower is approximately given by :

$$Z_{out} = r'_e + \frac{R'_{in}}{\beta}$$

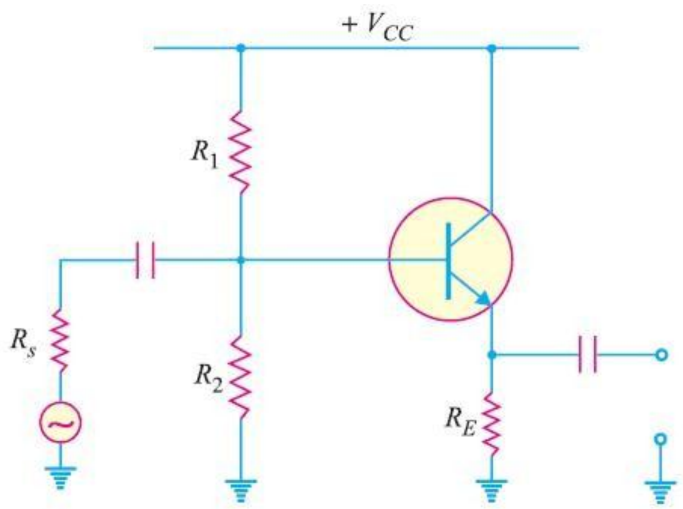


Figure 1.15

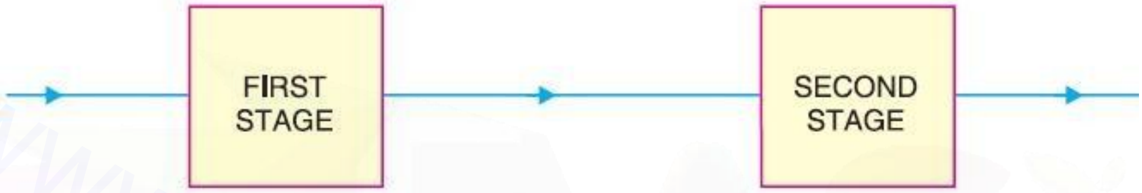
1.14 Applications of Emitter Follower

The emitter follower has the following principal applications :

(i) To provide current amplification with no voltage gain.

(ii) Impedance matching.

(i) Current amplification without voltage gain. We know that an emitter follower is a current amplifier that has no voltage gain (Aν = 1). There are many instances (especially in digital electronics) where an increase in current is required but no increase in voltage is needed. In such a situation, an emitter follower can be used. For example, consider the two stage amplifier circuit as shown in Fig. 1.16. Suppose this 2 stage amplifier has the desired voltage gain but current gain of this multistage amplifier is insufficient. In that case, we can use an emitter follower to increase the current gain without increasing the voltage gain.



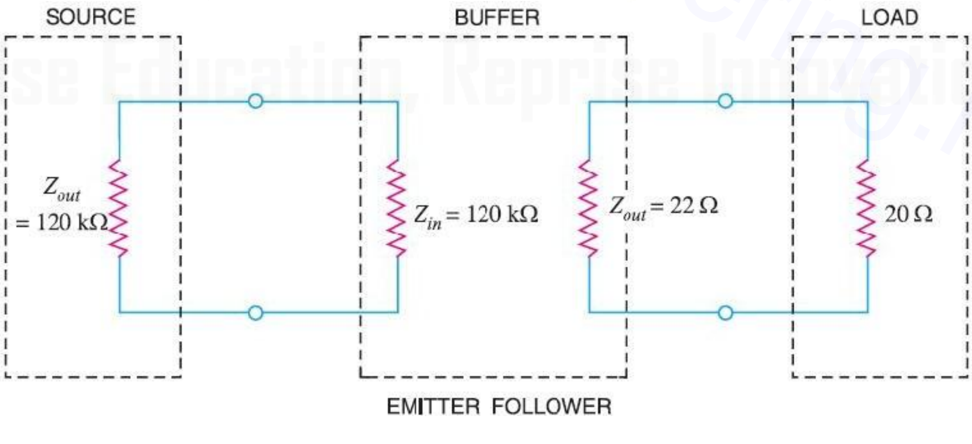


Figure 1.17

It may be noted that the job of impedance matching can also be accomplished by a transformer. However, emitter follower is preferred for this purpose. It is because emitter follower is not only more convenient than a transformer but it also has much better frequency response i.e., it works well over a large frequency range.

1.15 <u>Nyquist Criterion</u>

Criterion Of Nyquist:

- ✓ The Aβ is a function of frequency. Points in the complex plane are obtained for the values of Aβ corresponding to all values of 'f' from -∞ to ∞. The locus of all these points forms a closed curve.
- ✓ The criterion of nyquist is that amplifier is unstable if this curve encloses the point (-1+j0), and the amplifier is stable if the curve does not enclose this point.

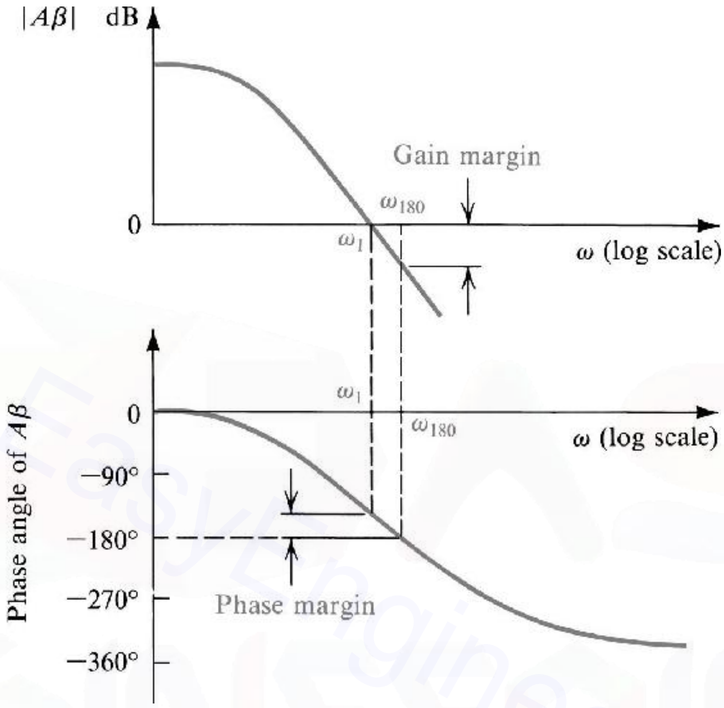


Figure 1.18 Nyquist Plot

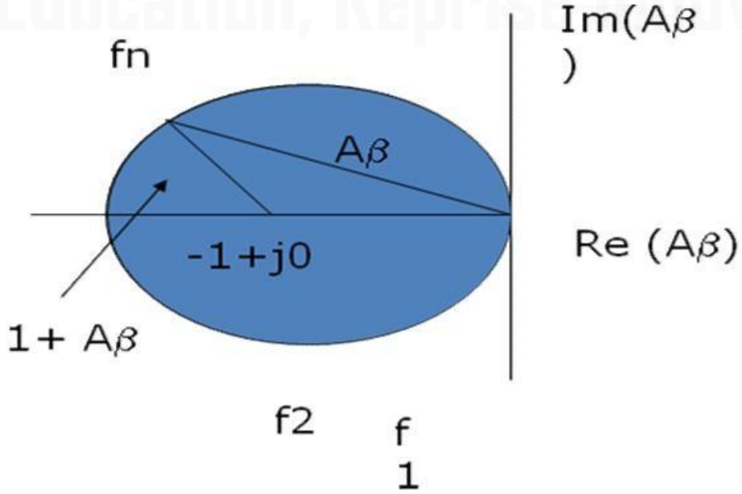


Figure 1.19 Locus of 1+Aβ is a circle of radius unity and centre (–1+j0**)**

The amplifier is unstable if this curve encloses the point –1+j0 and the amplifier is stable if the curve does not enclose this point

① Sketch the block diagram of a feedback amplifier & derive the expression for gain (i) With positive feedback & State its advantages
(ii) With negative feedback

[APR/MAY 2011 R8/R9, NOV/DEC 2012 R8] [16 M]

General Feedback structure:

Basic Concept of feedback:

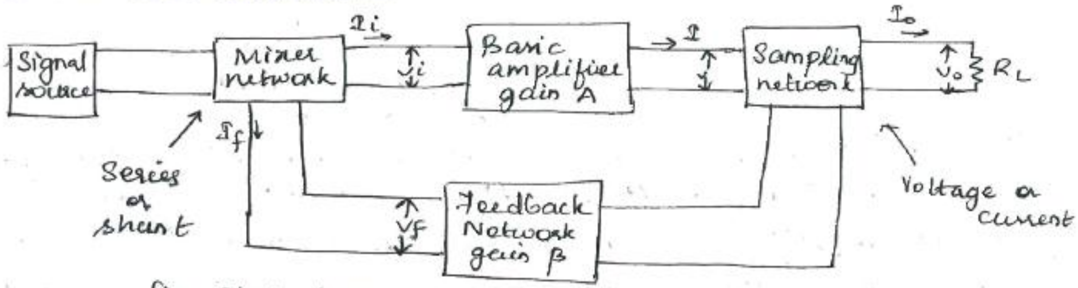


fig: Block diagram of an amplifier with feedback.

* The output quantity (V/I) is sampled by a suitable sampler which is of two types, namely voltage sampler and current sampler & fed to the feedback network

* Mixers are also known as comparator, it mixes an external source signal V_s with a fraction of an output signal from the o/p of feedback n

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* Mixer is of 2 types, Series mixer & shunt mixer.

Gain of Basic amplifier $A = \frac{V_o}{V_i}$

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

Gain of the feedback amplifier $A_f = \frac{V_o}{V_s}$ where V_s - AC signal at the input side (I/V)
V_f - feedback signal (I/V.

There are two types of feedback
(i) Positive feedback
(ii) Negative feedback

Positive feedback:

If the feedback signal V_f is in phase with input signal V_s, then the net voltage $V_i = V_s + V_f$. Hence the input voltage applied to the basic amplifier is increased thereby increasing V_o exponentially. This type of feedback is said to be positive or regenerative feedback. Gain of the amplifier with positive feedback is given by $A_f = \frac{V_o}{V_s}$

$$= \frac{V_o}{V_i - V_f}$$

$$= \frac{1}{\frac{V_i}{V_o} - \frac{V_f}{V_o}} = \frac{1}{\frac{1}{A} - \beta} \Rightarrow \boxed{\frac{A}{1-A\beta} = A_f}$$

Here $|A_f| > |A|$. The product of the open gain & the feedback is called the loop gain = $A\beta$. If $|A\beta| = 1$, then $A\beta = \infty$. Hence the gain of the amplifier with positive feedback is infinite & the amplifier gives an A.C. output without A.C input signal. Thus the amplifier acts as an oscillator.

Adv. & Disadv.

* The positive feedback
  - increases the instability of an amplifier,
  - reduces the bandwidth
  - increases the distortion & noise

This property of positive feedback is utilized in oscillators.

Negative feedback:

If the feedback signal V_f is out of phase with input signal V_s, then $V_i = V_s - V_f$. Hence the input voltage applied to the basic amplifier is decreased & correspondingly the o/p is decrea[sed]. Hence, the voltage gain is reduced. This type of feedback is known as negative or degenerative feedback. Gain of the amplifier with negative feedback is $A_f = \frac{V_o}{V_s} = \frac{V_o}{V_i + V_f}$

$$A_f = \frac{1}{\frac{V_i}{V_o} + \frac{V_f}{V_o}} = \frac{1}{1/A + \beta}$$

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

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Here $|A_f| < |A|$. If $|A\beta| >> 1$, then $A_f = \frac{1}{\beta}$, where β is a feedback ratio

- Gain depends entirely on the feedback network. If the feedback n/w contains only stable passive elements, the gain of the amplifier using negative feedback is also stable.

Adv.

Negative feedback is used to improve the performance of an electron amplifier

- always helps to increase the Bandwidth.
- decrease distortion & noise.
- helps to modify input & output resistances as desired

(ii) An amplifier, without feedback, has a voltage gain of 400, lower cut off frequency $f_1 = 50Hz$, upper cut off frequency $f_2 = 200kHz$ & a distortion of 10%. Determine the amplifier voltage gain, lower cut-off frequency & upper cut-off frequency and distortion, when a negative feedback is applied with feedback ratio is 0.01. (5 m) [NOV/DEC 2012 - R8

Gn:

$A = \frac{V_o}{V_i} = 400$ $f_L = f_1 = 50Hz$ $f_H = f_2 = 200kHz$

$D = 10\%$ $\beta = 0.01$ - feedback ratio.

① $A_f = \frac{A}{1+A\beta}$

$= \frac{400}{1+400\times0.01}$

$A_f = 80$

Voltage gain with feedback (-ve)

② $f_{Lf} = \frac{f_L}{1+A\beta}$

$= \frac{50}{1+400\times0.01}$

$f_{Lf} = 10$

③ $f_{Hf} = (1+A\beta)\times f_H$

$= (1+400\times0.01)\times 200\times10^3$

$f_{Hf} = 1\ MHz$

④ Distortion $D_f = \frac{D}{1+A\beta}$

$= \frac{10}{1+400\times0.01} = 2\%$

$D_f = 2\%$ ⑥

(iii) An amplifier, with feedback has voltage gain of 100. When the gain without feedback changes by 20% and the gain with feedback should not vary more than 2%. If so, determine the values of open loop gain A and feedback ratio fi(β)(5M)

[Nov/Dec 2012-R8]

An: $A_f = 100$ $\frac{dA_f}{A_f} = 2\% = 0.02$ $\frac{dA}{A} = 20\% = 0.2$

W.k.t

$$\frac{dA_f}{dA_f} = \frac{dA}{A} \times \frac{1}{1+A\beta}$$

$$0.02 = 0.2 \times \frac{1}{1+A\beta}$$

$$1+A\beta = \frac{0.02}{0.2} = 10 \rightarrow \boxed{1+A\beta = 10}$$

Also w.k.t, Gain with feedback is $A_f = \frac{A}{1+A\beta}$

$$100 = \frac{A}{10}$$

Open loop gain or Gain without feedback $\boxed{A = 1000}$

Feedback ratio (β):

w.k.t $1+A\beta = 10$

$A\beta = 9$

$\beta = \frac{9}{A} = \frac{9}{1000}$

$\boxed{\beta = 0.009}$

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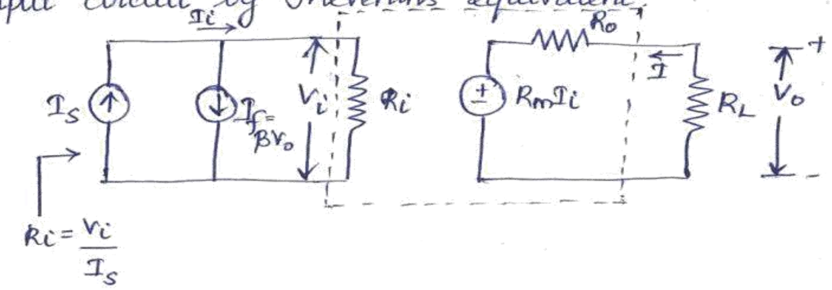
(2) a. Draw the circuit of voltage shunt & current series feedback amplifier and derive the expression for i/p & o/p resistance/impedance $R_{if/o}$. [APR/MAY 2011/R8, NOV/DEC 2012/R8] [10M] & also for voltage series, current shunt feedback Amplifier.

Voltage shunt feedback:

This topology is shown in the below fig. with the amplifier input circuit represented by Norton's model and the output circuit by Thevenin's equivalent.



Applying KCL to the input side, we get

$$I_s = I_i + I_f$$

$$\boxed{I_s = I_i + \beta V_o} \longleftarrow (1)$$

The output voltage is written as $V_o = \dfrac{R_m I_i R_o}{R_o + R_L}$

where $R_M = \dfrac{V_o}{I_i} = \dfrac{R_m R_o}{R_o + R_L}$ $\quad \boxed{V_o = R_M I_i} \longleftarrow (2)$

Sub. Eqn-(2) in (1).

$$I_s = I_i + \beta R_M I_i$$

$$I_s = I_i(1 + \beta R_M)$$

The input resistance with feedback is gn. as

$$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta R_M)} = \frac{R_i}{(1+\beta R_M)}$$

where R_m is the open circuit trans-resistance & R_M is the transresistance without feedback taking the load R_L into account.

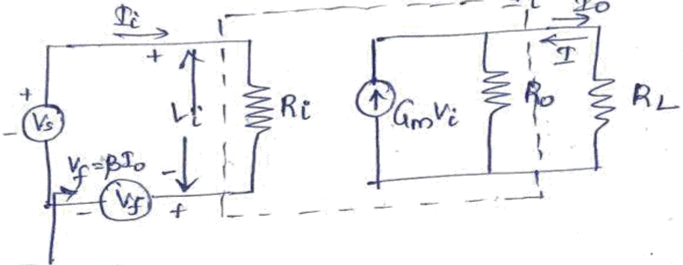
$$\boxed{\therefore R_m = \lim_{R_L \to \infty} R_M}$$

(8)

(6)

Current-series feedback: [6M – NOV/DEC 2010]-R8] (6)

This topology is shown in the below fig. with the amplifier input circuit represented by Thevenin's model & output circuit by Norton's equivalent circuit. Here the input impedance with feedback is given by $R_{if} = \frac{V_s}{I_i}$.



$$R_{if} = \frac{V_s}{I_i}$$

Applying KVL to the input side, we get

$$V_s = I_i R_i + V_f$$

$$\boxed{V_s = I_i R_i + \beta I_o} \quad \text{——} (1)$$

The output current is written as

$$I_o = \frac{G_m V_i R_o}{R_o + R_L} \qquad \text{where } G_M = \frac{G_m R_o}{R_o + R_L}.$$

$$\boxed{I_o = G_M V_i} \quad \text{——} (2)$$

sub. (2) in (1).

$$(1) \Rightarrow V_s = I_i R_i + G_M V_i \beta.$$

$$V_s = I_i R_i + G_M \beta . I_i R_i$$

$$R_{if} = \frac{V_s}{I_i} = R_i (1 + G_M \beta)$$

where G_m represents the short-circuit transconductance without feedback & G_M indicates the transconductance without feedback taking the load R_L into account.

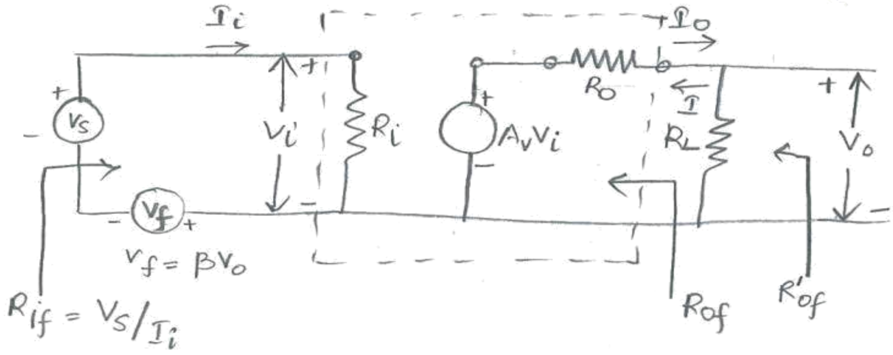
$$\boxed{G_m = \lim_{R_L \to 0} G_M}$$

(9)

Voltage series feedback:

The foll. fig. shows the voltage-series feedback topology with the amplifier i/p & o/p circuit replaced by its Thevenin's model.

Here A_v represents the open-circuit voltage gain taking R_s into account.



Here, the input impedance with feedback is gn. by $R_{if} = \frac{V_s}{I_i}$.

Applying KVL to the input side, we get

$$V_s = I_i R_i + V_f$$

$$V_s = I_i R_i + \beta V_o \quad \text{—(1)}$$

The o/p voltage is written as $V_o = \frac{A_v V_i R_L}{R_o + R_L} = A_v V_i$ —(2)

where $A_v = \frac{V_o}{V_i} = \frac{A_v R_L}{R_o + R_L}$

Sub. (2) in (1).

$$V_s = I_i R_i + \beta A_v V_i$$

$$V_s = I_i R_i (1 + \beta A_v)$$

$$R_{if} = \frac{V_s}{I_i} = R_i (1 + \beta A_v)$$

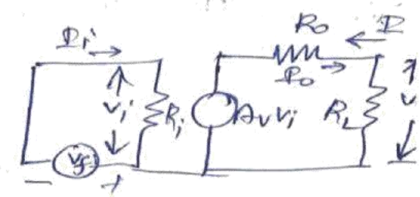
$$\boxed{R_{if} = R_i (1 + \beta A_v)}$$

$\therefore \quad A_v = \lim_{R \to \infty} A_v$

(10)

O/p Impedance: In the last fig V_o is replaced by V.
Applying KVL to the o/p side, we get

$$I = \frac{V - A_v V_i}{R_o} \quad \text{--- (3)}$$



The i/p voltage is written as

$$V_i = -V_f = -\beta V \quad \text{--- (4)} \quad (\text{with } V_s = 0)$$

Sub. (4) in (3)

$$I = \frac{V + A_v \beta V}{R_o} \Rightarrow \frac{V(1 + \beta A_v)}{R_o}$$

The o/p resistance with feed back is gn/- by.

$$\frac{1}{R_{of}} = \frac{I}{V} = \frac{(1+\beta A_v)}{R_o}$$

$$\boxed{R_{of} = \frac{R_o}{1 + \beta A_v}}$$

A_v - opencircuit voltage gain without taking R_L into account.
The output resistance with feedback R'_{of} including R_L as part of amplifier is gn. by $R'_{of} = R_{of} \,//\, R_L$

$$R'_{of} = \frac{R_{of} R_L}{R_{of} + R_L} = \frac{R_{of} R_L}{(1+\beta A_v)\left[\frac{R_o}{1+\beta A_v} + R_L\right]}$$

$$= \frac{R_o R_L}{R_o + R_L(1+\beta A_v)}$$

$$= \frac{R_o R_L}{R_o + R_L + R_L \beta A_v}$$

÷ Num & Deno. by $R_o + R_L$

$$R'_{of} = \frac{R_o R_L / (R_o + R_L)}{1 + \beta A_v R_L / (R_o + R_L)}$$

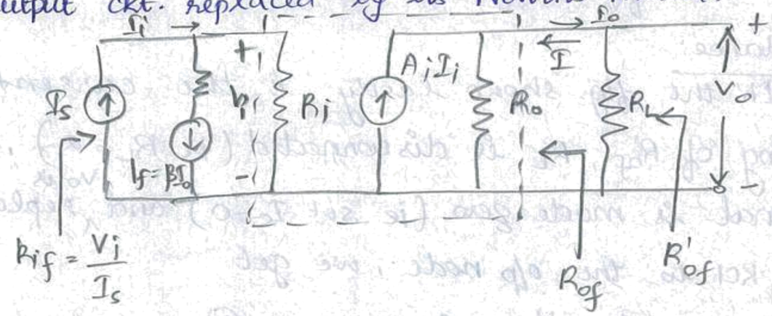
where $R_o' = \frac{R_o R_L}{R_o + R_L}$ & $A_V = \frac{A_v R_L}{R_o + R_L}$

$$R'_{of} = \frac{R_o'}{1 + \beta A_V}$$

where A_V indicates the opencircuit voltage gain taking the load R_L into account.

Current shunt feedback:

This topology is shown in the below fig with amplifier input & output ckt. replaced by its Norton's model.



Applying KCL to the i/p side, we get.

$$I_s = I_i + I_f$$

$$\boxed{I_s = I_i + \beta I_o} \quad \text{---} (5)$$

The o/p current is $I_o = \dfrac{A_i I_i R_o}{R_o + R_L}$

$$I_o = A_I I_i \quad \text{---} (6)$$

where $A_I = \dfrac{A_i R_o}{R_o + R_L} = \dfrac{I_o}{I_i}$.

Sub. (6) in (5).

$$I_s = I_i + \beta A_I I_i$$

$$= I_i (1 + \beta A_I)$$

The input resistance with feedback is given as

$$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta A_I)}$$

$$\boxed{R_{if} = \frac{R_i}{(1+\beta A_I)}}$$

where A_i represents the short circuit current gain without feedback and A_I is the current gain without feedback taking the load R_L into account.

$$\therefore A_i = \lim_{R_L \to 0} A_I$$

~~Effect of Negative feedback on i/p impedance (Pg. No: 5 & 6) & the o/p impedance is as follows.~~

O/p Impedance:

In the fig shown lastly is the current shunt feedback for finding of R_{of}, R_L is disconnected (ie. $R_L = \infty$), the external source signal is made zero (ie. set $I_s = 0$) and V_o is replaced with V.

Applying KCl to the o/p node, we get

$$I = \frac{V}{R_o} - A_i I_i \quad \text{--- (7)}$$

The input current is written as,

$$I_i = -I_f = -\beta I_o \Rightarrow +\beta I \quad \text{(8)} \quad (\text{with } I_s = 0 \text{ and } I = -I_o)$$

Sub. (8) in (7)

$$I = \frac{V}{R_o} - A_i \beta I$$

$$I(1 + A_i\beta) = \frac{V}{R_o}$$

The o/p resistance with feedback is $R_{of} = \frac{V}{I} = R_o(1 + A_i\beta)$

A_i is the short circuit current gain without taking the load R_L into account.

The o/p resistance with feedback R'_{of} including R_L is

$$R'_{of} = R_{of} \parallel R_L.$$

$$\text{where } R'_{of} = \frac{R_{of} R_L}{R_{of} + R_L} = \frac{R_o(1 + A_i\beta) R_L}{R_o(1 + A_i\beta) + R_L} \Rightarrow \frac{R_o R_L (1 + A_i\beta)}{R_o + R_L + R_o A_i \beta}$$

÷ Num & Den. by $(R_o + R_L)$

$$R'_{of} = \frac{\frac{R_o R_L}{(R_o + R_L)}(1 + A_i\beta)}{1 + \frac{R_o A_i \beta}{(R_o + R_L)}}$$

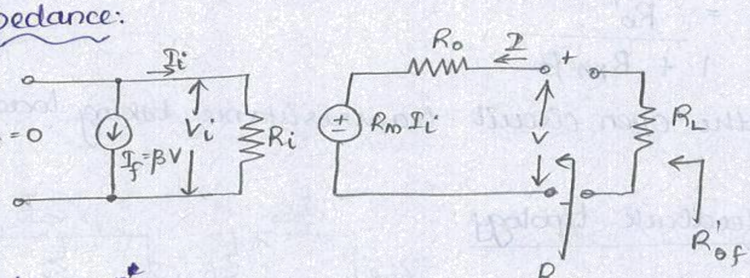
where $R'_o = \frac{R_o R_L}{R_o + R_L}$ & $A_I = \frac{A_i R_o}{R_o + R_L}$

$$\boxed{R'_{of} = \frac{R'_o(1 + A_i\beta)}{1 + A_I\beta}}$$

Effect of Negative feedback on i/p impedance (See Pg. No: 5 & 6) & the o/p impedance is as follows for current series & voltage shunt feedback amplifiers.

Voltage shunt feedback topology

O/p impedance:



For finding R_{of}, R_L is disconnected (ie $R_L = \infty$) & the external source signal is made zero (i.e. set $I_s = 0$) & V_o is replaced with V.

Applying KVL to the o/p side, we get

$$I = \frac{V - R_m I_i}{R_o} \quad \text{--- (9)}$$

The i/p current is written as

$$I_i = -I_f = -\beta V \ (\text{with } I_s = 0) \quad \text{--- (10)}$$

Substituting I_i in (9) eqn -

$$I = \frac{V - R_m(-\beta V)}{R_o}$$

$$I = \frac{V + \beta V R_m}{R_o} \Rightarrow \frac{V(1+\beta R_m)}{R_o}$$

$$\frac{V}{I} = \frac{R_o}{(1+\beta R_m)}$$

The o/p resistance with feedback is given as

$$R_{of} = \frac{V}{I} = \frac{R_o}{1+\beta R_m}$$

R_m - open circuit transresistance without taking the load R_L into account

O/p resistance with feedback R'_{of} including R_L & gn. by

$$R'_{of} = R_{of} \parallel R_L \Rightarrow \frac{R_{of} R_L}{R_{of} + R_L}$$

$$\therefore R'_{of} = \frac{R_o R_L}{1+\beta R_m} \bigg/ \frac{R_o}{(1+\beta R_m)} + R_L \Rightarrow \frac{R_o R_L}{R_o + R_L(1+\beta R_m)} \quad \text{(14)}$$

÷ Num & Den. by R_o+R_L

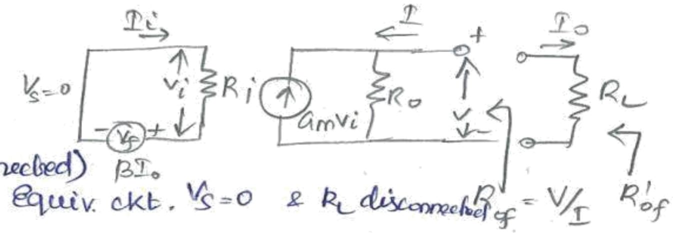
we get $R'_{of} = \dfrac{\dfrac{R_oR_L}{(R_o+R_L)}}{1+\dfrac{\beta R_m R_L}{(R_o+R_L)}}$ where $R_o' = \dfrac{R_oR_L}{(R_o+R_L)}$ & $R_M = \dfrac{R_mR_L}{R_o+R_L}$

$$R'_{of} = \frac{R_o'}{1+R_M\beta}$$

R_M indicates the open circuit transresistance taking load R_L into account.

Current series feedback topology:

O/p impedance:



To find R_{of} ($R_L \to \infty$ i.e. R_L disconnected) the external source signal is made zero. (ie. $V_s=0$) & V_o is replaced with V.

Equiv. ckt. $V_s=0$ & R_L disconnected $R_{of} = V/I$

Applying KCL to the o/p node, we get

$$I = \frac{V}{R_o} - G_m V_i \quad \text{——— (11)}$$

The i/p voltage is written as $V_i = V_f = -\beta I_o = \beta I$ (with $V_s=0$ & $I = -I_o$) — (12)

Sub. (12) in (11) we get $I = \dfrac{V}{R_o} - G_m\beta I$

$$\frac{V}{R_o} = I(1+G_m\beta) \Rightarrow \frac{V}{I} = R_o(1+\beta G_m)$$

The o/p resistance with feedback is gn. as $R_{of} = \dfrac{V}{I} = R_o(1+\beta G_m)$

where G_m represents short ckt. transconductance without R_L.

The o/p resistance with feedback R'_{of} including R_L is $R'_{of} = R_{of} \parallel R_L$

$$R'_{of} = \frac{R_{of}R_L}{R_{of}+R_L} = \frac{R_o(1+\beta G_m)R_L}{R_o(1+\beta G_m)+R_L} \Rightarrow \frac{R_oR_L(1+\beta G_m)}{R_o+R_L+R_o\beta G_m}$$

÷ Num & Den. by (R_o+R_L) $\Rightarrow \dfrac{\dfrac{R_oR_L}{(R_o+R_L)}\cdot(G_m\beta+1)}{1+\dfrac{R_o}{(R_o+R_L)}\beta G_m}$ where $\dfrac{R_oR_L}{R_o+R_L} = R_o'$ & $G_M = \dfrac{R_oG_m}{R_o+R_L}$

$$\therefore R'_{of} = \frac{R_o'(1+G_m\beta)}{1+G_M\beta}$$

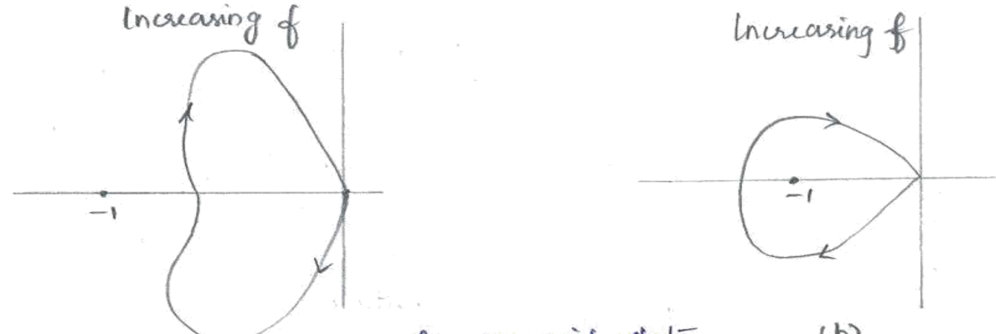
b. Discuss nyquist criterion for stability of feedback amplifiers with the help of nyquist plot & bode plots. [NOV/DEC 2012, R8 ; APR/MAY 11 R8] (13) [6M]

Nyquist criterion:

To check the stability of feedback amplifiers, a popular technique known as Nyquist method is used.

For a system to be stable, all poles of the transfer fn/- or the zeros of (1+Aβ) must lie in the left half of the complex frequency plane. Some compensation techniques are employed to prevent a feedback amplifier from becoming unstable i.e. oscillator.

Nyquist diagram is used to plot gain and phase shift as a function of frequency, on a complex plane. Since the pdt. Aβ is a complex number and function of frequency, points in the complex plane are obtained for the value of Aβ corresp. to values of f from −∞ to ∞. The locus of all these points forms a closed curve.



(a) Stable — fig: Nyquist plots Stability conditions — (b) Unstable.

Nyquist criterion for stability states that an amplifier is unstable if the nyquist curve encloses the −1+j0 point, and the amplifier is stable if the curve doesn't enclose this point which is shown in the above fig.

The criterion also represents in the complex plane positive & negative feedback. |1+Aβ| = 1 represents a circle with unit radius is shown in the foll. figure with the centre @ −1+j0. point.

For any freq, if Aβ extends outside this circle, the feedback is negative i.e. |1+Aβ| > 1, If Aβ lies inside the circle then |1+Aβ| < 1 & feedback is +ve.

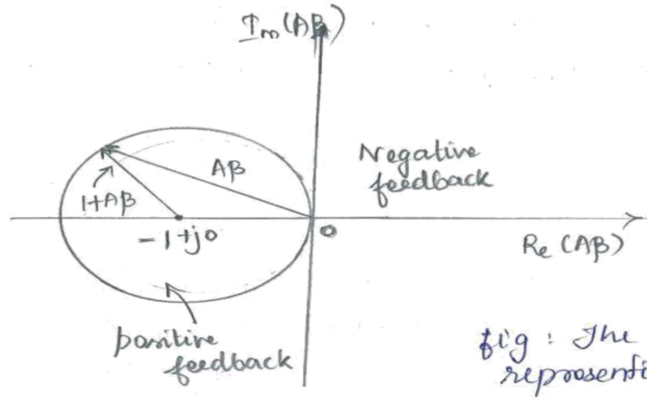


fig: The locus of |1 + βA| = 1 representing the type of feedback

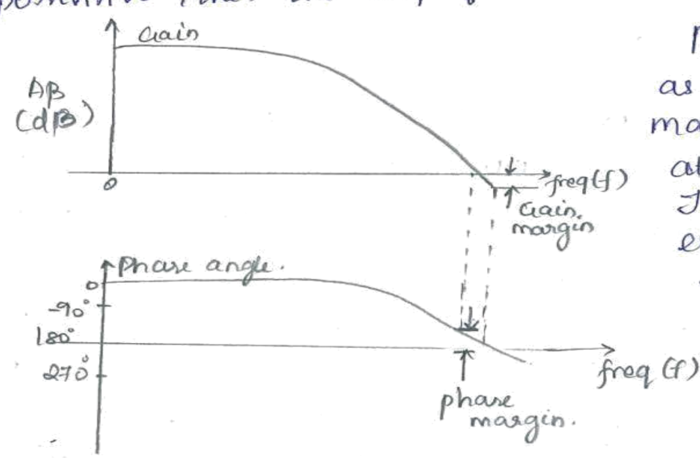
If the locus Aβ doesnot enclose the point −1+j0, i.e. |1 + Aβ| > 1, then the amplifier is stable & the feedback is negative for all frequencies.

Gain & phase margins:

– With the help of nyquist criterion, it is evident that a feedback amplifier is stable if the loop gain Aβ is less than unity (0dB) when its phase angle is 180°.

– Using Bode plots, we can determine some margins of stability to indicate how close to instability the system is.

Gain margin: (GM) is defined as the ~~angle of zero~~ value of |Aβ| in (dB) at the frequency at which the phase angle of Aβ is 180°; If the GM is negative, then the amplifier is stable. If GM is posivitive, then the amplifier is unstable.



Phase margin (PM) is defined as the angle of 180° minus the magnitude of the angle of Aβ at which |Aβ| is unity (0dB). The GM & PM are directly evaluated from the curves shown in current fig.

Bode plots showing Gain & Phase Margin. (17)

(15)

3. Explain the impact of negative feedback on Bandwidth, Harmonic distortion, Stability, input impedances, output impedances of an amplifier [16M] [APR/MAY '15 /R9] [MAY/JUNE '13 R8/10] [10M]

Bandwidth:

The gain with feedback for an amplifier is gn. by

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

Using the above eqn. we can write

$$A_{f\,mid} = \frac{A_{mid}}{1+\beta A_{mid}}$$

$$A_{f\,low} = \frac{A_{low}}{1+\beta A_{low}}$$

$$A_{f\,high} = \frac{A_{high}}{1+\beta A_{high}} \quad \&$$

The effect of negative feedback on lower cut off & upper cut off frequencies of the amplifier is analyzed here.

Lower cut-off frequency: The relation btw. gain at lower cut-off frequency and gain at mid freq. for an amplifier is given as

$$\frac{A_{low}}{A_{mid}} = \frac{1}{1-j\left(f_L/f\right)}$$

$$\therefore \quad A_{low} = \frac{A_{mid}}{1-j\left(f_L/f\right)}$$

Substituting A_{low} in $A_{f\,low}$,

$$A_{f\,low} = \frac{\dfrac{A_{mid}}{1-j\left(f_L/f\right)}}{1+\beta \dfrac{A_{mid}}{1-j\left(f_L/f\right)}}$$

$$A_{f\,low} = \frac{A_{mid}}{1-j\left(\frac{f_L}{f}\right) + A_{mid}\beta}$$

$$A_{f\,low} = \frac{A_{mid}}{(1 + A_{mid}\beta) - j\left(\frac{f_L}{f}\right)}$$

÷ Num & Den. by $(1+A_{mid}\beta)$, we have

$$A_{f\,low} = \frac{\frac{A_{mid}}{1+A_{mid}\beta}}{1 - j\left[\frac{f_L}{(1+A_{mid}\beta)f}\right]} \qquad \left[\because A_{fmid} = \frac{A_{mid}}{1+A_{mid}\beta}\right]$$

$$A_{f\,low} = \frac{A_{f\,mid}}{1-j\left[\frac{f_L}{(1+A_{mid}\beta)f}\right]}$$

$$\therefore \frac{A_{f\,low}}{A_{f\,mid}} = \frac{1}{1-j\left(\frac{f_{Lf}}{f}\right)}$$

where the lower cut-off frequency with feedback is given as

$$f_{Lf} = \frac{f_L}{(1+A_{mid}\beta)}$$

Thus from the above eqn- f_{Lf} we can say that lower-cut off frequency with feedback is less than the lower cut off frequency without feedback by factor $(1+A_{mid}\beta)$ ∴ By introducing negative feedback we can improve the low frequency response.

Upper-cut off frequency:

The relation btw. gain at upper cut off frequency & gain at mid frequency for an amplifier is given as

$$\frac{A_{high}}{A_{mid}} = \frac{1}{(1+j\,f/f_H)}$$

$$A_{high} = \frac{A_{mid}}{(1+j\,f/f_H)}$$

Sub. A_{high} in $A_{f\,high}$ eqn we have

$$A_{f\,high} = \frac{\frac{A_{mid}}{1+j(f/f_H)}}{1+\beta\left[\frac{A_{mid}}{1+j(f/f_H)}\right]} = \frac{A_{mid}}{1+j\left(f/f_H\right)+\beta A_{mid}}.$$

÷ Num & Den. by $(1 + A_{mid}\beta)$ (19)

$$A_{f\,high} = \frac{\dfrac{A_{mid}}{1 + A_{mid}\beta}}{1 + j\left[\dfrac{f}{(1 + A_{mid}\beta) f_H}\right]}$$

$$A_{f\,high} = \frac{A_{f\,mid}}{1 + j\left[\dfrac{f}{(1 + A_{mid}\beta) f_H}\right]}$$

Since $A_{f\,mid} = \dfrac{A_{mid}}{1 + A_{mid}\beta}$

$$\therefore \frac{A_{f\,high}}{A_{f\,mid}} = \frac{1}{1 + j\left(f/f_{Hf}\right)}$$

where the upper cut off frequency with feedback is gn. by

$$\boxed{f_{HL} = (1 + A_{mid}\beta) f_H}$$

upper cut off frequency with feedback is greater than upper cut off frequency without feedback by factor $(1 + A_{mid}\beta)$.
$\therefore$ By introducing negative feedback, high frequency response of an amplifier is improved.

Bandwidth without feedback is gn. by $BW = f_H - f_L$
Bandwidth with feedback is gn. by $BW_f = f_{Hf} - f_{Lf}$.

$$BW_f = (1 + A_{mid}\beta) f_H - \frac{f_L}{(1 + A_{mid}\beta)}$$

$$BW_f = BW(1 + A_{mid}\beta)$$

from the freq. response graph it. is clear that $(f_{Hf} - f_{Lf}) > (f_H - f_L)$ & hence $BW_f > BW$. As the A_{vf} reduces by the factor $(1 + A\beta)$, its Bandwidth increases by $(1 + A\beta)$. This shows that

$$\boxed{A_f \times BW_f = A \times BW.}$$

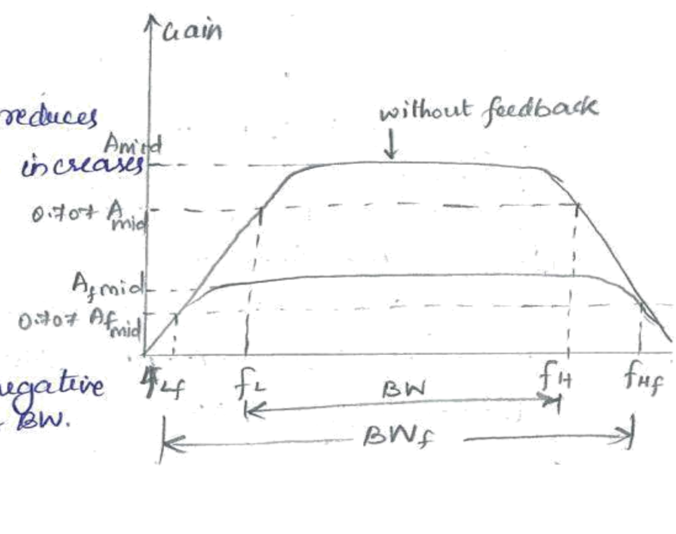
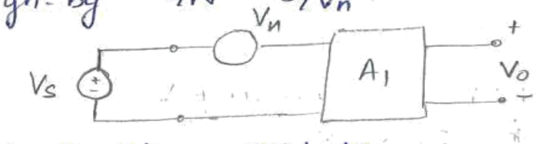


fig: Effect of negative feedback on Gain & BW.

Noise Reduction:

Negative feedback reduces the noise / interference in an amplifier, by increasing the ratio of signal to noise, which is possible only under certain conditions. The block diagram of an amplifier is shown below with input signal V_s, noise signal V_n and gain A_1. Assume that noise is introduced at the input of an amplifier & SNR of this amplifier is gn. by $S/N = V_s/V_n$.



Reduction in Nonlinear Distortion

The transfer characteristics of an amplifier is piecewise linear, with the voltage gain changing from 1000 to 100 and then to 0. This non-linear transfer characteristics of an amplifier generates a large amount of non-linear distortion at the o/p.

This can be linearized by applying negative feedback to the amplifier.

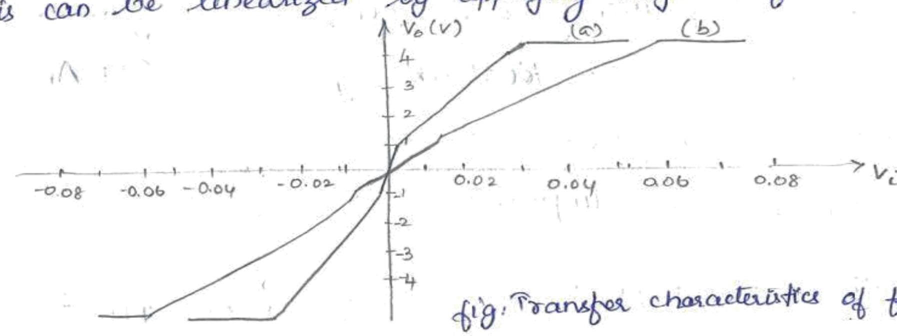


fig: Transfer characteristics of feedback amplifier.

Consider an amplifier with an open-loop voltage gain (A) & a total harmonic distortion without feedback (D). Then due to introduction of negative feedback, with the feedback ratio (B), the distortion D is reduced by a factor of $1+A\beta$ and the distortion with feedback (D_f) is gn. by

$$D_f = \frac{D}{1+A\beta}$$

High Input Impedance: The input impedance of the amplifier with feedback is given as $Z_{if} = Z_i(1+A\beta)$, when a factor is product to without feedback impedance, then the i/p impedance with fb. is increased by $(1+A\beta)$

Low Output impedance:

The output impedance of an amplifier with feedback is given as $Z_{of} = Z_o/(1+A\beta)$. Here the o/p impedance with fb. is decreased by $(1+A\beta)$.

Desenitization or Stabilization of Gain :

The closed-loop gain of the amplifier with negative feedback is given by

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

Diff. the above eqn- we get $\left|\frac{dA_f}{dA}\right| = \frac{(1+\beta A).1 - \beta A}{(1+\beta A)^2}$

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

Dividing both sides by A_f, we get

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

$$= \frac{dA}{(1+\beta A)^2} \times \frac{(1+\beta A)}{A} \qquad \left[\because A_f = \frac{A}{1+\beta A}\right]$$

$$\frac{dA_f}{A_f} = \frac{dA/A}{(1+\beta A)}$$

where $\frac{dA_f}{A_f}$ = Fractional change in amp., voltage gain with feedback

$\frac{dA}{A}$ = fractional change in voltage gain without feedback.

The term $\frac{1}{(1+A\beta)}$ is called Sensitivity

∴ Sensitivity is defined as ratio of percentage change in voltage gain with feedback to the percentage change in voltage gain without feedback.

$$\text{Sensitivity} = \frac{\left(\frac{dA_f}{A_f}\right)}{\left(\frac{dA}{A}\right)} = \frac{1}{1+A\beta}$$

Reciprocal of sensitivity is called desenstivity D or densensitivity $D = (1+\beta A)$. Hence the transfer gain divided by densensitivity is called the closed-loop gain & it can be written as $A_f = \frac{A}{1+\beta A}$

$$\boxed{A_f = \frac{A}{D}}$$

if $|A\beta| >> 1$

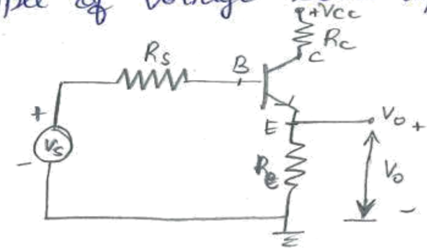
then $A_f = \frac{A}{1+\beta A} \approx \frac{A}{\beta A} = \frac{1}{\beta}$ ∴ Thus Gain entirely depends on fB n/w.

If fB n/w contains only stable passive elements, the improvement in stability is high. Increase in stability shows that the gain is made insensitive to changes in transistor parameters.

Ex. In voltage series fb, $A_{vf} = 1/\beta$ shows A_v is stabilized; In current series fb $G_{mf} = 1/\beta$ shows the transconductance gain is desensitized, In current shunt fb, A_I is densitized ($A_{If} = 1/\beta$) & in voltage shunt fB, the transresistance gain is stabilized/desensitized ($R_{mf} \approx 1/\beta$)

4(b) Draw the circuit of an Emitter follower. Identify the type of negative feedback. Cal. the gain i/p resistances with & without feedback [NOV/DEC '11, '14 R] [16M]

Example of voltage series topology, Emitter follower of BJT.



This fig. shows the BJT Emitter follower ckt. This feedback signal is the voltage V_f across R_E and the sampled signal is V_o across R_E.

This configuration conforms to voltage series feedback topology, as the sampled signal is taken directly from o/p node and the feedback signal is applied in series with the external excitation. Using the analysis steps, approximate expressions for voltage gain, input resistance & output resistance with feedback are obtained.

Now the basic amplifier without feedback is drawn.

To find the i/p ckt: Set $V_o = 0$, & hence V_s in series with R_s, appears between base B and emitter E.

To find the o/p ckt: Set $I_i = I_b = 0$ (i.e the input loop is opened), & hence R_E appears only in the o/p loop.

Following the above rules, the circuit shown below can be drawn

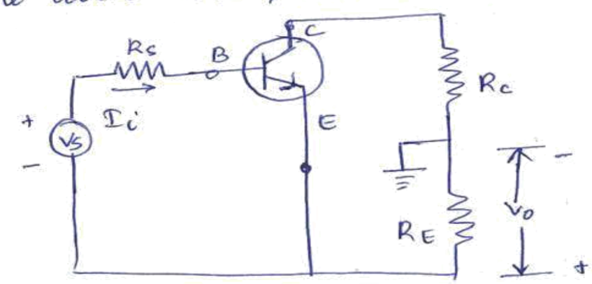


fig: Amplifier without feedback

The foll. fig. shows the equivalent ckt. after replacing the transistor by its low frequency approximate h-parameter model.

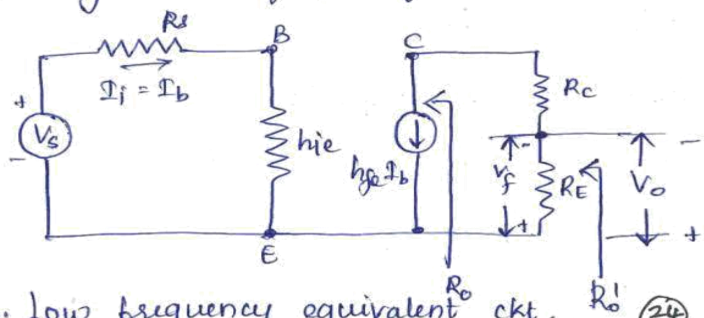


fig: Low frequency equivalent ckt.

In the above (last) fig. V_f & V_o are equal and hence $\beta = V_f / V_o = 1$
This topology stabilizes the voltage gain. Since R_s is considered as a part of the amplifier, then $V_i = V_s$, and the voltage gain without feedback is given by

$$A_V = \frac{V_o}{V_i} = \frac{h_{fe} I_b R_E}{V_s} = \frac{h_{fe} R_E}{R_s + h_{ie}} \quad \text{where} \quad V_s = I_b (R_s + h_{ie})$$

The desensitivity is given by

$$D = 1 + \beta A_V = 1 + \beta \frac{h_{fe} R_E}{R_s + h_{ie}} \qquad \because \beta = \frac{V_f}{V_o} = 1.$$

$$D = \frac{R_s + h_{ie} + h_{fe} R_E}{R_s + h_{ie}}$$

Then the voltage gain with feedback can be written as

$$A_{Vf} = \frac{A_V}{D} = \frac{h_{fe} R_E}{R_s + h_{ie} + h_{fe} R_E}$$

If $h_{fe} R_E >> R_s + h_{ie}$ then $A_{Vf} \propto 1$. This unity gain shows that it is an emitter follower ckt.

From the low freq. equiv. ckt., the input resistance without feedback is gn. by,

$$R_i = R_s + h_{ie}.$$

Hence, for voltage series feedback amplifier, the i/p resistance with feedback increases due to series mixing at the i/p & It is given by

$$R_{if} = R_i D = (R_s + h_{ie}) \times \frac{R_s + h_{ie} + h_{fe} R_E}{R_s + h_{ie}}$$

$$R_{if} = R_s + h_{ie} + h_{fe} R_E$$

The o/p resistance of the amplifier with feedback, without considering the external load resistance ($R_L = R$) is gn. by

$$R_{of} = \frac{R_o}{1 + \beta A_V} = \frac{\infty}{\infty} \quad \text{where } R_o = \infty \text{ and } A_V = \lim_{R_L \to \infty} A_V = \infty$$

This indeterminacy can be resolved by first evaluating R'_{of} an then apply the limit $R_L \to \infty$

The output resistance of the amplifier with feedback by considering the external load, can be written as

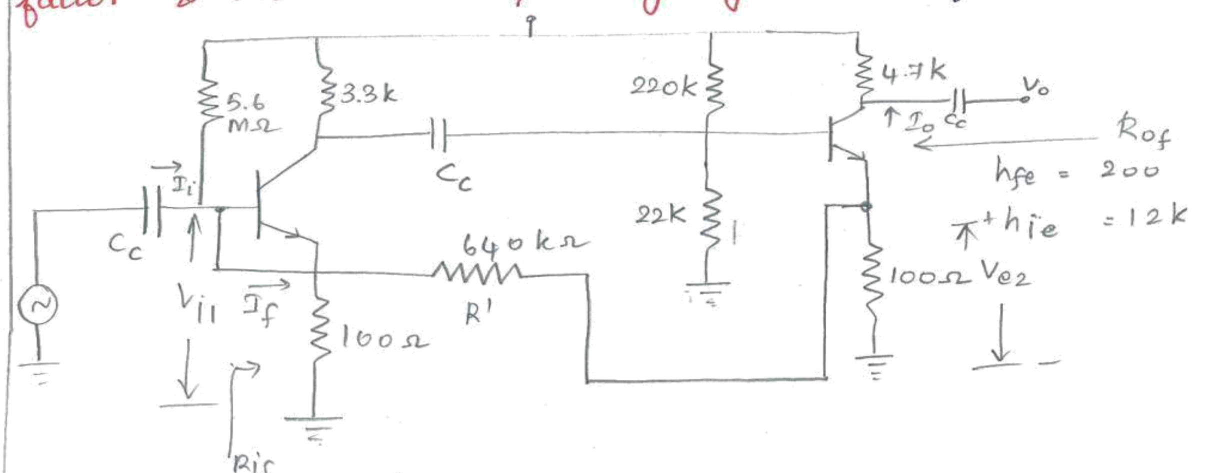
$$R'_{of} = \frac{R'_o}{D} = \frac{R_E\,(R_s + h_{ie})}{R_s + h_{ie} + h_{fe} R_E}$$

where $R'_o = R_L = R_E$

$$R_{of} = \lim_{R_L \to \infty} R'_{of} = \frac{R_s + h_{ie}}{h_{fe}}$$

Hence the feedback desensitizes voltage gain w.r.t. to changes in h_{fe} and it increases the input resistance & decreases the o/p resistance for voltage series feedback topology.

5. Find the type of amplifier shown in the diagram given below and draw the basic amplifier without feedback and find its gain with equiv. ckt. Also find feedback factor & its closed loop voltage gain. [16 M] [MAY/JUNE'14 R8/RC



Identifying the topology:

In this circuit, the sampling parameter is the current which is the emitter current of the second stage flowing through R_E and the mixing parameter is current which is the feedback current I_f flowing thro' the feedback resistance R'. Hence it is a case of current shunt feedback topology.

From the fig.

$$I_f = \frac{V_{i1} - V_{e2}}{R'} \quad \text{——— (1)}$$

Since $V_{e2} >> V_{i1}$ & neglecting the base current of Q_2 compared with the collector current.

$$I_f \approx \frac{-V_{e2}}{R'} \quad \text{——— (2)}$$

$$\therefore \quad I_f = -\frac{I_E R_E}{R'} \quad \text{——— (3)}$$

$$I_f = I_E + I_o$$

$$\therefore \quad I_E = I_f - I_o \quad \text{——— (4)}$$

Sub. (4) in (3)

$$I_f = \frac{-I_f R_E + I_o R_E}{R'} \quad \text{——— (5)}$$

$$I_f R' + I_f R_E = I_o R_E$$

$$I_f (R_E + R') = I_o R_E$$

$$I_f = \frac{I_o R_E}{R_E + R'}$$

$$I_f = \beta I_o \quad \text{——— (6)}$$

where $\beta = \frac{R_E}{R_E + R'}$ ——— (7)

$$A_{If} = \frac{I_o}{I_s} \simeq \frac{1}{\beta} \approx \frac{R' + R_E}{R_E} \quad \text{——— (8)}$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{I_o}{I_s} \cdot \frac{R_{C2}}{R_s} \quad \text{——— (9)}$$

$$A_{vf} = A_{If} \cdot \frac{R_{C2}}{R_s} \quad \text{——— (10)}$$

Sub. (8) in (10)

$$A_{vf} = \left(\frac{R' + R_E}{R_E}\right)\left(\frac{R_{C2}}{R_s}\right) \quad \text{——— (11)}$$

Finding the input circuit.

Since it is a current sampling set $I_o = 0$. This places R' in series with R_E between the base and the emitter of Q_1 as in foll. fig.

Finding the output circuit.

Since it is a shunt mixing set $V_i = 0$. This places R' in parallel with R_E as in the foll. fig.

∴ The resultant ckt. is shown below.

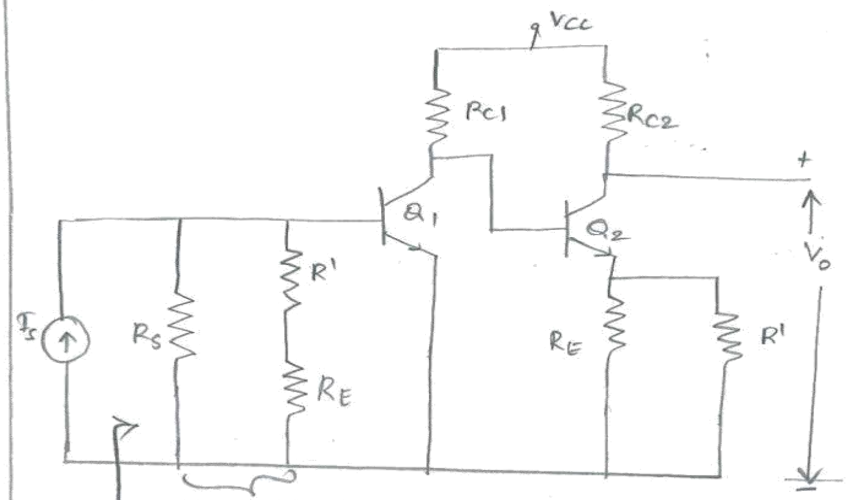


fig: Amplifier without feedback but including the loading of R'.

We know that,

$$R_{if} = \frac{R_i}{D} \quad \text{where} \quad R_i = R \parallel h_{ie}$$

$$R = R_s \parallel (R' + R_E)$$

Considering R_{C2} as external load, then R_o is the resistance seen looking into collector of Q_2. Since $h_{oe} = 0$ $R_o = \infty$.

W.K.T $R_{of} = R_o \cdot D$

$$\boxed{\therefore R_{of} = \infty}$$

$$R'_{of} = R_{of} \parallel R_L$$

$$= R_{of} \parallel R_{C2}$$

$$= \infty \parallel R_{C2}$$

$$\boxed{R'_{of} = R_{C2}}$$

(i) Feedback factor β

$$\beta = \frac{R_E}{R' + R_E}$$

an: $R_E = 100\Omega$

$R' = 640\Omega$

$$\therefore \beta = \frac{100}{640+100} = \frac{100}{740}$$

$$\boxed{\beta = 0.135}$$

(ii) Closed loop Voltage Gain (A_{vf})

$$A_{vf} = \left(\frac{R' + R_E}{R_E}\right)\left(\frac{R_{C2}}{R_S}\right)$$

$$= \left(\frac{640+100}{100}\right)\left(\frac{4.7k}{1k}\right)$$

$$= \left(\frac{740}{100}\right)\left(\frac{4.7\times10^3}{1\times10^3}\right) \quad \text{(Assume } R_s = 1k\text{)}$$

$$\boxed{A_{vf} = 34.78}$$

6. (i) Consider a three pole feedback amplifier with loop gain given by $T(f) = \frac{5\times10^5}{\left(1+j\frac{f}{10^6}\right)\left(1+j\frac{f}{10^7}\right)\left(1+j\frac{f}{10^8}\right)}$. Determine the frequency of dominant pole to stabilize the feedback s/m. Assume the phase margin is atleast 45° (6M) [APR/MAY 15 - R13]

(ii) A negative feedback amplifier has an open loop gain of 60,000 & closed loop gain of 300. If the open loop upper cut off frequency is 15 KHz, Estimate the closed loop upper cut off freq. Also calculate the total harmonic distortion with feedback if there is 10% harmonic without feedback (6M)

$$\boxed{f_{Hf} = f_H(1 + A_v\beta)}$$

an: $A_v = 60{,}000$; $A_{vf} = 300$

$f_H = 15\,kHz$ $\quad f_{Hf} = ?$ & $D_f = ?$ if $D = 10\%$.

Closed loop upper cut off freq. $A_{vf} = \frac{A_v}{1+\beta A_v} = \frac{60{,}000}{1+\beta(60{,}000)} = 300$ $\quad \boxed{\beta = 0.003316}$

$$D_f = \frac{D}{1+\beta A_v} = \frac{10/100}{1+(60{,}000\times0.003316)} \quad \boxed{D_f = 0.049\%}$$

$$f_{Hf} = f_H \times (1+A_v\beta) = 15\times10^3 \times 1+(60{,}000\times0.003316) \quad \boxed{f_{Hf} = 3\times10^6 Hz}$$

(i) $$T(f) = \frac{5 \times 10^5}{\left(1+j\frac{f}{10^6}\right)\left(1+j\frac{f}{10^7}\right)\left(1+j\frac{f}{10^8}\right)}$$

Solution:

$$\phi = 180 - \text{Phase Margin}$$

$$= 180 - 45$$

$$\phi = 135^\circ$$

$$\phi = \tan^{-1}\left(\frac{f}{10^6}\right) + \tan^{-1}\left(\frac{f}{10^7}\right) + \tan^{-1}\left(\frac{f}{10^8}\right)$$

$$135 = \tan^{-1}\left(\frac{f}{10^6}\right) + \tan^{-1}\left(\frac{f}{10^7}\right) + \tan^{-1}\left(\frac{f}{10^8}\right)$$

$$\boxed{f = 10\text{MHz}}$$

the equation for gain, input and output impedance ...

[MAY /JUNE 2013] - [16M]
R8/R10

Common Emitter (CE) amplified with unbypassed emitter resistance R_E is an example of current series feedback amplifier. is shown in foll. fig.

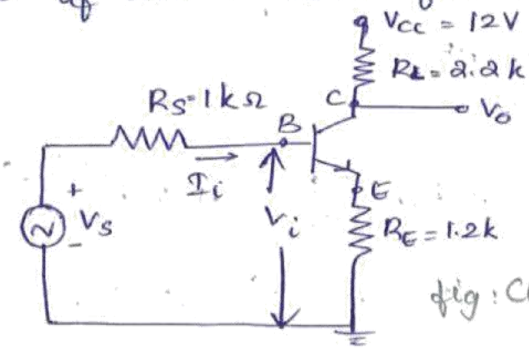


fig: CE amplifier with unbypassed emitter resistor R_E.

To draw the Basic amplifier without feedback as below, the i/p circuit of the amplifier is obtained by opening the output loop Here R_E appears in the input side. ||arly the o/p circuit is obtained by opening the input loop, and this place R_E again in the o/p side.

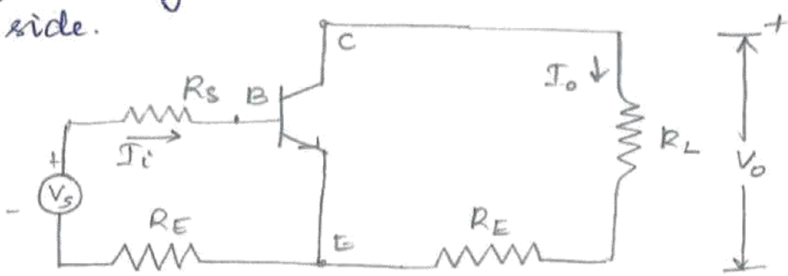


fig: Amplifier without feedback.

The resultant equivalent circuit is obtained by after replacing the transistor by its low frequency h-parameter model. No ground can be indicated in this ckt. because this couples the i/p to the o/p via R_E i.e. it would reintroduce FB, by taking the loading of β n/w into account. This topology stabilizes the transconductance gain G_M. Since the feedback voltage V_f appears across R_E in the o/p ckt.

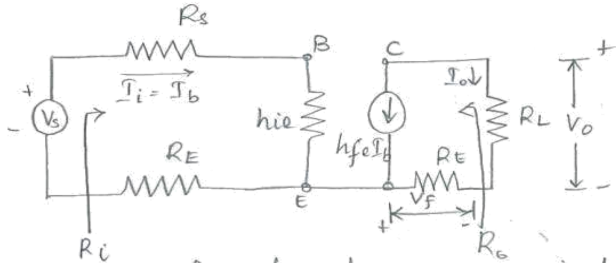


fig: low frequency equivalent ckt.

Then from fig above shown, $\beta = \frac{V_f}{V_o} = \frac{-I_o R_E}{I_o} = -R_E$

Since the i/p signal V_i without feedback is equal to V_s, then the transconductance without feedback is given by

$$G_M = \frac{I_o}{V_i} = \frac{-h_{fe} I_b}{V_s} = \frac{-h_{fe}}{R_s + h_{ie} + R_E} \quad \text{where } V_s = I_b(R_s + h_{ie} + R_E)$$

The desensitivity is given by

$$D = 1 + \beta G_M = 1 + \frac{R_E h_{fe}}{R_s + h_{ie} + R_E} \qquad [\because \beta = -R_E]$$

$$D = \frac{R_s + h_{ie} + R_E(1+h_{fe})}{R_s + h_{ie} + R_E}$$

$\therefore$ The transconductance with feedback can be written as

$$G_{Mf} = \frac{G_M}{D} = \frac{-h_{fe}}{R_s + h_{ie} + R_E(1+h_{fe})}$$

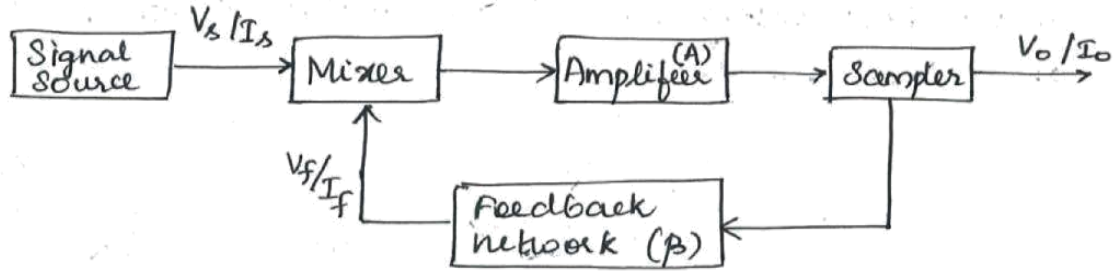
If $(1+h_{fe})R_E >> R_s + h_{ie}$ & since $h_{fe} >> 1$, then $G_{Mf} \simeq \frac{-1}{R_E} \simeq \frac{1}{\beta}$.

If R_E is a stable resistor, the transconductance gain with feedback is stabilized (desensitized). The load current is given by $I_o = G_{Mf} V_s = \frac{-h_{fe} V_s}{R_s + h_{ie} + (1+h_{fe}) R_E} \approx \frac{-V_s}{R_E}$.

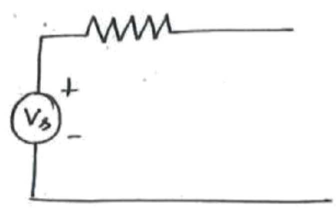
Under the conditions $(1+h_{fe}) R_E >> R_s + h_{ie}$ and $h_{fe} >> 1$, the load current is directly proportional to the input voltage, & this current depends only upon R_E & not upon any other ckt. / transistor parameter. The voltage gain is given by

$$A_{vf} = \frac{I_o R_L}{V_s} = G_{Mf} R_L = \frac{-h_{fe} R_L}{R_s + h_{ie} + (1+h_{fe}) R_E}$$

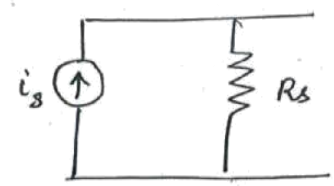
The basic functional blocks are a source, an amplifier with a transfer ratio (A), a mixer, a sampler and a feedback network with a feedback factor of β.



(i) Source: The source can either be a voltage source (or) current source. A voltage source is represented as thevenin's network and current source is represented as a Norton's network.



(a) Voltage source



(b) Current source

(ii) Amplifier: Basically classified into four types. Based on the input given & output obtained, they are

1. Voltage amplifier
2. Current Amplifier
3. Transconductance amplifier
4. Transresistance amplifier

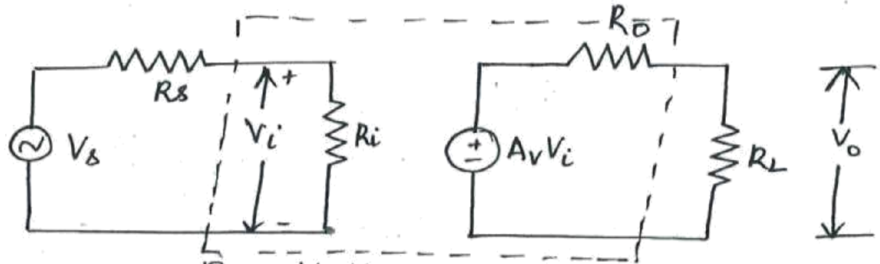
Voltage amplifier:

It is a two port network & it consists of thevenin's network both in input & output stage (Thevenin's voltage in series with thevenin's resistance).

The input impedance R_i of the amplifier should be large compared to the source Resistance R_s so that the drop

$$V_o = A_V \cdot V_i$$
$$= A_V V_s$$

Thus the amplifier provides an output voltage which is proportional to the input voltage. Such an amplifier is with high R_i (i.e. $R_i = \infty$) and Low R_o ($R_o = 0$).



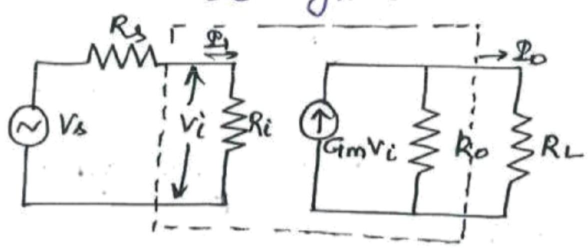
(i) $V_o \propto V_i$
(ii) $R_i > R_s$
(iii) $R_L > R_o$

Fig : Voltage amplifier

Transconductance Amplifier :

It is a two port network & consists of thevenin's network at the input stage & Norton's network in the output stage.

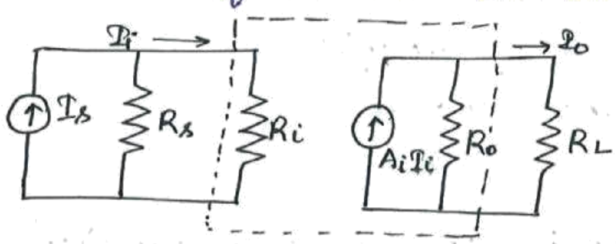
$$I_o = g_m V_i$$



(i) $I_o \propto V_i$
(ii) $R_i > R_s$
$R_o > R_L$
(iii) $R_i = \infty$, $R_o = \infty$

Current amplifier :

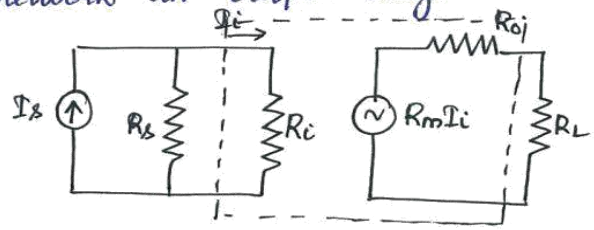
It consists of Norton's network both in the input & output stage.



(i) $I_o \propto I_i$
(ii) $R_i < R_s$ & $R_o < R_L$
(iii) $R_i = 0$, $R_o = \infty$

($\because I_s = I_i$) $\quad I_o = A_i I_i$

It consists of Norton's network at the input stage and thevenin's network in output stage.



(i) $V_o \propto I_i$

(ii) $R_i < R_s$

$R_L > R_o$

(iii) $R_i = R_o = \infty$

$V_o = R_m I_i$

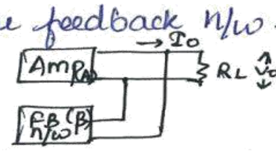
$R_m = \dfrac{V_o}{I_i}$

(iii) Sampling network:

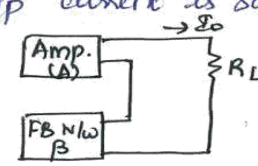
This functional block samples either the o/p current or o/p voltage and gives to feedback n/w.

1. Voltage sampler 2. Current sampler

* In case of voltage sampling (node sampling), the feedback n/w is connected across the output load R_L.



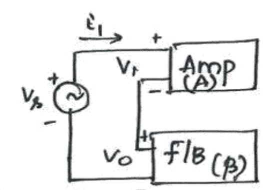
* In case of current sampler (loop sampling), the o/p current is sampled & so the feedback n/w is in series with the load.



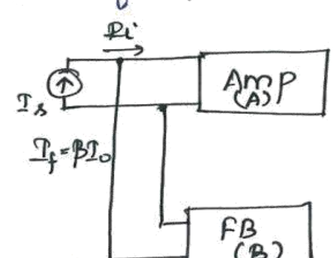
(iv) Mixer: It explains how the feedback signal mixes with the i/p signal.

1. Series Mixer 2. Shunt (parallel) mixer.

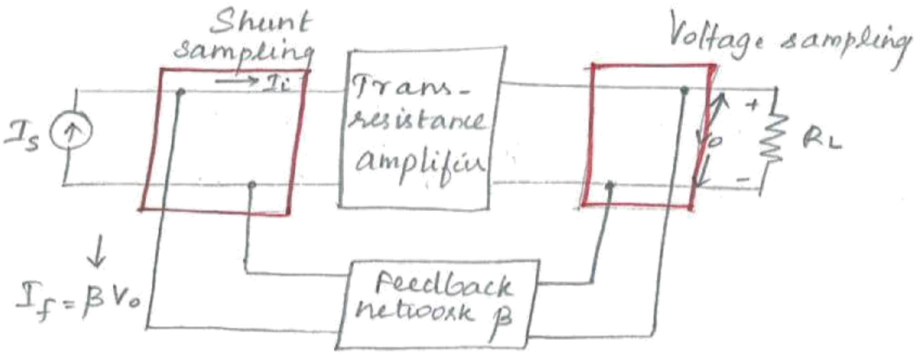
* Incase of series mixing, feedback signal is a voltage V_f & it is connected in series with the source voltage V_s. $R_i = \dfrac{V_s}{I_i} = \dfrac{V_i}{I_i}$; $R_{if} = \dfrac{V_s}{I_i} = \dfrac{V_i + V_f}{V_i}$ $R_{if} > R_i$



* In case of shunt mixing, the feedback current I_i is connected in shunt with the source current I_s.



$R_i = \dfrac{V_i}{I_s} = \dfrac{V_i}{I_i}$; $R_{if} = \dfrac{V_i}{I_s} = \dfrac{V_i}{I_i + I_f}$ $R_{if} < R_i$



(d) Transresistance amplifier with voltage shunt fB.

For voltage series feedback $A_{vf} = \frac{1}{\beta}$ (V_o/V_s), Voltage gain is stabilized

for current series feedback $G_{mf} = \frac{1}{\beta}$ (I_o/V_s), transconductance gain is stabilized.

For voltage shunt feedback $R_{mf} = \frac{1}{\beta}$ (V_o/I_s), transresistance gain is stabilized.

For current shunt feedback $A_{If} = \frac{1}{\beta}$ (I_o/I_s), Current gain is stabilized

Effect of Negative Feedback on Amplifier

Parameter	Voltage Series	Current series	Current shunt	Voltage shunt
Gain with feedback	$A_{vf} = \frac{A_v}{1+\beta A_v}$ decreases	$G_{mf} = \frac{G_m}{1+\beta G_m}$ decreases	$A_{if} = \frac{A_i}{1+\beta A_i}$ decreases	$R_{mf} = \frac{R_m}{1+\beta R_{mf}}$ decreases.
Stability & frequency response	Improves	Improves	Improves	Improves
Frequency distortion and Noise & Non linear distortion	Reduces	Reduces	Reduces	Reduces
I/p resistance	$R_{if} = R_i(1+\beta A_v)$ increases	$R_{if} = R_i(1+\beta G_m)$ increases	$R_{if} = \frac{R_i}{1+\beta A_I}$ decreases	$R_{if} = \frac{R_i}{1+\beta R_m}$ decreases
O/p resistances	$R_{of} = \frac{R_o}{1+\beta A_v}$ decreases	$R_{of} = R_o(1+\beta G_m)$ increases	$R_{of} = R_o(1+\beta A_i)$ increases	$R_{of} = \frac{R_o}{1+\beta R_m}$ decreases
	$R'_{of} = \frac{R_o'}{1+\beta A_v}$	$R'_{of} = \frac{R_o'(1+\beta G_m)}{(1+\beta G_M)}$	$R'_{of} = \frac{R_o'(1+\beta A_i)}{1+\beta A_I}$	$R'_{of} = \frac{R_o'}{1+\beta R_m}$

Unit - 1

PART-C

1. A feedback amplifier has open loop gain of 600, $\beta = 0.01$. Find the closed loop gain with negative feedback.

Soln.

$$A = 600$$

$$\beta = 0.01$$

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

$$= \frac{600}{1+(600 \times 0.01)}$$

$$\boxed{A_f = 85.7}$$

2. An amplifier has voltage gain of 1000 without feedback $f_c = 50\ Hz$, $f_H = 50\ kHz$ if 5% feedback is applied. Calculate the gain with feedback, lower & upper cut off frequency with feedback.

Soln.

$$A = 1000$$

$$\beta = \frac{5}{100} = 0.05.$$

$$F_L = 50\,Hz$$

$$F_H = 50\,kHz.$$

4. An amplifier has voltage gain of 1000. With negative feedback the voltage reduces to 100. Calculate β.

Soln.

$$A = 1000$$

$$A_f = 100$$

$$A_f = \frac{A}{1 + A\beta} \Rightarrow \beta = \frac{A/A_f - 1}{A}$$

$$\beta = \left(\frac{1}{A_f} - \frac{1}{A}\right) = \left(\frac{1}{100} - \frac{1}{1000}\right)$$

$$\boxed{\beta = 9 \times 10^{-3}}$$

5. If an amplifier has gain 400; $\beta = 0.1$. Find the gain with feedback.

Soln.

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

$$= \frac{400}{1 + (400 \times 0.1)}$$

$$\boxed{A_f = 9.756}$$

6. Voltage gain with feedback is 83.33 and $\beta = 0.01$. Calculate the gain of the amplifier without feedback.

Soln.

$$A_f = 83.33$$

$$\beta = 0.01$$

$A_f = \frac{A}{\ldots} \Rightarrow 1 + A\beta - \frac{A}{\ldots} = 0.$

$$A = \frac{-1}{\beta - 1/A_f}$$

$$A = \frac{-1}{0.01 - \frac{1}{83.33}}$$

$$= 499.8$$

$$\boxed{A \simeq 500}$$

7. An amplifier has mid band voltage gain of 1k with $F_L = 50 Hz$ and $F_H = 50 Hz$. If 0.05 feedback is supplied then calculate corresponding cut off frequencies.

Soln.

$F_L = F_H = 50 Hz$.

$A_{mid} = 1000$.

$\beta = 0.05$.

$$F_{LF} = \frac{F_L}{1 + A_{mid}\beta}$$

$$= \frac{50}{1 + (0.05)(1000)}$$

$$\boxed{f_{LF} = 0.98 Hz}$$

$$F_{HF} = F_H(1 + A_{mid}\beta)$$

$$= 50(1 + 50)$$

8. If an amplifier has BW of 300 k and voltage gain of 100. What will be the gain BW and gain of 10% -ve feedback is introduced? What is the gain BW product before and after feedback? What could be the amount of f/b if the BW is to be limited to 800k.

Soln.

$\beta = 0.1$

$A = 100$

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

$$= \frac{100}{1+(100)(0.1)}$$

$$\boxed{A_f = 9.09}$$

Bandwidth with f/b:

$$BW_f = BW(1+A\beta)$$

$$= 300k(1+[100\times0.1])$$

$$= 300\times10^3[1+(100\times0.1)]$$

$$\boxed{BW_f = 3300\ kHz}$$

Gain BW before feedback:

$$= A \times BW$$

$$= 100\times 300\times10^3$$

$= 30,000\ kHz$

$= 30\ MHz.$

Gain BW after feedback $= A_f \times BW_f$

$= 9.09 \times 3300 \times 10^3$

$= 29,997\ kHz.$

$\approx 30\ MHz.$

After feedback, $BW_F = 800\ kHz$; $A = 100$

$BW = 300kHz.$

$$BW_F = BW(1 + A\beta)$$

$$(800 \times 10^3) = (300 \times 10^3)[1 + (100)\beta]$$

$$1 + 100\beta = \frac{800}{300}$$

$$100\beta = \frac{8}{3} - 1$$

$$100\beta = \frac{5}{3}$$

$$\beta = \frac{5}{300}$$

$$\boxed{\beta = 0.0167}$$

9. Single stage RC coupled amplifier has midband gain 1k is made from negative f/b by feeding 10% of the ~~input~~ out voltage in series with input to find: (i) What is the ratio of half power frequency with f/b to without f/b? (ii) If $F_L = 20$ Hz & $F_H = 50$ kHz for amplifier without f/b, Find the corresponding values after f/b.

Soln.

$$A_{mid} = 1000$$

$$\beta = 0.1$$

$$F_L = 20\text{ Hz}$$

$$F_H = 50\text{ kHz}.$$

$$F_{LF} = \frac{F_L}{1 + A_{mid}\,\beta}$$

$$= \frac{20}{1 + (1000)(0.1)}$$

$$\boxed{F_{LF} = 0.198\text{ Hz}}$$

$$F_{HF} = F_H\,(1 + A_{mid}\,\beta)$$

$$= 50k\,[1 + (1000 \times 0.1)]$$

$$\boxed{F_{HF} = 5050\text{ kHz}}$$

Half power frequency / ratio = $\frac{F_{HF}}{F_H}$, $\frac{F_{LF}}{F_L}$

$$\frac{F_{HF}}{F_H} = \frac{5050 \times 10^3}{50 \times 10^3}$$

$$\boxed{\frac{F_{HF}}{F_H} = 101}$$

$$\frac{F_{LF}}{F_L} = \frac{0.198}{20}$$

$$\boxed{\frac{F_{LF}}{F_L} = 9.9 \times 10^{-3}}$$

10. When RC coupled amplifier has mid band gain 400, $F_L = 100$ Hz, $F_H = 15$ kHz & $\beta = 0.01$. Calculate gain with feedback & new BW.

Soln.

$F_L = 100$ Hz

$F_H = 15$ kHz

$\beta = 0.01$

$A_{mid} = 400.$

$$F_{LF} = \frac{F_L}{(1 + A_{mid}\beta)}$$

$$= \frac{100}{[1+(400\times 0.01)]}$$

$$\boxed{f_{LF} = 20\ Hz}$$

$$F_{HF} = F_H(1+A_{mid}\beta)$$

$$= (15\times 10^3)[1+(400\times 0.1)]$$

$$\boxed{F_{HF} = 75\ kHz}$$

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

$$= \frac{400}{1+(400\times 0.1)}$$

$$\boxed{A_f = 80}$$

New Bandwidth is:

$$BW = F_{HF} - F_{LF}$$

$$= (75\times 10^3) - 20$$

$$\boxed{BW = 74.98\ kHz}$$

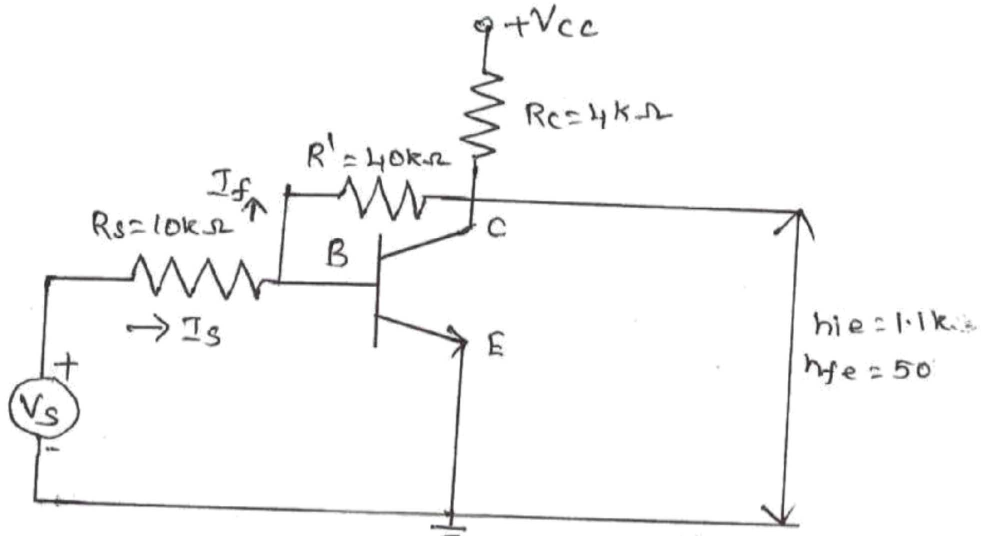


Fig shows the common emitter amplifier with a resistor R' connected from the output to the input.

Step 1: Identify topology:

The feedback current I_f is given as

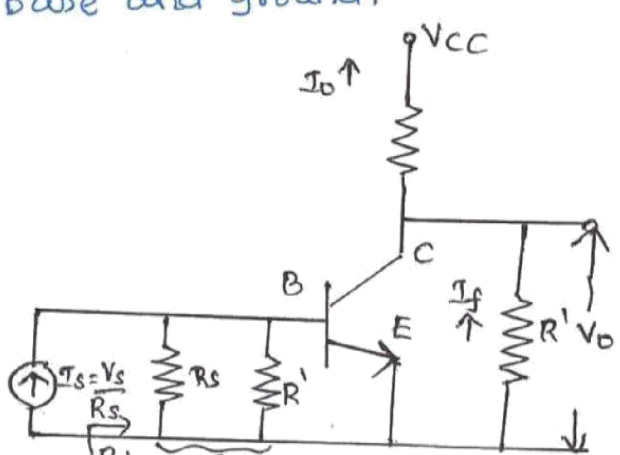
$$I_f = \frac{V_i - V_o}{R'} \quad \text{But } V_o > \beta V_i \quad \therefore \quad I_f = -\frac{V_o}{R'}$$

By shorting the output voltage ($V_o = 0$). feedback reduces to zero and hence it is a voltage sampling. As $I_i = I_s - I_f$ the mixing is shunt type and topology is "voltage shunt feedback amplifier".

Step: 2 and 3

Find input and output circuit.

→ To find the input circuit, set $V_o = 0$, this places R' between base and ground.



We have seen that

$$I_f = \frac{V_i - V_o}{R'} = -\frac{V_o}{R'}$$

$$\frac{I_f}{V_o} = \beta = \frac{-1}{R'}$$

$$R_{mf} = \frac{R_M}{1 + \beta R_M} = \frac{1}{\beta} = -R'$$

$$\overline{V_s} \quad \overline{I_s R_s} \quad \beta R_s \quad \overline{R_s}$$

Step 4: Find the open circuit transresistance

$$R_M = \frac{V_o}{I_s} = \frac{I_o R_c'}{I_s} = -\frac{I_c R_c'}{I_s}$$

where $R_c' = R_c \| R' = 4k \| 40k = 3.636\,k\Omega$ &

$$-\frac{I_c}{I_s} = -\frac{I_c}{I_b} \cdot \frac{I_b}{I_s}, \quad -\frac{I_c}{I_b} = A_i = -h_{fe} = -50 \;\&$$

$$\frac{I_b}{I_s} = \frac{R}{R+h_{ie}} = \frac{8k}{8k+1.1k} \Rightarrow 0.879$$

Now $R_M = -\frac{I_c R_c'}{I_s} = -\frac{I_c}{I_b} \cdot \frac{I_b}{I_s} \times R_c' = (-50) \times (0.879) \times 3.636k$

$$R_M = -159.8\,k\Omega$$

Step 5: calculate β: $\beta = -\frac{1}{R'} = -\frac{1}{40k} = -2.5 \times 10^{-5}$

Step 6: calculate D, R_{Mf}, A_{vf}, R_{if}, R_{of} and R_{of}'

$$D = 1 + \beta R_M = (1 + (-2.5 \times 10^{-5}) \times (-159.8 \times 10^{3})) = 4.995.$$

$$R_{Mf} = R_M / D = \frac{-159.8k}{4.995} \simeq -32k$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{V_o}{I_s R_s} = \frac{R_{Mf}}{R_s} = \frac{-32k}{10k} = -3.2$$

from fig(2)

$R_i = R \| h_{ie} = 0.967k$, $\quad R_{if} = R_i / D = \frac{0.967k}{4.995} = 193.59\,\Omega$

$$R_o = \infty \quad [h_{oe} = 0]$$

$$R_{of} = \infty / D = \infty$$

$$R_o' = R_o \| R_c' = \infty \| 3.636k = 3.636k$$

$$R_{of}' = \frac{R_o'}{D} = \frac{3.636k}{4.995} = 785\,\Omega$$ //.

EC 8452 - ELECTRONIC CIRCUIT II

2 MARKS WITH ANSWERS

UNIT-I

FEEDBACK AMPLIFIERS

1. Define stability factor? (Nov/dec-2009)

Stability factor is defined as the rate of change of collector current with respect to the rate of change reverse saturation current.

2. Justify that negative feedback amplifier increases bandwidth?(Nov/dec-2010)

The bandwidth of an amplifier is the difference between the upper cut-off frequency f2 and the lower cut-off frequency f1. The product of voltage gain and bandwidth of an amplifier without feedback and with feedback remains the same. i.e. Af*BWf=A*BW. Due to the negative feedback in the amplifier, the upper cut-off frequency is increased by the factor (1+Aβ) and the lower cut-off frequency is decreased by the same factor(1+Aβ).

3. State the Nyquist criterion to maintain the stability of negative feedback amplifier?(Nov/dec-2010)

Nyquist method is used to investigate the stability. Nyquist criterion for stability states that an amplifier is unstable if the Nyquist curve encloses the -1+0j point, and the amplifier is stable if the curve does not encloses this point.

4. What is the impact of negative feedback on noise in circuits?(Apr/may-2010)

With using the negative feedback with the feedback ratio, β, the noise N can be reduced by a factor of 1/(1+Aβ) in a similar manner to non-linear distortion. Thus the noise with feedback is given by,

$$N_f = \frac{N}{1+A\beta}$$

5. **Define sensitivity and de-sensitivity of gain in feedback amplifiers.(Apr/may-2010)** Sensitivity is defined as the ratio of percentage change in voltage gain with feedback to the percentage change in voltage gain without feedback.

Desensitivity is defined as the ratio of percentage change in voltage gain without feedback to the percentage change in voltage gain with feedback. The reciprocal of sensitivity.

6. What is return ratio of a feedback amplifier?(Nov/dec-2011)(Apr/may-2011) Feedback circuits consisting of bilateral feedback network (usually formed by passive elements) are analyzed using return ratio and loop gain. The closed-loop gain of the amplifier can be derived As

$$A=\frac{S0}{Si}=\frac{a}{1+a\,\beta}$$

where a is the forward gain, β is the feedback factor, and βa is the loop gain or the return ratio of the controlled source embedded in the forward amplifier.

7.Draw the block diagram of voltage shunt feedback amplifier and write the expressions for its input and output resistance.(Nov/dec-2011)

Input resistance with feedback $R_{if} = \frac{Ri}{1+A\beta}$

Output resistance with feedback $R_{of} = \frac{Ro}{1+A\beta}$

8. Define feedback factor of a feedback amplifier?(May/june-2012)

Feedback factor is given by

$$\beta=\frac{Vf}{Vo}$$

β- Feedback factor, Vf=feedback voltage, Vo= Output voltage

9.State the effect on output resistance and on input resistance of amplifier when current shunt feedback is employed.(May/june-2012)(Nov/dec-2012)(Apr/may-2011)

Characteristics	Current-shunt Feedback
Voltage gain	Decreases
Bandwidth	Increases
Input resistance	Decreases
Output resistance	Increases

10.Mention the three networks that are connected around the basic amplifier to Implement feedback concept.(Nov/dec-2012)

- Mixer
- Feedback network
- Sampler

11.List the characteristics of an amplifier which are modified by negative feedback.(Nov/dec-2013)

Characteristics	Type of Feedback			
	Current-shunt	Current - series	Voltage-series	Voltage- shunt
Voltage gain	Decreases	Decreases	Decreases	Decreases
Bandwidth	Increases	Increases	Increases	Increases
Input resistance	Decreases	Increases	Increases	Decreases
Output resistance	Increases	Increases	Decreases	Decreases

12.In a negative feedback amplifier. A=100,β=0.04 and Vs=50mv,find (a) gain with feedback (b) output voltage (c) feedback factor (d)feedback voltage.(Nov/dec-2013)

a). Af=20 b).Vo=1V c).Feedback factor β=0.04 d).Vf=0.04V

13. Negative feedback stabilises the gain-Justify the statement.(Apr/may-2014)

The negative feedback amplifier stabilises the gain. The gain of the amplifier with negative feedback is $A_f = \frac{A}{1+A\beta}$

Differentiating equation with respect to A, we get

$$\frac{dAf}{Af} = \frac{dA}{A}\frac{1}{1+A\beta}$$

15. Give an example for voltage-series feedback.

The Common collector or Emitter follower amplifier is an example for voltage series feedback.

16. Define negative feedback?

If the feedback signal is out of phase with the input signal then the input voltage applied to the basic amplifier is decreased and correspondingly the output is decreased. This type of feedback is known as negative or degenerative feedback.

17. What are the types of feedback?

i. Voltage-series feedback

ii. Voltage-shunt feedback

iii. Current-series feedback

iv. Current-shunt feedback

18. Give the properties of negative feedback.

i. Negative feedback reduces the gain

ii. Distortion is very much reduced

19. Define feedback?

A portion of the output signal is taken from the output of the amplifier and is combined with the normal input signal. This is known as feedback.

20. Write the expression for input and output resistance of voltage series feedback amplifier.

Input resistance with feedback, $R_{if} = (1+A\beta)R_i$

Output resistance with feedback, $R_{of} = \frac{R_o}{(1+A\beta)}$

16 Marks

1.(i) Explain how negative feedback acts on bandwidth, distortion, input impedance and output impedance of a circuit. (8) (April/May 2010)

(ii)An amplifier has a mid-frequency gain of 100 and a bandwidth of 200KHz.(1)What will be the new bandwidth and gain, if 5% negative feedback is introduced? (2) What should be the amount of feedback, if the bandwidth is to restricted to 1MHz? (8)

2. (i). Explain voltage series and voltage shunt feedback connections. (8) (April/May 2010) (ii). Explain Nyquist criterion to analyze the stability of feedback amplifiers.

3.(i). Sketch the block diagram of a feedback amplifier and derive the expressions for gain

(1) with positive feedback and (2) with negative feedback. State the advantages of negative feedback. (6) (April/May 2011)

(ii)An amplifier, without feedback, has a voltage gain of 400, lower cut off frequency f1=50Hz, upper cut off frequency f2=200 KHz and a distortion of 10%. Determine the amplifier voltage gain, lower cut off frequency and distortion, when a negative is applied with feedback ratio of 0.01. (5) (April/May 2011) (Nov/Dec 2012)

(iii)An amplifier, with feedback,has voltage gain of 100. When the gain without feedback changes by 20% and the gain with feedback should not vary more than 2%. If so, determine the values of open loop gain A and feedback ratio fi. (5) (April/May 2011) (Nov/Dec 2012)

4.(i). Draw the circuits of voltage shunt and current series feedback amplifiers and derive the expressions for input impedance Rif. (10) (April/May 2011) (Nov/Dec 2012)

(ii)Discuss Nyquist criterion for stability of feedback amplifier, with the help of Nquist plot and Bode plot. (6) (April/May 2011) (Nov/Dec 2012)

5.(i). Draw the block diagram of feedback amplifier and discuss the effect of negative feedback with respect to closed loop gain, bandwidth and distortion. (10) (Nov/Dec 2011)

(ii).An amplifier has a mid band gain of 125 and a bandwidth 250KHz. If 4% negative feedback is introduced, find the new bandwidth and gain. If the bandwidth is to be restricted to 1 MHz , find the feedback ratio. (6)

6.(i). With a neat circuit diagram, explain which type of feedback is employed in a BJT emitter follower and obtain the expressions for A_v,A_I,R_i &R_0. (8) (Nov/Dec 2011)

(ii). The voltage shunt feedback amplifier has the following values of circuit parameters. R_s=600Ω,h_{ie}=5KΩ,h_{fe}=80,R_L=2KΩ,R_B=40KΩ. Calculate A_v ,R_{if} ,A_{vf} ,R_{of} and R'_{of}. (8)

7.(i). For a feedback amplifier, derive the expressions for (1) the gain with feedback, (2)

Lower cut off frequency and (3) Upper cut off frequency. (8) (May/June 2012)

(ii). If an amplifier has a bandwidth of 300 KHz and a voltage gain of 100, what will be the new bandwidth and gain if 10% negative feedback is introduced? What will be the gain bandwidth product before and after feedback? What about be the amount of feedback if the bandwidth is to be limited to 800Khz? (8)

8.(i).Draw the block diagram of a voltage series feedback amplifier and derive the equation for input impedance and the voltage gain. (10) (Nov/Dec 2013)

(ii).Explain how a negative feedback in an amplifier helps in reduction of distortion and noise. (6)

9.(i). Draw the typical circuit for current series feedback confirmation and derive the expressions for voltage gain, current gain, input impedance and output impedance. (10) (Nov/Dec 2013) (Nov/Dec 2010)

(ii). Discuss the effect of negative feedback of stabilization of gain. (6)

(1) (i). Sketch the block diagram of a feedback amplifier and derive the expressions for gain With positive feedback and

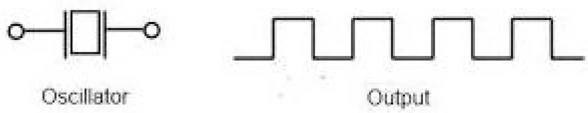
(2) With negative feedback. State the advantages of negative feedback. (Nov/Dec 2012)

(ii) Draw the block diagram of voltage series amplifier and derive for A_{vf2} ,R_{if} R_{of} . Draw a two stage amplifier with voltage series feedback. (10) (May/June 2014)

(iii) Derive for Bandwidth with feedback BW_f. (6) (May/June 2014)

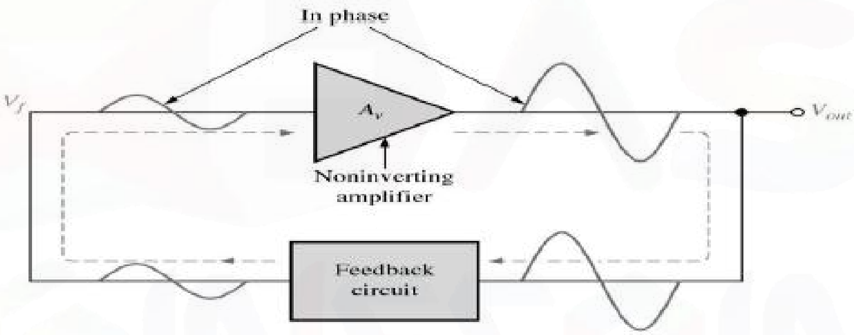
Unit II OSCILLATORS

An oscillator is a circuit that produces a repetitive signal from a dc voltage. The feedback type oscillator which rely on a positive feedback of the output to maintain the oscillations. The relaxation oscillator makes use of an RC timing circuit to generate a non-sinusoidal signal such as square wave.



The requirements for oscillation are described by the Baukhausen criterion:

- ✓ The magnitude of the loop gain Aβ must be 1
- ✓ The phase shift of the loop gain A β must be 0^o or 360^o or integer multiple of 2pi



Amplitude stabilization:

- ✓ In both the oscillators above, the loop gain is set by component values
- In practice the gain of the active components is very variuable
- If the gain of the circuit is too high it will saturate
- If the gain of the circuit is too low the oscillation will die

Real circuits need some means of stabilizing the magnitude of the oscillation to cope with variability in the gain of the circuit

Barkhausan criterion

The conditions for oscillator to produce oscillation are given by Barkhausan criterion. They are :

- The total phase shift produced by the circuit should be 360^o or 0^o
- ✓ The Magnitude of loop gain must be greater than or equal to 1 (ie)$|A\beta| \geq 1$

In practice loop gain is kept slightily greater than unity to ensure that oscillator work even if there is a slight change in the circuit parameters

2.2 Mechanism of start of oscillation

The starting voltage is provided by noise, which is produced due to random motion of electrons in resistors used in the circuit. The noise voltage contains almost all the sinusoidal frequencies. This low amplitude noise voltage gets amplified and appears at the output terminals. The amplified noise drives the feedback network which is the phase shift network. Because of this the feedback voltage is maximum at a particular frequency, which in turn represents the frequency of oscillation.

LC Oscillator:

Oscillators are used in many electronic circuits and systems providing the central “clock” signal that controls that controls the sequential operation of the entire system. Oscillators convert a DC input (the supply voltage) into an AC output (the waveform), which can have a wide range of different wave shapes and frequencies that can be either complicated in nature or simple sine waves depending upon the application.

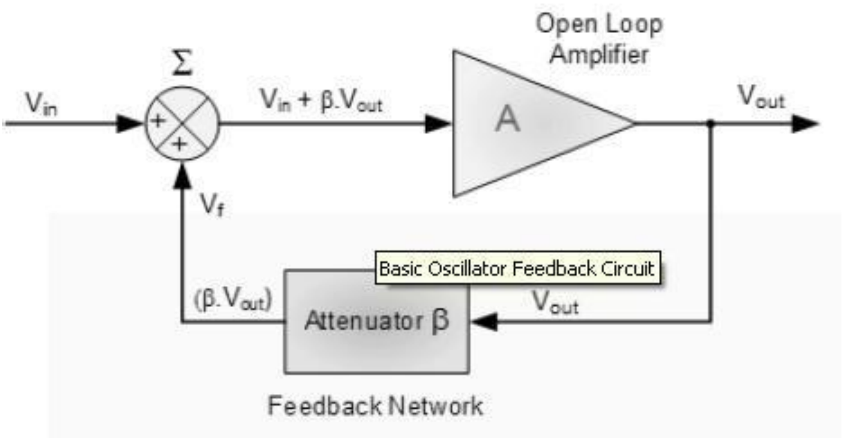
Oscillators are also used in many pieces of test equipment producing either sinusoidal sine wave, square, sawtooth or triangular shaped waveforms or just a train of pulse of a variable or constant width. LC Oscillators are commonly used in radio-frequency circuits because of their good phase noise characteristics and their ease of implementation.

An Oscillator is basically anAmplifier with “Positive Feedback”, or regenerative feedback (in-phase) and one of the many problems in electronic circuit design is stooping amplifiers from oscillating while trying to get oscillators to oscillate. Oscillators work because they overcome the losses of their feedback resonator circuit either in the form of a capacitor or both in the same circuit by applying DC energy at the required frequency into this resonator circuit.

In other words, an oscillator is a an amplifier which uses positive feedback that generates an output frequency without the use of an input signal.

It is self sustaining. Then an oscillator has a small signal feedback amplifier with an open-loop gain equal too or slightly greater than one for oscillations to start but to continue oscillations the average loop gain must return to unity. In addition to these reactive components, an amplifying device such as an Operational Amplifier or Bipolar Transistors required. Unlike an amplifier there is no external AC input required to cause the Oscillator to work as the DC supply energy is converted by the oscillator into AC energy at the required frequency.

2.3 Basic Oscillator Feedback Circuit



Where: β is a feedback fraction.

2.3.1 Without Feedback

$$\text{Gain, } Av = \frac{Vout}{Vin} \qquad A = \text{open loop voltage gain}$$

$$Av \times Vin = Vout$$

2.3.2 With Feedback

$$Av(Vin - \beta Vout) = Vout \qquad \beta \text{ is the feedback fraction}$$

$$Av.Vin - Av.\beta.Vout = Vout \qquad A\beta = \text{the loop gain}$$

$$Av.Vin = Vout(1 + A\beta) \qquad 1 + A\beta = \text{the feedback factor}$$

$$\therefore \frac{Vout}{Vin} = Gv = \frac{A}{1 + A\beta} \qquad Gv = \text{the closed loop gain}$$

Oscillators are circuits that generate a continuous voltage output waveform at a required frequency with the values of the inductors, capacitors or resistors forming a frequency selective LC resonant tank circuit and feedback network. This feedback network is an attenuation network which has a gain of less than one ($\beta<1$) and starts oscillations when A $\beta>1$ which returns to unity (A $\beta=1$) once oscillations commence. The LC oscillators frequency is controlled using a tuned or resonant inductive/capacitive (LC) circuit with the resulting output frequency being known as the Oscillation Frequency.

By making the oscillators feedback a reactive network the phase angle of the feedback will vary as a function of frequency and this is called Phase-shift.

There are basically types of Oscillators:\

1. Sinusoidal Oscillators - these are known as Harmonic Oscillators and are generally a :LC Tuned-feedback" or "RC tuned-feedback" type Oscillator that generates a purely sinusoidal waveform which is of constant amplitude and frequency.

2. Non-Sinusoidal Oscillators – these are known as Relaxation Oscillators and generate complex non-sinusoidal waveforms that changes very quickly from one condition of stability to another such as "Square-wave", "Triangular-wave" or "Sawtoothed-wave" type waveforms.

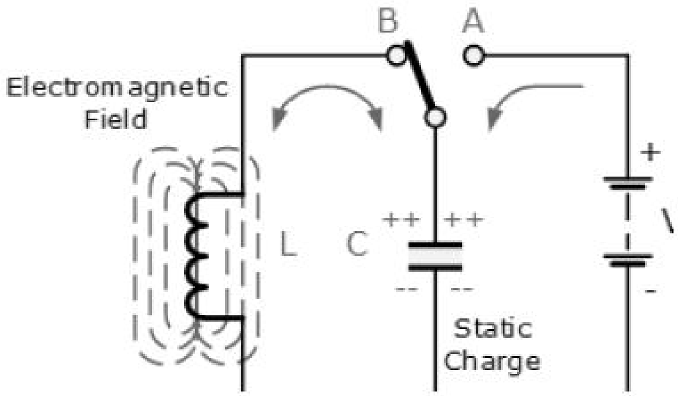
2.3.3 Resonance

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 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.

2.4 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 filed 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. This current now charges up the capacitor, c with the opposite polarity to its original charge.

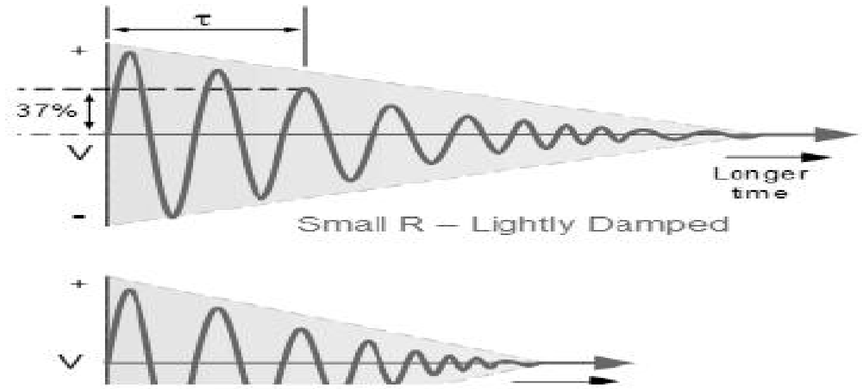
C continues to chare up until the current reduces to zero and the electromagnetic field of the coil has collapsed completely. The energy originally introduced into the circuit through the switch, has been returned to the capacitor which again has an electrostatic voltage potential across it, although it is now of the opposite polarity. The capacitor now starts to discharge again back through the coil and the whole process os 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 then forms the basis of an LC oscillators tank circuit and theoretically this cycling back and forth will continue indefinitely. However, every time energy is transferred from C to L or from L to C losses occur which decay the oscillations.

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.

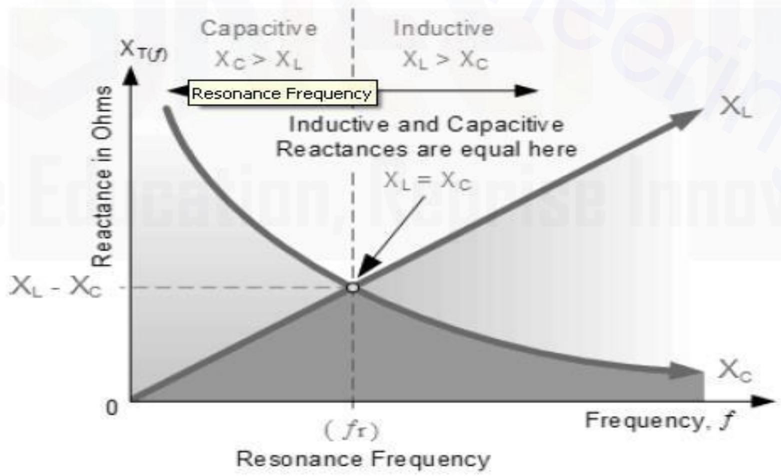
2.41 Damped Oscillations



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 on top of the curve for capacitive reactance so that both curves are on the same axes, the point of intersection will give us the resonance frequency point, (fr or ωr) as shown below.

2.4.2 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:

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

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

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

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

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

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

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

<u>2.4.3 Resonant Frequency of a LC Oscillator</u>

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

Where:

L is the Inductance in Henries

C is the Capacitance in Farads

ƒr 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 (ƒr) 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.

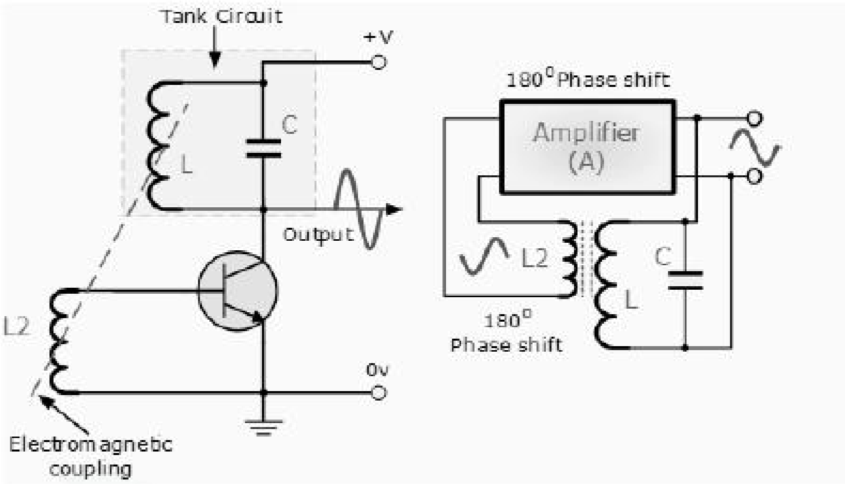
The amount of energy replaced must therefore be equal to the energy lost during each cycle. If the energy replaced is too large the amplitude would increase until clipping of the supply rails occurs. Alternatively, if the amount of energy replaced is too small the amplitude would eventually decrease 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.

2.5 Basic Transistor LC Oscillator Circuit



A Bipolar Transistor is used as the LC oscillators 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.

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 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 Oscillatorscircuit or by connecting a phase shift network between the output and input of the amplifier.

TheLC 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 OscillatorandClapp Oscillator to name a few.

2.6 The Hartley Oscillator

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.

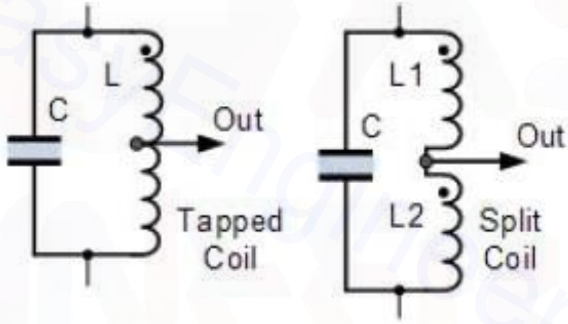
If the cumulative electromagnetic coupling between L1 and L2 is too small there would be insufficient feedback and the oscillations would eventually die away to zero Likewise if the feedback was too strong the oscillations would continue to increase in amplitude until they were limited by the circuit 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.



2.6.1 Hartley Oscillator Tuned Circuit

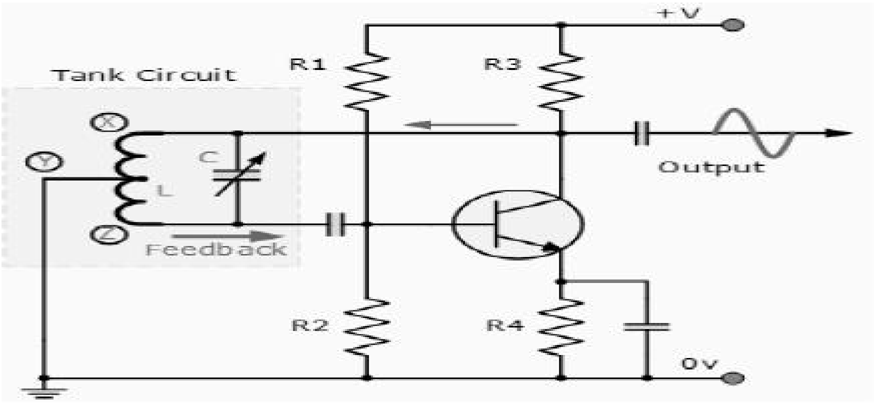
In the Hartley Oscillator the tuned LC circuit is connected between the collector and the base of the transistor amplifier. As far as the oscillatory voltage is concerned, the emitter is connected to a tapping point on the tuned circuit coil.

The feedback of the tuned 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.

2.6.2 <u>Basic Hartley Oscillator Circuit</u>



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 phase change in the voltage between the Base and Collector and this along with the original 180 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.

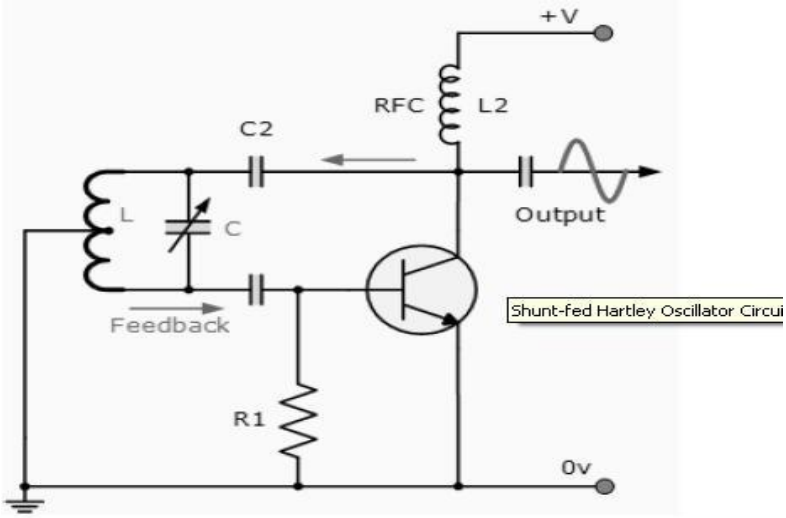
$$f = \frac{1}{2\pi\sqrt{L_T C}}$$

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

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.

2.6.3 **Shunt-fed Hartley Oscillator Cricuit**



In the Shunt-fed Hartley Oscillator 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.

2.7 Armstrong oscillator

The **Armstrong oscillato** (also known as **Meissneroscillator**) is named after the electrical engineer Edwin Armstrong, its inventor. It is sometimes called a tickler oscillator because the feedback needed to produce oscillations is provided using a tickler coil via magnetic coupling between coil L and coil T.

Assuming the coupling is weak, but sufficient to sustain oscillation, the frequency is determined primarily by the tank circuit (**L** and **C** in the illustration) and is approximately given by. In a practical circuit, the actual oscillation frequency will be slightly different from the value provided by this formula because of stray capacitance and inductance, internal losses (resistance), and the loading of the tank circuit by the tickler coil.

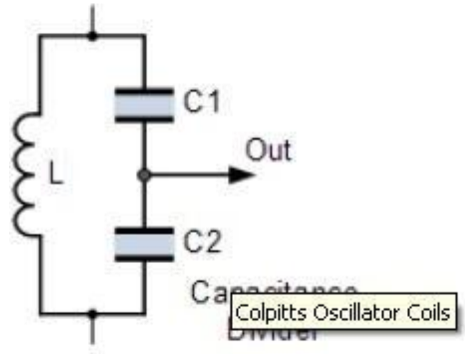
This circuit is the basis of the regenerative receiver for amplitude modulated radio signals. In that application, an antenna is attached to an additional tickler coil, and the feedback is reduced, for example, by slightly increasing the distance between coils **T** and **L**, so the circuit is just short of oscillation.

The result is a narrow-band radio-frequency filter and amplifier. The non- linear characteristic of the transistor or tube provides the demodulated audio signal.

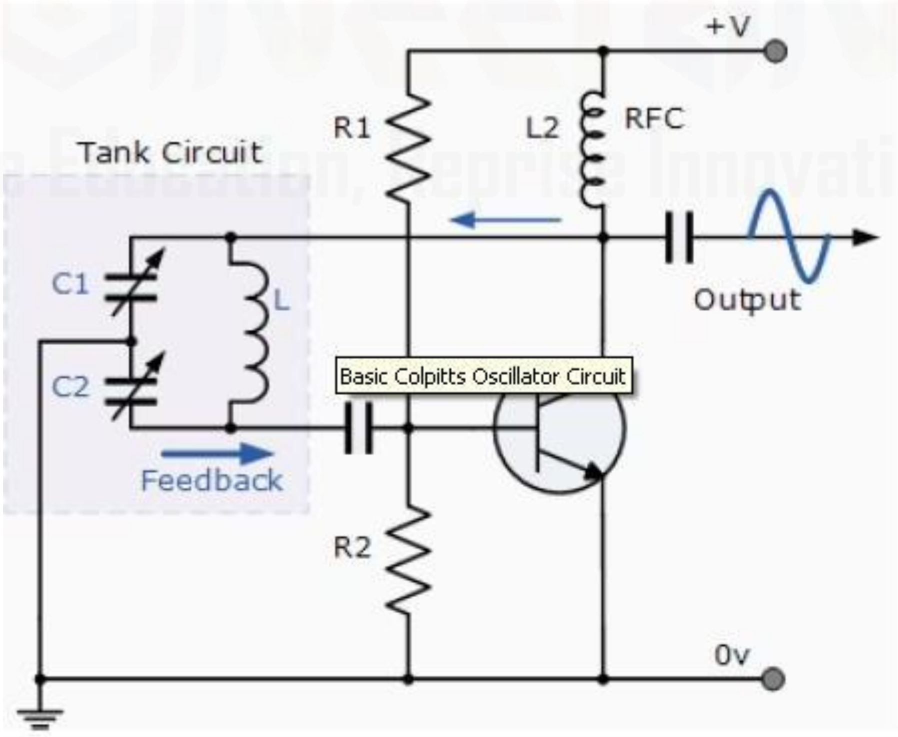
2.8 The Colpitts Oscillator

The Colpitts Oscillator, named after its inventor Edwin Colpitts is another type of LC oscillator design. 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.



2.9 Colpitts Oscillator Circuit



The transistor amplifiers emitter is connected to the junction of 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. The amount of feedback depends on the values of C1 and C2 with the smaller the values of C the greater will be the feedback.

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 un-damped oscillations. The amount of feedback is determined by the ratio of C1 and C2 which are generally "ganged" together to provide a constant amount of feedback so 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:.

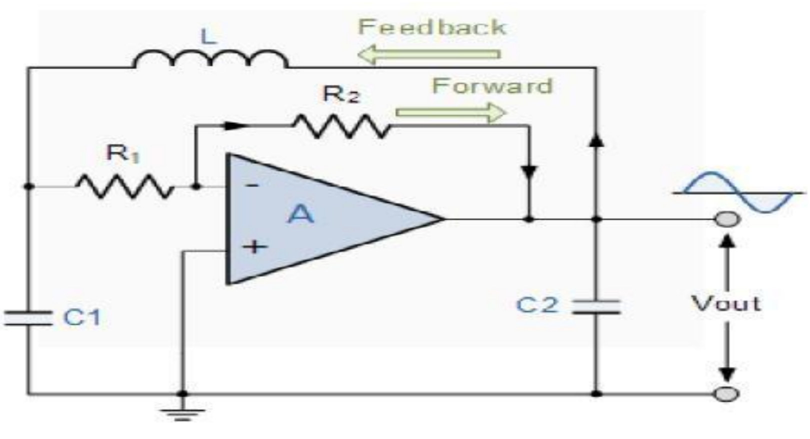
$$\frac{1}{C_T} = \frac{1}{C_1} + \frac{1}{C_2} \quad \text{or} \quad C_T = \frac{C_1 \times C_2}{C_1 + C_2}$$

The configuration of the transistor amplifier is of a Common Emitter Amplifier with the output signal 180^o out of phase with regards to the input signal. The additional 180^o 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^o. Resistors, R1 and R2 provide the usual stabilizing DC bias for the transistor in the normal manner while the capacitor acts as a DC-blocking capacitors. The radio-frequency choke (RFC) is used to provide a high reactance (ideally open circuit) at the frequency of oscillation, (fr) and a low resistance at DC.

2.9.2 Colpitts Oscillator using an Op-amp

As well as using a bipolar junction transistor (BJT) as the amplifiers active stage of the Colpitts oscillator, we can also use either a field effect transistor, (FET) or an operational amplifier, (op-amp). The operation of an **Op-amp Colpitts Oscillator** is exactly the same as for the transistorised version with the frequency of operation calculated in the same manner. Consider the circuit below.

2.9.3 Colpitts Oscillator Op-amp Circuit



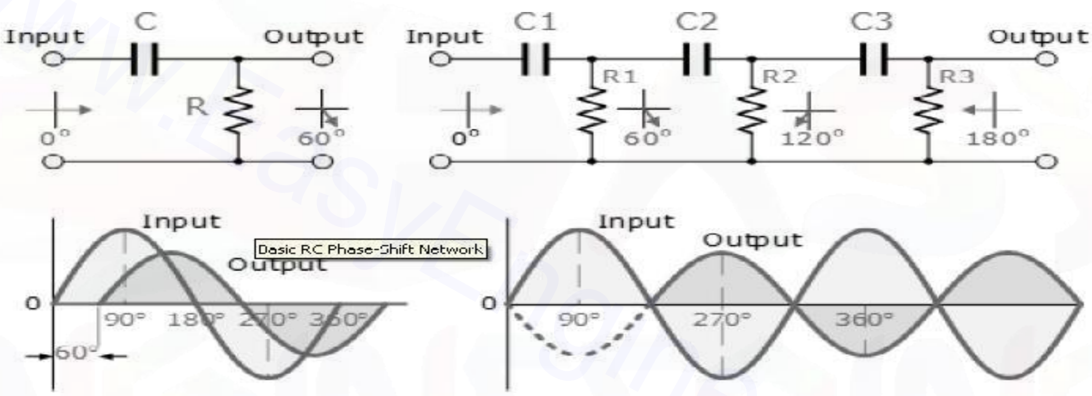
The advantages of the Colpitts Oscillatorover the Hartley oscillators are that the Colpitts oscillator produces a more purer sinusoidal waveform due to the low impedance paths of the capacitors at high frequencies. Also due to these capacitive reactance properties the Colpitts oscillator can operate at very high frequencies into the microwave region.

2.10 RC Phase-Shift Oscillator

In a RC Oscillator the input is shifted 180^o through the amplifier stage and180^o again through a second inverting stage giving us "$180^o + 180^o = 360^o$" of phase shift which is 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.

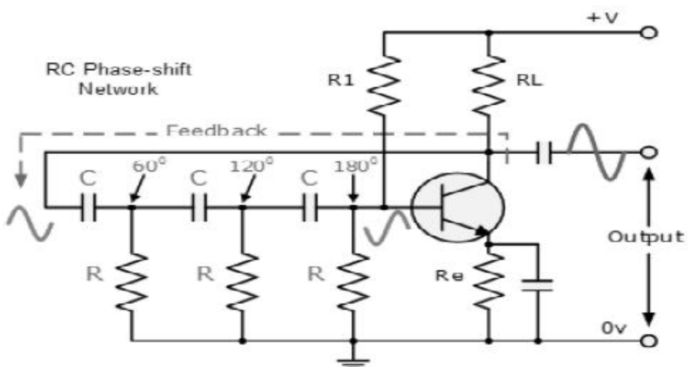
RC Phase-Shift Network



The circuit on the left shows a single resistor -capacitor network and whose output voltage "leads" the input voltage by some angle less than 90^o. An ideal RC circuit would produce a phase shift of exactly 90^o.The amount of actual phase shift in the circuit depends upon the values of the resistor and the capacitor, and the chosen frequency of oscillati ons with the phase angle (Φ) being given as:

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

RC Oscillator Circuit

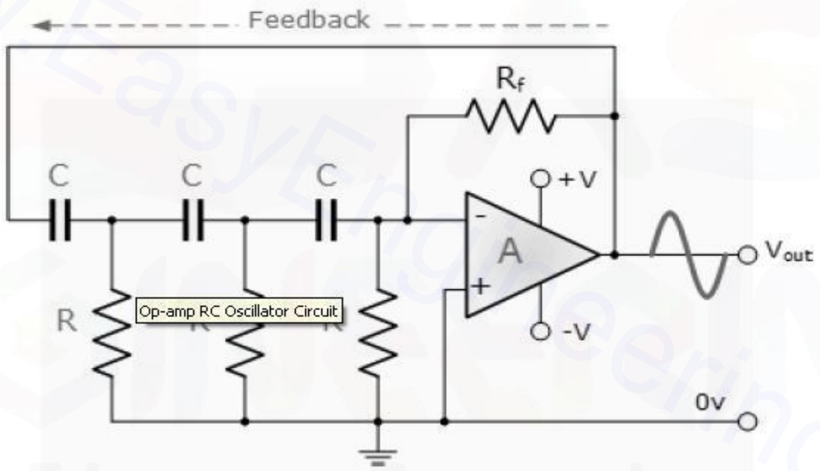


The RC Oscillator which is also called a Phase Shift Oscillator**,** produces a sine wave output signal using regenerative feedback from the resistor- capacitor combination. This regenerative feedback from the RC network is due to the ability of the capacitor to store an electric charge, (similar to the LC tank circuit).

This resistor-capacitor feedback network can be connected as shown above to produce a leading phase shift (phase advance network) or interchanged to produce a lagging phase shift (phase retard network) the outcome is still the same as the sine wave oscillations only occur at the frequency at which the overall phase-shift is 360^{o}. By varying one or more of the resistors or capacitors in the phase-shift network, the frequency can be varied and generally this is done using a 3-ganged variable capacitor

If all the resistors, R and the capacitors, C in the phase shift network are equal in value, then the frequency of oscillations produced by the RC oscillator is given as: $f = \frac{1}{2\pi CR\sqrt{6}}$

2.10.1 Op-amp RC Oscillator Circuit



As the feedback is connected to the non-inverting input, the operational amplifier is therefore connected in its "inverting amplifier" configuration which produces the required 180^{o} phase shift while the RC network produces the other 180^{o} phase shift at the required frequency ($180^{o} + 180^{o}$).

Although it is possible to cascade together only two RC stages to provide the required 180^{o} of phase shift ($90^{o} + 90^{o}$), the stability of the oscillator at low frequencies is poor.

One of the most important features of an RC Oscillator is its frequency stability which is its ability too provide a constant frequency output under varying load conditions. By cascading three or even four RC stages together (4×45^{o}), the stability of the oscillator can be greatly improved.

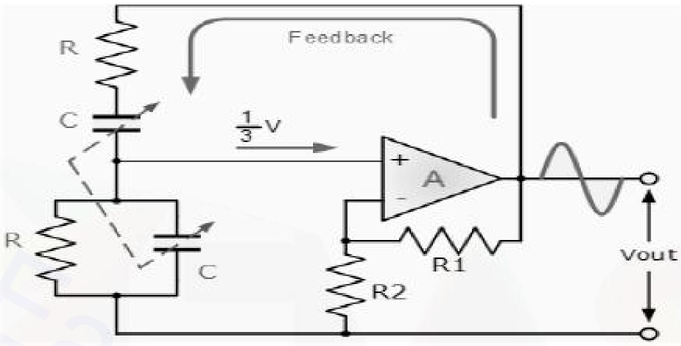
RC Oscillators with four stages are generally used because commonly available operational amplifiers come in quad IC packages so designing a 4- stage oscillator with 45^{o} of phase shift relative to each other is relatively easy.

2.11 WIEN BRIDGE OSCILLATOR

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 the Wien Bridge Oscillator.

The Wien Bridge Oscillator is so called because the circuit is based on a frequency-selective form of the Whetstone 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

Wien Bridge Oscillator



The output of the operational amplifier is fed back to both the inputs of the amplifier. One part of the feedback signal is connected to the inverting input terminal (negative feedback) via the resistor divider network of R1 and R2 which allows the amplifiers voltage gain to be adjusted within narrow limits.

The other part is fed back to the non-inverting input terminal (positive feedback) via the RC Wien Bridge network. The RC network is connected in the positive feedback path of the amplifier and has zero phase shift a just one frequency. Then at the selected resonant frequency, (fr) the voltages applied to the inverting and non-inverting inputs will be equal and "in-phase" so the positive feedback will cancel out the negative feedback signal causing the circuit to oscillate.

Also the voltage gain of the amplifier circuit MUST be equal to three "Gain =3" for oscillations to start. This value is set by the feedback resistor network, R1 and R2 for an inverting amplifier and is given as the ratio -R1/R2.

Also, due to the open-loop gain limitations of operational amplifiers, frequencies above 1MHz are unachievable without the use of special high frequency op-amps. Then for oscillations to occur in a Wien Bridge Oscillator circuit the following conditions must apply.

1. With no input signal the Wien Bridge Oscillator produces output oscillations.

2. The Wien Bridge Oscillator can produce a large range of frequencies.

3. The Voltage gain of the amplifier must be at least 3.

4. The network can be used with a Non-inverting amplifier.

5. The input resistance of the amplifier must be high compared to R so that the RC network is not overloaded and alter the required conditions.

6. The output resistance of the amplifier must be low so that the effect of external loading is minimised.

7. Some method of stabilizing the amplitude of the oscillations must be provided because if the voltage gain of the amplifier is too small the desired oscillation will decay and stop and if it is too large the output amplitude rises to the value of the supply rails, which saturates the op-amp and causes the output waveform to become distorted.

8. With amplitude stabilisation in the form of feedback diodes, oscillations from the oscillator can go on indefinitely.

2.12 Quartz Crystal Oscillators

One of the most important features of any oscillator is its frequency stability, or in other words its ability to provide a constant frequency output under varying load conditions. Some of the factors that affect the frequency stability of an oscillator include: temperature, variations in the load and changes in the DC power supply.

Frequency stability of the output signal can be improved by the proper selection of the components used for the resonant feedback circuit including the amplifier but there is a limit to the stability that can be obtained from normal LC and RC tank circuits.

To obtain a very high level of oscillator stability a Quartz Crystalis generally used as the frequency determining device to produce another types of oscillator circuit known generally as a Quartz Crystal Oscillator, (XO).

Crystal Oscillator

When a voltage source is applied to a small thin piece of quartz crystal, it begins to change shape producing a characteristic known as the Piezo-electric effect.

This piezo-electric effect is the property of a crystal by which an electrical charge produces a mechanical force by changing the shape of the crystal and vice versa, a mechanical force applied to the crystal produces an electrical charge.

Then, piezo-electric devices can be classed as Transducersas they convert energy of one kind into energy of another (electrical to mechanical or mechanical to electrical).

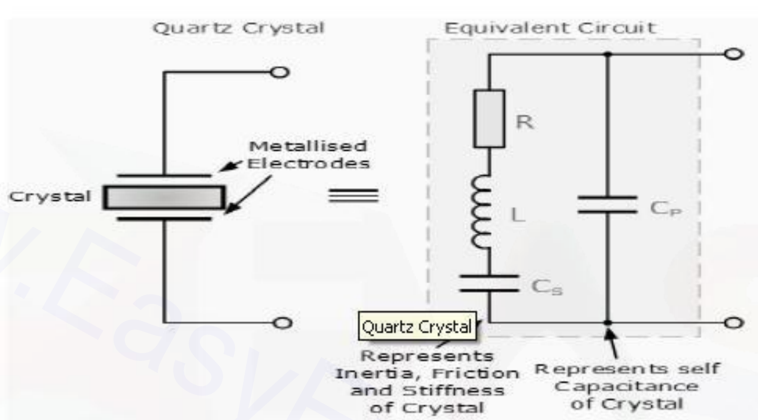
This piezo-electric effect produces mechanical vibrations or oscillations which are used to replace the LC tank circuit in the previous oscillators.

There are many different types of crystal substances which can be used as oscillators with the most important of these for electronic circuits being the quartz minerals because of their greater mechanical strength.

The quartz crystal used in a Quartz Crystal Oscillator is a very small, thin piece or wafer of cut quartz with the two parallel surfaces metallised to make the required electrical connections. The physical size and thickness of a piece of quartz crystal is tightly controlled since it affects the final frequency of oscillations and is called the crystals "characteristic frequency". Then once cut and shaped, the crystal can not be used at any other frequency. In other words, its size and shape determines its frequency.

The crystals characteristic or resonant frequency is inversely proportional to its physical thickness between the two metallised surfaces. A mechanically vibrating crystal can be represented by an equivalent electrical circuit consisting of low resistance, large inductance and small capacitance as shown below.

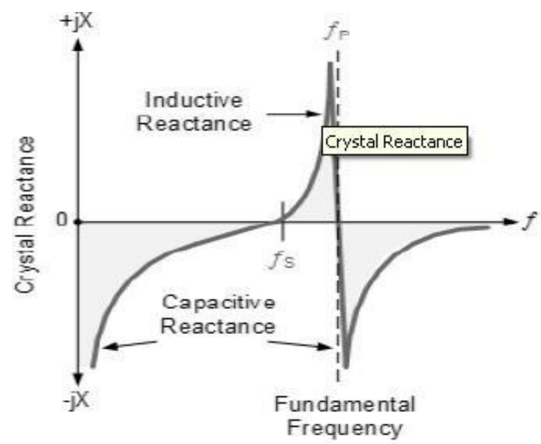
Quartz Crystal



The equivalent circuit for the quartz crystal shows an RLC series circuit, which represents the mechanical vibrations of the crystal, in parallel with a capacitance, Cp which represents the electrical connections to the crystal. Quartz crystal oscillators operate at "parallel resonance", and the equivalent impedance of the crystal has a series resonance where Cs resonates with inductance, L and a parallel resonance where L resonates with the series combination of Cs and Cp as shown.

Crystal Reactance

The slope of the reactance against frequency above, shows that the series reactance at frequency fs is inversely proportional to Cs because below fs and above fp the crystal appears capacitive, i.e. dX/df, where X is the reactance.



he slope of the reactance against frequency above, shows that the series reactance at frequency fs is inversely proportional to Cs because below fs and above fp the crystal appears capacitive, i.e. dX/d f, where X is the reactance. Between frequencies fs and fp, the crystal appears inductive as the two parallel capacitances cancel out. The point where the reactance values of the capacitances and inductance cancel each other out Xc = XL is the fundamental frequency of the crystal.

A quartz crystal has a resonant frequency similar to that of a electrically tuned tank circuit butwith a much higher Q factor due to its low resistance, with typical frequencies ranging from 4kHz to 10MHz. The cut of the crystal also determines how it will behave as some crystals will vibrate at more than one frequency. Also, if the crystal is not of a parallel or uniform thickness it has two or more resonant frequencies having both a fundamental frequency and harmonics such as second or third harmonics. However, usually the fundamental frequency is more stronger or pronounced than the others and this is the one used. The equivalent circuit above has three reactive components and there are two resonant frequencies, the lowest is a series type frequency and the highest a parallel type resonant frequency.

We have seen in the previous tutorials, that an amplifier circuit will oscillate if it has a loop gain greater or equal to one and the feedback is positive. In a Quartz Crystal Oscillator circuit the oscillator will oscillate at the crystals fundamental parallel resonant frequency as the crystal always wants to oscillate when a voltage source is applied to it.

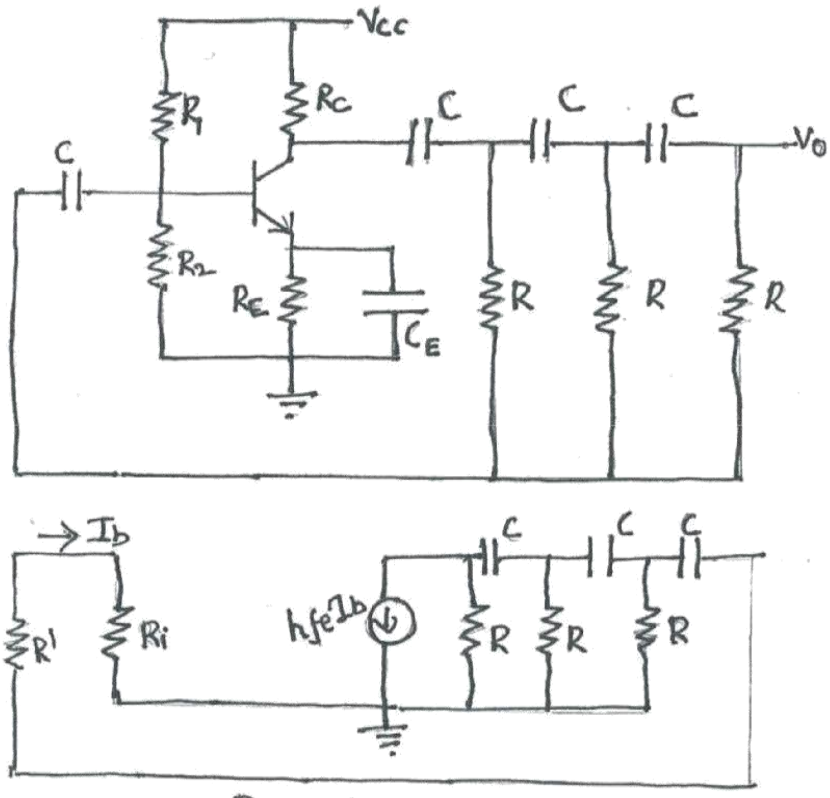
However, it is also possible to "tune" a crystal oscillator to any even harmonic of the fundamental frequency, (2nd, 4th, 8th etc.) and these are known generally as Harmonic Oscillators while Overtone Oscillators vibrate at odd multiples of the

16 Marks Questions

1) Explain the operation of Rc phase shift Oscillator and Wein Bridge Oscillator. Derive its frequency of Oscillation.

Rc Oscillators :-

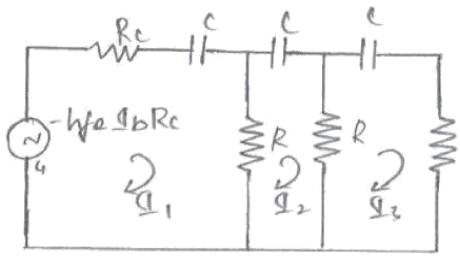
i) Rc phase shift Oscillator :-



$R_i = R_1 || R_2 || h_{fe}$

→ It is a audio frequency or low frequency Oscillator. It uses a Common emitter amplifier Whose Output is given to three Rc networks. The phase shift Produced by the Common emitter

emitter amplifier is 180°. Since an oscillator requir a phase shift of 360° and the additional 180° phase shift is obtained using three RC network with an individual phase shift of 60° each.



At loop B

$$(I_3 - I_2)R + \frac{I_3}{j\omega C} + I_3 R = 0$$

$$I_3(2R - 1/j\omega C) - I_2 R = 0 \rightarrow ①$$

At loop 2:- $(I_2 - I_1)R + I_2/j\omega C + (I_2 - I_3)R = 0$

$$-I_1 R + I_2(2R + 1/j\omega C) - I_3 R = 0 \rightarrow ②$$

At loop 1: $(I_1 - I_2)R + I_1/j\omega C + I_1 R_C = -h_{fe} I_b R_C$

$I_1(R + 1/j\omega C) + R_C) - I_1 R = -h_{fe} I_b R_C \Rightarrow h_{fe} I_3 R_C - I_2 R + I_1(R + R_C + 1/j\omega C) = 0$

$$\begin{vmatrix} 2R + 1/j\omega C & -R & 0 \\ -R & 2R + 1/j\omega C & -R \\ h_{fe} R_C & -R & R + R_C + 1/j\omega C \end{vmatrix} = (2R + 1/j\omega C)\left[(2R + 1/j\omega C) \times (R + R_C + 1/j\omega C)\right] - +R\left[-R(R + R_C + 1/j\omega C) + R h_{fe} R_C\right]$$

$$= (2R + 1/j\omega C)\left[2R^2 + 2RR_C + \frac{2R}{j\omega C} + \frac{R}{j\omega C} + \frac{R_C}{j\omega C} - 1/\omega^2 C^2 - R^2\right] + R\left[-R^2 - RR_C - \frac{R}{j\omega C} + R h_{fe} R_C\right]$$

$$= (2R + 1/j\omega C)\left[2R^2 + \frac{3R}{j\omega C} + 2RR_C - 1/\omega^2 C^2 - R^2 + \frac{R_C}{j\omega C}\right] + \left[-R^3 - R^2 R_C - \frac{R^2}{j\omega C} + R^2 h_{fe} R_C\right]$$

$$= 4R^3 + \frac{6R^2}{j\omega c} + 4R^2R_c - \frac{2R}{\omega^2c^2} - 2R^3 + \frac{2RR_c}{j\omega c} - R^3 - R^2R_c - \frac{R^2}{j\omega c} + R^2 h_{fe} R_c$$

$$+ \frac{2R^2}{j\omega c} - \frac{3R}{\omega^2c^2} + \frac{2RR_c}{j\omega c} - \frac{1}{j\omega^3c^3} - \frac{R^2}{j\omega c} - \frac{R_c}{\omega^2c^2}$$

$$= R^3 + 3R^2R_c + R^2 h_{fe} R_c - \frac{1}{j\omega^3c^3} - \frac{5R}{\omega^2c^2} + \frac{6R^2}{j\omega c} + \frac{4RR_c}{j\omega} - \frac{R_c}{\omega^2c^2}$$

$$= \left[R^3 + 3R^2R_c + R^2R_c h_{fe} - \frac{5R}{\omega^2c^2} - \frac{R_c}{\omega^2c^2}\right] + \left[\frac{6R^2}{j\omega c} + \frac{4RR_c}{j\omega c} - \frac{1}{j\omega^3c^3}\right] \rightarrow (1)$$

Equating imaginary part to zero

ie $\frac{6R^2}{j\omega c} + \frac{4RR_c}{j\omega c} - \frac{1}{j\omega^3c^3} = 0$, $\frac{6R^2 + 4RR_c}{j\omega c} - \frac{1}{\omega^3c^3} = 0$

$$\left[\omega^2c^2(6R^2 + 4RR_c) - 1\right] = 0 \Rightarrow \omega^2c^2(6R^2 + 4RR_c) = 1$$

$$\Rightarrow \omega^2c^2 = \frac{1}{6R^2 + 4RR_c}$$

$$\omega^2 = \frac{1}{c^2(6R^2 + 4RR_c)} \Rightarrow \frac{1}{c^2R^2\left[6 + \frac{4R_c}{R}\right]} \rightarrow$$

$$\omega = \frac{1}{RC\sqrt{6 + 4K}} \qquad K = R_c/R$$

$$f = \frac{1}{2\pi RC\sqrt{6 + 4K}} \rightarrow (3)$$

If R_c/R is very small

$$f = \frac{1}{2\pi RC\sqrt{6}}$$

Equating real part to zero

$$R^3 + 3R^2 R_C + R^2 R_C h_{fe} - \frac{5R}{\omega^2 C^2} - \frac{R_C}{\omega^2 C^2} = 0$$

$$R^3 + 3R^2 R_C + R^2 R_C h_{fe} - 5R(6R^2 + 4RR_C) - R_C(6R^2 + 4RR_C) = 0$$ from eqn (

$$R^3 + 3R^2 R_C + R^2 R_C h_{fe} - 30R^3 - 20R^2 R_C - 6R^2 R_C - 4RR_C^2 = 0$$

$$-29R^3 - 23R^2 R_C + R^2 R_C + 4RR_C^2 = 0$$

$$R^2 R_C h_{fe} = 29R^3 + 23R^2 R_C + 4RR_C^2$$

$$h_{fe} = \frac{29R}{R_C} + 23 + \frac{4R_C}{R} \rightarrow (4)$$

Let $k = R_C/R \Rightarrow h_{fe} = \frac{29}{k} + 23 + 4k$

diff w.r.t to 'k'. $\frac{dh_{fe}}{dk} = -\frac{29}{k^2} + 4$

$$\frac{dh_{fe}}{dk} = 0$$

$$\frac{29}{k^2} = 4 \Rightarrow k = 2.7 \rightarrow (5)$$

Substituting k value in (4) $h_{fe} = \frac{29}{2.7} + 23 + 4 \times 2.7$

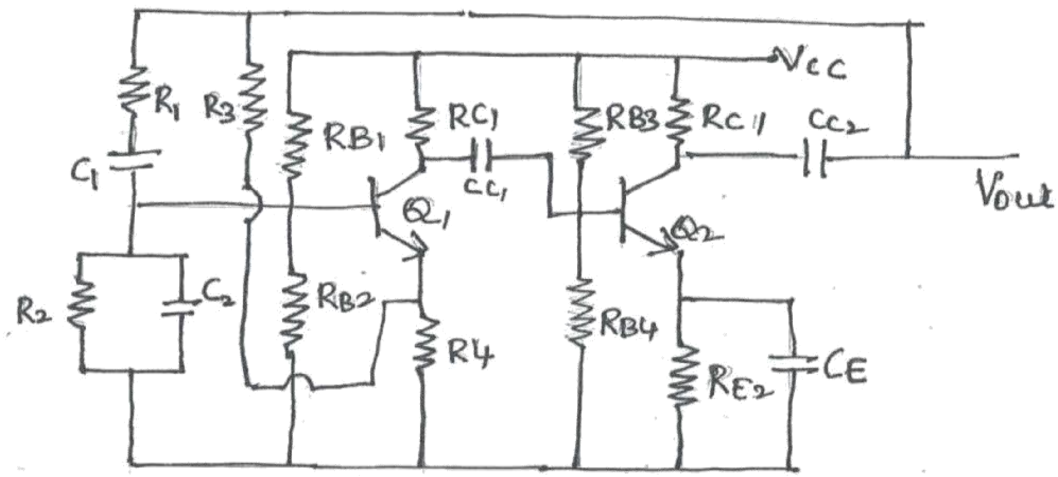
$$= 44.54$$

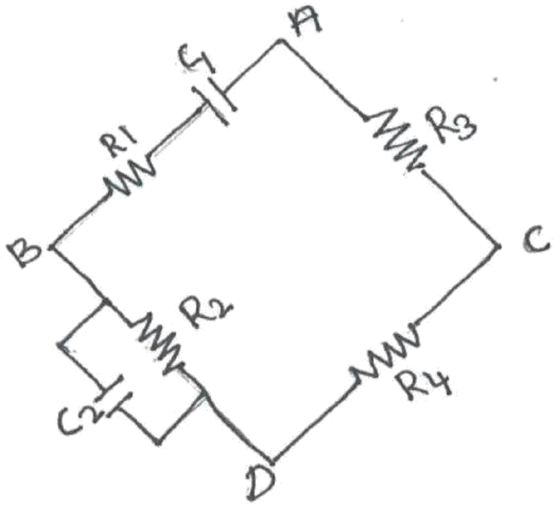
$h_{fe} \geq 44.54$ for sustained Oscillation

For Barkhausen Criterion $A\beta = 1$, $\beta = -1/29$

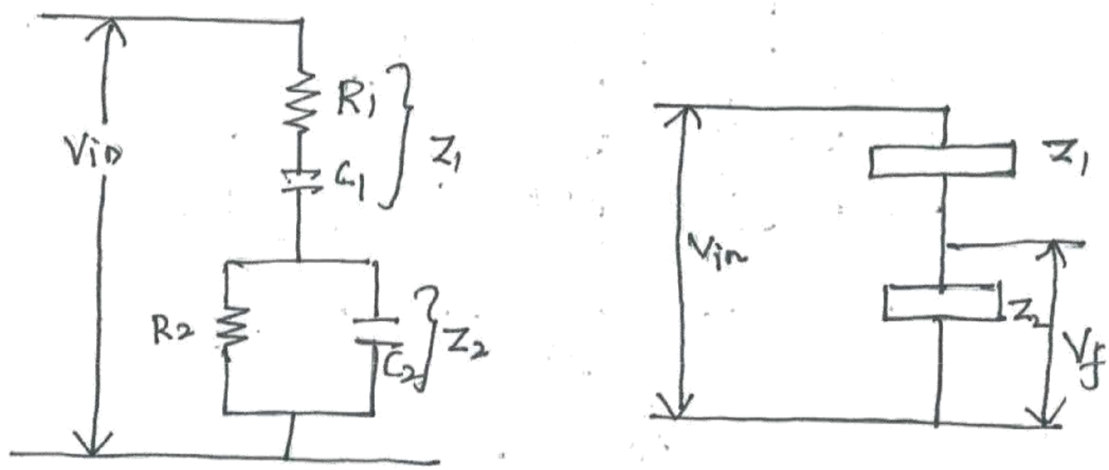
Wein Bridge Oscillator :-

→ It is an audio frequency Oscillator. It is the only one Oscillator involves both positive and negative feedback Negative feedback provides stability and positive feedback provides Oscillation.





The bridge circuit consists of two arms the resistive arm R_3 & R_4 and reactive arms R_1C_1 and R_2C_2. The resistive arm consisting of R_4 introduces negative feedback to the circuit of transistor Q. It improves bias stability. Since arm ACD contains only resistance the feedback provided by the arm is not sensitive to changes in frequency. The amount of feedback is determined only by the voltage divider $R_3 R_4$. The reactive arm contains two RC networks out of which one is and other is in series.



$$Z_1 = R_1 + \frac{1}{j\omega C_1} = \frac{R_1 j\omega C_1 + 1}{j\omega C_1}, \quad Z_2 = R_2 \,||\, 1/j\omega C_2 = \frac{R_2/j\omega C_2}{R_2 + \frac{1}{j\omega C_2}}$$

$$= \frac{R_2}{1 + j\omega C_2 R_2}$$

$$I = \frac{V_{in}}{Z_1 + Z_2}, \quad V_f = I \cdot Z_2 = \frac{V_{in} Z_2}{Z_1 + Z_2}$$

$\beta = \frac{V_f}{V_o}$ Here the output V_o is fed to the input $\therefore V_o = V_{in}$

$$\beta = \frac{Z_2}{Z_1 + Z_2} \rightarrow (3)$$ substituting Z_1, Z_2 in (3)

$$\beta = \frac{\frac{R_2}{1+j\omega R_2 C_2}}{\frac{1+j\omega R_1 C_1}{j\omega C_1} + \frac{R_2}{1+j\omega R_2 C_2}} = \frac{R_2 j\omega C_1}{(1+j\omega R_1 C_1)(1+j\omega R_2 C_2) + R_2 j\omega C_1}$$

$$= \frac{j\omega R_2 C_1}{1 + j\omega R_2 C_2 + j\omega R_1 C_1 + j^2\omega^2 R_1 C_1 R_2 C_2 + j\omega R_2 C_1}$$

$$= \frac{j\omega R_2 C_1}{1 + j\omega R_2 C_2 + j\omega R_1 C_1 - \omega^2 R_1 C_1 R_2 C_2 + j\omega R_2 C_1}$$

$$\beta = \frac{j\omega R_2 C_1}{1 + j\omega (R_2 C_2 + R_1 C_1 + R_2 C_1) - \omega^2 R_1 R_2 C_1 C_2}$$

$$= \frac{j\omega R_2 C_1}{(1 - \omega^2 R_1 R_2 C_1 C_2) + j\omega (R_1 C_1 + R_2 C_2 + R_2 C_1)} \rightarrow (4)$$

multiplying by its conjugate

$$\beta = \frac{j\omega R_2 C_1 [(1 - \omega^2 R_1 R_2 C_1 C_2) - j\omega (R_1 C_1 + R_2 C_2 + R_2 C_1)]}{(1 - \omega^2 R_1 R_2 C_1 C_2)^2 + \omega^2 (R_1 C_1 + R_2 C_2 + R_2 C_1)^2} \rightarrow (5)$$

$$\beta = \frac{j\omega R_2 C_1 (1 - \omega^2 R_1 C_1 R_2 C_2) + \omega^2 R_2 C_1 (R_1 C_1 + R_2 C_2 + R_2 C_1)}{(1 - \omega^2 R_1 C_1 R_2 C_2)^2 + \omega^2 (R_1 C_1 + R_2 C_2 + R_2 C_1)^2} \rightarrow ($$

To have zero phase shift of the feedback network, its imaginary part must be zero. Equating the imaginary part of (6) to be zero, we get

$$\omega R_2 C_1 \left[1 - \omega^2 R_1 C_1 R_2 C_2\right] = 0$$

As $\omega \neq 0$, $R_2 \neq 0$, $C \neq 0$

$$1 - \omega^2 R_1 C_1 R_2 C_2 = 0$$

$$\omega^2 = \frac{1}{R_1 R_2 C_1 C_2} \Rightarrow \omega = \frac{1}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$f = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}} \rightarrow (7)$$

If $R_1 = R_2 = R$, $C_1 = C_2 = C$

$$f = \frac{1}{2\pi RC} \rightarrow (8)$$

If $R_1 = R_2 = R$, $C_1 = C_2 = C$ in eqn (6)

$$\beta = \frac{\omega^2 RC(3RC) + j\omega RC(1 - \omega^2 R^2 C^2)}{(1 - \omega^2 R^2 C^2)^2 + \omega^2 (3RC)^2} \rightarrow (9)$$

Substituting $\omega = \frac{1}{RC}$

$$\beta = \frac{\frac{3R^2C^2}{R^2C^2} + j\frac{RC}{RC}\left(1 - \frac{R^2C^2}{R^2C^2}\right)}{\left(1 - \frac{R^2C^2}{R^2C^2}\right) + \frac{1}{R^2C^2}(3RC)^2} = \frac{3}{9} = 1/3$$

For Sustained Oscillators

$|A\beta| \geq 1 \quad \therefore |A(1/3)| \geq 1 \quad = A \geq 3$

Advantages :-

→ Good stability, By replacing R_2 with a thermistor, the amplitude stability of oscillator output voltage can be increased.

→ Overall gain is high because of two transistors employed in the circuit.

→ Frequency of oscillation can be changed varying R and C

→ Good sine wave output. It does not require inductors.

Disadvantages :-

→ Circuit requires two transistors and large number of components

→ It can't generate very high frequencies.

Frequency Stability of an Oscillator :-

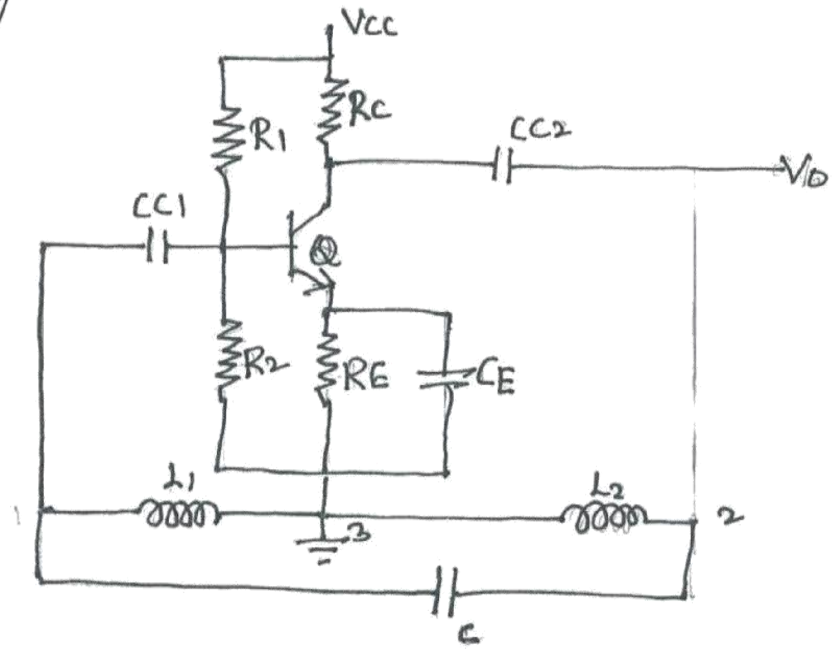
The ability of an oscillator to maintain a constant frequency of oscillation is called frequency stability

1) Operating point of the active devices
2) Inter element capacitances
3) Variations in power supply
4) Temperature variation
5) Output error
6) Mechanical vibrations.

2) Explain the Hartley and Colpits Oscillator and derive its frequency of Oscillation

LC Oscillators :-

i) Hartley Oscillators :-



→ It is an example of LC Oscillator Z_1, Z_2 are inductor and Z_3 is a Capacitor. Resistors R_1 & R_2 & R_E provide the necessary dc bias to the transistor the feedback network consisting of L_1, L_2 and C determines the frequency of Oscillation

→ C_E is the bypass capacitor, CC1 and CC2 are coupling capacitors.

Operation :→ When the supply Voltage +Vcc is switched ON, a transient current is produced in the tank circuit & consequent damped harmonic oscillation are set up in the circuit

→ The oscillatory circuit in the tank circuit produces ac voltage across L_1 and L_2. as terminal 3 is earthed at zero potential → If terminal 1 is at +ive potential with respect to 3 at any instant terminal 2 will be at a negative potential with respect to 3 at the same instant. Thus the phase difference between terminals 1 & 2 is always 180°.

→ In the CE mode, the transistor provides a phase shift of 180°. Thus the total phase shift of 360° is achieved that satisfies barkhausen criterion. Thus the frequency of oscillation is determined by the tank consisting of L_1, C_2 & C. The frequency of Oscillation is given as $f = \frac{1}{2\pi\sqrt{L_{eq}}}$

Where $L_{eq} = L_1 + L_2 + 2M$

$M \rightarrow$ mutual inductance L_1 & L_2

The condition for sustained Oscillation is $h_{fe} \geq \frac{L_1 + M}{L_2 + M}$

Analysis :-

In this Oscillator Z_1 & Z_2 are inductive reactances & Z_3 is capacitance reactance

$Z_1 = j\omega L_1 + j\omega M$, $Z_2 = j\omega L_2 + j\omega M$

$Z_2 = \frac{1}{j\omega C} = -\frac{j}{\omega C}$

Substituting for Z_1 & Z_2 and Z_3 in general eqn

$(Z_1 + Z_2 + Z_3) h_{ce} + Z_1 Z_3 + Z_1 Z_2 (1 + h_{fe}) = 0 \rightarrow (1)$

$\left[(j\omega L_1 + j\omega L_2 + j\omega M + j\omega M - \frac{j}{\omega C}\right] h_{ce} + \left[(j\omega L_1 + j\omega M) \times -\frac{j}{\omega C}\right]$

$+ (j\omega L_1 + j\omega M)(j\omega L_2 + j\omega M)(1 + h_{fe}) = 0 \rightarrow (2)$

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

$+ \left[j^2\omega^2 (L_1 + M)(L_2 + M)(1 + h_{fe})\right] = 0$

$$j\omega h_{ie}\left[(L_1+L_2+2M)-\frac{1}{\omega^2 C}\right]+j^2\omega\left[(L_1+M)\left(-\frac{1}{\omega C}\right)-\omega^2(1+h_{fe})\right.$$
$$\left.(L_1L_2+L_1M+L_2M+M^2)\right]=0$$

Multiplying and dividing by ω throughout

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

$$j\omega h_{ie}\left[(L_1+L_2+2M)-\frac{1}{\omega^2 C}\right]-\omega^2(L_1+M)\left[(L_2+M)(1+h_{fe})-\frac{1}{\omega^2 C}\right]=0 \rightarrow (3)$$

Equating imaginary part eqn (3) to zero

$$L_1+L_2+2M-\frac{1}{\omega^2 C}=0 \rightarrow (4) \quad [\because \omega\neq 0 \text{ \& } h_{ie}\neq 0$$

$$\frac{1}{\omega^2 C}=L_1+L_2+2M \qquad \omega=\frac{1}{\sqrt{(L_1+L_2+2M)C}} \rightarrow (5)$$

$$f=\frac{1}{2\pi\sqrt{(L_1+L_2+2M)C}} \rightarrow (6)$$

$$f=\frac{1}{2\pi\sqrt{L_{eq}C}} \rightarrow (7) \quad \text{where } L_{eq}=L_1+L_2+2M$$

The condition for sustained oscillation is obtained by equating the real part of eqn(3) to zero.

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

$\because \omega^2 \neq 0$, $L_1+M \neq 0$, we get

$(L_2+M)(1+h_{fe}) - 1/\omega^2 C = 0$, substituting for ω^2 from eqn ①

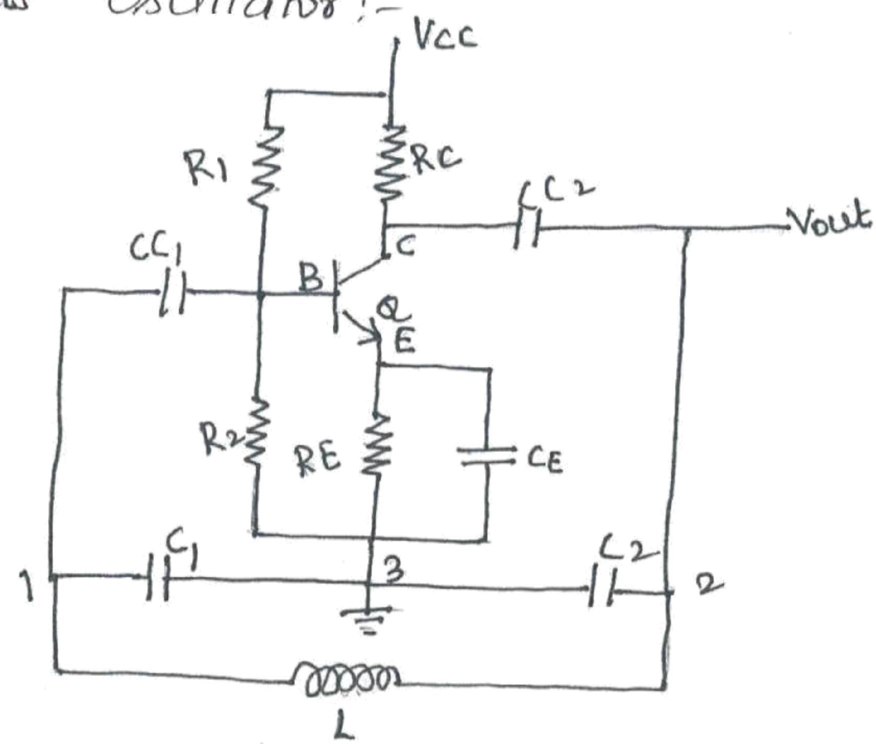
$$(L_2+M)(1+h_{fe}) - \frac{(L_1+L_2+2M)C}{C} = 0 \Rightarrow L_2 + L_2 h_{fe} + M + M h_{fe} - L_1 - L_2 - 2M = 0$$

$h_{fe}(L_2+M) - M - L_1 = 0 \quad \therefore h_{fe}(L_2+M) = L_1+M$

$$h_{fe} = \frac{L_1+M}{L_2+M}$$

It is the condition for sustained oscillation.

Colpitts Oscillator:-



→ In colpits Z_1 & Z_2 are the capacitors and Z_3 is an inductor. The resistors R_1, R_2 and R_E provides the necessary dc bias to the transistor. The bypass capacitor is C_E. The coupling capacitors are C_{C1} & C_{C2}. The feedback network consisting of C_1, C_2 & L determines the frequency of oscillator. Operation is similar to the Hartley oscillator.

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}} \quad \text{Where } C_{eq} = \frac{C_1 C_2}{C_1 + C_2}$$

Application:-

1) It is used for commercial signal generators for freq 1M to 800 MHz

2) It is used as local oscillator in super heterodyne radio receiver.

Analysis:-

$$Z_1 = \frac{1}{j\omega C_1} = -\frac{j}{\omega C_1}, \quad Z_2 = \frac{1}{j\omega C_2} = -\frac{j}{\omega C_2}$$

$$Z_3 = j\omega L$$

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

Substituting Z_1, Z_2 + Z_3 in above eqn

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

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

$$j h_{ie}\left[\omega L - 1/\omega C_1 - 1/\omega C_2\right] + \left[\frac{L}{C_1} - \frac{(1+h_{fe})}{\omega^2 C_1 C_2}\right] = 0 \rightarrow (2)$$

Equating the imaginary part of eqn to zero

$$\omega L - 1/\omega C_1 - 1/\omega C_2 = 0 \rightarrow (3)$$

$$\omega L = 1/\omega C_1 + 1/\omega C_2, \quad \omega L = 1/\omega\left[1/C_1 + 1/C_2\right]$$

$$\omega^2 = \frac{1}{L}\left[\frac{C_1 + C_2}{C_1 C_2}\right] \rightarrow (4)$$

$$\omega = \sqrt{\frac{C_1 + C_2}{L C_1 C_2}} \rightarrow (5)$$

$$f = \frac{1}{2\pi}\sqrt{\frac{C_1 + C_2}{L C_1 + C_2}}$$

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}} \rightarrow (6)$$

Where $C_{eq} = \dfrac{C_1 C_2}{C_1 + C_2} \rightarrow (7)$

The condition for sustained oscillation is

Obtained by equating the real part of eqn(2) to zero

$$\frac{L}{C_1} - \frac{(1+h_{fe})}{\omega^2 C_1 C_2} = 0 \rightarrow (8)$$

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

$$L = \frac{(1+h_{fe})}{\omega^2 C^2}$$

Substituting for ω^2 from eqn (4)

$$L = \frac{(1+h_{fe}) L C_1 C_2}{C_2 (C_1 + C_2)}$$

$$L = \frac{(1+h_{fe}) L C_1}{C_1 + C_2}$$

$$= C_1 C_2 = (1+h_{fe}) C_1$$

$$C_1 \left[1 + \frac{C_2}{C_1}\right] = (1+h_{fe}) C_1$$

$$h_{fe} = \frac{C_2}{C_1} \rightarrow (9)$$

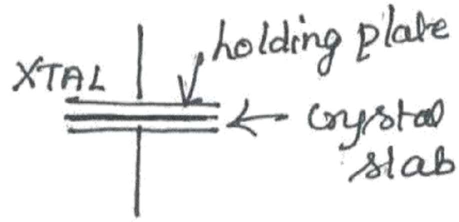
This is the condition for sustained oscillation.

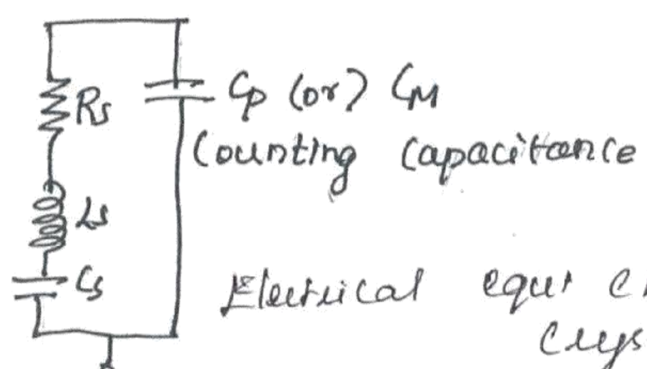
3) Explain the Construction of Quartz Crystal and derive their frequency of oscillation of Crystal Oscillator.

Crystal Oscillator:-

→ A crystal oscillator is basically a tuned circuit oscillator using a piezoelectric crystal as its resonant tank circuit. The crystal has a great stability in holding constant at whatever frequency the crystal is originally cut to operate.

→ They are used whenever great stability is required for ex in watches. communication transmitters and receivers.





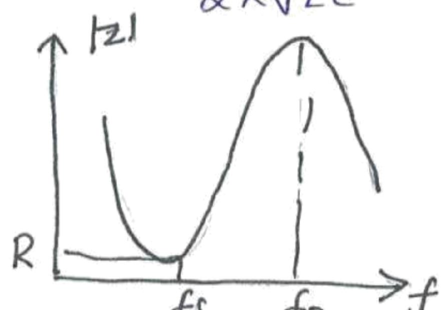
Electrical equi ckt of crystal

Characteristics of Crystal:-

→ A quartz crystal exhibits the property that when mechanical stress is applied across the faces of the crystal, a difference of potential across opposite faces of the crystal. This property of a crystal is called piezoelectric effect.

The series resonant freq is given as

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



Analysis:-

The net reactance of the circuit is given as

$$Z = \frac{\left(j\omega L_s + \frac{1}{j\omega C_s}\right) \cdot \frac{1}{j\omega C_p}}{j\omega L_s + \frac{1}{j\omega L_s} + \frac{1}{j\omega C_p}} \text{ (neglecting } R_s)$$

$$= \frac{\frac{j^2\omega^2 L_s C_s + 1}{(j\omega C_s)(j\omega C_p)}}{\frac{j^3\omega^3 L_s C_s C_p + j\omega C_p + j\omega C_s}{(j\omega C_s)(j\omega C_p)}}$$

$$= \frac{j^2\omega^2 L_s C_s + 1}{-j\omega^3 L_s C_s C_p + j\omega(C_p + C_s)} = \frac{1 - \omega^2 L_s C_s}{-j\omega L_s C_s C_p\left[\omega^2 - \frac{(C_p + C_s)}{\ldots}\right]}$$

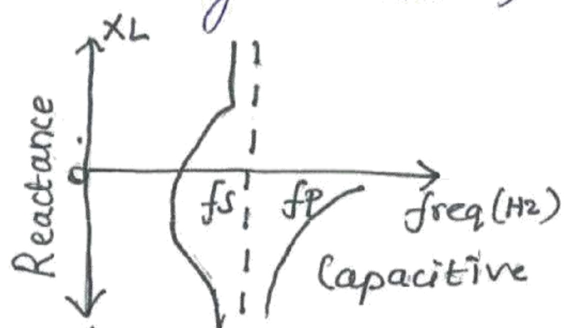
dividing numerator & denominator by L_sC_s

$$Z = \frac{\frac{1}{L_sC_s} - \omega^2}{-j\omega C_p\left[\omega^2 - \frac{(C_p+C_s)}{L_sC_sC_p}\right]} = \frac{\omega^2 - \omega_s^2}{j\omega C_p(\omega^2 - \omega_p^2)}$$

Where $\omega_s^2 = \frac{1}{L_sC_s}$, $\omega_p^2 = \frac{1}{L_s\left(\frac{C_sC_p}{C_s+C_p}\right)}$

Series resonant frequency is given as } $f_s = \frac{1}{2\pi\sqrt{L_sC_s}}$ &

parallel resonant frequency is given as } $f_p = \frac{1}{2\pi\sqrt{L_sC_{eq}}}$; $C_{eq} = \frac{C_sC_p}{C_s+C_p}$



Freq response curve

When $\omega < \omega_s$ → the reactance is capacitance i.e crystal act as a capacitive reactance

$\omega_s < \omega < \omega_p$ → the reactance is inductive

$\omega = \omega_s$ → the reactance is zero

$\omega > \omega_p$ → the reactance is capacitive

If $f_s < f < f_p$ then crystal can be used as an inductance osc ckt.

Where f_s, f_p Series and parallel resonant freq, f → freq of vibration

Q factor at $f_s = \frac{\omega_s L}{R_s}$; Q factor at $f_p = \frac{\omega_p L}{R_s}$

The freq stability of the crystal depends on
i) High value of Q, ii) Drift with time temperature

Effect of temperature :-

→ The resonant frequency of a crystal varies with temperature. A x-cut crystal have -ive temperature co-efficient i.e the resonant frequency decreases with increase in temperature.

→ A Y cut XTAL have +ive tempe coefficient i.e the resonant freq increases with increase in temp

∴ to stabilise the resonant frequencies against variation in temperature.

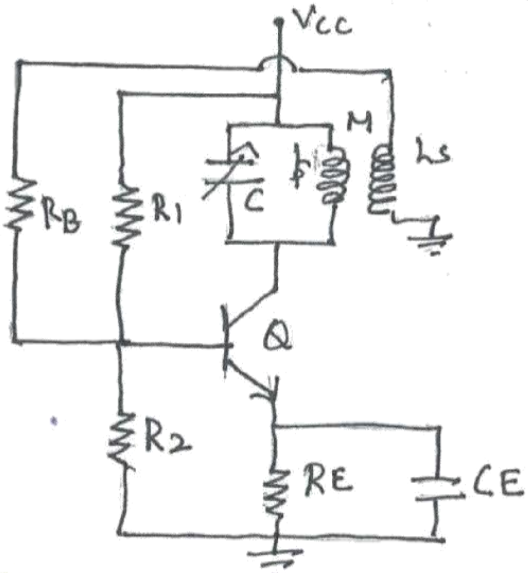
i) The angle of cut of XTAL should have zero temperature co-efficient.

2) The crystal can be kept in thermostatically controlled container. Since the Q factor is very high and if the resonant freq is stabilised against variation in temperature then the freq stability is high for a crystal oscillator which is greater

4) Write short notes on

a) Tuned collector and Miller Oscillator

Tuned Collector Oscillator :-



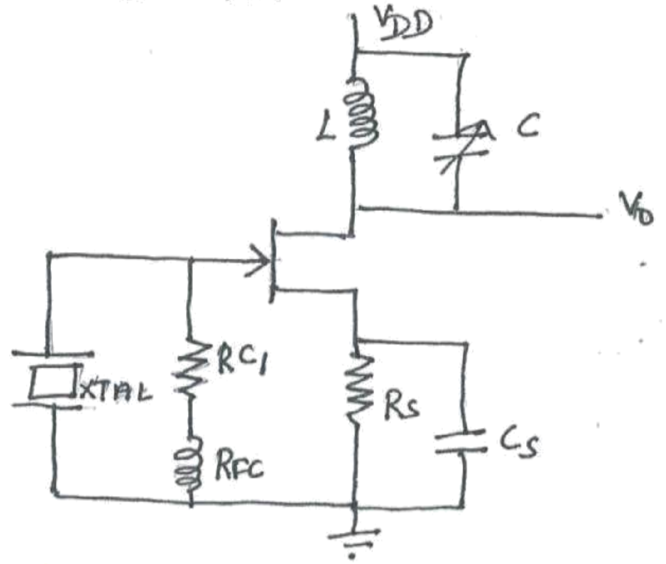
→ There is a tuned LC circuit in the collector branch which is connected to VCC.

→ Here the CE amplifier introduces a 180° phase shift & additional P.S required for sustained oscillation is introduced by inductor Ls which is coupled with inductor Lp of tank circuit $f = \frac{1}{2\pi\sqrt{L_p C}}$

→ Capacitor C_E is connected to emitter by bypass capacitor

→ Resistor R_B is used to control the amou of Fb.

Miller Oscillator:-



→ This is a crystal controlled Oscillator Circuit the crystal is operated in its parallel response mode where it offers maximum impedance

→ The crystal is connected across the gate to source of the FET amplifier. The tank circuit is connected at the drain terminal

→ At the resonant frequency of the crystal the gate source bias is minimum and so the drain current is maximum. Hence the LC tank circuit is set into sustained oscillations.

→ At frequencies other than resonant frequ the gate to source voltage is maximum & hence the drain current is low which results in damped oscillations.

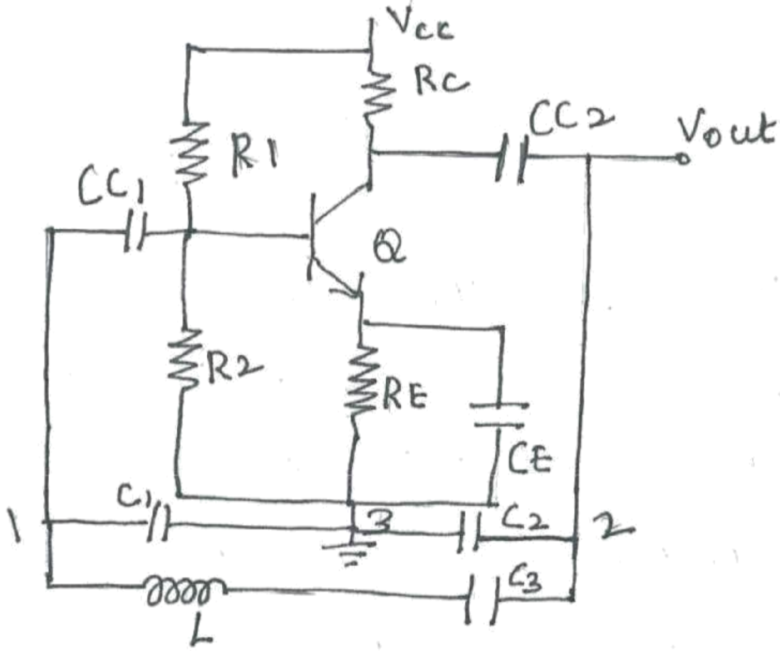
→ Hence the circuit results in undamped or sustained oscillations only at the crystal resonant frequency which results in better freq stability. The freq of oscillations is decided by the LC value of the tank circuit.

→ In the fig. the crystal forms one arm and the tank circuit forms the other arm. The interjunction capacitance Cgd acts as the capacitor between D and G. Since the ckts involves the all ack is Cgd it is called as miller oscillator.

b) Clapp Oscillator and Franklin Oscillator

Clapp Oscillator:-

→ It is also a LC oscillator it is modified version of colpitts oscillator. A capacitor C_3 is added in series with the inductor in the tank circuit.



The frequency of oscillation is given as

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}}$$

Where $C_{eq} = \frac{C_1 C_2 C_3}{C_1 C_2 + C_2 C_3 + C_3 C_1}$

Since C_3 is small compared to C_1 and C_2 at high frequencies. The resonant frequency is

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}}$$

Analysis :-

In Clapp Oscillator Z_1 and Z_2 are capacitor and Z_3 is the series combination of an Inductor L and Capacitor C_3.

Therefore $Z_1 = \frac{1}{j\omega C_1}$, $Z_2 = \frac{1}{j\omega C_2}$, $Z_3 = j\omega L + \frac{1}{j\omega C_3}$

Substituting all the eqn in general eqn,

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

$$\left(\frac{1}{j\omega C_1} + \frac{1}{j\omega C_2} + j\omega L + \frac{1}{j\omega C_3}\right) h_{ie} + \frac{1}{j\omega C_1}\left(j\omega L + \frac{1}{j\omega C_3}\right) + \frac{1}{j\omega C_1} \cdot \frac{1}{j\omega C_2}(1 + h_{fe}) = 0$$

$$\left(\frac{-j}{\omega C_1} - \frac{j}{\omega C_2} + j\omega L - j/\omega C_3\right) h_{ie} + \left(\frac{-j}{\omega C_1}\right)\left(j\omega L - j/\omega C_3\right) + \left(\frac{-j}{\omega C_1}\right)\left(\frac{-j}{\omega C_2}\right)(1 + h_{fe}) = 0$$

$$j\left[\omega L - \left(\frac{1}{\omega C_1} + \frac{1}{\omega C_2} + \frac{1}{\omega C_3}\right)\right] h_{ie} + \frac{\omega L}{\omega C_1} - \frac{1}{\omega^2 C_1 C_3} - \frac{(1 + h_{fe})}{\omega^2 C_1 C_2} = 0$$

$$j\left[\omega L - \left(\frac{1}{\omega C_1} + \frac{1}{\omega C_2} + \frac{1}{\omega C_3}\right)\right] h_{ie} + \frac{L}{C_1} - \frac{1}{\omega^2}\left(\frac{1}{C_1 C_3} + \frac{(1 + h_{fe})}{C_1 C_2}\right) =$$

$$j\omega h_{ie}\left(L - \frac{1}{\omega^2}\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right)\right) + \frac{L}{C_1} - \frac{1}{\omega^2}\left(\frac{1}{C_1C_3} + \frac{(1+h_{fe})}{C_1C_2}\right) = 0 \quad \to (2)$$

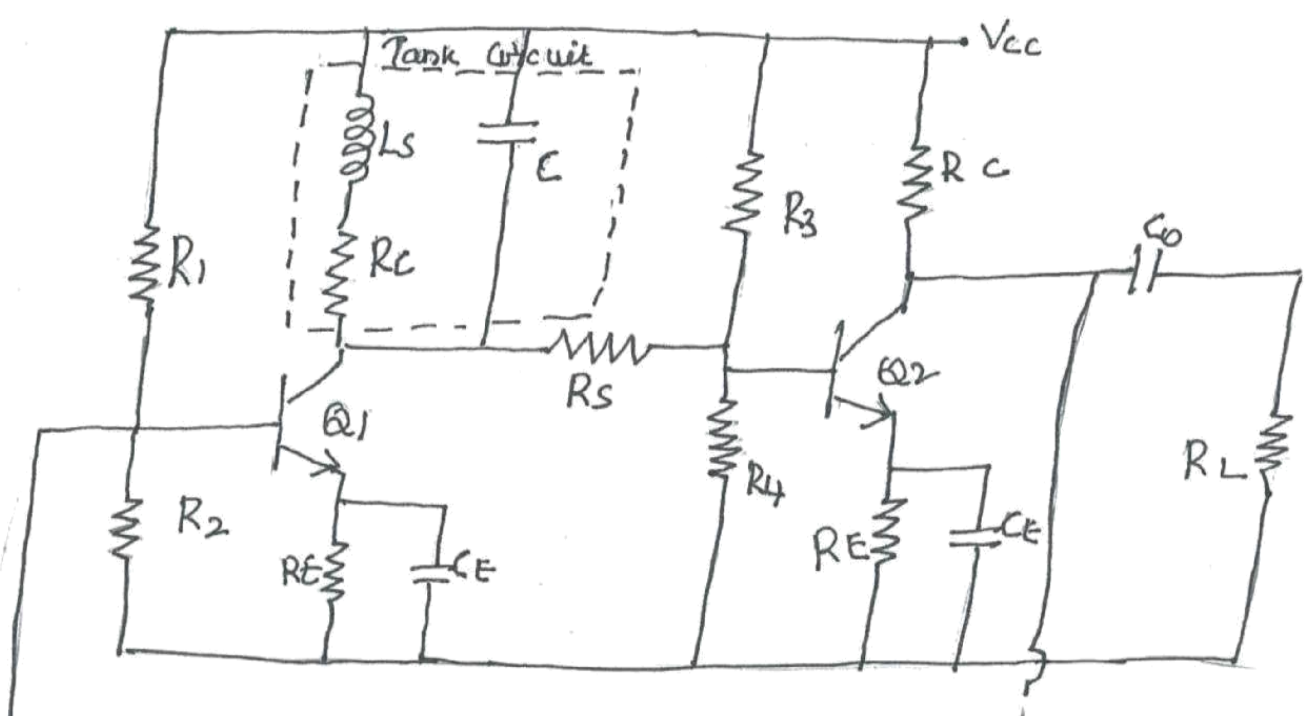
Equating the imaginary part of eqn (2) to zero

$$L - \left[\frac{1}{\omega^2}\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right)\right] = 0 \Rightarrow L = \frac{1}{\omega^2}\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right)$$

$$\omega = \sqrt{\frac{C_2C_3 + C_1C_3 + C_1C_2}{L C_1 C_2 C_3}}, \quad f = \frac{\omega}{2\pi} = \frac{1}{2\pi}\sqrt{\frac{C_1C_2 + C_2C_3 + C_3C_1}{L C_1 C_2 C_3}}$$

$$f = \frac{1}{2\pi\sqrt{L C_{eq}}} \quad \text{where } C_{eq} = \frac{C_1C_2C_3}{C_1C_2 + C_2C_3 + C_3C_1}$$

Franklin Oscillator:-



→ On applying the bias the devices conduct and they result in an o/p at the collector of Q_2 which is fedback to the base of Q_1 in phase. This increases the base bias resulting in an increased collector current of Q_1. Since the feedback is +ive (or) regenerative feedback the magnitude of collector current keeps increasing.

→ Meanwhile the capacitor of the tank ckt will be charged towards the biasing voltage and due to difference in concentration of e-ns in the plates of the capacitor the e-n move from the region of higher to the lower concentration. While moving they setup a magnetic field in the inductor.

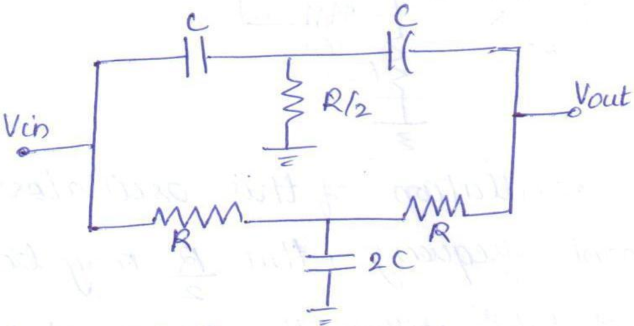
→ Once the charge concentration of both the plates are equal. the transfer of e-ns is stopped and the magnetic fields stop expanding. Since there is no further expansion the magnetic field collapses. As per Lenz law the collapsing magnetic field forces the e-n movement in the same direction so that the voltage across the capacitor is of opposite polarity. The above process is repeated in the opposite direction and the charges back to the original polarity.

→ Thees the capacitor charges bin +V and -V resulting in an oscillatory output. the frequency of oscillation is given as

$f = \frac{1}{2\pi\sqrt{LC}}$. The losses across the series resistance of the inductor R_2 is compensated by the collector current obtained from Q_1. Thus the oscillations of the circuit are undamped.

Twin_T_Oscillator:

-It is a basic lead lag network (i.e. a combinati of low pass & high pass network) its phase characteristics is shown below.



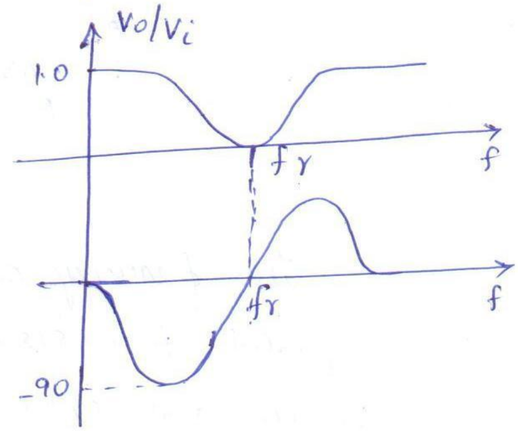


fig: Twin T n/w.

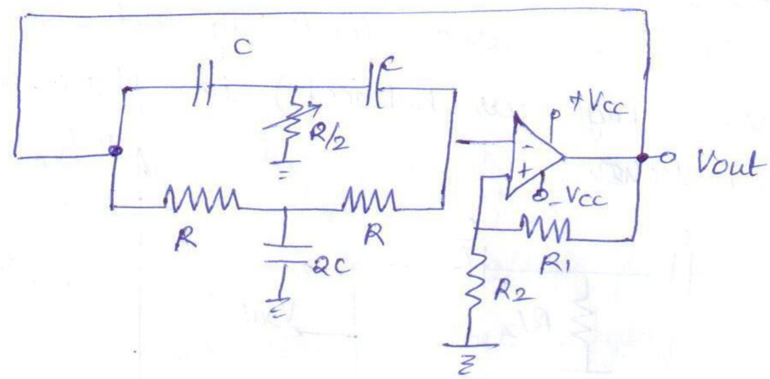
fr is the resonant frequency at which the phase shift equals 0°. The voltage gain is maximum at low & high frequencies and minimum at the resonant freq(..)

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

The Twin-T n/w shows that R₁ & (acts as a potential divider & provides +ve feedback & the n/w acts as a -ve feedback.

The n/w acts

When Vcc is applied the resistance R is low & the +ve fb is maximum, hence the capacitor is charged toward the maximum voltage, it will cause damped oscillati As oscillation builds up, R1 ↑es & +ve fb ↓es. If +ve f ↓, the oscillations become constant. Oscillation cannot occur at any other freq. other than fr, becoz the R/2 -ve fb provided by Twin-T n/w not allowed for oscillation i.e. only at fr, the -ve fb is neglected. Thus to ensure tha the oscillation freq. is near fr, R/2 of the twin T n/w is kept variable.



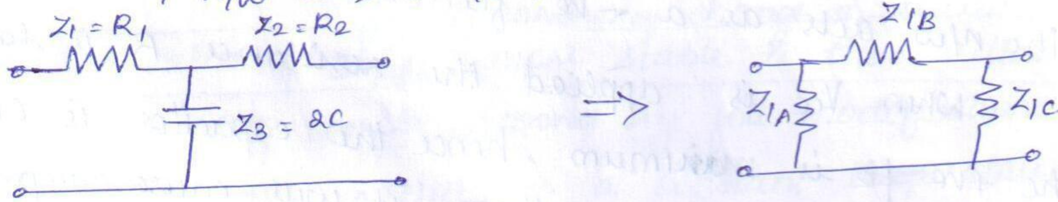
The frequency of oscillation of this oscillator is slig[htly] different from resonant frequency, thus $\frac{R}{2}$ may be slightly varied and ratio of R_1/R_2 within the range of 10 to 1000. This results in the oscillator forced to operate near the resonant or notch frequency.

Main drawback:

It is suitable for single frequency & can't be easily adjusted to other large frequency range.

Analysis:

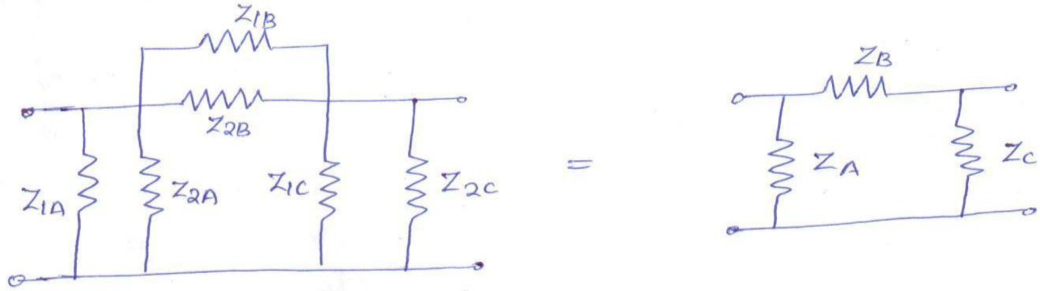
T N/w ⟹ π N/w.



$$Z_{1A} = \frac{\Sigma Z_1 Z_2}{Z_3} \qquad Z_{1B} = \frac{\Sigma Z_2 Z_3}{Z_1} \qquad Z_{1C} = \frac{\Sigma Z_3 Z_1}{Z_2}$$

$$Z_{2A} = \frac{\Sigma Z_1' Z_2}{Z_3} \qquad Z_{2B} = \frac{\Sigma Z_2' Z_3}{Z_1} \qquad Z_{2C} = \frac{\Sigma Z_3' Z_2}{Z_2}$$

The 2 π networks are combined & simplified as follows.



$$Z_A = \frac{Z_{1A} Z_{2A}}{Z_{1A} + Z_{2A}} \; ; \; Z_B = \frac{Z_{1B} Z_{2B}}{Z_{1B} + Z_{2B}} \; ; \; Z_C = \frac{Z_{1C} Z_{2C}}{Z_{1C} + Z_{2C}}$$

At f_o, no signal txmtd. thus $Z_A = Z_C = 0$ and $Z_B = \infty$, where Z_A & Z_C is zero, if denominator of above expression are ∞ & Z_B is infinite if denominator term is zero then we get,

$$Z_{1B} + Z_{2B} = 0$$

$$Z_{2B} = \frac{-X_2 X_1}{R_3} - j(X_1 + X_2) \; ; \; Z_{1B} = R_1 + R_2 + j\frac{R_1 R_2}{X_3}$$

thus $$\left(R_1 + R_2 + j\frac{R_1 R_2}{X_3}\right) - j\left[\frac{X_1 X_2}{R_3} - j(X_1 + X_2)\right] = 0.$$

(or) $$\left[R_1 + R_2 - \frac{X_1 X_2}{R_2}\right] - j\left[\frac{R_1 R_2}{X_3} - (X_1 + X_2)\right] = 0.$$

To find the frequency of oscillation equating real part to zero, thus

$$(R_1 + R_2) R_2 = X_1 X_2$$

If $R_1 = R_2$ & $X_1 = X_2$ then $(R_1 + R_2) R_3 = X_1 X_2 = X_1^2$.

$$2 R_1 R_3 = \frac{1}{\omega^2 C_1^2}$$

$$\omega^2 = \frac{1}{2 R_1 R_3 C_1^2}$$

$$\omega = \frac{1}{\sqrt{2 R_1 R_3 C_1^2}}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{2 R_1 R_3 C_1^2}}}$$

1. In a certain oscillator circuit, the gain of the amplifier is $\left[-\frac{16\times10^6}{j\omega}\right]$ and the feedback factor of the feedback network is $\frac{10^3}{[2\times10^3+j\omega]^2}$. Verify the Barkhausen Criterion for the sustained oscillations. Also find the frequency at which circuit will oscillate. [NOV/DEC 2011, 2012, 6 Marks]

Soln: Gn: $A = -\frac{16\times10^6}{j\omega}$ and $\beta = \frac{10^3}{[2\times10^3+j\omega]^2}$

To satisfy the Barkhausen condn, $|A\beta|=1$ at a frequency for which $\angle A\beta = 0°$.

$A\beta$ is rectangular form, $A\beta = \frac{-16\times10^6\times10^3}{j\omega[2\times10^3+j\omega]^2}$

$$= \frac{-16\times10^9}{j\omega[4\times10^6+4j\omega\times10^3+(+j^2\omega^2)]} = \frac{-16\times10^9}{4\times10^6 j\omega + j^2\omega^2 4\times10^3 - j\omega^3}$$

$$= \frac{-16\times10^9}{j\omega[4\times10^6-\omega^2]-[\omega^2\times4\times10^3]}$$

Rationalizing the denominator function we get,

$$A\beta = -\frac{16\times10^9[-4\times10^3\omega^2 - j\omega(4\times10^6-\omega^2)]}{\{(-4\times10^3\omega^2)+j\omega(4\times10^6-\omega^2)\}\{-(4\times10^3\omega^2)-j\omega(4\times10^6-\omega^2)\}}$$

Using $(a-b)(a+b)=a^2-b^2$ in the denominator,

$$A\beta = \frac{16\times10^9[4\times10^3\omega^2+j\omega(4\times10^6-\omega^2)]}{(-4\times10^3\omega^2)^2-[j\omega(4\times10^6-\omega^2)]^2}$$

$$A\beta = \frac{16\times10^9[4\times10^3\omega^2+j\omega(4\times10^6-\omega^2)]}{16\times10^6\omega^4+\omega^2(4\times10^6-\omega^2)^2}$$

To have $\angle A\beta = 0°$, the imaginary part of $A\beta$ should be zero. This is possible when

$$\omega(4\times10^6-\omega^2)=0.$$

$\omega = 0$; $4\times10^6-\omega^2=0$

$\omega^2 = 4\times10^6$

$\boxed{\omega = 2\times10^3}$

At $\omega = 2\times10^3$ rad/sec $|A\beta|$ can be obtained as

$$|A\beta| = \frac{16\times10^9[4\times10^3\omega^2]}{16\times10^6\omega^4+\omega^2(4\times10^6-\omega^2)^2}$$

$$= \frac{2.56\times10^{20}}{2.56\times10^{20}} = 1.$$

$\therefore$ At $\omega = 2\times10^3$ rad/sec, $\angle A\beta = 0°$ as imaginary part is zero while $|A\beta| = 1$. Thus Barkhausen criterion is satisfied.

$$\omega = 2\times10^3 \text{ rad/sec}$$

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

$$f = \frac{\omega}{2\pi} = \frac{2\times10^3}{2\times\pi} = 318.309\ Hz.$$

$$\boxed{f = 318.309\ Hz}$$

---

2. Find the capacitor C and h_{fe} for the transistor to provide a resonating frequency of 10 KHz of a transistorised phase shift oscillator. Assume $R_1 = 25k\Omega$, $R_2 = 57k\Omega$, $R_C = 20k\Omega$, $R = 7.1k\Omega$ & $h_{ie} = 1.8k\Omega$.

Sol:- For a transistorised phase shift oscillator

$$\boxed{f_r = \frac{1}{2\pi RC\sqrt{6+4k}}}$$

* $k = R_C/R$
* where $R = R_i' + R_3$
* $R_i' = R_1 || R_2 || h_{ie}$

Gn: $R_1 = 25k\Omega$, $R_2 = 57k\Omega$, $R_C = 20k\Omega$, $R = 7.1k\Omega$
$h_{ie} = 1.8k\Omega$ & $f_r = 10kHz$.

To find R:

$$R = R_i' + R_3.$$

$$R_i' = R_1 || R_2 || h_{ie}$$

$$= 25\times10^3 || 57\times10^3 || 1.8\times10^3$$

$$R_i' = 1.631\ k\Omega$$

$$k = R_C/R = \frac{20}{7.1} = 2.816$$

$$R_3 = R - R_i'$$

$$R_3 = 7.1\times10^3 - 1.631\times10^3$$

$$R_3 = 5.47\ k\Omega$$

$$f_r = \frac{1}{2\pi RC\sqrt{6+4k}} \Rightarrow C = \frac{1}{2\pi R f_r\sqrt{6+4k}} \Rightarrow C = \frac{1}{2\pi\times7.1\times10^3\times10\times10^3\sqrt{6+4(2.8}}$$

$$\boxed{C = 539.45\ PF}$$

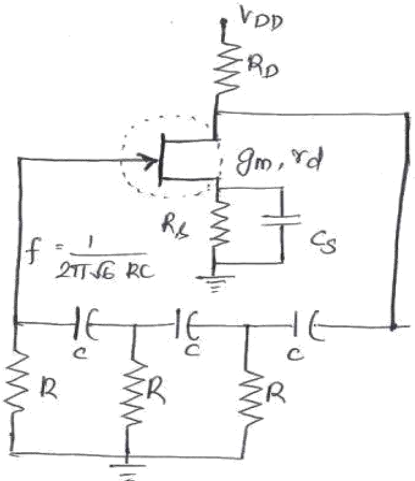
To find h_{fe}:

$$\boxed{h_{fe} \geq 4k + 23 + \frac{29}{k}}$$

$$h_{fe} \geq 4\times2.816 + 23 + \frac{29}{2.816}$$

$$\boxed{h_{fe} \geq 44.562}$$

8. It is desired to design a phase shift oscillator using a FET shown in fig. below having $g_m = 5000\mu S$, $r_d = 40k\Omega$, and a feedback circuit value of $R = 10k\Omega$. Select the value of C for oscillator operation at 1kHz and R_D for $A > 29$ to ensure oscillator action.



Gn: $R = 10k\Omega$, $r_d = 40k\Omega$, $g_m = 5000\mu S$, $f = 1kHz$

W.k.t. $|A| = g_m R_L$ $\quad R_L = R_D \| r_d$

$$f = \frac{1}{2\pi RC\sqrt{6}} \qquad R_L = \frac{R_D r_d}{R_D + r_d}$$

To find C: $C = \dfrac{1}{2\pi R f\sqrt{6}}$

$$C = \frac{1}{2\pi \times 10 \times 10^3 \times 1 \times 10^3 \sqrt{6}}$$

$$C = 6.497 \times 10^{-9} F$$

$$\boxed{C = 6.5nF}$$

To find R_D: First have to calculate - $R_L = \dfrac{|A|}{g_m}$

Assuming $A = 40$ $(A > 29)$ $\quad R_L = \dfrac{40}{5000 \times 10^{-6}}$

$$\boxed{R_L = 8k\Omega}$$

W.k.t. $R_L = \dfrac{R_D r_d}{R_D + r_d}$.

$$R_L R_D + r_d R_L = R_D r_d$$

$$R_D(R_L - r_d) + r_d R_L = 0.$$

$$R_D = \frac{-r_d R_L}{R_L - r_d}$$

$$R_D = \frac{r_d R_L}{r_d - R_L}$$

$$R_D = \frac{40 \times 10^3 \times 8 \times 10^3}{(40-8) \times 10^3}$$

$$\boxed{R_D = 10k\Omega}$$

Ans: $\boxed{C = 6.5nF \;;\; R_D = 10k\Omega}$ (98)

4. Design a RC phase shift oscillator to generate 5KHz sine wave with 20V peak to peak amplitude. Draw the design circuit. Assume $h_{fe} = 150$ [APR/MAY 2006]

Sln/-: For RC phase shift oscillator,

Gn: $h_{fe} = 150$
$f = 5kHz$.

$$h_{fe} = 4K + 23 + \frac{29}{K}$$

$$150 = \frac{4K^2 + 23K + 29}{K} \Rightarrow 4K^2 - 127K + 29 = 0.$$

i.e. $K = 31.52$
$K = 0.23$

Choosing lower value of $K = 0.23$

$$f = \frac{1}{2\pi RC\sqrt{6+4K}}$$

Choosing $C = 1000\ pF$

$$5 \times 10^3 = \frac{1}{2\pi R \times 1000 \times 10^{-12} \times \sqrt{6 + 4 \times 0.23}}$$

$$R = 12.1\ k\Omega$$

$$\boxed{R \simeq 12\ k\Omega}$$

$$K = \frac{R_C}{R} \Rightarrow R_C = KR$$

$$= 0.23 \times 12 \times 10^3$$

$$\boxed{R_C = 2.7\ k\Omega}$$

Assuming the biasing resistances ~~affects~~ very large & neglecting it.

Select the transistor with $h_{ie} = 2k\Omega$.

$$R_i' = h_{ie} = 2k\Omega$$

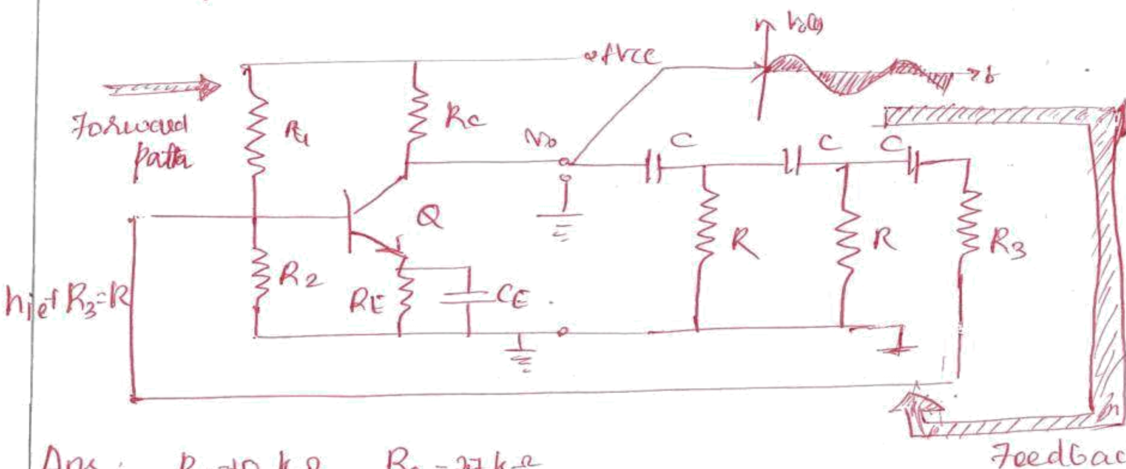
∴ The feedback resistance R_3 i.e. last resistance in the phase shift network is $R_3 = -R_i' + R$

$$R_3 = 12 \times 10^3 - 2k.$$

$$\boxed{R_3 = 10\ k\Omega}$$

The output amplitude can be controlled 10 V i.e. 20 V peak to peak, by using back to back connected zener diodes of 9.3 V (V_z) each at the output of emitter follower. The zener diode 9.3 V & forward biased diode of 0.7 V gives the total 10V.

The designed circuit:



Ans: $R_3 = 10\ k\Omega$ $\quad R_C = 27\ k\Omega$

$R = 12\ k\Omega$ $\quad C = 1000\ \mu F$.

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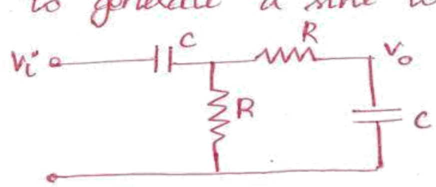
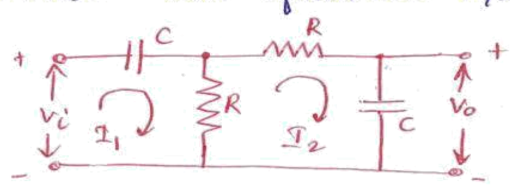
5. Design an oscillator with network shown in the feedback path of the amplifier to generate a sine wave of 2 kHz. (12 Marks) APR/MAY '06



Consider the feedback n/w as shown below,



Applying KVL to the 2 loops,

$$V_i = I_1 \frac{1}{j\omega C} + R(I_1 - I_2) \quad \text{——— (1)}$$

$$0 = (I_2 - I_1)R + I_2 R + I_2 \frac{1}{j\omega C} \quad \text{——— (2)}$$

$$\therefore I_1\left[R-\frac{j}{\omega C}\right]-I_2 R = V_i \qquad [1/j=-j] \quad \text{—— (1).a.}$$

$$-I_1 R + I_2\left[2R-\frac{j}{\omega C}\right]=0 \qquad [1/j=-j] \quad \text{—— (2).a.}$$

$$\therefore \quad D = \begin{vmatrix} R-j/\omega C & -R \\ -R & 2R-j/\omega C \end{vmatrix}$$

$$D = R^2-\frac{1}{\omega^2C^2}-j\frac{3R}{\omega C}$$

$$D_2 = \begin{vmatrix} R-j/\omega C & V_i \\ -R & 0 \end{vmatrix} = V_i R$$

But $V_o = I_2 \dfrac{1}{j\omega C}$

$$= -I_2\frac{j}{\omega C} \qquad \text{where } I_2=\frac{D_2}{D}.$$

$$\therefore \quad V_o = \frac{-j}{\omega C} \times \frac{V_i R}{R^2-\frac{1}{\omega^2C^2}-j\frac{3R}{\omega C}}$$

$$\therefore \beta = \frac{V_o}{V_i} = \frac{-j\,R/\omega C}{R^2-\frac{1}{\omega^2C^2}-j\frac{3R}{\omega C}}$$

Rationalize the expression for β,

$$\therefore \quad \beta = \frac{-j\frac{R}{\omega C}\left[\left[R^2-\frac{1}{\omega^2C^2}\right]+j\frac{3R}{\omega C}\right]}{\left[\left(R^2-\frac{1}{\omega^2C^2}\right)-j\frac{3R}{\omega C}\right]\left[\left(R^2-\frac{1}{\omega^2C^2}\right)+j\frac{3R}{\omega C}\right]}$$

$$= \frac{\frac{3R^2}{\omega^2C^2} - \frac{jR}{\omega C}\left(R^2-\frac{1}{\omega^2C^2}\right)}{\left[\left(R^2-\frac{1}{\omega^2C^2}\right)^2+\frac{9R^2}{\omega^2C^2}\right]}$$

Use two stage amplifier which gives 360° phase shift hence accdg. to Barkhausen criteria, phase shift of β must be 0°.
Thus the imaginary part of β must be zero.

But ω can not be infinite for imaginary part zero, hence

$$R^2 - \frac{1}{\omega^2 C^2} = 0.$$

$$\omega^2 = \frac{1}{R^2C^2} \quad \text{i.e} \quad \omega = \frac{1}{RC}$$

$$\therefore \boxed{f = \frac{1}{2\pi RC}} \quad \dots \text{Frequency of oscillations.}$$

At f_0,

$$|\beta| = \left| \frac{\dfrac{3R^2}{\left(\frac{1}{R^2C^2} \times C^2\right)}}{R^2 - \dfrac{1}{\frac{1}{R^2C^2} \times C^2} + \dfrac{9R^2}{\frac{1}{R^2C^2} \times C^2}} \right|$$

$$|\beta| = \left| \frac{3R^4}{9R^4} \right| = \frac{1}{3}.$$

Let A be the effective gain of the amplifier, then

$$|A\beta| \geq 1 \quad \text{i.e.} \quad |A| \geq \frac{1}{|\beta|} \geq 3.$$

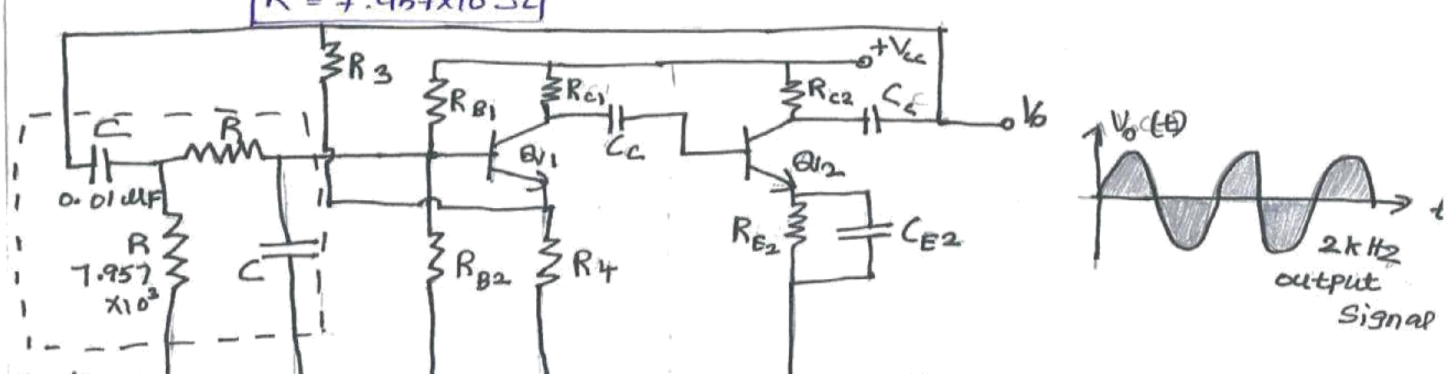
Here Gn : $f = 2\,kHz$

$$f = \frac{1}{2\pi RC} = 2 \times 10^3$$

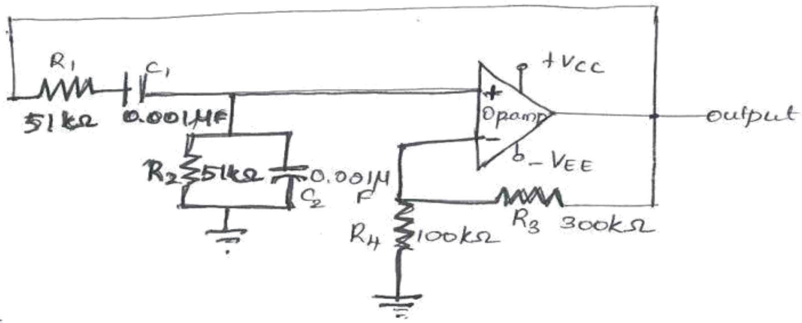
Choose $C = 0.01\,\mu F$ $\quad \frac{1}{2\pi \times R \times 0.01 \times 10^{-6}} = 2 \times 10^3.$

$$R = \frac{1}{2\pi \times 0.01 \times 10^{-6} \times 2 \times 10^3}$$

$$\boxed{R = 7.957 \times 10^3\,\Omega}$$



1. Calculate the resonant frequency of the Wien bridge oscillator of the fig. shown.



Soln:-

W.K.T. If $R_1 = R_2 = R$ & $C_1 = C_2 = C$, then the Oscillating/resonating frequency is $f = \frac{1}{2\pi RC} = \frac{1}{2\pi \times 51\times10^3 \times 0.001\times10^{-6}}$

$$\boxed{f_{o/r} = 3120.68\ Hz}$$

---

2. The frequency sensitive arms of the Wien bridge oscillator uses $C_1 = C_2 = 0.001\ \mu F$ & $R_1 = 10k\Omega$ while R_2 is kept variable. The frequency is to be varied from 10kHz to 50kHz by varying R_2. Find the minimum & Maximum values of R_2.

Soln- The frequency of the oscillator is given by

$$f = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}$$

for $f = 10kHz$.

$$10\times10^3 = \frac{1}{2\pi\sqrt{10\times10^3 \times R_2 \times (0.001\times10^{-6})^2}}$$

$\boxed{R_2 = 25.33\ k\Omega}$ for $f = 10kHz$

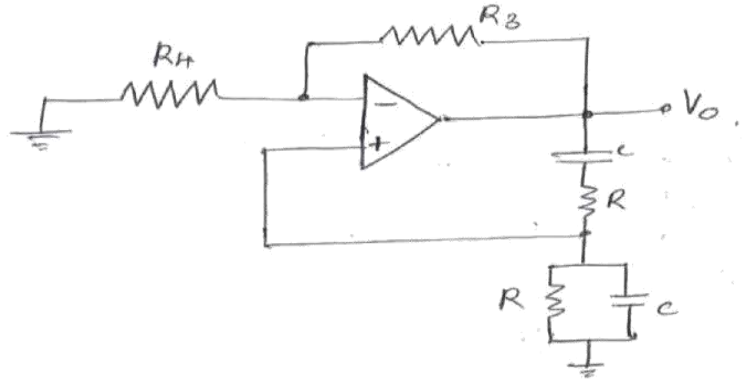
for $f = 50kHz$

$$50\times10^3 = \frac{1}{2\pi\sqrt{10\times10^3 \times R_2 \times (0.001\times10^{-6})^2}}$$

$\boxed{R_2 = 1.013\ k\Omega}$ for $f = 50kHz$.

Thus, the minimum value of R_2 is $1.013k\Omega$ & maximum value of R_2 is $25.33k\Omega$.

3. Determine whether the circuit shown below will work as an oscillator or not. If yes, determine the frequency of the oscillator.



$R = 5.1\,k\Omega$
$C = 0.001\,\mu F$
$R_3 = 6\,k\Omega$
$R_4 = 2\,k\Omega$.

Soln :-

The circuit is a Wein Bridge Oscillator using Op.amp.

The gain of the op.amp is $A = 1 + \frac{R_3}{R_4}$

$$A = 1 + \frac{6}{2} = 4$$

$A = 4$ Thus $A > 3$.

This satisfies the required oscillating condition. The feedback is given to the non-inverting terminal ensuring the zero phase shift. Hence the circuit will work as the oscillator.

$$f = \frac{1}{2\pi RC} = \frac{1}{2\pi \times 5.1 \times 10^3 \times 0.001 \times 10^{-6}}$$

$$\boxed{f = 31.2068\ kHz}$$

This is the frequency of oscillation.

Hartley Oscillator:

Transistorised:

Frequency of oscillation is gn. by $f = \dfrac{1}{2\pi\sqrt{C L_{eq}}}$

$L_{eq} = L_1 + L_2$ (or)
$L_{eq} = L_1 + L_2 + 2M$ where M - Mutual Inductance.

Value of h_{fe} required to satisfy the oscillating condns/- is

$h_{fe} = \dfrac{L_1}{L_2}$ $\quad$ $h_{fe} = \dfrac{L_1+M}{L_2+M}$ with mutual inductance M

FET Hartley oscillator:

If $L_1 = L_2 = L$ then $f = \dfrac{1}{2\pi\sqrt{2}\sqrt{LC}}$

Using OP. amp $f = \dfrac{1}{2\pi\sqrt{C L_{eq}}}$ where $L_{eq} = L_1 + L_2$ or $L_{eq} = L_1 + L_2 + 2M$

1. In a Hartley oscillator $L_1 = 15mH$ and $C = 50pF$. Calculate L_2 for a frequency of 168kHz. The mutual inductance btw. L_1 & L_2 is $5\mu H$. Also find the required gain of the transistor to be used for oscillations.

Sln/-: Gn: $L_1 = 15mH$, $C = 50pF$, $M = 5\mu H$, $f = 168\,KHz$

$f = \dfrac{1}{2\pi\sqrt{L_{eq}C}}$ where $L_{eq} = L_1 + L_2 + 2M$

$L_{eq} = ?$

$$168\times10^{3} = \frac{1}{2\pi\sqrt{L_{eq}\times 50\times10^{-12}}}$$

$L_{eq} = 17.95\ mH$

$17.95\times10^{-3} = 15\times10^{-3} + L_2 + 5\mu\times 2$

$L_2 = 2.945\ mH$

Gain h_{fe} = $\dfrac{L_1+M}{\ldots}$ = $\dfrac{15\times10^{-3}+5\times10^{-6}}{\ldots}$ = 5.08

2. A FET Hartley oscillator circuit uses coupled coils in the tank circuit, each with inductance of 0.1 mH and mutual inductance of 0.025 mH. The circuit uses a fixed capacitor of 100pF in series with a variable capacitor of 100 pF (trimmer):

(i) Calculate % change in frequency when direction of coupling between coils is reversed, trimmer capacitance set to zero.

(ii) Repeat calculations in part (i) when trimmer capacitance is changed from 0 to 100 pF, assume any one direction of coupling.

Sln: an: $L_1 = 0.1\,mH$, $L_2 = 0.1\,mH$, $M = 0.025\,mH$, $C = 100\,pF$

(i) Assume one particular coupling direction for which,

$$L_{eq} = L_1 + L_2 + 2M = 0.25\,mH$$

$$\therefore f = \frac{1}{2\pi\sqrt{L_{eq}C}} = \frac{1}{2\pi\sqrt{0.25\times10^{-3}\times100\times10^{-12}}}$$

$$\boxed{f = 1.00658\ MHz.}$$

Direction of coupling is reversed,

$$L_{eq} = L_1 + L_2 - 2M = 0.15\,mH.$$

$$f' = \frac{1}{2\pi\sqrt{L_{eq}C}} = \frac{1}{2\pi\sqrt{0.15\times10^{-3}\times100\times10^{-12}}}$$

$$\boxed{f' = 1.2994\ MHz}$$

$$\%\ \text{change in frequency} = \frac{f'-f}{f}\times100$$

$$= \frac{1.2994 - 1.00658}{1.00658}\times100$$

$$\boxed{\%\ \text{freq. change} = 29.09\%.}$$

Transistorized:

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}}$$

where $C_{eq} = \frac{C_1 C_2}{C_1 + C_2}$ $\omega = \frac{1}{\sqrt{LC_{eq}}}$

$$h_{fe} \; (A_v) \; \text{Gain} = \frac{C_2}{C_1}$$

1. In a Colpitts oscillator $C_1 = 0.001\,\mu F$, $C_2 = 0.01\,\mu F$ and $L = 10\,\mu H$, Find the frequency of oscillation, feedback factor and the voltage Gain [NOV/DEC 2004 - 8Marks]

Soln:

Gn: $C_1 = 0.001\,\mu F$, $C_2 = 0.01\,\mu F$, $L = 10\,\mu H$.

$$\therefore C_{eq} = \frac{C_1 C_2}{C_1 + C_2} \Rightarrow \frac{0.001\times10^{-6} \times 0.01\times10^{-6}}{(0.001 \times 0.01)\times10^{-6}}$$

$$C_{eq} = 9.09\times10^{-10} F.$$

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}} = \frac{1}{2\pi\sqrt{10\times10^{-6} \times 9.09\times10^{-10}}}$$

$$f = 1.6692\,MHz$$

Voltage Gain $A_V = \frac{C_2}{C_1} = \frac{0.01\times10^{-6}}{0.001\times10^{-6}} = 10$

$$A_V = 10$$

Feedback factor (β)

For oscillations, $A_v \beta = 1$ $\therefore \beta = \frac{1}{A_v}$

$$\beta = \frac{1}{10}$$

$$\beta = 0.1$$

Ans: Frequency of oscillation $\Rightarrow f = 1.6692\,MHz$

Voltage Gain $\Rightarrow A_V = 10$

Feedback Factor $\Rightarrow \beta = 0.1$

(ii) Let us assume direction of coupling such that,

$$L_{eq} = L_1 + L_2 + 2M = 0.25\,mH$$

C_t – Trim capacitor = 100 pF

$$C_{eq} = \frac{C \times C_t}{C + C_t} = 50\,pF$$

$$f = \frac{1}{2\pi\sqrt{C_{eq} L_{eq}}} = \frac{1}{2\pi\sqrt{0.25\times10^{-3}\times50\times10^{-12}}}$$

$$f = 1.4235\ MHz$$

Reversing the direction of coupling,

$$L_{eq} = L_1 + L_2 - 2M = 0.15\,mH$$

$$f' = \frac{1}{2\pi\sqrt{L_{eq} C_{eq}}} = \frac{1}{2\pi\sqrt{0.15\times10^{-3}\times50\times10^{-12}}}$$

$$f' = 1.83776\ MHz$$

$$\%\ \text{change} = \frac{f'-f}{f}\times100 = \frac{1.83776 - 1.4235}{1.4235}\times100$$

$$\%\ \text{change} = 29.101\ \%$$

2. In a colpitts oscillator, the values of the inductors & capacitors in the tank circuit are $L = 40mH$, $C_1 = 100pF$, and $C_2 = 500pF$. (i) Find the frequency of oscillation (ii) If the o/p vge. is 10V, find the feedback voltage. (iii) Find the minimum gain, if the frequency is changed by changing 'L' alone. (iv) Find the value of C_1 for a gain of 10. (v) Also, find the new frequency of oscillation.

[8 Marks – MAY'11, DEC'12 ; 10 Marks – MAY'07]

Soln:- Gn: $L = 40mH$ $C_1 = 100pF$ $C_2 = 500pF$.

(i) $f = \frac{1}{2\pi\sqrt{LC_{eq}}}$ where $C_{eq} = \frac{C_1C_2}{C_1+C_2} \Rightarrow \frac{100\times10^{-12}\times500\times10^{-12}}{100\times10^{-12}+500\times10^{-12}}$

$C_{eq} = 83.333\times10^{-12}F$

$$f = \frac{1}{2\pi\sqrt{40\times10^{-3}\times83.333\times10^{-12}}}$$

$$\boxed{f = 87.1727\,kHz}$$

(ii) Feedback voltage is a part of output voltage, where i/p voltage is not required for an oscillator.

$V_o = 10V$

Feedback Voltage $= \frac{V_o}{Gain}$ [ie. $Gain = \frac{V_o}{V_f}$]

$Gain = \frac{C_2}{C_1} = \frac{500}{100} = 5$

$V_f = \frac{V_o}{Gain} = \frac{10}{5} \Rightarrow 2V$

$$\boxed{V_f = 2V}$$

(iii) Minimum Gain $= \frac{C_2}{C_1} = 5$ Thus $\boxed{h_{fe(min)} = 5}$

(iv) $Gain = 10 = \frac{C_2}{C_1}$

$10 = \frac{500pF}{C_1} \Rightarrow \boxed{C_1 = 50pF}$

(v) New frequency of oscillation $f = \frac{1}{2\pi\sqrt{LC_{eq}}}$ $C_{eq} = \frac{C_1C_2}{C_1+C_2} = \frac{50p\times500p}{550p}$

$C_{eq} = 45.4545\,pF$

$f = \frac{1}{2\pi\sqrt{40\times10^{-3}\times45.4545\times10^{-12}}}$ $\boxed{f = 118.032\,kHz}$

3. A Colpitts oscillator is designed with $C_1 = 100pF$ and $C_2 = 7500pF$. The inductance is variable. Determine the range of inductance value if the frequency of oscillations is to vary between 950 KHz & 2050KHz [APR/MAY 2010 - Marks 8]

Soln:- Gen: $C_1 = 100pF$ $C_2 = 7500pF$ $f_1 = 950kHz$, $f_2 = 2050kHz$.

$$C_{eq} = \frac{C_1C_2}{C_1+C_2} = \frac{100\times10^{-12}\times7500\times10^{-12}}{(100\times10^{-12}+7500\times10^{-12})}$$

$$\boxed{C_{eq} = 98.6842\ pF}$$

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}}$$

$$f_1 = \frac{1}{2\pi\sqrt{L_1C_{eq}}} \quad \& \quad f_2 = \frac{1}{2\pi\sqrt{L_2C_{eq}}}$$

$$\therefore\ 950\times10^3 = \frac{1}{2\pi\sqrt{L_1\times98.6842\times10^{-12}}} \quad \& \quad 2050\times10^3 = \frac{1}{2\pi\sqrt{L_2\times98.6842\times10^{-12}}}$$

$$\boxed{L_1 = 284.41\mu H} \qquad \boxed{L_2 = 61.07\ \mu H.}$$

Range of inductance : $61.07\mu H$ to $284.41\ \mu H$.

Clapp Oscillator

Oscillator frequency $\boxed{f_0 = \frac{1}{2\pi\sqrt{LC_{eq}}}.}$ where $C_{eq} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$

Here C_{eq} is replaced by C_3 neglecting C_1 & C_2.

$$f = \frac{1}{2\pi\sqrt{LC_3}}$$

1. Calculate the frequency of oscillation for the Clapp oscillator with $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100\ pF$ and $L = 470\ \mu H$ [APR/MAY 07 Marks 6]

Gn: $C_1 = 0.1\mu F$ $C_2 = 1\mu F$, $C_3 = 100pF$ & $L = 470\ \mu H$

Soln:-
$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3} = \frac{1}{0.1\times10^{-6}} + \frac{1}{1\times10^{-6}} + \frac{1}{100\times10^{-12}}$$

$$C_{eq} = 99.89\ pF.$$

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}} = \frac{1}{2\pi\sqrt{470\times10^{-6}\times99.89\times10^{-12}}}$$

$$\boxed{f = 734.53kHz}$$

of oscillation in Clapp oscillator having $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100\mu F$ [APR/MAY 08, Marks 4]

Sln: Gn: $C_1 = 0.1\mu F$, $C_2 = 1\mu F$, $C_3 = 100\mu F$, $f = 734.5\,kHz$.

$$\frac{1}{C_{eq}} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

$$\frac{1}{C_{eq}} = \frac{1}{0.1\times10^{-6}} + \frac{1}{1\times10^{-6}} + \frac{1}{100\times10^{-6}}$$

$\therefore$ $C_{eq} = 9.0826\times10^{-8}\,F$

$$f = \frac{1}{2\pi\sqrt{LC_{eq}}} \Rightarrow 734.5\times10^3 = \frac{1}{2\pi\sqrt{L\times9.0826\times10^{-8}}}$$

$$\boxed{L = 0.5169\,\mu H}$$

Crystal oscillators:

1. A crysta has $L = 0.33\,H$, $C = 0.065\,pF$ and $C_M = 1\,pF$ with $R = 5.5\,k\Omega$. Find (i) Series resonant frequency (ii) Parallel resonant frequency (iii) By what percent does the parallel resonant frequency exceed the series resonant frequency? (iv) Find the Q factor of the crystal. NOV/DEC 2008 Marks 8.

Sln: Gn: $L = 0.33\,H$, $C = 0.065\,pF$, $C_M = 1\,pF$, $R = 5.5\,k\Omega$

(i)

$$\boxed{f_s = \frac{1}{2\pi\sqrt{LC}}} = \frac{1}{2\pi\sqrt{0.33\times0.065\times10^{-12}}}$$

$$\boxed{f_s = 1.087\,MHz}$$

(ii)

$$C_{eq} = \frac{C\,C_M}{C + C_M} = \frac{0.065\times1}{0.065+1} = 0.061\,pF$$

$$\boxed{f_p = \frac{1}{2\pi\sqrt{LC_{eq}}}} = \frac{1}{2\pi\sqrt{0.33\times0.061\times10^{-12}}}$$

$$\boxed{f_p = 1.121\,MHz}$$

(iii) % increase $= \dfrac{f_p - f_s}{f_s} = \dfrac{1.121\times10^6 - 1.087\times10^6}{1.087\times10^6} = \boxed{3.127\,\%}$

(iv) $Q = \dfrac{\omega_s L}{R} = \dfrac{2\pi f_s L}{R} = \dfrac{2\pi\times1.087\times10^6\times0.33}{5.5\times10^3} \Rightarrow \boxed{Q = 409.789}$

Tuned collector oscillator:

1. A tuned collector oscillator in a radio receiver has a fixed inductance of 60 μH & has to be tunable over the frequency band of 400 kHz to 1200 kHz. Find the range of variable capacitor to be used. [MAY/JUNE 2011, DEC 2011]

Soln: Gn: $L = 60\mu H$, $f_1 = 400 kHz$, $f_2 = 1200 kHz$.

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

$$400 \times 10^3 = \frac{1}{2\pi\sqrt{60\times10^{-6} \times C_1}} \quad ; \quad 1200\times10^3 = \frac{1}{2\pi\sqrt{60\times10^{-6}\times C_2}}$$

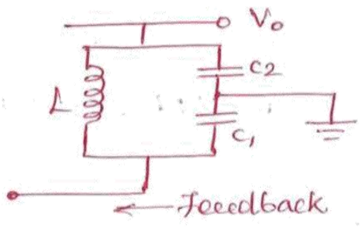
$$\therefore \boxed{C_1 = 2.638\, nF} \qquad \boxed{C_2 = 0.2931\, nF}$$

So the capacitor ranges between 0.2931 nF & 2.638 nF.

Other problems:

1. Draw the feedback circuit of a Colpitts oscillator. Obtain the value of the equivalent series capacitance required if it uses as L of 100 mH & is to oscillate at 40 kHz [2M, APR/MAY 13]

The feedback ckt. of Colpitts oscillator is shown below.



$$f = \frac{1}{2\pi\sqrt{LC_{eq}}}$$

$$40\times10^3 = \frac{1}{2\pi\sqrt{100\times10^{-3}\times C_{eq}}}$$

$$\therefore \boxed{C_{eq} = 158.314\, pF}$$

UNIT-II
OSCILLATORS

1. A crystal has the following parameters L=0.5H, C= 0.05 pF and mounting capacitances is 2pF. Calculate its series and parallel resonant frequencies. (Nov/Dec 2010)

Series resonance frequency fs=1007 KHz

Parallel resonant frequency fp=1019 KHz

2. In an RC phase shift oscillator, If its frequency of its oscillation is 955 Hz and R1=R2=R3= 680 KΩ. Find the value of capacitor.(Nov/Dec 2010)

$$F=\frac{1}{2\Pi RC\sqrt{6}}$$

Capacitor $C=1.0005*10^{-4}$

3. State the essential conditions for maintaining oscillation. (Nov/ Dec 2011) (May/ June 2012)

The total phase shift of an oscillator should be 360_{o}. For feedback oscillator it should satisfies Barhausen criterion.

4. A tuned collector oscillator in a radio receiver has a fixed inductance of 60 µH and has to be tunable over the frequency band of 400 kHz to 1200 kHz. Find the range of variable capacitor to be used. (Nov/Dec 2011) (May/ June 2012)

Fo=400KHz, C=2641pF

Fo=1200KHz, C=293pF

5. What is the major disadvantage of Twin T Oscillator?(Nov/Dec 2012)

- Higher amplitude distortion
- With amplitude stabilisation in the form of feedback diodes, oscillations from the oscillator can go on indefinitely.

- Some method of stabilizing the amplitude of the oscillations must be provided

  because if the voltage gain of the amplifier is too small the desired oscillation will decay and stop and if it is too large the output amplitude rises to the value of the

  supply rails, which saturates the op-amp and causes the output waveform to become distorted.

6. In a Hartley oscillator if L1= 0.2mH L2=0.3mH and c=0.003µF. Calculate the frequency of oscillations. (Nov/Dec 2012)

Fo=130mHz

7. Mention two essential conditions for a circuit to maintain oscillations.(April/ May 2010)

- The total phase shift of an oscillator should be 360^o.
- For feedback oscillator it should satisfies Barhausen criterion.i.e. $|A\beta|=1$ the magnitude of loop gain must be unity

8. In a RC phase shift oscillator, if R1=R2=R3= 200 KΩ and C1=C2=C3= 100 pF find the

frequency of the oscillator. (April/May 2010)

F0= 3.248KHz

9. A Weinbridge oscillator is used for operations at 9 kHz. If the value of the resistance R is

100 kohm, what i s the value of C required? (April/May 2011)

$F=\frac{1}{2\pi RC\sqrt{6}}$

C=176pF

10. **Define Barkhausen criterion for oscillators (May/June 2014)**

The product βA_V is greater than one this is called barkhausen criterion

$$A_{vf} = \frac{Av}{(1-\beta Av)}$$

$$A_{vf} = \frac{1}{0} = \text{infinite}$$

$$(1 - \beta Av) < 0$$

βAv >1 this is the condition for feedback Oscillator.

An Oscillator which follows Barkhausen criterion is called the Feedback Oscillator.

10. Differentiate oscillator and amplifier (Nov/ Dec 2013)

Oscillator	Amplifier
A circuit with an active device is used to produce an alternating current is called an oscillator circuit	An amplifier is a device which produces a large electrical output of similar characteristics to that of the input parameters.
Generate periodic signal without a.c input	Provide electrical signal with input

11. State the Barkhausen criterion for sustained oscillation. What will happen to the oscillations if magnitude of the loop gain is greater than unity? (Nov/ Dec 2013)

The product βA_V is greater than one this is called barkhausen criterion

$$A_{vf} = \frac{Av}{(1-\beta Av)} \qquad \frac{1}{0}$$

$$A_{vf} = \frac{}{(1 - \beta A_v)} = \text{infinite}$$

$\beta A_v > 1$ this is the condition for feedback Oscillator.

An Oscillator which follows Barkhausen criterion is called the Feedback Oscillator.

The loop gain is greater than unity so that the amplitude of oscillation will continue to increase without limit.

14. What are the classifications ofOscillators? *Based on wave generated:

i. Sinusoidal Oscillator,

ii. Non-sinusoidal Oscillator or Relaxation Oscillator

Ex: Square wave, Triangular wave, Rectangular wave etc.

*According to principle involved:

i. Negative resistance Oscillator,

ii. Feedback Oscillator.

*According to frequency generated:

i. Audio frequency oscillator 20 Hz – 20 kHz

ii. Radio frequency Oscillator 30 kHz – 30 MHz

iii. Ultrahigh frequency Oscillator 30 MHz – 3 GHz

iv. Microwave Oscillator 3 GHz – above.

* Crystal Oscillators

15. What are the types of feedback oscillators?

* RC-Phase shift Oscillator,

* LC-Oscillators

i. Tuned collector Oscillator

ii. Tuned emitter Oscillator

iii. Tuned collector base Oscillator

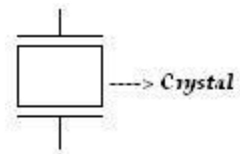
iv. Hartley Oscillator

v. Colpits Oscillator

vi. Clap Oscillator

16. Define Piezoelectric effect.

When applying mechanical energy to some type of crystals called piezoelectric crystals the mechanical energy is converted into electrical energy is called piezoelectric effect.



F= 1/T

17. What is Miller crystal oscillator? Explain its operation.

It is nothing but a Hartley oscillator its feedback Network is replaced by a crystal. Crystal normally generate higher frequency reactance due to the miller capacitance are in effect between the transistor terminal.

18. State the frequency for RC phase shift oscillator.

The frequency of oscillation of RC-phase shift oscillator is

$$F=\frac{1}{2\Pi RC\sqrt{6}}$$

Where k=2.639.

19. What happens to the circuit above and below resonance?

Above resonance the circuit acts as capacitive and below resonance the circuit acts as inductive

<u>16Marks:</u>

1.(i). Drive the general condition for oscillation for a LC oscillator and derive the frequency of oscillation for the colpitts oscillator? (10)
(ii). How is a clap oscillator modified from a colpitts oscillator? (6) (Nov/Dec 2010)

2.What is a wien bridge? How is it used to work as oscillator? Support it with necessary derivations. (16) (Nov/Dec 2010)

3.(i). Draw the circuit of wien bridge oscillator using BJT. Show that the gain of the amplifier must be atleast 3 for oscillation to occur. (10)

(ii). In a certain oscillator circuit the gain of the amplifier circuit is $\frac{-16*10^6}{jw}$ **and the feedback factor of the feedback network is** $\frac{10^8}{[2*10^8+jw]^2}$ **verify the Barkhausen criterion for the sustained oscillations. Also find the frequency at which circuit will oscillate. (6) (Nov/Dec 2011)**

4. (i). With the help of neat circuit diagram. Explain the following oscillators. Also explain how the frequency is found in each case.

(1). Clapp Oscillator (6)
(2). Miller Crystal Oscillator (6)

(ii). The equivalent circuit of a crystal has the values of L=0.7 H, C=0.01pF, R=1000Ω and Cm= 2pF, Calculate series resonant frequency, Parallel resonant frequency and Quality factor of the crystal (4) (Nov/Dec 2011)

5. (i) Explain Armstrong oscillator and derive its frequency of oscillation. (8)

(ii) A Colpitts oscillator is designed with C1 = 100 pF and C2 = 7500 pF. The inductance is variable. Determine the range of inductance values, if the frequency of oscillation is to vary between 950 KHz and 2050 KHz. (8) (April/May 2010)

6.(i) Explain Wien bridge oscillator and derive its frequency of oscillation. (10)

(ii) Write a note on frequency stability of oscillators. (6) (April/May 2010)

7.(i) Explain the working of a Colpitts oscillator with a neat circuit diagram and derive the frequency of oscillation. (8)

(ii) In a Colpitt's oscillator, the value of the inductor and capacitors in the tank circuit are L = 40mH, C1 = 100 pF and C2 = 500 pF. (8)

Find the frequency of oscillation. If the output voltage is 10 V, find the feedback voltage at the input side of the amplifier. Find the minimum gain, if the frequency is changed by charging `L' alone. Find the value of C1 for a gain of 10 if C2 i s kept constant as 500 pF. Also find the resulting new frequency. (April/ May 2011)

8.(i) Sketch the basic block diagram of an oscillator and explain how it works. If the gain of the amplifier is A and the feedback factor is fi, sketch the output waveforms for the three cases (1) jAfij > 1, (2) jAfij = 1 and (3) jAfij < 1. Derive the conditions of sustained oscillations. (10)

(ii) Make a table of comparison of RC phase shift oscillator and Wien-bridge oscillator bringing out the similarities and differences. (6) (April/ May 2011)

3

Tuned Amplifiers

3.1 Introduction

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

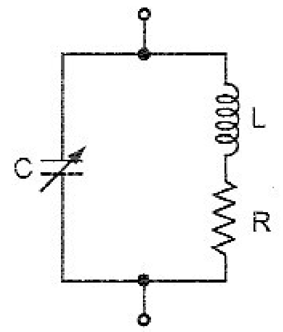


Fig. 3.1 Tuned circuit

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

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

and $$Z_r = \frac{L}{CR} \qquad \dots (2)$$

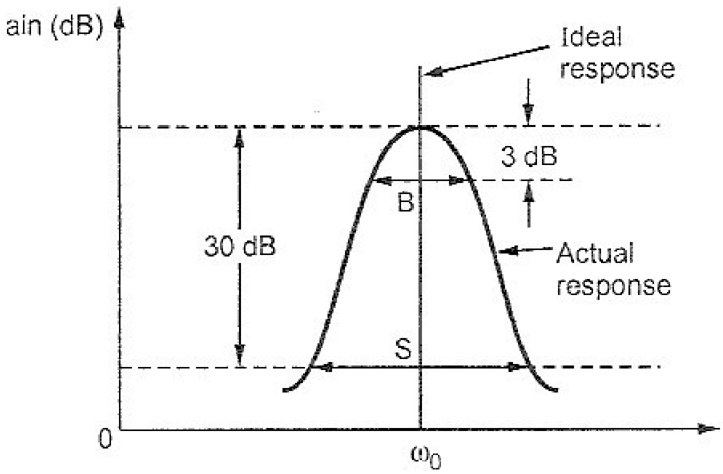


Fig. 3.2 Frequency response of a tuned amplifier

The response of tuned amplifiers is maximum at resonant frequency and it falls sharply for frequencies below and above the resonant frequency, as shown in the Fig. 3.2.

As shown in the Fig. 3.2, 3 dB bandwidth is denoted as B and 30 dB bandwidth is denoted as S. The ratio of the 30 dB bandwidth (S) to the 3 dB bandwidth (B) is known as **skirt selectivity.**

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

3.1.1 Coil Losses

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

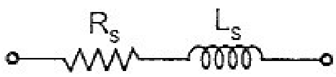
Fig. 3.3 Inductor with leakage resistance

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

3.1.2 Q Factor

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

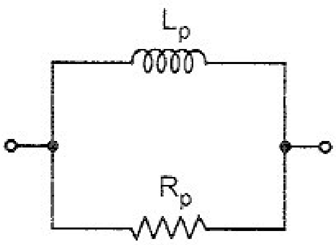
Series circuit



Inductive impedance

$$\frac{\omega L_s}{R_s}$$

Parallel circuit



Inductive admittance

$$\frac{R_p}{\omega L_p}$$

Fig. 3.4 Quality factor equations

Quality factor equation $Q = \frac{1}{D} = \frac{\omega L_s}{R_s} = \frac{R_p}{\omega L_p}$

3.1.3 Unloaded and Loaded Q

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

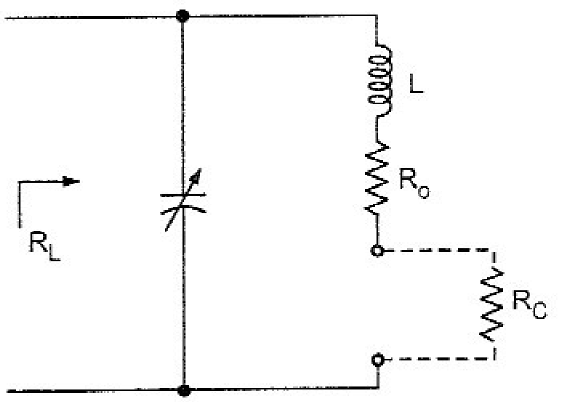


Fig. 3.5 Tuned load circuit

Let us consider the tuned load circuit as shown in the Fig. 3.5. Here, L and C represents tank circuit. The internal circuit losses of inductor are represented by R_o and R_C represents the coupled in load. For this circuit, we can write

$$R_o = \frac{\omega_o L}{Q_U} \text{ and } R_C = \frac{\omega_o L}{Q_L}$$

where Q_U is unloaded Q and Q_L is loaded Q.

The circuit efficiency for the above tank circuit is given as,

$$\eta = \frac{I^2 R_C}{I^2 (R_C + R_o)} = \frac{Q_U}{Q_U + Q_L} \times 100 \%$$

From above equation it can be easily realized that for high overall power efficiency, the coupled-in load R_C should be large in comparison to the internal circuit losses represented by R_o of the inductor.

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

$$BW = \frac{f_r}{Q_L}$$

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

If Q is large, bandwidth is small and circuit will be highly selective. For small Q values bandwidth is high and selectivity of the circuit is lost, as shown in the Fig. 3.6.

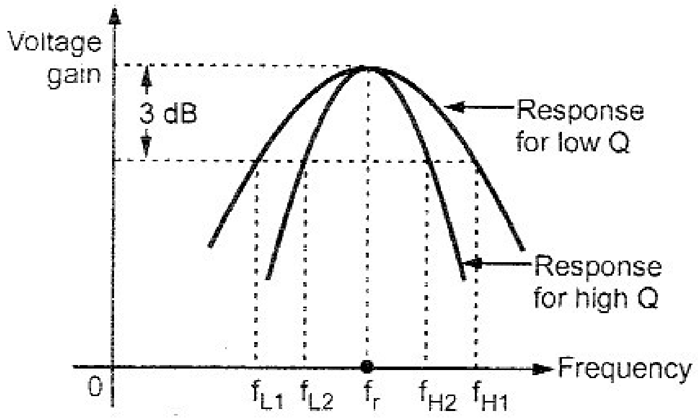


Fig. 3.6 Variation of 3dB bandwidth with variation in quality factor

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

3.1.4 Requirements of Tuned Amplifier

The basic requirements of tuned amplifiers are :

- The amplifier should provide selectivity of resonant frequency over a very narrow band.
- The signal should be amplified equally well at all frequencies in the selected narrow band.
- The tuned circuit should be so mounted that it can be easily tuned. If there are more than one circuit to be tuned, there should be an arrangement to tune all circuit simultaneously.
- The amplifier must provide the simplicity in tuning of the amplifier components to the desired frequency over a considerable range or band of frequencies.

3.1.5 Classification of Tuned Amplifier

We know that, multistage amplifiers are used to obtain large overall gain. The cascaded stages of multistage tuned amplifiers can be categorized as given below :

- Single tuned amplifiers
- Double tuned amplifiers
- Stagger tuned amplifiers.

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

- Capacitive coupled
- Inductive coupled
- Transformer coupled.

3.2 Small Signal Tuned Amplifier

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

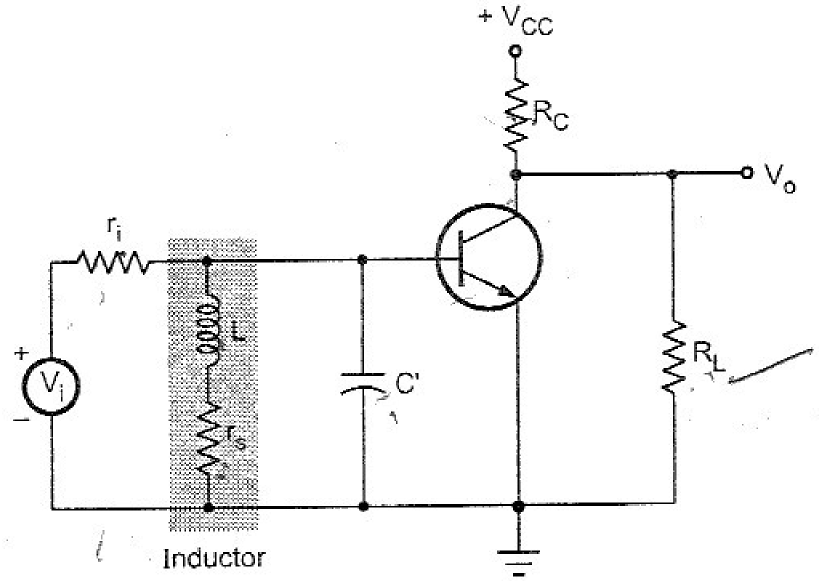


Fig. 3.7 Single tuned transistor amplifier

Before going to study the analysis of this amplifier we see the several practical assumptions to simplify the analysis.

Assumptions :

1. $R_L << R_C$
2. $r_{bb'} = 0$

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

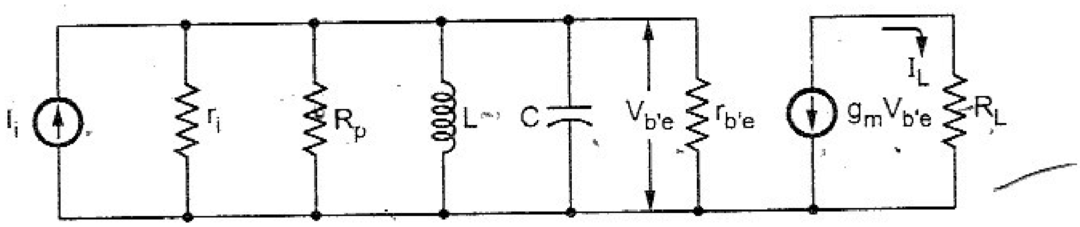


Fig. 3.8 Equivalent circuit of single tuned amplifier

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

C' : External capacitance used to tune the circuit

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

r_s : Represents the losses in coil

The series RL circuit in Fig. 3.7 is replaced by the equivalent RL circuit in Fig. 3.8 assuming coil losses are low over the frequency band of interest, i.e., the coil Q high.

$$\therefore \quad \boxed{Q_c \equiv \frac{\omega L}{r_c} >> 1} \qquad \ldots (1)$$

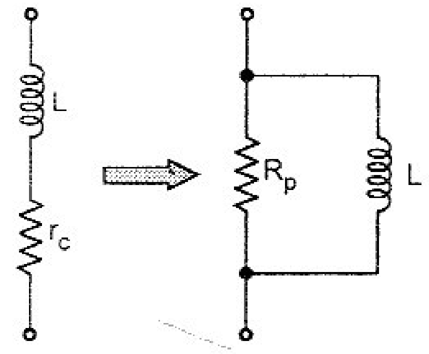


Fig. 3.9 Equivalent circuits

The conditions for equivalence are most easily established by equating the admittances of the two circuits shown in Fig. 3.9.

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

$$\approx \frac{r_c}{r_c^2 + \omega^2 L^2} - \frac{j\omega L}{r_c^2 + \omega^2 L^2}$$

$\because \omega L >> r_c$ from equation (1)

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

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

$\therefore$ Therefore, equating Y_1 and Y_2 we get,

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

$$\therefore \quad \frac{1}{R_p} = \frac{r_c}{\omega^2 L^2}$$

$$\boxed{\frac{1}{Q_c} = \frac{r_c}{\omega L}} \Rightarrow Q_c = \frac{\omega L}{r_c}$$

$$= \frac{r_c^2}{r_c\, \omega^2 L^2} = \frac{1}{r_c Q_c^2}$$

$$\therefore \quad \boxed{R_p = r_c Q_c^2 = \omega L Q_c} \qquad \because \omega L = Q_c r_c \text{ from equation (1)} \qquad \ldots (2)$$

Looking at Fig. 3.8 we have,

$$\therefore \quad \boxed{R = r_i \parallel R_p \parallel r_{b'e}} \qquad \ldots (3)$$

The current gain of the amplifier is then

$$A_i = \frac{-g_m R}{1 + j(\omega RC - R/\omega L)} = \frac{-g_m R}{1 + j\,\omega_o RC(\omega/\omega_o - \omega_o/\omega)} \qquad \ldots (4)$$

where $\quad \omega_o^2 = \dfrac{1}{LC}$

We define the Q of the tuned circuit at the resonant frequency ω_o to be

$$Q_i = \frac{R}{\omega_o L} = \omega_o RC \quad \text{...(5)}$$

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

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

$$\therefore \quad \boxed{A_{i\,(max)} = -g_m R} \quad \text{...(6)}$$

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

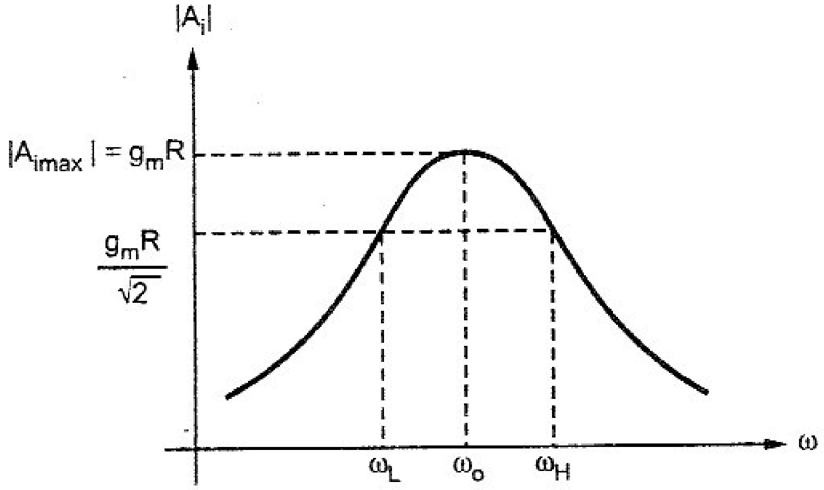


Fig. 3.10 Gain versus frequency for single tuned amplifier

At 3 dB frequency,

$$|A_i| = \frac{g_m R}{\sqrt{2}} \quad \text{... (7)}$$

$\therefore$ At 3 dB frequency

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

$$\therefore \quad 1 + Q_i^2\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)^2 = 2 \quad \text{... (8)}$$

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

$$BW = f_H - f_L = \frac{\omega_o}{2\pi Q_i} = \frac{1}{2\pi RC} \quad \text{... (9)}$$

$$\therefore \quad \boxed{BW = \frac{1}{2\pi RC}}$$

Example 3.1 : *Design a single tuned amplifier for following specifications :*

1. Centre frequency = 500 kHz

2. Bandwidth = 10 kHz

Assume transistor parameters : g_m = 0.04 S, h_{fe} = 100, $C_{b'c}$ = 1000 pF and $C_{b'c}$ = 100 pF. The bias network and the input resistance are adjusted so that r_i = 4 kΩ and R_L = 510 Ω.

Solution : From equation (9) we have,

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

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

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

From equation (3) we have,

$$R = r_i \parallel R_p \parallel r_{b'e}$$

where

$$r_i = 4\text{ k}\Omega$$

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

$$R_p = Q_c\,\omega_o L = \frac{Q_c}{\omega_o C}$$

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

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

$$\therefore \quad C = \frac{1}{2\pi \times 10 \times 10^3 \times \left[4 \times 10^3 \parallel 2500 \parallel \dfrac{Q_c}{2\pi \times 500 \times 10^3 \times C}\right]}$$

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

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

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

Solving for C we get,

$$C = 0.02\ \mu F$$

We have,

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

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

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

$$\therefore \quad C' = \mathbf{0.01686\ \mu F}$$

We have,

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

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

$$= \mathbf{5\ \mu H}$$

From equation (2) we have,

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

$$= \mathbf{1570\ \Omega}$$

$$\therefore \quad R = r_i \parallel R_p \parallel r_{b'e}$$

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

$$= 777\ \Omega$$

We have mid frequency gain as,

$$A_{i\,max} = -g_m R = (-0.04)(777) = \mathbf{-31}$$

3.3 Single Tuned FET Amplifier

The Fig. 3.11 shows the single tuned FET amplifier.

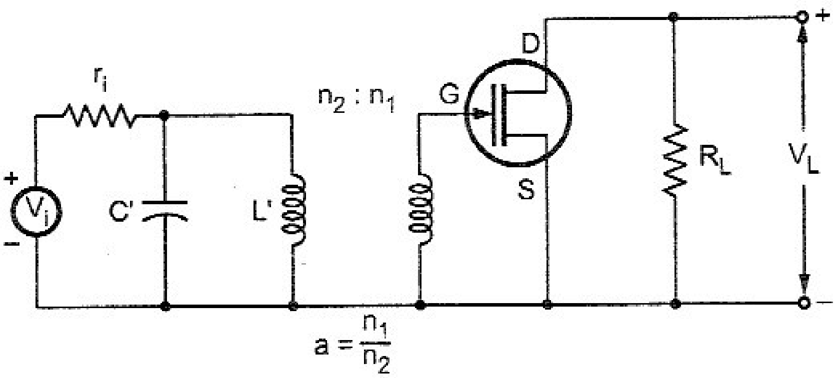


Fig. 3.11 Single tuned FET amplifier

The equivalent circuit for the given amplifier is as shown in the Fig. 3.12.

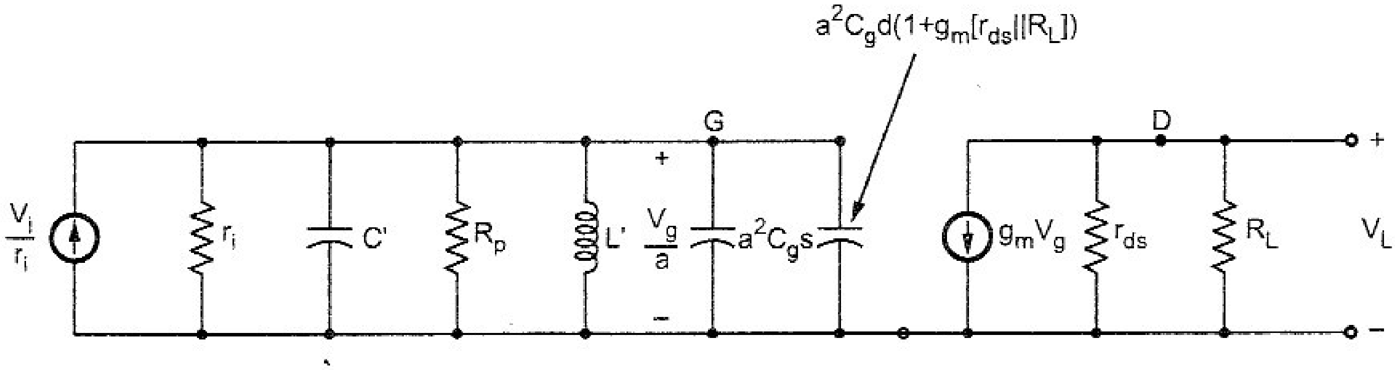


Fig. 3.12 Equivalent circuit of single tuned FET amplifier

The voltage gain is given by,

$$A_v = -a\, g_m\, (r_{ds} \,\|\, R_L)\, [(r_i \,\|\, R_p)/\, r_i] \qquad \dots (1)$$

where

$$C_i = a^2\, \{C_{gs} + C_{gd}\, [1 + g_m\, (r_{ds} \,\|\, R_L)]\} \qquad \dots (2)$$

$$Q_i = \omega_o\, (r_i \,\|\, R_p)\, (C' + C_i) \qquad \dots (3)$$

$$\omega_o^2 = \frac{1}{L\,(C' + C_i)} \qquad \dots (4)$$

At centre frequency, i.e., at $\omega = \omega_o$ gain is

$$A_{v\,max} = -a\, g_m\, (r_{ds} \,\|\, R_L)\, \frac{R_p}{r_i + R_p} \qquad \dots (5)$$

The 3 dB bandwidth is given by,

$$BW = \frac{1}{2\pi\,(r_i \,\|\, R_p)\,(C' + C_i)} \qquad \dots (6)$$

3.4 Single Tuned Capacitive Coupled Amplifier

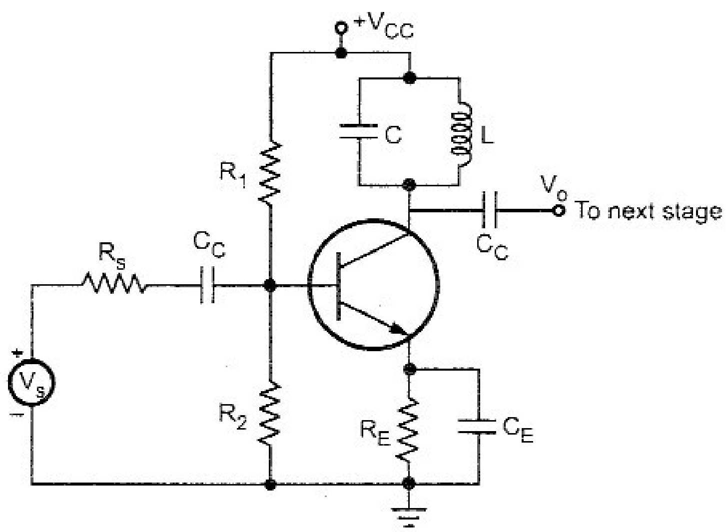


Fig. 3.13 Single tuned capacitive coupled transistor amplifier

Single tuned multistage amplifier circuit uses one parallel tuned circuit as a load in each stage with tuned circuits in all stages tuned to the same frequency. Fig. 3.13 shows a typical single tuned amplifier in CE configuration.

As shown in Fig. 3.13 tuned circuit formed by L and C acts as collector load and resonates at frequency of operation. Resistors R_1, R_2 and R_E along with capacitor C_E provides self bias for the circuit.

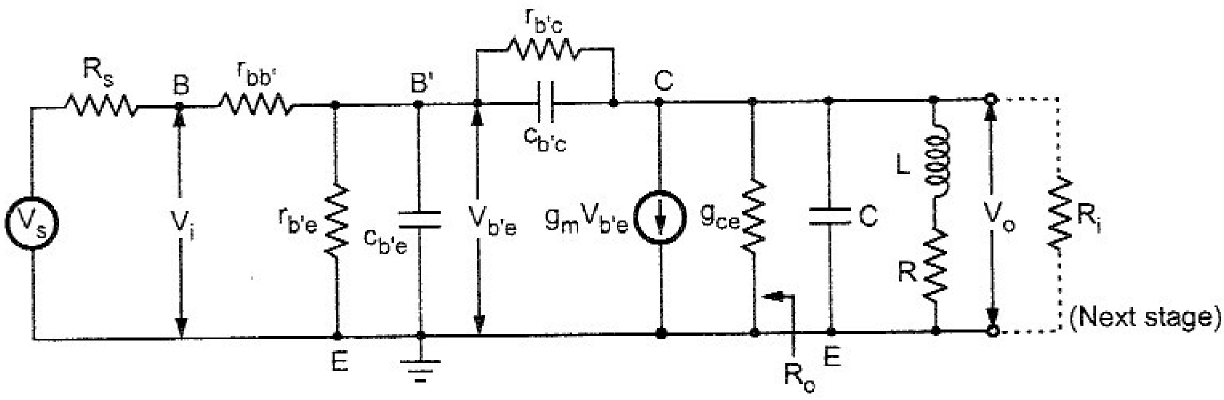


Fig. 3.14 Equivalent circuit of single tuned amplifier

The Fig. 3.14 shows the equivalent circuit for 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.

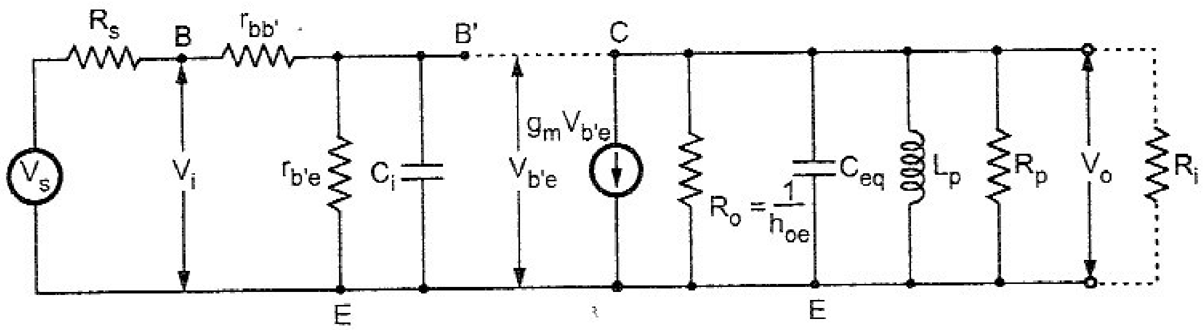


Fig. 3.15 Simplified equivalent circuit for single tuned amplifier

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

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

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

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

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

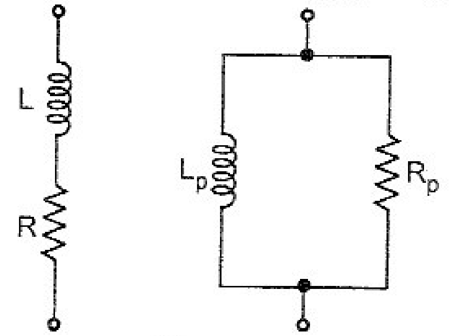


Fig. 3.16

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

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

Centre frequency

The centre frequency or resonant frequency is given as,

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

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

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

$$= C_o + C$$

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,

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

where ω_r is the centre frequency or resonant frequency.

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

The Q of the coil is usually large so that $\omega L >> R$ in the frequency range of operation.

From equation (4) we have,

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

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

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,

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

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

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.

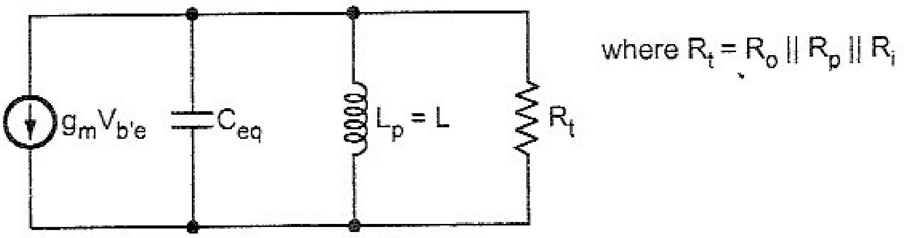


Fig. 3.17 Simplified output circuit for single tuned amplifier

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

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

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,

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

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

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

Example 3.2 : *A single tuned RF amplifier uses a transistor with an output resistance of 50 K, output capacitance of 15 pF and input resistance of next stage is 20 kΩ. The tuned circuit consists of 47 pF capacitance in parallel with series combination of 1 μH inductance and 2 Ω resistance. Calculate*

i) Resonant frequency

ii) Effective quality factor

iii) Bandwidth of the circuit

Solution : i) Resonant frequency f_r is given as,

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

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

$$= \mathbf{20.2\ MHz}$$

ii) Effective quality factor is given as,

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

$$= 2\pi f_r C_{eq} \times (R_o || R_p || R_i)$$

where $$R_p = \frac{\omega_r^2 L^2}{R} = \frac{(2\pi \times 20.2 \times 10^6)^2 (1 \times 10^{-6})^2}{2}$$

$$= 8054\ \Omega$$

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

$$= \mathbf{40.52}$$

iii) Bandwidth of the circuit is given as,

$$BW = \frac{f_r}{Q_{eff}} = \frac{20.2 \times 10^6}{40.52}$$

$$= \mathbf{498.5\ kHz}$$

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

Solution : Given : f_r = 600 kHz

BW = 15 kHz

R_t = 20 kΩ

C_o = 50 pF

$$\therefore \quad C_{eq} = (50\text{ pF} + C)$$

$$Q_{eff} = \frac{f_r}{BW} = \frac{600\text{ kHz}}{15\text{ kHz}}$$

$$= 40$$

i) We know that,

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

$$\therefore \quad C_{eq} = \frac{Q_{eff}}{\omega_r R_t} = \frac{40}{2\pi \times 600 \times 10^3 \times 20 \times 10^3}$$

$$= 530.5\text{ pF}$$

$$C_{eq} = (50\text{ pF} + C)$$

$$\therefore \quad C = 530.5\text{ pF} - 50\text{ pF}$$

$$= \mathbf{480.5\ pF}$$

ii) We know that,

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

$$\therefore \quad L = \frac{1}{(2\pi f_r)^2 C_{eq}} = \frac{1}{(2\pi \times 600 \times 10^3)^2 \times 530.5 \times 10^{-12}}$$

$$= \mathbf{132.6\ \mu H}$$

3.5 Double Tuned Amplifier

Fig. 3.18 shows double tuned RF amplifier in CE configuration. Here, voltage developed across tuned circuit is coupled inductively to another tuned circuit. Both tuned circuits are tuned to the same frequency.

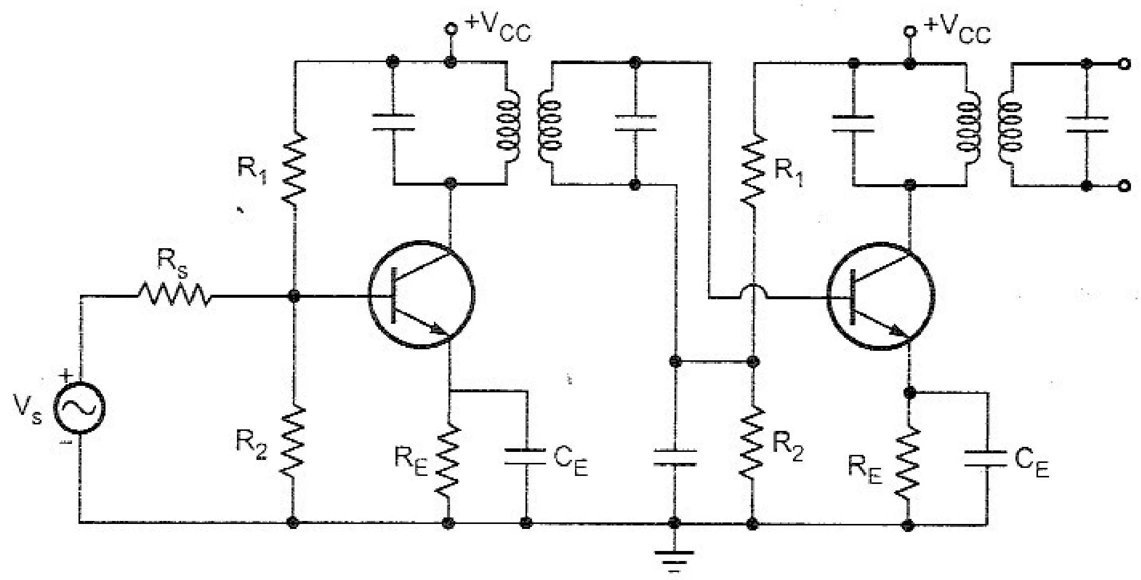


Fig. 3.18 Double tuned amplifier

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

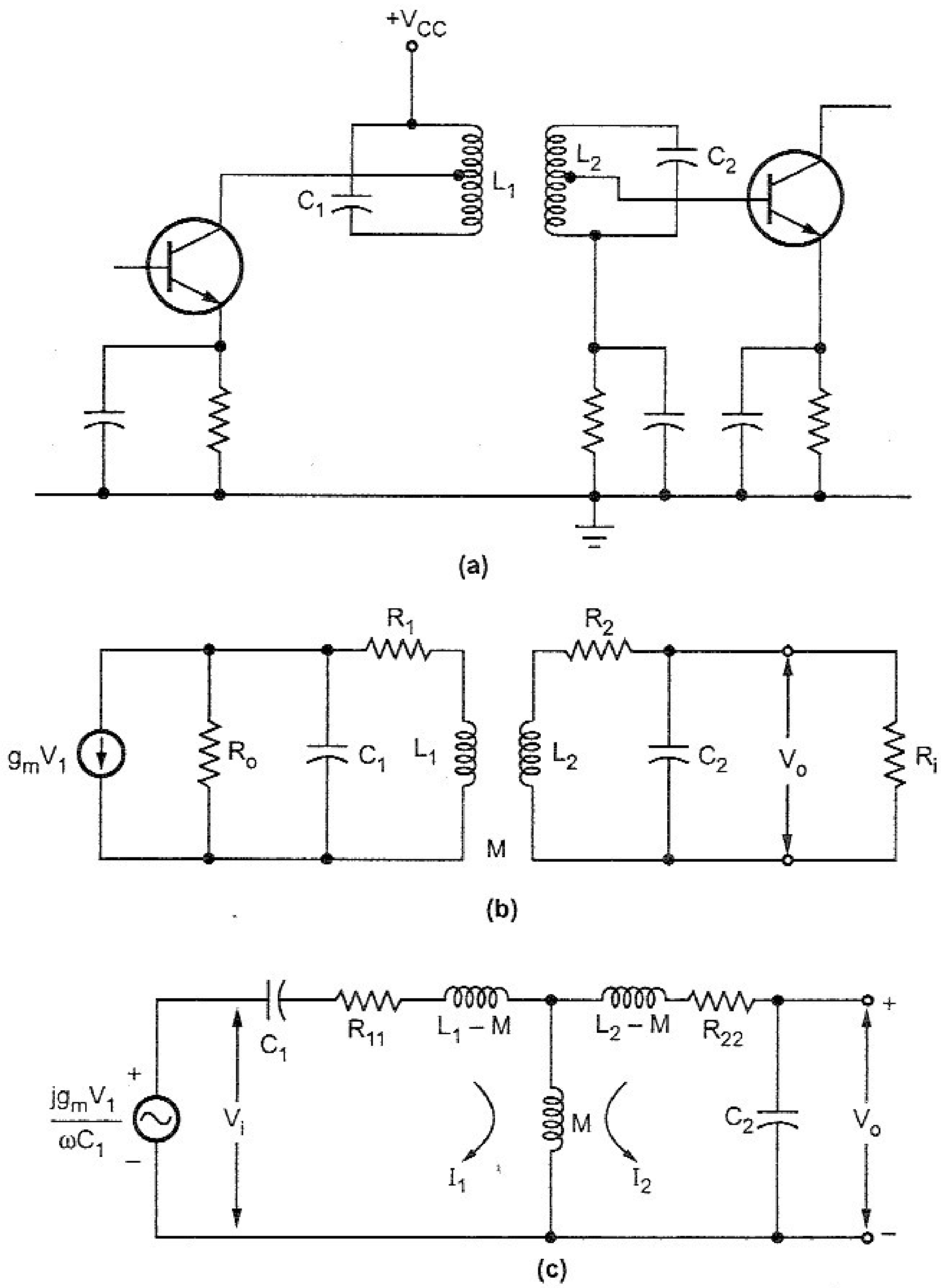
Analysis

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

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

Usually, the Q factors for both circuits are kept same. Therefore, $Q_1 = Q_2 = Q$ and the resonant frequency $\omega_r^2 = 1/L_1 C_1 = 1/L_2 C_2$.

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

$$V_o = -\frac{j}{\omega_r C_2} I_2 \quad ... (2)$$

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,

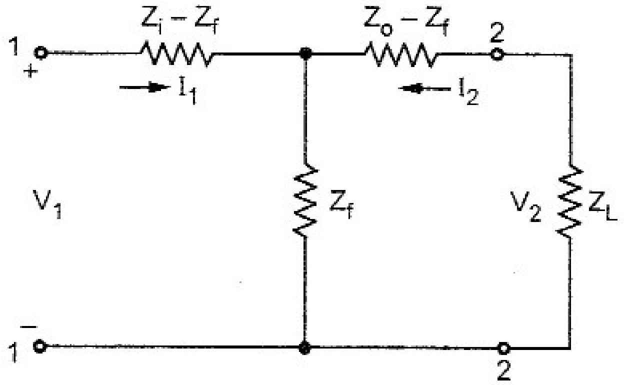


Fig. 3.20

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

$$= \frac{Z_f}{Z_f^2 - Z_i (Z_o + Z_L)}$$

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

$$Y_T = \frac{Z_f}{Z_f^2 - Z_i\,(Z_o + Z_L)} = \frac{1}{Z_f - Z_i\,(Z_o + Z_L)/Z_f}$$

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

$$Y_T = \frac{kQ^2}{\omega_r\,\sqrt{L_1\,L_2}\left[4\,Q\delta - j\,(1+k^2Q^2 - 4\,Q^2\,\delta^2)\right]} \qquad \dots (3)$$

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

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

$$\because\ V_i = \frac{j\,g_m\,V_1}{\omega C_1}$$

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

$$\because\ \frac{1}{\omega_r\,C} = \omega_r\,L$$

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

Taking the magnitude of equation (4) we have,

$$|A_v| = g_m\,\omega_r\,\sqrt{L_1\,L_2}\;Q\,\frac{kQ}{\sqrt{1+k^2Q^2 - 4\,Q^2\,\delta^2 + 16\,Q^2\,\delta^2}} \qquad \dots (5)$$

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.

$$4Q\delta - j\,(1+k^2Q^2 - 4Q^2\delta^2) = 0 \qquad \dots (6)$$

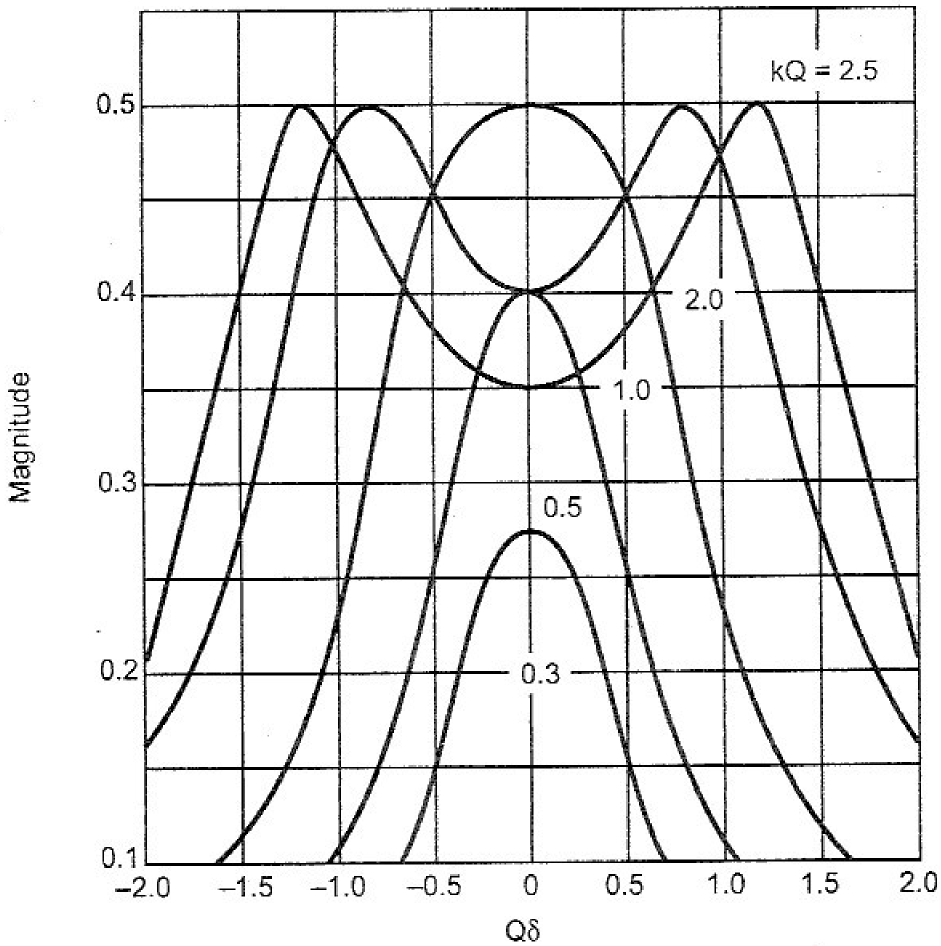


Fig. 3.21

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

$$f_1 = f_r\left(1-\frac{1}{2Q}\sqrt{k^2Q^2-1}\right) \text{ and}$$

$$f_2 = f_r\left(1+\frac{1}{2Q}\sqrt{k^2Q^2-1}\right) \quad \ldots (7)$$

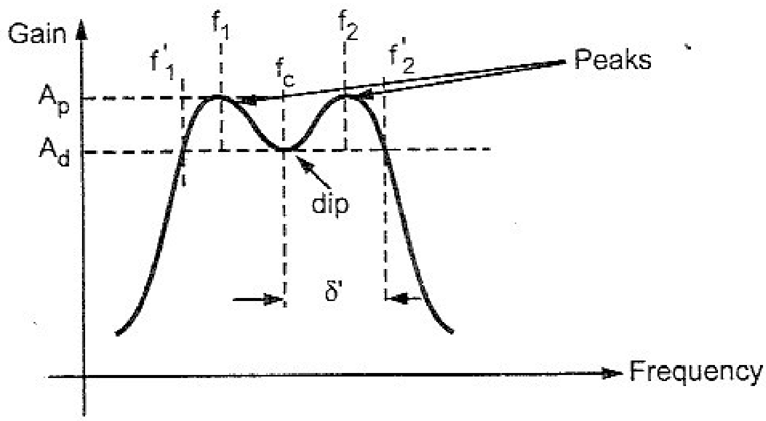


Fig. 3.22

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.

$$\gamma = \left|\frac{A_p}{A_d}\right| = \frac{1+k^2Q^2}{2\,kQ} \qquad \text{... (10)}$$

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,

$$BW = 2\,\delta' = \sqrt{2}\,(f_2 - f_1) \qquad \text{... (12)}$$

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.

3.6 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

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

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

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

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

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

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

Substituting for δ, the fractional frequency variation, i.e.

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

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

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

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

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

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

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

The bandwidth of n stage identical amplifier is given as,

$$\begin{aligned} BW_n &= f_2 - f_1 = (f_2 - f_r) + (f_r - f_1) \\ &= \frac{f_r}{2\,Q_{eff}}\sqrt{2^{\frac{1}{n}}-1} + \frac{f_r}{2\,Q_{eff}}\sqrt{2^{1/n}-1} \\ &= \frac{f_r}{Q_{eff}}\sqrt{2^{\frac{1}{n}}-1} \\ &= BW_1\sqrt{2^{\frac{1}{n}}-1} \qquad \ldots (1) \end{aligned}$$

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

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

Solution : i) We know that,

$$BW_n = BW_1\sqrt{2^{\frac{1}{n}}-1} = 20\times10^3\times\sqrt{2^{\frac{1}{3}}-1}$$

$$= \mathbf{10.196\ kHz}$$

ii) $$BW_n = 20\times10^3\times\sqrt{2^{\frac{1}{4}}-1} = \mathbf{8.7\ kHz}$$

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

3.7 Effect of Cascading Double Tuned Amplifiers on Bandwidth

When a number of identical double tuned amplifier stages are connected in cascade, the overall bandwidth of the system is thereby narrowed and the steepness of the sides of the response is increased, just as when single tuned stages are cascaded. The quantitative relation between the 3 dB bandwidth of n identical double tuned critically coupled stages compared with the bandwidth Δ_2 of such a system can be shown to be 3 dB bandwidth for

$$\text{n identical stages double tuned amplifiers} = \Delta_2 \times \left(2^{\frac{1}{n}} - 1\right)^{\frac{1}{4}} \quad \ldots (1)$$

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

Key Point: *The equation (1) assumes that the bandwidth Δ_2 is small compared with the resonant frequency.*

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

Solution : We know that for double tuned cascaded stages,

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

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

$$= \mathbf{14.28\ kHz}$$

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

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

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

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

For $n = 3$

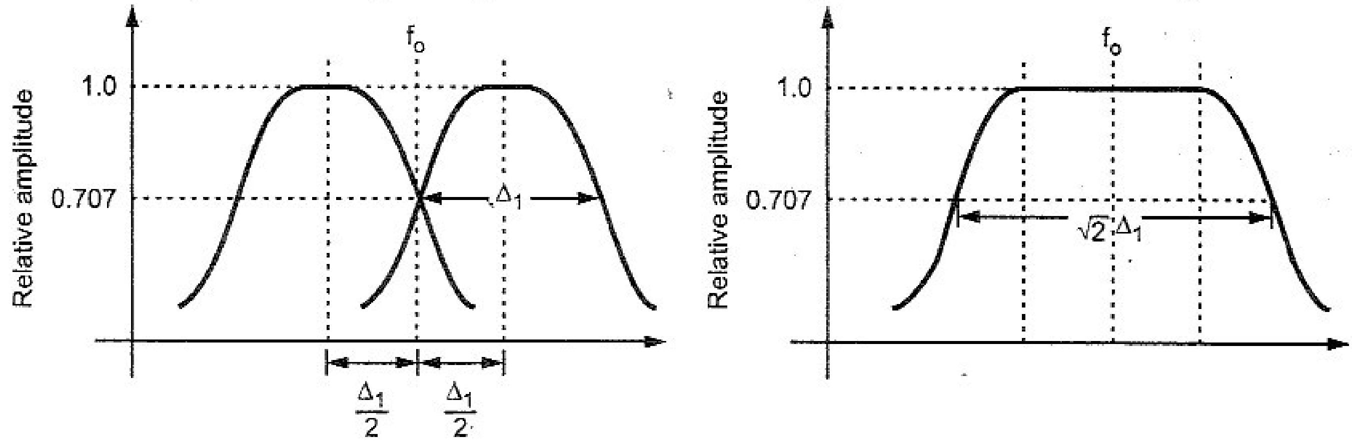
$$BW_n = \frac{BW_n}{(2^{1/3} - 1)^{1/4}}$$

$$= \frac{20 \times 10^3}{(2^{1/3} - 1)^{1/4}}$$

$$= \mathbf{28.01\ kHz}$$

3.8 Staggered Tuned Amplifier

We have seen that double tuned amplifier gives greater 3 dB bandwidth having steeper sides and flat top. But alignment of double tuned amplifier is difficult. To overcome this problem two single tuned cascaded amplifiers having certain bandwidth are taken and their resonant frequencies are so adjusted that they are separated by an amount equal to the bandwidth of each stage. Since the resonant frequencies are displaced or staggered, they are known as staggered tuned amplifiers. The advantage of staggered tuned amplifier is to have a better flat, wideband characteristics in contrast with a very sharp, rejective, narrow band characteristic of synchronously tuned circuits (tuned to same resonant frequencies). Fig. 3.23 shows the relation of amplification characteristics of individual stages in a staggered pair to the overall amplification of the two stages.



(a) Response of individual stages (b) Overall response of staggered pair

Fig. 3.23

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,

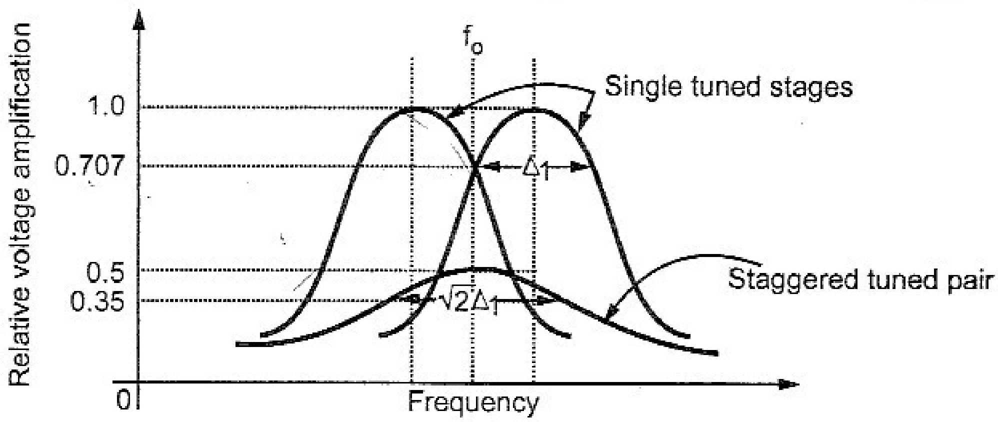


Fig. 3.24 Response of individually tuned and staggered tuned pair

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

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.

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

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

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

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

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

Substituting the value of X we get,

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

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

3.9 Large Signal Tuned Amplifiers

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

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

3.9.1 Class B Tuned Amplifier

The Fig. 3.25 shows the class B tuned amplifier. It works with a single transistor by sending half sinusoidal current pulses to the load. The transistor is biased at the edge of the conduction. Eventhough 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

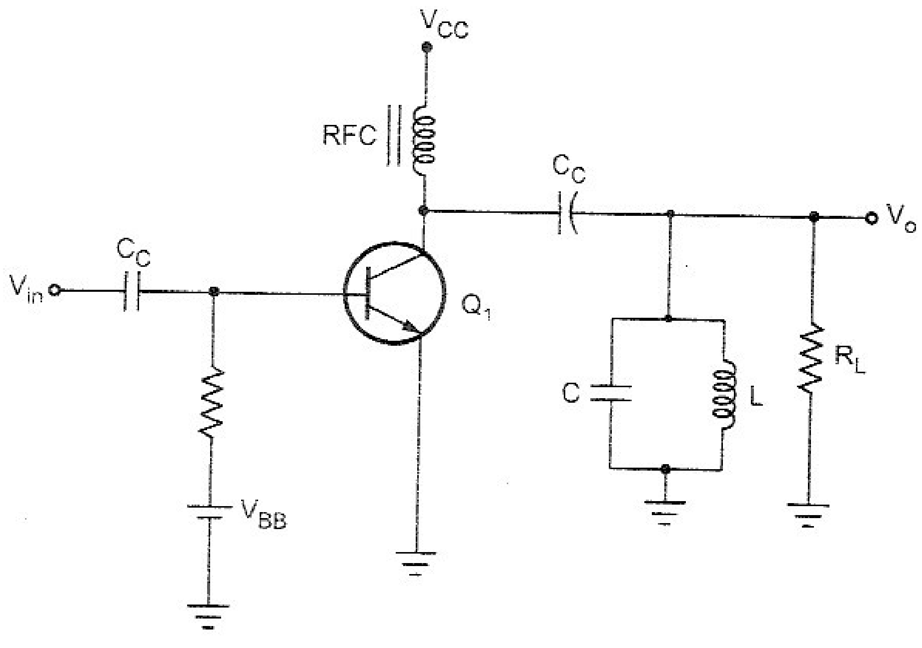


Fig. 3.25 Class B tuned amplifier

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.

3.9.2 Class C Tuned Amplifier

The amplifier is said to be class C amplifier, if the Q point and the input signal are selected such 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. Hence 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.

The current and voltage waveforms for a class C amplifier operation are shown in the Fig. 3.26.

Looking at Fig. 3.26, it is apparent that the total angle during which current flows is less than 180°. This angle is called the conduction angle, θ_c.

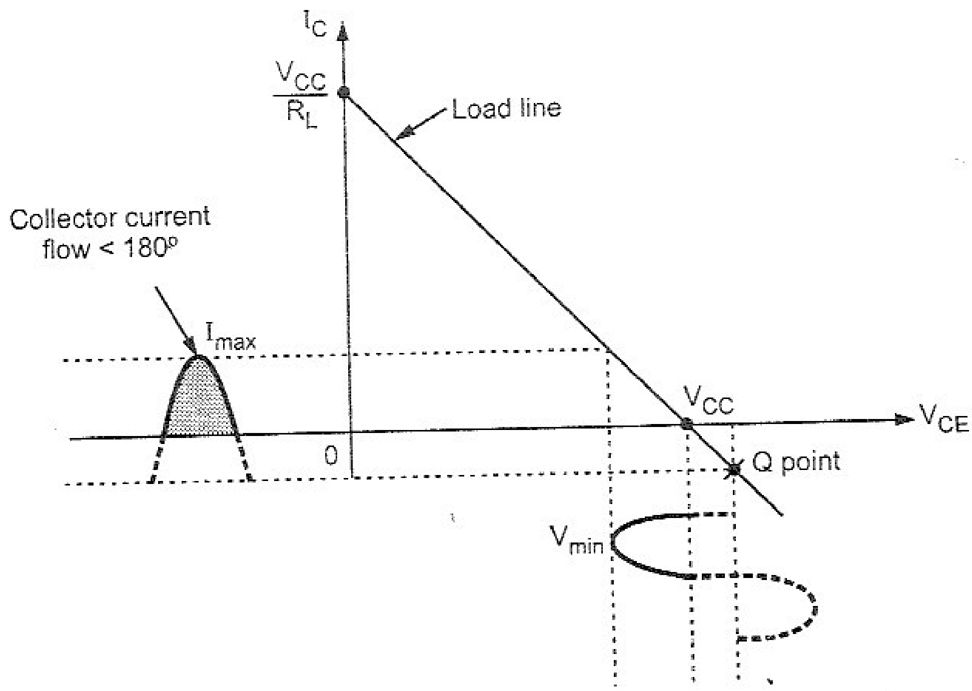


Fig. 3.26 Waveform representing class C operation

Fig. 3.27 shows the class C tuned amplifier. Here a parallel resonant circuit acts as a 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 a load impedance is tuned to the input frequency. Therefore, it filters the harmonic frequencies and produce a sine wave output voltage consisting of fundamental component of the input signal.

UNIT IV

WAVE SHAPING AND MULTIVIBRATOR CIRCUITS

Linear wave shaping :Process by which the shape of a non sinusoidal signal is changed by passing the signal through the network consisting of linear elements Diodes can be used in wave shaping circuits.

- ✓ Either limit or clip signal portion--- clipper
- ✓ shift the dc voltage level of the signal --- clampers

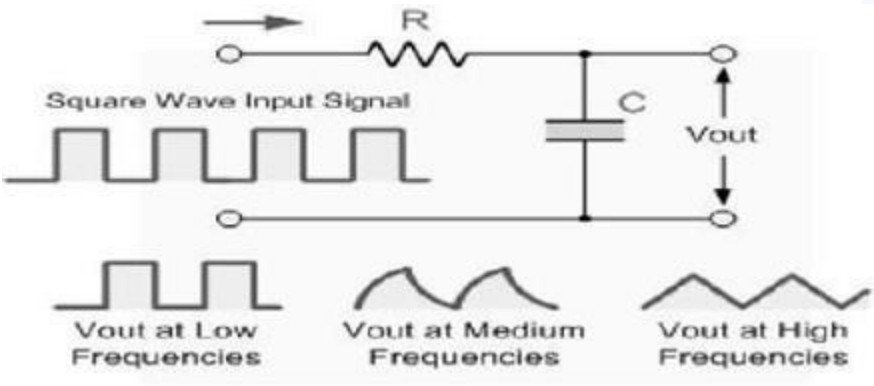
Types of non sinusoidal input

- ✓ Step
- ✓ pulse
- ✓ square
- ✓ Ramp input

➢ Integrator

The Integrator is basically a low pass filter circuit operating in the time domain that converts a square wave "step" response input signal into a triangular shaped waveform output as the capacitor charges and discharges.

A Triangular waveform consists of alternate but equal, positive and negative ramps. As seen below, if the RC time constant is long compared to the time period of the input waveform the resultant output waveform will be triangular in shape and the higher the input frequency the lower will be the output amplitude compared to that of the input.



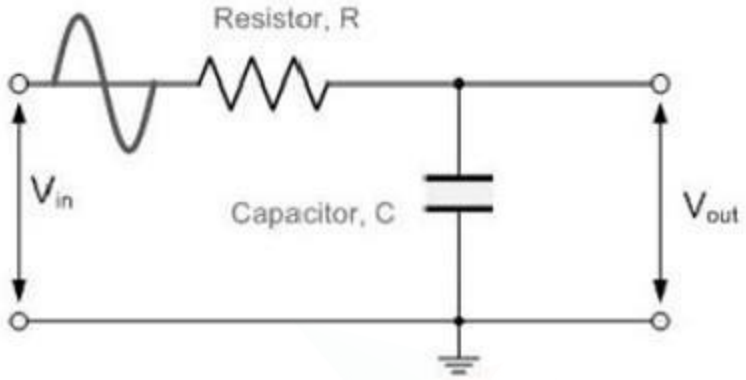
This makes this type of circuit ideal for converting one type of electronic signal to another for use in wave-generating or wave-shaping circuits.

The Low Pass Filter

A simple passive Low Pass Filter or LPF, can be easily made by connecting together in series a single Resistor with a single Capacitor as shown below. In this type of filter arrangement the input signal (Vin) is applied to the series combination (both the Resistor and Capacitor together) but the output signal (Vout) is taken across the capacitor only.

This type of filter is known generally as a "first-order filter" or "one-pole filter", why first-order or single-pole, because it has only "one" reactive component in the circuit, the capacitor.

Low Pass Filter Circuit



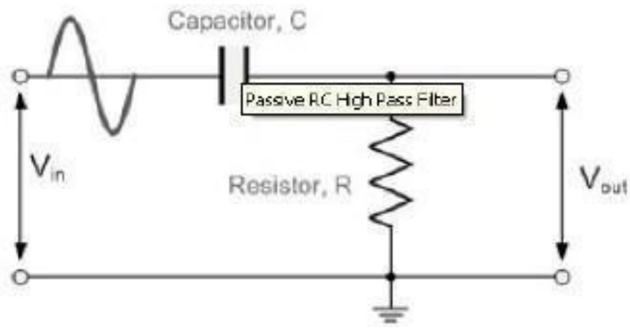
The reactance of a capacitor varies inversely with frequency, while the value of the resistor remains constant as the frequency changes. At low frequencies the capacitive reactance, (Xc) of the capacitor will be very large compared to the resistive value of the resistor, R and as a result the voltage across the capacitor, Vc will also be large while the voltage drop across the resistor, Vr will be much lower. At high frequencies the reverse is true with Vc being small and Vr being large.

High Pass Filters

A High Pass Filter or HPF, is the exact opposite to that of the Low Pass filter circuit, as now the two components have been interchanged with the output signal (Vout) being taken from across the resistor as shown.

Where the low pass filter only allowed signals to pass below its cut-off frequency point, fc. The passive high pass filter circuit as its name implies, only passes signals above the selected cut-off point, fc e l i m i n a t i n g a n y l o w frequency signals from the waveform. Consider the circuit below.

The High Pass Filter Circuit



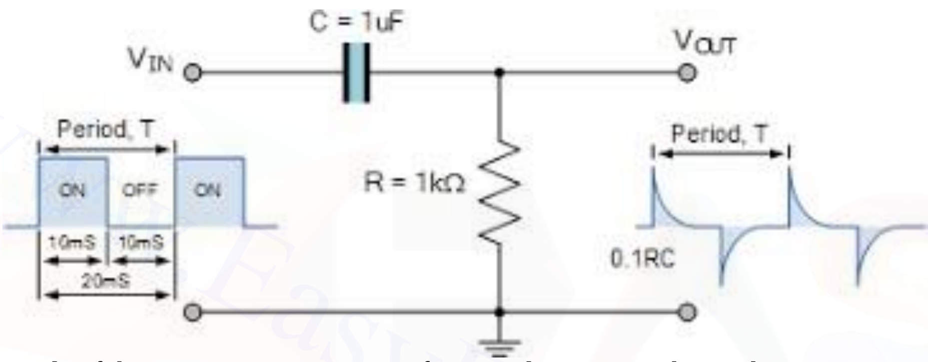
In this circuit arrangement, the reactance of the capacitor is very high at low frequencies so the capacitor acts like an open circuit and blocks any input signals at Vin until the cut-off frequency point (fc) is reached.

Above this cut-off frequency point the reactance of the capacitor has reduced sufficiently as to now act more like a short circuit allowing all of the input signal to pass directly to the output as shown below in the High Pass Frequency Response Curve.

RC Differentiator

Up until now the input waveform to the filter has been assumed to be sinusoidal or that of a sine wave consisting of a fundamental signal and some harmonics operating in the frequency domain giving us a frequency domain response for the filter.

However, if we feed the High Pass Filter with a Square Wave signal operating in the time domain giving an impulse or step response input, the output waveform will consist of short duration pulse or spikes as shown.



Each cycle of the square wave input waveform produces two spikes at the output, one positive and one negative and whose amplitude is equal to that of the input. The rate of decay of the spikes depends upon the time constant, (RC) value of both components, (t = R x C) and the value of the input frequency. The output pulses resemble more and more the shape of the input signal as the frequency increases

Therefore, the value of the voltage drop across the series resistor at that first instant must be 0 volts because there is no current flow through it. As time passes, current begins to flow through the circuit and voltage develops across the resistor. Since the circuit has a long time constant, the voltage across the resistor does NOT respond to the rapid changes in voltage of the input square wave. Therefore, the conditions for integration in an RL circuit are a long time constant with the output taken across the resistor.

There are a variety of diode network called clippers that have the ability to-―clip‖ off a portion of the input signal without distorting the remaining part of the alternating waveform. The half wave rectifier is an example of the simplest form of diode clipper one resistor and diode.

Depending on the orientation of the diode, the positive or negative region of the input signal is-―clipped‖ off. There are two general categories of clippers: series and parallel. The series configuration is defined as one where the diode is in series with the load, while the parallel variety has the diode in a branch parallel to the load.

Multivibrators

The type of circuit most often used to generate square or rectangular waves is the multivibrator. A multivibrator, is basically two amplifier circuits arranged with regenerative feedback. One of the amplifiers is conducting while the other is cut off When an input signal to one amplifier is large enough, the transistor can be driven into cutoff, and its collector voltage will be almost V CC. However, when the transistor is driven into saturation, its collector voltage will be about 0 volts.

A circuit that is designed to go quickly from cutoff to saturation will produce a square or rectangular wave at its output. This principle is used in multivibrators. Multivibrators are classified according to the number of steady (stable) states of the circuit. A steady state exists when circuit operation is essentially constant; that is, one transistor remains in conduction and the other remains cut off until an external signal is applied.

The three types of multivibrators :

- ✓ ASTABLE
- ✓ MONOSTABLE
- ✓ BISTABLE.

The astable circuit has no stable state. With no external signal applied, the transistors alternately switch from cutoff to saturation at a frequency determined by the RC time constants of the coupling circuits.

The monostable circuit has one stable state; one transistor conducts while the other is cut off. A signal must be applied to change this condition. After a period of time, determined by the internal RC components, the circuit will return to its original condition where it remains until the next signal arrives.

The bistable multivibrator has two stable states. It remains in one of the stable states until a trigger is applied. It then FLIPS to the other stable condition and remains there until another trigger is applied. The multivibrator then changes back (FLOPS) to its first stable state.

Astable Multivibrator

A multivibrator which generates square waves of its own (i.e. without any external trigger pulse) is known as astable multivibrator. It is also called free ramming multivibrator. It has no stable state but only two quasi-stables (half-stable) makes oscillating continuously between these states. Thus it is just an oscillator since it requires no external pulse for its operation of course it does require D.C power.

In such circuit neither of the two transistors reaches a stable state. It switches back and forth from one state to the other, remaining in each state for a time determined by circuit constants. In other words, at first one transistor conducts (i.e. ON state) and the other stays in the OFF state for some time. After this period of time, the second transistor is automatically turned ON and the first transistor turned OFF. Thus the multivibrator will generate a square wave of its own. The width of the square wave and it frequency will depend upon the circuit constants.

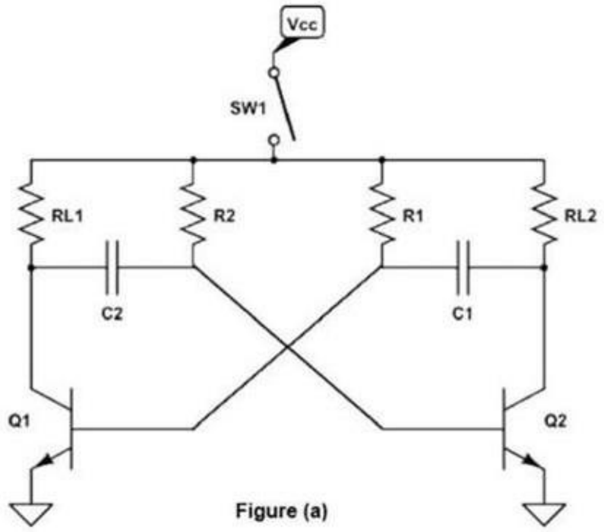


Figure (a)

Here we like to describe.

- ✓ Collector - coupled Astabe multivibrator
- ✓ Emitter - coupled Astable multivibrator

Figure (a) shows the circuit of a collector coupled astable multivibrator using two identical NPN transistors Q_1 and Q_2. It is possible to have $R_{L1} = R_{L2} = R_L = R_1 = R_2 = R$ and $C_1 = C_2 = C$. In that case , the circuit is known as symmetrical astable multivibrator. The transistor Q_1 is forward biased by the V_{cc} supply through resistor R_2. Similarly the transistor Q_2 is forward biased by the V_{cc} supply through resistor R_1. The output of transistor Q_1 is coupled to the input of transistor Q_2 through the capacitor C_2. Similarly the output of transistor Q_2 is coupled to the input of transistor Q_1 through the capacitor C_1.

It consists of two common emitter amplifying stages. Each stage provides a feedback through a capacitor at the input of the other. Since the amplifying stage introduces a 180^o phase shift and another 180^o phase shift is introduced by a capacitor, therefore the feedback signal and the circuit works as an oscillator. In other words because of capacitive coupling none of the transistor can remain permanently out-off or saturated, instead of circuit has two quasi-stable states (ON and OFF) and it makes periodic transition between these two states.

The output of an Astable multivibrator is available at the collector terminal of the either transistors as shown in figure (a). However, the two outputs are 180^o out of phase with each other. Therefore one of the outputs is said to be the complement of the other.

Let us suppose that

When $Q_1$is ON, Q_2 is OFF and

When Q_2 is ON, Q_1 is OFF.

When the D.C power supply is switched ON by closing S, one of the transistors will start conducting before the other (or slightly faster than the other). It is so because characteristics of no two similar transistors can be exactly alike suppose that Q_1 starts conducting before Q_2 does. The feedback system is such that Q_1 will be very rapidly driven ton saturation and Q_2 to cut-off. The circuit operation may be explained as follows.

Since Q_1 is in saturation whole of V_{CC} drops across R_{L1}. Hence $V_{C1} = 0$ and point A is at zero or ground potential. Since Q_2 is in cut-off i.e. it conducts no current, there is no drop across R_{L2}. Hence point B is at V_{CC}. Since A is at 0V C_2 starts to charge through R_2 towards V_{CC}.

When voltage across C_2 rises sufficiently (i.e. more than 0.7V), it biases Q_2 in the forward direction so that it starts conducting and is soon driven to saturation.V_{CC} decreases and becomes almost zero when Q_2 gets saturated. The potential of point B decreases from V_{CC} to almost 0V.

This potential decrease (negative swing) is applied to the base of Q_1 through C_1. Consequently, Q_1 is pulled out of saturation and is soon driven to cut-off.

Since, now point B is at 0V, C_1 starts charging through R_1 towards the target voltage V_{CC}.

When voltage of C1 increases sufficiently. Q_1 becomes forward-biased and starts conducting. In this way the whole cycle is repeated.

It is observed that the circuit alternates between a state in which Q_1 is ON and Q_2 is OFF and the state in which Q_1 is OFF and Q_2 is ON. This time in each state depends on RC values. Since each transistor is driven alternately into saturation and cut-off. The voltage waveform at either collector (points A and B in figure (b)) is essentially a square waveform with peak amplitude equal to V_{CC}.

Calculation of switching times and frequency of oscillations:

The frequency of oscillations can be calculated by charging and discharging capacitances and its base resistance R_B.

The voltage across the capacitor can be written as

$$V_o = V_f - (V_f - V_i)e^{\frac{-t}{RC}} = V_B$$

V_i= intial voltage = V_B =-V_{CC} thus the transistors enters from ON to OFF state

V_f = final voltage = V_B = -V_{CC} then the resistor enters from OFF to ON state

T_1 is ON & T_2 is OFF the above equation can be written as

$V_{B1}=V_{CC}$

substitute at $t=T_1$, $V_{B1}=0$ hence this equation becomes

$T_1=.69R_{B2}C_2$

The total time period $T=.694(R_{B1}C_1+R_{B2}C_2)$

When $R_{B!}=R_{B2}=R$ & $C_1=C_2=C$

$T=1.39RC$

Frequency of free running multivibrator is given by

$$f = \frac{1}{T} = \frac{1}{1.38RC}$$

the frequency stability of the circuit is not good as only the function of the product of RC but also depends on load resistances, supply voltages and circuit parameters. In order to stabilize the frequency, synchronizing signals are injected which terminate the unstable periods earlier than would occur naturally.

Bistable multivibrator

The **Bistable Multivibrator** is another type of two state device similar to the

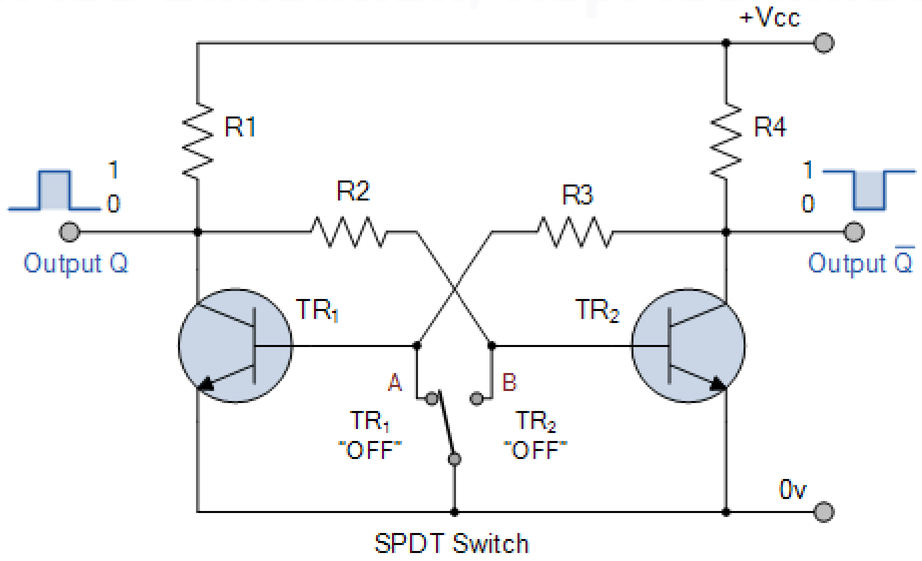
Monostable Multivibrator we looked at in the previous tutorial but the difference this time is that BOTH states are stable.

Bistable Multivibrators have **TWO** stable states (hence the name: "Bi" meaning two) and maintain a given output state indefinitely unless an external trigger is applied forcing it to change state.The bistable multivibrator can be switched over from one stable state to the other by the application of an external trigger pulse thus, it requires two external trigger pulses before it returns back to its original state. As bistable multivibrators have two stable states they are more commonly known as Latches and Flip-flops for use in sequential type circuits.

The discrete **Bistable Multivibrator** is a two state non-regenerative device constructed from two cross-coupled transistors operating as "ON-OFF" transistor switches. In each of the two states, one of the transistors is cut-off while the other transistor is in saturation, this means that the bistable circuit is capable of remaining indefinitely in either stable state.

To change the bistable over from one state to the other, the bistable circuit requires a suitable trigger pulse and to go through a full cycle, two triggering pulses, one for each stage are required. Its more common name or term of "flip-flop" relates to the actual operation of the device, as it "flips" into one logic state, remains there and then changes or "flops" back into its first original state. Consider the circuit below.

Bistable Multivibrator Circuit



The **Bistable Multivibrator** circuit above is stable in both states, either with one transistor “OFF” and the other “ON” or with the first transistor “ON” and the second “OFF”. Lets suppose that the switch is in the left position, position “A”.

The base of transistor TR_1 will be grounded and in its cut-off region producing an output at Q. That would mean that transistor TR_2 is “ON” as its base is connected to Vcc through the series combination of resistors R1 and R2. As transistor TR_2 is “ON” there will be zero output at Q, the opposite or inverse of Q.

If the switch is now move to the right, position “B”, transistor TR_2 will switch “OFF” and transistor TR_1 will switch “ON” through the combination of resistors R3 and R4 resulting in an output at Q and zero output at Q the reverse of above.

Then we can say that one stable state exists when transistor TR_1 is “ON” and TR_2 is “OFF”, switch position “A”, and another stable state exists when transistor TR_1 is “OFF” and TR_2 is “ON”, switch position “B”.

Then unlike the monostable multivibrator whose output is dependent upon the RC time constant of the feedback components used, the bistable multivibrators output is dependent upon the application of two individual trigger pulses, switch position “A” or position “B”.

So **Bistable Multivibrators** can produce a very short output pulse or a much longer rectangular shaped output whose leading edge rises in time with the externally applied trigger pulse and whose trailing edge is dependent upon a second trigger pulse

Bistable Multivibrator Triggering

To change the stable state of the binary it is necessary to apply an appropriate pulse in the circuit, which will try to bring both the transistors to active region and the resulting regenerative feedback will result on the change of state.

Triggering may be of two following types:

Asymmetrical triggering

Symmetrical triggering

<u>Asymmetrical triggering</u>

In asymmetrical triggering, there are two trigger inputs for the transistors Q1 and Q2. Each trigger input is derived from a separate triggering source. To induce transition among the stable states, let us say that initially the trigger is applied to the bistable. For the next transition, now the identical trigger must appear at the transistor Q2. Thus it can be said that the asymmetrical triggering the trigger pulses derived from two separate source and connected to the two transistors Q1 and Q2 individually, sequentially change the state of the bistable.

Figure shows the circuit diagram of an asymmetrically triggered bistable multivibrator.

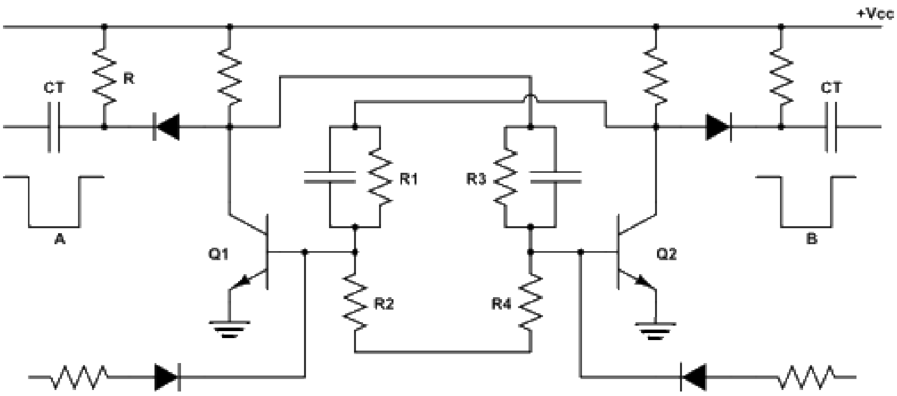


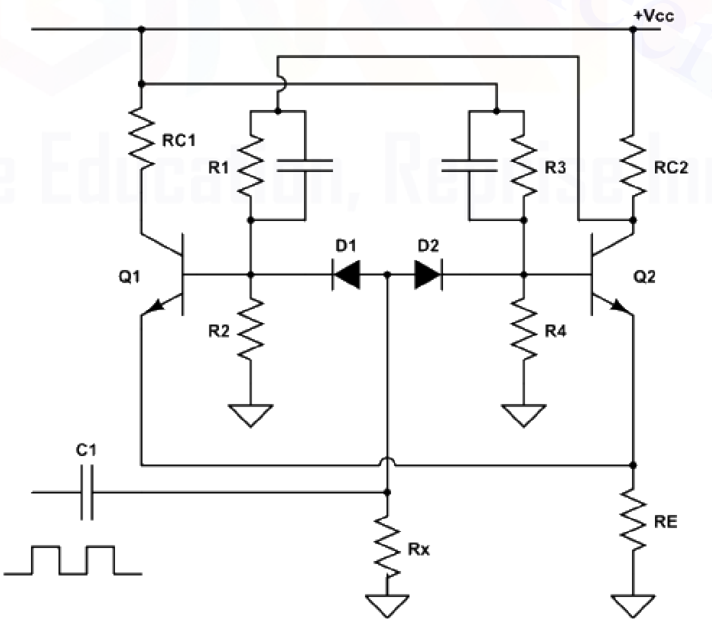
Figure: (b) Asymmetrical triggered bistable multivibrator

Bistable Astable Multivibrator

Initially Q_1 is OFF and transistor Q_2 is ON. The first pulse derived from the trigger source A, applied to the terminal turn it OFF by bringing it from saturation region to active transistor Q_1 is ON and transistor Q_2 is OFF. Any further pulse next time then the trigger pulse is applied at the terminal B, the change of stable state will result with transistor Q_2 On and transistor Q_1 OFF.

Asymmetrical triggering finds its application in the generation of a gate waveform, the duration of which is controlled by any two independent events occurring at different time instants. Thus measurement of time interval is facilitated.

symmetrical triggering



There are various symmetrical triggering methods called symmetrical collector triggering, symmetrical base triggering and symmetrical hybrid triggering. Here we would liked to explain only symmetrical base triggering (positive pulse) only as given under symmetrical Base Triggering.

Figure shows the circuit diagram of a binary with symmetrical base triggering applying a positive trigger pulses.

Diodes D_1 and D_2 are steering diodes. Here the positive pulses, try to turn ON and OFF transistor. Thus when transistor Q_1 is OFF and transistor Q_2 is ON, the respective base voltages and $V_{B1N, OFF}$ and $V_{B2N, ON}$. It will be seen that $V_{B1N, OFF} > V_{B1N, ON}$. Thus diode D_2 is more reverse-biased compared to diode D_1.

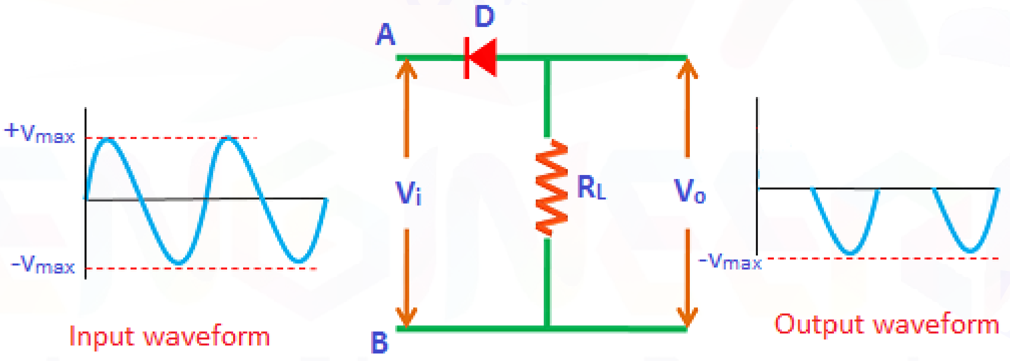
When the positive differentiated pulse of amplitude greater than ($V_{B1N, OFF} + V_\gamma$) appears, the diode D_1 gets forward biased, and transistor Q_1 enters the active region and with subsequent regenerative feedback Q_1 gets ON, and transistor Q_2 becomes OFF. On the arrival of the next trigger pulse now the diode D_2 will be forward biased and ultimately with regenerative feedback it will be in the ON state.

Clippers

Series clipper

The response of the series configuration to a variety of alternating waveforms is provided although first introduced as a half-wave rectifier (for sinusoidal waveforms); there are no boundaries on the type of signals that can be applied to a clipper. The addition of a dc supply can have a pronounced effect on the output of a clipper. Our initial discussion will be limited to ideal diodes, with the effect of VT reserved for a concluding example.

Circuit diagram:



The clipper circuit does not contain energy storage elements such ascapacitor but contains both linear and no-linear elements. The linear elements used in the clippers include resistors and the non-linear elements used in the clippers include diodes or transistors.

One of the basic clipping device is the half wave rectifier. A half wave rectifier removes either the positive half cycle or negative half cycle of the input AC signal and allows the remaining half cycle of the input AC signal. Thus, a half wave rectifier acts as a clipper circuit.

The half wave rectifier (clipper circuit) is made up of one diode and a resistor. Depending on the orientation of the diode, either the positive or the negative half cycle is removed.

The resistor is mainly used to limit the current flowing through the diode when it is forward biased.The clipping (removal) of the input AC signal is done in such a way that the remaining part of the input AC signal will not be distorted.

Clippers are often referred to as voltage limiters, current limiters, slicers, or amplitude selectors. Clipper circuits are extensively used in digital computers, radars, television receivers, radio receivers and other electronic systems for removing unwanted portion of the input AC signal.

Types of clippers

The clipper circuits are generally categorized into three types: series clippers, shunt clippers and dual (combination) clippers. In series clippers, the diode is connected in series with the output load resistance. In shunt clippers, the diode is connected in parallel with the output load resistance.

The series clippers are again classified into four types: series positive clipper, series positive clipper with bias, series negative clipper and series negative clipper with bias. The shunt (parallel) clippers are again classified into four types: shunt positive clipper, shunt positive clipper with bias, shunt negative clipper, and shunt negative clipper with bias.

The various types of clippers are as follows:

1) Positive clipper
2) Negative clipper
3) Biased clipper with
4) Dual (combination) clipper

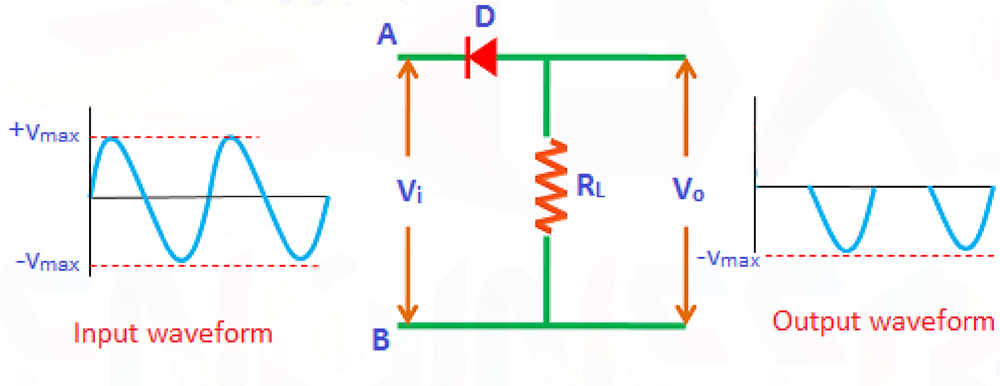
Series positive clipper

In series positive clipper, the positive half cycles of the input AC signal is removed. If the diode is arranged in such a way that the arrowhead of the diode points towards the input and the diode is in series with the output load resistance, then the clipper is said to be a series positive clipper.

In the circuit diagram, the diode D is connected in series with the output load resistance RL and the arrowhead of the diode is pointing towards the input. So the circuit is said to be a series positive clipper.

The vertical line in the diode symbol represents the cathode (n-side) and the opposite end represents the anode (p-side).

During positive half cycle:



Series positive clipper

During the positive half cycle, terminal A is positive and terminal B is negative. That means the positive terminal A is connected to n-side and the negative terminal B is connected to p-side of the diode. As we already know that if the positive terminal is connected to n-side and the negative terminal is connected to p-side then the diode is said to be reverse biased. Therefore, the diode D is reverse biased during the positive half cycle.

During reverse biased condition, no current flows through the diode. So the positive half cycle is blocked or removed at the output.

During negative half cycle:

During the negative half cycle, terminal A is negative and terminal B is positive. That means the negative terminal A is connected to n-side and the positive terminal B is connected to p-side of the diode. As we already know that if the negative terminal is connected to n-side and the positive terminal is connected to p-side then the diode is said to be forward biased. Therefore, the diode D is forward biased during the negative half cycle.

During forward biased condition, electric current flows through the diode. So the negative half cycle is allowed at the output.

Thus, a series of positive half cycles are completely removed at the output.We know that a clipper either clips a portion of half cycle or clips a complete half cycle. In this case, complete half cycles are removed.

Thus, a series positive clipper removes the series of positive half cycles.

- **<u>Series positive clipper with bias</u>**

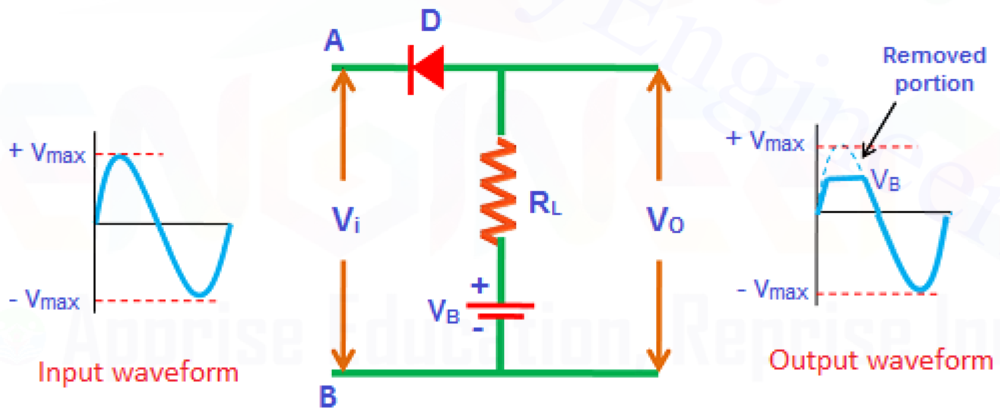
Sometimes it is desired to remove a small portion of positive or negative half cycles. In such cases, the biased clippers are used.

The construction of the series positive clipper with bias is almost similar to the series positive clipper. The only difference is an extra element called battery is used in series positive clipper with bias.

<u>Series positive clipper with positive bias</u>

<u>During positive half cycle:</u>

During the positive half cycle, terminal A is positive and terminal B is negative. That means the positive terminal is connected to n-side and the negative terminal is connected to p-side. As we already know that if the positive terminal is connected to n-side and the negative terminal is connected to p-side then the diode is said to be reverse biased. Therefore, the diode is reverse biased by the input supply voltage V_i.



Series positive clipper with positive bias

However, we are supplying the voltage from another source calledbattery. As shown in the figure, the positive terminal of the battery is connected to p-side and the negative terminal of the battery is connected to n-side of the diode. Therefore, the diode is forward biased by the battery voltage V_B.

That means the diode is reverse biased by the input supply voltage (V_i) and forward biased by the battery voltage (V_B).

Initially, the input supply voltage V_i is less than the battery voltage V_B($V_i < V_B$). So the battery voltage dominates the input supply voltage. Hence, the diode is forward biased by the battery voltage and allows electric current through it. As a result, the signal appears at the output.

When the input supply voltage V_i becomes greater than the battery voltage V_B, the diode D is reverse biased. So no current flows through the diode. As a result, input signal does not appear at the output.

Thus, the clipping (removal of a signal) takes place during the positive half cycle only when the input supply voltage becomes greater than the battery voltage.

During negative half cycle:

During the negative half cycle, terminal A is negative and terminal B is positive. That means the diode D is forward biased due to the input supply voltage. Furthermore, the battery is also connected in such a way that the positive terminal is connected to p-side and the negative terminal is connected to n-side. So the diode is forward biased by both battery voltage V_B and input supply voltage V_i.

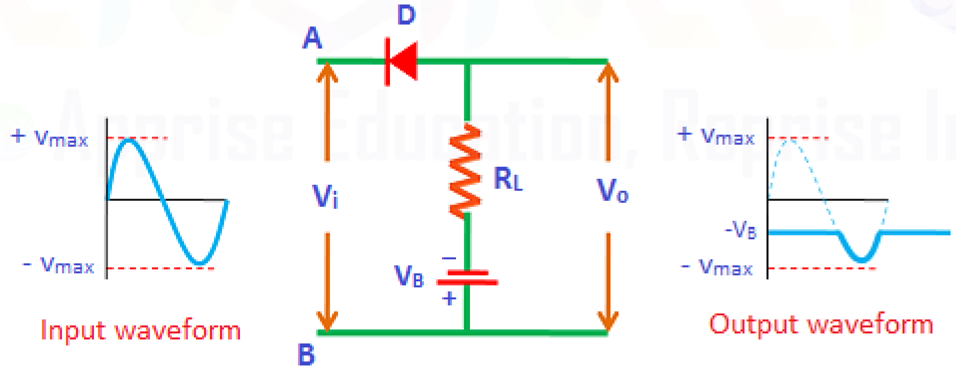
That means, during the negative half cycle, it doesn't matter whether the input supply voltage is greater or less than the battery voltage, the diode always remains forward biased. So the complete negative half cycle appears at the output.

Thus, the series positive clipper with positive bias removes a small portion of positive half cycles.

Series positive clipper with negative bias

During positive half cycle:

During the positive half cycle, the diode D is reverse biased by both input supply voltage V_i and battery voltage V_B. So no signal appears at the output during the positive half cycle. Therefore, the complete positive half cycle is removed.



Series positive clipper with negative bias

During negative half cycle:

During the negative half cycle, the diode is forward biased by the input supply voltage V_i and reverse biased by the battery voltage V_B. However, initially, the battery voltage V_B dominates the input supply voltage V_i. So the diode remains to be reverse biased until the $V_i$becomes greater than V_B. When the input supply voltage V_i becomes greater than the battery voltage V_B, the diode is forward biased by the input supply voltage V_i. So the signal appears at the output.

➢ **Series negative clipper**

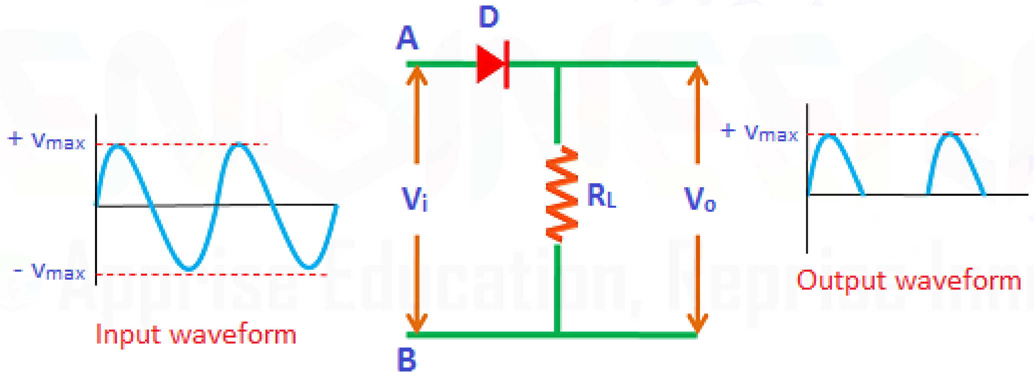
In series negative clipper, the negative half cycles of the input AC signal is removed at the output. The circuit construction of the series negative clipper is shown in the figure.

If the diode is arranged in such a way that the arrowhead of the diode points towards the output and the diode is in series with the output load resistance, then the clipper is said to be a series negative clipper. In simple words, in a series negative clipper, the diode is connected in a direction opposite to that of the series positive clipper.

The vertical line in the diode symbol represents the cathode (n-side) and the opposite end represents the anode (p-side).

During positive half cycle:

During the positive half cycle, terminal A is positive and terminal B is negative. That means the positive terminal A is connected to p-side and the negative terminal B is connected to n-side of the diode. As we already know that if the positive terminal is connected to p-side and the negative terminal is connected to n-side then the diode is said to be forward biased. Therefore, the diode D is forward biased during the positive half cycle.



During forward biased condition, electric current flows through the diode. So the positive half cycle is allowed at the output. Therefore, a series of positive half cycles appears at the output.

During negative half cycle:

During the negative half cycle, the terminal A is negative and the terminal B is positive. That means the negative terminal A is connected to p-side and the positive terminal B is connected to n-side of the diode. As we already know that if the negative terminal is connected to p-side and the positive terminal is connected to n-side then the diode is said to be reverse biased. Therefore, the diode D is reverse biased during the negative half cycle.

During reverse biased condition, no current flows through the diode. So the negative half cycle is completely blocked or removed at the output. In other words, a series of negative half cycles are removed at the output.

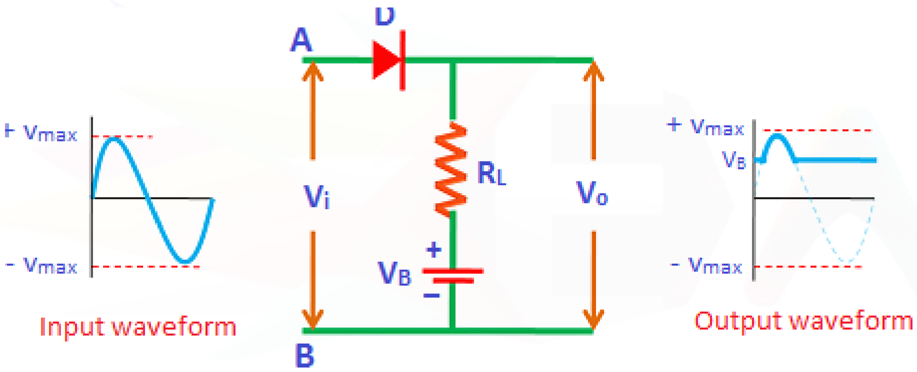
Thus, the series negative clipper removes the series of negative half cycles.

Series negative clipper with bias

Sometimes it is desired to remove a small portion of positive or negative half cycles of the input AC signal. In such cases, the biased clippers are used.

The construction of the series negative clipper with bias is almost similar to the series negative clipper. The only difference is an extra element called battery is used in series negative clipper with bias.

Series negative clipper with positive bias



Series negative clipper with positive bias

During the positive half cycle, terminal A is positive and terminal B is negative. That means the positive terminal A is connected to p-side and the negative terminal B is connected to n-side. As we already know that if the positive terminal is connected to p-side and the negative terminal is connected to n-side then the diode is said to be forward biased.

However, we are also supplying the voltage from another source called battery. As shown in the figure, the positive terminal of the battery is connected to n-side and the negative terminal of the battery is connected to p-side of the diode.

That means the diode is forward biased by input supply voltage $V_i$and reverse biased by battery voltage V_B. Initially, the battery voltage is greater than the input supply voltage. Hence, the diode is reverse biased and does not allow electric current. Therefore, no signal appears at the output.

When the input supply voltage V_i becomes greater than the battery voltage V_B, the diode is forward biased and allows electric current. As a result, the signal appears at the output.

<u>During negative half cycle:</u>

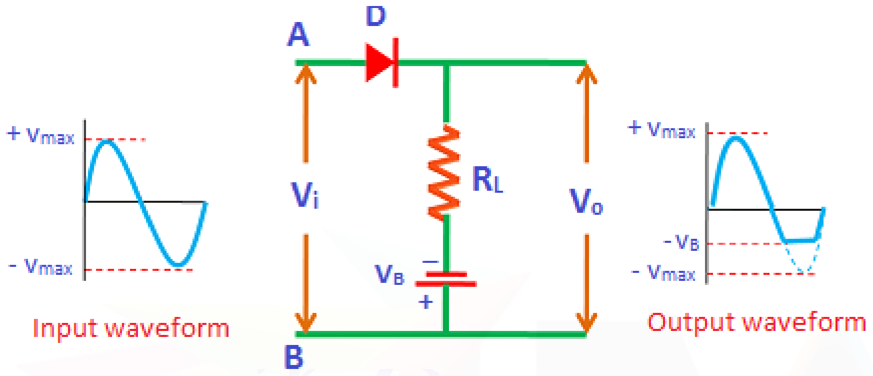
During the negative half cycle, the diode is reverse biased by both input supply voltage V_i and battery voltage V_B. So it doesn’t matter whether the input supply voltage is greater or less than the battery voltage V_B, the diode always remains reverse biased.

Therefore, during the negative half cycle, no signal appears at the output.

Series negative clipper with negative bias

During positive half cycle:

During the positive half cycle, the diode D is forward biased by both input supply voltage V_i and the battery voltage V_B. So it doesn't matter whether the input supply voltage is greater or less than battery voltage V_B, the diode always remains forward biased. Therefore, during the positive half cycle, the signal appears at the output.



During negative half cycle:

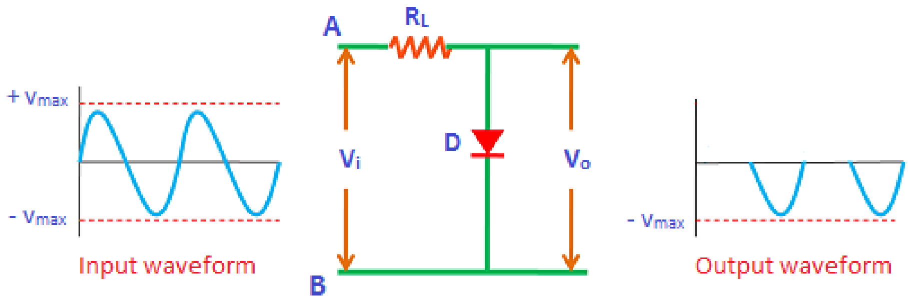
During the negative half cycle, the diode D is reverse biased by the input supply voltage V_i and forward biased by the battery voltage V_B. Initially, the input supply voltage V_i is less than the battery voltage V_B. So the diode is forward biased by the battery voltage V_B. As a result, the signal appears at the output.

When the input supply voltage V_i becomes greater than the battery voltage V_B, the diode will become reverse biased. As a result, no signal appears at the output.

- ## Shunt positive clipper

In shunt clipper, the diode is connected in parallel with the output load resistance. The operating principles of the shunt clipper are nearly opposite to the series clipper.The series clipper passes the input signal to the output load when the diode is forward biased and blocks the input signal when the diode is reverse biased.

The shunt clipper on the other hand passes the input signal to the output load when the diode is reverse biased and blocks the input signal when the diode is forward biased.



Shunt positive clipper

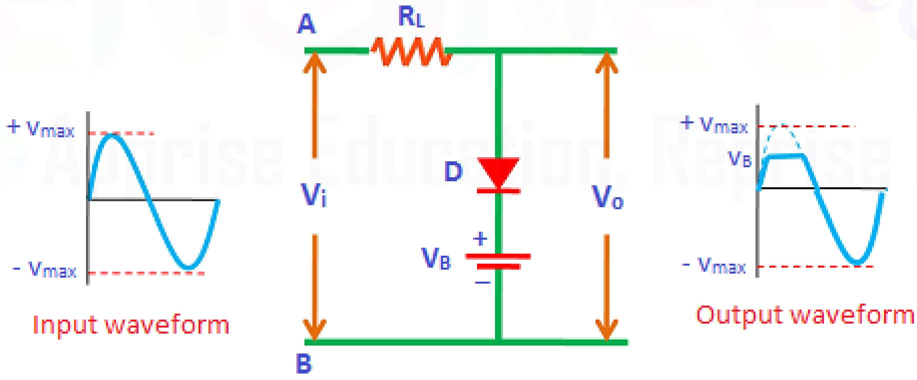
In shunt positive clipper, during the positive half cycle the diode is forward biased and hence no output is generated. On the other hand, during the negative half cycle the diode is reverse biased and hence the entire negative half cycle appears at the output.

- **Shunt positive clipper with bias**

Shunt positive clipper with positive bias

During the positive half cycle, the diode is forward biased by the input supply voltage Vi and

reverse biased by the battery voltage VB. However, initially, the input supply voltage Vi is less than the battery voltage VB. Hence, the battery voltage VB makes the diode to be reverse biased. Therefore, the signal appears at the output. However, when the input supply voltage Vi becomes greater than the battery voltage VB, the diode D is forward biased by the input supply voltage Vi. As a result, no signal appears at the output.

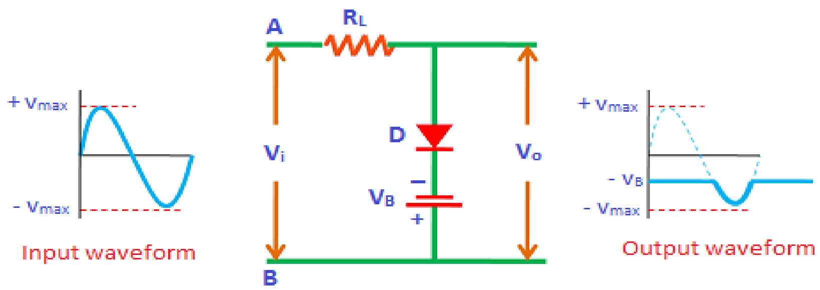


Shunt positive clipper with positive bias

During the negative half cycle, the diode is reverse biased by both input supply voltage and battery voltage. So it doesn't matter whether the input supply voltage is greater or lesser than the battery voltage, the diode always remains reverse biased. As a result, a complete negative half cycle appears at the output.

Shunt positive clipper with negative bias

During the positive half cycle, the diode is forward biased by both input supply voltage V_i and battery voltage V_B. Therefore, no signal appears at the output during the positive half cycle.

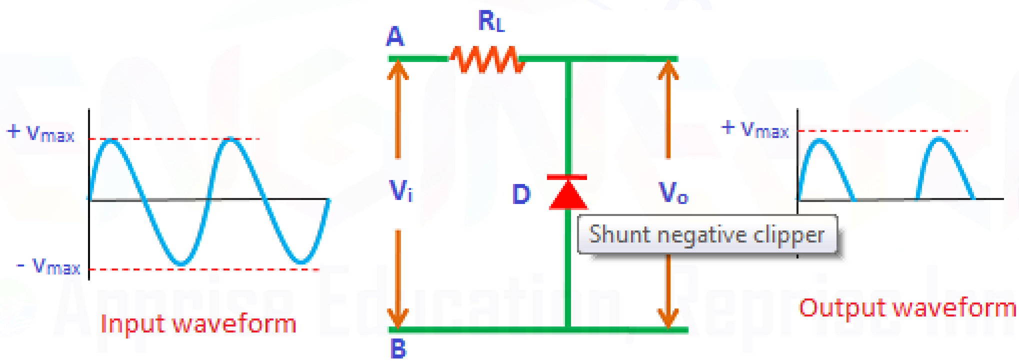


Shunt positive clipper with negative bias

During the negative half cycle, the diode is reverse biased by the input supply voltage and forward biased by the battery voltage. However, initially, the input supply voltage V_i is less than the battery voltage V_B. So the battery voltage makes the diode to be forward biased. As a result, no signal appears at the output. However, when the input supply voltage V_i becomes greater than the battery voltage V_B, the diode is reverse biased by the input supply voltage V_i. As a result, the signal appears at the output.

- **Shunt negative clipper**

In shunt negative clipper, during the positive half cycle the diode is reverse biased and hence the entire positive half cycle appears at the output. On the other hand, during the negative half cycle the diode is forward biased and hence no output signal is generated.

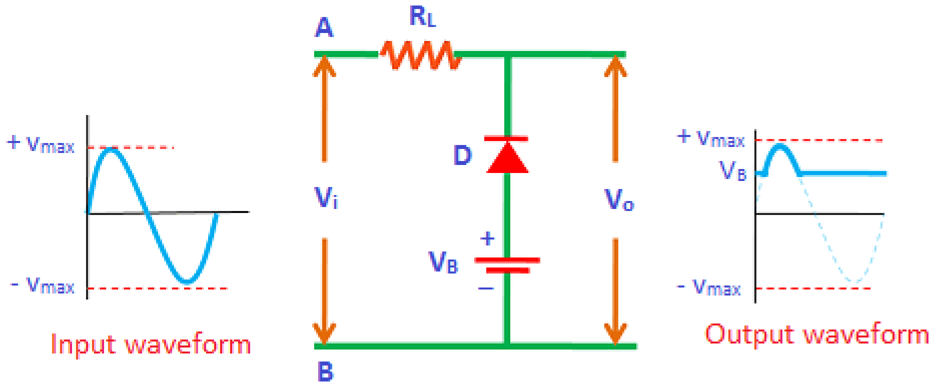


Shunt negative clipper

- **Shunt negative clipper with bias**

Shunt negative clipper with positive bias

During the positive half cycle, the diode is reverse biased by the input supply voltage V_i and forward biased by the battery voltage V_B. However, initially, the input supply voltage is less than the battery voltage. So the diode is forward biased by the battery voltage. As a result, no signal appears at the output. However, when the input supply voltage becomes greater than the battery voltage then the diode is reverse biased by the input supply voltage. As a result, the signal appears at the output.

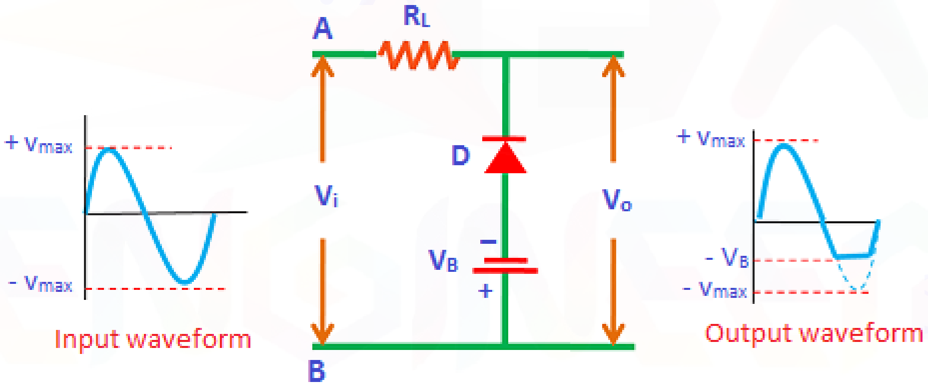


Shunt negative clipper with positive bias

During the negative half cycle, the diode is forward biased by both input supply voltage V_i and battery voltage V_B. So the complete negative half cycle is removed at the output.

Shunt negative clipper with negative bias

During the positive half cycle, the diode is reverse biased by both input supply voltage V_i and battery voltage V_B. As a result, the complete positive half cycle appears at the output.



Shunt negative clipper with negative bias

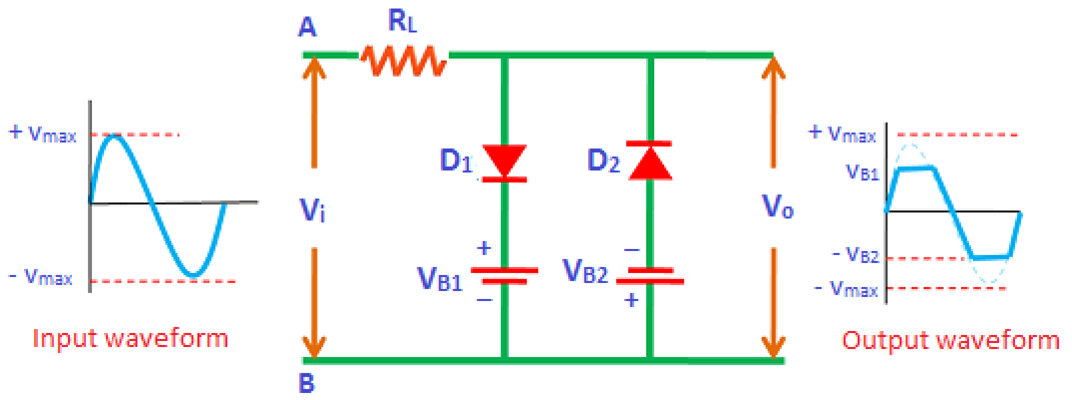
During the negative half cycle, the diode is forward biased by the input supply voltage V_i and reverse biased by the battery voltage V_B. However, initially, the input supply voltage is less than the battery voltage. So the diode is reverse biased by the battery voltage. As a result, the signal appears at the output. However, when the input supply voltage becomes greater than the battery voltage, the diode is forward biased by the input supply voltage. As a result, the signal does not appear at the output.

➢ **<u>Dual (combination) clipper</u>**

Sometimes it is desired to remove a small portion of both positive and negative half cycles. In such cases, the dual clippers are used.The dual clippers are made by combining the biased shunt positive clipper and biased shunt negative clipper.

Let us consider a dual clipper circuit in which a sinusoidal ac voltage is applied to the input terminals of the circuit.

<u>During positive half cycle:</u>



Dual (Combination) clipper

During the positive half cycle, the diode D_1 is forward biased by the input supply voltage V_i and reverse biased by the battery voltage V_{B1}. On the other hand, the diode D_2 is reverse biased by both input supply voltage V_i and battery voltage V_{B2}.

Initially, the input supply voltage is less than the battery voltage. So the diode D_1 is reverse biased by the battery voltage V_{B1}. Similarly, the diode D_2 is reverse biased by the battery voltage V_{B2}. As a result, the signal appears at the output. However, when the input supply voltage V_i becomes greater than the battery voltage V_{B1}, the diode D_1 is forward biased by the input supply voltage. As a result, no signal appears at the output.

<u>During negative half cycle:</u>

During the negative half cycle, the diode D_1 is reverse biased by both input supply voltage V_i and battery voltage V_{B1}. On the other hand, the diode D_2 is forward biased by the input supply voltage V_i and reverse biased by the battery voltage V_{B2}.

Initially, the battery voltage is greater than the input supply voltage. Therefore, the diode D_1 and diode D_2 are reverse biased by the battery voltage. As a result, the signal appears at the output.

When the input supply voltage becomes greater than the battery voltage V_{B2}, the diode D_2 is forward biased. As a result, no signal appears at the output.

<u>Applications of clippers</u>

- Clippers are commonly used in power supplies.
- Used in TV transmitters and Receivers
- They are employed for different wave generation such as square, rectangular, or trapezoidal waves.
- Series clippers are used as noise limiters in FM transmitters.

Schmitt Trigger

Schmitt trigger belongs to a class of bistable multivibrator circuits. In a bistable, there exist two D.C. couplings from each output to input of the other. But in Schmitt trigger circuit, there exists only one coupling. It can be recalled that if in the emitter coupled bistable the feedback network from the collector of transistor Q_2 to the base of transistor Q_1 is removed , it becomes a Schmitt trigger circuit.

The Schmitt trigger is used for wave shaping circuits. It can be used for generation of a square wave from a sine wave input. Basically, the circuit has two opposite operating states like in all other multivibrator circuits. However, the trigger signal is not, typically, a pulse waveform but a slowly varying A.C. Voltage. The Schmitt trigger is level sensitive and switches the output state at two distinct trigger levels. One of the triggering levels is called a lower trigger level (abbreviated as L.T.L) and the other as upper trigger level (abbreviated as U.T.L).

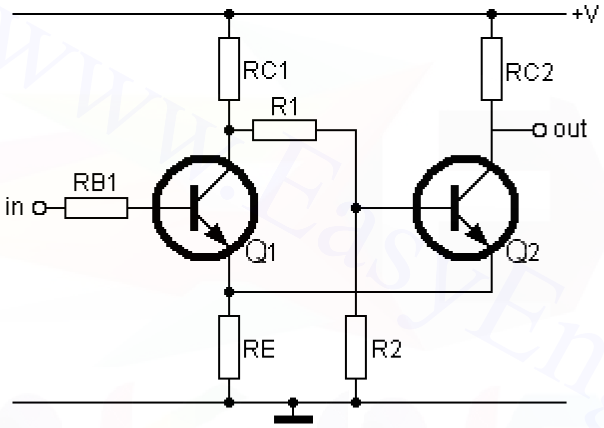
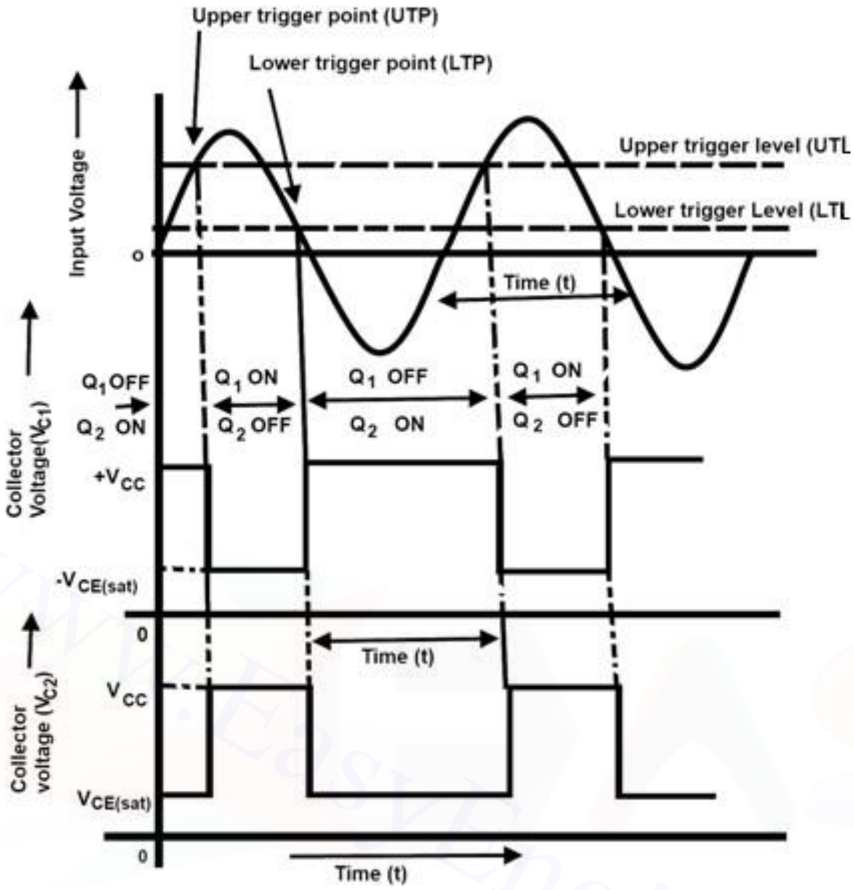


Figure 1 shows the circuit of a Schmitt trigger, the circuit of Schmitt trigger contains of two identical transistors Q_1 and Q_2 coupled through an emitter R_E. The resistor R_1 and R_2 form a voltage divider across the V_{CC} supply and ground. These resistors provide a small forward bias on the base of transistor Q_2.

Let us suppose that initially there is no signal at the input. Then as soon as the power supply V_{CC} is switched on, the transistor Q_2 starts conducting. The flow of its current through resistor R_E produces a voltage drop across it. This voltage drop acts as a reverse bias across the emitter junction of transistor Q_1 due to which it cuts-off. As a result of this, the voltage at its collector rises to V_{CC}. This rising voltage is coupled to the base of transistor Q_2 through the resistor R_1. It increases the forward bias at the base of transistor Q_2 and therefore drives it into saturation and holds it there. At this instant, the collector voltage, level are $V_{C1} = V_{CC}$ and $V_{C2} = V_{CE(sat)}$ as shown in Figure 2.

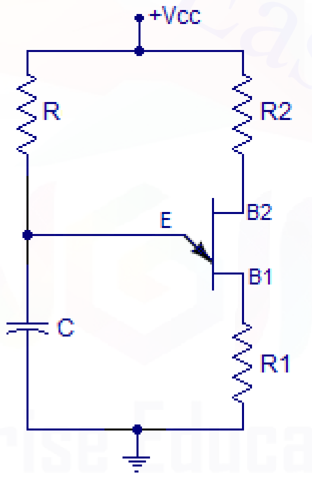


Now suppose an A.C. signal is applied at the input of the Schmitt trigger (i.e. at the base of the transistor Q_1). As the input voltage increases above zero, nothing will happen till it crosses the upper trigger level (U.L.T). As the input voltage increases, above the upper trigger level, the transistor Q_1 conducts. The point, at which it starts conducting, is known as upper trigger point (U.T.P). As the transistor Q_1 conducts, its collector voltage falls below V_{CC}. This fall is coupled through resistor R_1 to the base of transistor Q_2 which reduces its forward bias. This in turn reduces the current of transistor Q_2 and hence the voltage drop across the resistor R_E. As a result of this, the reverse bias of transistor Q_1 is reduced and it conducts more. As the transistor Q_1 conducts more heavily, its collector further reduces due to which the transistor Q_1 conducts near cut-off. This process continues till the transistor Q_1 is driven into saturation and Q_2 into cut-off. At this instant, the collector voltage levels are $V_{C1} = V_{CE(sat)}$ and $V_{C2} = V_{CC}$ as shown in the figure.

The transistor Q_1 will continue to conduct till the input voltage falls below the lower trigger level (L.T.L). It will be interesting to know that when the input voltage becomes equal to the lower trigger level, the emitter base junction of transistor Q_1 becomes reverse biased. As a result of this, its collector voltage starts rising toward V_{CC}. This rising voltage increases the forward bias across transistor Q_2 due to which it conducts. The point, at which transistor Q_2 starts conducting, is called lower trigger point (L.T.P). Soon the transistor Q_2 is driven into saturation and Q_1 to cur-off. This completes one cycle. The collector voltage levels at this instant are $V_{C1} = V_{CC}$ and $V_{C2} = V_{CE(sat)}$. No change in state will occur during the negative half cycle of the input voltage.

It is proved from the above discussion that the output of a Schmitt trigger is a positive going pulse, whose width depends upon the time during which transistor Q_1 is conducting. The conduction time is set by the upper and lower trigger levels.

UJT relaxation oscillator



UJT relaxation oscillator www.circuitstoday.com

The circuit diagram of a UJT relaxation oscillator is given shown above. R1 and R2 are current limiting resistors. Resistor R and capacitor C determines the frequency of the oscillator.

The frequency of the UJT relaxation oscillator can be expressed by the equation

F = 1/ (RC ln(1/(1-η)) where η is the intrinsic standoff ratio and *ln* stand for natural logarithm.

When power supply is switched ON the capacitor C starts charging through resistor R. The capacitor keeps on charging until the voltage across it becomes equal to 0.7V plus ηVbb.

This voltage is the peak voltage point "Vp" denoted in the characteristics curve After this point the emitter to RB1 resistance drops drastically and the capacitors starts discharging through this path.

When the capacitor is discharged to the valley point voltage "Vv" (refer Fig : 1) the emitter to RB1 resistance climbs again and the capacitor starts charging. This cycle is repeated and results in a sort of sawtooth waveform across the capacitor.

The saw tooth waveform across the capacitor of a typical UJT relaxation oscillator is shown in the figurebelow.

The UJT has negative resistance characteristic, because of this character the UJT provides trigger pulse. Any one of the three terminals can be taken for triggering pulse. The UJT can be used as relaxation oscillator i.e. it produces non-sinusoidal waves. The circuit diagram of relaxation oscillator

First the capacitor 'C' starts charging through the resistor R when V_{BB} is switched on. During the charging of the capacitor, the voltage across it increases exponentially until it reaches to the peak point voltage V_P. Up to now, the UJT is in off state, i.e non- conducting state at which R_{B1} value is high.

When the voltage across the capacitor reaches to peak point voltage (V_P) then, UJT comes into conducting state as the junction is forward biased and R_{B1} falls to low value (50Ω). Then the capacitor 'C' quickly discharges through UJT that means the discharging time is very less as the capacitor discharges through the low resistance UJT.

When the voltage across the capacitor decreases to valley point voltage (V_V) then the UJT shifts to off state and once again the capacitor gets charged through the resistor R and this process is repeated. This generates saw-tooth wave form (Fig.2) across the capacitor which can be viewed on the CRO screen.

We know in RC circuit the instant charge $q = q_0(1 - e^{-t/RC})$

Multiplying this equation by R\t $\frac{Rq}{t} = \frac{Rq_0}{t}(1 - e^{-t/RC})$

$RI = RI_0(1 - e^{-t/RC})$ $\because I = q / t$

$V = V_0(1 - e^{-t/RC})$ ------- (1) $\because RI = V$

This is the general equation for voltage in the charging of a capacitor in RC circuit.

But the maximum voltage across the condenser $V_0 = V_{BB}$ and $V = V_P - V_B$

The equation (1) becomes $V_P - V_B = V_{BB}(1 - e^{-t/RC})$

Here, t is the time taken by the condenser to get a charging potential = $(V_P - V_B)$

$$\therefore (1 - e^{-t/RC}) = \frac{V_P - V_B}{V_{BB}} = \eta$$ Where η is intrinsic stand off ratio.

$$\therefore e^{-t/RC} = (1 - \eta)$$

OR $$\therefore e^{t/RC} = \frac{1}{(1 - \eta)}$$

$$t = RC\log_e\left(\frac{1}{1-\eta}\right) \quad \text{------------ (2)}$$

The time period of the saw tooth wave

T = time of charging + time of discharging

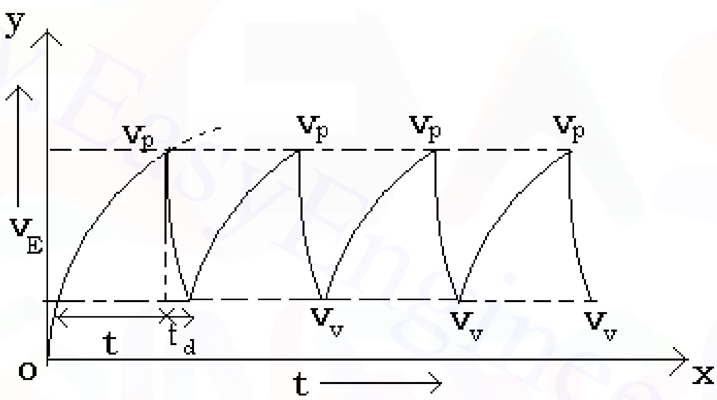
$T = t + t_d$

t_d value is negligibly small When compared to the value of t, as the discharge takes through the low resistance R_{B1}.

$T \approx t$ --------------- (3)

From the equations (2) and (3) the time period of the saw tooth wave

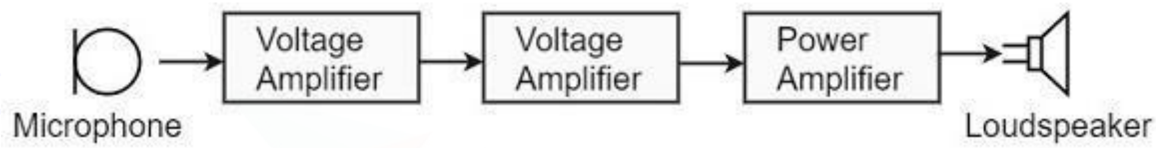
$$T = 2.303\, RC \log_{10}\left(\frac{1}{1-\eta}\right) \quad \text{or frequency } f = 1/T$$



Unit 5
POWER AMPLIFIERS AND DC CONVERTERS

Power Amplifier

After the audio signal is converted into electrical signal, it has several voltage amplifications done, after which the power amplification of the amplified signal is done just before the loud speaker stage. This is clearly shown in the below figure.



While the voltage amplifier raises the voltage level of the signal, the power amplifier raises the power level of the signal. Besides raising the power level, it can also be said that a power amplifier is a device which converts DC power to AC power and whose action is controlled by the input signal.

The DC power is distributed according to the relation,

DC power input = AC power output + losses

PowerTransistor

For such Power amplification, a normal transistor would not do. A transistor that is manufactured to suit the purpose of power amplification is called as a **Power transistor**.A Power transistor differs from the other transistors, in the following factors.It is larger in size, in order to handle large powers.The collector region of the transistor is made large and a heat sink is placed at the collector-base junction in order to minimize heat generated.

The emitter and base regions of a power transistor are heavily doped.Due to the low input resistance, it requires low input power.Hence there is a lot of difference in voltage amplification and power amplification. So, let us now try to get into the details to understand the differences between a voltage amplifier and a power amplifier.

Difference between Voltage and Power Amplifiers

Let us try to differentiate between voltage and power amplifier.

Voltage Amplifier

The function of a voltage amplifier is to raise the voltage level of the signal. A voltage amplifier is designed to achieve maximum voltage amplification.

The voltage gain of an amplifier is given by

$$Av=\beta(RcRin)Av=\beta(RcRin)$$

The characteristics of a voltage amplifier are as follows −

The base of the transistor should be thin and hence the value of β should be greater than 100.The resistance of the input resistor R_{in} should be low when compared to collector load R_C.The collector load R_C should be relatively high. To permit high collector load, the voltage amplifiers are always operated at low collector current.The voltage amplifiers are used for small signal voltages.

Power Amplifier

The function of a power amplifier is to raise the power level of input signal. It is required to deliver a large amount of power and has to handle large current.

The characteristics of a power amplifier are as follows −

The base of transistor is made thicken to handle large currents. The value of β being (β > 100) high.The size of the transistor is made larger, in order to dissipate more heat, which is produced during transistor operation.Transformer coupling is used for impedance matching.

Collector resistance is made low.he comparison between voltage and power amplifiers is given below in a tabular form.

S.No	Particular	Voltage Amplifier	Power Amplifier
1	β	High (>100)	Low (5 to 20)
2	R_C	High (4-10 KΩ)	Low (5 to 20 Ω)
3	Coupling	Usually R-C	Invariably transformer
4	Input voltage	Low (a few m V)	High (2-4 V)
5	Collector current	Low (≈ 1 mA)	High (> 100 mA)
6	Power output	Low	High
7	Output	High (≈ 12 K Ω)	Low (200 Ω)

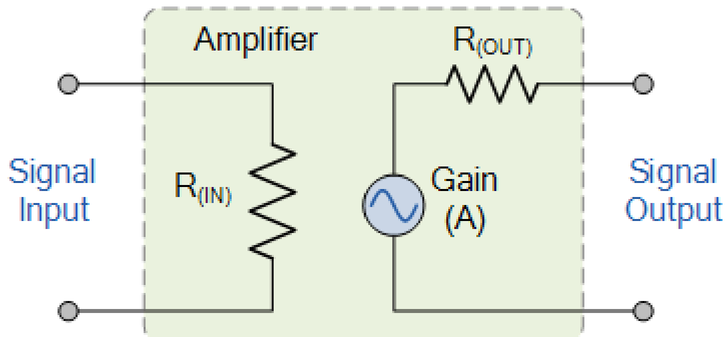
Classification of Signal Amplifier

Type of Signal	Type of	Classification	Frequency of
Small Signal	Common Emitter	Class A Amplifier	Direct Current (DC)
Large Signal	Common Base	Class B Amplifier	Audio Frequencies (AF)
	Common Collector	Class AB Amplifier	Radio Frequencies (RF)
		Class C Amplifier	VHF, UHF and SHF

Amplifiers can be thought of as a simple box or block containing the amplifying device, such as a Bipolar Transistor, Field Effect Transistor or Operational Amplifier, which has two input terminals and two output terminals (ground being common) with the output signal being much greater than that of the input signal as it has been "Amplified".

An ideal signal amplifier will have three main properties: Input Resistance or (RIN), Output Resistance or (ROUT) and of course amplification known commonly as Gain or (A). No matter how complicated an amplifier circuit is, a general amplifier model can still be used to show the relationship of these three properties.

Ideal Amplifier Model



The amplified difference between the input and output signals is known as the Gain of the amplifier. Gain is basically a measure of how much an amplifier "amplifies" the input signal.

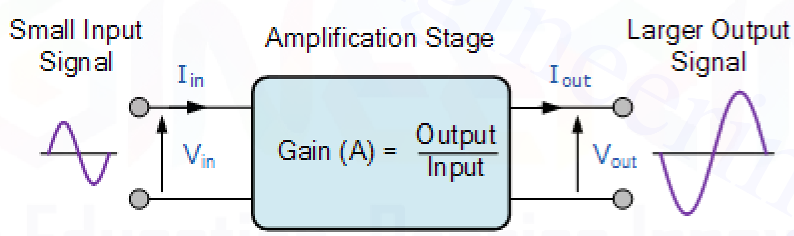
For example, if we have an input signal of 1 volt and an output of 50 volts, then the gain of the amplifier would be “50”. In other words, the input signal has been increased by a factor of 50. This increase is called Gain.

Amplifier gain is simply the ratio of the output divided-by the input. Gain has no units as its a ratio, but in Electronics it is commonly given the symbol “A”, for Amplification. Then the gain of an amplifier is simply calculated as the “output signal divided by the input signal”.

Amplifier Gain

The introduction to the amplifier gain can be said to be the relationship that exists between the signal measured at the output with the signal measured at the input. There are three different kinds of amplifier gain which can be measured and these are: *Voltage Gain* (Av), *Current Gain* (Ai) and *Power Gain* (Ap) depending upon the quantity being measured with examples of these different types of gains are given below.

Amplifier Gain of the Input Signal



Voltage Amplifier Gain

$$Voltage\,Gain\,(A_v) = \frac{Output\ Voltage}{Input\ Voltage} = \frac{Vout}{Vin}$$

Current Amplifier Gain

$$Current\,Gain\,(A_i) = \frac{Output\ Current}{Input\ Current} = \frac{Iout}{Iin}$$

Power Amplifier Gain

$$Power Gain\,(A_p) = A_v \times A_i$$

Note that for the Power Gain you can also divide the power obtained at the output with the power obtained at the input. Also when calculating the gain of an amplifier, the subscripts v, i and p are used to denote the type of signal gain being used.

The power gain (Ap) or power level of the amplifier can also be expressed in Decibels, (dB). The Bel (B) is a logarithmic unit (base 10) of measurement that has no units. Since the Bel is too large a unit of measure, it is prefixed with deci making it Decibels instead with one decibel being one tenth (1/10th) of a Bel. To calculate the gain of the amplifier in Decibels or dB, we can use the following expressions.

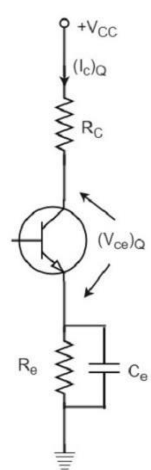
Voltage Gain in dB: av = 20*log(Av)

Current Gain in dB: ai = 20*log(Ai)

Power Gain in dB: ap = 10*log(Ap)

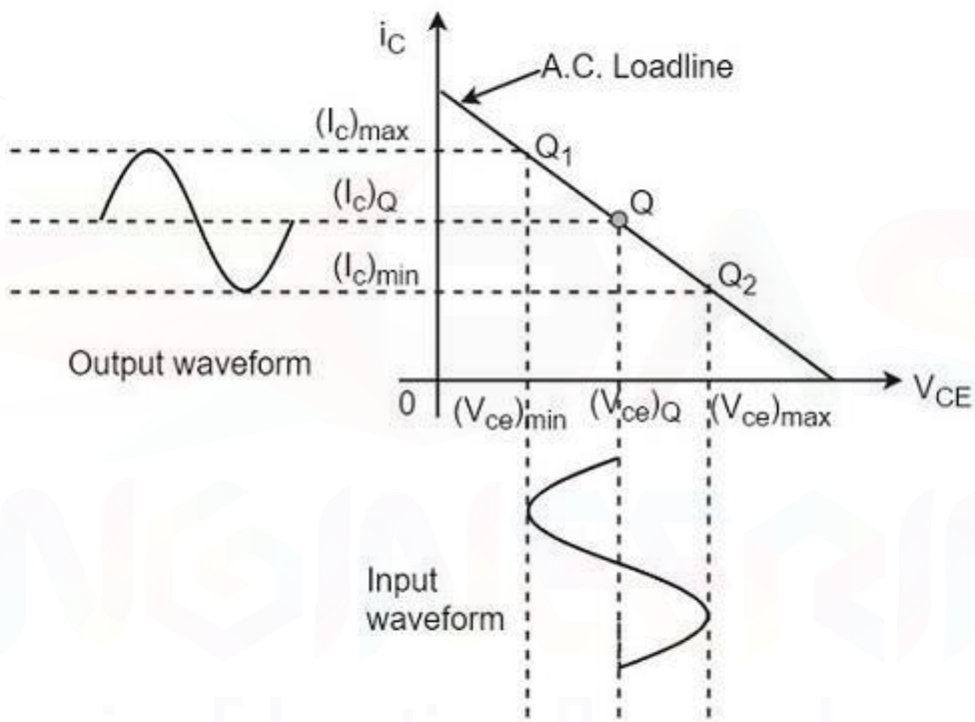
We have already come across the details of transistor biasing, which is very important for the operation of a transistor as an amplifier. Hence to achieve faithful amplification, the biasing of the transistor has to be done such that the amplifier operates over the linear region.

A Class A power amplifier is one in which the output current flows for the entire cycle of the AC input supply. Hence the complete signal present at the input is amplified at the output. The following figure shows the circuit diagram for Class A Power amplifier.



From the above figure, it can be observed that the transformer is present at the collector as a load. The use of transformer permits the impedance matching, resulting in the transference of maximum power to the load e.g. loud speaker.

The operating point of this amplifier is present in the linear region. It is so selected that the current flows for the entire ac input cycle. The below figure explains the selection of operating point.



The output characteristics with operating point Q is shown in the figure above. Here $(I_c)_Q$ and $(V_{ce})_Q$ represent no signal collector current and voltage between collector and emitter respectively. When signal is applied, the Q-point shifts to Q_1 and Q_2. The output current increases to $(I_c)_{max}$ and decreases to $(I_c)_{min}$. Similarly, the collector-emitter voltage increases to $(V_{ce})_{max}$ and decreases to $(V_{ce})_{min}$.

D.C. Power drawn from collector battery V_{cc} is given by

Pin=voltage×current=VCC(IC)QPin=voltage×current=VCC(IC)Q

This power is used in the following two parts –

Power dissipated in the collector load as heat is given by

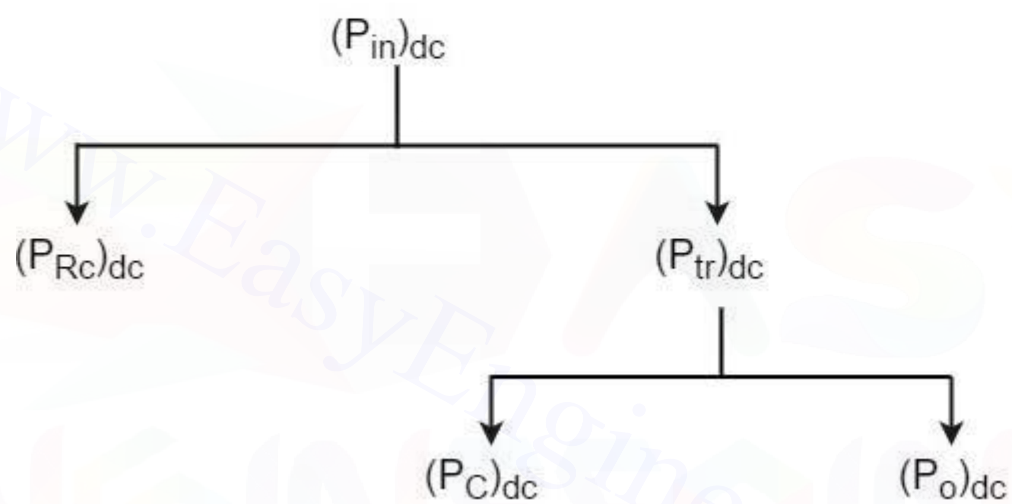
PRC=(current)2×resistance=(IC)2QRC

Power given to transistor is given by

Ptr=Pin−PRC=VCC−(IC)2QRCPtr=Pin−PRC=VCC−(IC)Q2RC

When signal is applied, the power given to transistor is used in the following two parts −

The D.C. power dissipated by the transistor (collector region) in the form of heat, i.e., $(P_C)_{dc}$



This class A power amplifier can amplify small signals with least distortion and the output will be an exact replica of the input with increased strength.

Overall Efficiency

The overall efficiency of the amplifier circuit is given by

(η)overall=a.cpowerdeliveredtotheloadtotalpowerdeliveredbyd.csupply(η)overall=a.cpowerdeliveredtotheloadtotalpowerdeliveredbyd.csupply

=(PO)ac(Pin)dc=(PO)ac(Pin)dc

Collector Efficiency

The collector efficiency of the transistor is defined as

(η)collector=averagea.cpoweroutputaveraged.cpowerinputtotransistor(η)collector=averagea.cpoweroutputaveraged.cpowerinputtotransistor

$$=(PO)ac(Ptr)dc$$

Expression for overall efficiency

$$(PO)ac=Vrms \times Irms(PO)ac=Vrms \times Irms$$

$$=12–\sqrt{[(Vce)max-(Vce)min2]} \times 12–\sqrt{[(IC)max-(IC)min2]}=12[(Vce)max-(Vce)min2] \times 12[(IC)max-(IC)min2]$$

$$=[(Vce)max-(Vce)min] \times [(IC)max-(IC)min]8=[(Vce)max-(Vce)min] \times [(IC)max-(IC)min]8$$

Therefore

$$(\eta)overall=[(Vce)max-(Vce)min] \times [(IC)max-(IC)min]8 \times VCC(IC)Q(\eta)overall=[(Vce)max-(Vce)min] \times [(IC)max-(IC)min]8 \times VCC(IC)Q$$

Advantages of Class A Amplifiers

The advantages of Class A power amplifier are as follows −

- The current flows for complete input cycle
- It can amplify small signals
- The output is same as input
- No distortion is present

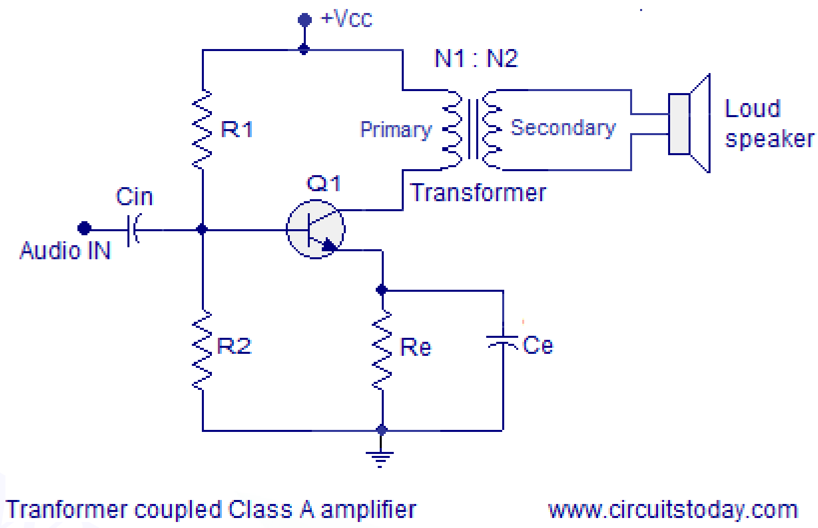
Disadvantages of Class A Amplifiers

The advantages of Class A power amplifier are as follows −

- Low power output
- Low collector efficiency

Transformer coupled Class A power amplifier.

An amplifier where the load is coupled to the output using a transformer is called a transformer coupled amplifier. Using transformer coupling the efficiency of the amplifier can be improved to a great extend. The coupling transformer provides good impedance matching between the output and load and it is the main reason behind the improved efficiency. Impedance matching means making the output impedance of the amplifier equal to the input impedance of the load and this is an important criteria for the transfer of maximum power. Circuit diagram of typical single stage Class A amplifier is shown in the circuit diagram below.



Impedance matching can be attained by selecting the number of turns of the primary so that its net impedance is equal to the transistors output impedance and selecting the number of turns of the secondary so that its net impedance is equal to the loudspeakers input impedance.

Advantages of transformer coupled amplifier.

- Main advantage is the improvement of efficiency.
- Provides good DC isolation as there is no physical connection between amplifier output and load. Audio signals pass from one side to other by virtue of induction.

Disadvantages of transformer coupled amplifier.

- It is a bit hard to make/find an exactly matching transformer.
- Transformers are bulky and so it increases the cost and size of the amplifier.
- Transformer winding does not provide any resistance to DC current. If any DC components if present in the amplifier output, it will flow through the primary winding and saturate the core. This will result in reduced transformer action.
- Transformer coupling reduces the low frequency response of the amplifier.
- Transformer coupling induces hum in the output.
- Transformer coupling can be employed only for small loads.

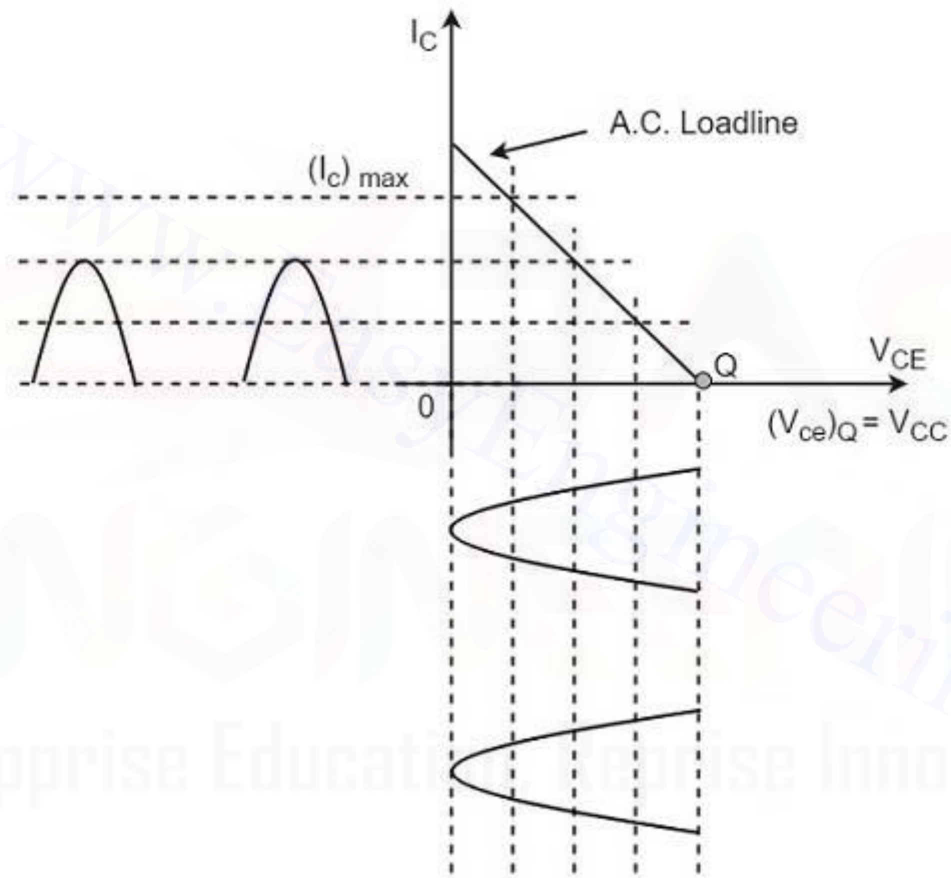
Class B Amplifier

When the collector current flows only during the positive half cycle of the input signal, the power amplifier is known as **class B power amplifier**.

Class B Operation

The biasing of the transistor in class B operation is in such a way that at zero signal condition, there will be no collector current. The **operating point** is selected to be at collector cut off voltage. So, when the signal is applied, **only the positive half cycle** is amplified at the output.

The figure below shows the input and output waveforms during class B operation.



When the signal is applied, the circuit is forward biased for the positive half cycle of the input and hence the collector current flows. But during the negative half cycle of the input, the circuit is reverse biased and the collector current will be absent. Hence **only the positive half cycle** is amplified at the output.

As the negative half cycle is completely absent, the signal distortion will be high. Also, when the applied signal increases, the power dissipation will be more. But when compared to class A power amplifier, the output efficiency is increased.

Well, in order to minimize the disadvantages and achieve low distortion, high efficiency and high output power, the push-pull configuration is used in this class B amplifier.

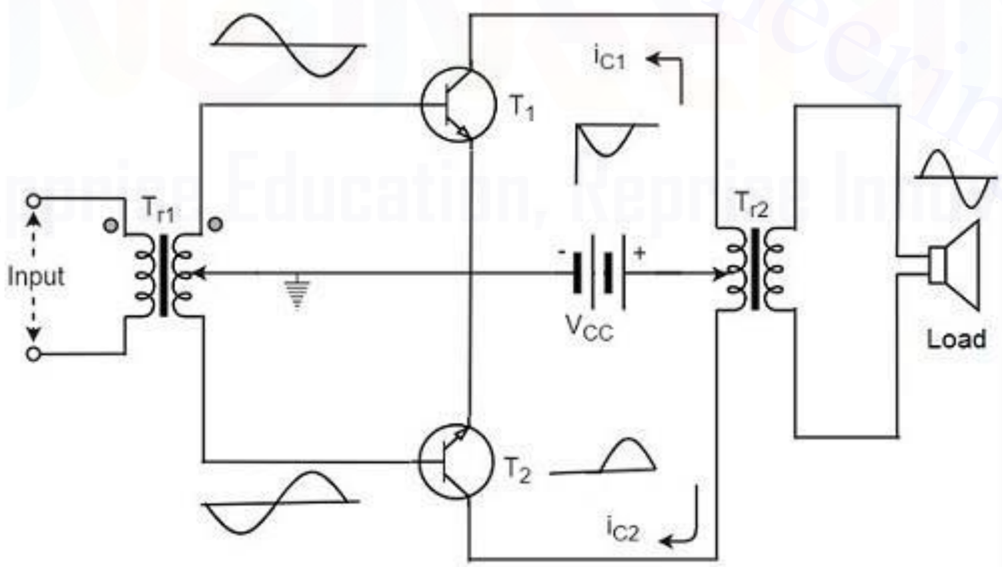
Class B Push-Pull Amplifier

Though the efficiency of class B power amplifier is higher than class A, as only one half cycle of the input is used, the distortion is high. Also, the input power is not completely utilized. In order to compensate these problems, the push-pull configuration is introduced in class B amplifier.

Construction

The circuit of a push-pull class B power amplifier consists of two identical transistors T_1 and T_2 whose bases are connected to the secondary of the center-tapped input transformer T_{r1}. The emitters are shorted and the collectors are given the V_{CC} supply through the primary of the output transformer T_{r2}.

The circuit arrangement of class B push-pull amplifier, is same as that of class A push-pull amplifier except that the transistors are biased at cut off, instead of using the biasing resistors. The figure below gives the detailing of the construction of a push-pull class B power amplifier.

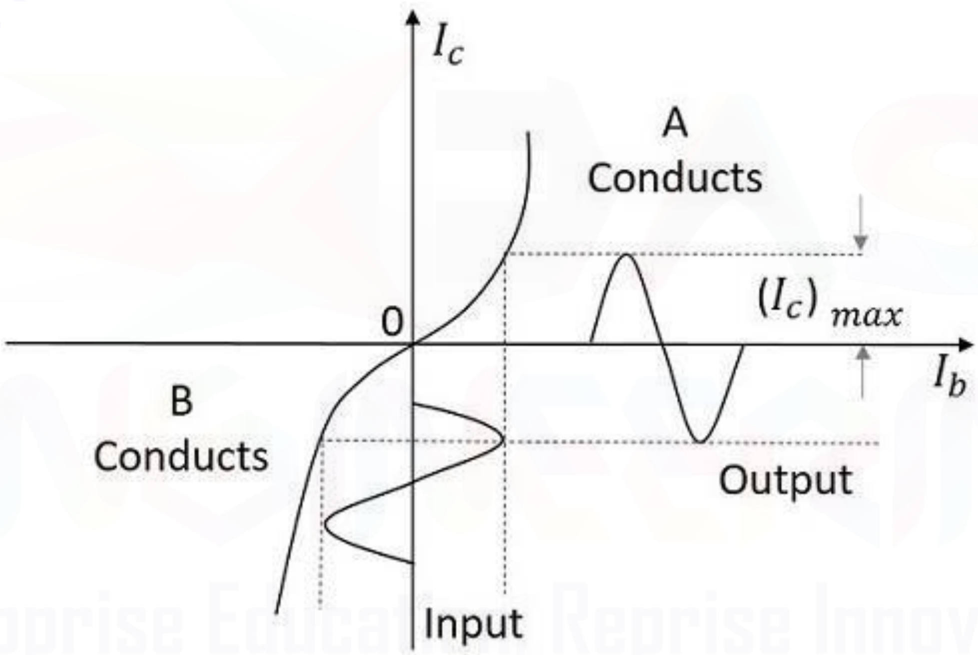


Operation

The circuit of class B push-pull amplifier shown in the above figure clears that both the transformers are center-tapped. When no signal is applied at the input, the transistors T_1 and

T_2 are in cut off condition and hence no collector currents flow. As no current is drawn from V_{CC}, no power is wasted.

When input signal is given, it is applied to the input transformer T_{r1} which splits the signal into two signals that are 180° out of phase with each other. These two signals are given to the two identical transistors T_1 and T_2. For the positive half cycle, the base of the transistor T_1 becomes positive and collector current flows. At the same time, the transistor T_2 has negative half cycle, which throws the transistor T_2 into cutoff condition and hence no collector current flows. The waveform is produced as shown in the following figure.



For the next half cycle, the transistor T_1 gets into cut off condition and the transistor T_2 gets into conduction, to contribute the output. Hence for both the cycles, each transistor conducts alternately. The output transformer T_{r3} serves to join the two currents producing an almost undistorted output waveform.

Power Efficiency of Class B Push-Pull Amplifier

The current in each transistor is the average value of half sine loop.

For half sine loop, I_{dc} is given by

Idc=(IC)maxπIdc=(IC)maxπ

Therefore,

(pin)dc=2×[(IC)maxπ×VCC](pin)dc=2×[(IC)maxπ×VCC]

Here factor 2 is introduced as there are two transistors in push-pull amplifier.

R.M.S. value of collector current = (IC)max/2–√(IC)max/2
R.M.S. value of output voltage = VCC/2–√VCC/2
Under ideal conditions of maximum power

Therefore,

(PO)ac=(IC)max2–√×VCC2–√=(IC)max×VCC2(PO)ac=(IC)max2×VCC2=(IC)max×VCC2

Now overall maximum efficiency

ηoverall=(PO)ac(Pin)dcηoverall=(PO)ac(Pin)dc

=(IC)max×VCC2×π2(IC)max×VCC=(IC)max×VCC2×π2(IC)max×VCC

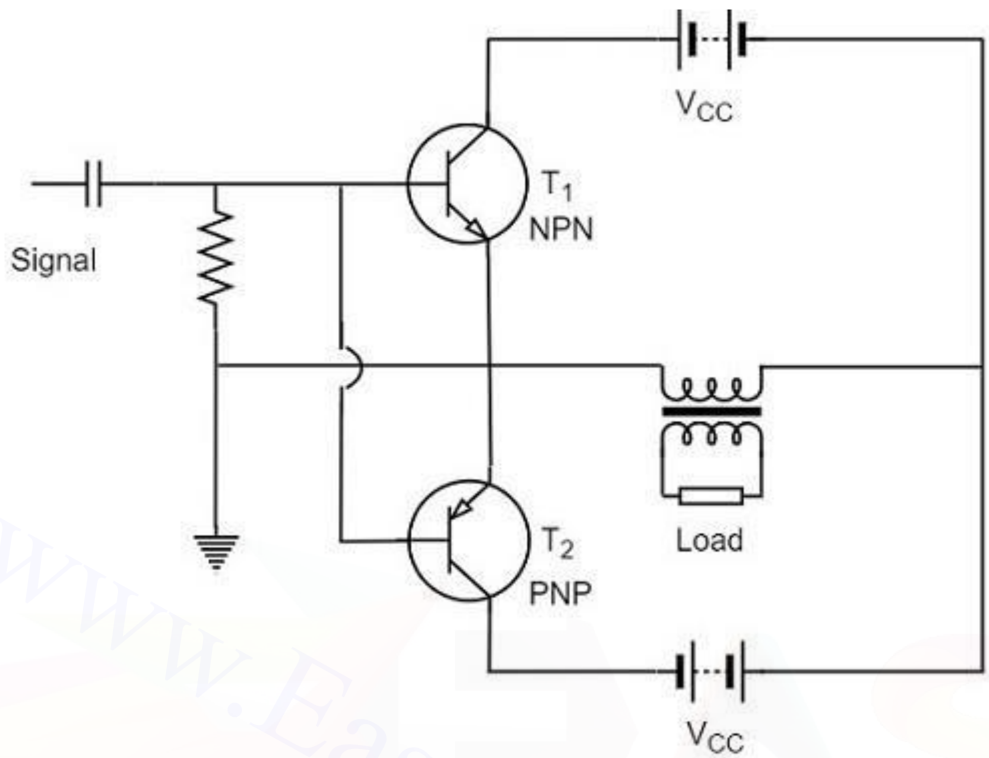
=π4=0.785=78.5%=π4=0.785=78.5%

The collector efficiency would be the same.

Hence the class B push-pull amplifier improves the efficiency than the class A push-pull amplifier.

Complementary Symmetry Push-Pull Class B Amplifier

The push pull amplifier which was just discussed improves efficiency but the usage of center-tapped transformers makes the circuit bulky, heavy and costly. To make the circuit simple and to improve the efficiency, the transistors used can be complemented, as shown in the following circuit diagram.



The above circuit employs a NPN transistor and a PNP transistor connected in push pull configuration. When the input signal is applied, during the positive half cycle of the input signal, the NPN transistor conducts and the PNP transistor cuts off. During the negative half cycle, the NPN transistor cuts off and the PNP transistor conducts.

In this way, the NPN transistor amplifies during positive half cycle of the input, while PNP transistor amplifies during negative half cycle of the input. As the transistors are both complement to each other, yet act symmetrically while being connected in push pull configuration of class B, this circuit is termed as

Complementary symmetry push pull class B amplifier.

Advantages

The advantages of Complementary symmetry push pull class B amplifier are as follows.

- As there is no need of center tapped transformers, the weight and cost are reduced.
- Equal and opposite input signal voltages are not required.

Disadvantages

The disadvantages of Complementary symmetry push pull class B amplifier are as follows.

- It is difficult to get a pair of transistors (NPN and PNP) that have similar characteristics.
- We require both positive and negative supply voltages.

Class AB and Class C Power Amplifiers

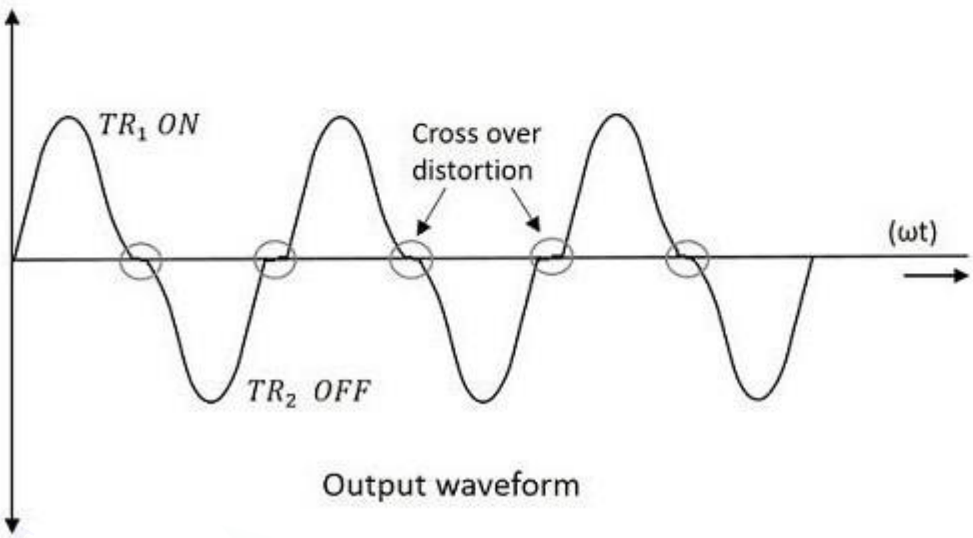
The class A and class B amplifier so far discussed has got few limitations. Let us now try to combine these two to get a new circuit which would have all the advantages of both class A and class B amplifier without their inefficiencies. Before that, let us also go through another important problem, called as **Cross over distortion**, the output of class B encounters with.

Cross-over Distortion

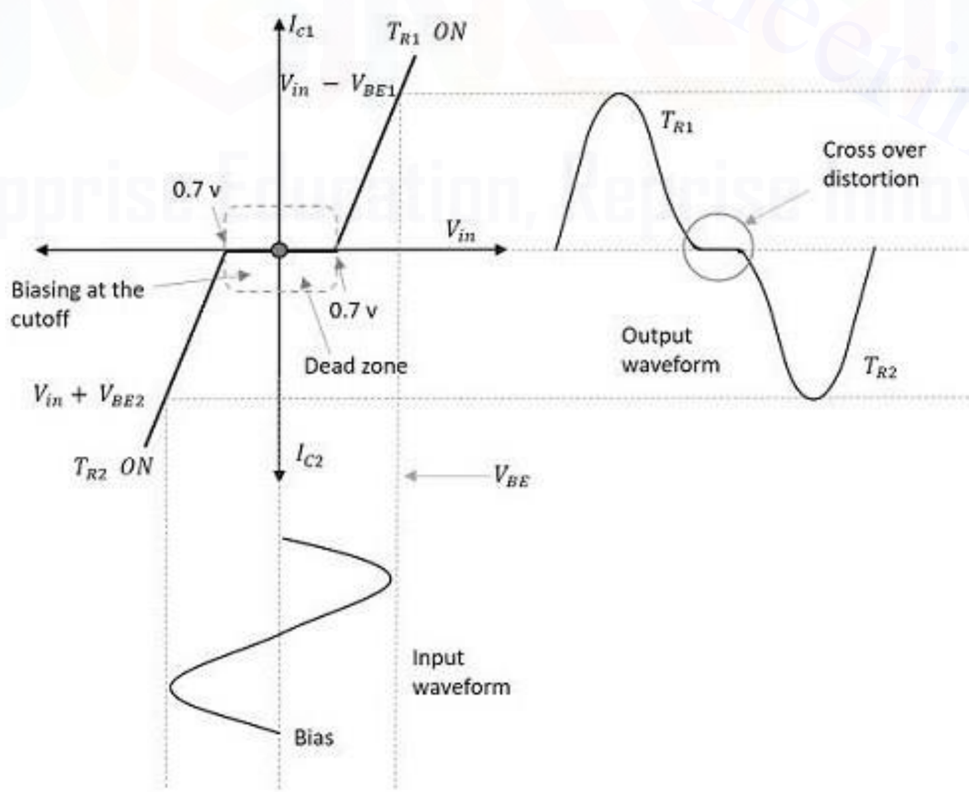
In the push-pull configuration, the two identical transistors get into conduction, one after the other and the output produced will be the combination of both.

When the signal changes or crosses over from one transistor to the other at the zero voltage point, it produces an amount of distortion to the output wave shape. For a transistor in order to conduct, the base emitter junction should cross 0.7v, the cut off voltage. The time taken for a transistor to get ON from OFF or to get OFF from ON state is called the **transition period**.

At the zero voltage point, the transition period of switching over the transistors from one to the other, has its effect which leads to the instances where both the transistors are OFF at a time. Such instances can be called as **Flat spot** or **Dead band** on the output wave shape.



The above figure clearly shows the cross over distortion which is prominent in the output waveform. This is the main disadvantage. This cross over distortion effect also reduces the overall peak to peak value of the output waveform which in turn reduces the maximum power output. This can be more clearly understood through the non-linear characteristic of the waveform as shown below.



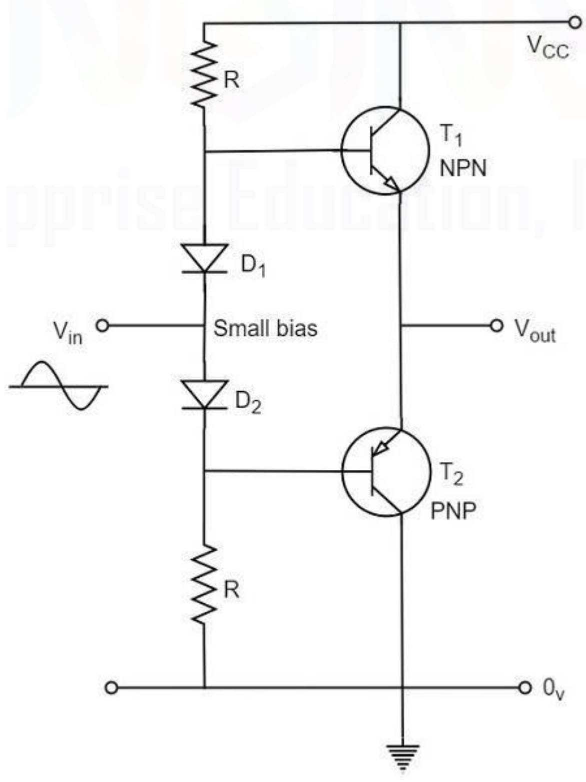
It is understood that this cross-over distortion is less pronounced for large input signals, where as it causes severe disturbance for small input signals. This cross over distortion can be eliminated if the conduction of the amplifier is more than one half cycle, so that both the transistors won’t be OFF at the same time.

This idea leads to the invention of class AB amplifier, which is the combination of both class A and class B amplifiers, as discussed below.

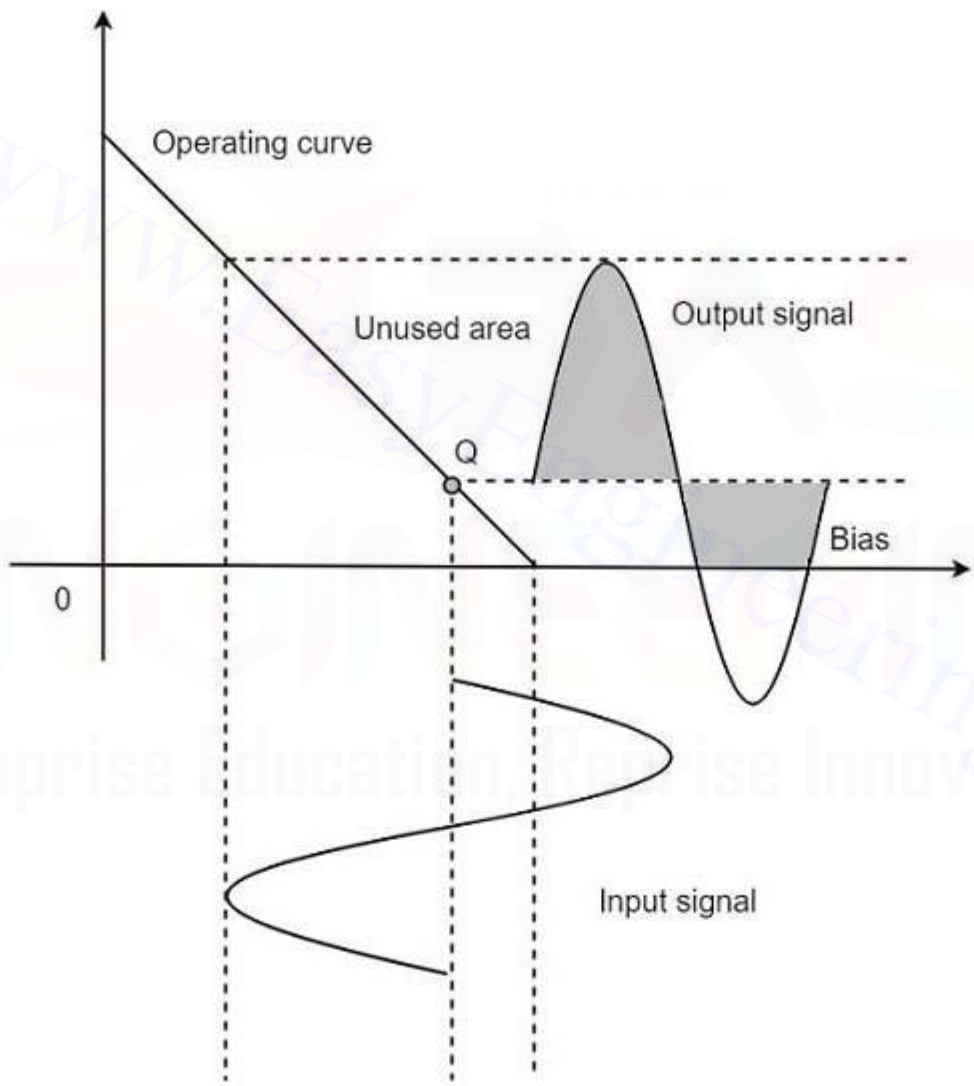
Class AB Power Amplifier

As the name implies, class AB is a combination of class A and class B type of amplifiers. As class A has the problem of low efficiency and class B has distortion problem, this class AB is emerged to eliminate these two problems, by utilizing the advantages of both the classes.

The cross over distortion is the problem that occurs when both the transistors are OFF at the same instant, during the transition period. In order to eliminate this, the condition has to be chosen for more than one half cycle. Hence, the other transistor gets into conduction, before the operating transistor switches to cut off state. This is achieved only by using class AB configuration, as shown in the following circuit diagram.



Therefore, in class AB amplifier design, each of the push-pull transistors is conducting for slightly more than the half cycle of conduction in class B, but much less than the full cycle of conduction of class A. The conduction angle of class AB amplifier is somewhere between 180^o to 360^o depending upon the operating point selected. This is understood with the help of below figure.



The small bias voltage given using diodes D_1 and D_2, as shown in the above figure, helps the operating point to be above the cutoff point. Hence the output waveform of class AB results as seen in the above figure. The crossover distortion created by class B is overcome by this class AB, as well the inefficiencies of class A and B don't affect the circuit.

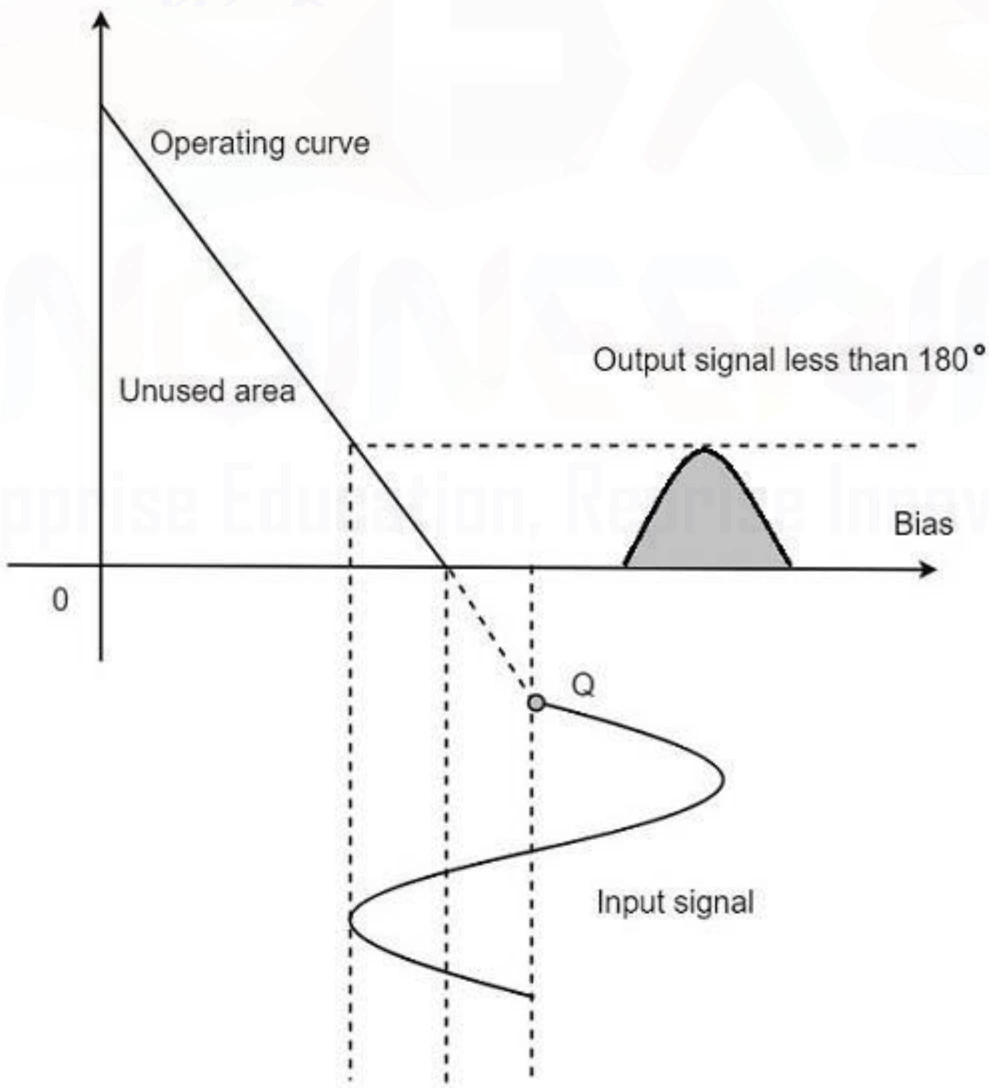
So, the class AB is a good compromise between class A and class B in terms of efficiency and linearity having the efficiency reaching about 50% to 60%. The class A, B and AB amplifiers are called as **linear amplifiers** because the output signal amplitude and phase are linearly related to the input signal amplitude and phase.

Class C Power Amplifier

When the collector current flows for less than half cycle of the input signal, the power amplifier is known as **class C power amplifier**.

The efficiency of class C amplifier is high while linearity is poor. The conduction angle for class C is less than 180^o. It is generally around 90^o, which means the transistor remains idle for more than half of the input signal. So, the output current will be delivered for less time compared to the application of input signal.

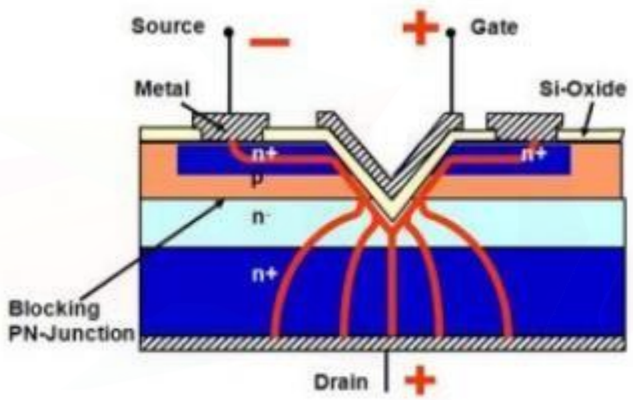
The following figure shows the operating point and output of a class C amplifier.



This kind of biasing gives a much improved efficiency of around 80% to the amplifier, but introduces heavy distortion in the output signal. Using the class C amplifier, the pulses produced at its output can be converted to complete sine wave of a particular frequency by using LC circuits in its collector circuit.

Power MOSFET

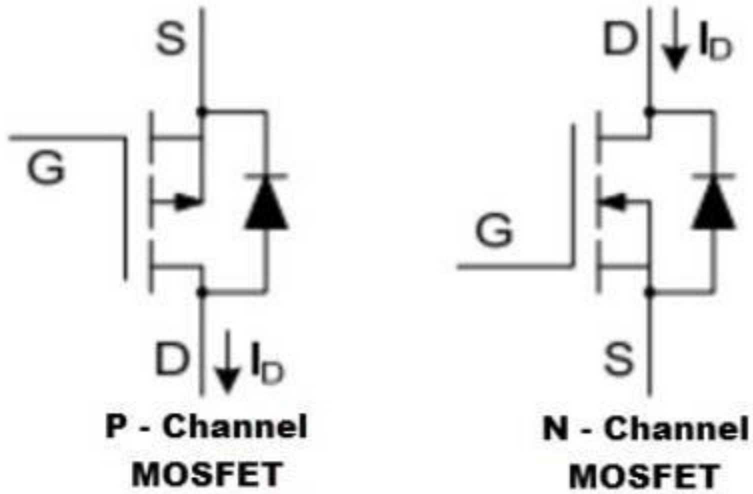
Working Principle and Applications



Power MOSFET

The Power MOSFET is a type of MOSFET. The operating principle of power MOSFET is similar to the general MOSFET. The power MOSFETS are very special to handle the high level of powers. It shows the high switching speed and by comparing with the normal MOSFET, the power MOSFET will work better. The power MOSFETs is widely used in the n-channel enhancement mode, p-channel enhancement mode, and in the nature of n-channel depletion mode. Here we have explained about the N-channel power MOSFET. The design of power MOSFET was made by using the CMOS technology and also used for development of manufacturing the integrated circuits in the 1970s.

A power MOSFET is a special type of metal oxide semiconductor field effect transistor. It is specially designed to handle high-level powers. The power MOSFET's are constructed in a V configuration. Therefore, it is also called as V-MOSFET, VFET. The symbols of N- channel & P- channel power MOSFET are shown in the below figure.



Power MOSFET

Basic Statures of Power MOSFET

There is three basic status in the power MOSFET which is following.

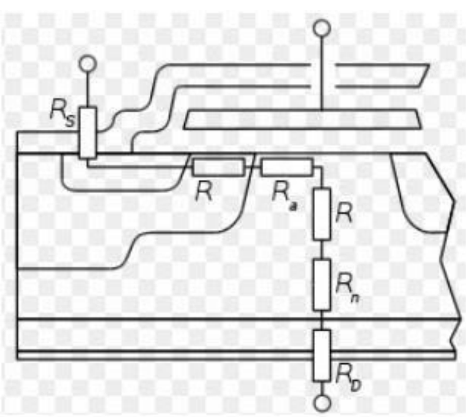
On sate resistance

Breakdown voltage

Body diode

On State Resistance

If the power MOSFET is in ON sate, then it produces the resistive behavior in-between the drain & source terminals. We can see in the following figure, that the resistance is the sum of many elementary contributions. The RS resistance is the source resistance. It will show all resistance between the source terminals of the package to the channel of the MOSFET.



On State Resistance

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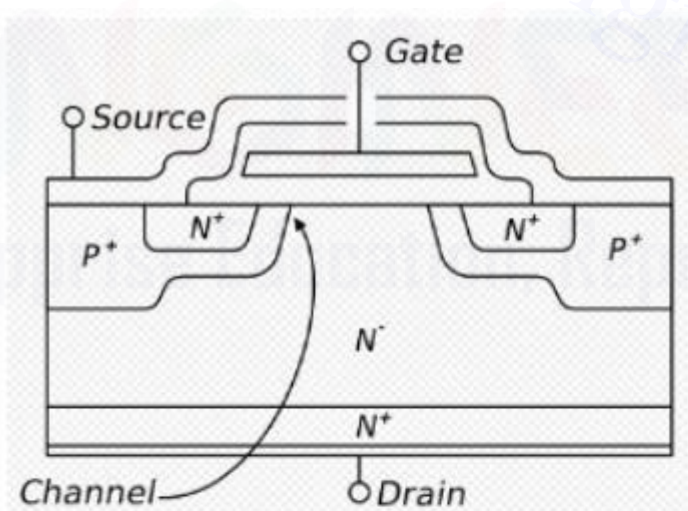
Body Diode

The body diode can be seen in the following figure that the source metallization is connected to both the N+ and P implantations. Even though the basic principle of the MOSFET requires only that the source should be connected to the N+ zone. Thus, this would result in a floating P zone between the N-doped source and drain. It is equivalent to an NPN transistor with a nonconnected base. Under some conditions like high drain current, in the order of the same volts of an on-state drain to source voltage, this parasitic transistor of NPN should be triggered and make the MOSFET uncontrollable.



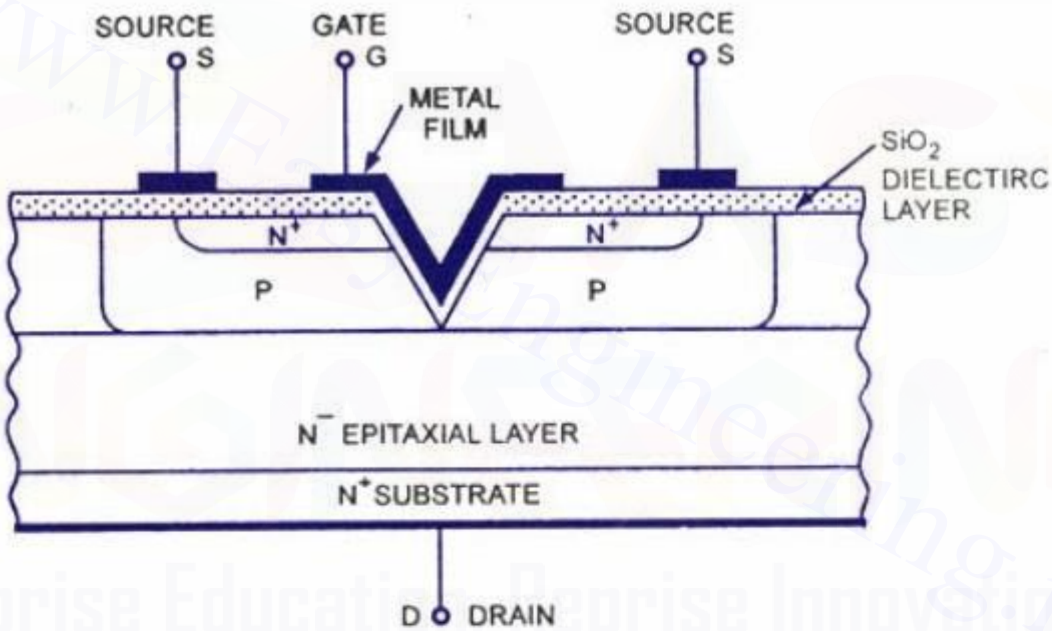
Body Diode

The connections of the P implantation to the source metallization short the base terminal of the transistor parasitic to its emitter and it prevents the latching. Hence this solution creates a diode between the cathode & anode of the MOSFET and the current blocks in one direction.

For inductive loads, the body diodes utilize the freewheeling diodes in the configuration of H Bridge & half bridge. Generally, these diodes will have a high forward voltage drop, the current is high. They are sufficient in many applications like reducing part count.

Working with Power MOSFET and Characteristics

The construction of the power MOSFET is in V-configurations, as we can see in the following figure. Thus the device is also called as the V-MOSFET or V-FET. The V- the shape of power MOSFET is cut to penetrate from the device surface is almost to the N+ substrate to the N+, P, and N – layers. The N+ layer is the heavily doped layer with a low resistive material and the N- layer is a lightly doped layer with the high resistance region.

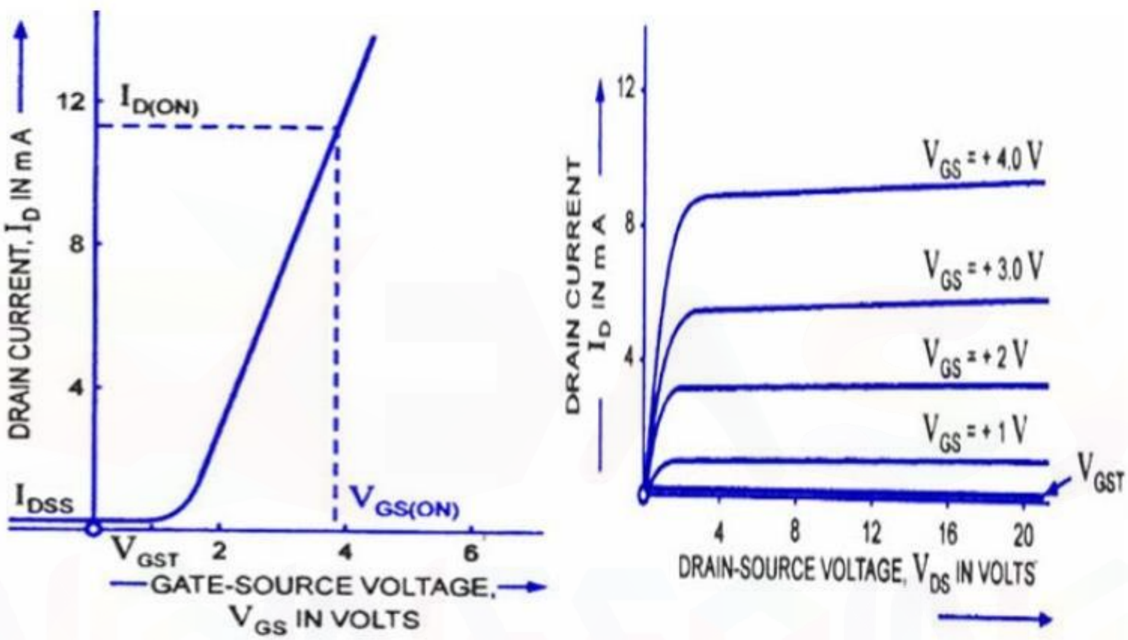


N – Channel Power MOSFET

Both the horizontal and the V cut surface are covered by the silicon dioxide dielectric layer and the insulated gate metal film is deposited on the SiO2 in the V shape. The source terminal contacts with the both N+ and P- layers through the SiO2 layer. The drain terminal of this device is N+.

The V-MOSFET is an E-mode FET and there is no exists of the channel in between the drain & source till the gate is positive with respect to the source. If we consider the gate is positive with respect to the source, then there is a formation of the N-type channel which is close to the gate and it is in the case of the E-MOSFET. In the case of E-MOSFET, the N-type channel provides the vertical path for the charge carriers. To flow between the drain and source terminals. If the VGS is zero or negative, then there is no channel of presence and the drain current is zero.

The following figures show the drain & transfer characteristics for the enhancement mode of N-channel power MOSFET is similar to the E-MOSFET. If there is an increase in the gate voltage then the channel resistance is reduced, therefore the drain current ID is increased. Hence the drain current ID is controlled by the gate voltage control. So that for a given level of VGS, ID is remaining constant through a wide range of VDS levels.



Transfer & Drain characteristics

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The channel length of the power MOSFET is in the diffusion process, but in the MOSFET the channel length is in the dimensions of the photographic masks employed in the diffusion process. By controlling the doping density and diffusion time, the channel length will become shorter. The shorter channels will give, the more current densities which will contribute again to larger power dissipation. It also allows a larger transconductance gm to be attained in the V-FET.

In the geometry of power MOSFET, there is an important factor which is the presence of lightly doped, N- epitaxial layer which is close to the N+ substrate. If the VGS is at zero or negative, then the drain is positive with respect to the source and there is a reverse biased between the P- layer & N- layer. At the junction the depletion region penetrates into the N- layer, therefore it punch-through the drain to the source are avoided. Hence, relatively high VDS are applied without any danger of device breakdown.

Applications of Power MOSFET

- The power MOSFET’s are used in the power supplies
- DC to DC converters
- Low voltage motor controllers
- These are widely used in the low voltage switches which are less than the 200V

DC-to-DC converter

Buck Converter

A **buck converter** (**step-down converter**) is a DC-to-DC power converter which steps down voltage (while stepping up current) from its input (supply) to its output (load). It is a class of switched-mode power supply (SMPS) typically containing at least two semiconductors (a diode and a transistor, although modern buck converters frequently replace the diode with a second transistor used for synchronous rectification) and at least one energy storage element, a capacitor, inductor, or the two in combination. To reduce voltage ripple, filters made of capacitors (sometimes in combination with inductors) are normally added to such a converter's output (load-side filter) and input (supply-side filter).[1]

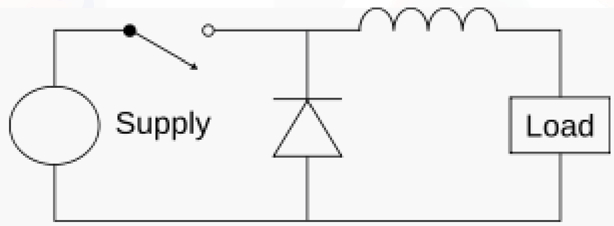
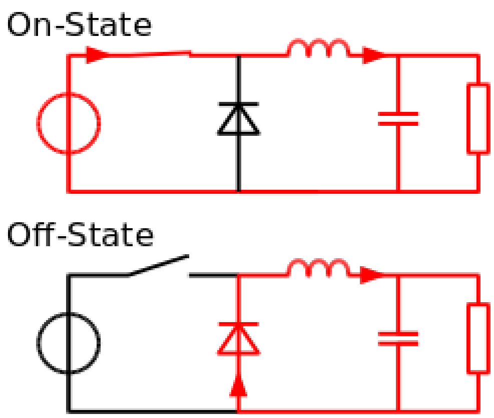


Fig. : Buck converter circuit diagram.

Switching converters (such as buck converters) provide much greater power efficiency as DC-to-DC converters than linear regulators, which are simpler circuits that lower voltages by dissipating power as heat, but do not step up output current.[2]

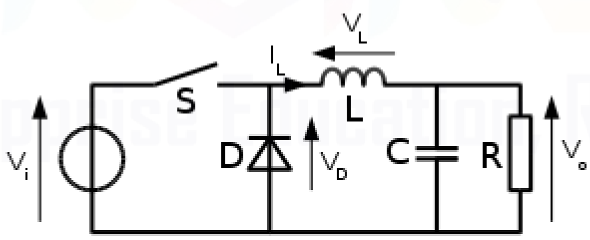
Buck converters can be highly efficient (often higher than 90%), making them useful for tasks such as converting a computer's main (bulk) supply voltage (often 12 V) down to lower voltages needed by USB, DRAM and the CPU (1.8 V or less).

The basic operation of the buck converter has the current in an inductorcontrolled by two switches (usually a transistor and a diode). In the idealised converter, all the components are considered to be perfect. Specifically, the switch and the diode have zero voltage drop when on and zero current flow when off, and the inductor has zero series resistance. Further, it is assumed that the input and output voltages do not change over the course of a cycle (this would imply the output capacitance as being infinite).



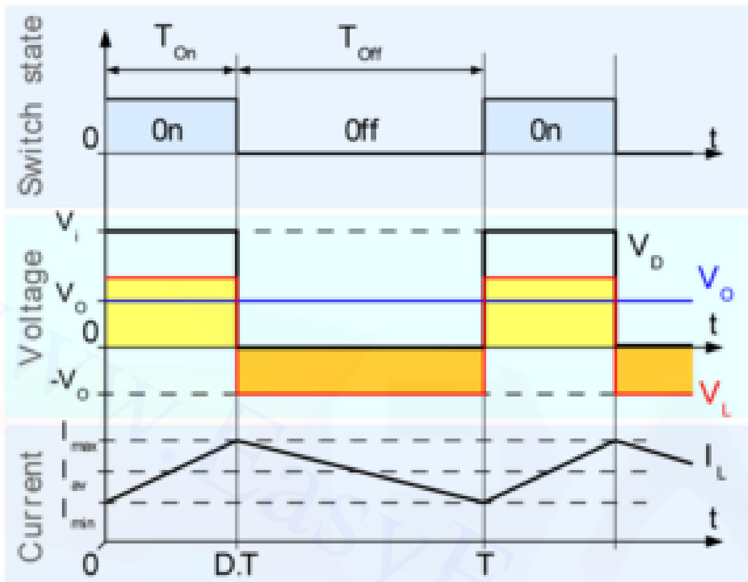
The conceptual model of the buck converter is best understood in terms of the relation between current and voltage of the inductor. Beginning with the switch open (off-state), the current in the circuit is zero. When the switch is first closed (on-state), the current will begin to increase, and the inductor will produce an opposing voltage across its terminals in response to the changing current. This voltage drop counteracts the voltage of the source and therefore reduces the net voltage across the load. Over time, the rate of change of current decreases, and the voltage across the inductor also then decreases, increasing the voltage at the load.

During this time, the inductor stores energy in the form of a magnetic field. If the switch is opened while the current is still changing, then there will always be a voltage drop across the inductor, so the net voltage at the load will always be less than the input voltage source. When the switch is opened again (off-state), the voltage source will be removed from the circuit, and the current will decrease.



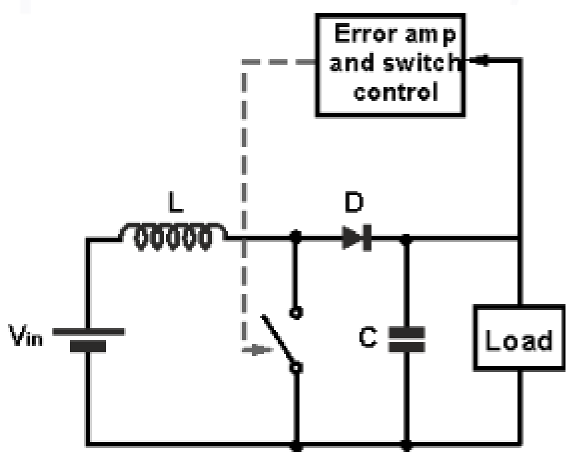
The decreasing current will produce a voltage drop across the inductor (opposite to the drop at on-state), and now the inductor becomes a Current Source. The stored energy in the inductor's magnetic field supports the current flow through the load. This current, flowing while the input voltage source is disconnected, when concatenated with the current flowing during on-state, totals to current greater than the average input current (being zero during off-state).

The "increase" in average current makes up for the reduction in voltage, and ideally preserves the power provided to the load. During the off-state, the inductor is discharging its stored energy into the rest of the circuit. If the switch is closed again before the inductor fully discharges (on-state), the voltage at the load will always be greater than zero.



Boost Converter

The boost converter circuit has many similarities to the buck converter. However the circuit topology for the boost converter is slightly different. The fundamental circuit for a boost converter or step up converter consists of an inductor, diode, capacitor, switch and error amplifier with switch control circuitry.



The circuit for the step-up boost converter operates by varying the amount of time in which inductor receives energy from the source.

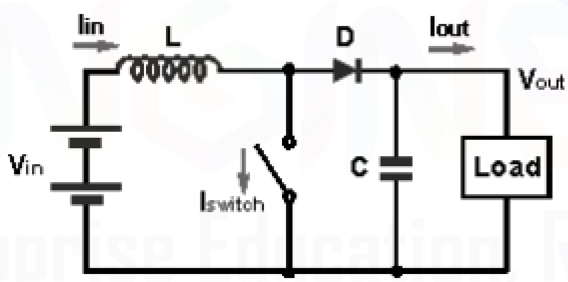
In the basic block diagram the operation of the boost converter can be seen that the output voltage appearing across the load is sensed by the sense / error amplifier and an error voltage is generated that controls the switch.

Typically the boost converter switch is controlled by a pulse width modulator, the switch remaining on of longer as more current is drawn by the load and the voltage tends to drop and often there is a fixed frequency oscillator to drive the switching.

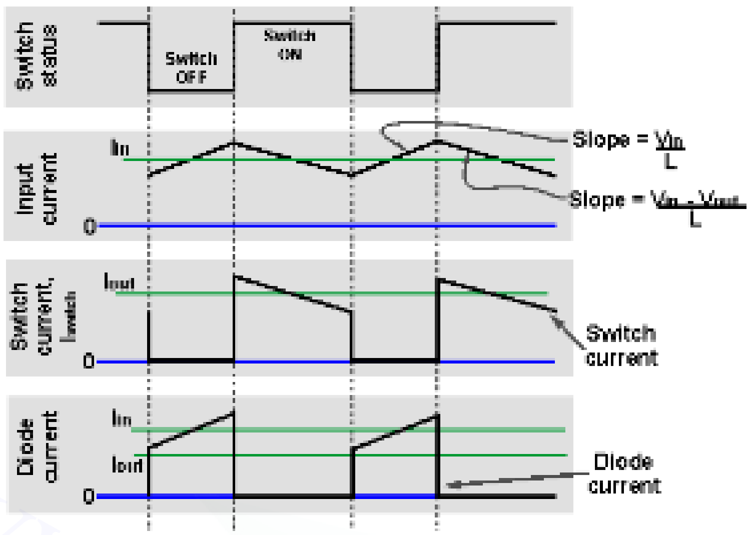
Boost converter operation

The operation of the boost converter is relatively straightforward.When the switch is in the ON position, the inductor output is connected to ground and the voltage Vin is placed across it. The inductor current increases at a rate equal to Vin/L.

When the switch is placed in the OFF position, the voltage across the inductor changes and is equal to Vout-Vin. Current that was flowing in the inductor decays at a rate equal to (Vout-Vin)/L.



Referring to the boost converter circuit diagram, the current waveforms for the different areas of the circuit can be seen as below.



It can be seen from the waveform diagrams that the input current to the boost converter is higher than the output current. Assuming a perfectly efficient, i.e. lossless, boost converter, the power out must equal the power in, i.e. Vin · Iin = Vout · Iout. From this it can be seen if the output voltage is higher than the input voltage, then the input current must be higher than the output current.

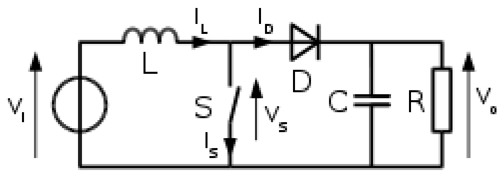
In reality no boost converter will be lossless, but efficiency levels of around 85% and more are achievable in most supplies.

The key principle that drives the boost converter is the tendency of an inductor to resist changes in current by creating and destroying a magnetic field. In a boost converter, the output voltage is always higher than the input voltage. A schematic of a boost power stage is shown in Figure

When the switch is closed, current flows through the inductor in clockwise direction and the inductor stores some energy by generating a magnetic field. Polarity of the left side of the inductor is positive.

When the switch is opened, current will be reduced as the impedance is higher. The magnetic field previously created will be destroyed to maintain the current towards the load. Thus the polarity will be reversed (means left side of inductor will be negative now). As a result, two sources will be in series causing a higher voltage to charge the capacitor through the diode D.

If the switch is cycled fast enough, the inductor will not discharge fully in between charging stages, and the load will always see a voltage greater than that of the input source alone when the switch is opened. Also while the switch is opened, the capacitor in parallel with the load is charged to this combined voltage. When the switch is then closed and the right hand side is shorted out from the left hand side, the capacitor is therefore able to provide the voltage and energy to the load. During this time, the blocking diode prevents the capacitor from discharging through the switch. The switch must of course be opened again fast enough to prevent the capacitor from discharging too much.



The basic principle of a Boost converter consists of 2 distinct states (see figure 2):

in the On-state, the switch S (see figure 1) is closed, resulting in an increase in the inductor current;in the Off-state, the switch is open and the only path offered to inductor current is through the flyback diode D, the capacitor C and the load R. This results in transferring the energy accumulated during the On-state into the capacitor.

The input current is the same as the inductor current as can be seen in figure 2. So it is not discontinuous as in the buck converterand the requirements on the input filter are relaxed compared to a buck converter.The buck boost converter is a DC to DC converter. The output voltage of the DC to DC converter is less than or greater than the input voltage. The output voltage of the magnitude depends on the duty cycle. These converters are also known as the step up and step down transformers and these names are coming from the analogous step up and step down transformer. The input voltages are step up/down to some level of more than or less than the input voltage. By using the low conversion energy, the input power is equal to the output power. The following expression shows the low of a conversion.

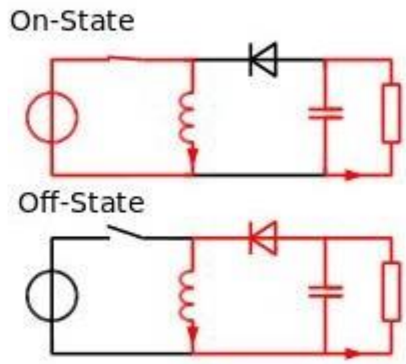
Input power (Pin) = Output power (Pout)

For the step up mode, the input voltage is less than the output voltage (Vin < Vout). It shows that the output current is less than the input current. Hence the buck booster is a step up mode. Vin < Vout and Iin > Iout

In the step down mode the input voltage is greater than the output voltage (Vin > Vout). It follows that the output current is greater the input current. Hence the buck boost converter is a step down mode.

Vin > Vout and Iin < Iout

It is a type of DC to DC converter and it has a magnitude of output voltage. It may be more or less than equal to the input voltage magnitude. The buck boost converter is equal to the fly back circuit and single inductor is used in the place of the transformer. There are two types of converters in the buck boost converter that are buck converter and the other one is boost converter. These converters can produce the range of output voltage than the input voltage. The following diagram shows the basic buck boost converter.



Buck Boost Converter

Working principle of Buck Boost Converter

The working operation of the DC to DC converter is the inductor in the input resistance has the unexpected variation in the input current. If the switch is ON then the inductor feed the energy from the input and it stores the energy of magnetic energy. If the switch is closed it discharges the energy. The output circuit of the capacitor is assumed as high sufficient than the time constant of an RC circuit is high on the output stage. The huge time constant is compared with the switching period and make sure that the steady state is a constant output voltage Vo(t) = Vo(constant) and present at the load terminal.

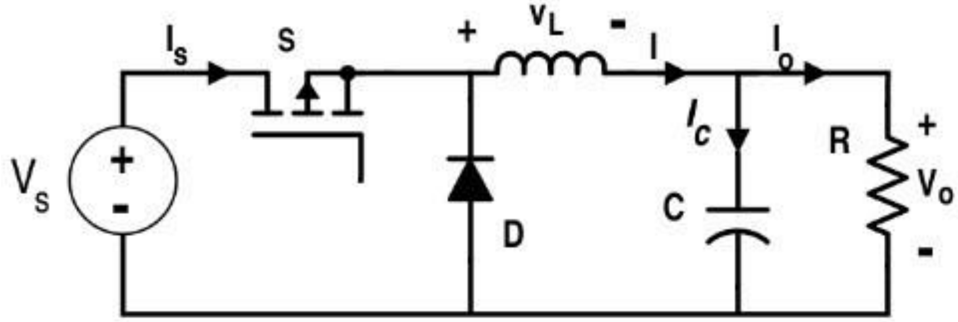
There are two different types of working principles in the buck boost converter.

Buck converter.
Boost converter.
Buck Converter Working

The following diagram shows the working operation of the buck converter.

In the buck converter first transistor is turned ON and second transistor is switched OFF due to high square wave frequency. If the gate terminal of the first transistor is more than the current pass through the magnetic field, charging C, and it supplies the load. The D1 is the Schottky diode and it is turned OFF due to the positive voltage to the cathode.

The inductor L is the initial source of current. If the first transistor is OFF by using the control unit then the current flow in the buck operation. The magnetic field of the inductor is collapsed and the back e.m.f is generated collapsing field turn around the polarity of the voltage across the inductor. The current flows in the diode D2, the load and the D1 diode will be turned ON



Buck Converter Working

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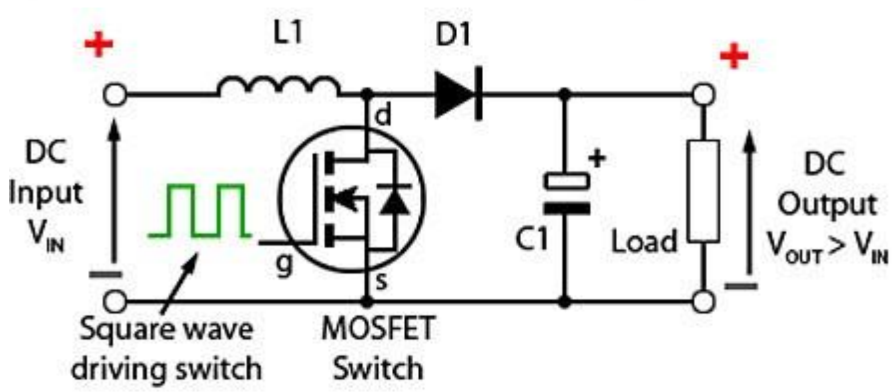
The discharge of the inductor L decreases with the help of the current. During the first transistor is in one state the charge of the accumulator in the capacitor. The current flows through the load and during the off period keeping Vout reasonably. Hence it keeps the minimum ripple amplitude and Vout closes to the value of Vs

Boost Converter Working

In this converter the first transistor is switched ON continually and for the second transistor the square wave of high frequency is applied to the gate terminal. The second transistor is in conducting when the on state and the input current flow from the inductor L through the second transistor. The negative terminal charging up the magnetic field around the inductor. The D2 diode cannot conduct because the anode is on the potential ground by highly onducting the second transistor.



Boost Converter Working

By charging the capacitor C the load is applied to the entire circuit in the ON State and it can construct earlier oscillator cycles. During the ON period the capacitor C can discharge regularly and the amount of high ripple frequency on the output voltage. The approximate potential difference is given by the equation below.

VS + VL

During the OFF period of second transistor the inductor L is charged and the capacitor C is discharged. The inductor L can produce the back e.m.f and the values are depending up on the rate of change of current of the second transistor switch. The amount of inductance the coil can occupy. Hence the back e.m.f can produce any different voltage through a wide range and determined by the design of the circuit. Hence the polarity of voltage across the inductor L has reversed now.

The input voltage gives the output voltage and atleast equal to or higher than the input voltage. The diode D2 is in forward biased and the current applied to the load current and it recharges the capacitors to VS + VL and it is ready for the second transistor.

Modes Of Buck Boost Converters

There are two different types of modes in the buck boost converter. The following are the two different types of buck boost converters.

Continuous conduction mode.

Discontinuous conduction mode.

Continuous Conduction Mode

In the continuous conduction mode the current from end to end of inductor never goes to zero. Hence the inductor partially discharges earlier than the switching cycle.

Discontinuous Conduction Mode

In this mode the current through the inductor goes to zero. Hence the inductor will totally discharge at the end of switching cycles.

Applications of Buck boost converter

- It is used in the self regulating power supplies.
- It has consumer electronics.
- It is used in the Battery power systems.
- Adaptive control applications.

- Power amplifier applications.

Advantages of Buck Boost Converter

- It gives higher output voltage.
- Low operating duct cycle.
- Low voltage on MOSFETs