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 and shunt-series feedback amplifiers - stablity problem - Gain and phase margins - Frequency compensation.

Introduction

A practical amplifier has a gain of nearly one million ie its output is one million times the i/p. Consequently, even a casual disturbance at the i/p 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.

∴ 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 a small a level as possible

The noise level in amplifier can be reduced considerably by the use of negative feedback.

Feedback.

The process of injecting a fraction of output energy of some device back to the input is known as Feedback.

Depending upon whether the feedback energy aids or opposes the input signal, there are two basic types of feedback in amplifiers

① positive feedback

② Negative feedback.

<u>positive Feedback.</u>

If the feedback signal (voltage or current) is applied in such a way that it is in phase with the input signal and thus increases it, then it is called a positive feedback. It is also known as regenerative feedback or direct feedback.

The positive feedback increases the gain of amplifier. However it produces excessive distortion due to which it is seldom used in amplifiers. The positive feedback is used in oscillators.

<u>Negative Feedback.</u>

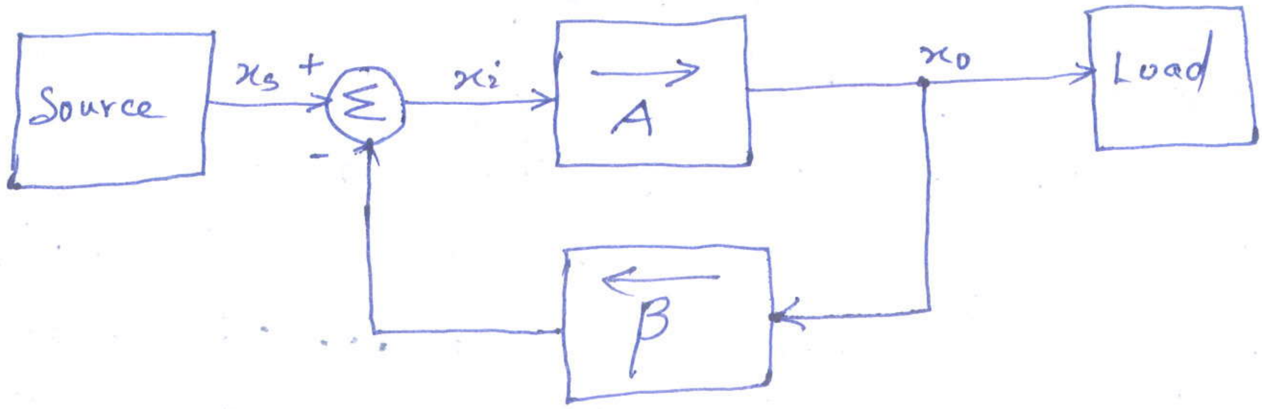
If the feedback signal (i.e, voltage or current) is applied in such away that it is out of phase with the input signal and thus decreases it, then it is called a negative feedback. Sometimes, it is also called as degenerative feedback or inverse feedback.

The negative feedback reduces gain of the amplifier. However, it improves the amplifier performance in many other respects. Thus a negative feedback is frequently used in small-signal as well as with large-signal amplifier circuits.

In amplifier design, negative feedback is applied to effect one or more of the following properties:

1. Desensitize the gain: ie, make the value of the gain less sensitive to variations in the value of circuit components, such as might be caused by changes in temperature.
2. Reduce nonlinear distortion: Make the o/p proportional to the input.
3. Reduce the effect of noise: Minimize the contribution to the output of unwanted electric signals generated, either by the circuit components themselves, or by extraneous interference.
4. Control the input and output impedances: ie, raise or lower the i/p and o/p impedances by the selection of an appropriate feedback topology.
5. Extend the bandwidth of the amplifier.

The General Feedback structure



The above figure shows the basic structure of a feedback amplifier. Rather than showing voltages and current the fig is a signal-flow diagram, where each of the quantities x can represent either voltage or a current signal.

The open-loop amplifier has a gain A; thus its output x_o is related to the input x_i by

$$x_o = Ax_i \qquad \text{——— (1)}$$

The output x_o is fed to the load as well as to a feedback network, which produces a sample of the output. The sample x_f is related to x_o by the feedback factor β.

$$x_f = \beta x_o \qquad \text{——— (2)}$$

The feedback signal x_f is subtracted from the source signal x_s, which is the input to the complete feedback amplifier, to produce the signal x_i, which is the input to the basic amplifier,

$$x_i = x_s - x_f \quad \text{———} \quad (3)$$

Here, the subtraction that makes the feedback negative. Negative feedback reduces the signal that appears at the input of the basic amplifier.

The gain of the feedback amplifier can be obtained by combining eq (1) & (3)

$$A_f \equiv \frac{x_o}{x_s} = \frac{A}{1+A\beta} \quad \text{———} \quad (4)$$

The quantity $A\beta$ is called the loop gain.

For the feedback to be negative, the loop gain $A\beta$ should be positive; ie, the feedback signal x_f should have the same sign as x_s, thus resulting in a smaller difference signal x_i.

Eq (4) indicates that for positive $A\beta$ the gain-with-feedback A_f will be smaller then the open loop gain A by the quantity $1+A\beta$, which is called the amount of feedback.

If the loop gain $A\beta$ is large, $A\beta \gg 1$, then from equ (4) it follows that

$$A_f \simeq 1/\beta$$

The gain of the feedback amplifier is almost entirely determined by the feedback network.

Properties of negative feedback.

① Gain Desensitivity

The transfer gain of the amplifier is not constant as it depends on the factors such as operating point, temperature, etc. This lack of stability in amplifiers can be reduced by introducing negative feedback.

WKt,

equ ④ $\Rightarrow$ $$A_f = \frac{A}{1+\beta A}$$

Assuming β is constant & Taking differentials of both sides w.r.t. A we get,

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

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

$$\boxed{dA_f = \frac{dA}{(1+\beta A)^2}} \quad \text{———} \quad (*)$$

Dividing equ (*) by A_f, we get

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

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

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

$$\left|\frac{dA_f}{A_f}\right| = \left|\frac{dA}{A}\right| \frac{1}{|(1+\beta A)|} \quad \text{———— (i)}$$

Where

$\frac{dA_f}{A_f}$ = Fractional change in amplification with feedback.

$\frac{dA}{A}$ = Fractional change in amplification without feedback.

From equ (i) it may be noted that change in the gain with feedback is less than the change in the gain with the feedback ~~wtd~~ by factor $(1+\beta A)$.

The fractional change in amplification with feedback divided by the fractional change without feedback is called the sensitivity of the transfer gain. Hence the sensitivity is $\frac{1}{(1+\beta A)}$. The reciprocal of the sensitivity is called the desensitivity D. It is given as.

$$\boxed{D = 1+\beta A}$$

∴ stability of the amplifier increases with increase in desensitivity.

Bandwidth Extension

WKT, the gain with feedback for an amplifier is given by

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

using the above equation, we can write

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

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

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

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

Lower cut-off frequency : WKT, the relation between gain at lower cut-off frequency and gain at mid frequency for an amplifier is given as

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

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

Substituting A_{low} in the $A_{f\,low}$ equation we get.

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

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

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

$\div$ Num. & deno. by $(1 + A_{mid} \cdot \beta)$ we have

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

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

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

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

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

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

From the above equation, we can say that lower cut-off frequency with feedback is less than the lower cut off frequency without feedback by factor $(1+A_{mid}\beta)$. Therefore, by introducing negative feedback, low frequency response of the amplifier is improved.

$|||^{ly}$ upper cut off frequency.

WKT, the relation between gain at upper cut-off frequency and gain at mid frequency for an amplifier is given as

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

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

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

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

$\div$ Num. and den. by $(1+A_{mid}\beta)$, we get

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

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

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

Where the upper cutoff frequency with feedback is given as

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

From the above equation, we can say that upper cut-off frequency with feedback is greater than upper cut-off frequency with out feedback by factor $(1+A_{mid}\beta)$.

∴ By introducing negative feedback, high frequency response of the amplifier is improved.

The bandwidth of the amplifier without feedback is given as

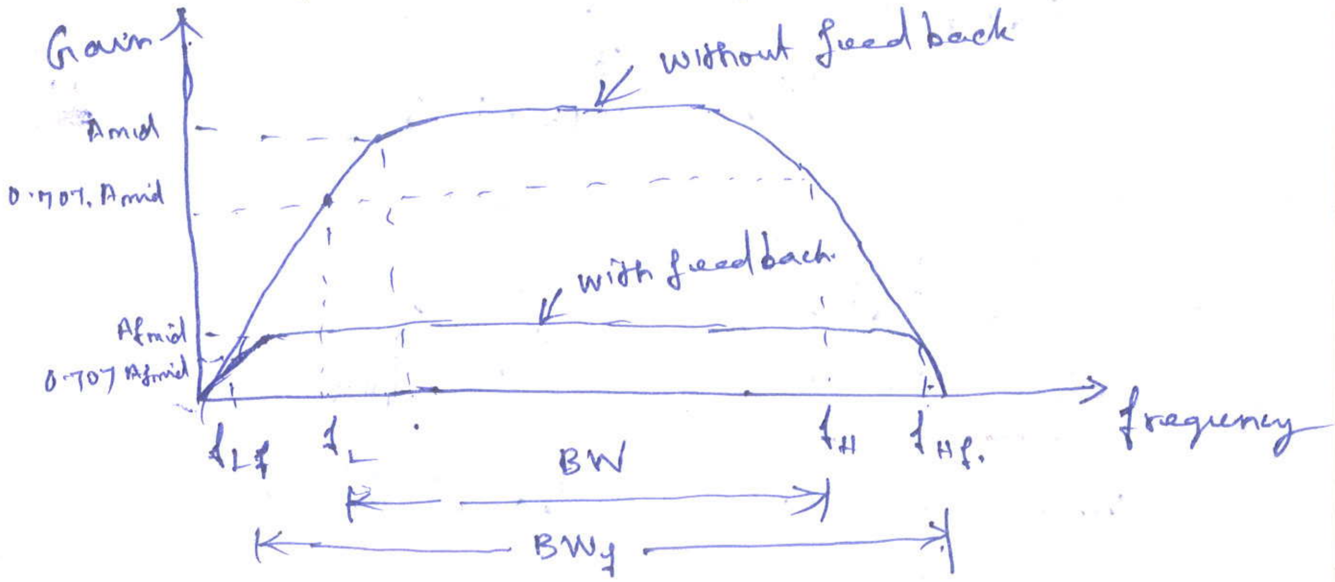
$$BW = f_H - f_L$$

BW of the amplifier with feedback can be written as

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

or, it can also be written as

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



From the freq. response shown above, it is clear that $(f_{Hf} - f_{Lf}) > (f_H - f_L)$ and hence the bandwidth of the amplifier is greater than the bandwith of the amplifier without feedback.

As the voltage gain of the feedback amp. reduces by the factor $(1 + A\beta)$ its BW increases by $(1 + A\beta)$. This shows that product of voltage gain of BW of an amplifier without feedback and with feedback remains the same.

i.e. $A_f \times BW_f = A \times BW$

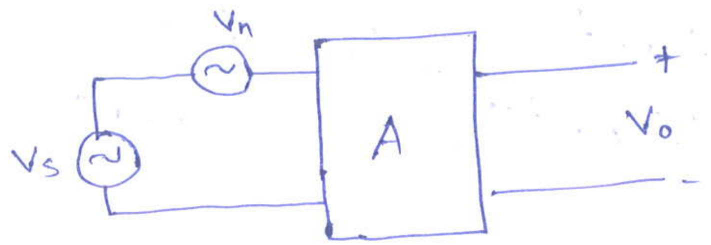
Reduction in Noise

Almost all amplifier circuit produce noise due to presence of active and passive components in it. During amplification process this noise is also amplified along with the signal. During amplification process this noise is also amplified along with the signal.

The negative feedback can be used to reduce in amplifiers. Thus negative feedback can play an important role in improving signal to noise ratio (SNR) of an amplifier.

It is possible to improve signal to noise ratio of an amplifier under certain conditions.

Let us consider, the model of noisy amplifier as shown below



Here, a noisy amplifier is modeled by a noiseless amplifier in series with noise source V_n,

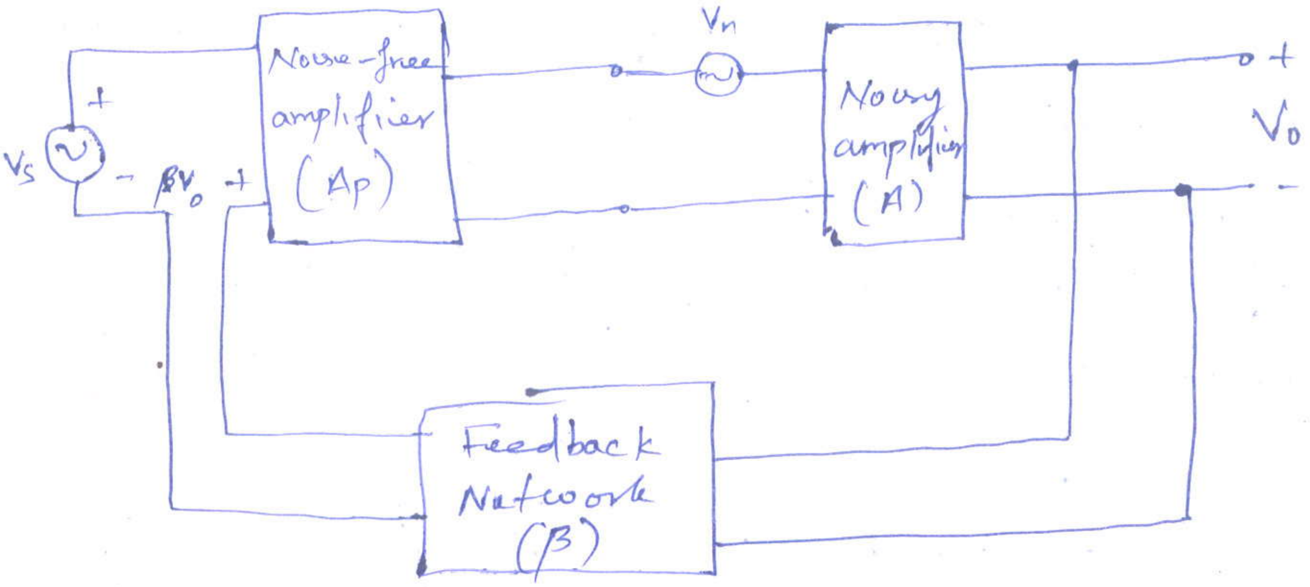
The output voltage of this noisy amplifier is given by

$$V_o = A(V_s + V_n)$$

and the signal to noise ratio is given by

$$S/N = V_s / V_n$$

It is possible to improve signal to noise ratio of this noisy amplifier by preceding a noise-free preamplifier as shown below.



Using superposition theorem, the output voltage of the feedback amplifier shown above can be given as

$$V_0 = V_s \cdot \frac{A_p \cdot A}{1 + A_p \cdot A \cdot \beta} + V_n \cdot \frac{A}{1 + A_p A \beta}$$

$$= \frac{A}{1 + A_p A\beta} \left(V_s A_p + V_n \right)$$

The signal to noise ratio (SNR) at the output is given as

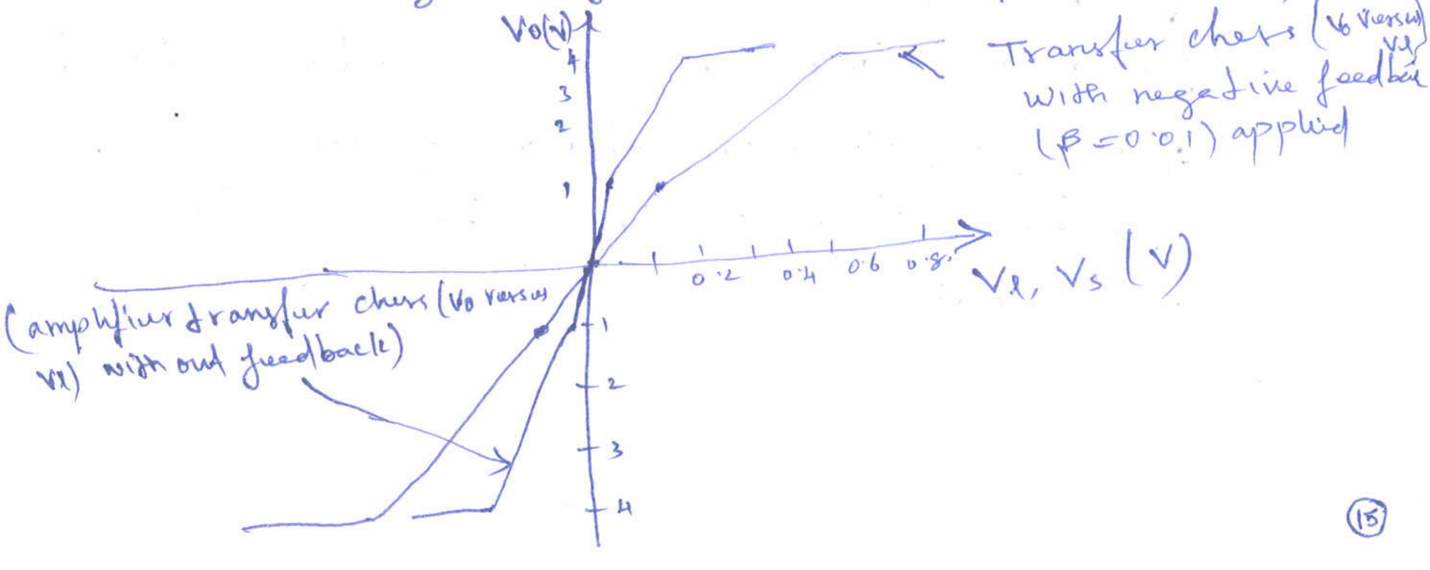
$$S/N = \frac{V_s \cdot A_p}{V_n}$$

The above equation shows that we can improve SNR of a noisy amplifier by factor of A_p if a noise free preamplifier with voltage gain A_p precedes a noisy amplifier. Fortunately, we can meet this situation in most of the practical amplifiers. For example power amplifier stage suffers from hum noise due to large currents. Removing this humming noise by filtering is not economical. In such situations we can considerably improve SNR by preceding a small-signal amplifier with large voltage gain.

- Reduction in Nonlinear Distortion

Non-linear distortion occurs when an active-device (transistor) in the amplifier has non-linear transfer characteristics. In case of non-linear distortion, additional frequency components at multiple of fundamental frequency are present in the output. Thus, such distortion is also known as harmonic distortion.

The negative feedback reduces this nonlinear distortion by the factor $D = 1 + A\beta$.



Basic Feedback Topologies

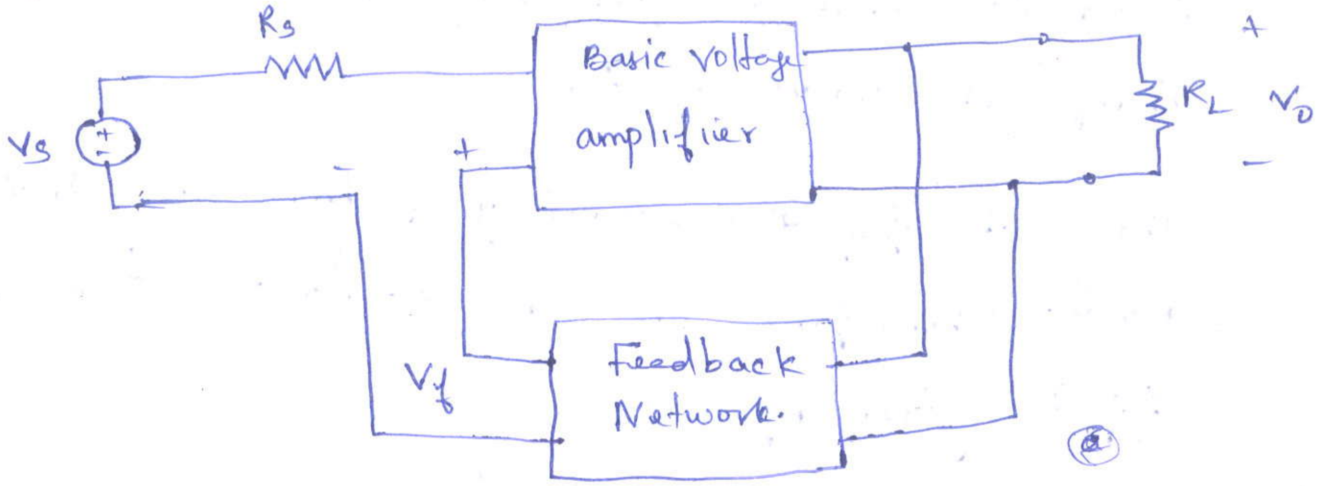
Based on the quantity to be amplified (voltage or current) and on the desired form of output (voltage or current), amplifiers can be classified into four categories.

(i) voltage Amplifiers

Voltage amplifiers are intended to amplify an input voltage signal and provide an output voltage signal. The voltage amplifier is essentially a voltage-controlled voltage source. The input resistance is required to be high and the output resistance is required to be low. Since the signal source is essentially a voltage source, it is convenient to represent it in terms of a Thevenin equivalent circuit.

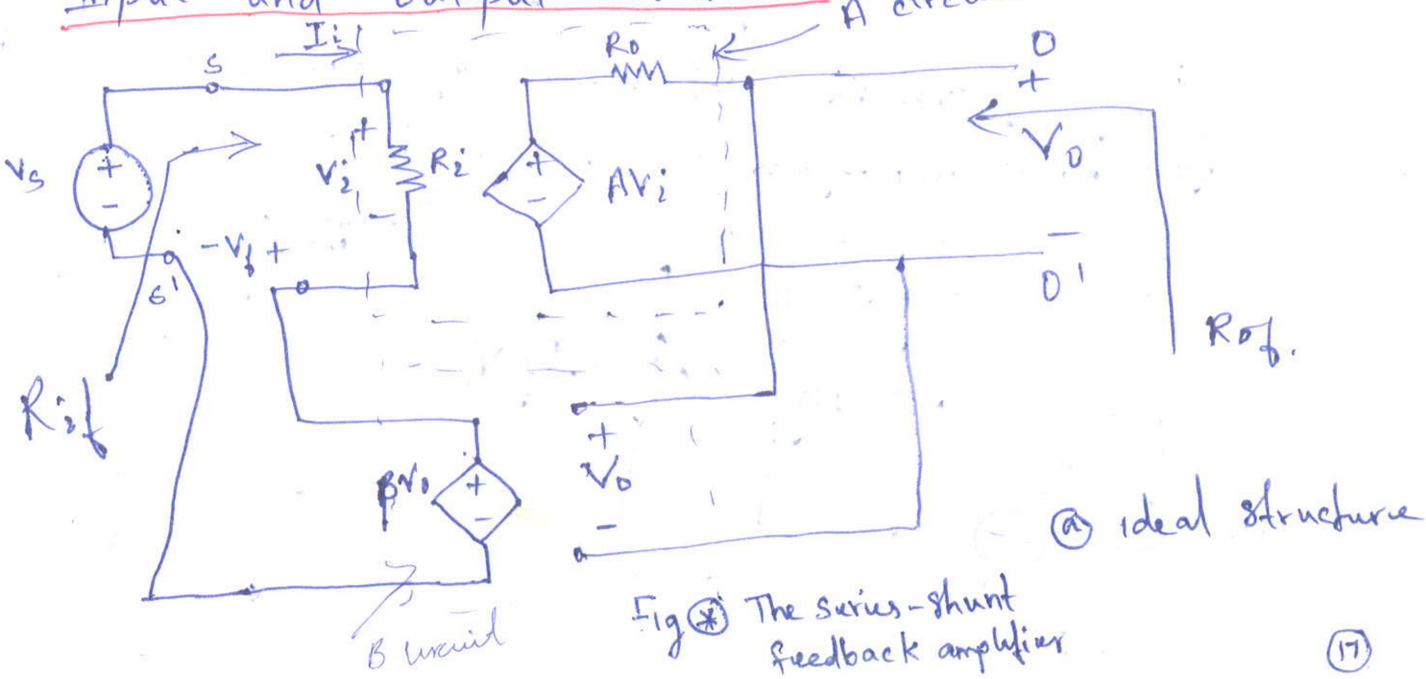
In a voltage amplifier the output Quantity of interest is the output voltage. It follows that the feedback network should sample the output voltage, just as a voltmeter measure a voltage. Also, because of the Thevenin representation of the source, the feedback signal x_f should be a voltage that can be mixed with the source voltage in series.

The most suitable feedback topology for the voltage amplifier is the voltage-mixing voltage sampling as shown below. Because of the series connection at the input and the parallel or shunt connection at the output, this feedback topology is also known as series-shunt feedback. This topology not only stabilizes the voltage gain but also results in a higher input resistance and a lower o/p resistance, which are the desirable properties for a voltage amplifier.



(a)

Input and output resistance



(a) ideal structure

Fig (*) The series-shunt feedback amplifier

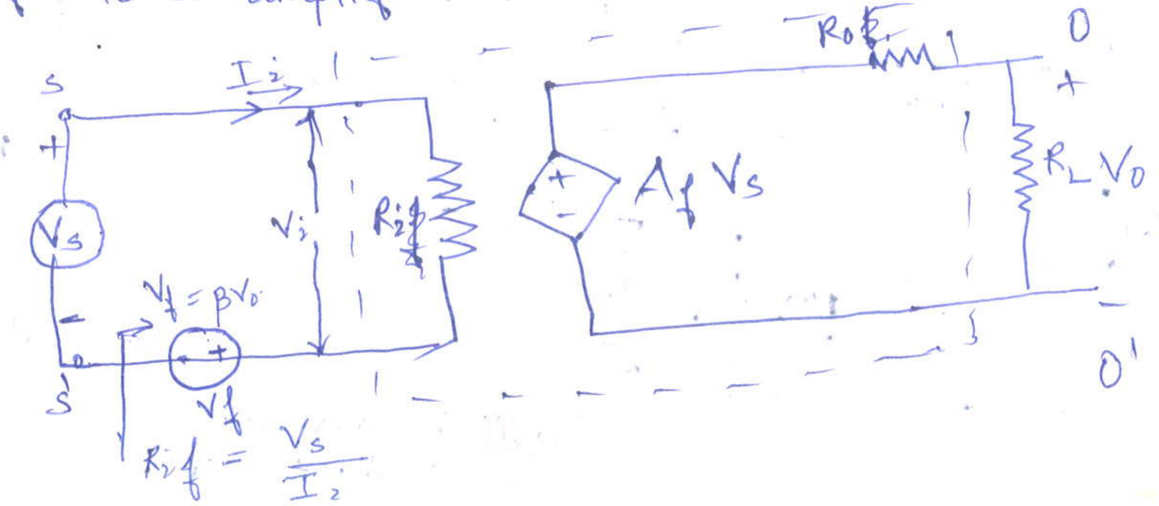
The series-shunt is the appropriate feedback topology for a voltage amplifier. The ideal structure of the series-shunt feedback is shown in fig. (a). It consists of unilateral open-loop amplifier (the A ckt) and an ideal voltage sampling, voltage-mixing feedback network (the β ckt). The A ckt has an input resistance R_i, an open-circuit voltage gain A, and an output resistance R_o. It is assumed that the source and load resistance have been absorbed inside the A circuit.

Furthermore, note that the β circuit does not load the A circuit; that is, connecting the β circuit does not change the value of A.

The ckt fig. a exactly follows the ideal feedback model. Therefore the closed-loop voltage gain A_f is given by

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

The equivalent ckt model of the series-shunt feedback amplifier is shown below.



A_f = open-ckt voltage gain of the feedback amplifier

R_{if} = input resistance

R_{of} = o/p resistance.

Expressions for R_{if} & R_{of}

The input resistance with feedback is given as

$$R_{if} = \frac{V_s}{I_i} \quad \text{——— (1)}$$

Applying KVL to the input side we get

$$V_s - I_i R_i - V_f = 0$$

$$\boxed{V_s = I_i R_i + V_f = I_i R_i + \beta V_o} \quad \because V_f = \beta V_o \quad \text{— (2)}$$

obtaining expression for V_o in terms of I_i

The o/p voltage V_o is given as

$$\boxed{V_o = A_d I_i R_i} \quad \text{——— (3)}$$

~~The~~ sub. the value of V_o from equ (3) in equ (2) we get.

$$V_s = I_i R_i + \beta A_f I_i R_i$$

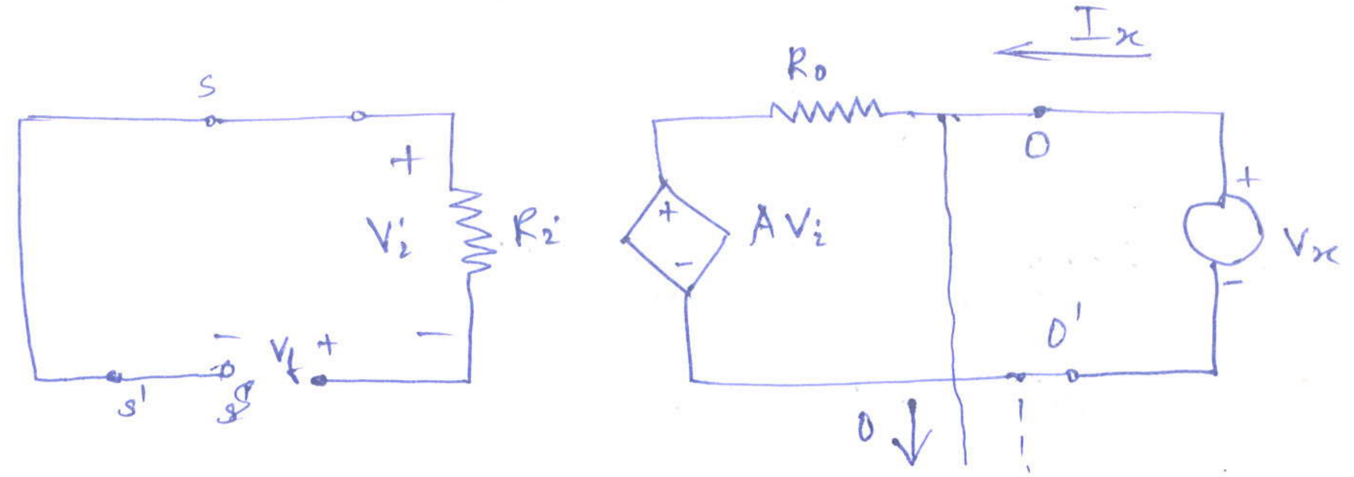
$$\therefore R_{if} = \frac{V_s}{I_i} = R_i + \beta A . R_i$$

$$\boxed{R_{if} = R_i (1 + \beta A)} \quad \text{——— (4)}$$

R_i - i/p resistance without feedback.

Thus, as expected, the series-mixing feedback results in an increase in the amplifier input resistance by a factor equal to the amount of feedback, $(1+A\beta)$, a highly desirable property for a voltage amplifier.

Output resistance of the feedback amplifier.



To determine the o/p resistance R_{of} of the feed back amplifier, we set $V_s = 0$ and apply a test voltage V_x between the output terminals. If the current drawn from V_x is I_x, the o/p resistance R_{of} is

$$R_{of} \equiv \frac{V_x}{I_x}.$$

To find I_x, Apply KVL to the o/p side

$$A V_i + I_x R_0 - V_x = 0$$

$$I = \frac{V_x - A V_i}{R_0} \quad \text{——— (1)}$$

The input voltage is given as

$$V_i = -V_f = -\beta V \qquad \therefore V_s = 0 \quad \text{——— (3)}$$

Sub. the V_i from equⁿ ⑧ in equⁿ ① we get

$$I_x = \frac{V_x + A\beta V_x}{R_0} = \frac{V_x(1+\beta A)}{R_0}.$$

To obtain expression for R_{of}.

$$R_{of} = \frac{V_x}{I_x}$$

$$\boxed{R_{of} = \frac{R_0}{1+\beta A}}$$

Thus, as expected, the shunt sampling (or voltage sampling) at the output results in a decrease in the amplifier output resistance by a factor equal to the amount of negative feedback, $(1+A\beta)$, a highly desirable property for a voltage amplifier.

The practical case

Analysis of practical voltage series Feedback amplifier ckt.

Fig ① shows a practical non-inverting op-amp circuit with feedback. Here, the op-amp is in the forward path while the feedback circuit consists of the resistance R_1 and R_2. The input is applied to the noninverting terminal.

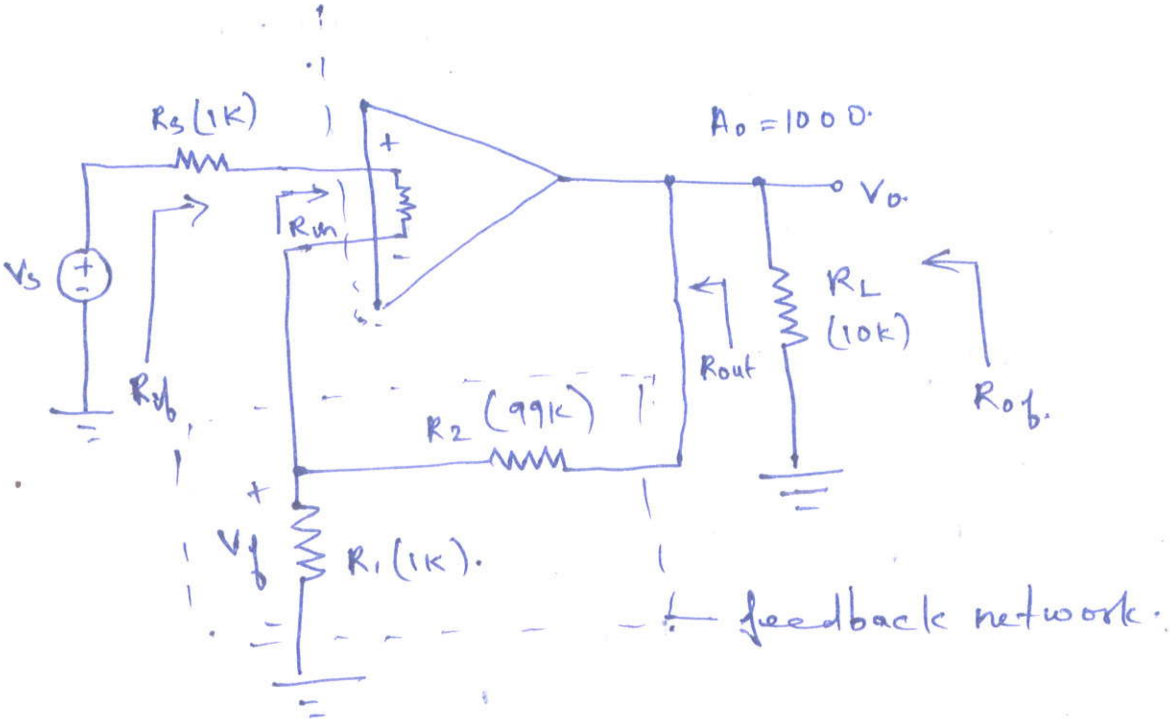


fig ①

The voltage across R_1 is the feedback voltage V_f while V_o is the input to the feedback circuit. The V_f is given to inverting terminal which opposes input voltage by 180° which ensures that the feedback is negative.

Analysis

Step 1: Identify topology

* By shorting output voltage ($V_o = 0$), feedback signal becomes zero and hence it is voltage sampling.

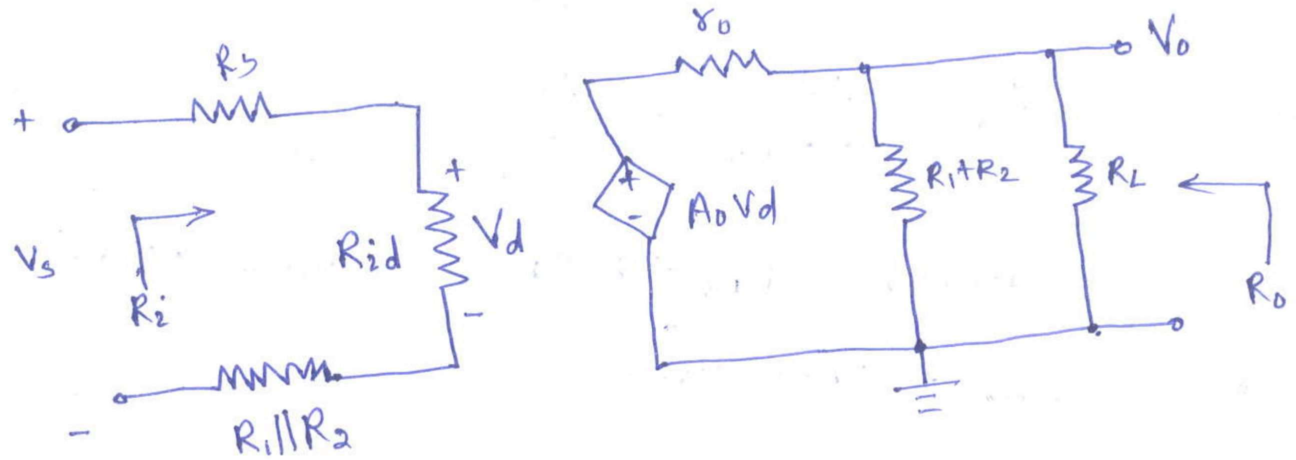
* Looking at fig ① we can see that feedback signal V_f is subtracted from the externally applied signal V_s and hence it is a series mixing. combining two

conclusions we can say that it is a voltage series feedback amplifier.

step 2 & step 3: Find input and output circuit.

* To find the input circuit, set $V_o = 0$, set $V_o = 0$.
By grounding V_o' we have R_s, R_{id} and parallel combination of R_1 & R_2 in series with V_s.

* To find the output circuit, set $I_i = 0$. Since $I_i = 0$, series combination of R_1 and R_2 appears in the output circuit. Following figure shows the obtained equivalent circuit.



step 4: Find open loop voltage gain

$$V_o = A_o V_d \times \frac{[(R_1+R_2) || R_L]}{[(R_1+R_2) || R_L] + r_o}$$

$$V_d = \frac{R_{id}}{R_s + R_{id} + (R_1 || R_2)} V_s$$

$$A_v = \frac{V_o}{V_s} = A_o \times \frac{[(R_1+R_2)||R_L]}{[(R_1+R_2)||R_L + r_o} \times \frac{R_{id}}{R_s + R_{id} + (R_1||R_2)}$$

$$= 1000 \times \frac{(1+99)||10}{[(1+99)||10]+1} \times \frac{100}{1+100+(1||99)}$$

$$= 883.32$$

Step 5 : calculate β

$$\beta = \frac{V_f}{V_o} = \frac{R_1}{R_1+R_2} = \frac{1}{1+99} = 0.01$$

Step 6 : Calculate D, A_{vf} , R_{if} and R_{of}

$$D = 1+\beta A_v = 1+0.01 \times 883.32$$

$$= 9.833$$

$$A_f = \frac{A_v}{1+\beta A_v} = \frac{A_v}{D} = \frac{883.32}{9.833} = 89.832$$

$$R_i = R_s + R_{id} + R_1||R_2 = 1+100+(1||99)$$

$$= 101.99\,k\Omega$$

$$R_{if} = R_i \times D = 101.99 \times 9.833$$

$$= 1002.87\,k\Omega$$

$$R_o = r_o||(R_1+R_2)||R_L = 1||(1+99)||10 = 0.9\,k\Omega$$

$$R_{of} = \frac{R_o}{D} = \frac{900}{9.833} = 91.53\,\Omega$$

Analysis of voltage series Feedback:

Eg: ① Transistor Emitter follower.

The below figure shows the transistor emitter follower circuit. Here the feedback voltage is the voltage across R_o and sampled signal is V_o across R_e.

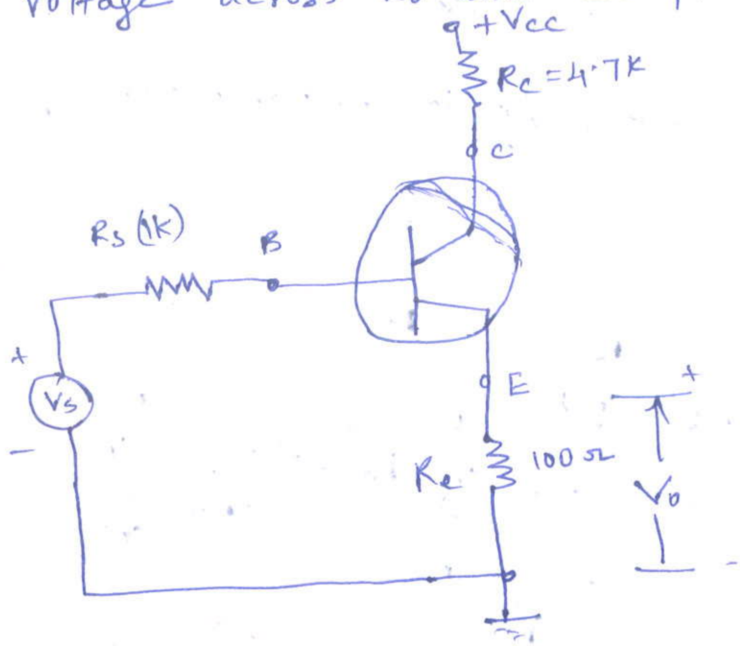


Fig ⊛

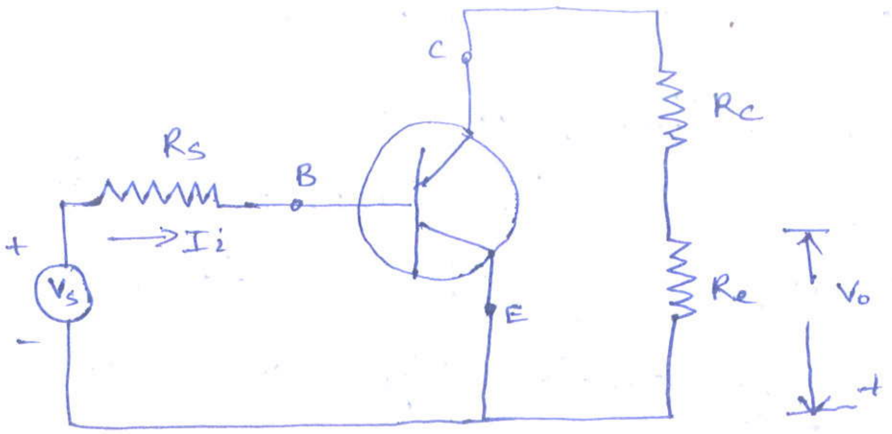
Step 1 : Identify topology

By shorting output voltage ($V_o = 0$), feedback signal becomes zero and hence it is voltage sampling. Looking at the fig ⊛ we can see that feedback signal V_f is subtracted from the externally applied signal V_s and hence it is a series mixing. Combining two conclusions we can say that it is a voltage series feedback amplifier.

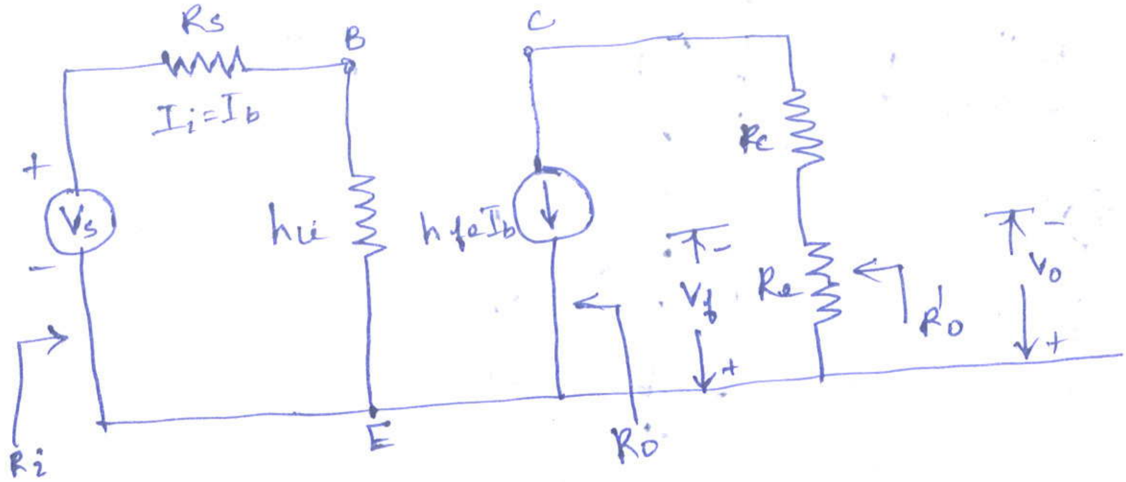
Step 2 & Step 3 : Find input & o/p ckt.

To find the input circuit, set $V_o = 0$, and hence V_s in series with R_s appears between B and E. To find the output circuit set $I_1 = I_b = 0$, and hence R_e appears only in the output loop.

With these connections we obtain the following ckt



Step 4: Replace transistor by its h-parameter equivalent ckt.



Step 5: Find open loop voltage gain

$$A = \frac{V_o}{V_s} = \frac{h_{fe} I_b R_e}{V_s} \quad \text{——— ①}$$

Applying KVL to i/p loop we get

$$V_s = I_b (R_s + h_{ie}) \quad \text{——— ②.}$$

Sub value of Vs in eq ① we get

$$A = \frac{h_{fe} R_e}{R_s + h_{ie}} = \frac{50 \times 100}{1 \times 10^3 + 1.1 \times 10^3} = 2.38$$

Step 6 : Indicate V_o and V_f and calculate β

$\beta = \frac{V_f}{V_o} = 1$ $\quad \therefore$ both voltage present across R_e.

Step 7 : Calculate D, A_{vf}, R_{if}, R_{of}

$$D = 1 + \beta A$$
$$= 1 + 1 \times 2.38 = 3.38$$

$$A_f = \frac{A}{1+\beta A} = \frac{A}{D} = \frac{2.38}{3.38} = 0.7$$

$$R_i = R_s + h_{ie} = 1k + 1.1k = 2.1k$$

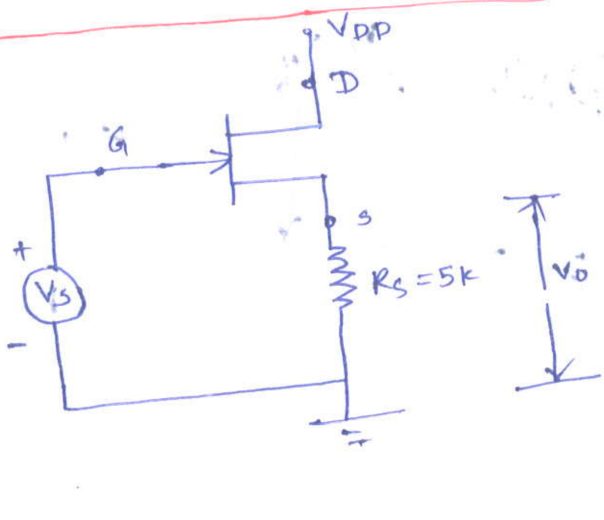
$$R_{if} = R_i D = 2.1k * 3.38 = 7.098\,k$$

$$R_o = \infty, \quad R_{of} = \infty$$

$$R'_{of} = \frac{R'_o}{D} \qquad \text{where } R'_o = R_e$$

$$R'_{of} = \frac{R_e}{D} = \frac{100}{3.38} = 29.58\,\Omega.$$

Eg. II FET Source follower.



r_d = 40K constant

Analysis

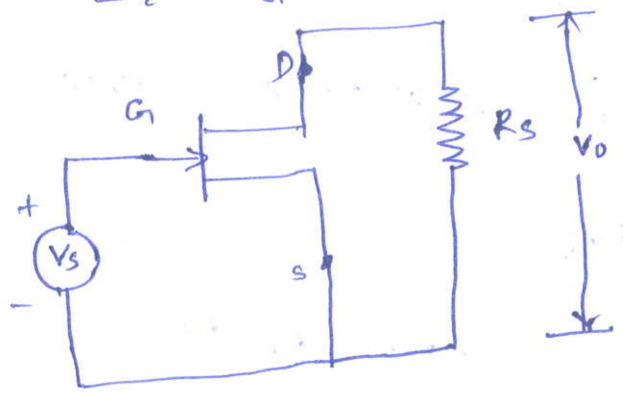
Step 1 : Identify topology

By shorting o/p voltage $V_o = 0$ feedback signal becomes zero. so it is a voltage sampling. feedback signal V_f is subtracted from the externally applied voltage V_s and it is series mixing. so it is voltage series feedback amplifier.

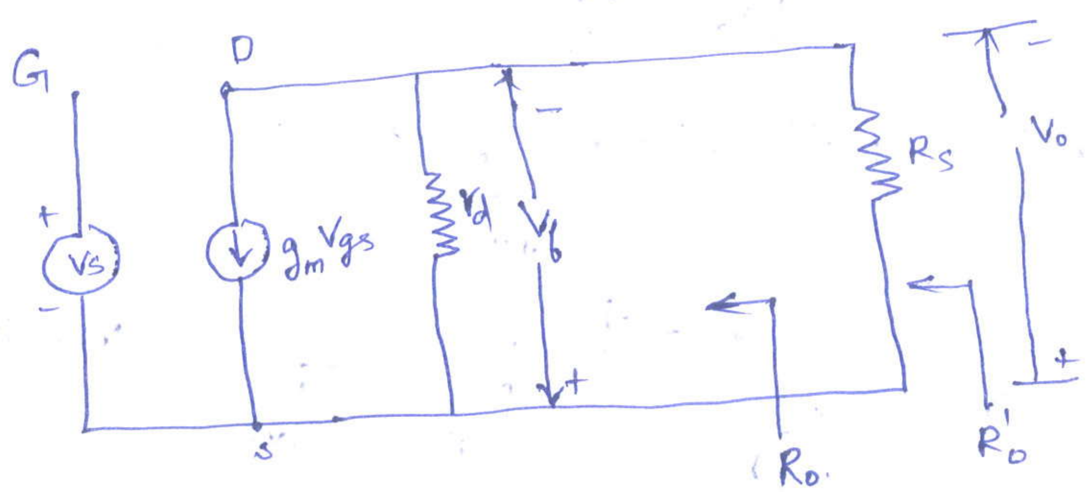
Step 2 & 3 Find i/p & o/p ckt.

i/p $V_o = 0$ hence V_s appears between G & S

o/p $I_i = I_G = 0$ hence R_s appears in the output loop.



Step 4 : Replace it by hybrid equivalent ck.



step 5 : Find the voltage gain

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

$$V_o = g_m \cdot V_{gs} \cdot (r_d \parallel R_s)$$

$$= \frac{r_d \cdot R_s}{r_d + R_s} (g_m \cdot V_{gs})$$

$V_{gs} = V_s$

$$A = \frac{\frac{r_d \cdot R_s}{r_d + R_s} \cdot (g_m \cdot \cancel{V_{gs}})}{\cancel{V_{gs}}} = g_m \cdot \frac{r_d \cdot R_s}{r_d + R_s}$$

$$\boxed{A = \frac{\mu \cdot R_s}{r_d + R_s}}$$

$\mu = g_m r_d$.

step 6 : To find β

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

$V_f = \beta \cdot V_o$ $\quad \beta = 1$

$$V_f = V_o$$

$R_i = \infty$ $\quad$ ($\because$ No i/p impedance, as series connected $R_i = \infty$]

$$R_{if} = R_i(1 + A\beta) = \infty$$

$$R_o = R_d$$

$$R_{of} = R_o / D = \frac{r_d}{1 + A\beta}.$$

step 7 : Calculate D, A_{vf}, R_{if}, R_{of} and R'_{of}.

$$A_v = \frac{\mu \cdot R_S}{R_S + r_d} = \frac{40 \times 5 \times 10^3}{5000 + 40 \times 10^3} = \frac{200}{45} = 4.44$$

$$\beta = \frac{V_f}{V_o} = 1$$

$$D = 1 + \beta A_v = 1 + 4.44 = 5.44$$

$$A_{vf} = \frac{A_v}{1 + \beta \cdot A_v} = \frac{A_v}{D} = \frac{4.44}{5.44} = 0.816.$$

$$R_i = \infty \qquad R_{if} = R_i D = \infty$$

$$R_o = r_d = 40\,k\Omega$$

$$R_{of} = \frac{R_o}{D} = \frac{40K}{5.44} = 7.35\,K$$

$$R'_{of} = \frac{R'_o}{D}$$

$$R'_o = R_S \,//\, r_d = \frac{5K \times 40K}{5K + 40K} = 4.44\,k$$

$$R'_o = 4.44\,K$$

$$R'_{of} = \frac{R'_o}{D} = \frac{4.44\,k}{5.44} = 816.2\,\Omega$$

$$R'_{of} = 816.2\,\Omega.$$

The Feedback Transconductance Amplifier (series-series Feedback topology).

The following figure shows the ideal structure for the series-series feedback amplifier. It consists of a unilateral open-loop amplifier (A ckt) and an ideal feedback network.

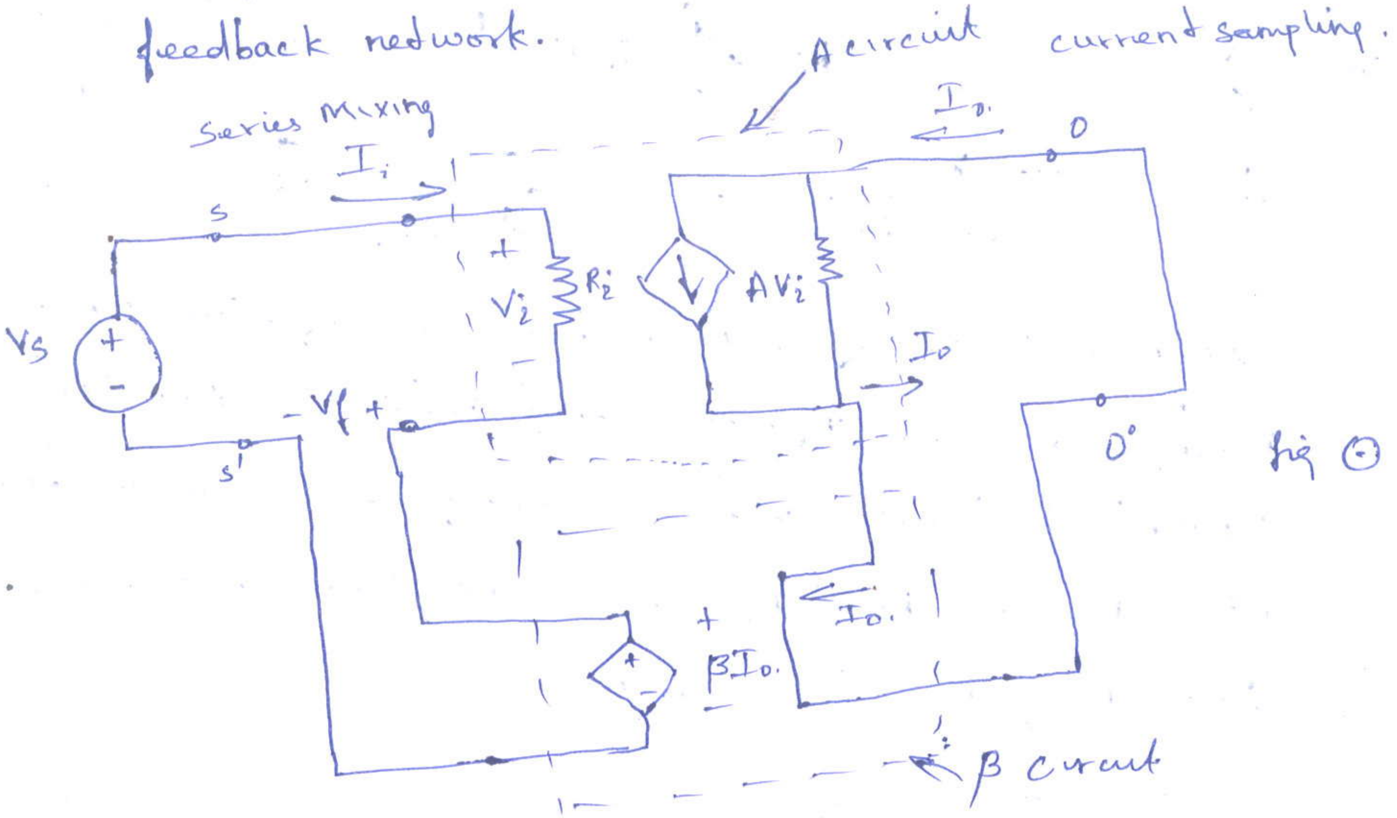
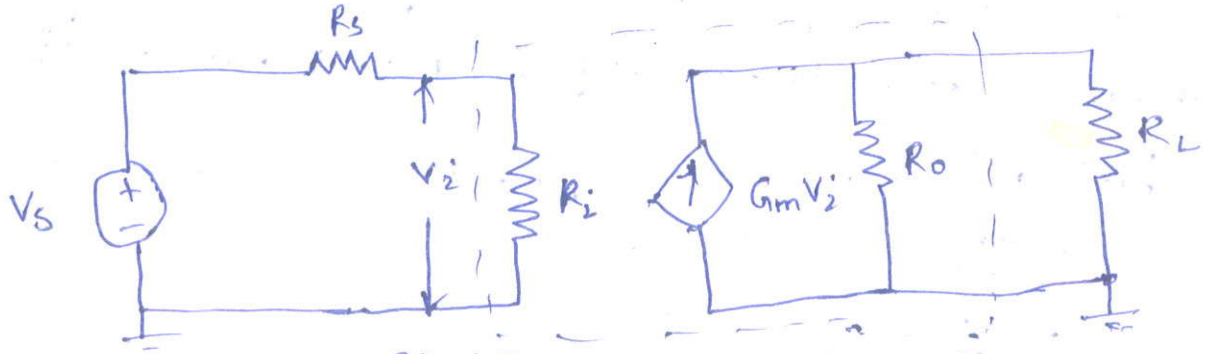


fig ①

The A circuit has an input resistance R_i, a short-ckt transconductance $A \equiv I_o/V_i$ and an output resistance R_o. The β ckt samples the short-circuit output current I_o and provides a feedback voltage V_f that is subtracted from V_s in the series input loop. The fig ① shows a transconductance amplifier with a thevenin's equivalent in its input ckt and Norton's equivalent in it output ckt. In this amplifier, an o/p current is proportional the input signal voltage and the proportionality factor is independent of the magnitudes of the source and load resistance.



$R_i >> R_s$. $R_o >> R_L$.

fig ⊛

Ideally, this amplifier must have an infinite input resistance R_i and infinite output resistance R_o. For practical transconductance amplifier we must have $R_i >> R_s$ and $R_o >> R_L$. Since $R_i >> R_s$, $V_i \simeq V_s$ and since $R_o >> R_L$, $I_L = G_m V_i$.

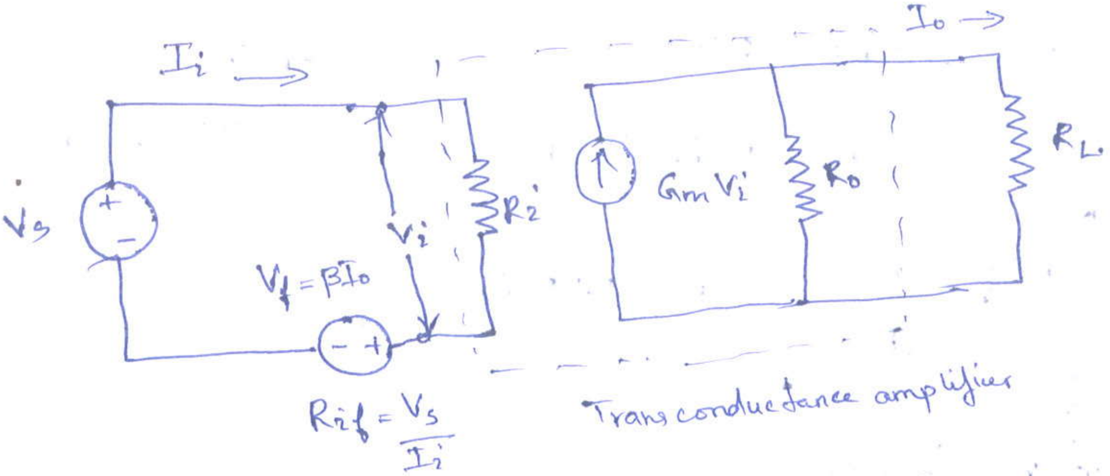
$I_L = G_m V_s$ Where $G_m = \frac{I_L}{V_s}$ is the transfer or Mutual conductance

As shown in fig ⊙ the feedback topology is suitable for transconductance amplifier is current sampling and series mixing. Such topology is known as series-series feedback or current series feedback.

input and output resistance

input resistance

The current series feedback topology is shown in figure below with amplifier input circuit is represented by Thevenin's equivalent circuit and output circuit by Norton's equivalent circuit.



Transconductance amplifier

From fig i/p resistance with feedback is given as.

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

* obtain the expression for V_s

Apply KVL to the input side we get.

$$V_s - I_i R_i - V_f = 0$$

$$V_s = I_i R_i + V_f$$

$$= I_i R_i + \beta I_o \qquad \because V_f = \beta I_o$$

* Obtain the expression for I_o in terms of V_i

The o/p current I_o is given as

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

G_m represents the open circuit transconductance without feedback and G_M is the transconductance without feedback taking the load R_L into account.

* obtain expression for R_{if}.

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

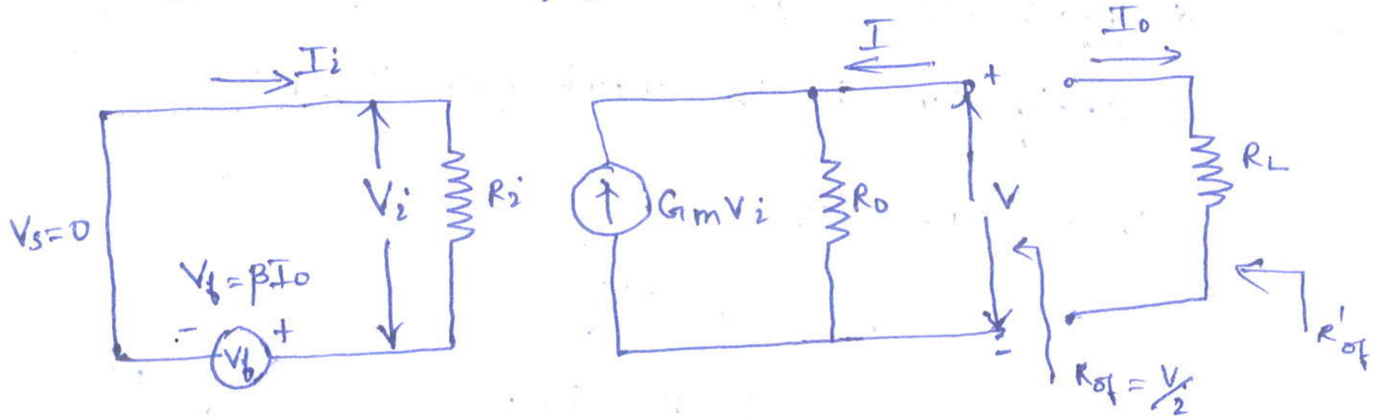
$$V_s = I_i R_i + \beta I_o \qquad I_o = G_M V_i$$

$$= I_i R_i + \beta G_M I_i R_i \qquad V_i = I_i R_i$$

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

output resistance

In this topology the output resistance can be measured by shorting the input source $V_s = 0$ and looking into the output terminals with R_L disconnected as shown in the figure below.



* obtain expression for I in terms of V

Applying KCL to the o/p node we get.

$$I = \frac{V}{R_o} - G_m V_i$$

The input voltage is given by

$$V_i = -V_f = -\beta I_o$$

$$= \beta I \qquad \therefore I_o = -I$$

Sub. the value

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

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

* Obtain expression for Rd

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

Here G_m is the open loop transconductance without taking R_L in account.

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

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

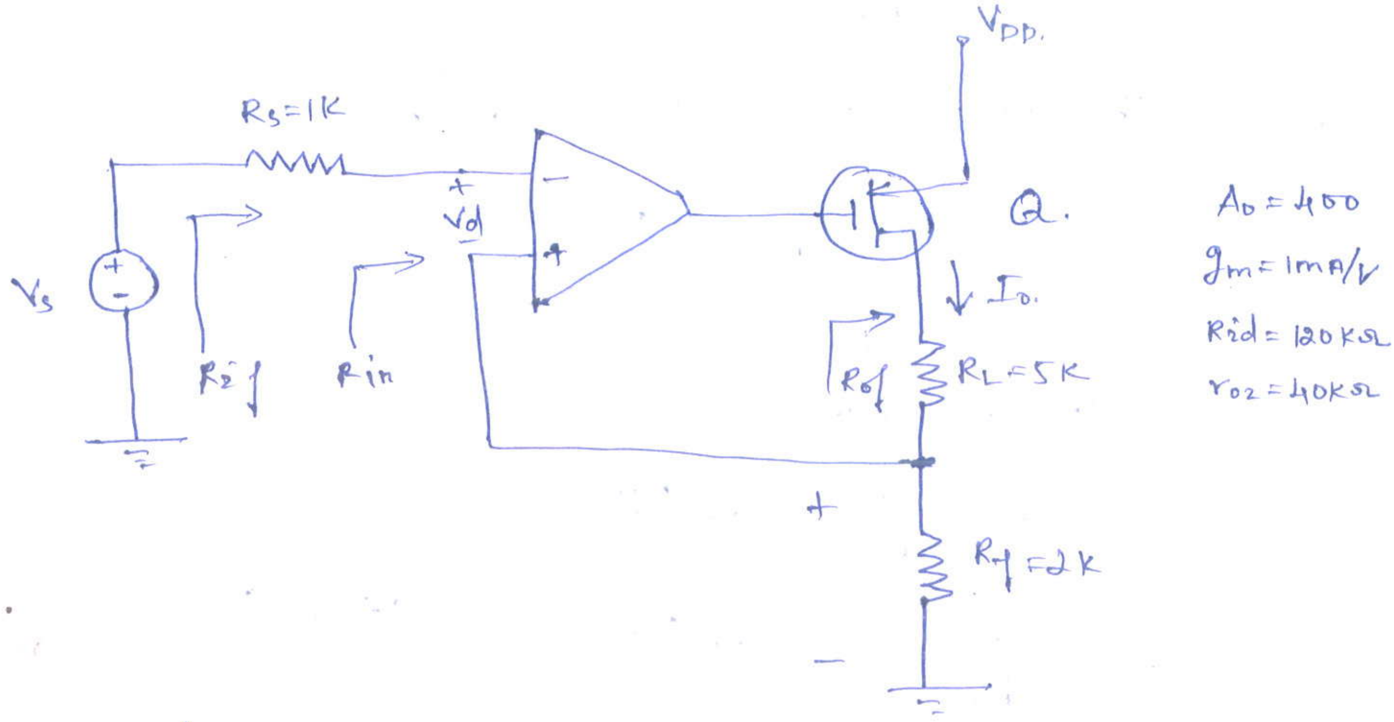
$\div$ num. & den. by $R_o + R_L$

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

$$R'_{of} = \frac{R'_o (1+\beta G_m)}{1+\beta G_m} \qquad \therefore R'_o = \frac{R_o R_L}{R_o + R_L} \quad G_m = \frac{G_m R_o}{R_o + R_L}$$

* Here, G_m is the open loop current gain taking R_L in account.

practical current series feedback amplifier circu...



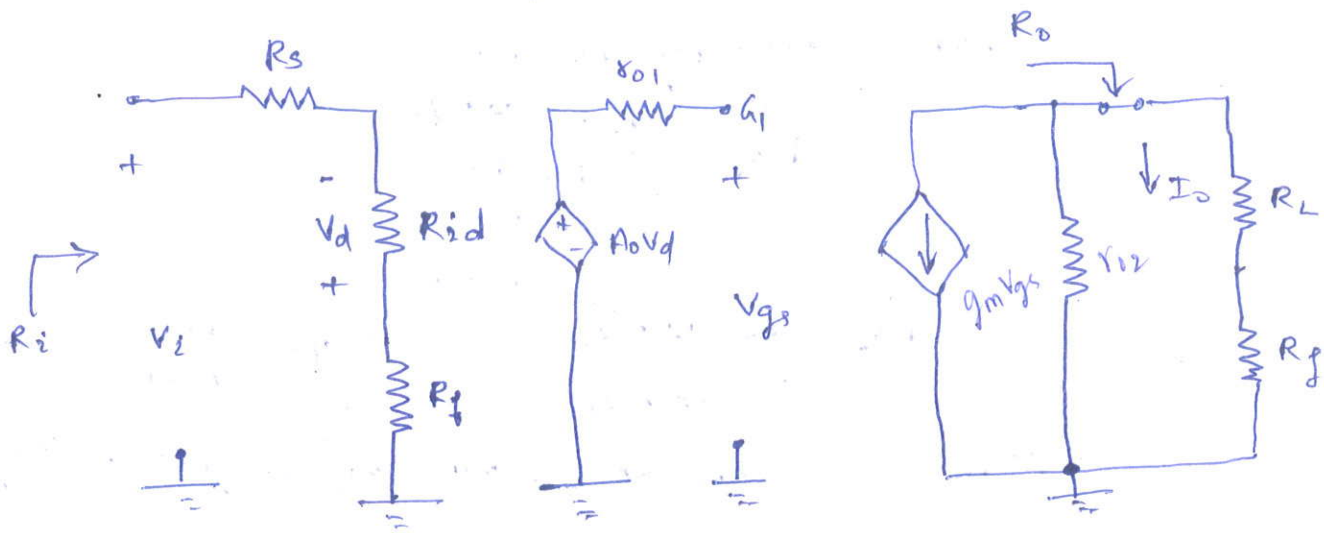
Analysis

Step 1: identify topology.

By opening the output loop, (o/p current, $I_o = 0$) feedback signal becomes zero and hence it is current sampling.

From fig, the feedback signal V_f is subtracted from the externally applied signal V_s and hence it is a series mixing. Combining two conclusions we can say it is a current series feedback amplifier.

Step 2 and step 3: Find the input and output circuit.



To find input circuit set $I_o = 0$. This results R_f to appear at the input side in series with R_{id}.

To find o/p circuit set $I_i = 0$. This results R_f to appear at the output side

Step 4 : Find open loop voltage gain

$$V_d = -V_s \frac{R_{id}}{R_s + R_{id} + R_f} \quad \text{——— (1)}$$

$$V_{gs} = A_o V_d \quad \text{——— (2)}$$

$$I_o = -g_m V_{gs} \frac{r_{o2}}{r_{o2} + R_L + R_f} \quad \text{——— (3)}$$

using eq (1), (2) & (3) we have

$$G_m = \frac{I_o}{V_i} = (g_m A_o)\left(\frac{r_{o2}}{r_{o2} + R_L + R_f}\right)\left(\frac{R_{id}}{R_s + R_{id} + R_f}\right) \quad \text{——— (4)}$$

usually, $R_{id} >> (R_s + R_f)$ and $r_{o2} >> (R_L + R_f)$

$$\therefore \quad G_m \simeq g_m A_o$$

$$\therefore \quad G_M = (1 \times 10^{-3} \times 400)\left(\frac{40}{40+5+2}\right)\left(\frac{120}{1+120+2}\right)$$

$$= 0.332.$$

Step 5 : Calculate β

$$\beta = \frac{V_f}{I_o} = \frac{I_o R_f}{I_o} = R_f = 2k$$

Step 6 : calculate D, G_{MF}, A_{vf}, R_{if} and R_{of}.

$$D = 1 + \beta G_M = 1 + 2000 \times 0.332$$

$$= 665$$

$$G_{MF} = \frac{G_M}{D} = \frac{0.332}{665} = 5 \times 10^{-4}$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{I_o R_L}{V_s} = G_{MF} R_L = 5 \times 10^{-4} \times 5 \times 10^{3}$$

$$= 2.5$$

$$R_i = R_s + R_{id} + R_f = 1 + 120 + 2$$

$$= 123 K$$

$$R_{if} = R_i D = 123 \times 665 = 81.8 M\Omega$$

$$R_o = r_{o2} + R_L + R_F = 40 + 5 + 2 = 47 K\Omega$$

$$R_{of} = R_o D = 47 \times 665$$

$$= 31.255 M\Omega .$$

The Feedback Transresistance Amplifier (shunt-shunt).

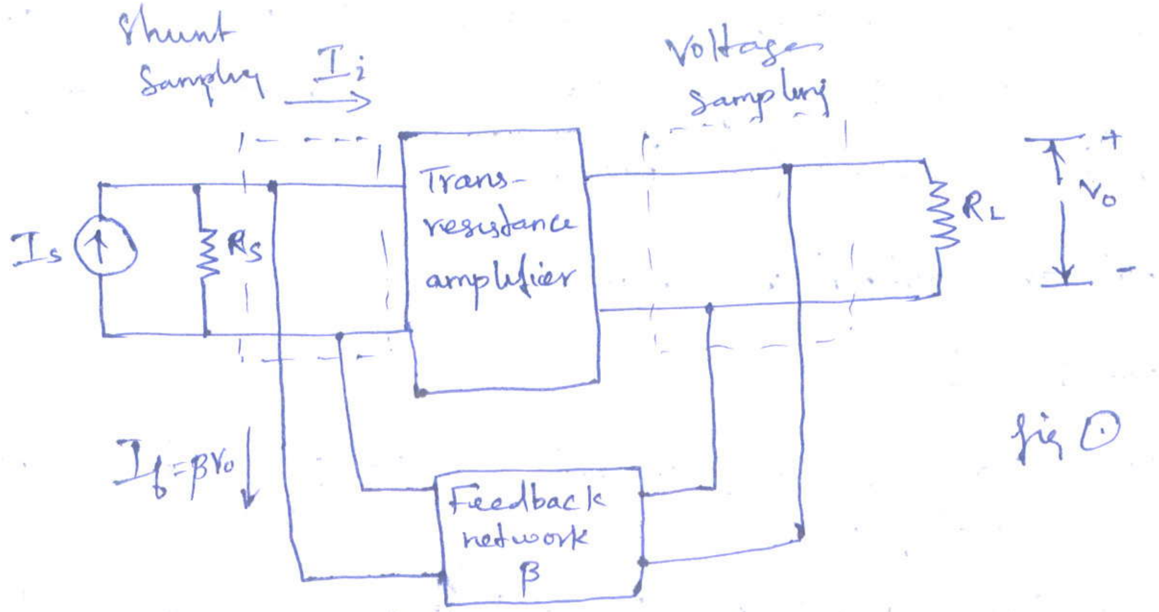
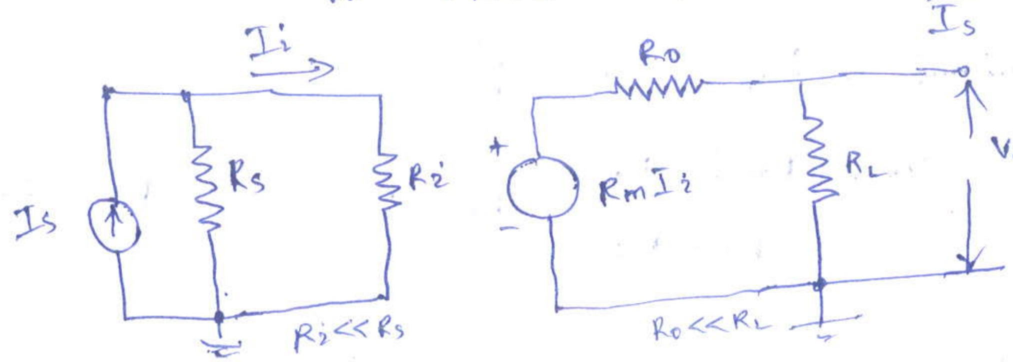


fig ①

The above figure shows the transresistance amplifier with voltage shunt feedback. The below figure shows a transresistance amplifier with a Norton's equivalent in its input circuit and a Thevenin's equivalent in its output circuit. In this amplifier an output voltage is proportional to the input signal current and the proportionality factor is independent on the source and load resistances. Ideally, this amplifier must have zero i/p resistance R_i zero o/p resistance R_o. For practical transresistance amplifier we must have $R_i \ll R_s$ and $R_o \ll R_L$. Since $R_i \ll R_s$, $I_i = I_s$ and since $R_o \ll R_L$, $V_o = R_m I_i$.

$$V_o = R_m I_s \qquad \text{Where } R_m = \frac{V_o}{I_s} \text{ is the transfer or mutual resistance.}$$



As shown in Fig ①, the feedback topology suitable for transresistance amplifier is voltage sampling and shunt mixing. Such topology is known as shunt-shunt feedback or voltage shunt feedback.

Input and output resistance.

Input Resistance.

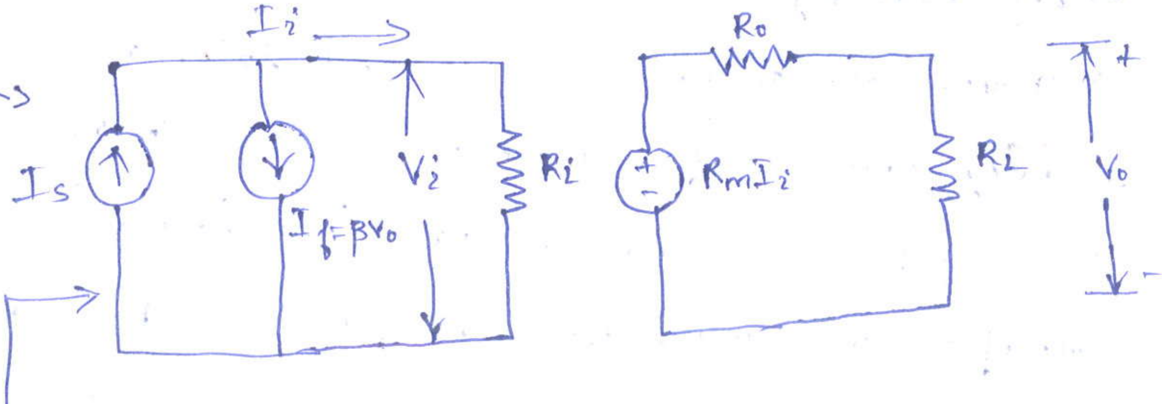
Step 1: Draw the equivalent circuit for voltage shunt amplifier.

- The voltage shunt feedback topology is shown below. With amplifier input circuit is represented by Norton's equivalent ckt and o/p ckt represented by Thevenin's equivalent.

Step 2: Obtain expression for I_s

Applying KCL at input node we get,

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



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

Step 3: Obtain expression for V_o interms of I_i

The output voltage V_o is given

$$V_o = \frac{R_m I_i R_o}{R_o + R_L} = R_M I_i \qquad R_M = \frac{R_m R_o}{R_o + R_L}$$

R_m represents the open circuit transresistance without feedback and R_M is the transresistance without feedback taking the load R_L into account.

Step 4: Obtain expression for R_{if}

The i/p resistance with feedback R_{if} is given as

$$R_{if} = \frac{V_i}{I_s}$$

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

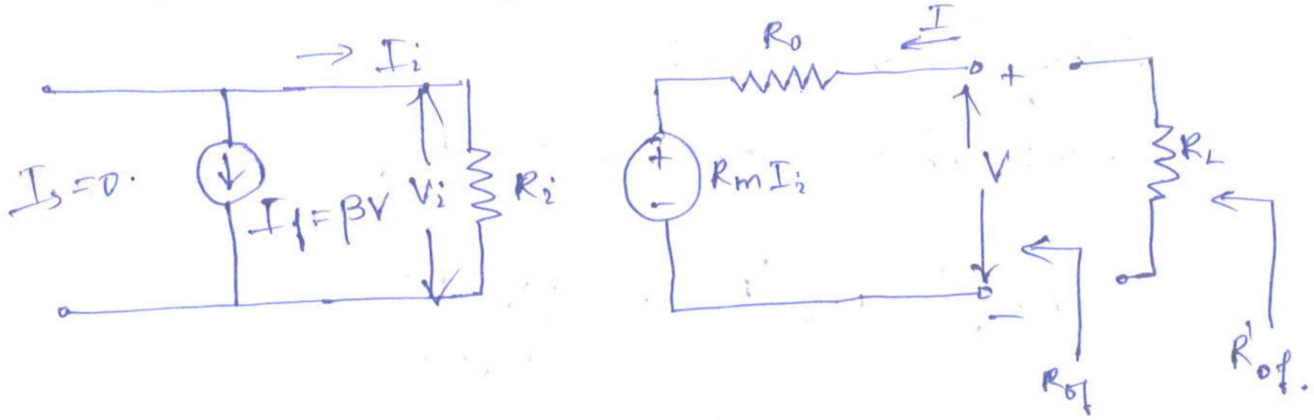
$$= I_i (1 + \beta R_M)$$

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

output resistance

Step 1: Draw equivalent circuit.

In this topology, the output resistance can be measured by making $I_s = 0$ and looking into the o/p terminals with R_L disconnected, as shown below



step 2:
obtain expression for I in terms of V

Applying KVL to the output side we get

$$R_m I_i + I R_o - V = 0$$

$$I = \frac{V - R_m I_i}{R_o} \quad \text{——— (1)}$$

The input current is given as

$$I_i = -I_f = -\beta V \quad \text{——— (2)}$$

Sub I_i from equ (2) in equ (1) we get

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

step 3 : Obtain expression for R_{of}

$$R_{of} = \frac{V}{I}$$

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

Here, R_m is the open loop transresistance without taking R_L in account.

step 4 : Obtain expression for R'_{of}

$$R'_{of} = R_{of} \| R_L = \frac{R_{of} \times R_L}{R_{of} + R_L}$$

$$= \frac{\dfrac{R_o \times R_L}{1 + R_m \beta}}{\dfrac{R_o}{1 + R_m \beta} + R_L} = \frac{R_o R_L}{R_o + R_L(1 + R_m \beta)}.$$

Dividing num. & deno. by $(R_o + R_L)$ we get,

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

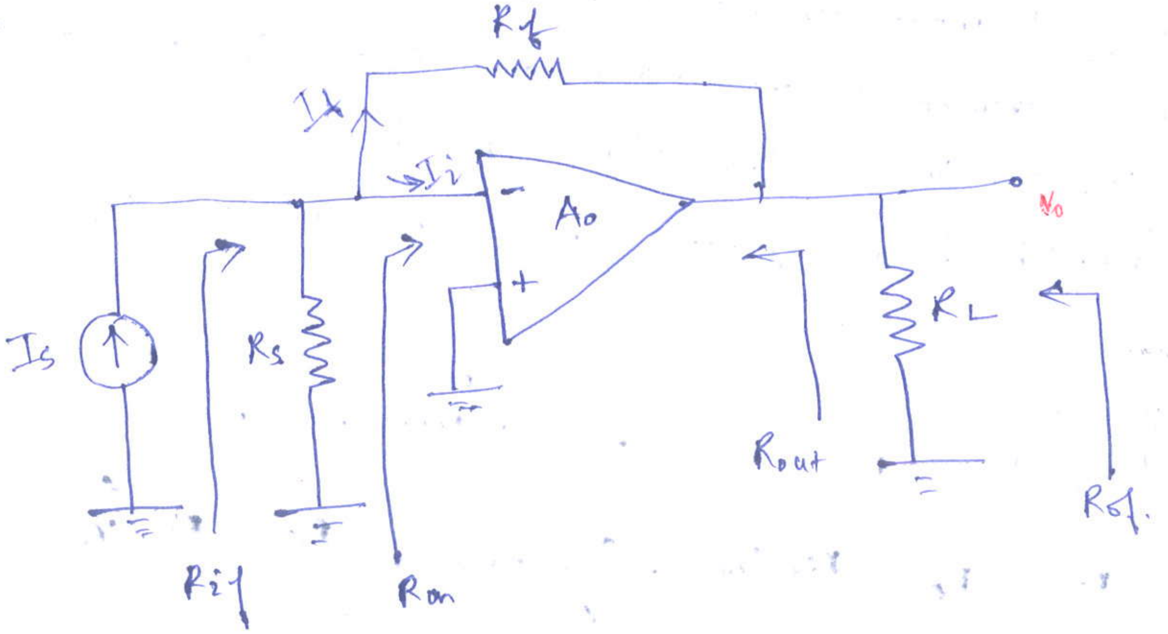
$$\boxed{R'_{of} = \frac{R'_o}{1 + \beta R_M}}$$

Where $R'_o = \frac{R_L + R_{of}}{R_L + R_{of}}$

$R_M = \frac{R_m R_L}{(R_o + R_L)}$.

Here R_M is the open loop transresistance taking R_L in account.

Analysis of practical voltage-shunt Feedback amplifier



The above figure shows a practical voltage-shunt feedback amplifier. This is a practical inverting amplifier with feedback. The resistance R_F forms the feedback circuit. The inverted output is applied to the inverting terminal through resistance R_F. The polarity of the output voltage which is feedback is opposite to the polarity of the input voltage. Thus the feedback is negative.

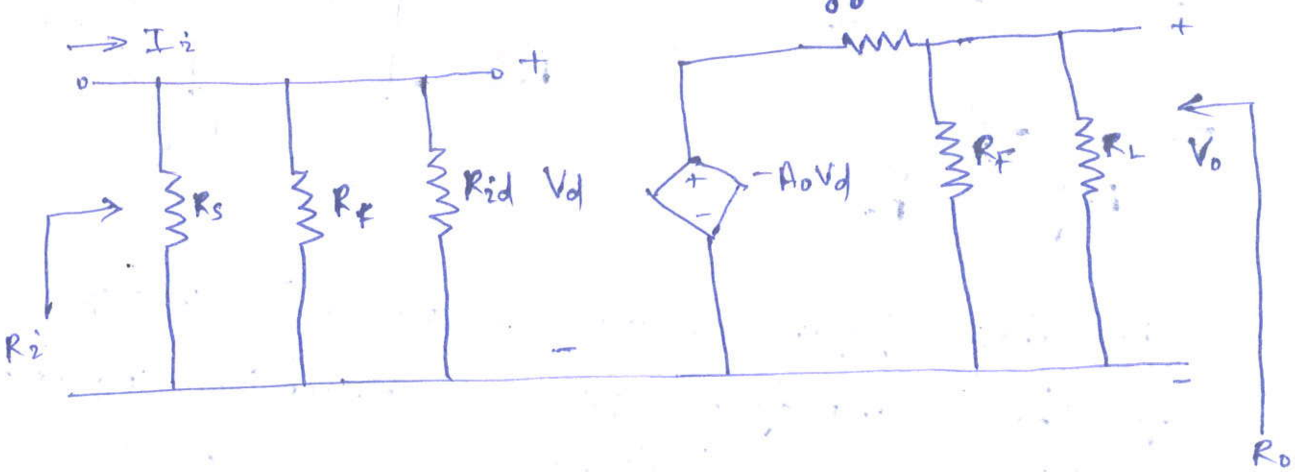
The voltage amplifier is implemented with op-amp having open-loop gain A_0, input resistance R_{id} and an o/p resistance r_0.

Step ① Identify topology:

By shorting output voltage ($V_0 = 0$), 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 step ③ Find input and output circuit.

To find input circuit, set $V_0 = 0$, this places R_F between inverted input and ground. To find output circuit, set $V_i = 0$, this places R_F between output and ground. The below figure shows the obtained equivalent circuit.



Step 4 : Find: open loop voltage gain

From fig we have

$$V_o = -A_o V_d \frac{(R_F \| R_L)}{(R_F \| R_L) + r_o} \quad \text{——— (1)}$$

$$V_d = I_i R_i \quad \text{——— (2)}$$

$$R_i = R_S \| R_F \| R_{id} \quad \text{——— (3)}$$

$$= 1 \| 10 \| 120 = 0.9 k\Omega$$

Using equations (1), (2) and (3) we have

$$R_M = \frac{V_o}{I_i} = -A_o \frac{(R_F \| R_L)}{(R_F \| R_L) + r_o} \cdot R_i$$

$$= -1000 \left[\frac{(10 \| 5)}{(10 \| 5) + 0.1} \right] \times 0.9 = -873.79 k\Omega$$

Step 5 : calculate β

$$I_f = \frac{V_i - V_o}{R_F} = -\frac{V_o}{R_F} \qquad \because V_o >> V_i$$

$$\beta = \frac{I_f}{V_o} = -\frac{1}{R_F} = -1/10 = -0.1 mA/V$$

Step 6 : calculate D, R_{MF}, A_{vf}, R_{if} and R_{of}.

$$D = 1 + \beta R_M = 1 + (-0.1 \times 10^{-3}) \times -873.79 \times 10^{3}$$
$$= 88.379$$

$$R_{Mf} = \frac{R_M}{1 + R_M \beta} = \frac{R_M}{D} = \frac{-873.79}{88.379} = -9.89 k\Omega$$

$$A_{vf} = \frac{V_o}{V_s} = \frac{V_o}{I_s R_s} = \frac{R_{mf}}{R_s} = \frac{-9.89\,K}{1\,K} = -9.89$$

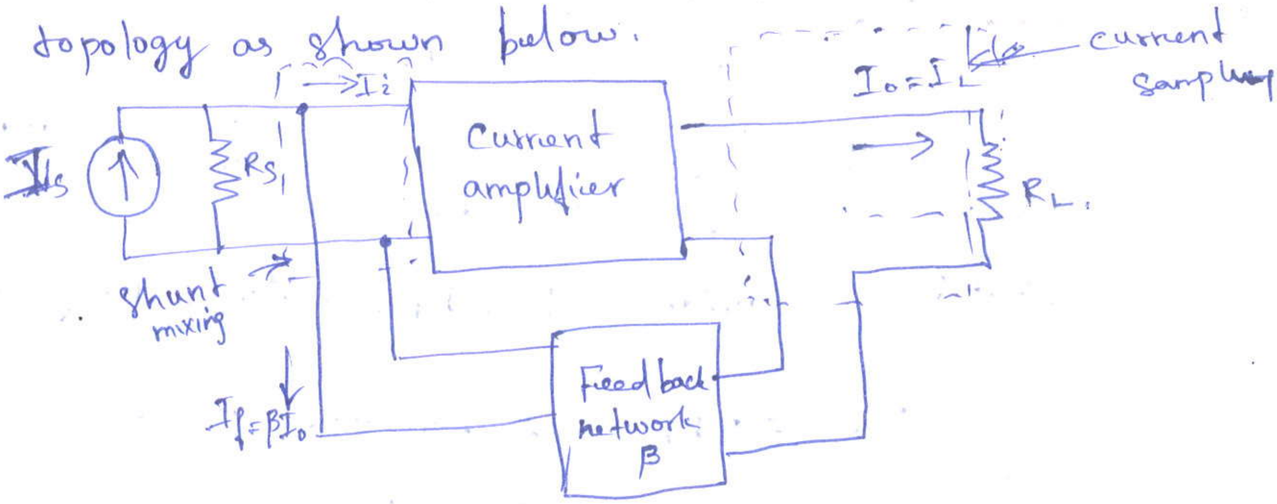
$$R_{if} = \frac{R_i}{D} = \frac{900\,\Omega}{88.379} = 10.18\,\Omega$$

$$R_o = r_o \,//\, R_F \,//\, R_S = 0.1 \,//\, 10 \,//\, 1 = 90\,\Omega$$

$$R_{of} = \frac{R_o}{D} = \frac{90}{88.379} = 1.02\,\Omega$$

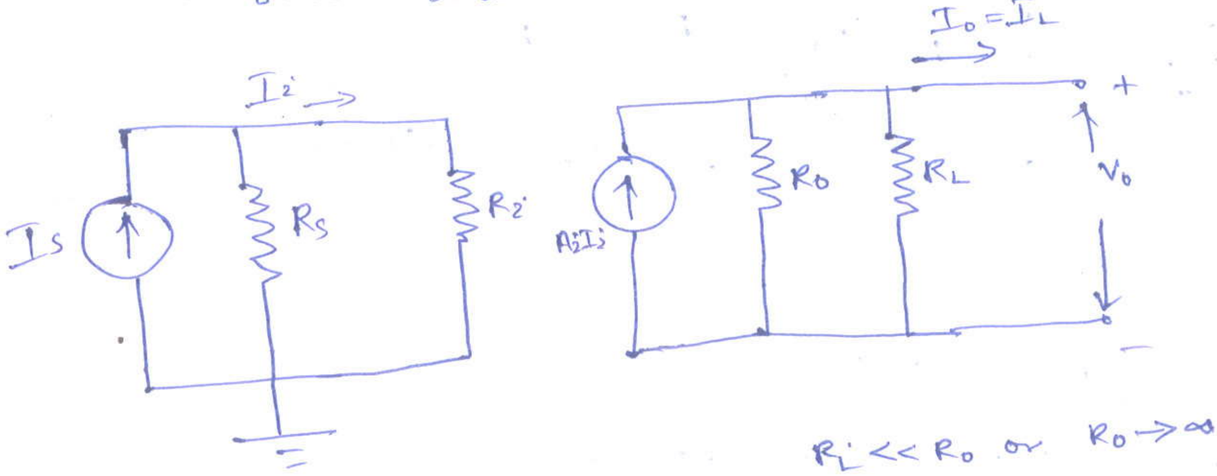
- The feedback current amplifier (shunt - series)

The input signal in a current amplifier is essentially a current, and thus the signal source is most conveniently represented by its Norton equivalent. The o/p quantity of interest is current; hence the feedback network should sample the output current, just as a current meter measures a current. The feedback topology most suitable for a current amplifier is the current mixing, current-sampling topology as shown below.



Because of the parallel (or shunt) connection at the input and the series connection at the output, this feedback topology is also known as shunt-series feedback.

Here, the signal source is a current a hence, it is convenient to represent it in Norton's equivalent circuit as shown below.



$R_L << R_o$ or $R_o \to \infty$

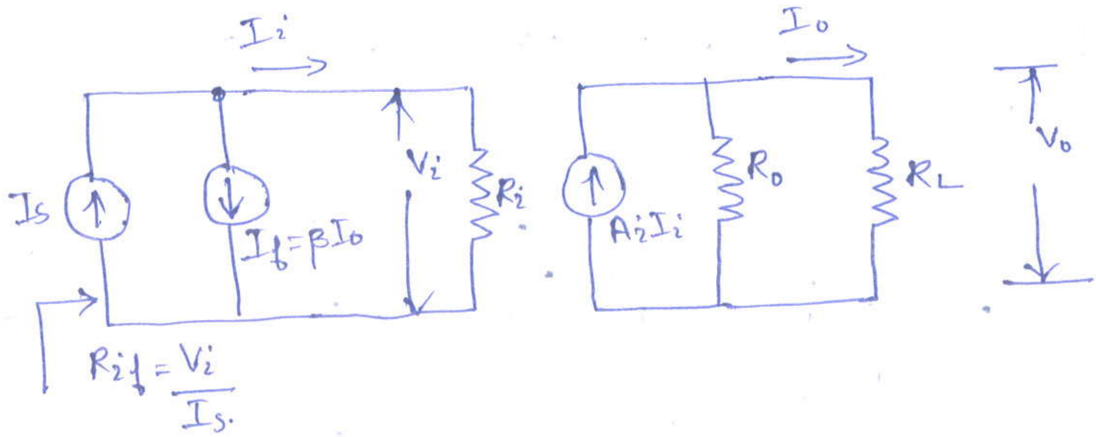
$R_i \to 0$ or $R_i << R_s$

For current amplifier, input resistance $R_i \to 0$ and hence $I_i \approx I_s$ and output resistance $R_o \to \infty$ and hence $I_L = A_i I_i$ (where A_i = current gain). Such amplifier provides a current output proportional to the input current, and the proportionality factor is independent on source and load resistance. This amplifier is called current amplifier. An ideal current amplifier must have zero input resistance R_i and infinite o/p resistance R_o. For practical current amplifier we must have $R_i << R_s$ and $R_o >> R_L$.

Input and output Resistances.

Input Resistance Expression

step ① Draw the equivalent circuit current shunt amplifier i/p & o/p ckt replaced by norton's equivalent ckt.



step 2: obtain expression for I_s.

Applying KCL to the input node we get.

$$I_s = I_f + I_i = I_i + \beta I_o \qquad \because I_f = \beta I_o \quad \text{——— ①}$$

step 3: Obtain expression for I_o interms of I_i.

The o/p current I_o is given as.

$$I_o = \frac{A_i I_i \cdot R_o}{R_o + R_L} = A_I I_i \quad \text{——— ②}$$

where $A_I = \frac{A_i R_o}{R_o + R_L}$

A_i represents the open circuit current gain without feedback and A_I is the current gain with out feedback taking the load R_L in to account.

step ④: obtain expression for R_{if}.

Sub. value of I_o from equ ② into equ ① we get,

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

The input resistance with feedback is given as

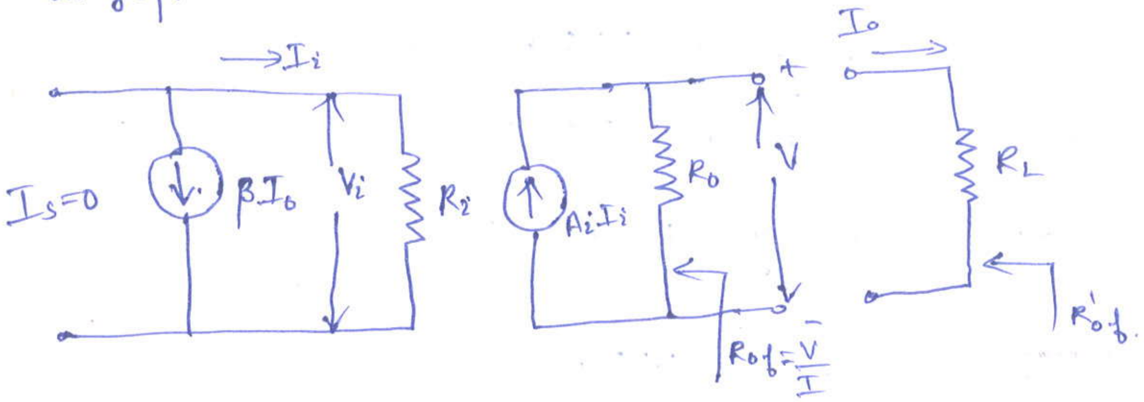
$$R_{if} = \frac{V_i}{I_s} = \frac{V_i}{I_i(1+\beta A_I)}$$

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

output resistance.

. Step 1 : Draw the equivalent circuit.

In this topology, the output resistance can be measured by open circuiting the input source $I_s = 0$ and looking into the output terminals, with R_L disconnected as shown in fig.



Step 2 : Obtain expression for I in terms of V

Apply the KCL to the output node we get

$$I = \frac{V}{R_o} - A_i I_i \qquad \text{——— (1)}$$

The input current is given as.

$$I_i = -I_f = -\beta I_o \qquad \because I_s = 0.$$

$$= \beta I \qquad I = -I_o \qquad \text{——— (2).}$$

sub. value of I_i from equ ② in equ ① we get

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

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

Step 3 : Obtain expression for R_{of}.

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

Here A_i is the open loop current gain without taking R_L in account.

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

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

÷ num & den by $(R_o + R_L)$ we get

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

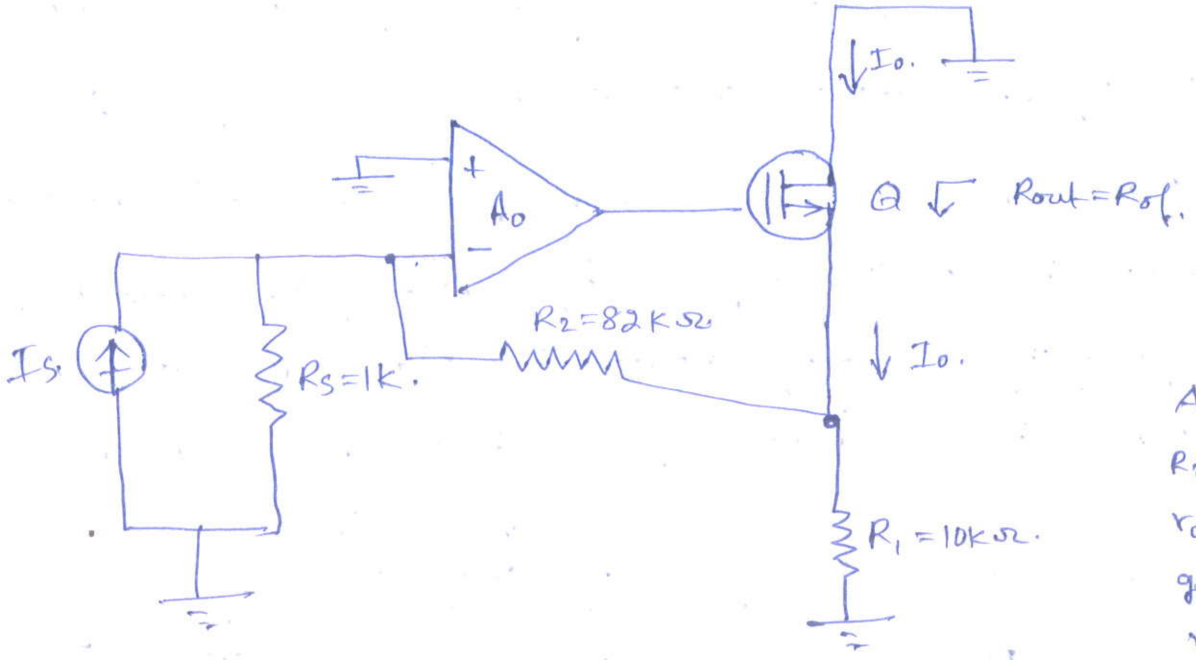
$$R'_{of} = \frac{R'_o(1+\beta A_i)}{(1+\beta A_I)}$$

$$\therefore R'_o = \frac{R_o R_L}{R_o + R_L}$$

$$A_I = \frac{A_i R_o}{R_o + R_L}.$$

Here A_I is the open loop current gain taking R_L in account.

Analysis of practical current shunt Feedback amplifier ckt



$A_o = 1000$
$R_{id} = \infty$
$r_{o1} = 1k\Omega$
$g_m = 5\,mA/V$
$r_{o2} = 25\,k\Omega$

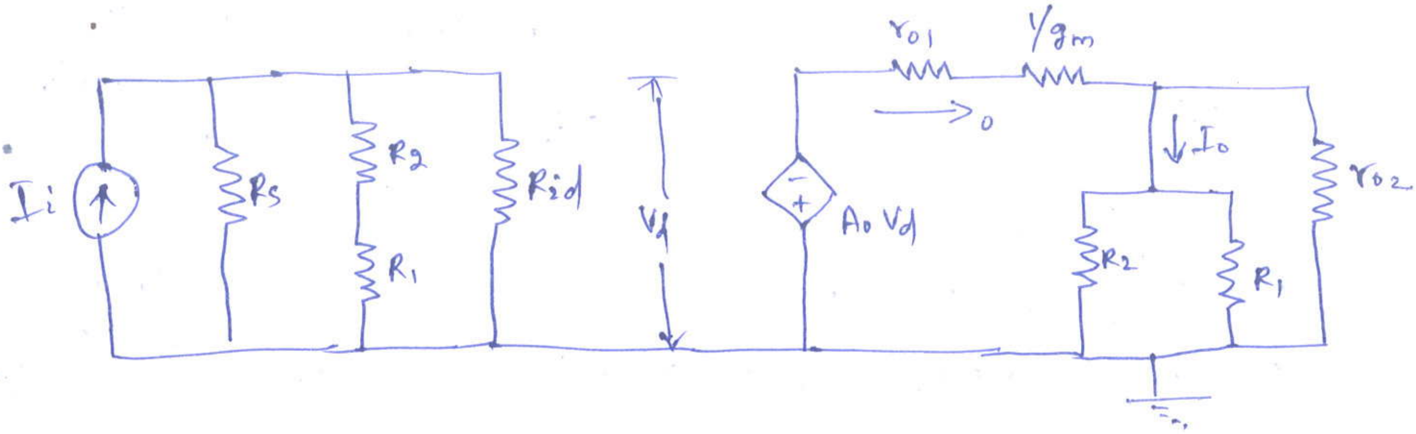
The above figure shows a practical current shunt feedback amplifier. It consists of an inverting voltage amplifier followed by a MOSFET. The resistance R_1 and R_2 forms a feedback network. The source current of MOSFET is sampled and applied at the inverting input $I_i = I_s - I_f$.

Step 1 : Identify the topology.

The output loop ($I_o = 0$), feedback signal becomes zero and hence it is a current feedback. The feedback signal appears in shunt with input ($I_i = I_s - I_f$), hence the topology is current shunt feedback amplifier.

Step 2 and step 3: Find the input & o/p circuit.

The input ckt of the amp without feedback is obtained by opening the output loop at the source of Q ($I_o = 0$). This places R_2 in series with R_1 from inverting input to ground. The o/p ckt is found by shorting the input node, i.e. grounding inverting input. This places R_2 in parallel with R_1. The resultant equ. ckt is shown below.



Step 4: Find open loop gain

From the above fig

$$R_i = R_s \,||\, (R_2 + R_1) \,||\, R_{id} \quad \text{——— (1)}$$

$$= 1 \,||\, (82 + 10) \,||\, \infty = 989\,\Omega.$$

$$V_d = I_i R_i \quad \text{——— (2)}$$

$$I_o = \frac{-A_o V_d}{\frac{1}{g_m} + (R_2 || R_1 || r_{o2})} \times \frac{r_{o2}}{r_{o2} + (R_1 || R_2)} \quad \text{——— (3)}$$

Using equ (1) (2) and (3) we have

$$A_I = \frac{I_o}{I_i} = \frac{-A_o R_i}{\frac{1}{g_m} + (R_2 || R_1 || r_{o2})} \times \frac{r_{o2}}{r_{o2} + (R_1 || R_2)}$$

$$= \frac{-1000 \times 989}{\frac{1}{5\times10^{-3}} \times (82\|10\|25)\times10^{3}} \times \frac{25}{25+(10\|82)} \times 10^{3}$$

$$= -554.8$$

Step 5 : calculate β.

$$\beta = \frac{I_f}{I_o} = -\frac{R_1}{R_1+R_2} = \frac{-10}{10+82} = -0.1087$$

Step 6 : calculate D, A_{If}, R_{if} and R_{of}.

$$D = 1 + A_I \beta = 1 + (-554.8) \times (-0.1087) = 61.3$$

$$A_{If} = \frac{A_I}{1+A_I\beta} = \frac{A_I}{D} = \frac{-554.8}{61.3} = -9.05$$

$$R_{if} = \frac{R_i}{1+A_I\beta} = \frac{R_i}{D} = \frac{989}{61.3} = 16.13\ \Omega.$$

Here, r_o is the output resistance of MOSFET and MOSFET has a resistance $(R_1\|R_2)$ in the source load

Therefore, we have

$$R_o = r_{o2} + (R_1\|R_2) + (g_m r_{o2})(R_1\|R_2)$$

$$= g_m r_{o2}(R_1\|R_2)$$

$$= 5\times10^{-3} \times 25\times10^{3} \times (10\|82)\times10^{3}$$

$$= 1114\ k\Omega$$

$$R_{of} = R_o(1+A_I\beta) = R_o D = 1114 \times 61.3 = 68.296\ M\Omega$$

Determining the Loop Gain

The loop gain $A\beta$ is an important quantity in feedback amplifiers. It determines whether the feedback amplifier is stable or not.

Alternative way for obtaining $A\beta$.

Consider the general feedback amplifier shown below (fig a) To find the loop gain we set the source x_s equal to zero, break the feedback loop at some point at amplifier input and apply a test signal x_t at this point as shown in fig (b).

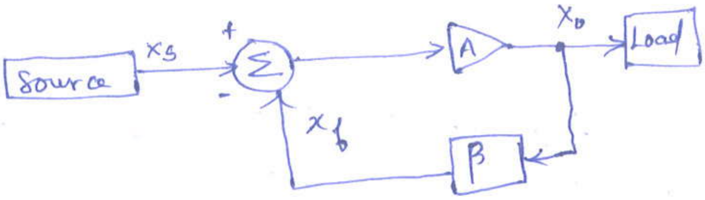


fig (a).

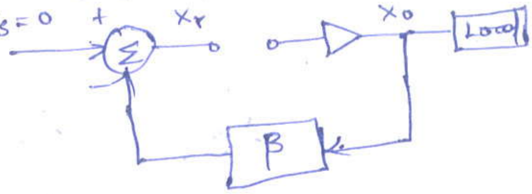


fig (b).

The return signal x_r is now $-A\beta x_t$. Therefore

$$A\beta = -\frac{x_r}{x_t} \qquad (1)$$

It is important to note that, the conditions that existed ~~that~~ prior to the loop being broken must remain unchanged after loop is broken. These conditions include: maintaining the same biasing conditions and maintaining the same resistance at the return point. To maintain such conditions it is necessary to insert an equivalent resistance at the point where the loop is broken as shown in fig (c)

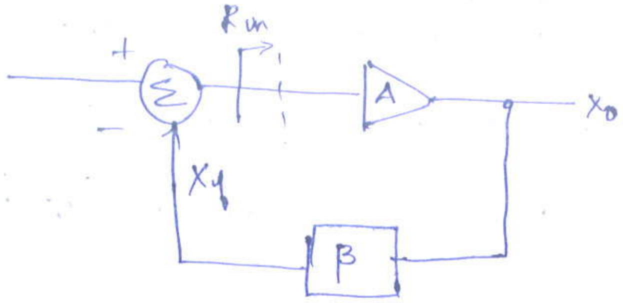


fig ①a

Equivalent impedance.

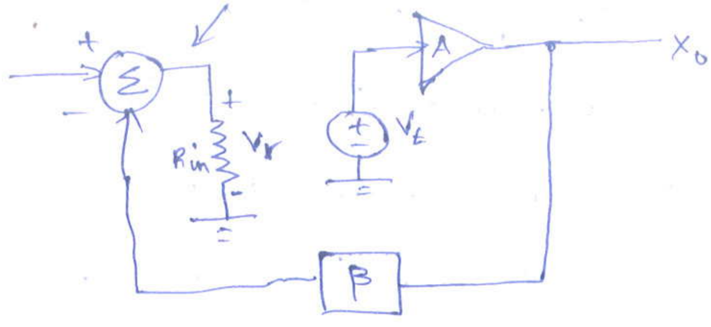


fig ①b

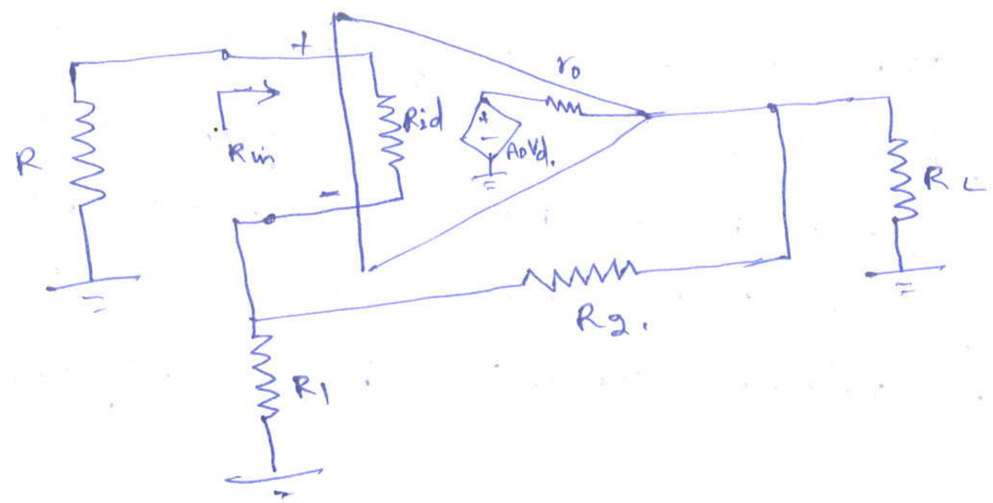
As shown in fig ①b, a test voltage, V_t is applied to the right-hand side terminal and a resistance R_{in} is inserted at the left had side terminals. The return voltage (V_r) is then measured at the output terminal. For this circuit, the loop gain is given by.

$$A\beta = -\frac{V_r}{V_t} \quad \text{——— ②}$$

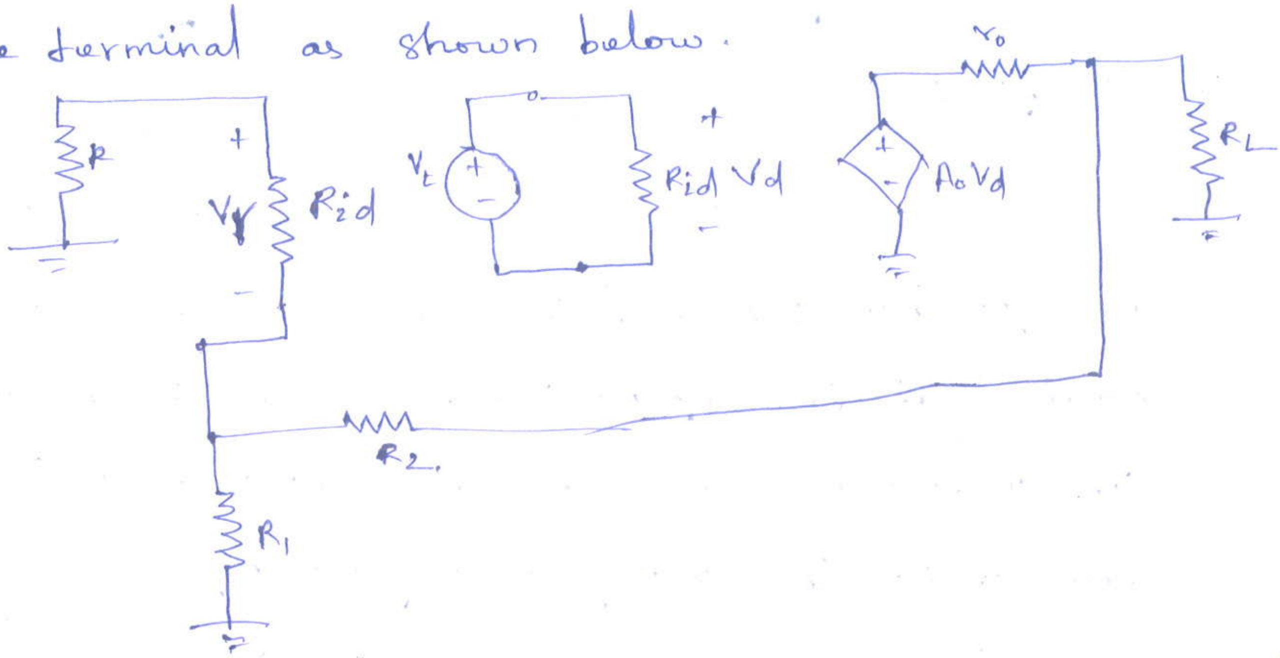
Instead of a test voltage, if we apply test current I_t and measure a return current I_r the loop gain will be

$$A\beta = -\frac{I_r}{I_t} \quad \text{——— ③}$$

Consider the feedback loop with op-amp shown below



The feedback loop is now broken at the op-amp input terminals. A test voltage is applied to the right hand side terminals and a resistance R_{id} is inserted at the left hand side terminal as shown below.



From the above fig we have

$$V_r = -A_o V_d \left(\frac{R_{id}}{R_{id}+R}\right)\left(\frac{R_1 \| (R_{id}+R)}{[(R_1 \| (R_{id}+R)) + R_2]}\right)\left(\frac{R_L \| [R_2 + R_1 \| (R_{id}+R)]}{R_L \| [R_2 + R_1 \| (R_{id}+R)] + r_o}\right)$$

once we calculate V_r, the loop gain can be determined by

$$A\beta = -\frac{V_r}{V_t} = -\frac{V_r}{V_d} \qquad \therefore V_t = V_d.$$

Significance of poles, zeros and Feedback loop.

The system poles are the roots of the denominator of $H(s)$. The system zeros are the roots of the numerator of $H(s)$. The system natural modes are the time functions corresponding to the poles. There is one natural mode per pole. The pole are determined by selecting the

the external excitation to zero and hence they are independent of the external excitation. Thus, poles of feedback amplifier depend only on the feedback loop. Because of this, it is possible to generate number of circuits having a same feedback loop with different transmission zeros. The transmission zero and the close loop gain of a feedback amplifier depend on how and where the input signal is applied to the loop.

Stability problem.

Transfer function of Feedback amplifier

The ideal close loop transfer function for the feedback amplifier is given by

$$A_f = \frac{X_o}{X_s} = \frac{A}{1+A\beta} \qquad \text{——— (1)}$$

Where A is the open-loop transfer function

β is the feedback transfer function.

Amplifier's components and parameters change with frequency and hence its open loop gain is a function of frequency. Considering this the ~~open~~ close-loop transfer function $A_f(s)$ is given by

$$A_f(s) = \frac{A(s)}{1+A(s)\beta(s)} = \frac{A(s)}{1+T(s)} \qquad \text{——— (2)}$$

Where $T(s)$ is the loop. gain.

For the negative feedback this loop gain should be positive. For physical frequencies, $s=j\omega$ and the loop gain $T(j\omega) = A(j\omega)\beta(j\omega)$ is a complex function. It can be represented by its magnitude and phase as follows

$$T(j\omega) = |T(j\omega)| \angle\phi \quad \text{--- (3)}$$

or $$A(j\omega)\beta(j\omega) = |A(j\omega)\beta(j\omega)| \angle\phi \quad \text{--- (4)}$$

Now, the close loop gain can be written as

$$A_f(j\omega) = \frac{A(j\omega)}{1+T(j\omega)} = \frac{A(j\omega)}{1+A(j\omega)\beta(j\omega)} \quad \text{--- (5)}$$

The stability of feedback amplifier is depend on the loop gain. The frequency at which phase angle ϕ becomes $180°$, the loop gain $A(j\omega)\beta(j\omega)$ will be a real number with negative sign. Thus at this frequency the feedback will become positive at this frequency if the magnitude of the loop gain is equal to unity ie $T(j\omega) = -1$, then from eq (5) we can say that at $T(j\omega) = -1$, the close-loop gain becomes infinity This means that the amplifier will have output for a zero input and it will behave like an oscullator.

Let us consider again the feedback amplifier with $X_s = 0$ and a presence of noise signal due to power-supply switching at the input of an amplifier. Such noise signal contains a wide range of frequencies. The noise frequency component ω with phase 180° is amplified and feedback at the input of the amplifier through feedback network. This feedback signal is given by

$$X_f = A(j\omega_{180})\,\beta(j\omega_{180})X_i = -X_i$$

The feedback signal, $X_f(-X_i)$ is further multiplied by -1 in summer block. This confirms the input of the amplifier is X_i. As a result, amplifier oscillates at frequency ω_{180}.

Nyquist plot.

Following figure shows the Nyquist plot. It can be used to test the stability of the feedback amplifier.

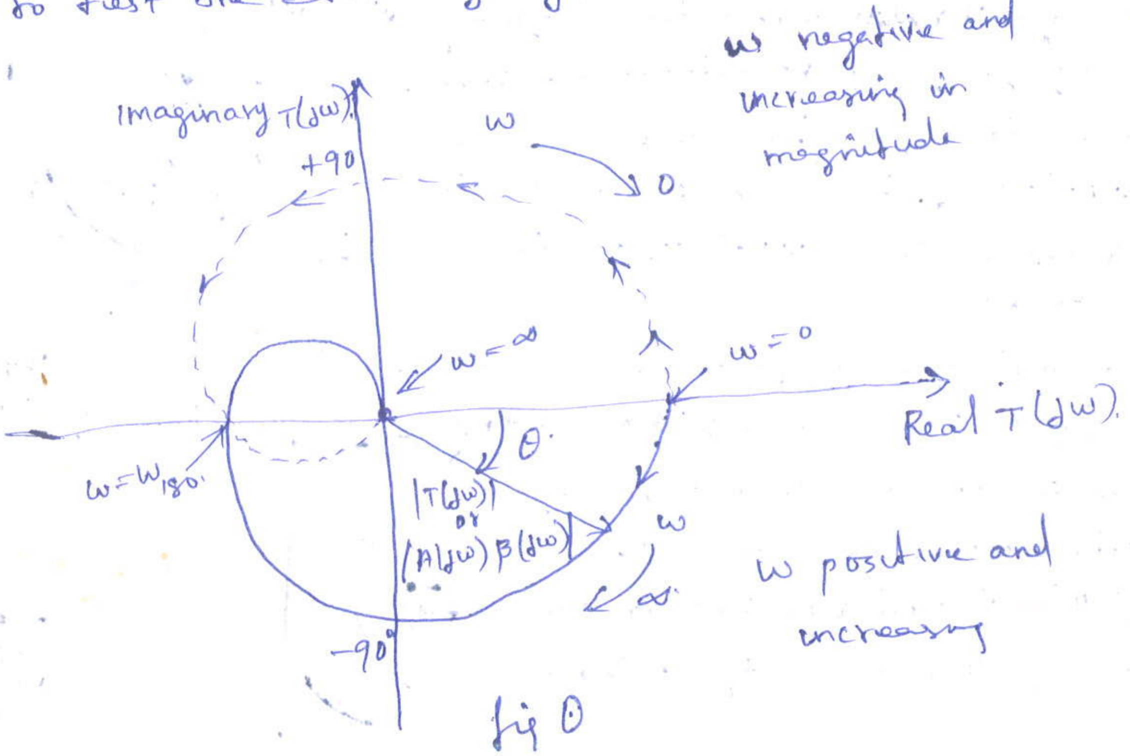
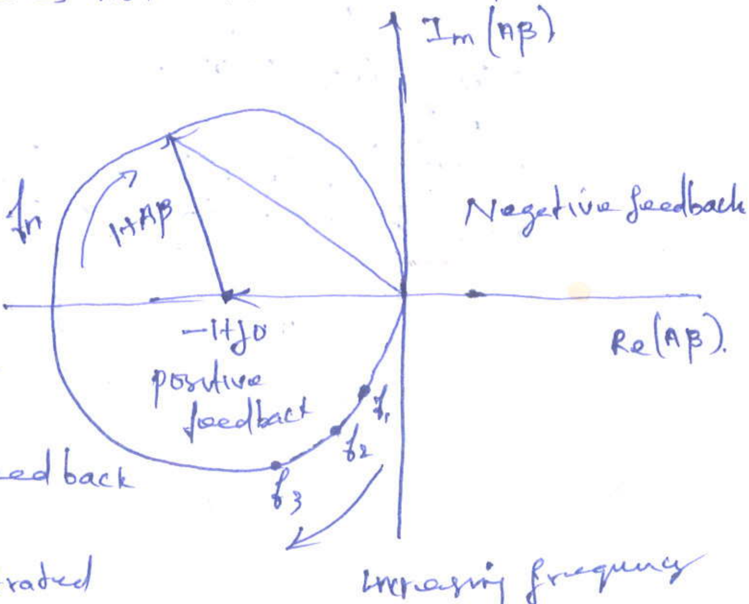


fig ①

The Nyquist plot not only determines if a system is stable, it also indicates the degree of system stability. As shown in fig ①, Nyquist plot is a plot of the real and imaginary component of a complex function $T(j\omega)$. It is a polar plot of loop gain with a frequency varies from minus infinity to plus infinity. The negative frequencies in the plot have no physical meaning. The negative frequency plot is a complex conjugate of the polar plot for positive frequencies. The plot for positive frequencies is drawn with a solid line and plot for negative frequencies is drawn with dashed line.

The criterion of Nyquist is that 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 as shown in fig below.



Locus of $|1+A\beta| = 1$.

Fig ① shows the locus of $|1+A\beta| = 1$. It is a circle of unit radius, with center at $-1+j0$. If for any frequency, $A\beta$ extends outside this circle the feedback is negative since $|1+A\beta| > 1$ if, however, $A\beta$ lies within this circle, then $|1+A\beta| < 1$, and the feedback is positive.

An example of the Nyquist criterion is illustrated in fig ②. The locus in fig ②a is stable since it does not enclose the $-1+j0$ point, whereas the locus shown in fig ② b is unstable since the curve does enclose the $-1+j0$ point.

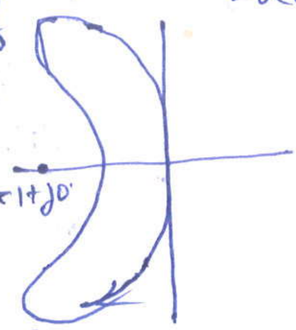


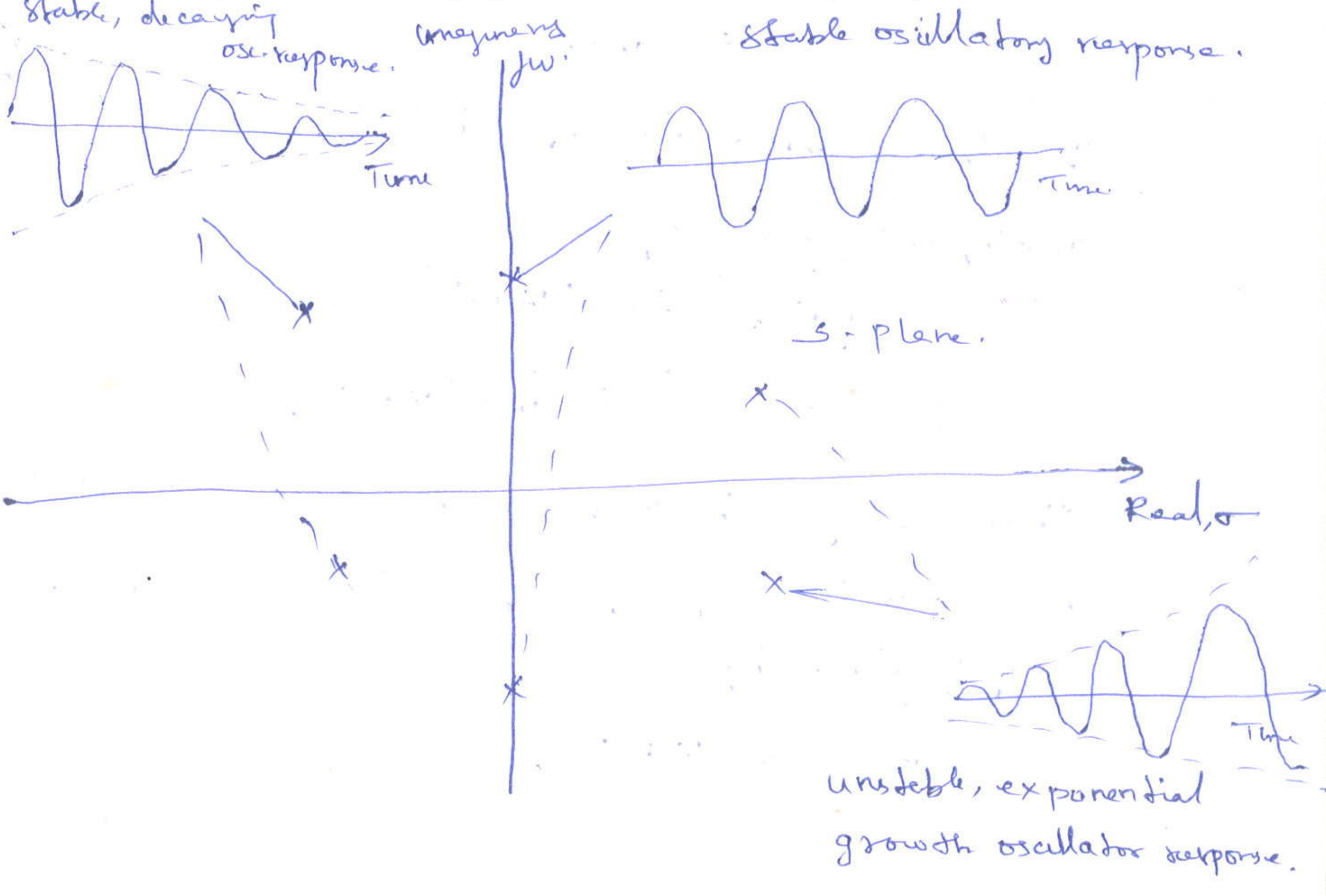
fig ②a stable.

fig ②(b) unstable.

Effect of Feedback on amplifier poles.

The frequency response of an amplifier and its stability can be determined directly by its poles.

For an amplifier to be stable, it is necessary that its poles should lie in the left half of the s plane. A pair of complex-conjugate poles on the jω (imaginary) axis gives rise to sustained oscillations. If poles are to the left half of the jω (imaginary) axis, the system is globally stable and they give decaying oscillations. On the other hand, if poles are to the right half due of the jω (imaginary) axis the system is unstable and they give rise to growing oscillation as shown in the figure below.



By solving the characteristic equation of the feedback loop given below, we obtain the poles of the feedback amplifier.

$$1 + A(s)\beta(s) = 0$$

The feedback affects the poles of amplifier.

Amplifier with single pole response.

The open loop gain of an amplifier is flat from DC to what is referred to as the dominant pole. From there it falls off at 6 dB/octave or 20 dB/decade. This is referred to as single pole response. The open-loop transfer function for an amplifier with a single pole is given by

$$A(s) = \frac{A_0}{1 + s/\omega_p}.$$

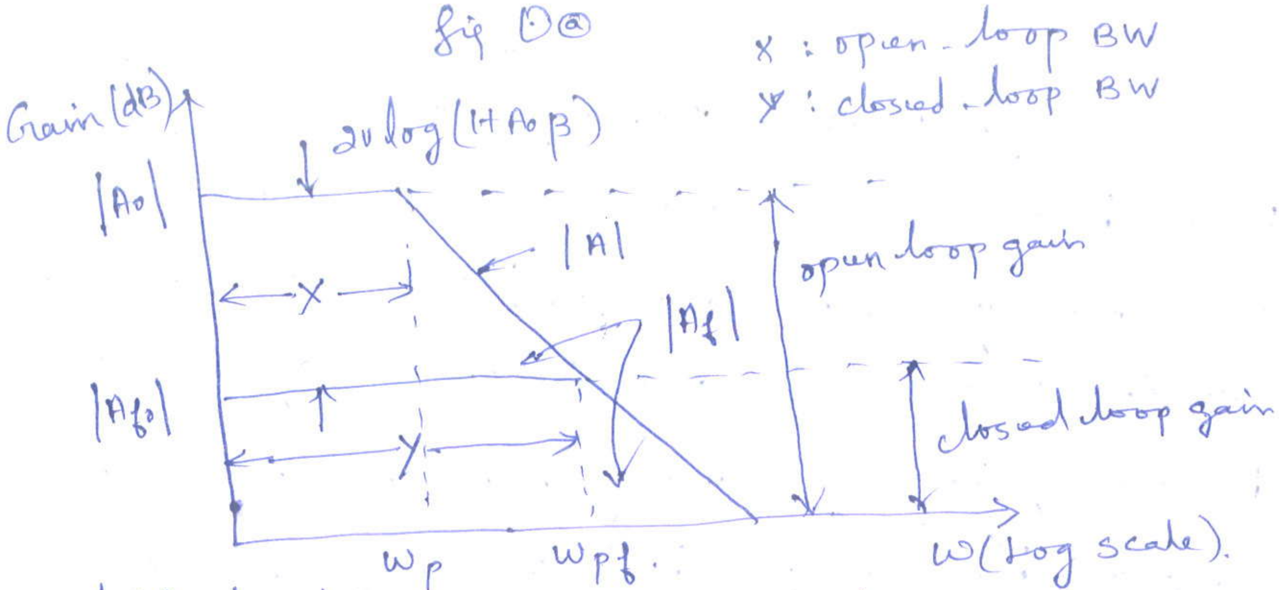
The closed loop transfer function for a feedback amplifier is given by

$$A_f(s) = \frac{A_0/(1+A_0\beta)}{1 + s/\omega_p(1+A_0\beta)}.$$

Thus the feedback moves the pole along the negative real axis to a frequency ω_{pf}. The frequency ω_{pf} is given by.

$$\omega_{pf} = \omega_p(1+A_0\beta).$$

fig 1 a



open-loop and closed loop freq response of an amplifier.

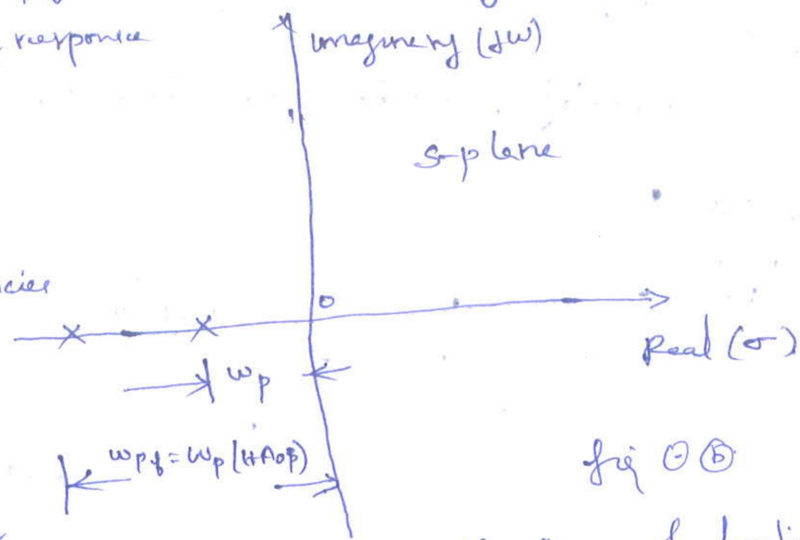


fig 1 b

Effect of feedback on pole location.

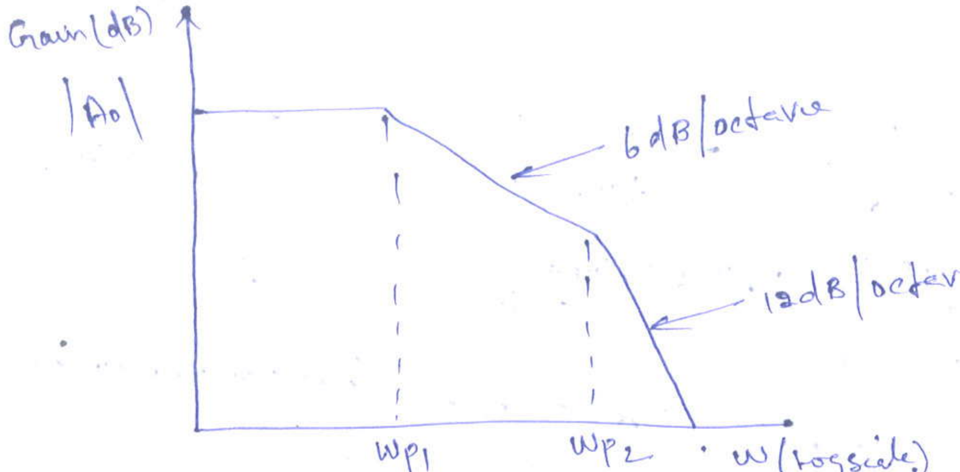
From fig 1 a it can be observed that at low frequencies the difference between plot of open-loop gain and closed loop gain is 20 log (1+A0β). However, these two plots concide at high frequencies. This can be presented by

$$A_f(s) \approx \frac{A_0 \omega_p}{s} \approx A(s)$$

$$\because \omega >> \omega_p (1+A_0\beta).$$

From fig 1 a, it is clear that when negative feedback is applied to an amplifier it gain reduces; however, its bandwidth increases. With negative feedback, the pole of the amplifier is shift towards left from imaginery axes and it never enters the right half of the s-plane. Because of this a single-pole amplifier is stable for any value of β and hence such an amplifier is also known as unconditionally stable amplifier. It is also important to note that with a single pole, the phase lag never exceeds 90°, Because of this the loop gain never achieves the 180° phase Shift and hence the feedback is always negative.

Amplifier with Two-pole frequency.



The open-loop transfer function for an amplifier with two poles is given by

$$A(s) = \frac{A_0}{\left(1 + s/\omega_{p1}\right)\left(1 + s/\omega_{p2}\right)}$$

$\therefore$, the characteristics equation $1 + A_s(\beta)$ takes the form

$$s^2 + s(\omega_{p1} + \omega_{p2}) + (1 + A_0\beta)\,\omega_{p1}\,\omega_{p2} = 0$$

By solving this quadratic equation the close-loop are

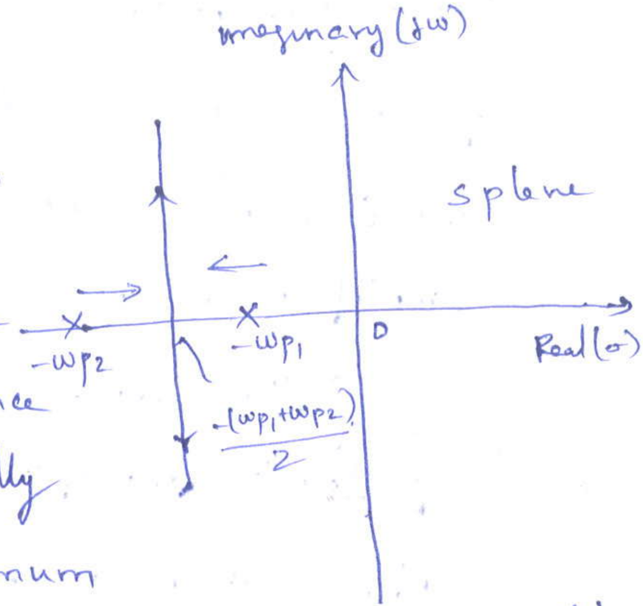
$$s = \frac{-(\omega_{p1} + \omega_{p2}) \pm \sqrt{(\omega_{p1} + \omega_{p2})^2 - 4(1)(1 + A_0\beta)\,\omega_{p1}\,\omega_{p2}}}{2}$$

$$= \frac{-(\omega_{p1}+\omega_{p2})}{2} \pm \frac{1}{2}\sqrt{(\omega_{p1}+\omega_{p2})^2 - 4(1+A_0\beta)\omega_{p1}\omega_{p2}}$$

From the above equation we can see that as loop as increases, the poles move close to each other and at point

$$s = \frac{-(\omega_{p1}+\omega_{p2})}{2}.$$

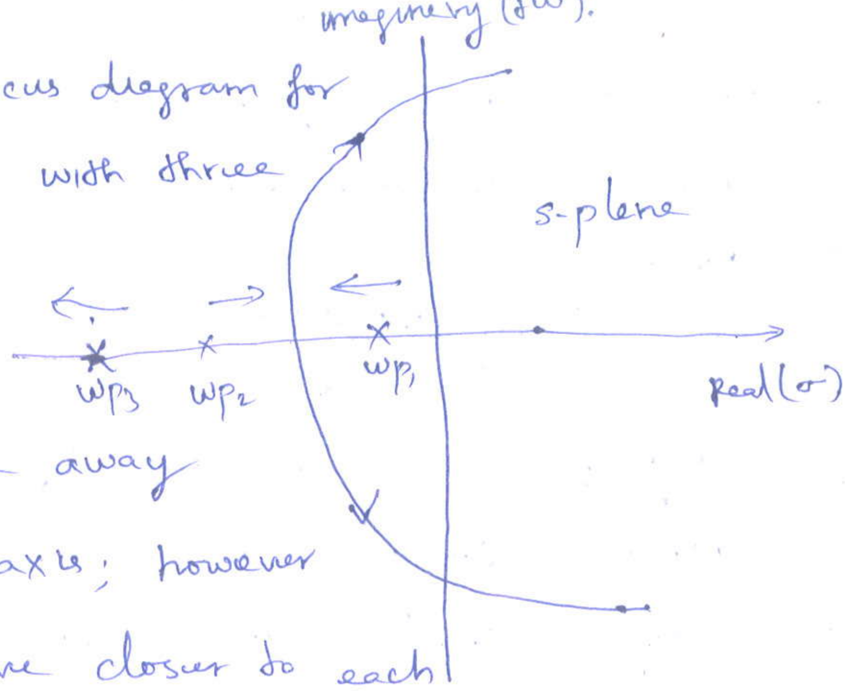
they become coincident. If the loop gain is further increased, the poles becomes complex conjugate and move along the vertical line as shown in the figure. Such a locus of the poles for increasing loop gain is known as root locus diagram



For a feedback amplifier with two poles, the poles are either negative real or complex conjugate with negative real part and hence this amplifier is also unconditionally stable. We know that, maximum phase-shift for two pole amplifier is 180° (90° per pole). However, this phase shift is reached at $\omega = \infty$. So there is no finite frequency at which the phase shift is 180° and hence the feedback of amplifier is always negative for entire frequency range.

Amplifiers with Three or More poles.

Fig. shows a root locus diagram for a feedback amplifier with three poles. As loop gain increases, the highest frequency pole move away from the imaginary axis; however two other poles move closer to each other and become coincident. If loop gain is further increased, the poles becomes complex conjugate and then move into the right half of s-plane causing amplifier to become unstable.



In other words, we can say that if a feedback amplifiers has more than two poles, the phase angle of the loop gain could exceed -180° beyond certain frequency and amplifier can be unstable.

Frequency compensation

If a feedback amplifier has more than two poles it can be unstable. The technique used to make unstable feedback amplifier stable is called frequency compensation technique. Here, the open-loop transfer function A(s) of an amplifier is modified by introducing compensating network to make the amplifier stable.

Dominant pole compensation

In this compensation technique a dominant pole is introduced in to the amplifier so that phase shift is less then -180° when the loop gain is unity. Consider a feedback amplifier with three-poles having a transfer function

$$A(s) = \frac{A_0}{(1+s/\omega_{p1})(1+s/\omega_{p2})(1+s/\omega_{p3})}$$

Where $\omega_{p1} < \omega_{p2} < \omega_{p3}$. By adding a new dominant pole ~~sub~~ such that $\omega_D < \omega_{p1} < \omega_{p2} < \omega_{p3}$ the transfer function becomes.

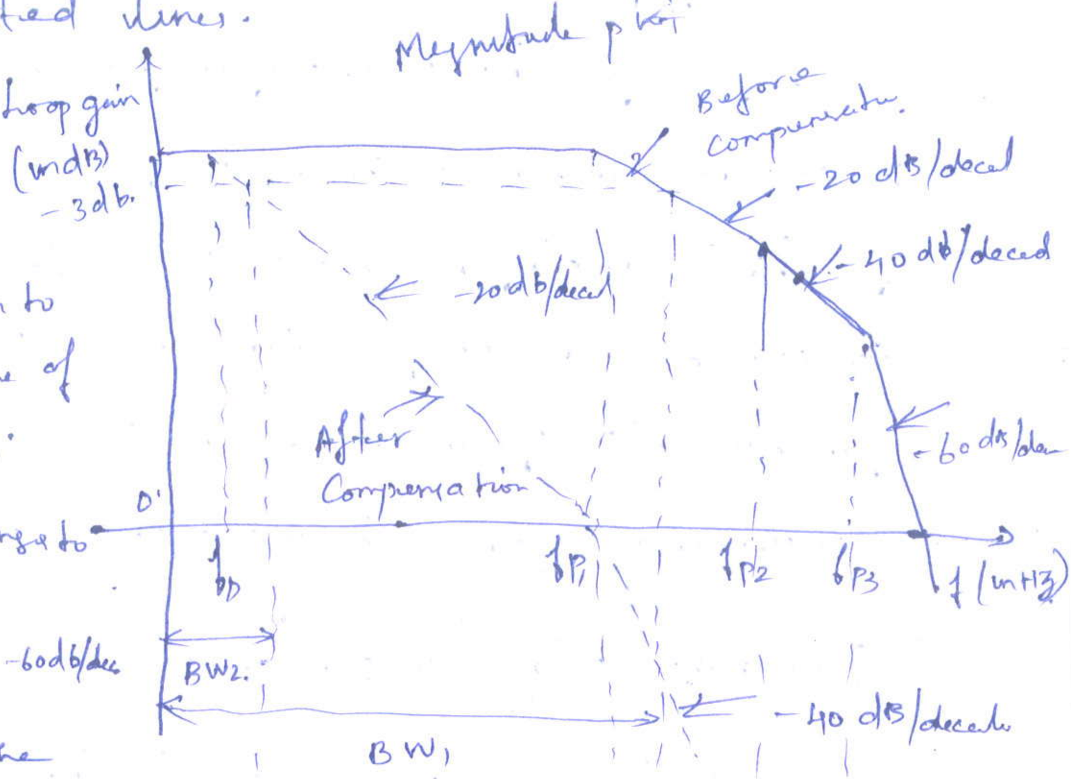
$$A'(s) = \frac{A_0}{(1+s/\omega_D)(1+s/\omega_{p1})(1+s/\omega_{p2})(1+s/\omega_{p3})}$$

Fig. shows the bode plot of this modified transfer function by dotted lines.

The introduction of the dominant pole causes the loop gain to roll-off with a slope of -20 dB/decade at f_D.

At f_{P_1} the slope changes to -40 dB/decade

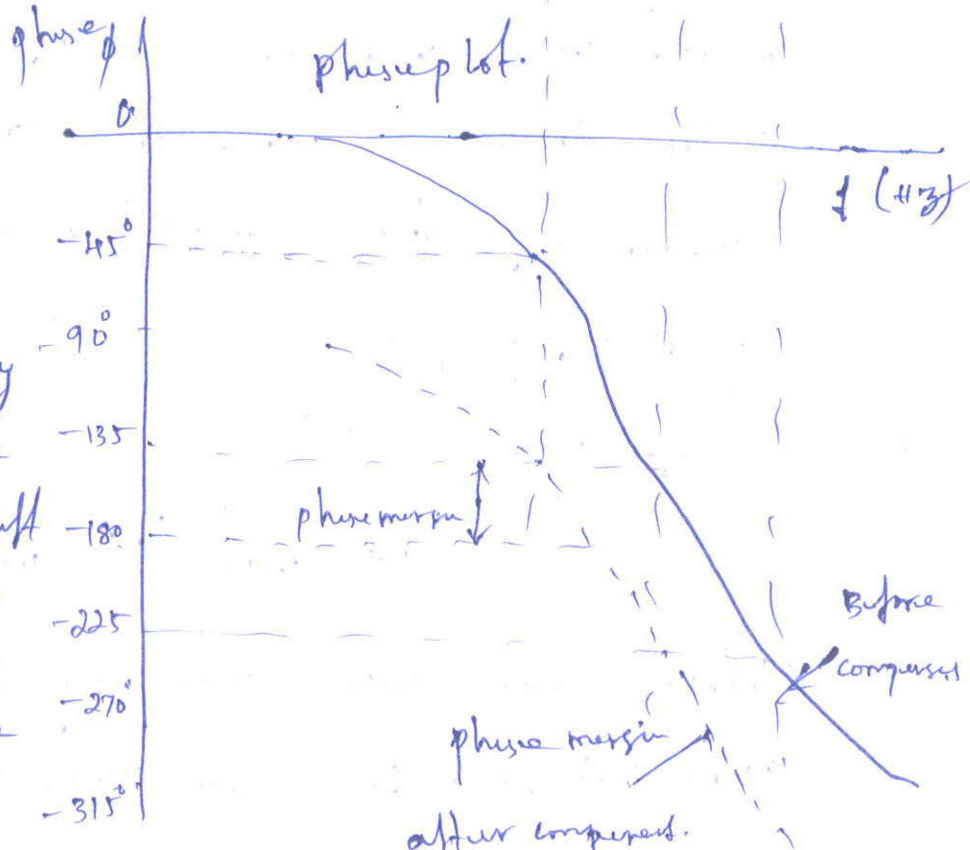
At f_{P_2} it changes to -60 db/dec



The frequency f_D of the dominant pole is chosen so that the loop gain is unity at frequency f_{P_1}.

Because of this the phase shift at frequency f_{P_1} due to the dominant pole f_D is $-90°$ and the phase shift due to first pole f_{P_1} is $-45°$.

At $f = f_{P_1}$, the total phase shift is $-135°$ and the phase margin is $\phi_m = 180 - |\phi| = 180 - 135 = 45°$. The phase margin is defined as the amount of degrees by which the phase angle ϕ of the loop gain fall short of $\pm 180°$ when the magnitude of loop gain is unity.



For adequate relative stability the phase margin should be least 45°. Thus by adding dominant pole it is possible to make amplifier stable.

It can be observed from the magnitude plot that 3db bandwidth for compensated amplifier (BW2) is considerably smaller then the 3db bandwidth for non compensated amplifier (BW1).

pbm.

The voltage gain of an amplifier without feedback is 3000. Calculate the voltage gain of the amplifier if negative voltage feedback is introduced in the circuit. Given that feedback fraction $\beta = 0.01$

$$A = 3000 \quad , \quad \beta = 0.01$$

Voltage gain with negative feedback is

$$A_f = \frac{A}{1+A\beta} = \frac{3000}{1+3000 \times 0.01} = 97.$$

* The overall gain of a multistage amplifier is 140. When negative voltage feedback is applied, the gain is reduced to 17.5. Find the fraction of the output that is fedback to the input.

$$A = 140 \qquad A_f = 17.5$$

Let β be the feedback fraction. Voltage gain with negative feedback is

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

$$17.5 = \frac{140}{1+140\beta}.$$

$$17.5 + 2450\beta = 140$$

$$\beta = \frac{140 - 17.5}{2450} = 1/20$$

* When negative voltage feedback is applied to an amplifier of gain 100, the overall gain fall to 50

(i) calculate the fraction of the output voltage feedba

(ii) if this fraction is maintained, calculate the value of the amplifier gain required if the overall stage gain is to be 75.

(i) Gain without feedback $A = 100$

Gain with feedback $A_f = 50$.

Let β be the fraction of the output voltage feedback

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

$$50 = \frac{100}{1 + 100\beta}$$

$$50 + 500\beta = 100$$

$$\beta = \frac{100 - 50}{5000} = 0.01$$

(ii) $A_f = 75$ $\beta = 0.01$ $A = ?$.

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

$$75 + \frac{A}{1 + 0.01A}$$

$$75 + 0.75A = A$$

$$A = \frac{75}{1 - 0.75} = 300.$$

The gain of an amplifier without feedback is 50 whereas with negative voltage feedback, it falls to 25. If due to ageing, the amplifier gain falls to 40, find the percentage reduction in stage gain (i) without feedback and (ii) with negative feedback.

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

$$25 = \frac{50}{1 + 50\beta}$$

$$\beta = 1/50.$$

(i) without feedback

The gain of the amplifier without feedback is 50. However due to ageing it falls to 40.

% of reduction in stage gain $= \frac{50 - 40}{50} \times 100 = 20\%$

(ii) with negative feed back

gain without feedback was 50, the gain with neg feedback was 25. Now the gain without feedback falls to 40

$$A_f = \frac{A}{1 + A\beta} = \frac{40}{1 + (40 \times 1/50)} = 22.2.$$

% reduction in stage gain $= \dfrac{25 - 22.2}{25} \times 100$

$= 11.2\%$

* An amplifier has a voltage amplification A and a fraction β of its o/p is fedback in opposition to the input. If $m_v = 0.1$ and $A_v = 100$, calculate the percentage change in the gain of the system if A falls 6 db due to ageing

* An amplifier has an open-loop gain of 1000 and a feedback ratio of 0·04. If the open-loop gain changes by 10% due to temperature, find the percentage change in gain of the amplifier with feedback.

$$A = 1000, \quad \beta = 0{\cdot}04 \quad \text{and} \quad \frac{dA}{A} = 10$$

WKT the % change in gain of the amplifier with feedback is

$$\frac{dA_f}{A_f} = \frac{dA}{A} \frac{1}{(1+A\beta)} = 10 \times \frac{1}{1+1000 \times 0{\cdot}04} = 0{\cdot}25\%$$

* An amplifier has voltage gain with feedback of 100. If the gain without feedback changes by 20% and the gain with feedback should not vary more than 2% determine the value of open-loop gain A and feedback ratio β.

$$A_f = 100, \quad \frac{dA_f}{A_f} = 2\% = 0{\cdot}02 \quad \text{\&} \quad \frac{dA}{A} = 20\% = 0{\cdot}2$$

WKT

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

$$0{\cdot}02 = 0{\cdot}2 \times \frac{1}{1+A\beta}$$

$$\therefore \quad (1+A\beta) = \frac{0{\cdot}2}{0{\cdot}02} = 10$$

The gain with feedback is.

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

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

$$A = \cancel{10,000}\ 1000$$

$$\therefore\quad 1+A\beta = 10$$

$$A\beta = 9$$

$$\beta = \frac{9}{1000} = 0.009$$

* An amplifier has a midband gain of 125 and a bandwidth 250 KHz.
  ① if 4% negative feedback is introduced, find the new bandwidth and gain.
  ② If the bandwidth is to be restricted to 1MHz find the feedback ratio.

Given $A = 125$ $\quad BW = 250\ KHz$ $\quad \beta = 0.04$

① $A_f = \dfrac{A}{1+\beta A} = \dfrac{125}{1+0.04 \times 125} = 20.83$

$$BW_f = BW \times (1+\beta A) = 250\times10^3 \times (1+0.04\times125)$$

$$= 1.5\ MHz$$

②

$$BW_f = 1MHz$$

$$1\times10^6 = BW \times (1+\beta A)$$

$$1\times10^6 = 250\times10^3(1+\beta\times125)$$

$$\beta = 0.024.$$

* A single stage RC coupled amplifier has a midband gain of 1000 is made into a negative feedback amplifier by feeding 10% of the output voltage in series with input opposing.

(i) What is the ratio of half power frequencies with feedback to those without feedback?

(ii) If $f_L = 20Hz$ and $f_H = 50KHz$ for the amplifier without feedback, find the corresponding values of f if feedback is incorporated.

$A = 1000 \quad \beta = 0.1$

(i) $$\frac{f_{Hf}}{f_H} = 1+\beta A = 1+0.1\times1000 = 101$$

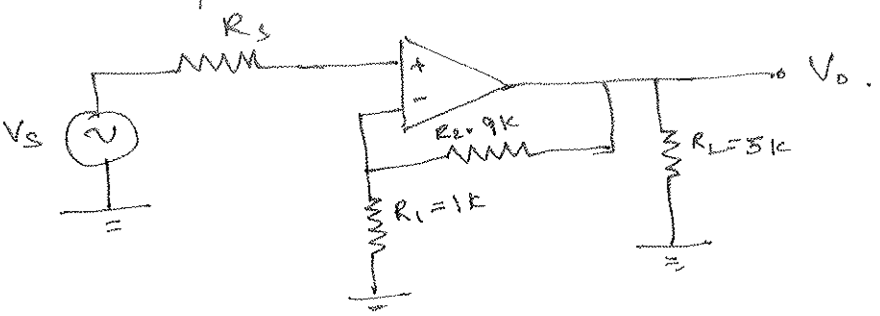
$$\frac{f_{Lf}}{f_L} = \frac{1}{1+\beta A} = \frac{1}{1+0.1\times1000} = \frac{1}{101} = 0.0099$$

(ii) with $f_L = 20Hz$ and $f_H = 50KHz$

$$f_{Lf} = 2\times0.0099 = 0.198Hz$$

$$f_{Hf} = 50KHz\times101 = 5.05MHz$$

* For a non inverting op-amp ckt shown below $R_1 = 2k$ and $R_2 = 8k$. For open-loop configuration, op-amp circuit has gain = 1000, lower 3dB frequency = 10Hz and upper 3dB frequency = 1KHz. Find the close loop gain close loop lower cut-off and upper cut of freq.



$$\beta = \frac{R_1}{R_1+R_2} = 0.2$$

$$A_f = \frac{A}{1+A\beta} = 4.975$$

$$f_{Lf} = \frac{f_L}{D} = \frac{f_L}{1+A\beta} = 0.04975\,Hz$$

$$f_{Hf} = f_H \times D = 201\,KHz.$$

Consider a SNR improvement circuit having output noisy power amplifier stage with gain 5, a small signal amplifier as a pre amplifier with gain 1000 and a feedback network with $\beta = 0.1$. If $V_s = 1V$ and $V_n = 0.5V$, calculate the signal and noise voltages at the output and improved SNR. Also calculate total o/p voltages.

Given $A_p = 1000$, $A = 5$, $\beta = 0.1$, $V_s = 1V$ & $V_n = 0.5V$

Signal o/p voltage $V_{os} = \frac{V_s A_p \cdot A}{1+A_p A\beta} = \frac{1 \times 1000 \times 5}{1+1000 \times 5 \times 0.1}$

$= 9.98V$

Noise o/p voltage $V_{no} = \frac{V_n A}{1+A_p A\beta} = \frac{1 \times 5}{1+1000 \times 5 \times 0.1} = 0.00998$

$$SNR = \frac{V_{so}}{V_{no}} = \frac{9.98}{0.00998} = 1000.$$

$= 60dB$

$V_o = V_{os} + V_{no} = 9.98 + 0.00998 = 9.98998V$

The gain and distortion of an amplifier are 100 and 4% respectively. If a negative feedback with $\beta = 0.3$ is applied, find the new distortion in the system

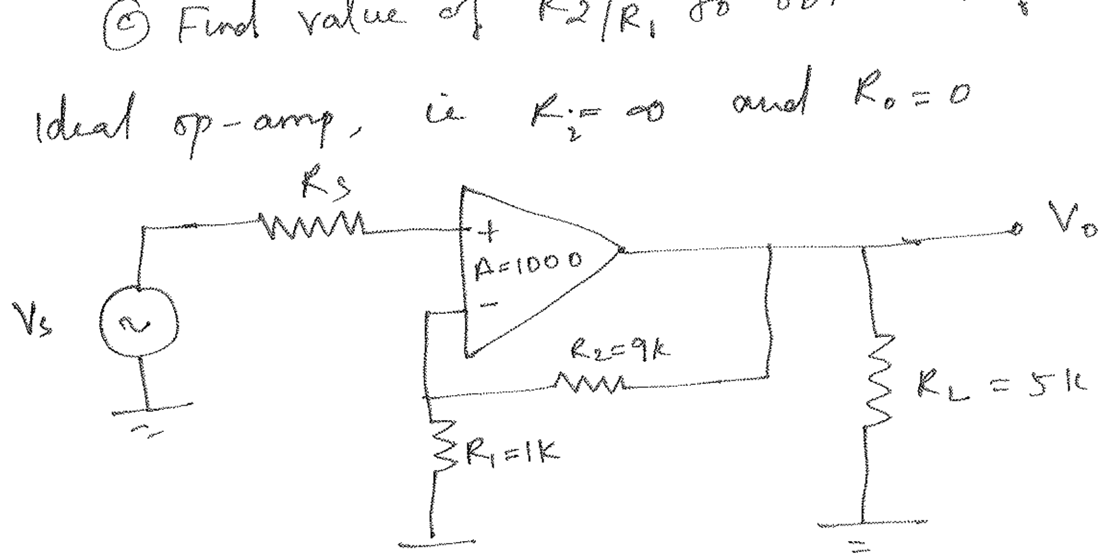
$$D = 1 + \beta A = 1 + 0.3 \times 100 = 31$$

$$\text{Distortion} = \frac{4\%}{31} = 0.129\%$$

Consider the noninverting op-amp circuit shown below

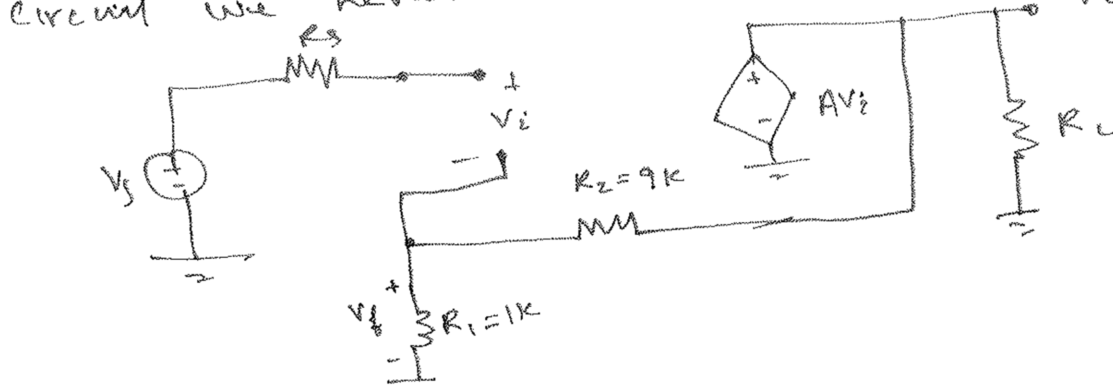
(a) Find β (b) Find A_f if A increases by 25%.

(c) Find value of R_2/R_1 to obtain $A_f = 50$. Assume ideal op-amp, ie $R_i = \infty$ and $R_o = 0$



Solution:

Replacing the op-amp with its equivalent circuit we have.



ⓐ $V_f = \beta V_0$

$= \dfrac{R_1}{R_1 + R_2} \cdot V_0$ $\qquad\qquad \beta = \dfrac{R_1}{R_1 + R_2}$

$= \dfrac{1k}{1k + 9k} = 0.1$

ⓑ $A = A \times 1.25 = 1250.$

$A_f = \dfrac{A}{1 + A\beta} = \dfrac{1250}{1 + 1250 \times 0.1} = 9.92$

ⓒ $A_f = \dfrac{A}{1 + A\beta}$

$50 = \dfrac{1000}{1 + 1000\beta}$

$\beta = 0.019$

$\beta = \dfrac{R_1}{R_1 + R_2}.$

$1/\beta = \dfrac{R_1 + R_2}{R_1} = 1 + \dfrac{R_2}{R_1} = \dfrac{1}{0.019} = 52.632$

$\dfrac{R_2}{R_1} = 51.632$

unit II : Oscillators.

Oscillators

classification, Barkhausen Criterion - Mechansim for start of oscillation and stabilization of amplitude, General form of an Oscillator, Analysis of LC oscillators - Hartley, colpitts, clapp, Franklin, Armstrong, Tuned collector oscillators, RC oscillators - phase shift - Wienbridge - Twin-T oscillators, Frequency range of RC and LC oscillators, Quartz Crystal construction, Electrical equivalent circuit of crystal, Miller and Pierce oscillators, frequency stability of oscillators.

Introduction:

Many electronic devices requires a source of energy at a specific frequency which may range from a few Hz to several MHz. This is achieved by an electronic device called oscillator. Oscillator are extensively used in electronic equipment.

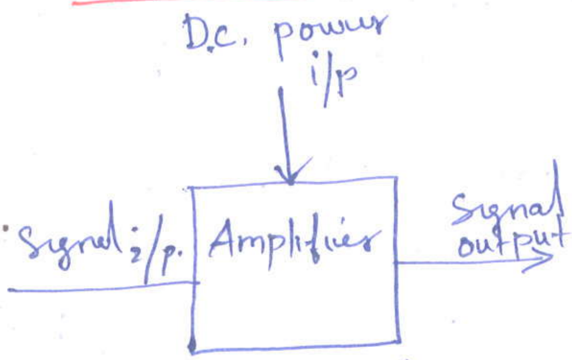
For example:

In radio and television receivers, oscillators are used to generate high frequency wave (carrier wave) in the tunning stages. oscillators are widely used in radar, electronic computers and other electronic devices.

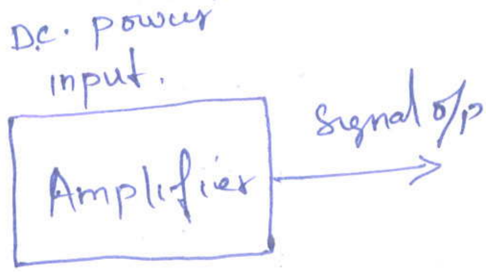
Oscillator

An electronic device that generates (sinusoidal) or any forms oscillations of desired frequency is known as oscillators.

Comparison between an Amplifier and an Oscillator



(A) Amplifier



(B) Oscillator.

The above fig (A) show the block diagram of an amplifier while (B) shows that of an oscillator. We know that an amplifier is a device, which produces an output signal with a similar waveform as that of the input. But its power level is generally high. This additional power is supplied by an external d.c. source. Thus an amplifier is essentially an energy convention device, i.e., a device, which gets energy from the d.c. source and converts it into an a.c. energy at the same frequency as that of the input signal. The d.c to a.c conversion is controlled by the i/p signal. It means that if there is no i/p signal the no energy convertion takes place. Thus there is no o/p signal

An oscillator is a device, which produces an output signal, without any input signal of any desired frequency. It keeps producing an output signal, so long as the d.c. power is supplied. It is evident from the above discussion that an oscillator does not require any external signal to start or maintain energy conversion process.

Classification of oscillators.

The electronic oscillators may be broadly classified into the following categories:

① Sinusoidal or Harmonic oscillators

The oscillators, which provide an output having a sine waveform, are called sinusoidal or harmonic oscillators. Such oscillators can provide output at frequencies ranging from 20HZ to 1GHZ.

② Non-Sinusoidal or Relaxation oscillators.

The oscillators, which provide an output having a square, rectangular or sawtooth waveform, are called non-sinusoidal or relaxation oscillators. Such oscillators can provide output at frequencies ranging from zero to 20 MHZ.

The sinusoidal oscillators may be further sub-divided into the following types.

ⓐ Tuned circuit oscillators:

These oscillators use a tuned-circuit consisting of inductors (L) and capacitors (C) and are used to generate high frequency signals. Thus they are also known as radio frequency (R.F) oscillators. Such oscillators are

Hartley oscillator
Colpitts oscillator
Clapp - oscillator.

ⓑ RC oscillators

These oscillators use resistors and capacitors and are used to generate low or audio-frequency signals. Thus they are also known as audio-frequency (A.F.) oscillators. Such oscillators are

phase shift oscillator
Wien-bridge oscillator

ⓒ Crystal oscillators.

These oscillators use quartz crystals are are used to generate highly stabilized output signal with frequencies up to 10MHz. Such oscillator is

crystal oscillator.

(d) Negative - resistance oscillators.

These oscillators use negative - resistance char's of the devices such as tunnel diodes. A tuned diode oscillator is an example of negative resistance oscillator.

Barkhausen Criterion

As we known in Feedback amplifiers, that the overall voltage gain of a positive feedback amplifier is given by the expression

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

A = Gain of an amplifier without feedback also called open loop gain &

$\beta \cdot A$ = product of feedback fraction and open loop-gain it is called loop gain.

If the product $(\beta \cdot A)$ is made equal to unity, then the denominator of the equation is zero. It means that the gain (A_f) will increase to infinity. of cource the o/p of the feedback amplifier, in actual practice cannot be infinite

∴ the condition (ie, $1 - \beta \cdot A = 0$) represents that there will be an output voltage, whose frequency is completely different from the input signal.

In otherwords, the circuit has stopped amplifying and started oscillating. It may be noted that the oscillations will not be maintained, if the value of $\beta \cdot A$ is less then unity.

WKT, an amplifier reverses the phase of an input signal at its output. It means that an amplifier causes a phase shift of $180°$ b/w the i/p & o/p signals. In order to provide positive feedback, the feedback network must provide a phase shift of $180°$, so as to provide a signal with a phase shift of $360°$ or $0°$ at the amplifier input.

The above two conditions for positive feedback ie ($\beta \cdot A = 1$ & the net phase shift around the loop equal to $360°$ or $0°$) are called Barkhausen criterion for oscillation. Mathematically, the Barkhausen criterion may be stated as follows.

$$\beta \cdot A = 1$$

$$\angle \beta \cdot A = 0°$$

Thus if $\beta \cdot A$ is a complex quantity, then its real part must be equal to unity and imaginary part equal to zero.

However, an oscillator in which the quantity $\beta \cdot A$ is exactly unity, is not realizable in practice.

In every pratical oscillator, the quantity $\beta \cdot A$ is slightly larger than unity. Thus an oscillator circuit must satisfy the two conditions mentioned above to produce sustained oscillations. The conditions are ⓐ loop gain ($\beta \cdot A \geq 1$) and ⓑ phase shift b/w the i/p and o/p signals must be equal to $360°$ or $0°$. It may be noted that if a circuit does not satisfy any one of these conditions, the oscillations will not be produced.

Mechanism for start of oscillation

Oscillator is a circuit which self gernerates some waveform like sine, triangular and square wave etc. It is basically an amplifier circuit with positive feedback introduced, through the feedback components like resistance, capacitance (RC), inductance capacitance (LC) or crystal circuits are used.

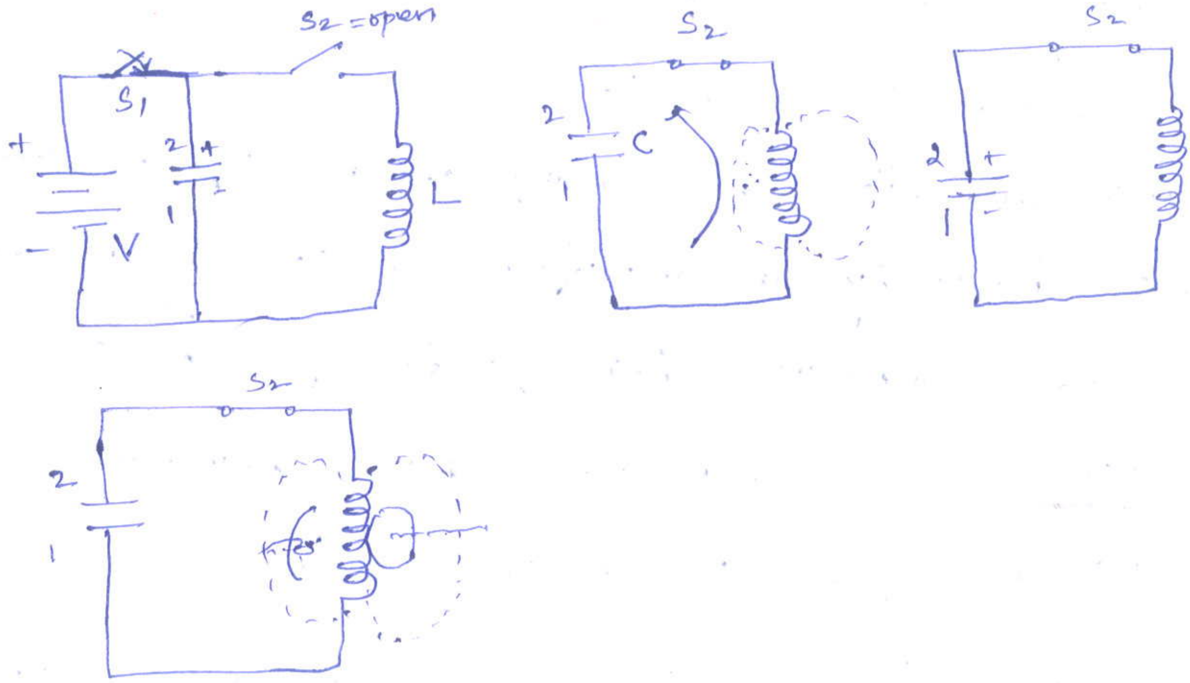
Oscillator circuits has two reactive elements inductance (L) and capacitance (C). This LC circuit is known as tank circuit. Both elements are capable of storing electrical energy.

* Inductors stores energy in the form of magnetic field whenever current flows through it.

* Capacitance stores energy in its electric field whenever a voltage is applied across the plates. we assume that they don't have any power loss.

<u>circuit operation (LC Tank circuit).</u>

Let us consider the tank circuit excited by dc source as shown below. Initially, the switch S_1 is closed, S_2 is open thus a capacitor be charged towards a maximum voltage from a dc voltage source with a polarity as shown in fig below.



After reaching the maximum voltage the switch 'S_2' is closed, and S_1 is open. Now the capacitance 'C' discharges through inductance, the discharging is due to flow of electrons from 1 to 2 through the coil L as given by arrow.

so conventional current direction is from 2 to 1. This current flow setup a magnetic field in the coil and it stores the energy released by the electric field. So, because of this inductive effects, current increases to maximum value, it occurs when 'c' is fully discharged.

At this instant electric field in 'c' is zero and magnetic field in the inductance is maximum. so, Now electric field is converted to magnetic field. Once capacitor is completely discharged, magnetic field begins to collapse and produce a counter emf.

According to Len's Law, the counter emf keeps electron moving in same direction. So capacitance starts to charge, in opposite direction. When capacitance is charged completely, magnetic field is collapsed completely. so now magnetic field is converted to electric field.

Now capacitance discharges in opposite direction. so electrons move from 2 to 1. Electric field is collapsing and magnetic field inbuilt.

The mentioned sequence of charging and discharging of capacitor results in an alternating motion of electrons as an alternating current. So energy is alternately stored in the electric field of the capacitor and magnetic field of the coil. The interchange of energy between capacitance and inductance is continuously repeated and this results in the production of damped electrical oscillations.

In order to make the oscillations in the tank circuit undamped the following conditions must be fulfilled.

(I) The amount of energy supplied should be such as to meet the losses in the tank circuit.

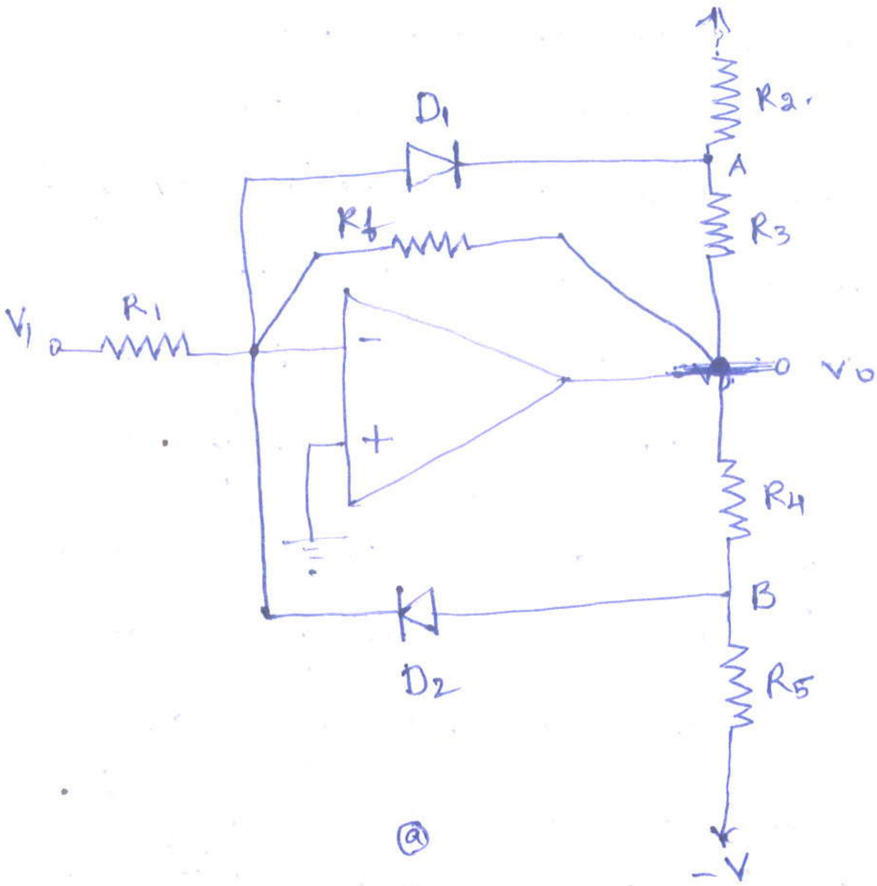
(II) The applied energy should be inphase with oscillations setup in the tank circuit.

(III) The frequency of the energy ~~is~~ supplied to the tank circuit should be same as that of oscillation produced by it.

Stabilization of amplitude.
(Nonlinear amplitude control)

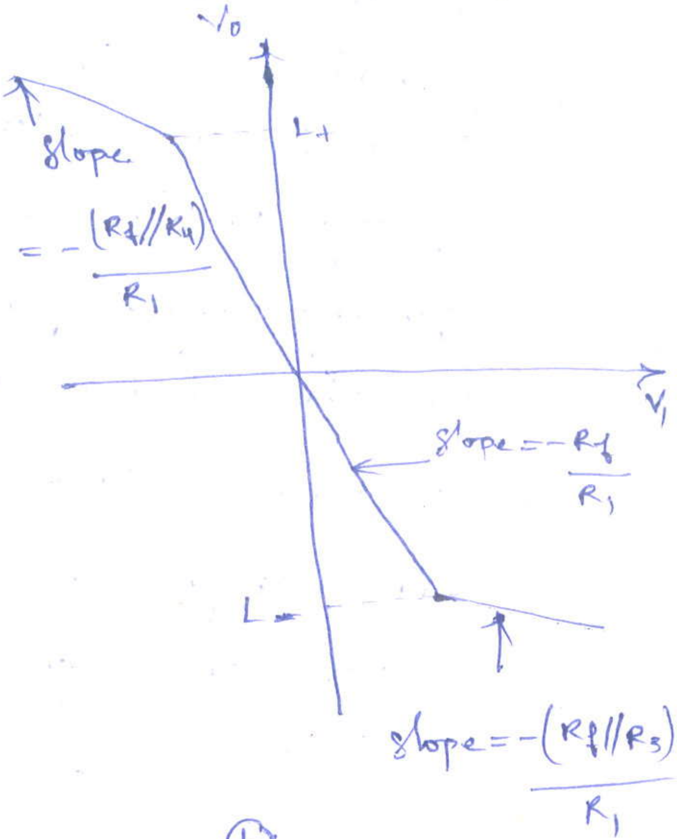
The oscillation condition, the Barkhausen criterion, guarantees sustained oscillations in a mathematical sense. It is well known, however, that the parameters of any physical system cannot be maintained constant for any length of time. In other words, suppose we work hard to make $A\beta = 1$ at $\omega = \omega_0$, and then the temperature changes and $A\beta$ becomes slightly less than unity. Obviously, oscillations will cease in this case. Conversely, if $A\beta$ exceeds unity, oscillations will grow in amplitude. We therefore need a mechanism for forcing $A\beta$ to remain equal to unity at the desired value of output amplitude. This task is accomplished by providing a nonlinear circuit for gain control.

A limiter circuit is employed for the amplitude control of op-amp oscillators. The limiter circuit is shown in fig ⊛a and its transfer is depicted in fig ⊛b. To see how the transfer characteristic is obtained, consider first the case of a small (close to zero) input signal V_1 and a small output voltage V_0, so that V_A is positive and V_B is negative. It can be easily seen that both diodes D_1 and D_2 will be off.



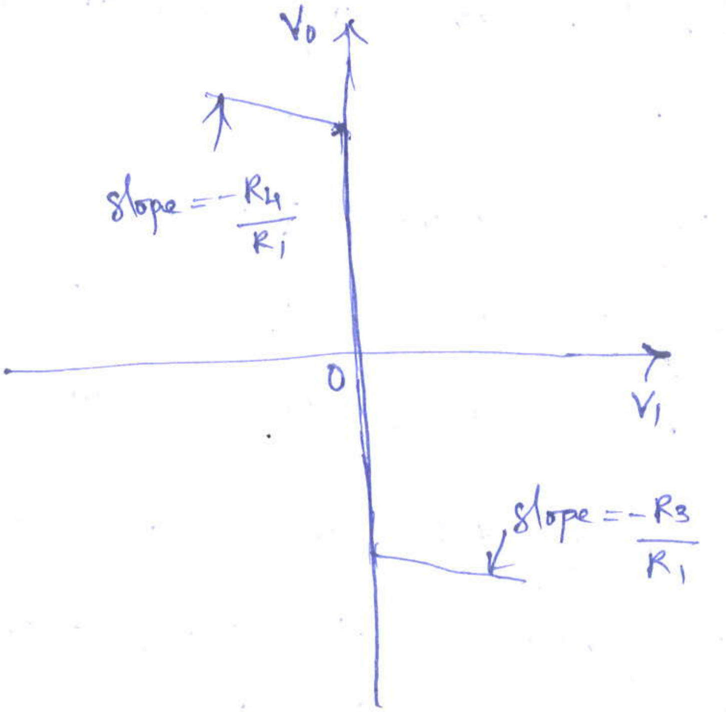
(a)

Limiter ckt.



(b)

Transfer chars of the limiter ckt.



(c)

When Rf is removed the limiter turns into a comparator

Thus all of the input current V_1/R_1 flows through the feedback resistance R_f and the output voltage is given by

$$V_0 = -\left(R_f/R_1\right) \cdot V_1 \quad \text{——— (1)}$$

This is the linear portion of the limiter transfer characteristic in fig (b). We can use superposition to find the voltages at node A and B in terms of $\pm V$ and V_0 as,

$$V_A = V \cdot \frac{R_3}{R_2+R_3} + V_0 \cdot \frac{R_2}{R_2+R_3} \quad \text{———} \quad (2)$$

$$V_B = -V \cdot \frac{R_4}{R_4+R_5} + V_0 \cdot \frac{R_5}{R_4+R_5} \quad \text{———} \quad (3)$$

As v_I goes positive, V_0 goes negative (equ (1)) and from eq (3) that V_B will become more negative, thus keeping D_2 off. Equ (2) shows, however, that V_A becomes less positive. Then, if we continue to increase v_I, a negative value of V_0 will be reached at which V_A becomes $-0.7V$ or so and diode D_1 conducts. If we use the constant voltage-drop model for D_1 and denote the voltage drop V_D, the value of V_0 at which D_1 conducts can be found from equ (2). This is the negative limiting level, which we denote L_-.

$$L_- = -V \cdot \frac{R_3}{R_2} - V_D\left(1+\frac{R_3}{R_2}\right) \quad \text{———} \quad (4)$$

The corresponding value of v_I can be found by dividing L_- by the limiter gain $-R_f/R_1$. If v_I is increased beyond this value, more current is injected into D_1, and V_A remains at approximately $-V_D$. Thus the current through R_2 remains constant, and the additional diode current flows through R_3. Thus R_3

appears in effect in parallel with R_f, and the incremental gain is $-(R_f \| R_3)/R_1$. To make the slope of the transfer characteristic small in the limiting region, a low value should be selected for R_3.

///ly L_+

$$L_+ = V \frac{R_4}{R_5} + V_D\left(1 + \frac{R_4}{R_5}\right) \quad \text{--- (5)}$$

and the slope of the transfer characteristic in the positive limiting region is $-(R_f \| R_4)/R_1$. We thus see that the the ckt in fig (a) functions as a soft limiter, with the limiting levels L_+ & L_- and the limiting gain independently adjustable by the selection of appropriate resistor values.

Finally, we note that increasing R_f results in a higher gain in the linear region while keeping L_+ and L_- unchanged. In the limit, removing R_f altogether results in the transfer characteristic of fig (c), which is that of a comparator. That is, the circuit compares V_I with the comparator reference value of 0V; $V_I > 0$ results in $V_O \simeq L_-$ and $V_I < 0$ yields $V_O \simeq L_+$.

General form of LC oscillator:

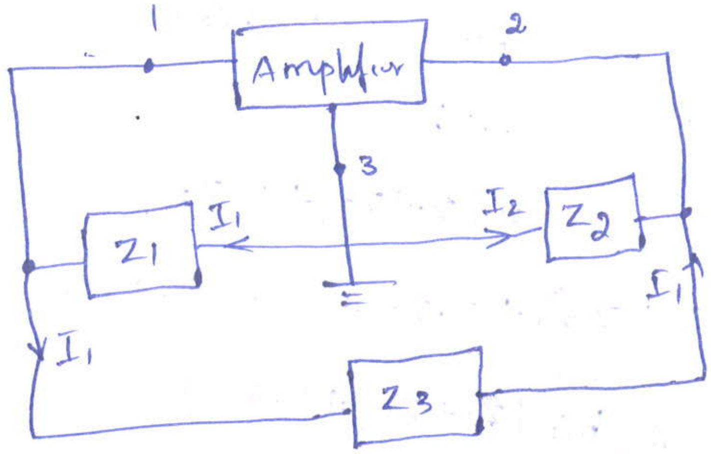
Oscillator is a combination of the amplifier and feedback components. The amplifier provides 180° phase shift and the feedback network produces additional phase shift of 180°, thus the total phase shift is 360°, It is the required condition for the oscillation.

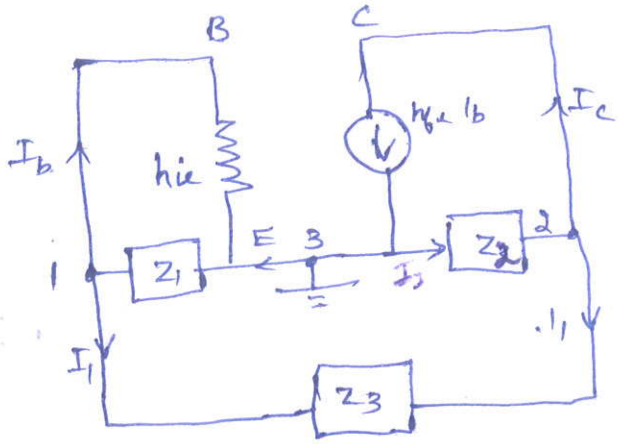
1, BJT LC oscillator provides finite input impedance.

2, FET or OP-AMP provides infinite impedance.

General form LC oscillator is shown below. Z_1, Z_2 & Z_3 are reactive elements constituting the feedback tank circuit which determines the frequency of oscillation. Here Z_1 and Z_2 serve as an a.c. voltage divider for the output voltage and feedback signal. Therefore, the voltage across Z_1 is the feedback signal. The frequency of oscillation of the LC oscillator is

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





The inductive or capacitive reactances are represented by Z_1, Z_2, Z_3.

The output terminals are 2 and 3 and input terminals are 1 and 3.

Load impedance

Z_1 & the input resistance hie of the transistor are in parallel, their equivalent impedance Z' is given by

$$\frac{1}{Z'} = \frac{1}{Z_1} + \frac{1}{h_{ie}}$$

$$Z' = \frac{Z_1 h_{ie}}{Z_1 + h_{ie}}$$

Now the Load impedance Z_L b/w the o/p terminal 2 and 3 is the equivalent impedance of Z_2 in parallel with the series combination of Z' and Z_3.

$$\frac{1}{Z_L} = \frac{1}{Z_2} + \frac{1}{Z' + Z_3}$$

$$= 1/Z_2 + \frac{1}{\frac{Z_1 h_{ie}}{Z_1 + h_{ie}} + Z_3}$$

$$= \frac{1}{Z_2} + \frac{Z_1 + h_{ie}}{Z_1 h_{ie} + Z_3(Z_1 + h_{ie})}$$

$$= \frac{1}{Z_2} + \frac{Z_1 + h_{ie}}{h_{ie}(Z_1 + Z_3) + Z_1 Z_3}$$

$$= \frac{h_{ie}(Z_1 + Z_3) + Z_1 Z_3 + Z_2(Z_1 + h_{ie})}{Z_2 [h_{ie}(Z_1 + Z_3) + Z_1 Z_3)}$$

$$\frac{1}{Z_L} = \frac{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}{Z_2\left[h_{ie}(Z_1+Z_3)+Z_1Z_3\right]}$$

$$Z_L = \frac{\cancel{h_{ie}}\cdot Z_2\left[h_{ie}(Z_1+Z_3)+Z_1Z_3\right]}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}.$$

Voltage gain without feedback.

$$A_V = \frac{V_o}{V_i} = \frac{I_C R_L}{I_b h_{ie}} = \frac{-h_{fe}\cancel{I_b}R_L}{\cancel{I_b}h_{ie}}$$

$$\boxed{A_V = \frac{-h_{fe}Z_L}{h_{ie}}}$$

Feedback fraction β

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

The o/p voltage between the terminals 3 & 2 in terms of the current I_1 is given by

$$V_o = -I_1(Z'+Z_3)$$

$$= -I_1\left(\frac{Z_1h_{ie}}{Z_1+h_{ie}}+Z_3\right)$$

$$= -I_1\left(\frac{Z_1h_{ie}+Z_3h_{ie}+Z_3Z_1}{Z_1+h_{ie}}\right)$$

$$V_o = -I_1\left(\frac{h_{ie}(Z_1+Z_3)+Z_1Z_3}{Z_1+h_{ie}}\right)$$

The feedback voltage input terminals 3 & 1 is given by

$$V_f = -I_1 Z' = -I_1\left(\frac{Z_1 h_{ie}}{Z_1 + h_{ie}}\right)$$

$$\beta = \frac{V_f}{V_o} = \frac{-\cancel{I_1}\,\dfrac{Z_1 h_{ie}}{\cancel{Z_1 + h_{ie}}}}{-\cancel{I_1}\left(\dfrac{h_{ie}(Z_1+Z_3)+Z_1Z_3}{\cancel{Z_1+h_{ie}}}\right)}$$

$$\boxed{\beta = \frac{Z_1 h_{ie}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}}$$

Equation for the oscillator. For oscillation, we must have

$$A_v \cdot \beta = 1$$

$$\left(\frac{-h_{fe} Z_L}{h_{ie}}\right)\left(\frac{Z_1 h_{ie}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}\right) = 1$$

$$\frac{\dfrac{-h_{fe} Z_2\left[h_{ie}(Z_1+Z_3)+Z_1Z_3\right]}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}}{\cancel{h_{ie}}}\left[\frac{Z_1 \cancel{h_{ie}}}{h_{ie}(Z_1+Z_3)+Z_1Z_3}\right] = 1$$

$$\left(\frac{-h_{fe} Z_2\left[h_{ie}(Z_1+Z_3)+Z_1Z_3\right]}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2}\right)\left(\frac{Z_1}{h_{ie}(Z_1+Z_3)+Z_1Z_3}\right) = 1$$

$$\left(\frac{h_{fe}Z_2\left[\cancel{Z_1h_{ie}+Z_3h_{ie}}\right]+Z_1Z_3}{h_{ie}Z_1+h_{ie}Z_2+h_{ie}Z_3+Z_1Z_2+Z_1Z_3}\right)\left(\frac{Z_1}{h_{ie}\cancel{(Z_1+Z_3)}+Z_1Z_3}\right)=-1$$

$$\frac{h_{fe}Z_2Z_1}{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3}=-1$$

$$-h_{fe}Z_2Z_1=h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2+Z_1Z_3$$

$$-h_{fe}Z_1Z_2-Z_1Z_2=h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_3$$

$$h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_3=-Z_1Z_2(1+h_{fe})$$

$$\boxed{h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2(1+h_{fe})+Z_1Z_3=0}$$

This is the general equation for the oscillator.

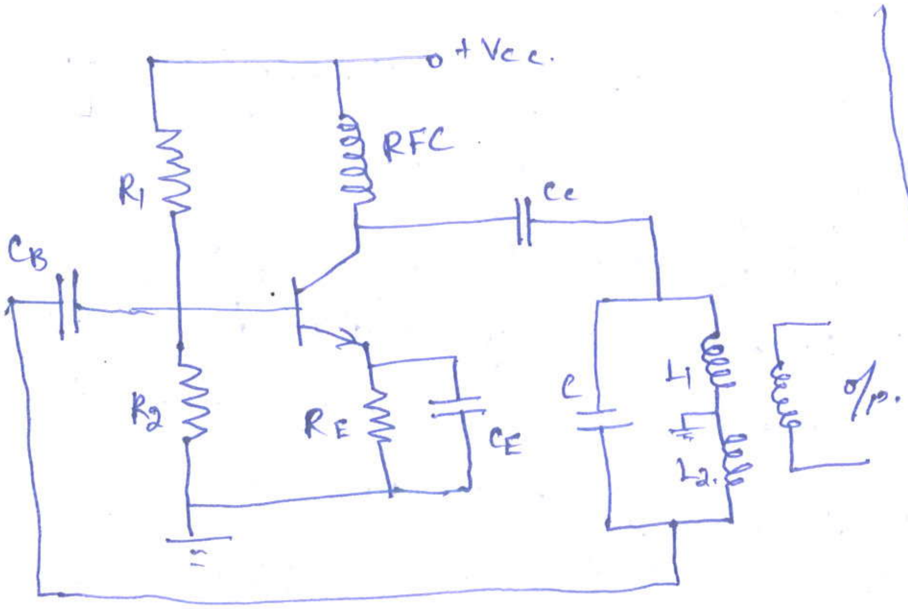
Hartley Oscillator.

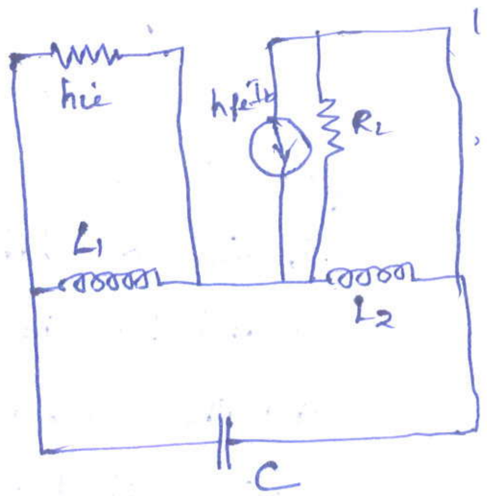
The fig ① shows the circuit of a Hartley oscillator, falls in the category of tuned oscillator or resonant circuit oscillator. It tank circuit consists of two coils L_1 and L_2. The coil L_1 is inductively coupled to coil L_2 and the combination works as an autotransformer. A coil called radio frequency choke (RFC) is connected between the collector and the Vcc supply.

It acts as a load for the collector and also permits an easy flow of d.c. current. But blocks a.c. current. The feed back b/w the output and input cks is accomplished through autotransformer action, which also introduces a phase shift of 180°. The phase reversal b/w the o/p & i/p voltage occurs because they are taken from the opposite ends of the coils (L_1 & L_2) with respect to the tap, which is grounded. It may be noted that the tap on the combination of L_1 and L_2 coils is actually connected to the transistor emitter terminal via ground and through the capacitance (C_E). Since the transistor also introduces a phase shift of 180°, therefore, the total phase shift is 360° and hence the feedback is positive. Ignoring the loading effects of the base, the feedback fraction is given by the relation,

$$\beta = L_2/L_1$$

For oscillations to start, the voltage gain must be greater than $1/\beta$, which is equal to L_1/L_2.





For oscillations to start, the voltage gain must be greater than $1/\beta$ which is equal to L_1/L_2

The capacitor C_C, connected b/w the collector and the tuned circuit, is called coupling capacitor. It permits only the a.c. current to pass to the tank circuit. In other words, the capacitor C_C blocks the d.c. currents. The capacitor C_B, called blocking capacitor, further blocks the d.c. currents reaching at the base. The resistors R_1, R_2 and R_E are used to provide d.c. bias to the transistor.

The oscillations are produced because of positive feedback from the tank circuit. The frequency of oscillation is given by relation

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

Where $L = L_1 + L_2 + 2M$

$= L_1 + L_2$ (if the mutual inductance M is neglected).

Operation :

* When V_{CC} is applied, the collector current begins to flow and the drop in collector voltage is coupled through capacitor C and L_2. Thus the capacitor charges to its maximum voltage. This voltage acts as initial excitation for the tank circuit causing a current to flow in the LC circuit. This current induces damped

oscillations across L_1, which drives the base of the transistor.

* This damped signal is amplified and appears at collector which is coupled or feedback to the tank circuit C & L_1. The feedback voltage across L_2 is in phase with the input voltage across L_1 results in sustained oscillation is obtained

* The feedback voltage is in phase with input voltage, since 180° phase shift produced by transistor and another 180° being provided by the fact that the two end coil L_1 & L_2 are connected in opposite polarity.

* The capacitor C_0 blocks d.c. components of the collector circuit but couples ac signal. The dc component supply through RFC which also prevents the a.c. component to flow through this path because RFC acts as short circuit for d.c. and open circuit for a.c.

* As a result of this, d.c. is out of tank circuit, thus energy loss due to tank circuit is reduced hence the oscillator is more stable.

Analysis

The general equation for LC oscillator is given by

$$h_{ie}(Z_1+Z_2+Z_3)+Z_1Z_2(1+h_{fe})+Z_1Z_3=0.$$

$$z_1 = j\omega L_1 + j\omega M$$

$$z_2 = j\omega L_2 + j\omega M$$

$$z_3 = \frac{1}{j\omega C} = \frac{-j}{\omega C}$$

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

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

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

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

$$h_{ie}\left(2j\omega M + j\omega(L_1+L_2) - j/\omega C\right) + j(\omega L_1 + M)\,j\,(\omega L_2 + M)$$
$$\left(1+h_{fe}\right) + \left(j\omega L_1 + j\omega M\right)\cdot -j/\omega C = 0.$$

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

$$h_{ie}\left[2jwM + jw(L_1+L_2) - j/wc\right] + j^2w^2(L_1+M)(L_2+M)(1+h_{fe})$$
$$+ j^2w^2\left[(L_1+M)\cdot \frac{1}{w^2c}\right]$$

$$h_{ie}jw\left(2M+(L_1+L_2) - \frac{1}{w^2c}\right] - w^2(L_1+M)\left[(L_2+M)(1+h_{fe}\right]$$
$$-\frac{1}{w^2c}\Big] = 0.$$

To find the frequency of oscillation

$f_0 = \frac{w_0}{2\pi}$ can be dermune by equating

Imaginary part $=0$

$$L_1+L_2+2M - \frac{1}{w^2c} = 0$$

$$w^2cL_1 + w^2cL_2 + 2Mw^2c - 1 = 0.$$

$$w^2c\left[L_1+L_2+2M\right] - 1 = 0$$

$$w^2c\left[L+L_2+2M\right] = 1$$

$w = 2\pi f$

$$w^2c\left[L_1+L_2+2M\right] = 1$$

$$(2\pi f)^2c(L_1+L_2+2M) = 1$$

$$4\pi^2f^2c\left[L_1+L_2+2M\right) = 1$$

$$f^2 = \frac{1}{4\pi^2c\left[L_1+L_2+2M\right]}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{L_1+L_2+2M\cdot C}}}$$

To find condition for oscillation the real part is equ to zero.

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

$$(1+h_{fe})(L_2+M)=\frac{1}{\omega^2 C}$$

$$1+h_{fe}=\frac{1}{\omega^2 C(L_2+M)}$$

$$=\frac{1}{\frac{1}{(L_1+L_2+2M)}(L_2+M)}=\frac{L_1+L_2+2M}{L_2+M}$$

$$1+h_{fe}=\frac{L_1+L_2+M+M}{L_2+M}=\frac{L_2+M}{L_2+M}+\frac{L_1+M}{L_2+M}$$

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

pbm 1 In the Hartley oscillator $L_2 = 0.4\,mH$ and $C = 0.004\,\mu F$. If the frequency of the oscillator is 120 KHz, find the value of L_1. Neglect the mutual inductance.

Sol: The frequency of Hartley oscillator is given by

$$f_0=\frac{1}{2\pi\sqrt{(L_1+L_2)C}}$$

$$L_1=\frac{1}{4\pi^2 f_0^2 C}-L_2.$$

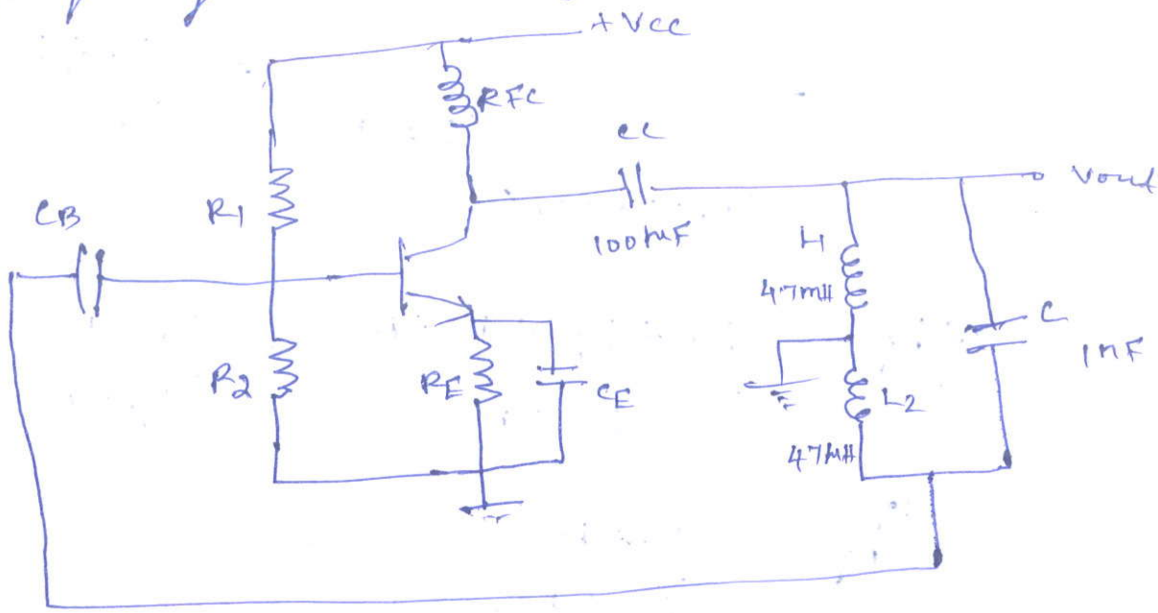
$$= \frac{1}{4\pi^2 (120 \times 10^3)^2 \times 0.004 \times 10^{-6}} - 4.4 \times 10^{-3}$$

$$= 0.44 \times 10^{-3} - 0.4 \times 10^{-3}$$

$$= 0.04\ mH.$$

pbm

For the Hartley oscillator shown below, find the frequency oscillation of the circuit.



Sol:

Given $C = 1nF = 1 \times 10^{-9} F$

$L_1 = 4.7 mH = 4.7 \times 10^{-3} H$

$L_2 = 47 \mu H = 47 \times 10^{-6}$

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{\sqrt{2\pi (L_1 + L_2) \cdot C}}$$

$$= \frac{1}{2\pi \sqrt{[(4.7 \times 10^{-3}) + (47 \times 10^{-6})] \times (1 \times 10^{-9})}\, C}$$

$$= \frac{1}{2\pi \times 2.179 \times 10^{-6}} = 0.0731 \times 10^{6}\ Hz$$

$$= 73.1\ KHz$$

pbm

A Hartley oscillator is designed with $L_1 = 2mH$, $L_2 = 20\mu H$ and a variable capacitance. Determine the range of capacitance values if the frequency of oscillation is varied between 950 and 2050 KHz.

Sol:

$L_1 = 2mH$

$L_2 = 20mH$

$f_0 = 950KHz$ to $2050 KHz$

Let

C_1 = value of capacitance required for a frequency of 950 KHz of

C_2 = value of capacitance required for a frequency of 2050 KHz.

Wkt the resonant frequency

$$950 \times 10^3 = \frac{1}{2\pi\sqrt{(L_1 + L_2) \cdot C_1}}$$

$$= \frac{1}{2\pi\sqrt{[(2\times10^{-3}) + (20\times10^{-6})] \cdot C_1}}$$

$$= \frac{1}{2\pi\sqrt{2.02\times10^{-3} \cdot C_1}}$$

$$C_1 = \frac{1}{4\pi^2 (2.02\times10^{-3}) \times (950\times10^3)^2}$$

$$= 13.9 \times 10^{-12} F = 13.9 PF.$$

$|||^{ly}$

$$C_2 = 2.98 PF$$

Hence the range of capacitance values required is 2.98 PF to 13.9 PF.

pbm In a Hartley oscillator, the value of the capacitor in the tuned circuit is 500 pF and the two sections of coil have inductances 38 μH and 12 μH. Find the freq. of osc. & the feedback factor β.

Sol
$$f_o = \frac{1}{2\pi\sqrt{LC}}$$

$$L = L_1 + L_2 = 38\times10^{-6} + 12\times10^{-6} = 50\times10^{-6}$$
$$= 50\,\mu H$$

$$C = 500\,pF$$

$$\therefore f_o = \frac{1}{2\pi\sqrt{50\times10^{-6}\times500\times10^{-12}}} = 1MHz$$

$$\text{Feedback factor } \beta = \frac{L_1}{L_2} = \frac{38\times10^{-6}}{12\times10^{-6}} = 3.166$$

Colpitts oscillator.

It is similar to the Hartley oscillator, except for one difference. The colpitts oscillator uses tapped capacitance instead of tapped inductance used in Hartley oscillator. The fig. ① shows the colpitts oscillator circuit, whereas fig ② shows its a.c. equivalent circuit. The tank circuit is made up of two capacitors C_1 and C_2 connected in series with each other across a fixed inductance (L). The resistance (R_1, R_2 and R_E) Capacitance (C_C & C_B) and RFC have the function as mentioned for Hartley oscillator. The feedback b/w the output and input circuit is accomplished by the voltage developed across the capacitor C_2. Ignoring the loading effect of the base, the feedback fraction

$$\beta = C_1/C_2$$

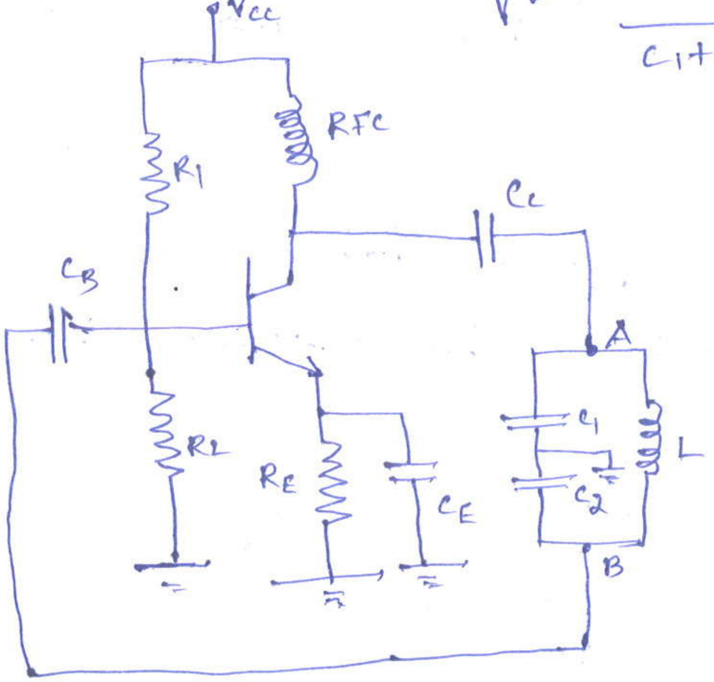
For oscillation to start, the voltage gain (A_v) must be greater than $1/\beta$ (or C_2/C_1). Or in other words $A_v > C_2/C_1$.

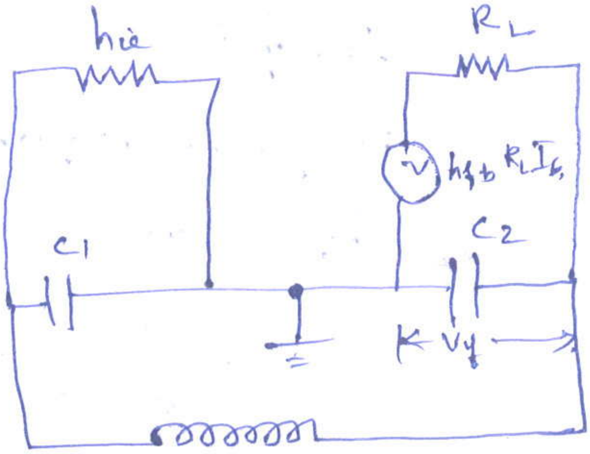
- The frequency of oscillations (neglecting mutual inductance) is give by the relation

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

Where C is the effective value of capacitance and equal to $\frac{C_1 \cdot C_2}{C_1 + C_2}$

$$f_0 = \frac{1}{2\pi\sqrt{L \times \frac{C_1 \cdot C_2}{C_1 + C_2}}}.$$





It may be noted that in a colpitts oscillator, the capacitors C_1 & C_2 act as a simple alternating voltage divider. Therefore points A and B are 180° out of phase with each other. Another phase shift of 180° is provided by the transistor itself. Thus there is a total phase shift of 360° b/w the emitter-base and collector-base circuits.

When the circuit is energised by switching on the supply, the capacitors C_1 & C_2 are charged. These capacitors discharge through the coil (L), which sets up the oscillations of frequency $f_0 = 1/2\pi\sqrt{LC}$

where $C = C_1 \cdot C_2 / (C_1 + C_2)$.

The oscillations across the capacitor C_2 are feedback to the base-emitter junction and appears in an amplified form at the collector. Because of the positive feedback the oscillations of constant amplitude are produced. The colpitts oscillators are widely used as a signal generator for frequencies between 1MHz and 500 MHz.

Analysis

The general equation for oscillator is

$$h_{fe} Z_1 Z_2 + Z_3 Z_1 + Z_1 Z_2 + h_{ie} (Z_1 + Z_2 + Z_3) = 0.$$

$$Z_1 = \frac{1}{jwC_1} = -j/wC_1 \qquad Z_3 = jwL$$

$$Z_2 = 1/jwC_2 = -j/wC_2$$

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

$$+\left(\frac{-j}{\omega C_1}\right)(j\omega L)=0$$

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

$$-\left(\frac{j}{\cancel{\omega} C_1}\right)\left(\frac{j}{\omega C_2}\right)(j\cancel{\omega}L)=0$$

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

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

$$jh_{ie}\left[\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right]+\left[\frac{1+h_{fe}}{\omega^2 C_1 C_2}-\frac{L}{C_1}\right]=0.$$

To find freq of oscillation equate the imaginary part to zero.

$$jh_{ie}\left[\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L\right]=0$$

$$\frac{1}{\omega C_1}+\frac{1}{\omega C_2}-\omega L=0$$

$$\omega L=\frac{1}{\omega C_1}+\frac{1}{\omega C_2}.$$

$$\omega^2 L=\frac{1}{C_1}+\frac{1}{C_2}$$

$$\omega^2=\frac{1}{C_{eq}\cdot L} \qquad \frac{1}{C_{eq}}=\frac{1}{C_1}+\frac{1}{C_2}.$$

$$f=\frac{1}{2\pi\sqrt{C_{eq}\cdot L}}=\frac{1}{2\pi\sqrt{L\cdot\frac{C_1C_2}{C_1+C_2}}}$$

To determine the condition for oscillation, real part is equated to zero.

This is the condition for the sustained oscillation.

pbm. In a transistor colpitts oscillator, $C_1 = 0.001\ \mu F$; $C_2 = 0.01\ \mu F$ and $L = 5\ \mu H$. Find the required gain for oscillation and the frequency of oscillation.

$C_1 = 0.0001\ \mu F$

$C_2 = 0.01\ \mu F$

$L = 5\ \mu H$

Voltage gain required to produce oscillations

$$A_v > \frac{C_2}{C_1} = \frac{0.01 \times 10^{-6}}{0.001 \times 10^{-6}} = 10$$

Capacitance

$$C = \frac{C_1 \cdot C_2}{C_1 + C_2} = \frac{(0.001 \times 10^{-6}) \times (0.01 \times 10^{-6})}{(0.001 \times 10^{-6}) + (0.01 \times 10^{-6})}$$

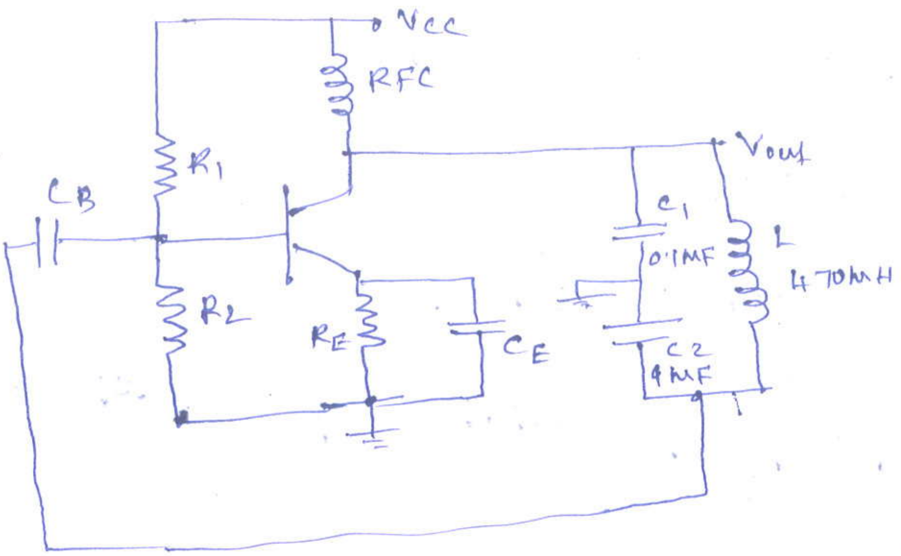
$$= 9.091 \times 10^{-10}\ F.$$

Frequency of oscillations

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi \times \sqrt{(5\times10^{-6})\times(9.091\times10^{-10})}}\ Hz$$

$$= 2.37\times10^{6}\ Hz = 2.37\ MHz$$

pbm For the transistor colpitts osc. shown below find the frequency of oscillations.



$$f_0 = \frac{1}{2\pi\sqrt{L\cdot C}} = \frac{1}{2\pi\sqrt{L\times\frac{C_1\cdot C_2}{C_1+C_2}}}$$

$$= 24.4\ KHz$$

pbm. A colpitts oscillator is designed with $C_1 = 100\,pF$ and $C_2 = 7500\,pF$. The inductance is variable. Determine the range of inductance values, if the frequency of oscillation is to vary between 950 KHz and 2050 KHz

Sol L_1 = value of inductance required for a freq. of 950×10^3 Hz

L_2 = " 2050×10^3 Hz

Capacitance $C = \frac{C_1\cdot C_2}{C_1+C_2} = \frac{(100\times10^{-12})\times(7500\times10^{-12})}{(100\times10^{-12})+(7500\times10^{-12})}\ F = 98.68\times10^{-12}\ F.$

and the resonant frequency.

$$950 \times 10^{3} = \frac{1}{2\pi\sqrt{L_1 \cdot C}} = \frac{1}{2\pi \times \sqrt{L_1 \times (98.68 \times 10^{-12})}}$$

$$(950 \times 10^{3})^2 = \frac{1}{4\pi^2 \cdot L_1 \times (98.68 \times 10^{-12})}$$

$$L_1 = \frac{1}{4\pi^2 \times (98.68 \times 10^{-12}) \times (950 \times 10^{3})^2}$$

$$= 284 \times 10^{-6}\,H = 284\,\mu H.$$

///ly and for 2050×10^{3} Hz

$$L_2 = 61 \times 10^{-6}\,H = 61\,\mu H.$$

pbm The frequency of oscillation of a colpitts osc is given by $f_0 = \frac{1}{2\pi\sqrt{L \cdot \left(\frac{C_1 C_2}{C_1 + C_2}\right)}}$

Where L, C_1 & C_2 are the frequency-determining components. Such a circuit operates at 450 KHz with $C_1 = C_2$. What will be the oscillation frequency if the value of C_2 is doubled.

Sol $f_0 = 450$ KHz & $C_1 = C_2 = C$ (say)

f_0' = The freq. when the value of C_2 is double

WKT

$$f_0 = \frac{1}{2\pi\sqrt{\left[L\left(\frac{C_1 C_2}{C_1 + C_2}\right)\right]}} = \frac{1}{2\pi\sqrt{L \cdot \frac{C \cdot C}{C + C}}}$$

$$= \frac{1}{2\pi\sqrt{\frac{L \cdot C}{2}}} = \frac{1}{2\pi} \cdot \sqrt{\frac{2}{L \cdot C}} \quad \text{--- (1)}$$

When the value of C_2 is doubled, the frequency of oscillation

$$f_0' = \frac{1}{2\pi\sqrt{L\left(\frac{C_1+2C_2}{C_1+2C_2}\right)}} = \frac{1}{2\pi\sqrt{L \cdot \frac{2C \cdot C}{3C}}}$$

$$= \frac{1}{2\pi \cdot \sqrt{\frac{2LC}{3}}} = \frac{1}{2\pi} \cdot \sqrt{\frac{3}{2LC}} \quad \text{———} \; (2)$$

Dividing (II) by (I)

$$\frac{f_0'}{f_0} = \frac{\frac{1}{2\pi} \cdot \sqrt{\frac{3}{2LC}}}{\frac{1}{2\pi} \cdot \sqrt{\frac{2}{LC}}} = \sqrt{\frac{3}{2LC} \times \frac{LC}{2}} = \sqrt{3/4}$$

$$= 0.86.$$

$$f_0' = 0.86 \times f_0 = 0.866 \times 450 \times 10^3$$

$$= 389 \times 10^3 \; Hz$$

$$= 389 \; KHz$$

Pbm In a colpitts oscillator, the values of the inductors and capacitors in the tank circuit are $L = 40mH$, $C_1 = 100pF$ and $C_2 = 500pF$.

(I) Find the frequency of oscillation.

(II) if the output voltage is 10V, find the feedback vol.

(III) Find the minimum gain if freq. is changed by changing L alone

(IV) Find the value of C_1 for a gain of 10.

(V) Also find the new frequency.

(i) $$f_0 = \frac{1}{2\pi\sqrt{LC_{eq}}} = \frac{1}{2\pi\sqrt{L \cdot \frac{C_1 + C_2}{C_1 \cdot C_2}}}$$

$$f_0 = 87.2\,KHz$$

(ii) $$V_f = \frac{V_0 C_1}{C_2} = \frac{10 \times 100 \times 10^{-12}}{500 \times 10^{-12}} = 2V$$

(iii) Since the gain depends upon C_1 & C_2 only and is independent of L

$$Gain = \frac{C_2}{C_1} = \frac{500 \times 10^{-12}}{100 \times 10^{-12}} = 5$$

(iv) $$\frac{C_2}{C_1} = 10$$

$$C_1 = \frac{C_2}{10} = \frac{500 \times 10^{-12}}{10} = 50\,pF$$

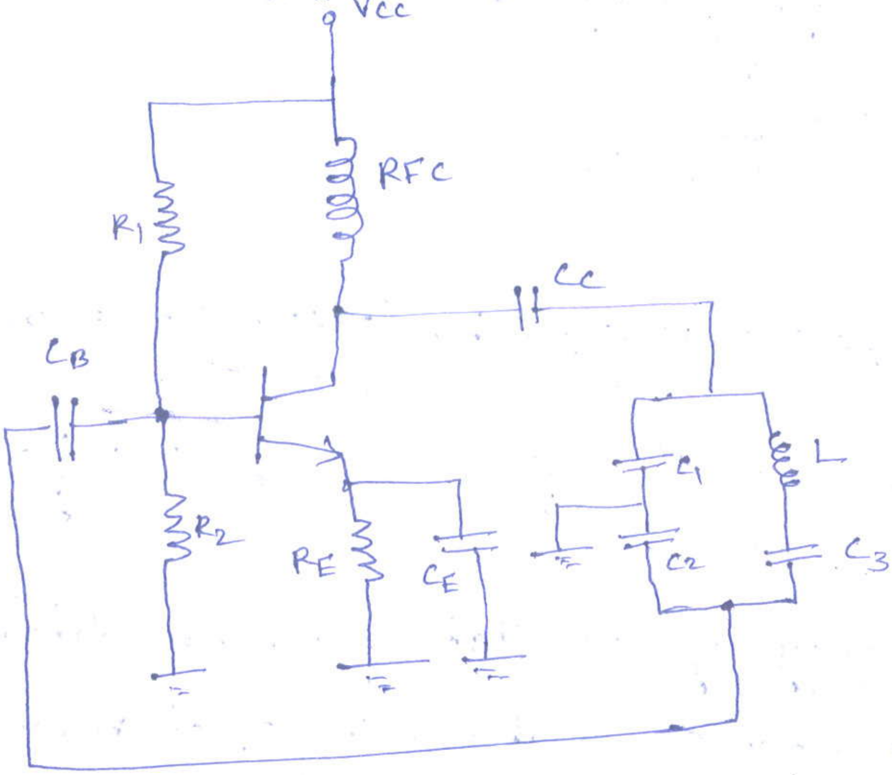
(v) The frequency of oscillation

$$f_0 = \frac{1}{2\pi\sqrt{\frac{40 \times 10^{-3} \times 50 \times 10^{-12} \times 500 \times 10^{-12}}{50 \times 10^{-12} + 500 \times 10^{-12}}}}$$

$$= 118.1\,KHz$$

Clapp Oscillator:

It is an improved version of the colpitts oscillator. Figure below shows the circuit of clapp oscillator.



The ckt differs from the colpitts oscillator only in one respect, that it contains one additional capacitor (C_3) connected in series with the inductor. The addition of capacitor (C_3) improves the frequency stability and eliminates the effect of transistor parameters and stray capacitances.

The operation of clapp oscillator ckt is in the same way as that of colpitts oscillator. The frequency of oscillator is given by the relation

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

Where $$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}}$$

usually, the value of C_3 is much smaller than C_1 & C_2. As a result of this, C is approximately equal to C_3. Therefore the frequency of oscillation

$$f_0 = \frac{1}{2\pi\sqrt{L \cdot C_3}}$$

Analysis

$$h_{fe} Z_1 Z_2 + Z_3 Z_1 + Z_1 Z_2 + h_{ie}(Z_1 + Z_2 + Z_3) = 0.$$

$$Z_1 = -j/\omega C_1 \qquad Z_2 = -j/\omega C_2 \qquad Z_3 = j\omega L + \frac{1}{j\omega C_3}$$

The frequency of oscillation can be determined from the resonance condition. WKT, at resonance the imaginary part of general LC oscillator is equated to zero.

hence $$-\frac{j}{\omega C_1} - \frac{j}{j\omega C_2} - \frac{j}{\omega C_3} + j\omega L = 0.$$

$$j\omega L = j\left(\frac{1}{\omega C_1} + \frac{1}{\omega C_2} + \frac{1}{\omega C_3}\right)$$

$$\omega^2 L = \left(1/C_1 + 1/C_2 + 1/C_3\right)$$

$$\omega^2 = \frac{1}{LC_T} \qquad \text{where } 1/C_T = \frac{1}{C_1} + 1/C_2 + 1/C_3$$

$$\boxed{f = \frac{1}{2\pi\sqrt{L \cdot C_T}}}$$

The capacitance C_3 is used for tuning purpose only hence to find the condition for the oscillation real part of general equation is equated to zero.

[C_3 neglected]

$$h_{fe} Z_1 Z_2 + Z_1 Z_3 + Z_1 Z_2 = 0.$$

$$Z_1 Z_2 (1 + h_{fe}) + Z_1 Z_3 = 0.$$

$$\cancel{h_{fe}(1 + Z_1 Z_2) + Z_1 Z_3 = 0.}$$

$$(h_{fe}+1)\left(\frac{-1}{\omega^2 C_1 C_2}\right) + \frac{-j}{\omega C_1} \cdot (j\omega L) = 0.$$

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

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

$$\omega^2 L C_2 = h_{fe} + 1$$

WK $\omega^2 L = 1/C_1 + 1/C_2$ for colpitts osc.

$$C_2\left(1/C_1 + 1/C_2\right) = h_{fe} + 1$$

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

$$\boxed{h_{fe} \geq \frac{C_2}{C_1}}$$

pbm 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 $L = 470 \mu H$.

$$C = \frac{1}{1/C_1 + 1/C_2 + 1/C_3} = \frac{1}{\frac{1}{(0.1\times10^{-6})} + \frac{1}{(1\times10^{-6})} + \frac{1}{(100\times10^{-12})}}$$

$$= \frac{1}{(10\times10^{6}) + (1\times10^{6}) + (10{,}000\times10^{6})}$$

$$= \frac{1}{10011\times10^{6}} = 99.9\times10^{-12} F = 99.9 PF.$$

Frequency of oscillation

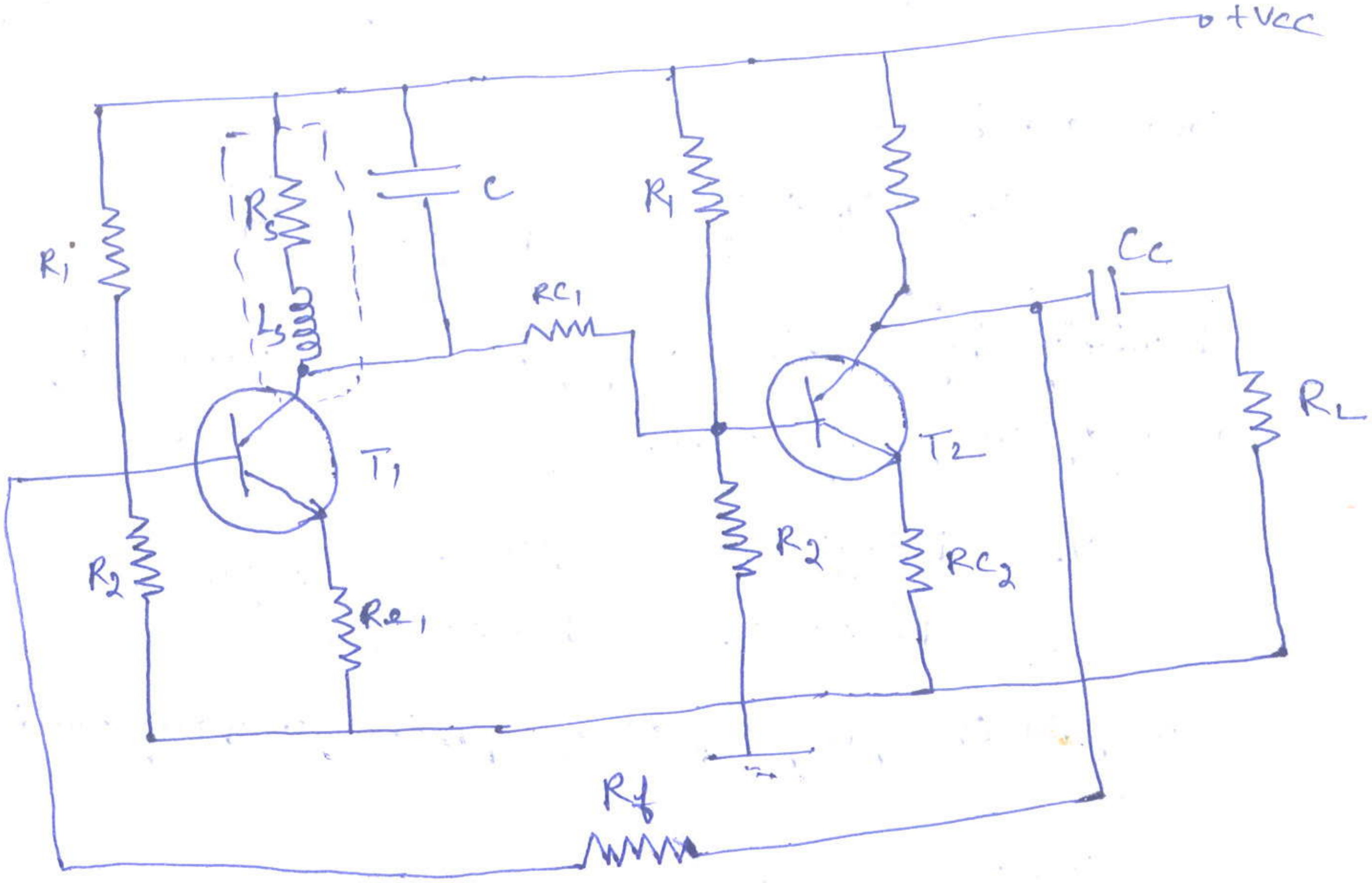
$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi \times \sqrt{(470\times10^{-6})\times(99.9\times10^{-12})}}$$

$$= 734.5 KHz.$$

Franklin Oscillator:

The figure ① shows a simple franklin oscillator. In this oscillator two CE amplifier stages are used to obtain 360° phase shift b/w i/p and output, as a result the loop gain is just greater than unity and barkhausen criteria is satisfied.

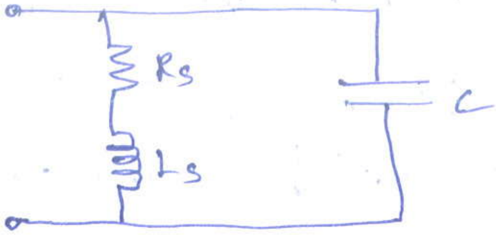
In this oscillator a parallel combination of 'Ls' and 'C' acts as tank circuit, when Vcc is applied the parallel tuned circuit will produce a damped oscillation at a resonant frequency. It is fed to the input of amplifier T_2, it amplifies and produce the phase shift of 180°. The out put of second amplifier is feedback to the input of amplifier T_1 through



Resistance Rf. Thus, the signal is further amplified and produce the another phase difference of 180°. Hence total phase shift of 360° or 0° is achieved as a result of this sustained oscillation is occured.

Frequency of oscillation

The frequency of oscillation is determined by the tank circuit at resonant condition.



WKT

$$Y_L = \frac{1}{R_s + j\omega L_s} \quad \text{d} \quad Y_C = \frac{1}{1/j\omega C}$$

Total admidance

$$Y_T = Y_L + \frac{1}{C} = \frac{1}{R_s + j\omega L_s} + j\omega C$$

Multiply by complex conjugate and separate by real of imaginery part.

$$Y_T = \frac{R_s \cancel{- j\omega L_s}}{R_s^2 + \omega^2 L_s^2} - \frac{j\omega L_s}{R_s^2 + \omega^2 L_s^2} + j\omega C$$

To find the freq. of osc. equ. the imaginery part to zero

$$-\frac{\omega L_s}{R_s^2 + \omega^2 L_s^2} + \omega C = 0$$

$$\omega C = \frac{\omega L_s}{R_s^2 + \omega^2 L_s^2}$$

$$(R_s^2 + \omega^2 L_s^2)\, C = L \quad (\text{or}) \quad R_s^2 + \omega^2 L_s^2 = \frac{L_s}{C}$$

$$\omega^2 L_s^2 = \frac{1}{L_s C} - R_s^2$$

$$\omega^2 = \frac{1}{L_s C} - \frac{R_s^2}{L_s^2}$$

$$\omega = \sqrt{\frac{1}{L_s C} - \frac{R_s^2}{L_s^2}}$$

$$f = \frac{1}{2\pi}\left(\sqrt{\frac{1}{L_s C} - \frac{R_s^2}{L_s^2}}\right)$$

usually $\frac{R_s^2}{L_s^2} << 1$ then $f = \frac{1}{2\pi\sqrt{L_s C}}$.

The impedance at resonance condition is determined from the real part.

$$Z_r = \frac{1}{Y_r} = \frac{R_s^2 + \omega^2 L_s^2}{R_s} = \frac{R_s + \omega^2 L_s^2}{R_s}$$

Substitute the value of ω^2 in the above equ we get

$$Z_r = R_s + \left(\frac{1}{L_s C} - \frac{R_s^2}{L_s^2}\right) L_s^2 = R_s + \frac{\left(\frac{L_s}{C} - R_s^2\right)}{R_s}$$

$$= R_s + \frac{L_s}{CR_s} - R_s = \frac{L_s}{CR_s}$$

thus

$$Z_r = \frac{L_s}{CR_s} = Z_L$$

Hence the voltage gain of 1st stage amplifier is

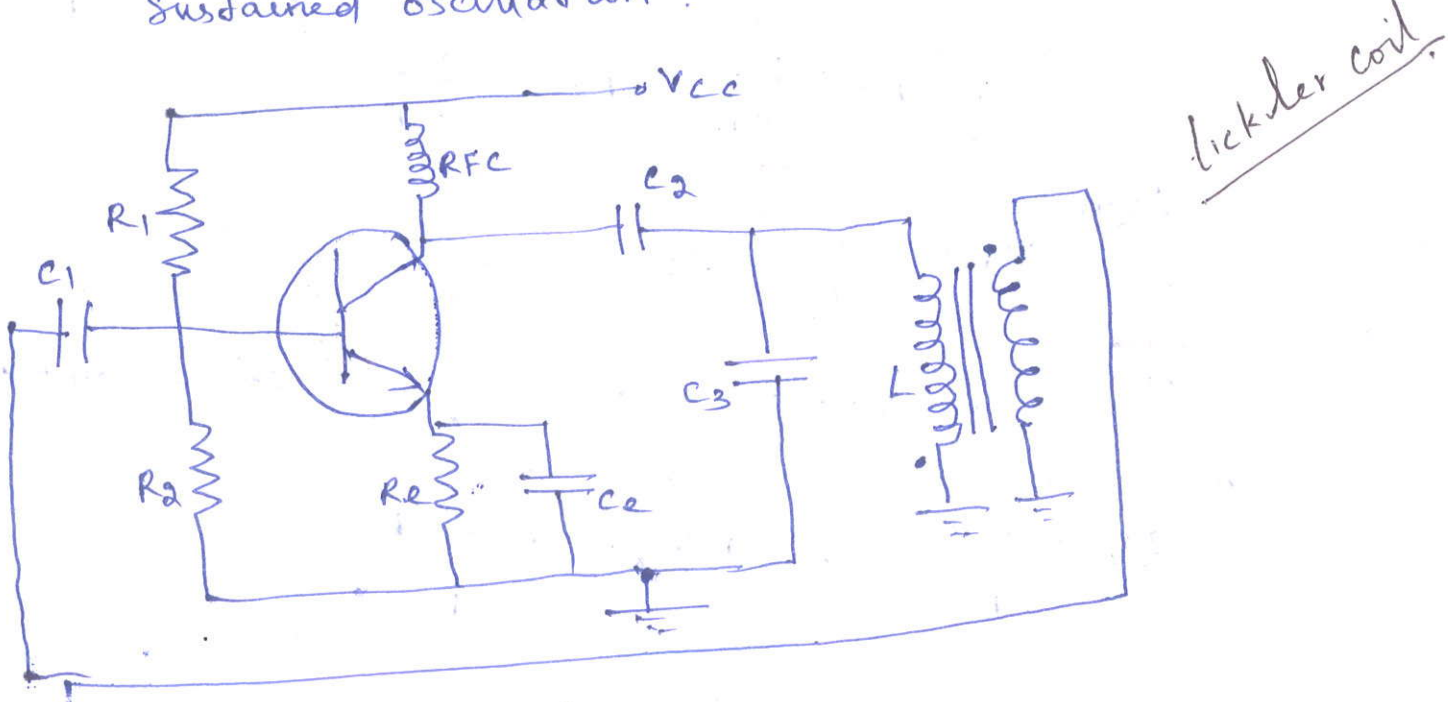
$$AV_1 = \frac{Z_L}{R_{e1}} \quad \& \quad AV_2 = \frac{R_{C2} \| R_L}{R_L}$$

and the overall gain.

$$\boxed{A_V = AV_1 \cdot AV_2}$$

Armstrong Oscillator.

The following fig. shows the armstrong oscillator. In this circuit the collector drives an LC resonant circuit. The primary winding of transformer L and the capacitor C_3 forms a resonant circuit. The feedback signal is taken from a small secondary winding and fed back to the base. There is a phase shift of 180° in the transformer and another 180° phase shift is produced by the amplifier itself, a a result of this total phase shift is 0° or 360° which satisfies the barkhausen criterion, thus the circuit produces sustained oscillation.



The frequency of oscillation is $f = \frac{1}{2\pi\sqrt{LC_3}}$ it is dependent of the primary winding inductance L and capacitor C.

The feedback factor can be defined as

$$\beta = \frac{M}{L}$$

Where M = Mutual inductance

L = Self inductance of the transformer windings

Thus to produce oscillations, the voltage gain must be greater than $1/\beta$

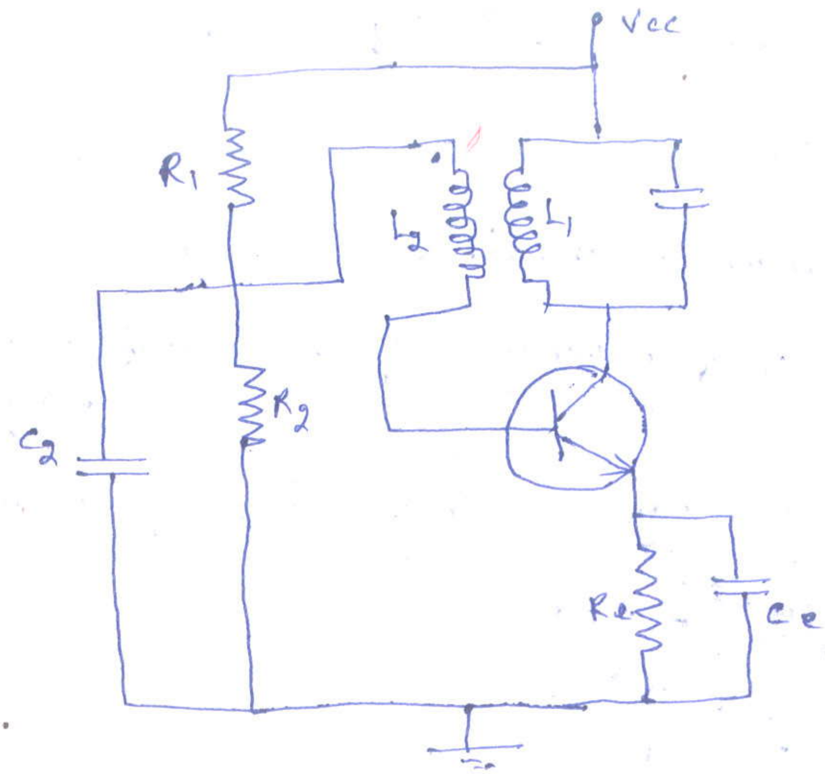
$$\therefore A > 1/\beta > \frac{L}{M}$$

The main drawback of this oscillator is that, it uses transformer as a coupled device, as a result some losses occurs.

Tuned collector oscillator:

In this oscillator, the tank circuit connected in the collector circuit act as load impedances, it determines the frequency of oscillation. The output developed across the tuned circuit is inductively coupled to the base circuit through L_2. The winding direction of the two coils are so adjusted that the positive feedback from the collector circuit to the base circuit.

The tuned circuit in the collector is resonates at w_0. At this frequency the impedance of this circuits is purely reactive.



$L_2 - L_p$

Thus the voltage ~~drop~~ across the inductor is 180° out of phase with the input. The transformer winding direction are so adjusted as to produce a another phase shift of 180°. Then the total loop phase shift is exactly zero.

Operation

When Vcc is applied, the collector current starts rising and the capacitor starts charging towards the maximum voltage. This voltage acts as initial excitation for the tank circuit L_1C_1 causes a current flow in the circuit, when the capacitor discharges. The action induces harmonic ~~operation~~ oscillation across L_1C_1. These oscillation which induces an emf in the coil L_2 magnetically linked with L_1. The frequency of emf induced in coil L_2 is the same as that of oscillation in the oscillatory circuit, but the magnitude will depend upon the coupling and turn ratio between L_1 and L_2.

This induced emf is applied at the input of amplifier, due to this, collector current increases but the frequency of amplified output is remain same as the resonant frequency of L_1C_1. The amplified output in the collector circuit will supply power to the oscillator circuit, thus compensating for the losses occurring in it. Hence the oscillation in the oscillatory circuit will be undamped.

In this circuit L_1C_1 and L_2 acts as feedback element. A phase difference of 180° phase shift is obtained by the amplifier as a result of this total phase shift is 0° or 360°. The frequency of oscillation is

$$f_0 = \frac{1}{2\pi\sqrt{L_p C}}$$

Example

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 to 1200 KHz. Find the range of variable capacitor to be used.

Sol: The resonant frequency is given by

$$f_0 = \frac{1}{2\pi\sqrt{L_p C}}$$

$$C = \frac{1}{4\pi^2 f_0^2 L_p}$$

When $f_0 = 400 \, KHz$

$$C = \frac{1}{4\pi^2 (400 \times 10^3)^2 \times 60 \times 10^{-6}} = 2641 \, pF$$

When $f_0 = 1200 \, KHz$

$$C = \frac{1}{4\pi^2 (1200 \times 10^3)^2 \times 60 \times 10^{-6}} = 293 \, pF$$

Hence, the capacitor range required is 293 pF - 2641 pF.

Example:

A tank circuit contains an inductance of 1mH. Find out the range of tuning capacitor value if the resonant frequency ranges from 540 - 1650 KHz.

Sol Given $L = 1 \, mH$

f_0 ranges from 540 KHz to 1650 KHz

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

$$C = \frac{1}{4\pi^2 f_0^2 L}$$

$$C_{max} = \frac{1}{4\pi^2 (540 \times 10^3)^2 \times 10^{-3}} = 86.86 \, pF$$

$$C_{min} = \frac{1}{4\pi^2 (1650 \times 10^3)^2 \times 10^{-3}} = 9.3 \, pF$$

Hence, the value of capacitance ranges from 9.3 pF to 86.86 pF.

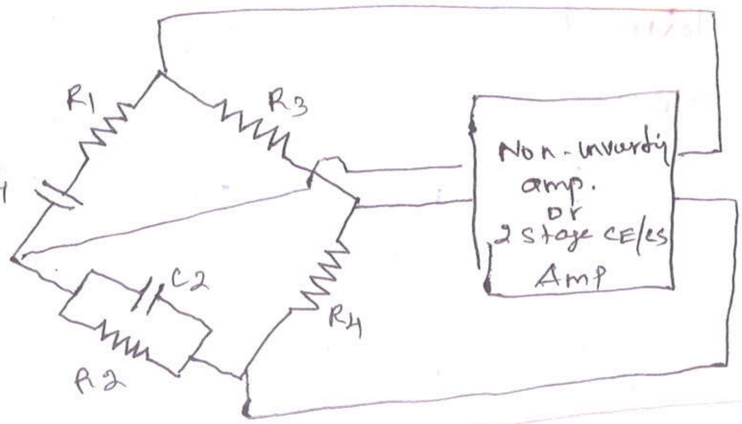
Wien Bridge oscillator

The wien-bridge oscillator is the standard oscillator circuit for all frequencies in the range of 10 HZ to about 1 MHZ. It is the most frequently used type of audio oscillator as the output is free from circuit fluctuations and ambient temperature.

In this oscillator the RC network or weins bridge does not produce any phase shift. Therefore to obtain total phase shift of 360° or 0° two stage CE (BJT) amplifier or ~~of twostage~~ CS (JFET) amplifier or non-inverting (op-amp) amplifier is required.

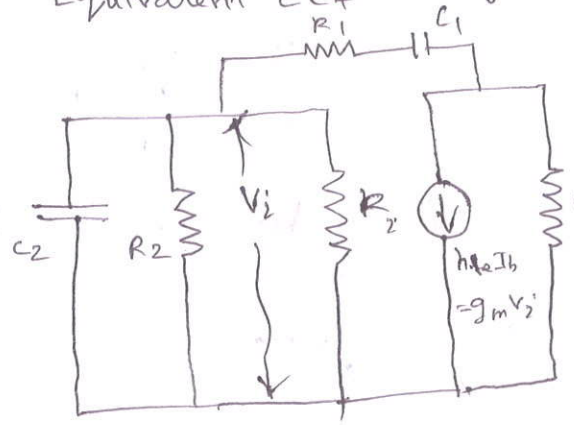
General form of weins bridge OSC.

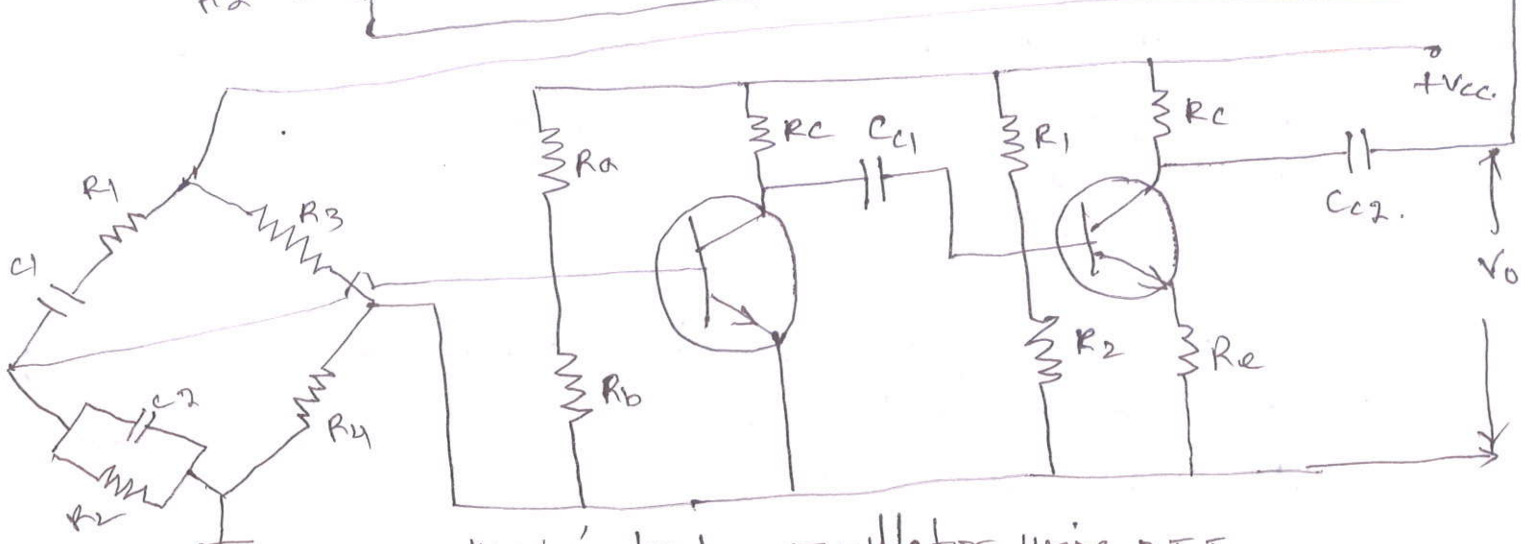
fig (a)



Equivalent ckt

fig (b)





Wein's bridge oscillator using BJT.

Theory

A block diagram of wein's bridge oscillator is shown in fig (a). It consists of a non inverting amplifier and a frequency determining network consisting of R_1C_1 in series R_2C_2 in shunt.

The equivalent circuit of weins bridge oscillation is shown in figure (b). Where R_i includes the input impedance and any biasing resistance of the active device used and R_L consists of output resistance of the device and any load connected to the amplifier. R_1C_1 & R_2C_2 acts as feedback network. The voltage across the parallel combination of R_2C_2 is fed to the input of the amplifier.

The frequency of oscillation is determined by R_1C_1 and R_2C_2. The desired frequency of oscillation can be obtained by varying 2 capacitors or resistors simultaneously. The feedback network provides the positive feedback. In addition to this the resistors R_3 and R_4 provides negative feedback hence this oscillator has better amplitude stability. The resistor R_4 is often a temperature sensitive resistor with a positive temperature coefficient. If the amplitude of oscillation increases the resistance R_2 increases. This reduces the negative feedback which reduces the amplitude of the gain and thus amplitude of oscillations is restored to stable value.

Analysis of wein's bridge oscillator.

The output voltage of the amplifier or across the $R_2 C_2$ is calculated using potential division rule.

$$V_0 = \frac{A V_i \, Z_{parallel}}{Z_{parallel} + Z_{series}} \quad \text{——— (1)}$$

Where $Z_{parallel} = \left(R_2 \parallel \frac{1}{jwC_2}\right) = \frac{R_2\left(\frac{1}{jwC_2}\right)}{R_2 + \frac{1}{jwC_2}}$

$$= \frac{R_2}{R_2 jwC_2 + 1}$$

& $Z_{series} = \left(R_1 + \frac{1}{jwC_1}\right)$

Hence

$$V_0 = \frac{A V_i \left(\frac{R_2}{R_2 jwC_2 + 1}\right)}{\frac{R_2}{R_2 jwC_2 + 1} + \left(R_1 + \frac{1}{jwC_1}\right)}$$

$$V_0 = \frac{A V_i R_2}{R_2 + \left(R_1 + \frac{1}{jwC_1}\right)\left(R_2 jwC_2 + 1\right)} \quad \text{——— (2)}$$

$$\frac{V_0}{V_i} = \frac{A R_2 jwC_1}{(R_1 jwC_1 + 1)(R_2 jwC_2 + 1) + R_2 jwC_1} \quad \text{——— (3)}$$

$$= \frac{A R_2 jwC_1}{-w^2 R_1 C_1 R_2 C_2 + 1 + jw(R_1C_1 + R_2C_2 + R_2C_1)}$$

Multiply the num. & den. by $-j$

$$\frac{V_o}{V_i} = \frac{A R_2 C_1 \omega}{j\omega^2 R_1 C_1 R_2 C_2 - j + \omega(R_1C_1 + R_2C_2 + R_2C_1)} \quad \text{——— (4)}$$

To determine the frequency of oscillation equating the imaginary part to zero.

$$\omega^2 R_1 C_1 R_2 C_2 - 1 = 0 \quad \text{——— (5)}$$

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

$$\omega^2 = \frac{1}{R_1 R_2 C_1 C_2} \quad \text{——— (6)}$$

$$\boxed{f = \frac{1}{2\pi\sqrt{R_1 R_2 C_1 C_2}}} \quad \text{——— 7}$$

If $R_1 = R_2 = R$ and $C_1 = C_2 = C$ then

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

To determine the condition for the oscillation equating real part to zero if $\frac{V_o}{V_i} = 1$

$$\frac{A R_2 C_1 \omega}{\omega(R_1C_1 + R_2C_2 + R_2C_1)} = 1 \quad \text{——— (9)}$$

$$R_1C_1 + R_2C_2 + R_2C_1 = A R_2 C_1 \quad \text{——— (10)}$$

$$\frac{R_1}{R_2} + \frac{C_2}{C_1} + 1 = A \quad \text{——— (11)}$$

If $R_1 = R_2 = R$ & $C_1 = C_2 = C$ then $A = 3$ ——— (12)

hence as per the barkhausen criterion

$$A\beta = 1 \quad \text{or} \quad \beta = 1/A = 1/3 \qquad \text{——— (13)}$$

It is desired value of A and β to obtain the sustained oscillation.

Advantages

* It provides a stable low distortion output over a wide range of frequency.
* It uses both positive and negative feedback hence it provides better stability.
* Overall gain is high because two stages amplifier is used.
* Frequency can be easily adjusted by varying either a resistor or capacitors.

Disadvantages.

* it is costlier because more components are used.
* It cannot be used to generate very high frequency.

Applications.

It is used as standard oscillator for generating frequencies with the range of 10 HZ to 100 KHZ. Hence it is used almost in all commercial audio signal generators.

Limitations of LC and RC oscillators.

The major problems in RC and LC oscillators is that their frequency of operation doesnot remain perfectly constant. This is due to the changes in the value of resistors, inductors and capacitors. However in some application, it is necessary to maintain constant frequency with an extremely low tolerance. Tolerance means the percentage deviation or change in either direction.

The tolerance should be as small as possible. In case of radio broadcasting the tolerance should not be more than 0.002% otherwise the signals of nearby stations will overlap. In such applications the use of LC and RC oscillators is avoided.

The solution to this problem is the use of crystal oscillation. In crystal oscillators, the piezoelectric crystals are used in place of LC or RC circuit. The frequency of crystal oscillators are perfectly constant.

pbm In a wien - bridge oscillator, if the value of R is 100 kΩ and frequency of oscillation is 10 KHz, find the value of capacitor C.

Sol

The operating frequency of a wien-bridge oscillator is given by

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

$$C = \frac{1}{2\pi R f_o}$$

$$= \frac{1}{2\pi \times 100 \times 10^3 \times 10 \times 10^3} = \underline{\underline{159\,PF}}$$

pbm

In the wien bridge oscillator shown in fig (a) $R_1 = R_2 = 220\,k\Omega$ and $C_1 = C_2 = 250\,pF$. Determine the frequency of oscillations.

Sol $R_1 = R_2 = R = 220\,k\Omega = 220 \times 10^3\,\Omega$

$C_1 = C_2 = C = 250\,pF = 250 \times 10^{-12}\,F$

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

$$= \frac{1}{2\pi \times 220 \times 10^3 \times 250 \times 10^{-12}}\ Hz$$

$$\boxed{f = 2892\ Hz.}$$

Twin T oscullator.

A twin T network is a basic lead lag network (ie a combination of low pass and high pass network) its phase characterstics is shown below.

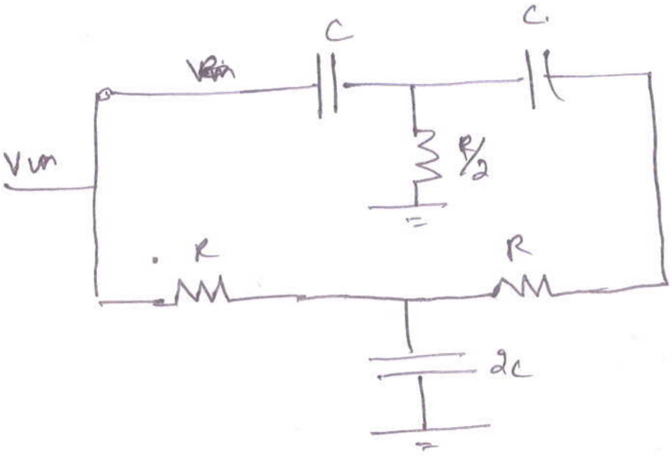
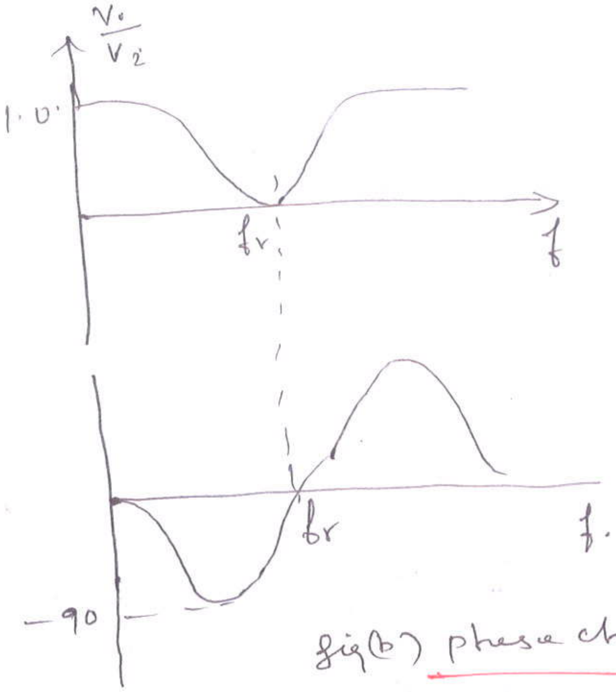


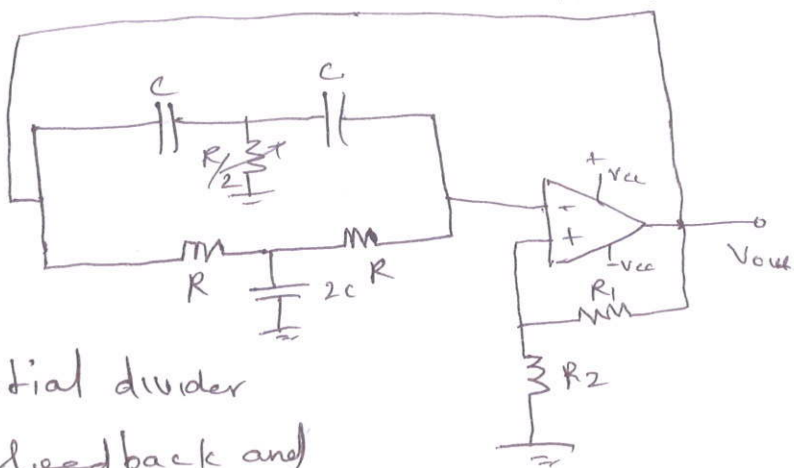
fig ⓐ Twin T Network



fig(b) phase chur

The frequency fr is the resonant frequency at which the phase shift equals 0°. The voltage gain is maximum at low and high frequencies and minimum at the resonant frequency fr.

Twin T-oscullator



The resonant frequency of the network is $f_r = \frac{1}{2\pi RC}$.

fig ⓒ shows Twin T oscullator. The resistor R1 and Rc acts as potential divider and provides positive feedback and Twin T network acts as negative feedback.

fig ⓒ Twin T oscullator.

When V_{cc} is applied the resistance R is low and the positive feedback is maximum, hence the capacitor is charged toward the maximum voltage, it will cause damped oscillations. As oscillation builds up, the resistance R_1 increases and positive feedback decreases. If the positive feedback decreases the oscillations become constant. The oscillation cannot occur at any frequency other than f_r, because the frequency other than f_r, the negative feedback provided by Twin T network not allowed for oscillation i.e., only at f_r the negative feedback is neglected. Thus to ensure that the oscillation frequency is near f_r, $R/2$ of the twin T network is kept variable.

The frequency of oscillation of Twin T oscillator is slightly different from resonant frequency thus the resistance $\frac{R}{2}$ may be slightly varied and the ratio $\frac{R_1}{R_2}$ with in the range of 10 to 1000. This results in the oscillator forced to operate near the resonant or notch frequency.

The main draw back of this oscillator is that it is suitable for single frequency and cannot be easily adjusted to other large frequency range.

Analysis.

By using the star - delta conversion formulae, the Twin-T network show in fig (a) can be modified as equivalent delta network shown below.

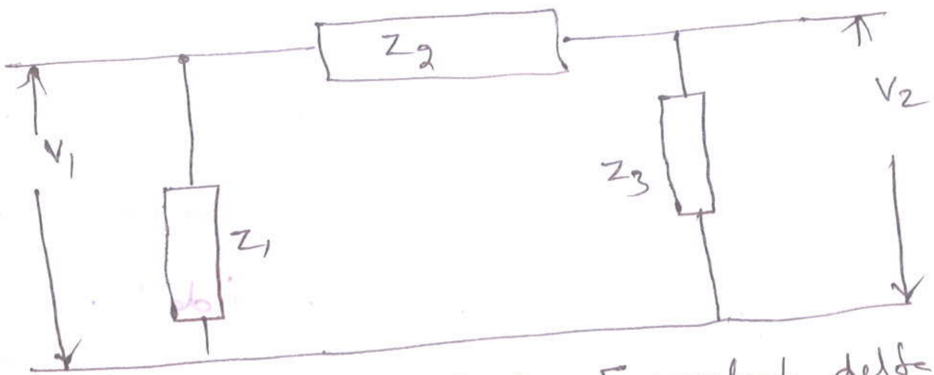


fig (d) Equivalent delta network

$$Z_1 = Z_3 = \frac{sCR+1}{2sC}$$

$$Z_2 = \frac{2R(sCR+1)}{s^2C^2R^2+1}$$

From fig (d) we can get the relation expressed by

$$\frac{V_2}{V_1} = \frac{s^2C^2R^2+1}{s^2C^2R^2+4sCR+1}$$

Sub $s = j\omega$

$$\frac{V_2}{V} = \frac{1-\omega^2C^2R^2}{1-\omega^2C^2R^2+j\omega RC}$$

By equ the real part to zero.

$$\boxed{f_0 = \frac{1}{2\pi RC}}$$

Frequency Range of RC and LC oscillator.

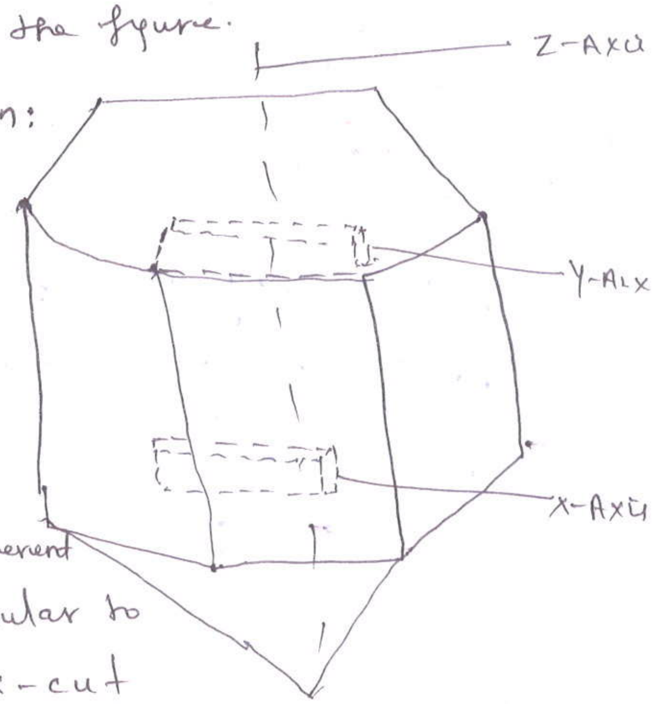
Type of oscillator	Frequency	Applications
① Rc oscillators		
ⓐ RC phase shift osc.	20 to 20 KHZ	AF appli.
ⓑ Wein bridge osc.	up to 100 KHZ	
② LC oscillators		
ⓐ Resonant circuit osc.	20 KHZ to 3 MHZ	RF. appli
ⓑ crystal oscillators.	3 MHZ to 30 MHZ	
③ Very high freq. osc.	30 MHZ to 300 MHZ	TV and Radio
④ UHF oscillators.	300 MHZ to 3 GHZ	TV, broadcasting
⑤ Microwave osc.	3 GHZ to several GHZ	Radar.

crystal oscillators.

The crystals are either naturally occuring or synthetically manufactured exhibiting the piezoelectric effect. The piezoelectric effect means under the influence of the mechanical pressure, the voltage gets generated across the opposite faces of the crystal. If the mechanical force is applied in such a way to force the crystal to vibrate, the a.c. voltage gets generated across it.

Conversely, if the crystal is subjected to a.c. voltage, it vibrates causing mechanical distortion in the crystal shape. Every crystal has its own resonating frequency depending on its cut. So under the influence of the mechanical vibrations, the crystal generates an electrical signal of very constant frequency. A crystal oscillator is basically a tuned-circuit oscillator using a piezoelectric crystal as its resonant tank circuit. The crystal oscillators are preferred when greater frequency stability is required.

Quartz is inexpensive and easily available in nature and hence very commonly used in the crystal oscillators. The natural shape of quartz crystal is hexagonal as shown in the figure.



The three axes are shown: the z-axis is called the optical axis, the x-axis is called the electrical axis and y-axis is called the mechanical axis. Quartz crystal can be cut in different ways. Crystal cut perpendicular to the the x-axis is called x-cut crystal, whereas the cut perpendicular to y axis is called y-cut crystal.

The piezoelectric properties of a crystal depend upon its cut.

Frequency of crystal.

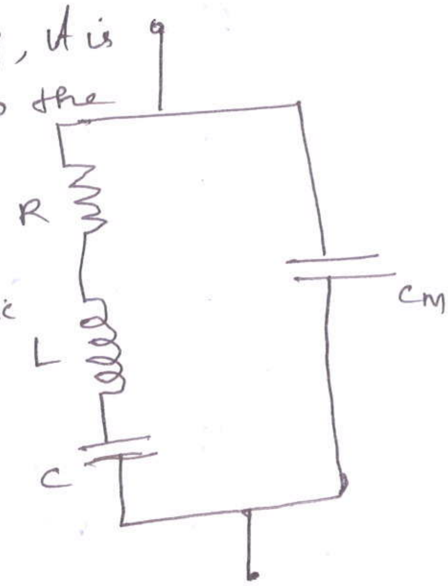
Each crystal has a natural frequency like pendulum. The natural frequency f of a crystal is given by;

$$\boxed{f = \frac{K}{t}}$$

Where k is a constant that depends upon the cut and t is the thickness of the crystal. It is clear that frequency is inversely proportional to crystal thickness. The thinner the crystal, the greater is its natural frequency and vice-versa. However, extremely thin crystal may break because of vibration. This puts a limit to the frequency obtainable. In practice, frequencies between 25 KHz to 5 MHz have been obtained with crystal.

Equivalent circuit of crystal.

When the crystal is not vibrating, it is equivalent to a capacitance due to the mechanical mounting of the crystal. Such a capacitance existing due to the two metal plates separated by a dielectric like crystal slab is called as mounting capacitance denoted as C_M or C''.



When it is vibrating, there are internal frictional losses which are denoted by a resistance R. While the mass of the crystal, which is indication of its inertia is represented by an inductance L. In vibrating condition, it is having some stiffness, which is represented by a capacitor C. The mounting capacitance is a shunt capacitance. And hence the overall equivalent circuit of a crystal can be shown in fig.

RLC forms a resonating circuit. The expression for the resonating frequency f_r is

$$\boxed{f_r = \frac{1}{2\pi\sqrt{LC}} \sqrt{\frac{Q^2}{1+Q^2}}}$$

Q = Quality factor of crystal.

$$\boxed{Q = \frac{\omega L}{R}}$$

The Q factor of the crystal is very high, typically 20000. Value of Q up to 10^6 also can be achieved.

Hence $\sqrt{\frac{Q^2}{1+Q^2}}$ factor approaches to unity and we get the resonating frequency as

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

The crystal has two resonating frequencies, series resonant frequency and parallel resonant freq

Series resonance frequency

The series resonance frequency is same as the resonating frequency given by equ ⊛

$$\boxed{f_s = \frac{1}{2\pi\sqrt{LC}}}.$$

parallel resonance frequency.

The other resonant condition occurs when the reactances of series resonant leg equals the reactance of the mounting capacitor Cm. This is parallel resonance or antiresonance condition. under this condition the impedance offered by the crystal to the external circuit is maximum.

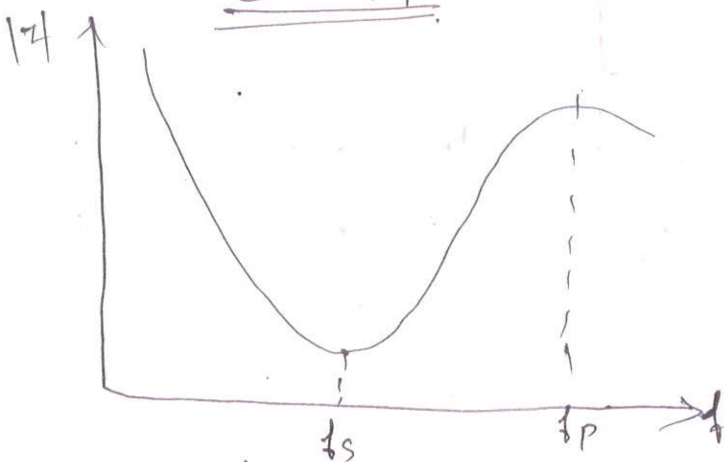
under parallel resonance, the equivalent capacitance is,

$$\boxed{C_{eq} = \frac{C_M C}{C_M + C}}$$

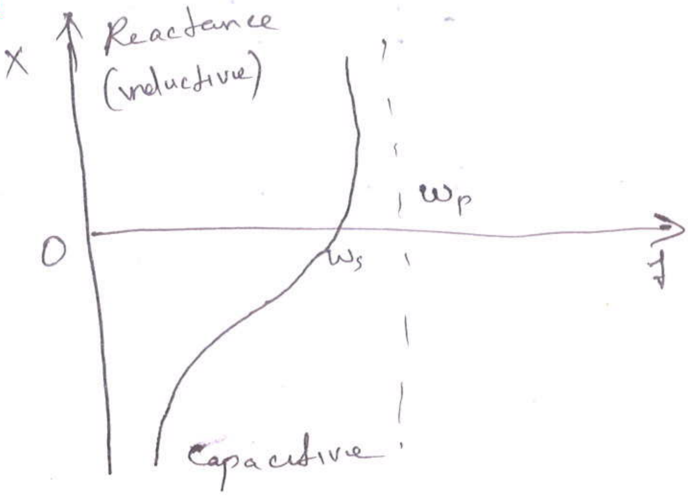
Hence the parallel resonating frequency is given by

$$\boxed{f_p = \frac{1}{2\pi\sqrt{LC_{eq}}}}$$

$C << M$



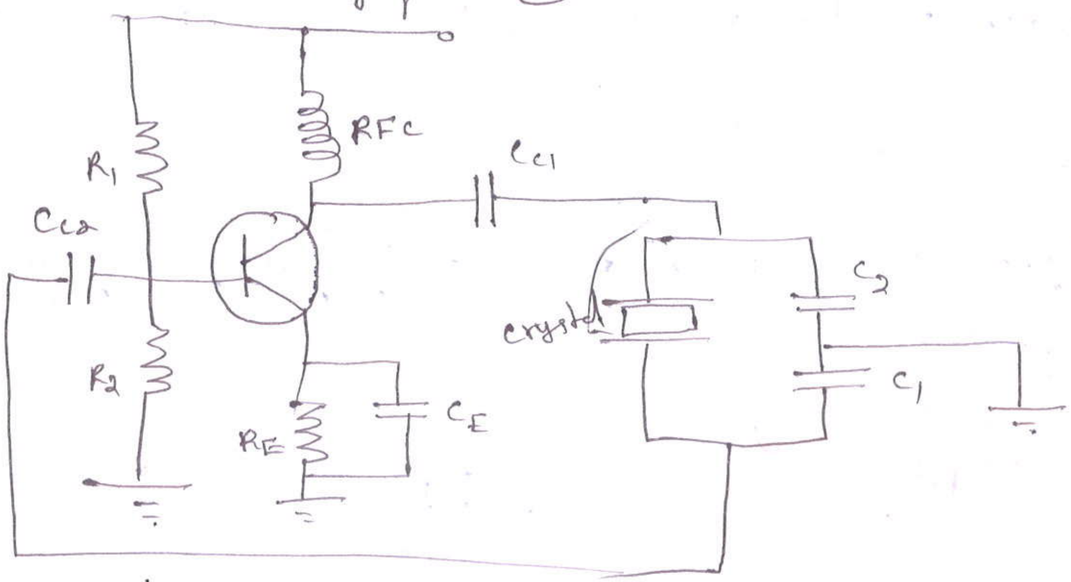
Crystal impedance Vs freq.



Reactance Vs freq.

pierce crystal oscillator.

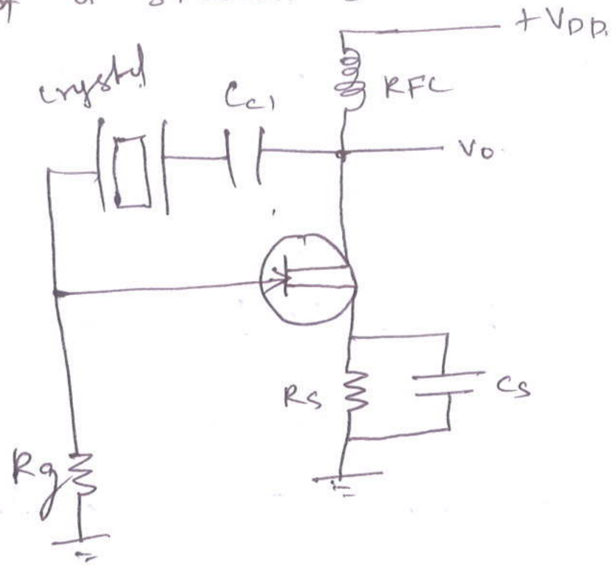
The colpitts oscillator can be modified by using the crystal to behave as an inductor. The circuit is called pierce crystal oscillator. The crystal behaves as an inductor for a frequency slightly higher than the series resonance frequency f_s. The two capacitors C_1 & C_2 required in the tank circuit along with an inductor are used, as they are used in colpitts oscillator circuit. As only inductor gets replaced by the crystal, which behaves as an inductor, the basic working principle of pierce crystal oscillator is same as that of colpitts oscillator. The practical transistorised pierce crystal oscillator ckt is shown in figure (a)



The resistances R_1, R_2, R_E provide d.c. bias while the capacitor C_E is emitter bypass capacitor. RFC (Radio frequency choke) provides isolation b/w a.c & d.c. operation C_{c1} & C_{c2} are coupling capacitors.

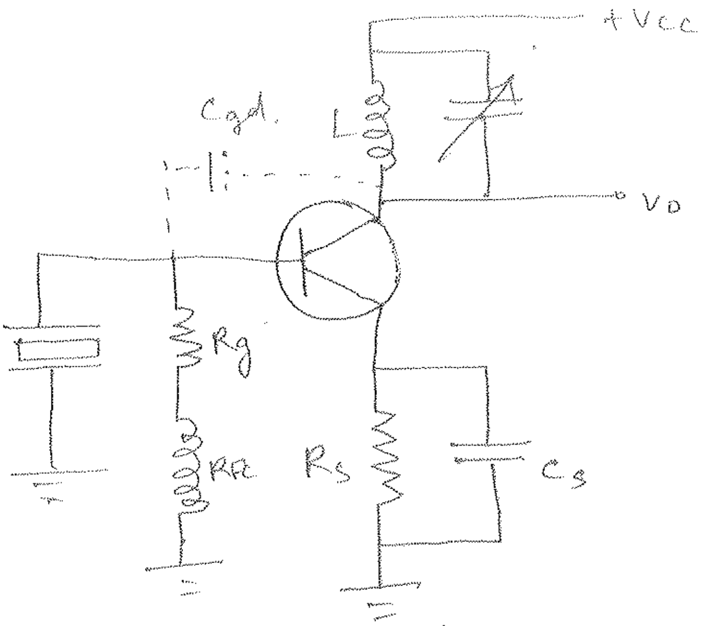
The resulting circuit frequency is set by the series resonant frequency of the crystal, change in the supply voltages, temperature, transistor parameters have no effect on the circuit operating conditions and hence good frequency stability is obtained.

The oscillator circuit can be modified by using the internal capacitors of the transistor used instead of C_1 and C_2. The separate capacitors C_1, C_2 are not required in such circuit. Such circuits using FET ~~and~~ is shown below.



Miller Crystal oscillator.

The Hartley oscillator circuit can be modified to get Miller crystal oscillator. In Hartley oscillator circuit, two inductors and one capacitor is required in the tank circuit. One inductor is replaced by the crystal which acts as an inductor for the frequencies slightly greater than the series resonant frequency. The transistorised Miller crystal oscillator circuit is shown in fig ①.



- The tuned circuit of L_1 and C is offtuned to behave as an inductor ie L_1. The crystal behaves as other inductance L_2 between base and ground. The internal capacitance of the transistor acts as a capacitor required to fulfill the elements of the tank circuit. The crystal decides the operating frequency of the oscillator. Due to its low output power, Miller oscillator is not often used in high frequency applications. It is used in the applications which require an o/p of moderate power and good stability at a specific freq. It is commonly employed in trigger circuits, Sawtooth generates, phase control and timing ckts.

Frequency stability of oscillators.

The frequency stability of an oscillator is a measure of its ability to maintain a constant frequency, over a long time interval. However, it has been found that if an oscillator is set at some particular frequency, it does not maintain it for a longer period. In other words, the frequency of an oscillator changes slowly from the initally set value. sometimes, the change in frequency is uniform in one direction. But at some other times it may be changing quite erratically. The change in oscillation frequency may arise due to the following factors.

① Operating point of the active device

The operating point of the active device is selected in such a way so that its operation in non-linear region, changes the values of device parameters which in turn, affects the frequency stability of the oscillator.

② Circuit components.

The values of circuit components (R, L & C) change with ~~temp~~ variation in temperature. Since such changes take place slowly, they also cause a drift in oscillator frequency.

③ Supply voltage.

The changes in d.c. supply voltage applied to the active device, shift the oscillator frequency. This problem can be avoided by using a highly regulated power supply.

④ output load.

A change in the o/p load may cause a change in the Q factor of the tank ckt, thereby causing a change in the oscillator o/p frequency.

⑤ Interelement capacitances and stray capacitances.

Any change in the interelement capacitance of a transistor, cause changes in the oscillator o/p frequency and thus affect the frequency stability. Similarly, the stray capacitances, also affect the frequency stability of an oscillator. The effect of change in interelement capacitances can be neutralized by putting an additional capacitor across the corresponding elements. However it is difficult to avoid the effect of stray capacitances.

Comparison b/w crystal and LC oscillators.

Sl.No	LC oscillators	crystal osc.
1	The separate L and C components are necessary in the tuned circuit	The single crystal serves the purpose of tuned circuit.
2	The Q value of LC tuned circuit is less as compared to the crystal	The Q value is much higher than LC tuned circuit
3	The frequency stability is less	Very high frequency stability
4	The bandwidth is more.	The bandwidth is very small.
5	The effect of temperature on the frequency is more severe	The effect of temperature on freq. is negligible
6	The freq. range which can be generated is more	The limit to the frequency generated due to thickness of the crystal.
7	used in general purpose application like signal generator.	used in specific appl. when need high freq. stability like watches, computers, counters.

A crystal has $L = 0.33\ H$, $C = 0.065\ pF$ and $C_m = 1\ pF$ with $R = 5.5\ k\Omega$. Find (i) f_s (ii) f_p (iii) By what % does the parallel resonant frequency exceed the series resonant frequency? (iv) Find the Q factor of the crystal.

Sol

(i) $$f = \frac{1}{2\pi\sqrt{LC}} = 1.087\ MHz$$

(ii) $$C_{eq} = \frac{C C_m}{C + C_m} = 0.061\ PF$$

$$f_p = \frac{1}{2\pi\sqrt{L C_{eq}}} = 1.121\ MHz$$

(iii) $$\%\ \text{increase} = \frac{1.121 - 1.087}{1.087} \times 100 = 3.127\%$$

(iv) $$Q = \frac{\omega_s L}{R} = \frac{2\pi f_s L}{R} = 409.789.$$

* A quartz crystal has the following constants, $L = 50\ mH$, $C_1 = 0.02\ pF$, $R = 500\ \Omega$ and $C_2 = 12\ pF$. Find the values of f_s and f_p. If external capacitance across the crystal changs from 5pF to 6pF, Find the change in frequency of oscillatori

Sol

$L = 50\ mH$, $C_1 = 0.02\ pF$, $R = 500\ \Omega$, $C_2 = 12\ pF$

$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2} = \frac{0.02 \times 12 \times 10^{-12} \times 10^{-12}}{[0.02 + 12] \times 10^{-12}} = 0.01996\ pF$$

$$f_s = \frac{1}{2\pi\sqrt{L C_1}} = 5.0329\ MHz \quad \text{and} \quad f_p = 1/2\pi\sqrt{L C_{eq}} = 5.0379\ MHz$$

Let $C_s = 5\ pF$ connected across the crystal

$$C_2' = C_2 + C_x = 12 + 5 = 17\ PF$$

$$C_{eq}' = \frac{C_1 C_2'}{C_1 + C_2} = 0.019976\ pF$$

$$f_p' = 1/2\pi\sqrt{L C_{eq}'} = 5.03588\ MHz$$

New $C_x = 6\ pF$ is connected then

$$C_2'' = C_2 + C_x = 12 + 6 = 18\ pF$$

$$C_{eq}'' = \frac{C_1 C_2''}{C_1 + C_2''} = 0.0199778\ pF$$

$$f_p'' = \frac{1}{2\pi\sqrt{L C_{eq}''}} = 5.035716\ MHz$$

$$\text{Change} = f_p' - f_p'' = (5.03588 - 5.035716) \times 10^6 = 164\ Hz$$

2-

pbm

* A crystal has a thickness of t mm. If the thickness is reduced by 1% what happens to frequency of oscillators ?.

$$\text{Frequency}, \ f = \frac{K}{t}$$

$$f \propto 1/t$$

If the thickness of the crystal is reduced by 1%, the frequency of oscillations will increase by 1%.

* The ac equivalent circuit of a crystal has these values $L = 1H$, $C = 0.01PF$, $R = 1000\Omega$ and mounting capacitance (C_m) is $20PF$. Calculate f_s and f_p of the crystal.

$L = 1H$

$C = 0.01PF = 0.01 \times 10^{-12} F$

$C_m = 20pF = 20 \times 10^{-12} F$

$$f_s = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{1 \times 0.01 \times 10^{-12}}} Hz$$

$$= 1589 \times 10^3 Hz = 1589 KHz$$

$$C_T = \frac{C \times C_m}{C + C_m} = \frac{0.01 \times 20}{0.01 + 20} = 9.99 \times 10^{-3} pF$$

$$= 9.99 \times 10^{-15} F$$

$$f_p = \frac{1}{2\pi\sqrt{LC_T}} = \frac{1}{2\pi\sqrt{1 \times 9.99 \times 10^{-15}}} Hz$$

$$= 1590 \times 10^3 Hz = 1590 KHz.$$

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 - large signal tuned amplifiers - class c tuned amplifier Efficiency and applications of class c tuned amplifier - Stability of tuned amplifiers - Neutralization - Hazeltine neutralization method.

Introduction

In order to pickup and amplify the desired frequency signal, the resistive load (R_C) in an amplifier is replaced by a tuned circuit. This tuned circuit is capable of selecting a particular frequency and rejecting all the other frequency. An amplifier with this tuned circuit as a load is known as Tuned amplifier.

These amplifier uses a parallel tuned circuit as a load because, it has high impedance at its frequency of resonance and the impedance falls off sharply as the frequency departs from the frequency of resonance.

Tuned amplifiers used for amplifying narrow band of frequencies hence it is also known as "narrow band amplifier" or Bandpass amplifier.

Need for Tuned amplifiers.

- In a radio recievers or TV receiver, it is necessary to select a particular selective circuit is needed that will allow as to amplify the frequency band required and reject all the other unwanted signals. Such a circuit is known as Tuned amplifiers.

A tuned circuit generally uses either a variable capacitor or variable inductor for adjusting the resonant frequency at the centre of the band of frequencies to be amplified. Over this narrow band of frequencies the gain of the circuit is more or less constant.

- In tuned amplifiers, harmonic distortion is very small because, the impedance and gain of the amplifier becomes negligibly small at hormonic frequencies.

Tuned amplifiers may be divided into two categories such as

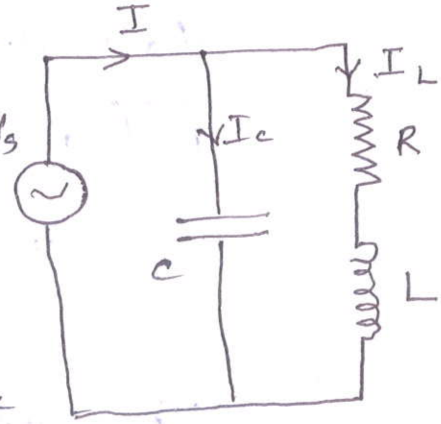
① small ~~single~~ signal tuned amplifier

② Large signal tuned amplifier

"Small Signal tuned amplifier" are operated in class A mode and power involved is small thus the distortion is also negligibly small.

• "Large signal tuned amplifier" amplifies the large signal at radio frequencies (RF) and power involved is large. Hence these amplifiers operated under class B or class C mode. However the distortion gets increased, but the tuned circuit itself eliminates most of the hormonic distortion.

Parallel resonant or Tuned circuit



The parallel resonant circuit consists of an inductor 'L' and capacitor 'C' connected in parallel to each other. "R represents the coil resistance as shown in the fig. The supply voltage Vs having constant amplitude, but the frequency is variable. If the frequency of applied voltage is equal to the resonant frequency of LC circuit, the electrical resonance occurs.

At resonance, the impedance of the circuit is maximum and the line current is minimum and the powerfactor is unity.

Resonant frequency.

The admittance of the inductive branch

$$Y_L = \frac{1}{R_L + j\omega L}$$

the admittance of capacitive branch

$$Y_c = \frac{1}{1/j\omega C} = j\omega C.$$

Total admittance

$$Y_t = Y_L + Y_c$$

$$= \frac{1}{R + j\omega L} + j\omega C$$

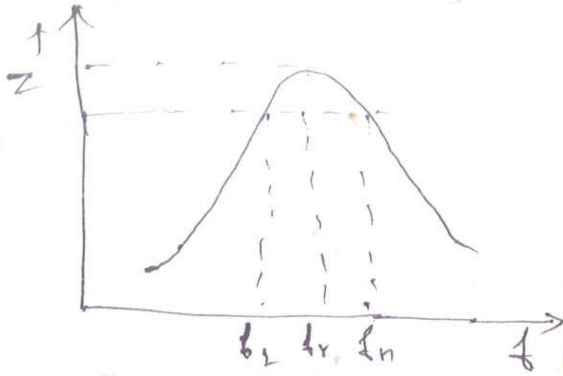
Rationalizing the above equation and separate real & imaginary part.

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

The condition for the resonance is imaginary part is equated to zero

$$\omega C - \frac{\omega L}{R^2 + \omega^2 L^2} = 0$$

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

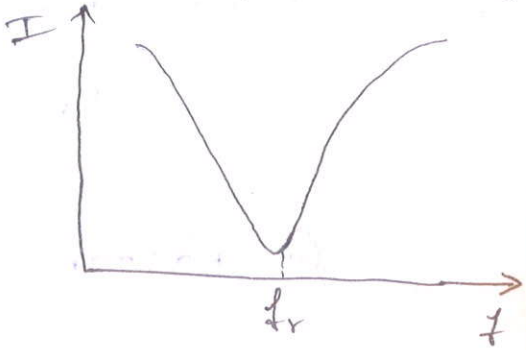


$$\omega C\left[R^2+\omega^2L^2\right] = \omega L$$

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

$$\omega^2L^2 = \frac{L}{C} - R^2$$

$$\omega^2 = \frac{1}{LC} - \left(\frac{R}{L}\right)^2$$



$$\omega = \frac{1}{\sqrt{LC}} - \frac{R}{L}$$

$$f = \frac{1}{2\pi\sqrt{\frac{1}{LC} - \left(\frac{R}{L}\right)^2}}$$ if $\frac{R}{L} << 1$ then

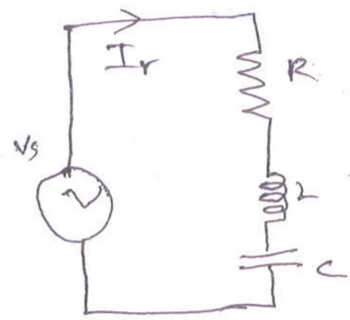
$$\boxed{f_r = \frac{1}{2\pi\sqrt{LC}}}$$ = resonant frequency.

Series resonant circuit.

Total impedance of the series resonant circuit is given by

$$Z_s = \sqrt{R^2+(X_L-X_c)^2}$$

As frequency increases, X_L increases and X_c decreases. At resonant frequency $X_L = X_c$ and two curves intersect each other.



condition for resonance is $X_L - X_c = 0$

$$X_L = X_c$$

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

thus resonant frequency $f = \frac{1}{2\pi\sqrt{LC}}$

At resonance $Z_R = R$ $(\because X_L - X_C = 0)$

so impedance is minimum and the current I_r is maximum

$$I_r = \frac{V}{R}$$

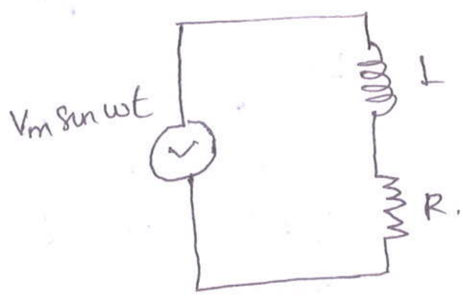
Q - Factor

In practice, the inductor possesses a small resistance in addition to its inductance. The lower the value of this resistance, the better the Q-factor of the inductor.

The Q-factor or quality factor of an inductor at operating frequency 'ω' is defined as the ratio of impedance of the coil to its resistance and can be defined as

$$Q = 2\pi \times \frac{\text{Maximum energy stored per cycle}}{\text{Energy dissipated per cycle}} \quad \text{——} (*)$$

let
Sinusoidal voltage - $V_m \sin \omega t$



I_m = peak value of the current in the circuit.

Maximum energy stored per cycle.

$$= \frac{1}{2} L I_m^2.$$

$P = V \times i$
$= L i \cdot \frac{di}{dt}$.

Average power dissipated in the inductor per cycle.

$$= \left(\frac{I_m}{\sqrt{2}}\right)^2 R.$$

Energy absorbed
$= \frac{1}{2} L i^2$

Energy = power x time.

Energy dissipated in inductor percycle.

$$= \text{power} \times \text{periodic time for one cycle.}$$

$$= \left(\frac{I_m}{\sqrt{2}}\right)^2 R \times T$$

$$= \left(\frac{I_m}{\sqrt{2}}\right)^2 R \times \frac{1}{f}$$

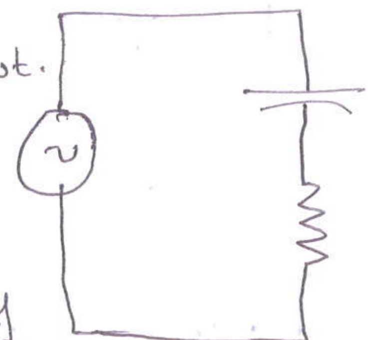
$$= \frac{I_m^2 R}{2f}$$

sub. in equ ⊛

$$\boxed{Q = 2\pi \times \frac{(1/2)\cancel{I_m^2} L}{\frac{\cancel{I_m^2} \cdot R}{2f}} = \frac{2\pi f L}{R} = \frac{\omega L}{R}}$$

$$\boxed{Q = \frac{X_L}{R}}$$

Q factor of a capacitor



The capacitor also possesses a small resistor in series with it.

The Q-factor of a capacitor at the operating frequency 'ω' is the ratio of reactance of the capacitor to its series resistor (R).

$i = C \frac{dv}{dt}$

$p = vi = vC\frac{dv}{dt}$ / $W = \frac{1}{2} C V^2$

Maximum energy stored in capacitor per cycle.

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

V_{max} = maximum value of capacitor

When $R << \left(\frac{1}{\omega C}\right)$

$$V_{max} = \frac{I_m}{\omega C}$$

Max. energy stored in capacitor per cycle

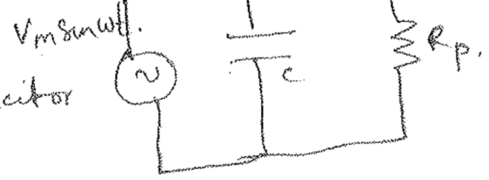
$$= \frac{1}{2} C V^2 max = \frac{1}{2} \cdot \frac{I_m^2}{\omega^2 \cdot C}.$$

$$\text{Energy dissipated per cycle} = \frac{I_m^2 R}{2f}$$

$$Q = 2\pi \left[\frac{\frac{\cancel{I_m^2}}{\cancel{2}\omega^2 \cdot C}}{\frac{\cancel{I_m^2} \cdot R}{\cancel{2} f}} \right] = \frac{1}{\omega C R}$$

$$\boxed{Q = \frac{1}{\omega C R}.}$$

But, often a leaky capacitor is represented by a capacitor 'c' with a high resistance 'R_p' in shunt as shown in fig.



Maximum energy stored in capacitor

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

$$= \frac{1}{2} C V_m^2$$ V_m = maximum value of applied voltage.

Average power dissipated per cycle in R_p

$$= \left(\frac{V_m}{\sqrt{2}}\right)^2 \times \frac{1}{R_p} = \frac{V_m^2}{2R_p}.$$

Total energy dissipated per cycle

$$= \frac{V_m^2}{2R_p} \times T$$

$$= \frac{V_m^2}{2R_p} \times \frac{1}{f} = \frac{V_m^2}{2R_p f}.$$

$$Q = 2\pi \cdot \left[\frac{\frac{1}{2} C V_m^2}{\frac{V_m^2}{2R_p f}}\right] = \omega C R_p.$$

$$\boxed{Q = \omega C R_p}$$

The above equ gives the quality factor of a lossy 'c' with high resistance "R_p" in shunt.

Bandwidth

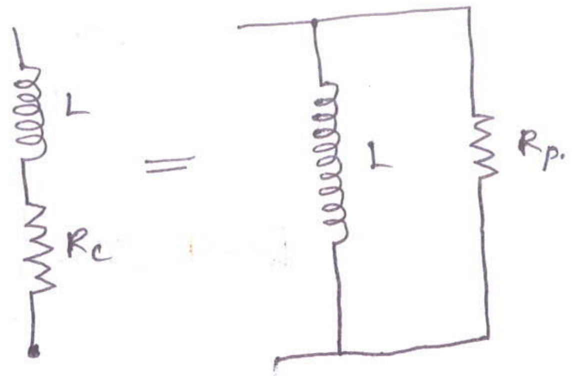
The bandwidth is given by

$$\boxed{B \cdot W = \frac{f_0}{Q_0}.}$$

Inductor (or) Coil losses.

The tuned circuit consists of 'L' and 'C', practically the inductor 'L' is not purely inductive. It consists of power loss in the inductor is usually represented by a series resistance r_s as shown in fig. below. However, rather than specifying the value of r_c the usual practice is to specify the quality factor.

$$Q = \frac{\omega_0 L}{r_c}$$



The analysis of a tuned amplifier is greatly simplified by representing the inductor loss by a parallel resistance R_p. The relationship between R_p and Q_0 can be derived as follows.

The admittance of the coil $Y_L = \frac{1}{r_c + j\omega L} = \frac{1}{j\omega L}\left[\frac{1}{1+\frac{r_c}{j\omega L}}\right]$ —① .

$$Y_L = \frac{1}{j\omega_0 L}\left[\frac{1}{1+j\frac{r_c}{\omega_0 L}}\right] = \frac{1}{j\omega_0 L}\left[\frac{1}{1-j\frac{1}{Q_0}}\right]$$

$$\boxed{Q_0 = \frac{\omega_0 L}{R}}$$

$$= \frac{1}{j\omega_0 L}\left[\frac{1+j\left(\frac{1}{Q_0}\right)}{1^2+\left(\frac{1}{Q_0}\right)^2}\right] \text{ —— ②.}$$

$Q_0 >> 1$ then $Y_L = \frac{1}{j\omega_0 L}\left[1 + j\left(\frac{1}{Q_0}\right)\right]$ ——— (3)

equating equ (1) and (3) we get

$$Y_L = \frac{1}{j\omega_0 L}\left[1 + j\left(\frac{1}{Q_0}\right)\right] = \frac{1}{r_s + j\omega_0 L} = \frac{r_s - j\omega_0 L}{r_s^2 + \omega_0^2 L^2}$$

$$\frac{1}{j\omega_0 L} + \frac{1}{\omega_0 L Q_0} = \frac{r_s}{r_s^2 + \omega_0^2 L^2} - \frac{j\omega L}{r_s^2 + \omega_0^2 L^2}$$

$$\frac{1}{j\omega_0 L} + \frac{1}{\omega_0 L Q_0} = \frac{1}{R_p} - \frac{1}{j\omega L_p} \quad \text{——— (4)}$$

equating real and imaginary parts we get

$$\omega_0 L Q_0 = R_p$$

$$\boxed{Q_0 = \frac{R_p}{\omega_0 L}}$$

Where $R_p = \frac{r_s^2 + \omega_0^2 c^2}{r_s}$ = coil loss.

unloaded and loaded Q of tank circuits

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

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

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

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

Due to the loading, the equivalent parallel resistance R_P is reduced by any external resistance R_{EXT} placed in parallel with the circuit. Therefore, the loaded Q-factor is given by

$$Q_L = \frac{R_T}{X_L} = \frac{R_T}{WL_P}$$

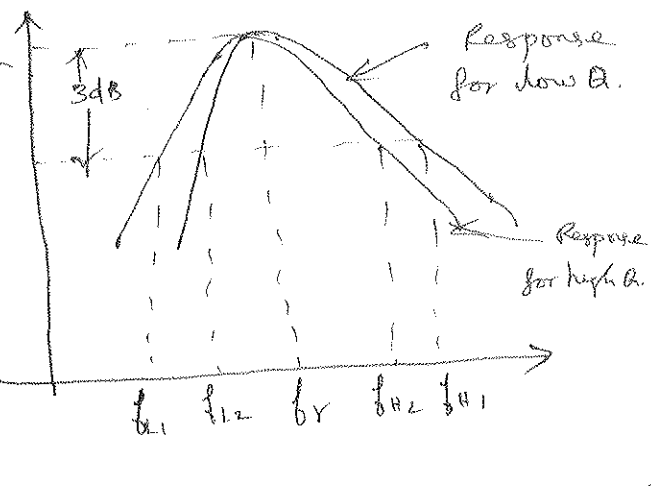
Where, $R_T = R_P \,||\, R_{EXT}$

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 fig.



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.

pbm

A tuned amplifier has its maximum gain at a frequency of 2 MHZ and has a bandwidth of 50 KHZ. calculate the Q-factor.

$$f_r = 2\ MHZ \qquad BW = 50\ KHZ$$

$$Q.\text{factor} - Q = \frac{f_r}{BW} = \frac{2 \times 10^6}{50 \times 10^3} = 40.$$

pbm

A tuned amplifier is designed to receive AM broadcast of speech signal at 650 KHZ. What is needed Q for amplifier ?

Given $f_c = 650\ HZ$

Assume max. modulating frequency for AM broadcast speech signal = 3KHZ.

$$\text{Bandwidth} = 2 f_m = 2 \times 3 = 6\ KHZ$$

$$Q = \frac{f_r}{BW} = \frac{650\ KHZ}{6\ KHZ} = 108.33.$$

pbm

A tank circuit contains an inductance of 1mH. Find out the range of tuning capacitor value if the resonant frequency ranges from 540 KHZ to 1650 KHZ.

For $f_r = 540\ KHz$

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

$$540 \times 10^3 = \frac{1}{2\pi\sqrt{1 \times 10^{-3} \times C}}$$

$$C = 86.86\ pF.$$

For $f_r = 1650\ KHz$

$$f_r = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{1 \times 10^{-3} \times C}} = 1650 \times 10^3$$

$$C = 9.3\ pF.$$

pbm

A parallel resonant circuit has an inductance of 150 μH and a capacitance of 100 pF. Find the resonant frequency.

$L = 150\ \mu H$ & $C = 100\ pF$

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{150 \times 10^{-6} \times 100 \times 10^{-12}}} = 1299.494\ KHz.$$

pbm

A parallel resonant circuit has a capacitor of 100 pF and an inductor of 100 micro H. The inductor has a resistance of 5 ohms. Find the value of frequency at which the circuit resonates and the circuit impedance at resonance.

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{2\pi\sqrt{100 \times 10^{-6} \times 100 \times 10^{-12}}} = 1.59155\ MHz$$

Impedance at resonance is given by

$$R_p = \frac{\omega^2 L^2}{\omega_p} = \frac{(2\pi \times 1.59155 \times 10^6)^2 \times (100 \times 10^{-6})^2}{5} = 200\ k\Omega$$

Small Signal tuned Amplifiers

In order to obtain a large overall voltage gain, it is required to use a number of tuned amplifiers stages in cascade. These cascaded tuned amplifiers may be classified as (I) single tuned amplifiers
(II) Double tuned amplifiers
(III) Stagger tuned amplifiers.

Single tuned amplifiers use one parallel resonant circuit as load impedance in each stage and all the tuned circuits are tuned in the same frequency.

Double tuned amplifiers uses two inductively coupled tuned circuits per stage, both the tuned circuits being tuned to the same frequency.

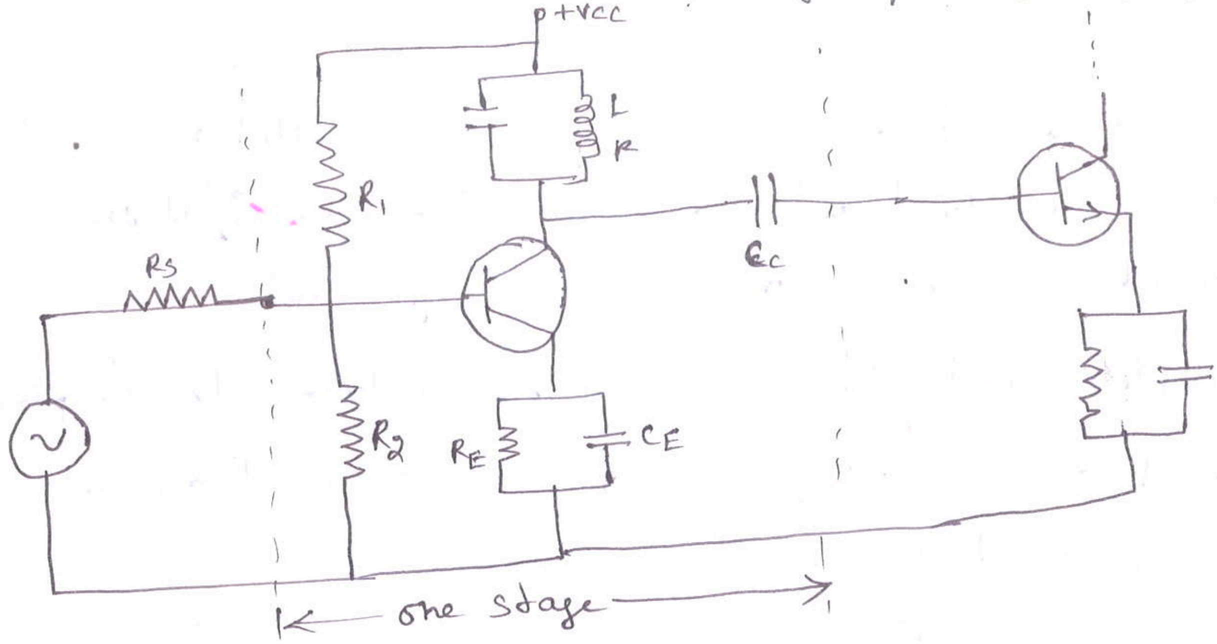
Stagger tuned amplifiers use a number of single tuned stages in cascade, the successive tuned circuits being tuned to slightly different frequencies

Single tuned amplifiers can be further classified as

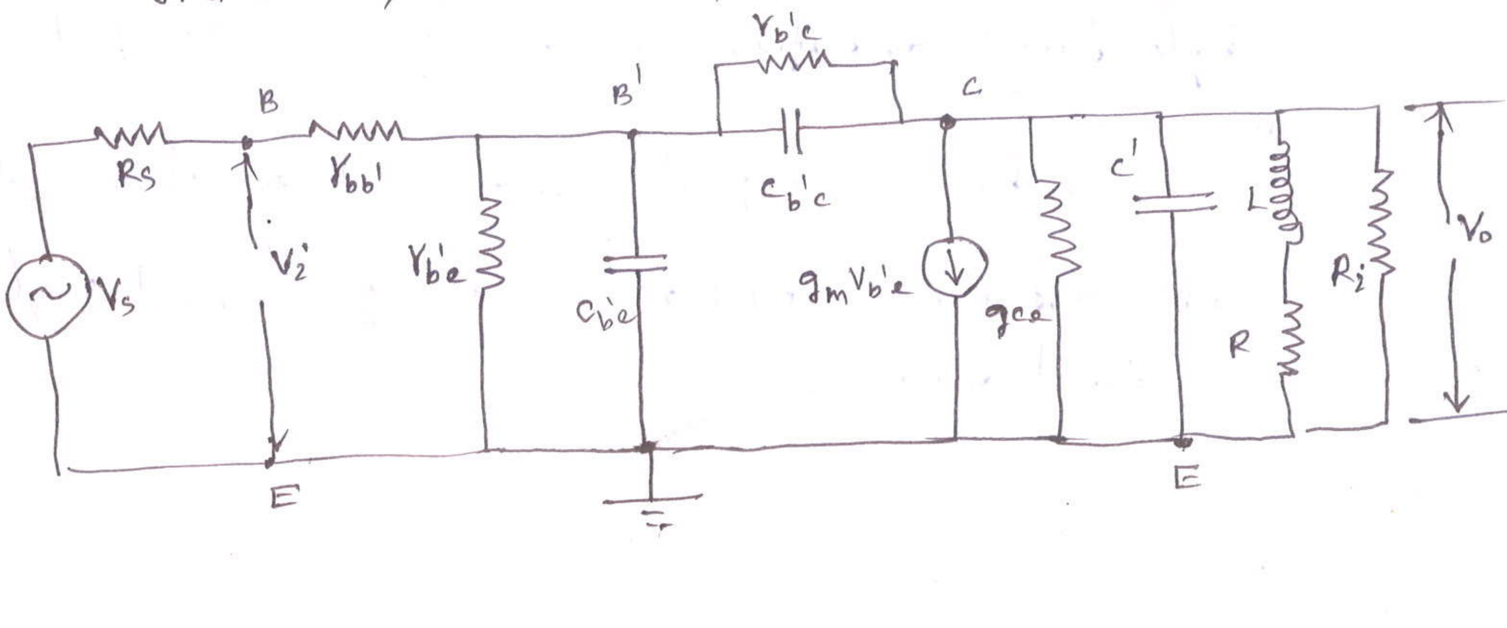
(a) Capacitance Coupled single tuned amplifier and
(b) Transformer coupled or inductively coupled single tuned amplifier.

Capacitance Coupled Single tuned amplifier

The following figure shows the circuit of a single tuned amplifier in which the output across the tuned circuit is coupled to the next stage through the coupling capacitor C_c. The tuned circuit formed by L and c' resonates at the frequency of operation.



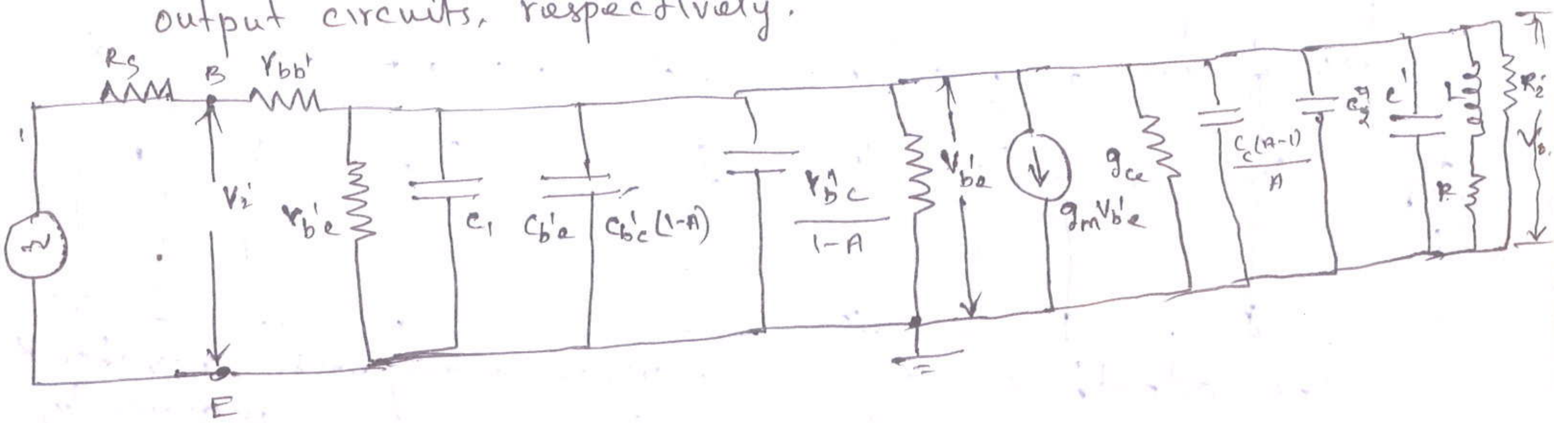
The following fig gives the equivalent ckt for the amplifier using high frequency hybrid π model for the transistor. In this circuit, R_i is the input resistance of the next stage.



Following fig. gives the modified equivalent circuit obtained by applying Miller's theorem. Here,

A – Voltage gain of the amplifier

C_1, C_2 are the stray wiring capacitances in the input and output circuits, respectively.



~~The equivalent circuit~~

i/p & o/p ckts, where all the inputs circuit can be grouped together to form C_s given by

$$C_s = C_{b'e} + C_1 + C_{b'c}(1-A).$$

All the capacitances in the output circuit can be grouped together to form C given by

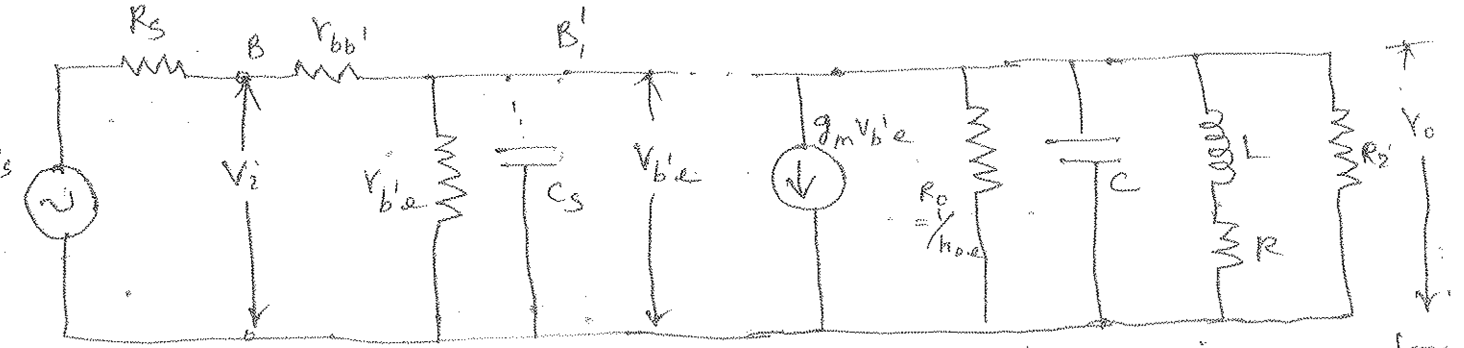
$$C = C_{b'c}\left(\frac{A-1}{A}\right) + C_2 + C'$$

Further, as simplified earlier,

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

R_o is the o/p resistance of current generator $g_m V_{b'e}$.

The reactances of the bypass capacitor C_E and the coupling capacitor C_c are negligibly small at the operating frequency and these elements can be neglected in the simplified ckt as shown below.



The admittance of the inductor along with resistor R is given by

$$Y_i = \frac{1}{R+j\omega L} = \frac{(R-j\omega L)}{(R+j\omega L)(R-j\omega L)}$$

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

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

$$= \frac{1}{R_p} + \frac{1}{j\omega L_p} \quad \text{——— (1)}$$

Where

$$R_p = \frac{R^2+\omega^2L^2}{R} \quad \& \quad L_p = \frac{R^2+\omega^2L^2}{\omega^2 L} \quad \text{——— (2)}$$

Thus the inductor branch may be represented by a resistor R_p and inductor L_p in shunt. Quality factor Q of the coil at resonance is give by.

$$\boxed{Q_0 = \frac{\omega_0 L}{R}}$$

Where $\omega_0 = \frac{1}{\sqrt{LC}}$ is the frequency of the resonance ckt.

Q_0 of the coil is usually large so that $\omega L >> R$ is in the frequency range of operation. As.

$$\frac{R}{\omega^2 L^2} << 1$$

From (2)

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

$$= \frac{\cancel{R}\left(R + \frac{\omega^2 L^2}{R}\right)}{\cancel{R}}$$

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

$$\boxed{R_p = \frac{\omega^2 L^2}{R}} \text{ ——— (3)}$$

From (2)

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

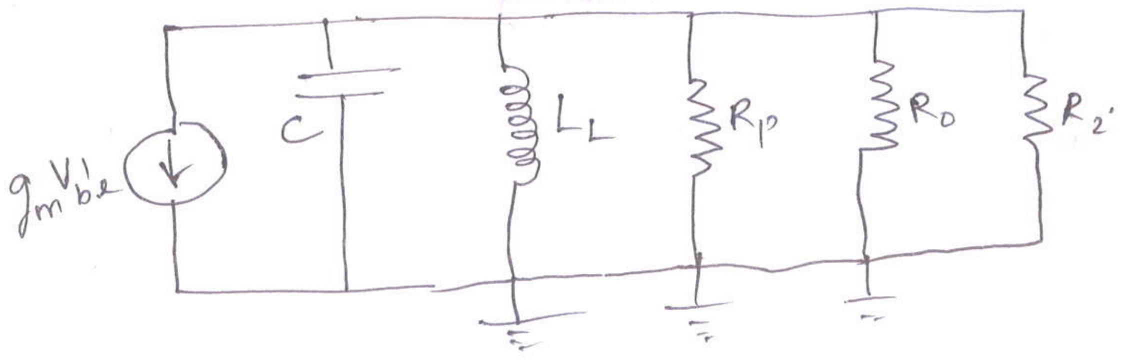
$\div$ N & D by $\omega^2 L$

$$= \frac{\left(\frac{R^2}{\omega^2 L} + \frac{\omega^2 L^2}{\omega^2 L}\right)\cancel{\omega^2 L}}{\cancel{\omega^2 L}}$$

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

$$L_p \simeq L \text{ ——— (4)}$$

The output circuit of the amplifier can be modified as shown below.



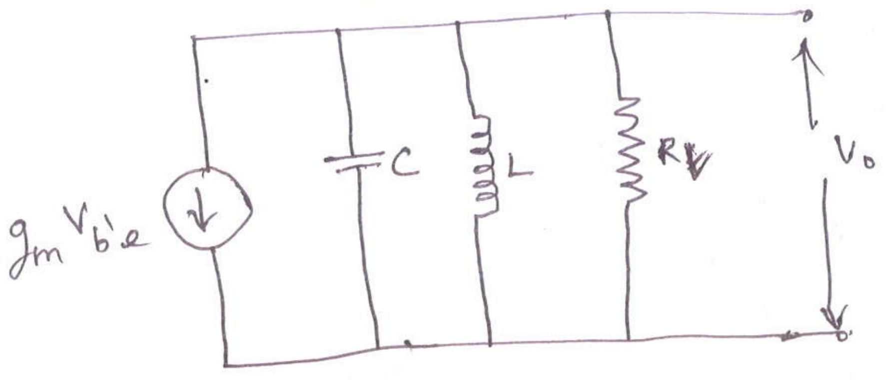
Taking R_t as the parallel combination of R_o, R_p and R_2'

$$\frac{1}{R_t} = \frac{1}{R_o} + \frac{1}{R_p} + 1/R_2'$$

The effective quality factor of the circuit magnification factor of the entire output circuit including R_p and R_i at resonant frequency ω_o is given by

$$Q_e = \frac{\text{Susceptance of inductance L (or) cap. C.}}{\text{Conductance of shunt Resistor } R_t.}$$

Simplified equivalent ckt.



$$Q_e = \frac{\frac{1}{X_L}}{1/R} = \frac{\frac{1}{\omega_0 L}}{1/R_t} = \frac{R_t}{\omega_0 L} \quad \text{——— (5)}$$

$$\boxed{\omega_0 C = \frac{1}{\omega_0 L}}$$

From the o/p circuit

$$V_0 = -g_m V_{b'e} Z \quad \text{——— (6)}$$

Where Z is the impedance of C, L and R_t in parallel. The admittance $Y \left(= \frac{1}{Z}\right)$ is given by

$$Y = \frac{1}{Z} = \frac{1}{R_t} + \frac{1}{j\omega L} + j\omega C$$

$$= \frac{1}{R_t}\left[1 + \frac{R_t}{j\omega L} + j\omega C . R_t\right]$$

X'ing num. & den. by ω_0.

$$Y = \frac{1}{R_t}\left[1 + \frac{R_t \omega_0}{j\omega L \omega_0} + \frac{j\omega C . R_t . \omega_0}{\omega_0}\right]$$

sub

$$\frac{R_t}{\omega_0 L} = \omega_0 C R_t = Q_e$$

$$Y = \frac{1}{R_t}\left[1 + \frac{\omega_0 . Q_e}{j\omega} + \frac{j\omega . Q_e}{\omega_0}\right]$$

$$Y = \frac{1 + j Q_e\left[\omega/\omega_0 - \frac{\omega_0}{\omega}\right]}{R_t}$$

$$Z = 1/Y = \frac{R_t}{1 + jQ_e\left[\omega/\omega_0 - \frac{\omega_0}{\omega}\right]} \quad \text{——— (7)}$$

Let δ indicate fractional frequency variation i.e. variation in frequency expressed as a fraction of the resonant frequency.

$$\delta = \frac{\omega - \omega_0}{\omega_0} = \frac{\omega}{\omega_0} - \frac{\omega_0}{\omega_0} = \frac{\omega}{\omega_0} - 1$$

$$\frac{\omega}{\omega_0} = 1 + \delta$$

$$Z = \frac{R_t}{1 + jQ_e\left[1 + \delta - \frac{1}{1+\delta}\right]}$$

$$= \frac{R_t}{1 + jQ_e\left[\frac{1 + \delta^2 + 2\delta - 1}{1+\delta}\right]}$$

$$= \frac{R_t(1+\delta)}{1 + jQ_e\left[1 + \delta^2 + 2\delta - 1\right]}$$

$$= \frac{R_t(1+\delta)}{1 + jQ_e\left[1 + 2\delta + \delta^2 - 1\right]}$$

Taking 2δ outside the denominator.

$$Z = \frac{R_t}{1 + j\,2Q_e\delta\left[\frac{\delta/2 + 1}{(1+\delta)}\right]}$$

At any frequency ω close to the frequency of resonance ω_0, $\delta \ll 1$

$$Z = \frac{R_t}{1 + j\,2Q_e\delta} \quad \text{—— (8)}$$

$$\frac{R_t(1+\delta)}{1 + jQ_e\left[\left(\frac{1}{2\delta} + 1 + \frac{\delta^2}{2\delta}\right)2\delta\right]}$$

$$\frac{R_t(1+\delta)}{1 + jQ_e\left[\left(1 + \frac{\delta}{2}\right)2\cdot\delta\right]}$$

At resonance, $\omega = \omega_0$ and $\delta = 0$ Hence the impedance Z becomes.

$$Z = R_t = R_o \parallel R_p \parallel R$$

Where

$$R_p = \frac{\omega_0^2 L^2}{R} = \frac{\omega_0 L}{\omega_0 C R} \quad \left(\text{since } \omega_0 L = \frac{1}{\omega_0 C}\right)$$

$$R_p = \frac{L}{C \cdot R}$$

From equ (3)

$$R_p = \frac{\omega_0^2 \cdot L^2}{R}$$

$$Q_0 = \frac{\omega_0 L}{R}$$

$$Q_0^2 = \frac{\omega_0^2 L^2}{R^2}.$$

$$R_p = Q_0^2 R = \omega_0 L Q_0.$$

Where Q_0 is the Q of the coil alone at resonance.

From i/p simplified equivalent ckt. neglecting C_s

$$V_{b'e} = V_i \frac{r_{b'e}}{r_{bb'} + r_{b'e}}.$$

$$V_o = -g_m V_{b'e} \cdot Z$$

$$= -g_m \left(V_i \frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) Z$$

Hence, voltage gain without considering the source resistance is given by

$$A = \frac{V_o}{V_i} = -g_m \left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) Z$$

sub.

$$A = -g_m \left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) \left(\frac{R_t}{1 + j2Q_e\delta} \right)$$

The voltage gain at resonance ($\delta = 0$) is given by

$$A_{res} = -g_m \left(\frac{r_{b'e}}{r_{bb'} + r_{b'e}} \right) \cdot R_t$$

Hence

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

$$\left| \frac{A}{A_{res}} \right| = \frac{1}{\sqrt{1 + (2\delta Q_e)^2}}$$

Phase angle of $\frac{A}{A_{res}}$ is given by

$$\phi = -\tan^{-1}(2\delta Q_e).$$

At a frequency ω_1 below the resonant frequency let $\delta = \frac{1}{2Q_e}$ then.

$$\left| \frac{A}{A_{res}} \right| = \frac{1}{\sqrt{2}} = 0.707$$

Thus, gain A is 3dB lower than A_{res}. This frequency ω_1 is the lower 3 dB frequency.

$|||^{ly}$ at a frequency ω_2 above ω_0 let

$$\delta = +\frac{1}{2Q_e}$$

So that $$\left|\frac{A}{A_{res}}\right| = 1/\sqrt{2} = 0.707.$$

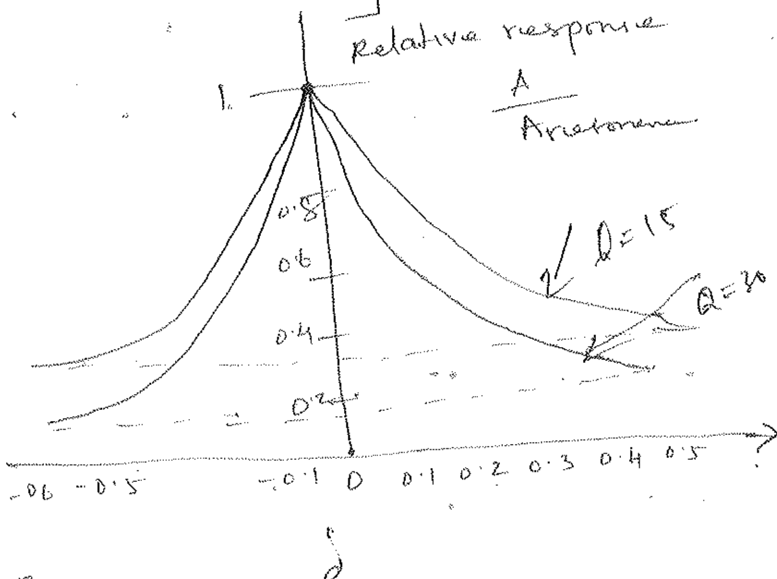
Hence this frequency ω_2 is the upper 3dB frequency.

The 3dB bandwidth $\Delta\omega$ (or B) $= \omega_2 - \omega_1$

$$= \frac{\left[(\omega_2 - \omega_0) + (\omega_0 - \omega_1)\right]\omega_0}{\omega_0}$$

$$\Delta\omega = \left[\frac{(\omega_2 - \omega_0)}{\omega_0} + \frac{(\omega_0 - \omega_1)}{\omega_0}\right].\omega_0 = [\delta + \delta]\omega_0$$

$$\Delta\omega = 2\delta\omega_0.$$



But $$\delta = \frac{1}{2Q_e}$$

$$2\delta = \frac{1}{Q_e}.$$

$$\therefore \Delta\omega = \frac{\omega_0}{Q_e}.$$

From (5) $$Q_e = \omega_0 C R_t = \frac{R_t}{\omega_0 L}$$

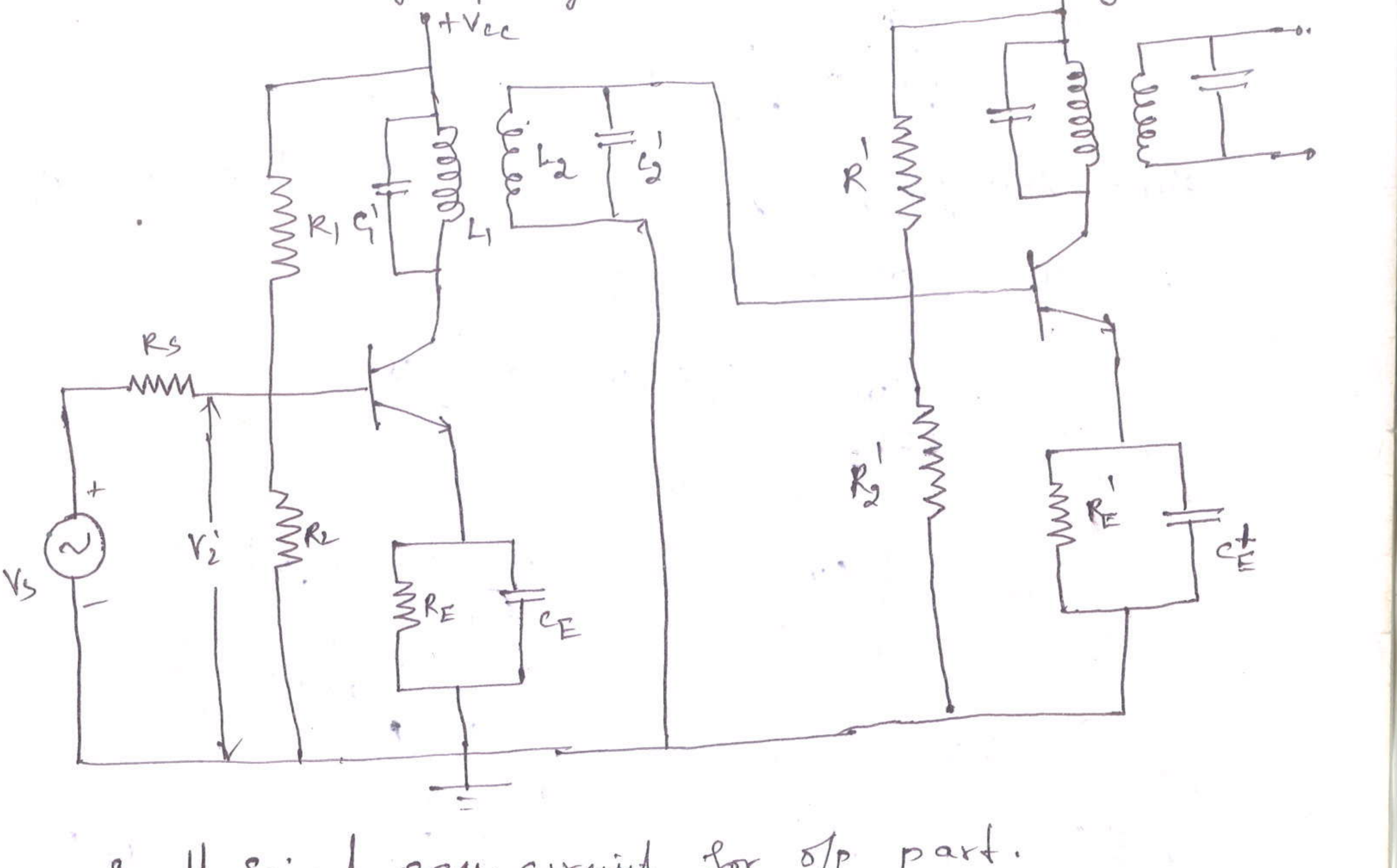
$$\Delta\omega = \frac{\omega_0}{R_t \cdot \omega_0 \cdot C} = \frac{1}{R_t \cdot C} \text{ rad/sec.}$$

$$\boxed{\Delta\omega = 1/R_t \cdot C} \text{ rad/s.}$$

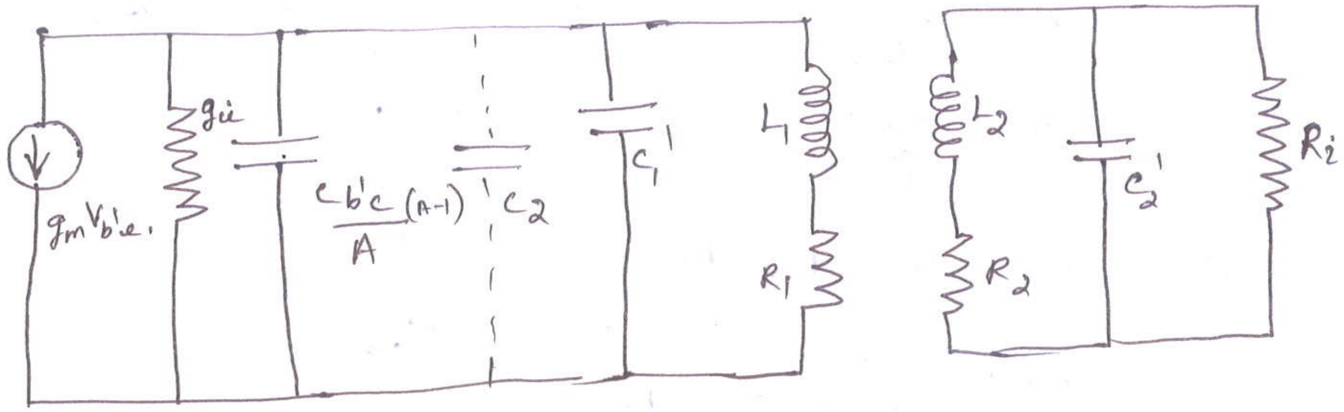
fig (D) shows the magnitude of relative gain $\frac{A}{A_{res}}$ plotted against δ.

Double tuned amplifier.

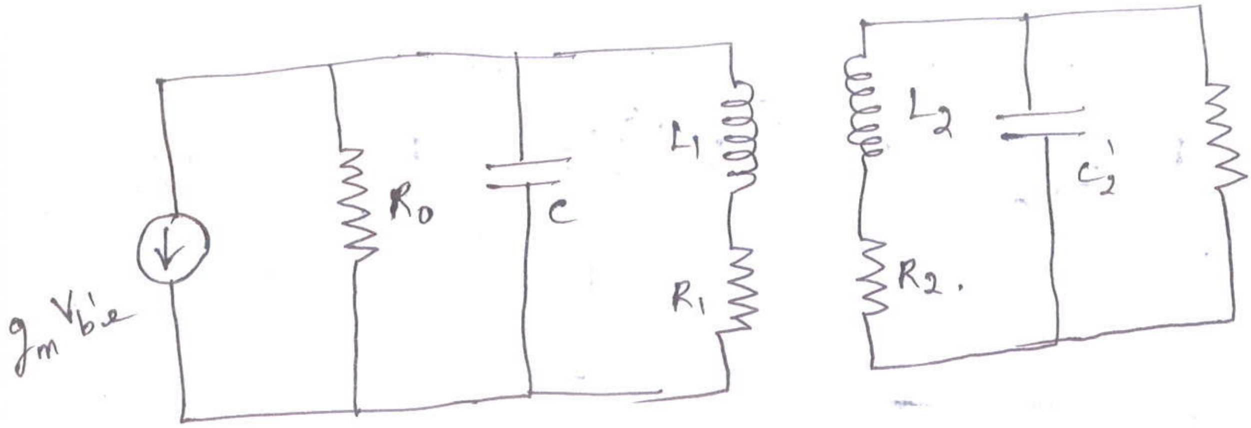
Voltage developed across the tuned circuit in the collector circuit is inductively coupled to another tuned circuit, both the tuned circuits being tuned to the same frequency, ie the frequency of the signal.



Small signal equ. circuit for o/p part.



Simplified equivalent circuit for output part.



The equivalent capacitance value.

$$C = \frac{C_{b'c}(A-1)}{A} + C_2 + C_1'$$

$$g_{ce} = \frac{1}{r_{ce}} \simeq h_{oe} = \frac{1}{R_o} \text{ (say)}.$$

The parallel connection of C_2' & R_i can be converted into series connection by using the formula,

$$C_s = C_p\left(1 + \frac{1}{Q_p^2}\right) \quad \& \quad R_s = \frac{R_p}{1+Q_p^2}.$$

Where $Q_p = \omega C_p R_p$

if $Q_p \geq 10$, $C_s \simeq C_p$ & $R_s \simeq \frac{R_p}{Q_p^2}$.

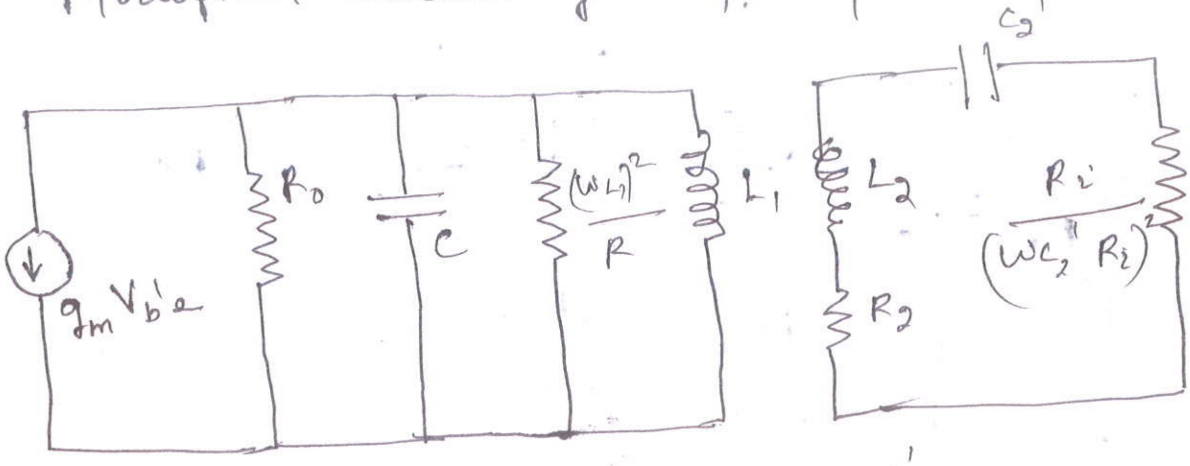
|||ly the series connection of L_1 & R_1 can be converted into parallel connection by using the formula

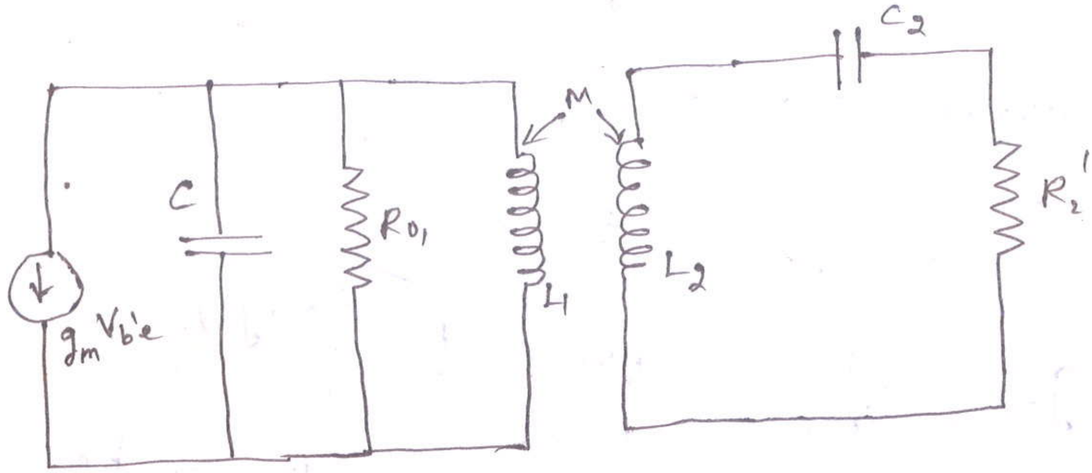
$$L_p = L_s\left(1 + 1/Q_s^2\right) \quad \& \quad R_p = R_s\left(1 + 1/Q_s^2\right)$$

Where $Q_s = \frac{\omega L_s}{R_s}$. if $Q_s \geq 10$, $L_p \simeq L_s$ &

$R_p \simeq R_s Q_s^2$.

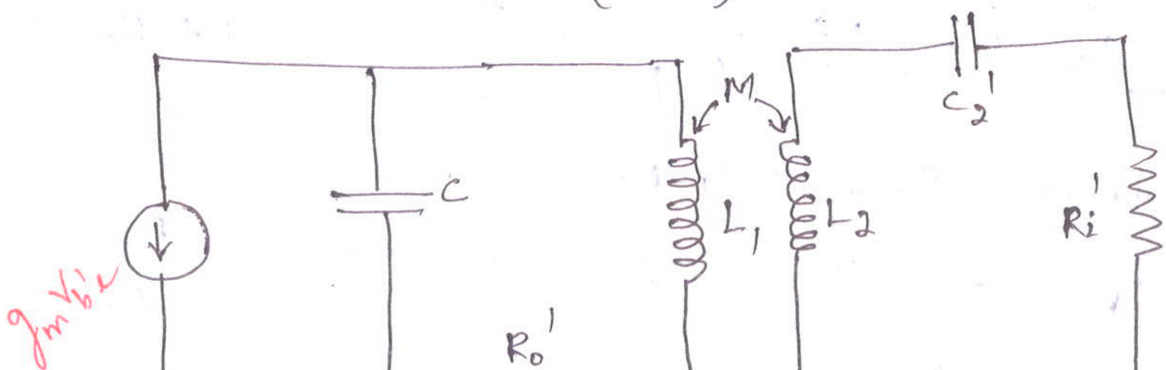
Modified circuit of output part.





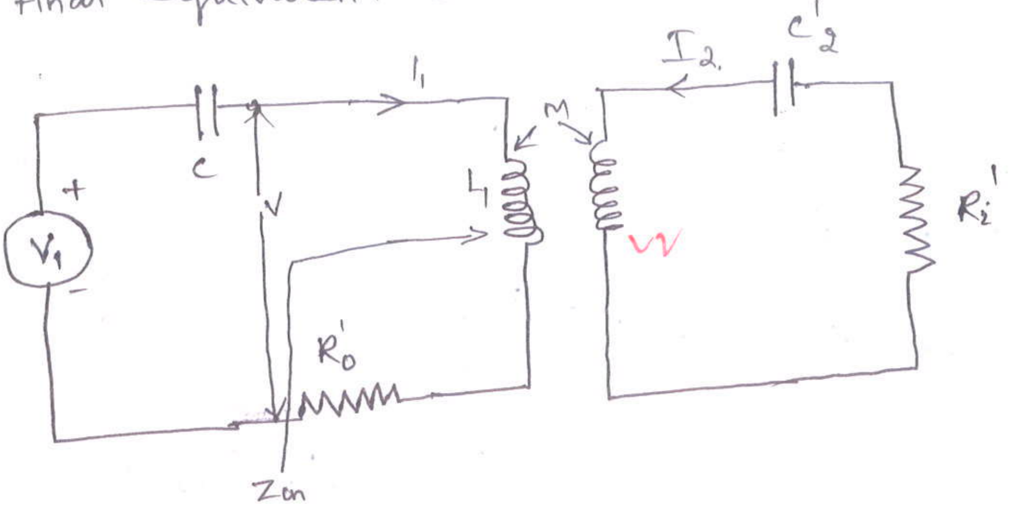
$$R_{o1} = R_o \,//\, \frac{(\omega L_1)^2}{R}$$

$$R_2' = R_2 + \frac{1}{(\omega C_2')^2 R_i}$$



$$R_o' = \frac{R_{o1}}{1 + \left(\omega L_1 / R_{o1}\right)^2}.$$

Final equivalent ckt.



$$V_1 = \frac{g_m V_{b'e}}{j\omega C}$$

To find Zin

Apply KVL to both the loops

To loop I_1

$$V = R_0' I_1 + j\omega L_1 I_1 + j\omega M I_2$$

$$V = (R_0' + j\omega L_1) I_1 + j\omega M I_2.$$

$$V = Z_{11} I_1 + Z_{12} I_2 \quad \text{———} (1).$$

$$\boxed{Z_{11} = R_0' + j\omega L_1} \; ; \; \boxed{Z_{12} = j\omega M}$$

To loop II_1

$$0 = j\omega M I_1 + R_i' I_2 + j\omega L_2 I_2 + \frac{1}{j\omega C_2'} \cdot I_2.$$

$$0 = j\omega M I_1 + \left(R_i' + j\omega L_2 - \frac{j}{\omega C_2'}\right) I_2.$$

$$0 = Z_{21} I_1 + Z_{22} I_2 \quad \text{———} (2)$$

$$\boxed{Z_{21} = j\omega M} \; ; \; \boxed{Z_{22} = R_i' + j\left(\omega L_2 - \frac{1}{\omega C_2'}\right)}.$$

From equ (2)

$$\boxed{I_2 = \frac{-Z_{21} I_1}{Z_{22}}} \quad \text{——— (3)}$$

Sub (3) in (1).

$$V = Z_{11} I_1 + Z_{12}\left(\frac{-Z_{21} I_1}{Z_{22}}\right)$$

$$= Z_{11} I_1 - \frac{Z_{12}^2 I_1}{Z_{22}} \qquad \left(\because \boxed{Z_{12} = Z_{21} = j\omega M}\right).$$

$$V = \left(\frac{Z_{11} Z_{22} - Z_{12}^2}{Z_{22}}\right) I_1$$

i/p impedance, $\boxed{Z_{in} = \frac{V}{I_1}}$

$$\therefore \boxed{Z_{in} = Z_{11} - \frac{Z_{12}^2}{Z_{22}}} \quad \text{——— (4)}$$

Substitute the values of Z_{11}, Z_{12} & Z_{22} in (4).

$$Z_{in} = R_0' + j\omega L_1 + \frac{\omega^2 M^2}{R_i' + j\left(\omega L_2 - \frac{1}{\omega C_2'}\right)} \quad \text{——— (5)}$$

At resonance, $\omega = \omega_0$, $\omega_0 L_2 = \frac{1}{\omega_0 C_2'}$

$$(5) \Longrightarrow R_0' + j\omega_0 L_1 + \frac{\omega_0^2 M^2}{R_i' + j\left(\omega_0 L_2 - \frac{1}{\omega_0 C_2'}\right)}.$$

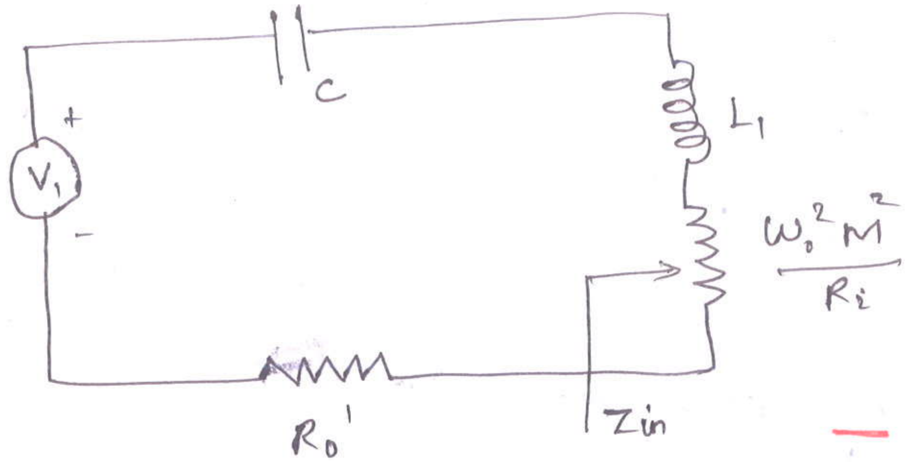
$$Z_{in} = R_0' + j\omega_0 L_1 + \frac{\omega_0^2 M^2}{R_i'}$$

When M is reasonably large.

$R_0' << \frac{\omega_0^2 M^2}{R_2'}$ of R_0' is neglected.

$\therefore \; Z_{in} \approx j\omega_0 L_1 + \frac{\omega_0^2 M^2}{R_2'}$ ——— (6)

Simplified equivalent ckt



For maximum transfer of power

$$R_0' = \frac{\omega_0^2 M^2}{R_2'} \quad \text{———} \quad (7)$$

When M is adjusted to the

At critical value of M_c. case. $M = M_c$

$$R_0' = \frac{\omega_0^2 M_c^2}{R_2'}$$

$R_0' R_2' = \omega_0^2 M_c^2 \quad ; \quad \sqrt{R_0' R_2'} = \omega_0 M_c.$

Where $M_c = K_c \sqrt{L_1 L_2}$

$K_c \rightarrow$ Coefficient of coupling

$$\sqrt{R_0' R_2'} = \omega_0 K_c \sqrt{L_1 L_2}.$$

WKT $Q = \frac{X_L}{R}$

$$k_c = \frac{\sqrt{R_0' R_2'}}{\omega_0 \sqrt{L_1 L_2}} = \frac{\sqrt{R_0'} \cdot \sqrt{R_2'}}{\sqrt{\omega_0} \cdot \sqrt{\omega_0} \cdot \sqrt{L_1} \sqrt{L_2}}.$$

$$k_c = \left(\frac{R_0'}{\omega_0 L_1}\right)^{1/2} \cdot \left(\frac{R_2'}{\omega_0 L_2}\right)^{1/2}.$$

k_c is the critical value of the coeff. of coupling corresp to the critical value M_c of mutual inductance.

$$k_c = \frac{1}{\sqrt{Q_{01} \cdot Q_{02}}} \quad \text{——— (8)}$$

At non-critical case $M \neq M_c$

By applying KVL at both loops.

To loop I_1

$$V_1 = \left(R_0' + j\omega L_1 + \frac{1}{j\omega C}\right) I_1 + j\omega M I_2.$$

$$V_1 = Z_{11} I_1 + Z_{12} I_2 \quad \text{——— (9)}$$

To loop I_2.

$$0 = j\omega M I_1 + \left(R_2' + j\omega L_2 + \frac{1}{j\omega C_2'}\right) I_2$$

$$0 = Z_{21} I_1 + Z_{22} I_2 \quad \text{——— (10)}$$

$$Z_{11} = R_0' + j\left(\omega L_1 - \frac{1}{\omega C}\right) ; \quad Z_{22} = R_2' + j\left(\omega L_2 - \frac{1}{\omega C_2'}\right).$$

$$Z_{12} = Z_{21} = j\omega M$$

At resonance $\omega = \omega_0$, $\omega_0 L_1 = \frac{1}{\omega_0 C}$ &

$\omega_0 L_2 = \frac{1}{\omega_0 C_2'}$

$M = M_c$.

Hence $Z_{11} = R_0'$; $Z_{22} = R_2'$ & $Z_{12} = Z_{21} = j\omega_0 M_c$

From eq (10) $I_1 = -\dfrac{Z_{22} I_2}{Z_{21}}$

Sub I_1 in (9)

$$V_1 = \frac{-Z_{11} Z_{22} I_2}{Z_{21}} + Z_{12} I_2$$

$$Z_{21} V_1 = -Z_{11} Z_{22} I_2 + Z_{12}^2 I_2 \qquad (Z_{12} = Z_{21})$$

$$-V_1 Z_{21} = (Z_{11} Z_{22} - Z_{12}^2) I_2.$$

$$I_2 = \frac{-V_1 Z_{21}}{Z_{11} Z_{22} - Z_{12}^2} \qquad \text{—— (11)}$$

Sub Z_{11}, Z_{22}, Z_{12}, & Z_{21} in (11).

$$I_2 max = -\frac{V_1 (j\omega_0 M_c)}{R_0' R_2' + \omega_0^2 M_c^2}$$

$$I_2 max = \frac{-j V_1 \omega_0 M_c}{R_0' R_2' + \omega_0^2 M_c^2}$$

sub $M_c = \dfrac{\sqrt{R_0' R_2'}}{\omega_0}$

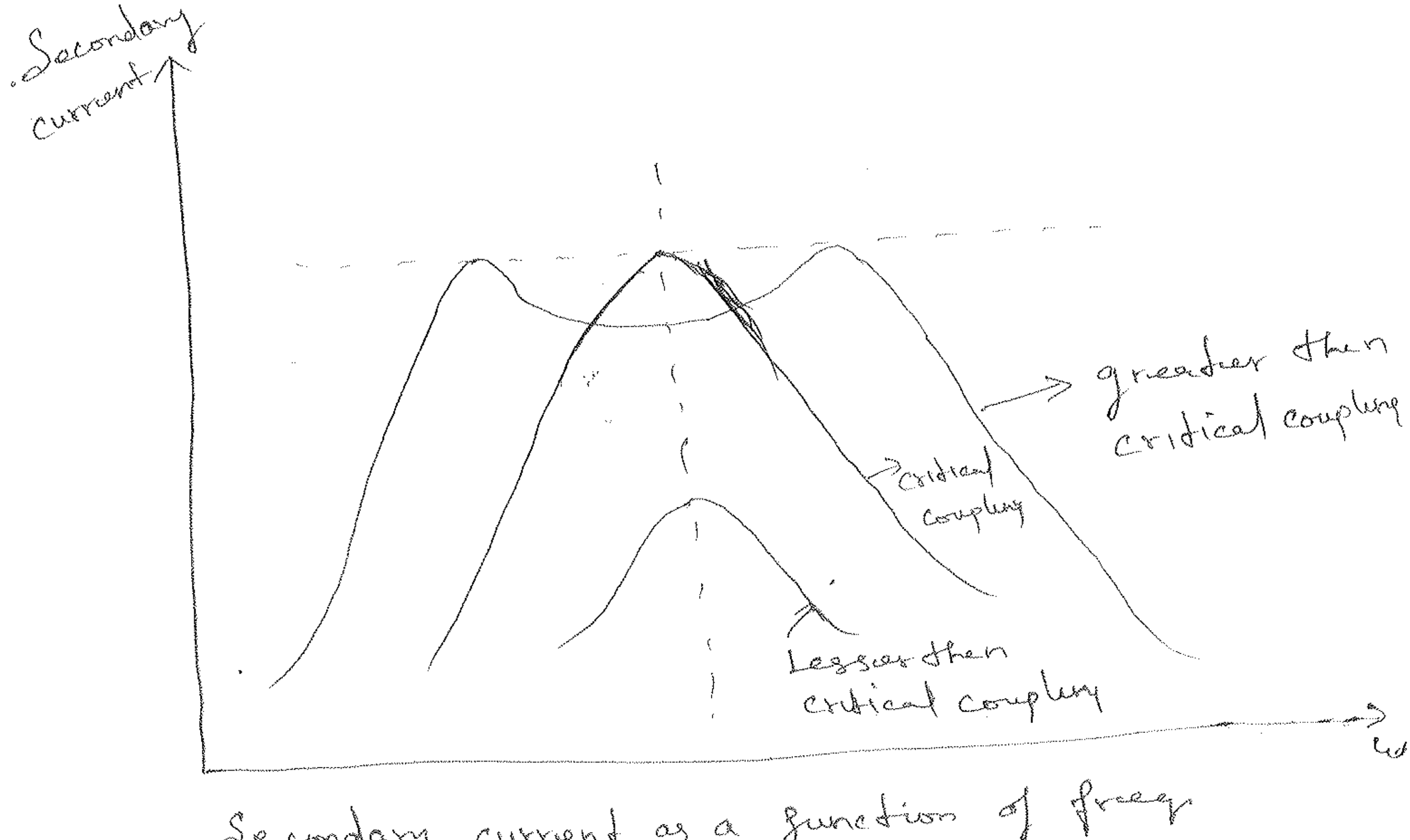
$$\omega_0 M_c = \sqrt{R_0' R_2'}$$

$$I_{2\,max} = \frac{-jV_1\omega_0 \cdot \sqrt{\frac{R_0' R_2'}{\omega_0}}}{R_0' R_2' + \omega_0^2 \frac{(R_0' R_2')}{\omega_0^2}}$$

$$= \frac{-jV_1 \sqrt{R_0' R_2'}}{R_0' R_2' (2)}$$

$$I_{2\,max} = \frac{-jV_1}{2\sqrt{R_0' R_2'}} \qquad (12)$$

$$|I_{2\,max}| = \frac{V_1}{2\sqrt{R_0' \cdot R_2'}}$$



Secondary current as a function of freq

For coupling greater than the critical coupling, power transfer is maximum at two other frequencies. These two frequencies can be obtained by equating $|I_2\,max|$ to $|I_2|$.

Thus $|I_{2max}| = |I_2|$

$$\left| \frac{-jV_1}{2\sqrt{R_0' R_2'}} \right| = \left| \frac{-V_1 Z_{21}}{Z_{11}Z_{22} - Z_{12}^2} \right|$$

Sub Z_{11}, Z_{22} & Z_{12}, Z_{21} values we get.

$$\left| \frac{-jV_1}{2\sqrt{R_0' R_2'}} \right| = \left| \frac{1}{\left(R_0' + j\omega L_1 + \frac{1}{j\omega C}\right)\left(R_2' + j\omega L_2 + \frac{1}{j\omega C_2'}\right) - (j\omega M)^2} \right| \quad \text{—— (13)}$$

The condition for maximum power transfer.

$$R_0' = \frac{\omega_0^2 M_c^2}{R_2'}$$

$$M = bM_c \quad ; \quad M_c = M/b$$

$$R_0' = \frac{\omega_0^2 M^2}{b^2 R_2'}$$

$$R_0' R_2' b = \omega_0^2 M^2 \quad ; \quad \omega_0 M = b\sqrt{R_0' R_2'} \quad \text{—— (14)}$$

sub (14) in (13)

$$\left|\frac{-jV_1}{2\sqrt{R_0' R_2'}}\right| = \left|\frac{-jV_1 b\sqrt{R_0' R_2'}}{\left(R_0'+j\omega L_1+\frac{1}{j\omega C}\right)\left(R_2'+j\omega L_2+\frac{1}{j\omega C_2'}\right)+b^2 R_0' R_2'}\right| \quad \text{(14a)}$$

Let $L_2 = L_1 = L$, $C_2' = C$, $R_0' = R_2' = R$ &

$$j\left(\omega L - \frac{1}{\omega C}\right) = jX$$

$$\therefore \left|\frac{-jV_1}{2R}\right| = \left|\frac{-jV_1\, bR}{(R+jX)^2 + b^2R^2}\right|.$$

$$\left|R^2 - X^2 + j2RX + R^2b^2\right| = \left|2bR^2\right|$$

$$2bR^2 = \sqrt{\left(R^2(1+b^2) - X^2\right)^2 + (2RX)^2}$$

Squaring both sides,

$$4b^2R^4 = \left(R^2(1+b^2) - X^2\right)^2 + 4R^2X^2$$

Solving for x, we get.

$$X = \pm\sqrt{b^2-1}\cdot R$$

$$\omega L - \frac{1}{\omega C} = \pm\left[\sqrt{b^2-1}\right]R$$

$$\omega^2 LC - 1 = \pm\left[\sqrt{b^2-1}\right]\omega CR \quad \text{(15)}$$

$$\frac{1}{Q} = \omega C R \simeq \omega_0 C R \quad \text{------} (16)$$

Also $\omega_0^2 = \frac{1}{LC}$ ------ (17)

Sub (16) & (17) in (15)

$$\frac{\omega^2}{\omega_0^2} - 1 = \frac{\pm\sqrt{b^2-1}}{Q}$$

$$\frac{\omega^2}{\omega_0^2} = 1 \pm \frac{\sqrt{b^2-1}}{Q}$$

$$\boxed{\omega = \pm\,\omega_0 \sqrt{1 \pm \frac{\sqrt{b^2-1}}{Q}}} \quad \text{------} (18)$$

if $b < 1$, ie, coefficient of coupling is less than critical value, then ω becomes complex. Thus for $b<1$ & $K < K_C$ there is no real frequency at which maximum power transfer can take place.

However for $b > 1$, there result two freq. at which max. power transfer take place

$$\left| \frac{-jV_1 b\sqrt{R_0' R_2'}}{\left(R_0' + j\omega L_1 + \frac{1}{j\omega C}\right)\left(R_2' + j\omega L_2 + \frac{1}{j\omega C_2'}\right) + b^2 R_0' R_1'} \right| = \frac{1}{\sqrt{2}} \left| \frac{-jV_1}{2\sqrt{R_0' R_1'}} \right| \quad \text{---} (19)$$

Let $L_2 = L_1 = L$, $C_2' = C$, $R_0' = R_2' = R$ &

$$j\left(\omega L - \frac{1}{\omega C}\right) = jX$$

Sub. these values in (19)

The 3 db frequency is frequency at which I_2 reduces to 0.707 of its maximum value are obtained by equating $|I_2|$ to $\frac{1}{\sqrt{2}}$ times $|I_{2max}|$

$$\left| \frac{V_1 b R}{(R+jX)^2 + b^2 R^2} \right| = \frac{1}{\sqrt{2}} \left| \frac{V_1}{2R} \right|$$

$$|2\sqrt{2}\, b R^2| = |-X^2 + 2jXR + R^2 + b^2 R^2|$$

$$2\sqrt{2}\, b R^2 = \sqrt{(R^2(1+b^2) - x^2)^2 + 4X^2R^2}.$$

Squaring both sides.

$$8b^2R^4 = (R^2(1+b^2) - X^2)^2 + 4X^2R^2$$

solving for X

$$X = \pm\sqrt{(b^2 - 1 \pm 2b)}\, R$$

Sub $X = WL - \frac{1}{Wc}$.

$$WL - \frac{1}{Wc} = \pm\sqrt{(b^2 - 1 \pm 2b)} \cdot (R).$$

Multiplying by c

$$WLC - 1/W = \pm\sqrt{(b^2 - 1 \pm 2b)} \cdot RC$$

Multiplying by W_0

$$WLCW_0 - \frac{W_0}{W} = \pm\sqrt{(b^2 - 1 \pm 2b} \cdot W_0 RC$$

Sub $W_0 RC = 1/2$ & $LC = 1/W_0^2$

~~wt~~ .

$$\frac{\omega\,\omega_0}{\omega_0^2} - \frac{\omega_0}{\omega} = \pm\frac{\sqrt{(b^2-1\pm 2b)}}{Q}.$$

$$\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega} = \pm\frac{\sqrt{(b^2-1\pm 2b)}}{Q} \qquad \text{--- (20)}$$

Equ (20) gives two 3 dB frequencies ω_2 & ω_1, one corresponding to the +ve sign and the other corresponding to the −ve sign

$$\left(\frac{\omega_2}{\omega_0} - \frac{\omega_0}{\omega_2}\right) = -\left(\frac{\omega_1}{\omega_0} - \frac{\omega_0}{\omega_1}\right)$$

$$\omega_0 = \sqrt{\omega_1\omega_2}.$$

Taking +ve sign in right hand side of equ (20.)

$$\frac{\sqrt{(b^2-1)\pm 2b}}{Q} = \left(\frac{\omega_2}{\omega_0} - \frac{\omega_0}{\omega_2}\right) = \frac{\omega_2^2-\omega_0^2}{\omega_0\omega_2}.$$

$$= \frac{\omega_2^2 - \omega_1\omega_2}{\omega_0\omega_2} = \frac{\omega_2-\omega_1}{\omega_0}$$

The 3 dB bandwidth $\Delta\omega$ is given by

$$\boxed{\Delta\omega = \omega_2-\omega_1 = \frac{\omega_0}{Q}\sqrt{(b^2-1)\pm 2b}} \qquad \text{--- (21).}$$

Thus the 3dB bandwidth $\omega_2-\omega_1$ is proportional to ω_0/Q.

Effect of cascading single tuned amplifiers on Bandwidth.

In order to obtain a high overall gain, several identical stages or tuned amplifiers can be used in cascade. The overall voltage gain is the product of the voltage gains of the individual stages. At the same time, the high voltage gain is accompanied by a narrower bandwidth than for a single stage.

Consider 'n' stages of single tuned direct coupled amplifiers connected in cascade. The aim here is to determine the overall gain and bandwidth of such an amplifier.

The relative gain of single tuned amplifier with respect to gain at resonant frequency f_o is given as

$$\left| \frac{A}{A_{res}} \right| = \frac{1}{\sqrt{1 + (2 \delta Q_e)^2}}$$

Now, the gain of n stage cascaded amplifier becomes.

$$\left| \frac{A}{A_{res}} \right|^n = \left[\frac{1}{\sqrt{1 + (2 \delta Q_e)^2}} \right]^n$$

$$= \left[\frac{1}{1 + (2 \delta Q_e)^2} \right]^{n/2}$$

The 3dB frequencies for the n stage cascaded amplifier can be found by equating $\left|\frac{A}{A_{res}}\right|^n$ to $\frac{1}{\sqrt{2}}$.

$$\left|\frac{A}{A_{res}}\right|^n = \frac{1}{\left(\sqrt{1+(2\delta Q_e)}\right)^n} = \frac{1}{\sqrt{2}}$$

$$\left[\sqrt{1+(2\delta Q_e)^2}\right]^n = \sqrt{2}$$

ie $$1+(2\delta Q_e)^2 = 2^{1/n}$$

$$2\delta Q_e = \pm\sqrt{2^{1/n}-1}$$

substituting for δ the fractional frequency variation

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

$$2\left(\frac{f-f_0}{f_0}\right)\cdot Q_e = \pm\sqrt{2^{1/n}-1}$$

$$2(f-f_0)Q_e = \pm f_0\cdot\sqrt{2^{1/n}-1}$$

Now $$f_2-f_0 = +\frac{f_0}{2Q_e}\cdot\sqrt{2^{1/n}-1}$$

III^{ly} $$f_0-f_1 = \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1}$$

Thus the bandwidth of the n-stage identical amplifier is

$$B_{2n} = f_2-f_1 = (f_2-f_0)+(f_0-f_1)$$

$$= \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1} + \frac{f_0}{2Q_e}\sqrt{2^{1/n}-1}$$

$$= \frac{f_0}{\cancel{2}Q_e}\times\cancel{2}\sqrt{2^{1/n}-1}$$

$$= \frac{f_0}{Q_e} \sqrt{2^{1/n} - 1}$$

$$= B_1 \sqrt{2^{1/n} - 1}$$

Where B_{In} is the bandwidth of ~~single~~ n stages ~~bandwidth of n stages~~ of cascade amplifier and B_1 is the bandwidth for single stage.

Bandwidth of n stages B_{In} is equal to B_I multiplied by a factor of $\sqrt{2^{1/n} - 1}$

When $n = 2$; $\sqrt{2^{1/n} - 1} = 0.643$.

$n = 3$; $\sqrt{2^{1/n} - 1} = 0.510$.

Thus, the BW is reduced to 64.3% for two stages and 51% ~~for~~ 3 stages of cascade amplifier.

In order to maintain a prescribed 3dB BW, Q of the tuned circuit should be reduced.

Effect of cascading double tuned amplifiers on band width.

The 3 dB bandwidth of the cascaded double tuned amplifier is given by

$$B_{2n} = B_2 \left[2^{1/n} - 1 \right]^{1/4}$$

Where B_2 is the 3 dB bandwidth of the single stage double tuned amplifier and n is the number of identical stages connected in cascade.

Stagger tuned amplifiers:

In order to increase bandwidth, double tuned amplifiers are preferred, but alignment of double tuned amplifiers is difficult. In stagger tuned circuits, two single tuned cascade amplifiers having a certain bandwidth are taken. The resonant frequencies of the two tuned circuits 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 stagger tuned circuits.

The resultant staggered pair will have a Bandwidth ie, $\sqrt{2}$ times that of each of the individual single tuned circuits.

The relative gain of single tuned direct coupled amplifier is given by

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

let $\frac{A}{A_{res}} = \frac{1}{1+jX}$, Where $X = 2\delta Q_e$.

As $B = \frac{f_0}{Q_e}$, under 3dB condition $\delta = \frac{1}{2Q_e}$.

The equation for bandwidth can be written as

$$B = 2\delta f_0$$

Since one stage is tuned to a frequency δf_0 below

f_0 and the other stage is tuned to a frequency δf_0 above f_0, the corresponding selectivity function of the circuits are.

$$\left(\frac{A}{A_{res}}\right)_1 = \frac{1}{1+j(x-1)} \quad \text{and} \quad \left(\frac{A}{A_{res}}\right)_2 = \frac{1}{1+j(x+1)}$$

By multiplying relative gains of the two amplifiers the overall gain function becomes

$$\left(\frac{A}{A_{res}}\right)_{pair} = \left(\frac{A}{A_{res}}\right)_1 \cdot \left(\frac{A}{A_{res}}\right)_2$$

$$= \frac{1}{2 - x^2 + 2jx}$$

The magnitude of the resulting function is

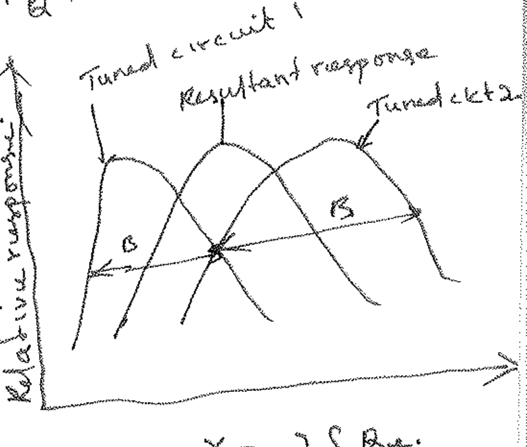
$$\left|\left(\frac{A}{A_{res}}\right)_1 \cdot \left(\frac{A}{A_{res}}\right)_2\right| = \frac{1}{\sqrt{(2-x^2)^2 + (2x)^2}} = \frac{1}{\sqrt{4 + x^4}}$$

$$= \frac{1}{\sqrt{4 + (2\delta_0 Q)^4}} = \frac{1}{2}\,\frac{1}{\sqrt{1 + 4\delta_0^4 Q^4}}$$

$$\left|\left(\frac{A}{A_{res}}\right)_1 \left(\frac{A}{A_{res}}\right)_2\right| = \frac{1}{2} \cdot \frac{1}{\sqrt{1+4\delta_0^4 Q^4}}.$$

Where δ_0 = value of δ at new freq. ω_0

Q = value of Q_e for each circuit refferred to ω_0



Large signal tuned amplifiers.

The large signal tuned amplifiers amplify high power signals of the radio frequency range using class c, operation rather than class-A as in the small signal tuned amplifiers.

class-c tuned amplifiers

The amplifier is said to be a class-c amplifier if the output signal is obtained for less than half a cycle for a full input cycle. The Q-point and the input signal are selected so as to achieve the conditions for a class-c amplifier. However, it should be noted that the fundamental frequency component for the class-c amplifier output is not linearly related to the fundamental components of the input signal. A typical bipolar transistor class-c-amplifier ckt and its chers' waveforms are shown below.

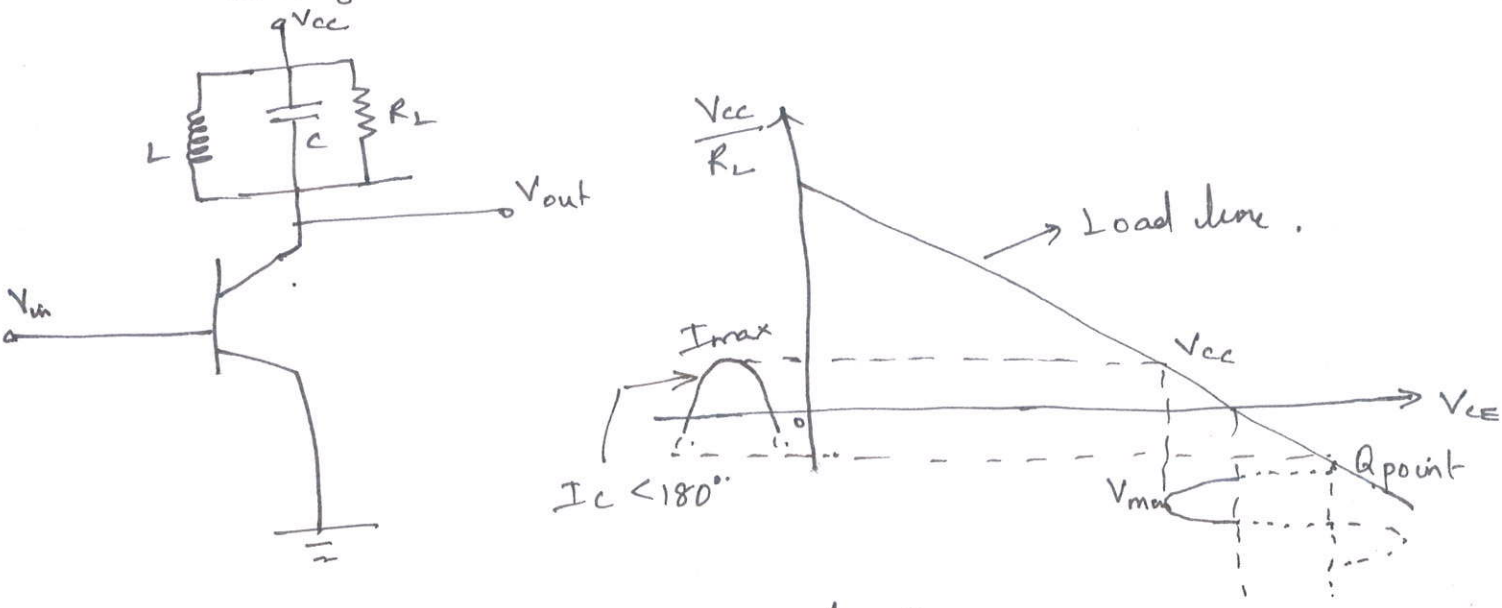


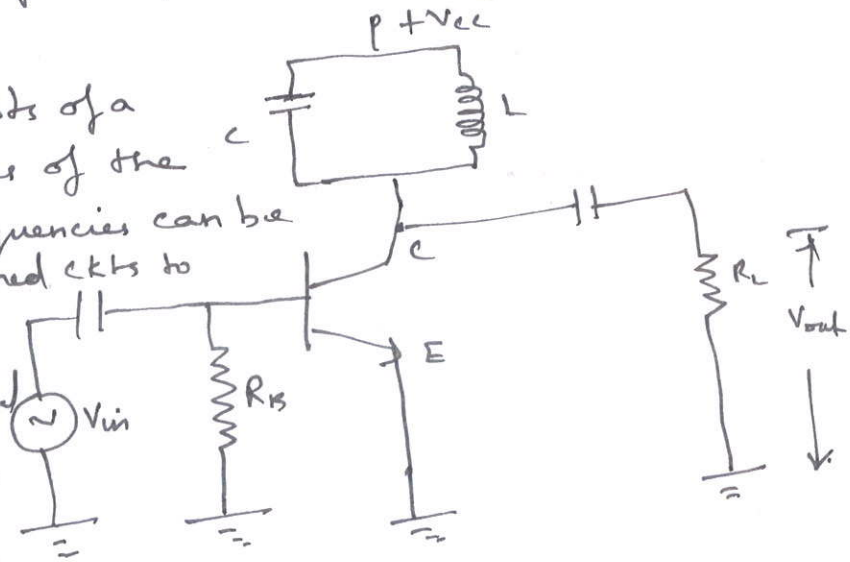
fig ①

The selection of Q point is made in such a way that the transistor remains active, for less than half a cycle so that only that much part of the input waveform is reproduced at the output with amplification. For the remaining part of the input, the transistor remains inactive and no corresponding o/p signals are produced. However the output pulses from the transistor triggers the tank circuit to produce continuous sinusoidal waveform.

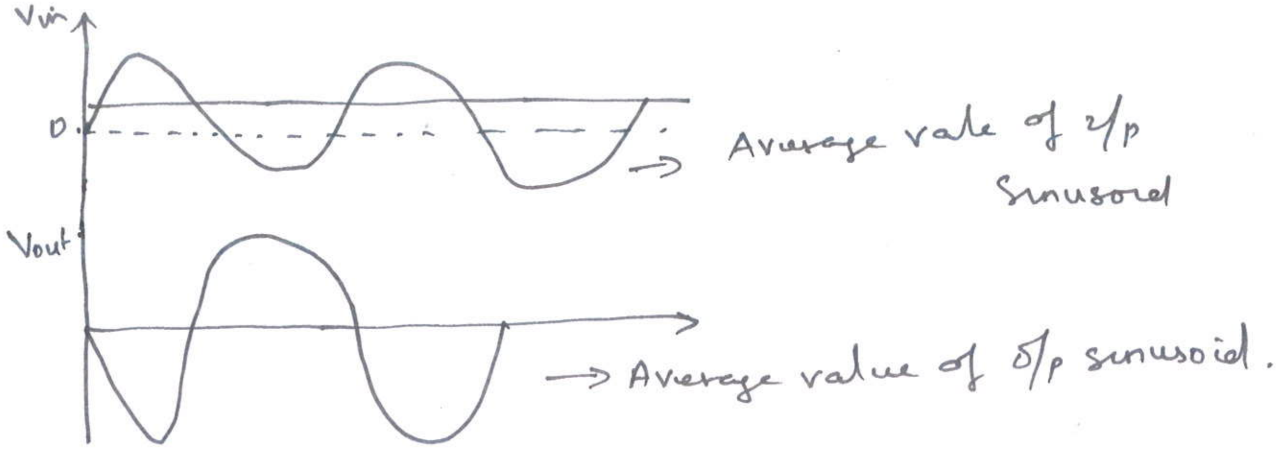
Conduction angle (θ_c)

The conduction angle θ_c is defined as the total angle during which the collector current exists at the output. From fig ① it is evident that θ_c is always less than 180°. In continuation of the basic class-c amplifier operation, large signal tuned amplifier having class c-amplifier as its load is shown below.

The o/p of class c amp. consists of a series of pulses with harmonics of the input signal. The harmonic frequencies can be filtered out using parallel tuned ckts to produce a sine wave o/p consisting of only the fundamental component of the input signal.



However, the analysis part remains identical to that of small signal tuned amp except for the o/p waveform, i.e θ_c is 360° in the small signal amp. where class A is generally used, and θ_c is less than 180° in the large signal tuned amplifier as shown below.



Efficiency of class-c Tuned amplifier.

The amplitude of the fundamental component of a class-c waveform depends on the conduction angle 'Q_c'. The greater the conduction angle, the greater the ratio of the amplitude of the fundamental component to the amplitude of the total waveform that includes harmonics.

Let

r_1 — peak value of fundamental component to the peak value of class c waveform.

The value of r_1 is closely approximated by

$$r_1 = \left(-3.54 + 4.1 Q_c - 0.0072 Q_c^2\right) \times 10^{-3}$$

Where $0^\circ \le Q_c \le 180^\circ$

The values r_1 varies from 0 to 0.5 as Q_c varies from 0° to 180°.

r_0 – peak value of the ratio of class c wave form to its peak value.

r_0 = d.c value / peak value θ_c/π (180).

Here the values 'r_0' varies from 0 to $1/\pi$ as θ_c varies from 0° to 180°.

The o/p power at the fundamental frequency under maximum drive condition is

$$P_0 = (r_1 I_p)\, V_{cc}/2.$$

I_p – peak o/p (collector) current.

Average supplied power (P_s) is

$$P_s = (r_0 I_p) V_{cc}$$

The efficiency is

$$\eta = P_0/P_s = (r_1 \cancel{I_p}) \cancel{V_{cc}} / r_0 \cancel{I_p} \cancel{V_{cc}}$$

$$\boxed{\eta = \frac{r_1}{2 r_0}}$$

Application of class c Tuned amplifier.

Stability of Tuned amplifier.

A parallel tuned circuit will be inductive at frequencies below resonant frequencies. ie when it is used 'L' is significant and 'C' is neglected.

In tuned amplifier there is a frequency at which tuned input and tuned output circuits remain inductive at which input tuned circuit is represented by a net inductance L_i in parallel with R_B (Biasing resistor) and output tuned circuit is represented as L_0 in parallel with R_c as shown in figure below. (fig ⊛)

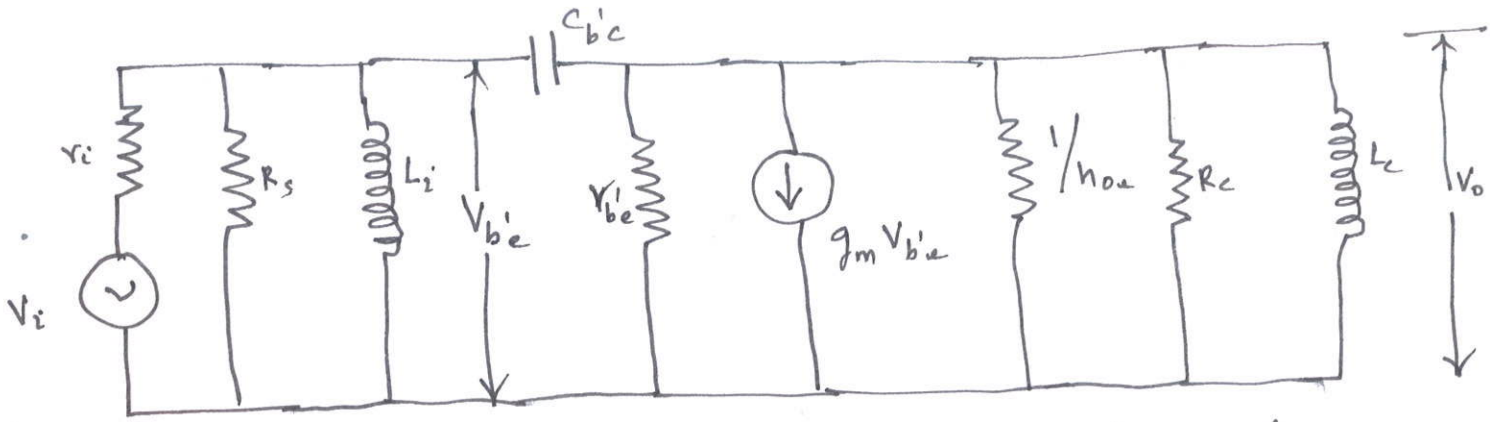


fig ⊛ Hybrid π equivalent circuit of Tuned amplifier

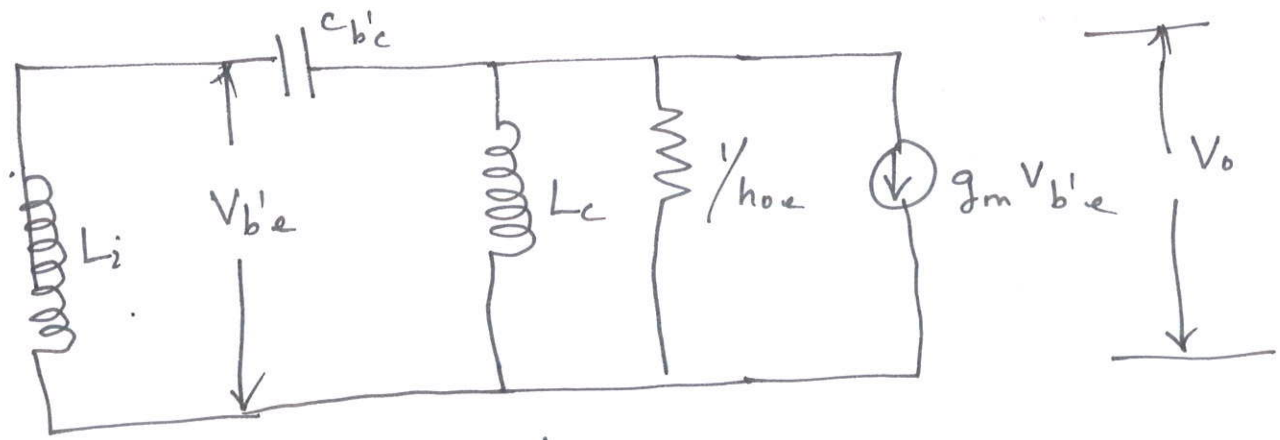


fig ⊙ : Simplified ckt.

If R_c and R_B are very large compared to L_i and L_o then R_c and R_B and the junction capacitance $C_{b'c}$ is effective at this frequency neglected hence the circuit is reduced as shown in fig ©.

The loop consisting of L_i, L_o and $C_{b'c}$ forms a resonant circuit for which the resonant frequency is

$$\omega_o = \frac{1}{\sqrt{(L_i + L_o) C_{b'c}}}$$

At resonant frequency, the loop of the system develops self sustained oscillations. The voltage $V_{b'e}$ being provided from the output voltage V_o. Initially it requires only a random voltage like noise to start the system into oscillatory mode.

So we know the condition for oscillation is $h_{fe} = \frac{L_o}{L_i}$

An amplifier bursting into oscillation instead of amplification thus it is said to be unstable.

To prevent instability, R_B and R_c is to provided damping. They should be made sufficiently small compared to X_L. so oscillations will not start and gain will be decreased.

In addition to that, the stability of bandpass or tuned amplifiers are affected by the local positive feedback provided by $C_{b'c}$ in BJT and $C_{g'd}$ in FET, it leads to unavoidable oscillations in the circuit. It can be avoided by applying the proper coupling between input and output. This is achieved by selecting a BJT or FET with small values of C'_{bc} or $C_{g's}$.

Tuned amplifiers constructed with FET are free from this oscillations. In addition there are RF transistors, working at radio frequency having low value of $C_{b'e}$. Thus circuits using them have less tendency to oscillate.

This effect can be reduced by one of the following three methods.

(I) Neutralization
(II) Unilateralization.
(III) Mismatch technique

Neutralization

Neutralization is one of the technique reducing the feedback, provided interelectrode capacitance $C_{g'd}$ or $C_{b'e}$ to obtain the better stability. In this method the feedback is cancelled by introducing a current that is equal to but 180°. out of phase with the feedback signal at the input of the active device. Thus the two signals will cancel and the effect of the feedback will be eliminated.

so as to neutralize the energy feed back taking place through inter electrode capacitance. ~~Differ~~

Hazeltine Neutralization

The following fig shows one variation of the Hazeltine circuit. In this circuit a small value of variable capacitance C_N is connected from the bottom of coil, point B, to the base. Therefore, the internal capacitance C_{bc} shown dotted, feeds a signal from the top end of the coil, point A, to the transistor base and the C_N feeds a signal of equal magnitude but opposite

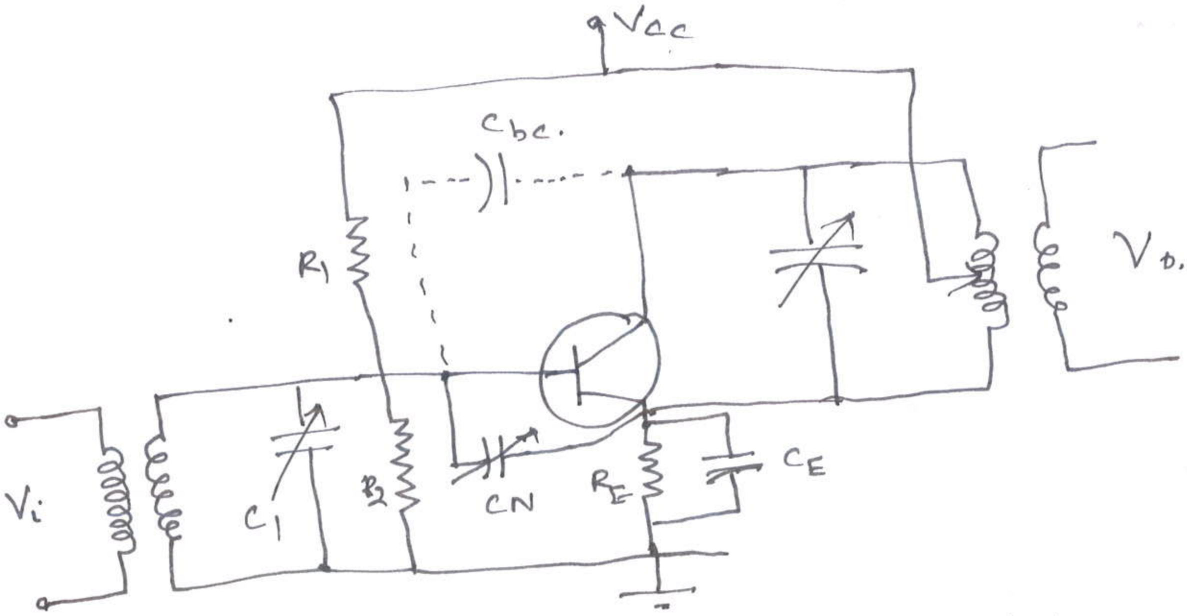
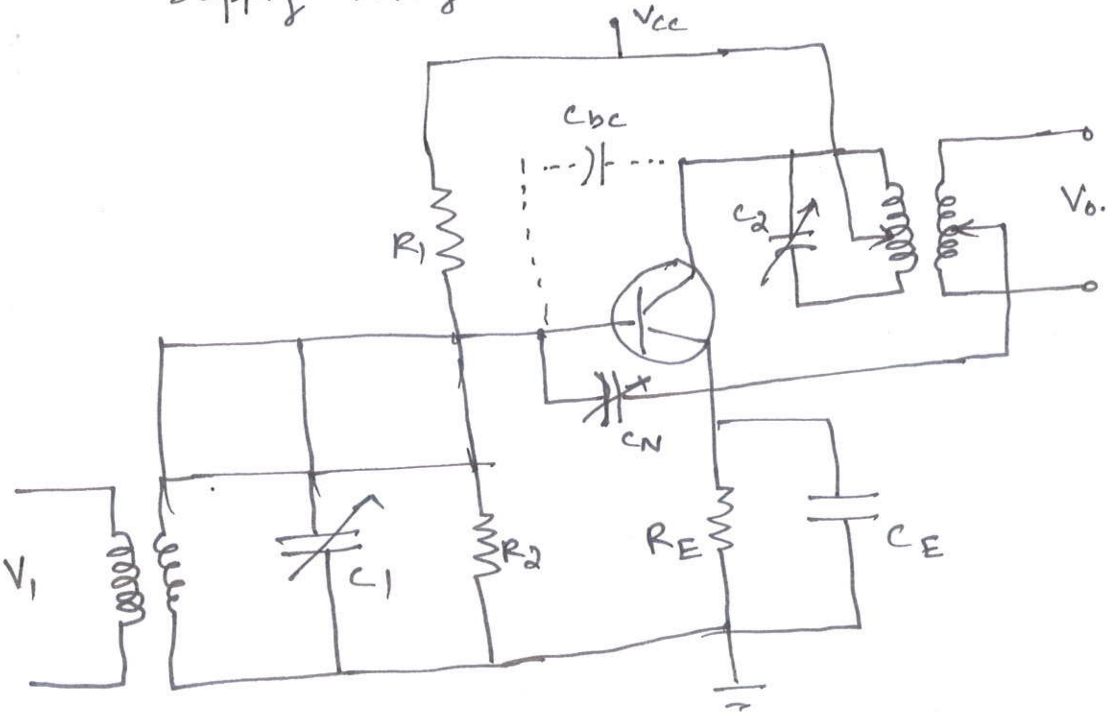


fig ① Tuned RF amplifier with Hazeltine neutralization polarity from the bottom of coil, point B, to the base. The neutralizing capacitor, C_N, can be adjusted correctly to completely nullify the signal fed through the C_{bc}.

Modified Hazeltine (Neutrodyne) Neutralization

The fig (i) shows typical neutrodyne circuit. In this circuit the neutralization capacitor is connected from the lower end of the base coil of the next stage to the base of the transistor. In principle, this circuit functions in the same manner as the Hazeltine neutralization circuit with the advantage that the neutralizing capacitor does not have the supply voltage across it.



Advantages and Disadvantages of Tuned amplifiers

Advantages

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

Disadvantages.

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

Applications of Tuned amplifier

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

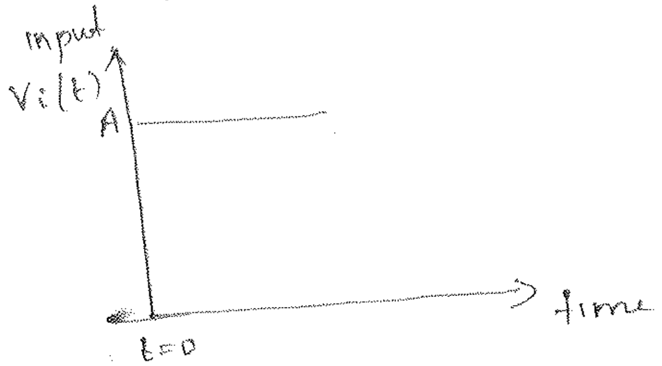
Unit - IV

Wave Shaping and Multivibrator Circuits

pulse circuits

Non Sinusoidal Signals such as Step, pulse, square wave, ramp are common in pulse circuits.

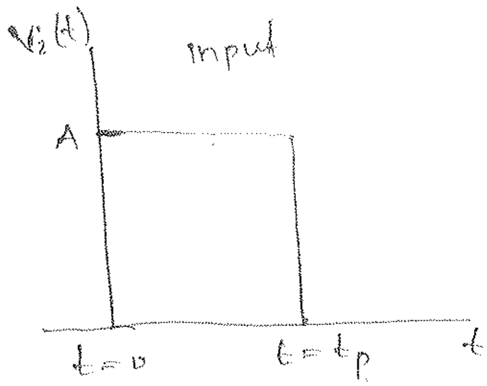
Step Waveform



$V_i(t) = 0 \quad t < 0$

$\quad\quad\quad A \quad t \geq 0$

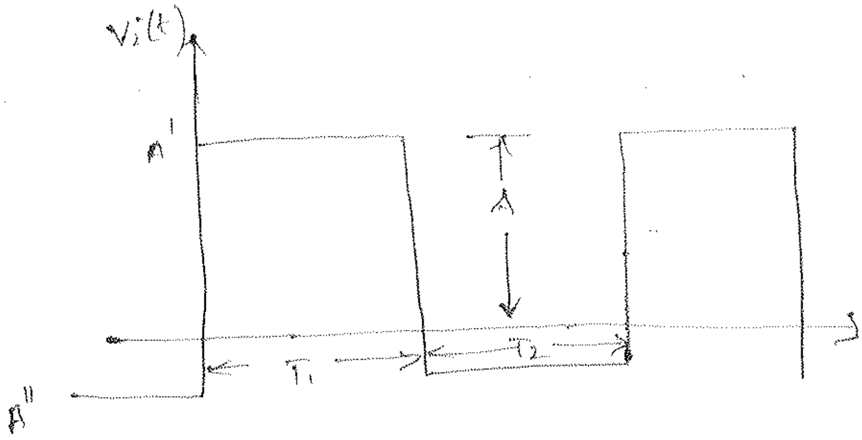
pulse waveform.



$V_i(t) = 0$ for $t < 0$ & $t > t_p$

$= A$ for $0 \leq t \leq t_p$.

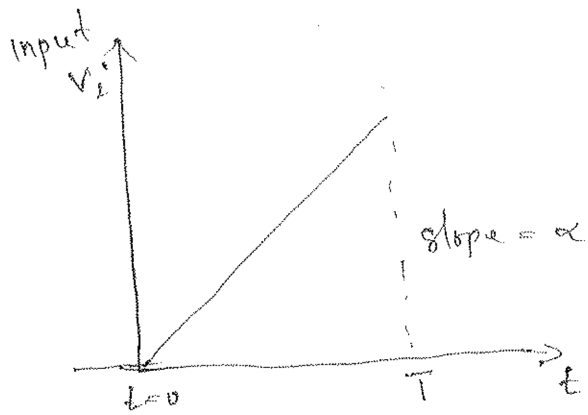
Square waveform



$V_i(t) = A' \quad 0 < t < T_1$

$\quad\quad\quad A'' \quad T_1 < t < T$

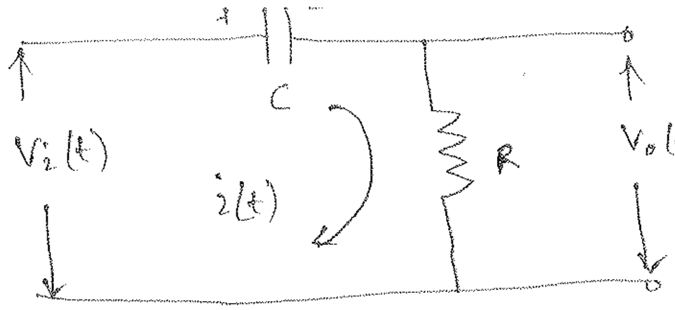
Ramp Waveform



$V_i(t) = 0$ for $t < 0$

$= \alpha t$, for $t \geq 0$

High pass RC circuit (Differentiator)



* For the RC circuit output is taken across R

* The capacitive reactance is given by

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

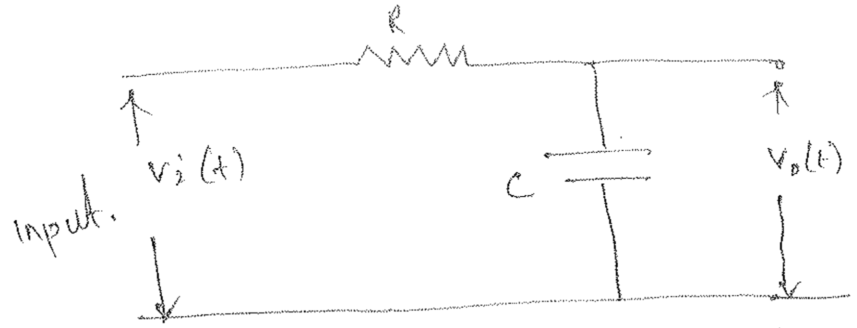
* At high frequency, the capacitor acts as short circuit and all the input appears at the output.
* The circuit attenuate the low frequencies and allows high frequencies.
* Hence the circuit is called high pass RC circuit
* if time constant $\tau = RC$ is very small compared to input it acts as differentiator.
* The drop across RC is negligible, the entire input appear across C.

The current i is

$$i = C.\frac{dV_c}{dt} = \frac{C dV_i}{dt}.$$

$$\boxed{V = iR = RC\frac{dV_i}{dt}}$$

Low pass RC circuit (Integrator)



* The output is taken across the capacitor
* The capacitive reactance depends on the frequency.
* At high frequencies, the capacitor acts as short circuit and output falls to zero.
* For low pass RC if the time constant is very large, the circuit acts as an integrator.
* The entire input appears across R, the current i is given by

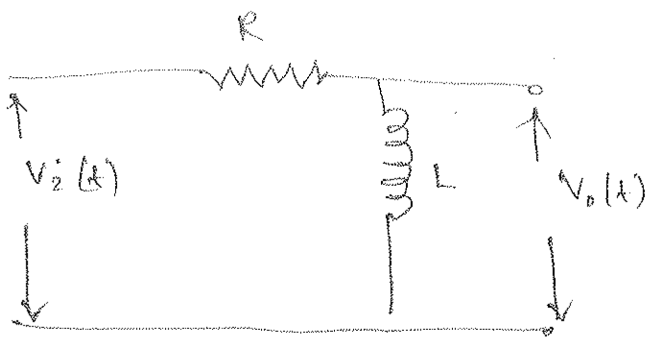
$$V_R = V_i = iR$$

$$i = \frac{V_i}{R}$$

$$V_o = V_c = \frac{1}{C} \int i \, dt$$

$$\boxed{V_o = \frac{1}{RC} \int v_i(t) \, dt}$$

High pass RL circuit

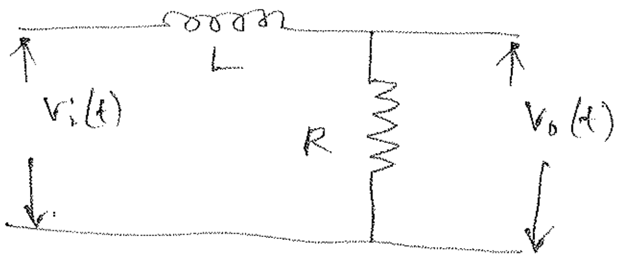


* The output is taken across inductor L
* The inductive reactance is

$$X_L = 2\pi f L$$

* At zero frequency, the inductance becomes zero and acts as short circuit and hence low frequency cannot reach the output

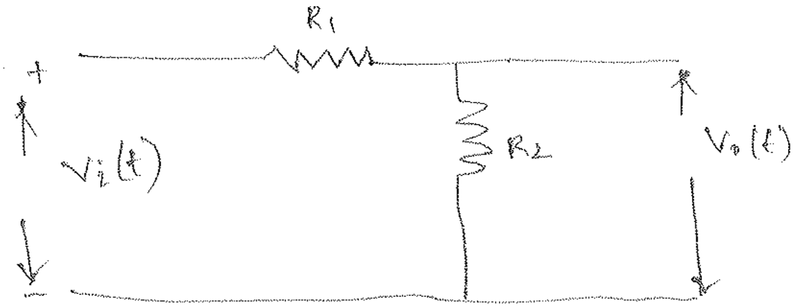
Low pass RL circuit



* The output is taken across the resistance
* The inductive reactance is $X_L = 2\pi f L$
* At low frequency the inductor acts as short circuit and hence output is maximum.
* At high frequency the reactance is very high and inductor acts as open circuit and output falls to zero.

Attenuator

* Attenuator is a device which is used to reduce the amplitude of signal waveform.



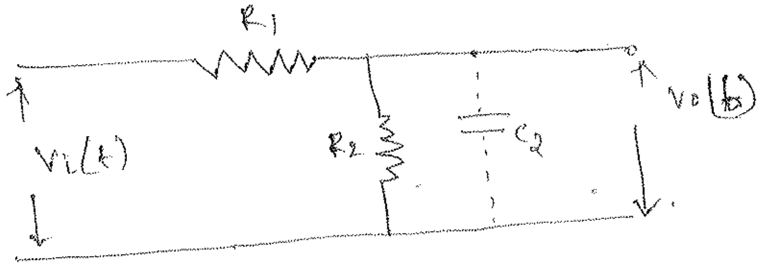
Simple attenuator

* The potential divider consisting of resistances R_1 and R_2 can be used as an attenuator.

* The input is attenuated by radio a

$$a = \frac{R_2}{R_1 + R_2}$$

* In practice a shunt capacitance exits across R_2 as C_2



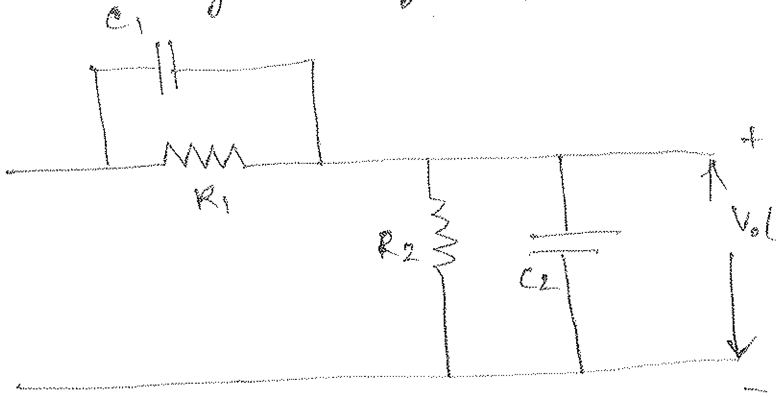
* The capacitance C_2 is the capacitance of the next circuit (ie) capacitance of output stage.

* If C_2 is absent the attenuation provided by the circuit is independent of frequency and there is no distortion.

* To get the attenuation independent of frequency the compensation is provided. Such attenuator is called Compensated attenuator.

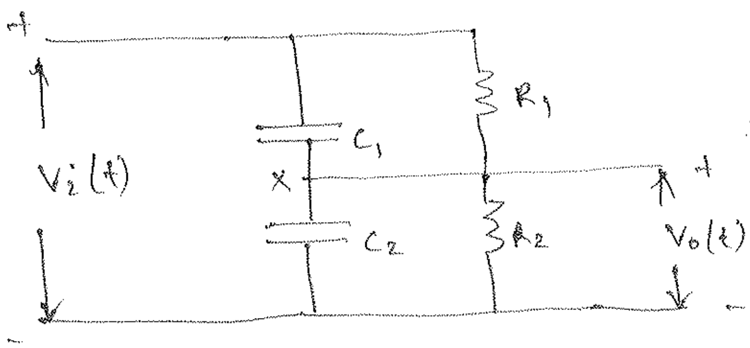
Compensated attenuator.

* The compensation is provided in actual attenuator by shunting R_1 by capacitance C_1



* The bridge will be balanced when

$$R_1 C_1 = R_2 C_2$$



* The compensation is dependent on the condition

$$R_1 C_1 = R_2 C_2.$$

* To achieve this C_1 is kept adjustable.

Clippers:

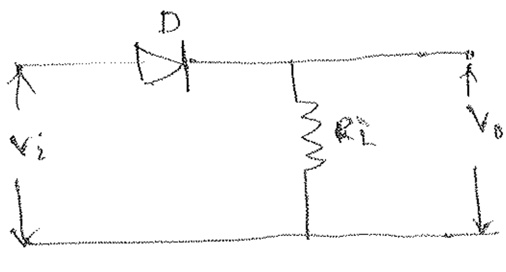
* The basic action of clipper is to remove certain portion of waveform above or below certain levels.
* The clippers are also called as limiters or slicers

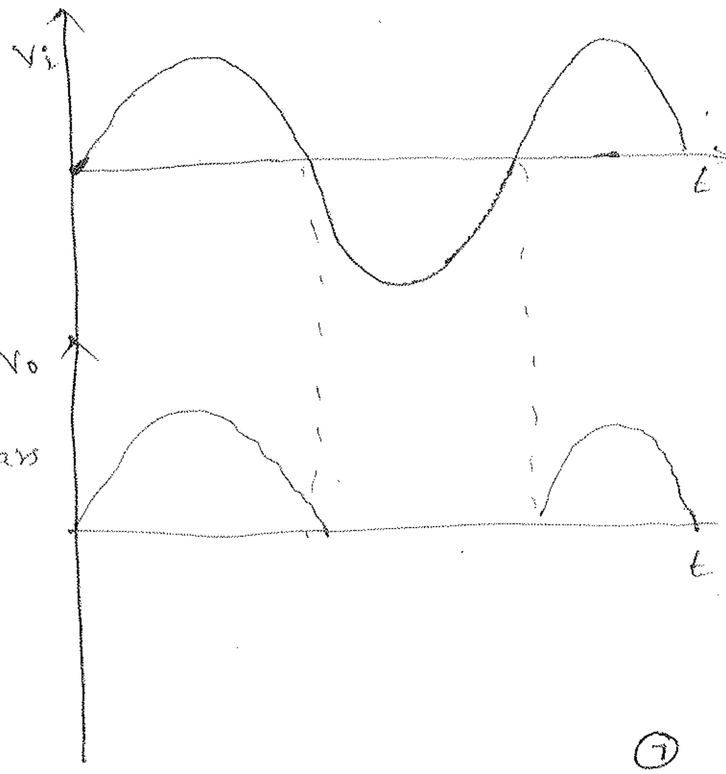
It is of two types.
1. Series clipper
2. parallel clipper

Series clipper

(I) It is used to clip off entire positive or negative half cycle of input waveforms.
(II) Diode is most important part of clipper
(III) It acts as a switch
(IV) It makes the circuit open when reverse biased and closed when forward biased.

Series negative clipper.

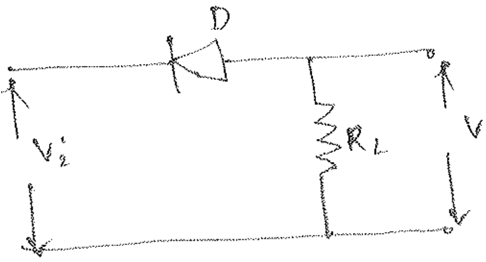


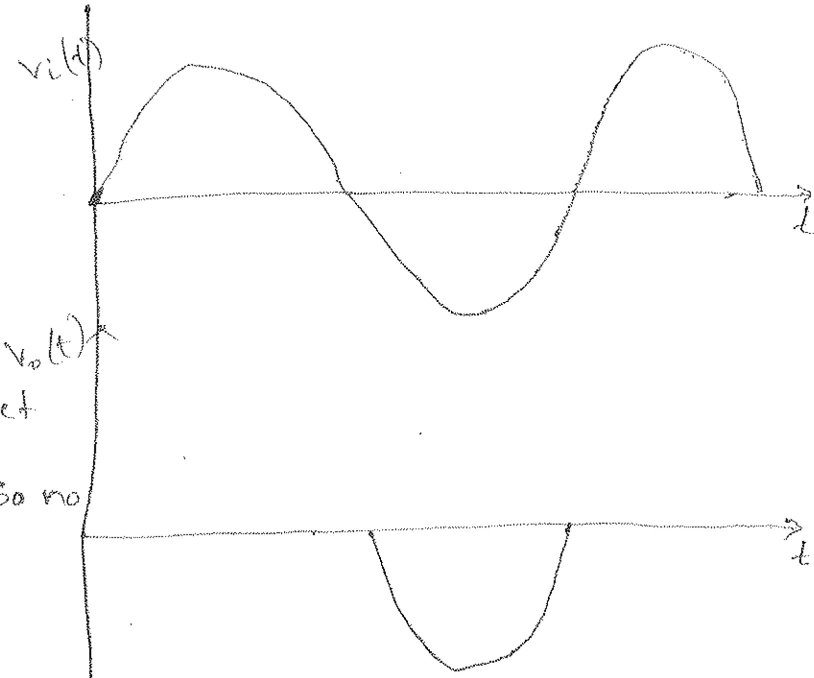


* Diode 'D' is forward biased and the voltage across R_L appears in positive half cycle.

* During negative half cycle diode 'D' is reverse biased and the voltage does not appear across R_L

Series positive clipper

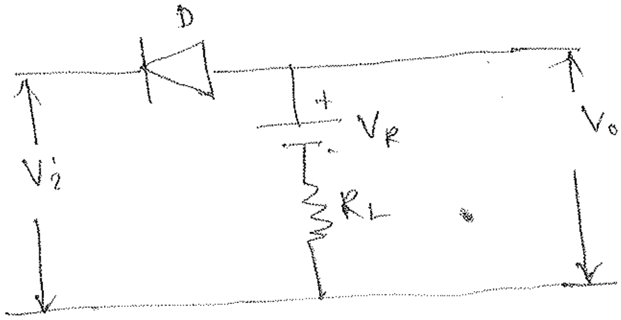


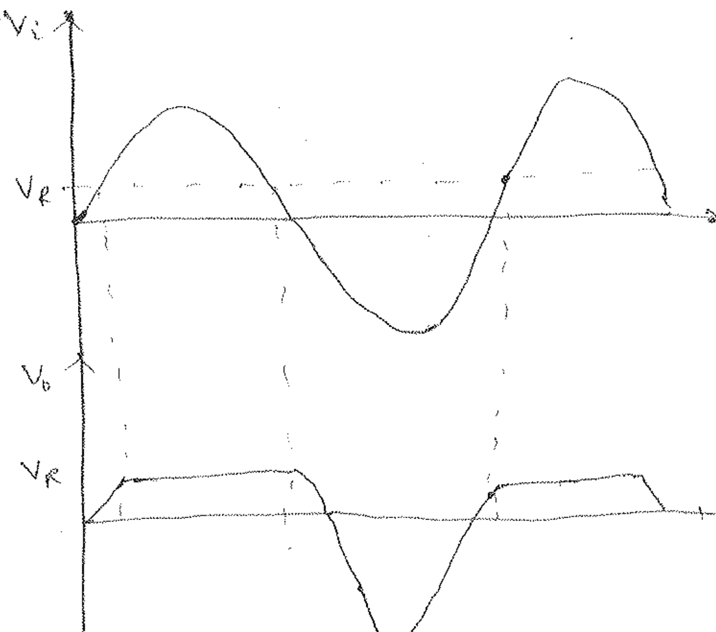


* The diode does not conduct during positive half cycle. So no voltage appears across R_L

* The diode conducts during negative half cycle and the voltage appears across R_L.

Biased Positive clipper.

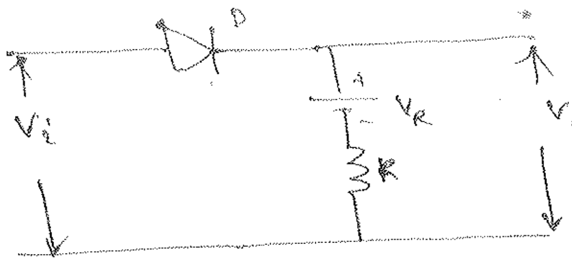


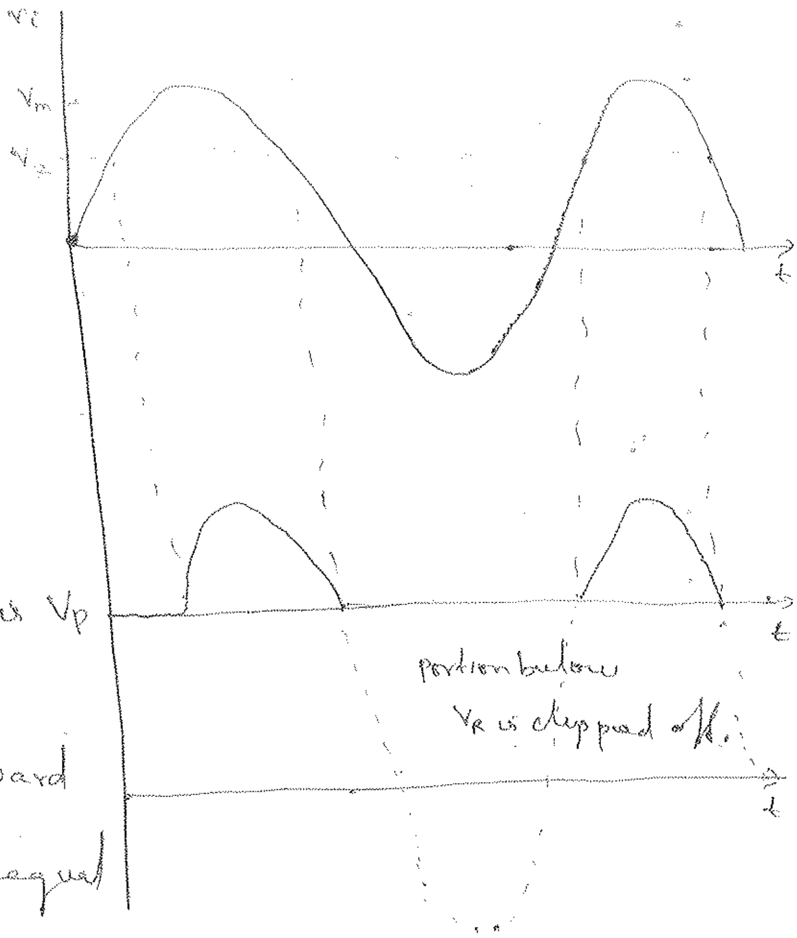


* When $V_i < V_R$ the diode becomes forward biased and the output voltage is equal to input voltage.

* When $V_i > V_R$ the diode is reverse biased.

Biased negative clipper.



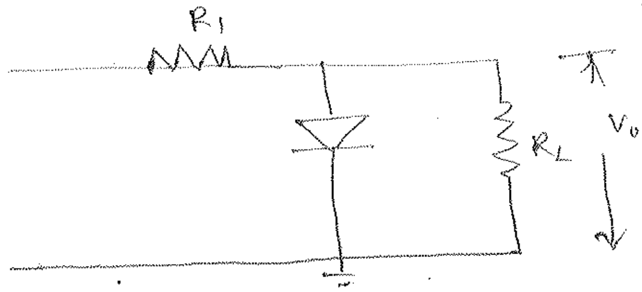


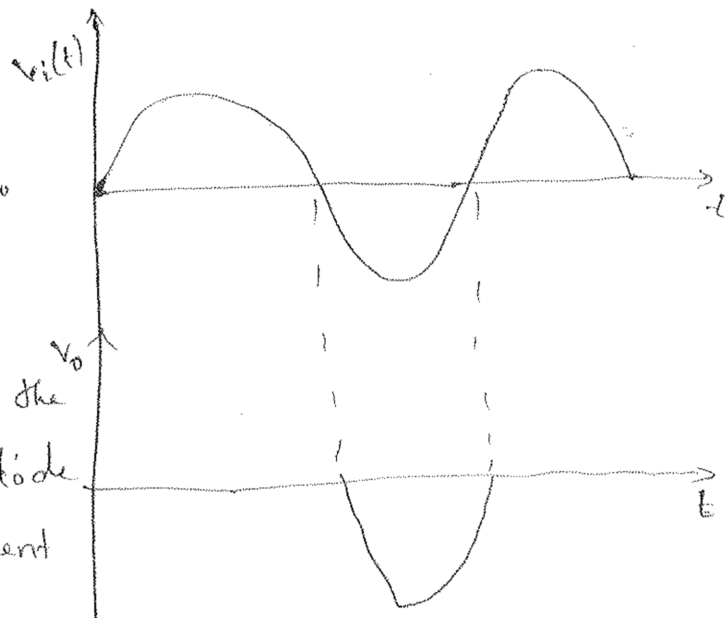
* When $V_i < V_R$ the diode is reverse biased and circuit is open.
* When $V_i > V_R$ diode is forward biased and output voltage is equal to input voltage.

parallel clipper

* In parallel clipper, diode is connected across load terminals

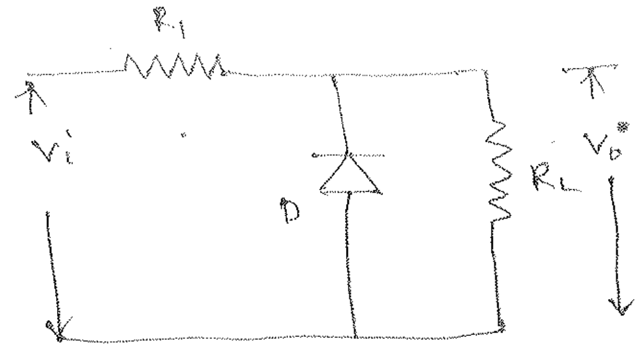
positive parallel clipper.

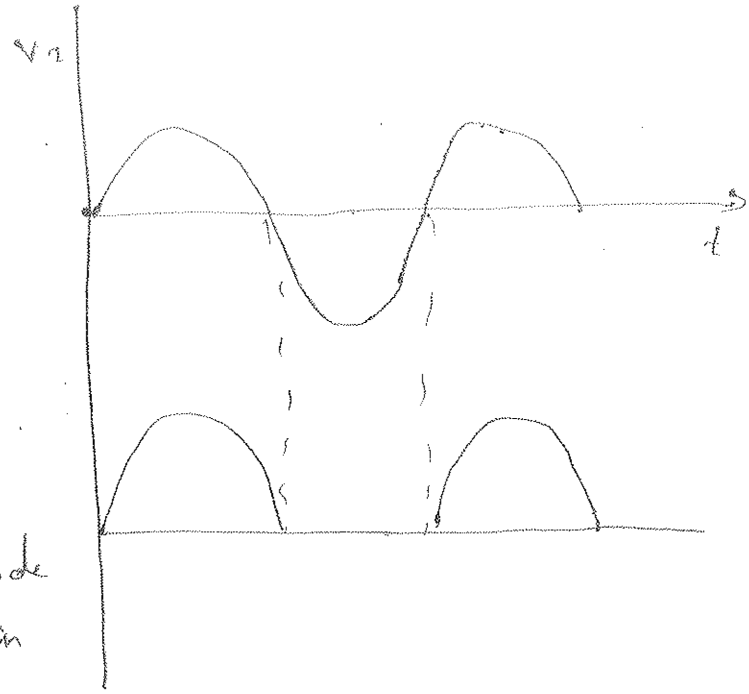




* During positive half cycle the diode is forward biased, the diode is short circuit and no current flows through R_L
* During negative half cycle the diode is open and the output appears across R_L

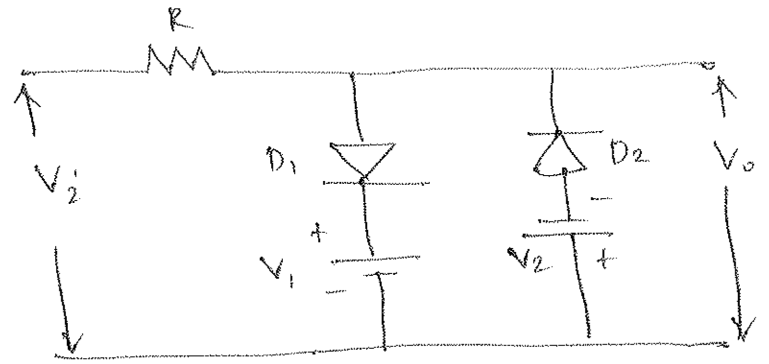
parallel negative clipper

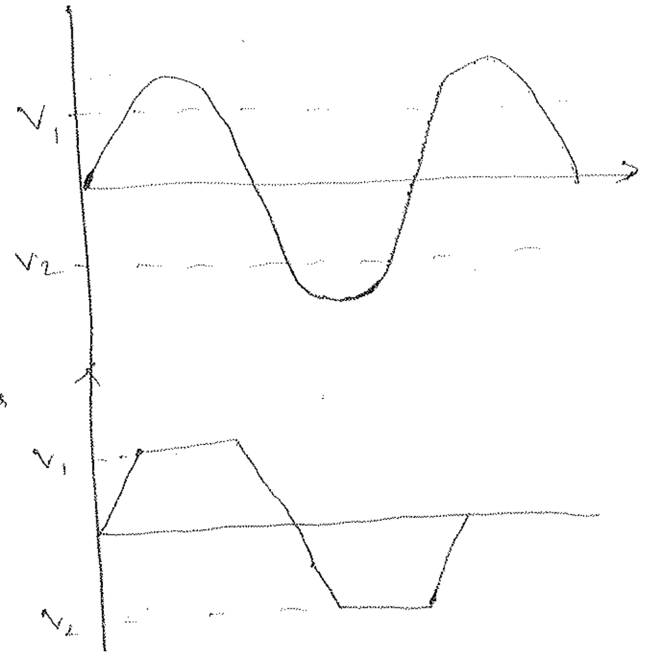




* When Vi is positive the diode is reverse biased and the Vin get reproduced at output
* When Vi is negative the diode is forward biased and short circuit and no voltage appear across R_L

Two way parallel clipper





* When $V_i > V_1$ the diode (D_1) is forward biased and conducts V
* When $V_i < V_2$ the diode D_2 is forward biased.

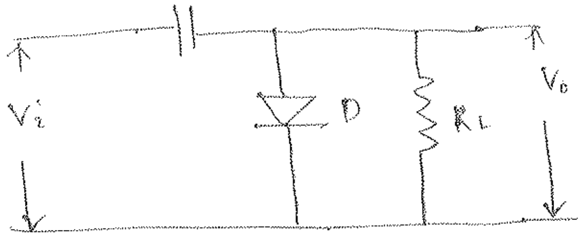
clampers.

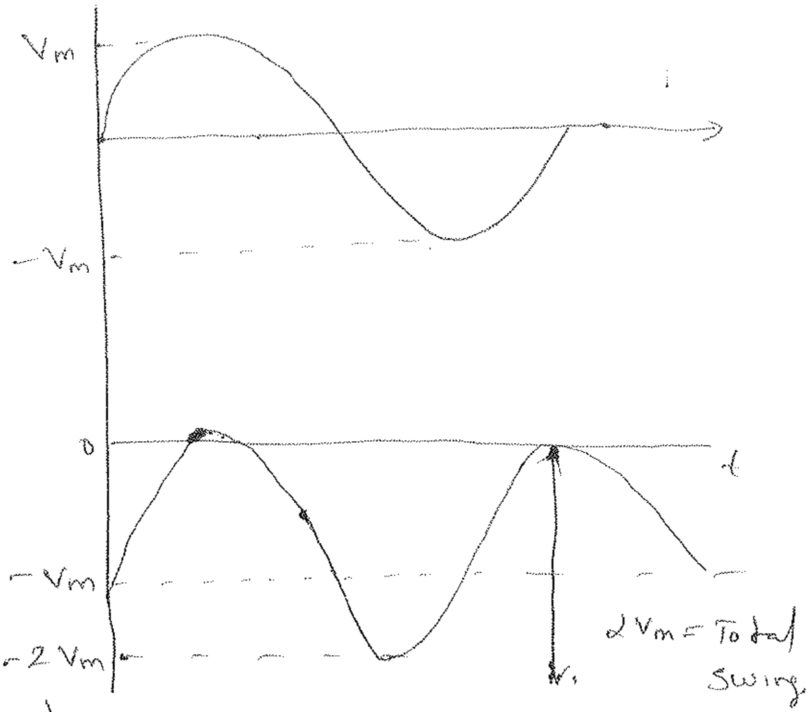
* It adds d.c level to the a.c output signal
* It is also called d.c restorer

It is of two types.

(I) Negative clamper
(II) positive clamper.

Negative clamper





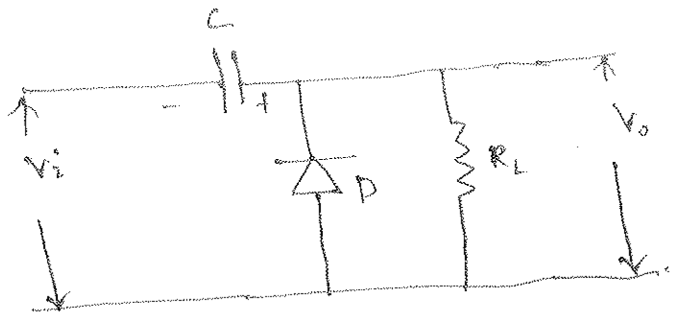
* it consists of capacitor c, diode D and Resistor R
* During the positive half cycle of input voltage V_i the capacitor get charged up to maximum value Vm of input signal.
* As diode 'D' is 'ON' the output voltage V_o is zero.
* In negative half cycle, the diode will remain reverse biased. The capacitor starts discharging through the resistance R_L
* The capacitor holds all the charge and remains charged to Vm during this period.

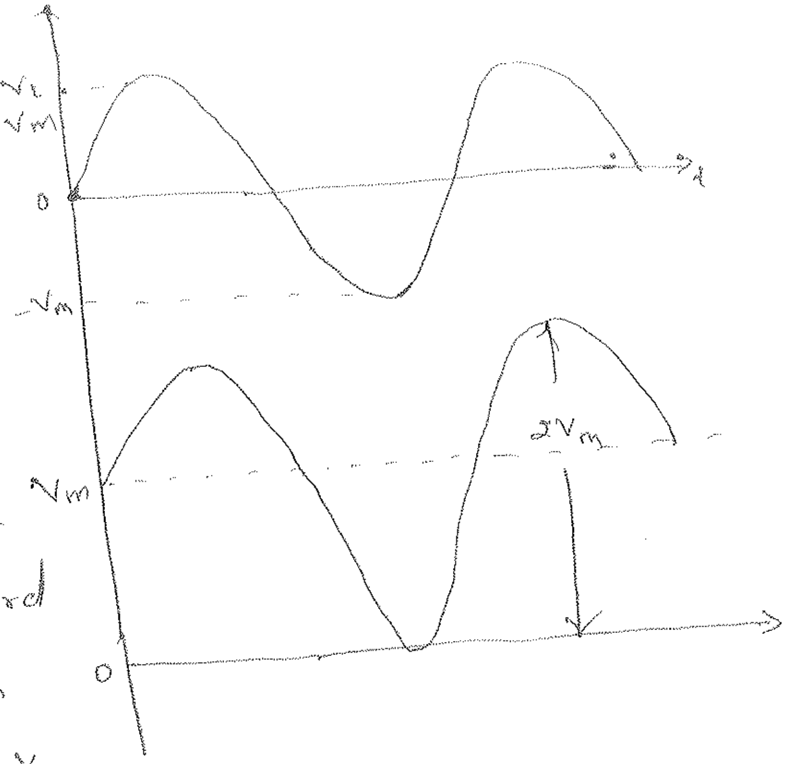
$$V_o = -V_m \; ; \text{ for } V_i = 0$$

$$V_o = 0 \; ; \text{ for } V_i = V_m$$

$$V_o = 2V_m \; ; \text{ for } V_i = -V_m.$$

positive clamper





* During the first negative half cycle, diode D is forward biased and the capacitor is charged to maximum value V_m

* The capacitor once charged acts as battery of voltage V_m

* In the positive half cycle the diode is reverse biased and capacitor starts discharging through R_L

$$V_o = V_i + V_m$$

$$V_o = V_m \quad \text{for} \quad V_i = 0$$

$$V_o = 2V_m \quad \text{for} \quad V_i = V_m$$

$$V_o = 0 \quad \text{for} \quad V_i = -V_m$$

Multivibrators

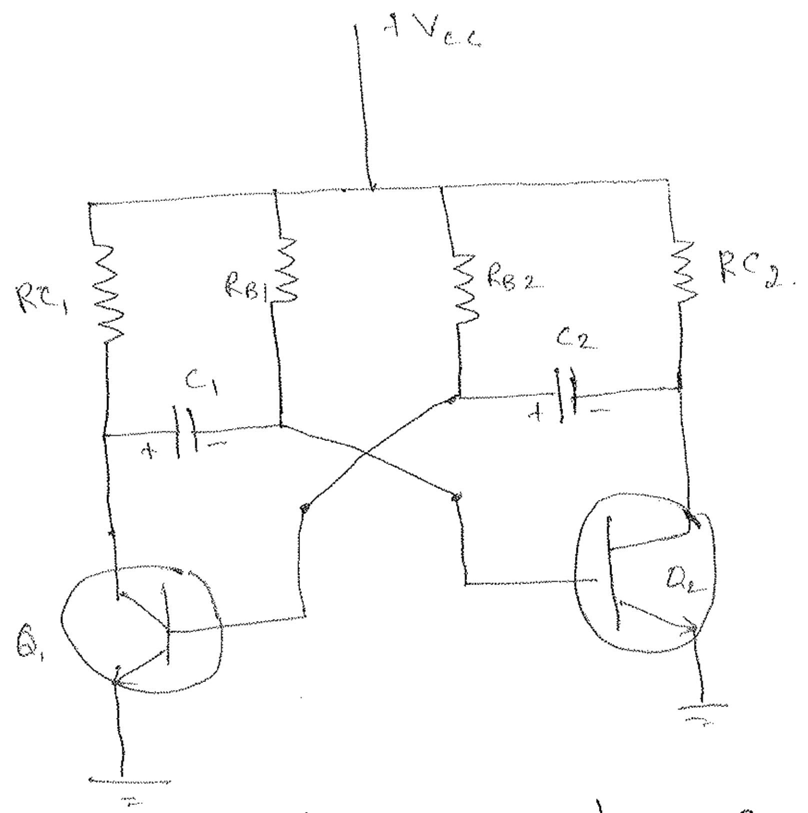
* It is a two stage switching circuit in which the o/p of first stage is fed to the input of II^{nd} stage
* The output of two stages are complementary

Three types of multivibrator.

(i) Astable multivibrator
(ii) Monostable multivibrator
(iii) Bistable multivibrator.

Astable Multivibrator

Collector Coupled Astable Multivibrator.



* Astable multivibrator generates square wave without any external triggering pulse
* It has no stable states (ie) 2 quasi stable state
* it switches back and forth from one state to other, remaining in each state for time

depending upon discharging of capacitive circuit

* The components in one half is identical to other

* When $+V_{cc}$ is applied Q_1 conducts and Q_2 is cutoff

$$V_{C1} = V_{CE(sat)} \text{ (ie) } 0V \; ; \; V_{C2} = +V_{cc}$$

* C_1 charges exponentially with time constant R_1C_1

* V_{B2} increases to V_{cc} when it reaches cut in and Q_2 starts conducting and V_{C2} falls to V_{CE} sat.

* V_{B1} falls due to capacitive coupling between collector of Q_2 & base of Q_1 to OFF.

* V_{C1} coupled through C_1 to base of Q_2 causing overshoot in V_{B2}

* $\{ V_{B1} = -Ve, \; V_{C1} = V_{cc} \; ; \; V_{B2} = V_{BE(sat)}$

$V_{C2} = V_{CE(sat)}$

* Q is OFF and Q_2 is ON, V_{B1} increases exponentially with time constant R_2C_2 to V_{1C}

* Q_1 is driven into saturation and Q_2 to cutoff

$V_{B1} = V_{BE(sat)}$; $V_{C1} = V_{CE(sat)}$ / V_{B2} is $-Ve$ & $V_{C2} = V_{cc}$.

Expression for time period T in astable multivibrator.

The equation of output can be written as

$$V_o = V_f - (V_f - V_i)e^{-t/\tau}$$

$V_o \rightarrow$ base voltage of B_2

As the capacitor C_2 discharges exponentially, the voltage V_{B_2} at B_2 increases exponentially

V_i – initial value of $V_{B2} = -V_{cc}$

V_f – final value of $V_{B2} = +V_{cc}$

$$V_{B2} = V_{cc} - (V_{cc} - (-V_{cc}))e^{-t/R_2C_2}$$

$$= V_{cc} - 2V_{cc}\, e^{-t/R_2C_2}$$

$$= V_{cc}\left(1 - 2e^{-t/R_2C_2}\right)$$

At switching time, $t = T_2$ and $V_{B2} = V_\gamma$

$$V_\gamma = V_{cc}\left(1 - 2e^{-T_1/R_1C_1}\right)$$

To obtain T_2 sub $V_\gamma = 0V$

$$0 = V_{cc}\left(1 - 2e^{-T_2/R_2C_2}\right)$$

$$1 - 2e^{-T_2/R_2C_2} = 0$$

$$e^{-T_2/R_2C_2} = 0.5$$

$$\ln\left(e^{-T_2/R_2C_2}\right) = \ln(0.5)$$

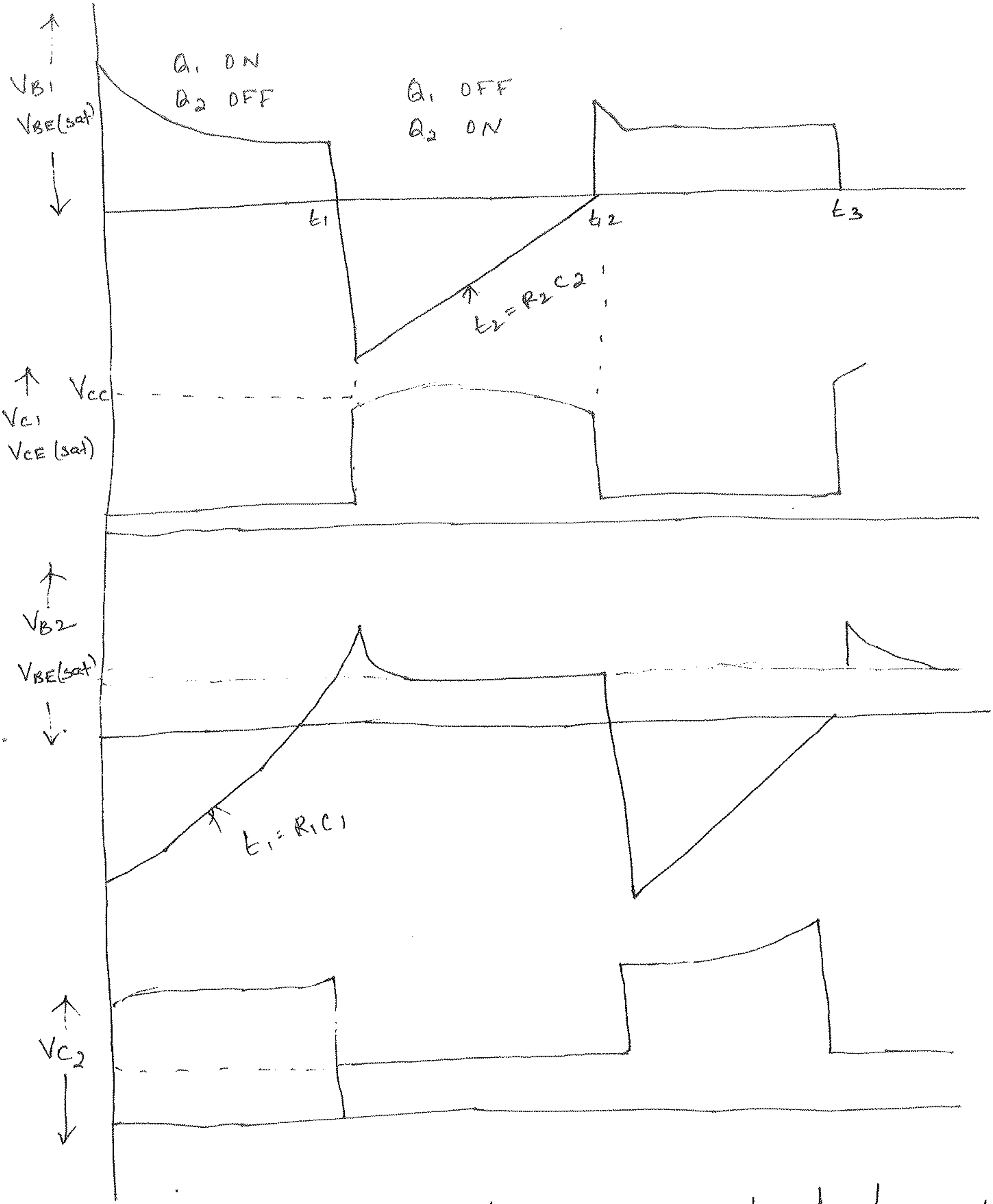
$$\frac{-T_2}{R_2C_2} = 0.693$$

$$T_2 = 0.693\, R_2C_2$$

ON time for Q_2 is $T_2 = 0.693\, R_2C_2$

$|||^{ly}$ ON time for Q_1 is $T_1 = 0.693\, R_1C_1$

Total period $T = T_{ON} + T_{OFF} = 0.69\,(R_1C_1 + R_2C_2)$



* Q_2 is ON falling of V_{C_2} permit discharges of capacitor C_2, drives Q_1 to cutoff.

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

$$T = 1.386\,RC \quad \text{and} \quad f = \frac{1}{T} = \frac{1}{1.386\,RC}$$

To ensure oscillation, the value of resistor should satisfy the following condition

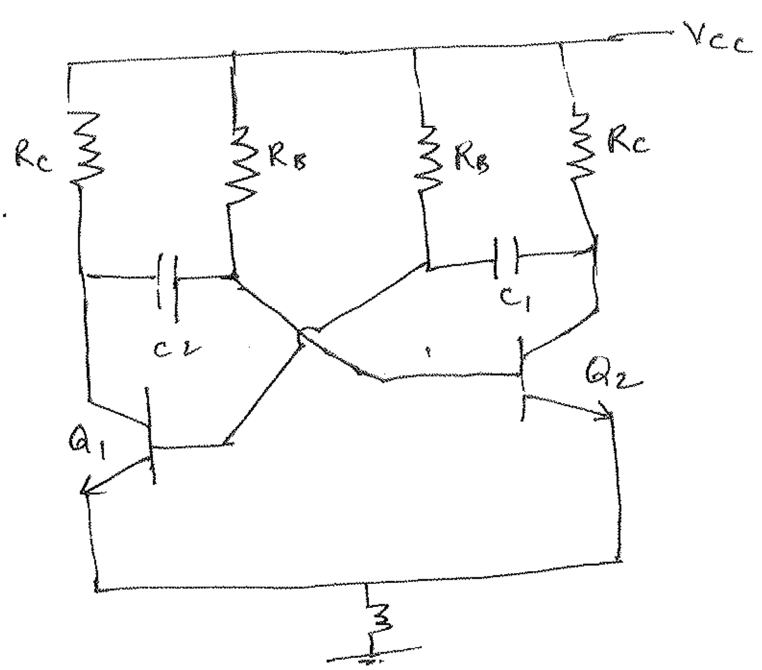
$$R_1 \leq h_{fe\,min} R_{C1} \quad \text{and} \quad R_2 \leq h_{fe\,min} R_{C2}.$$

$h_{fe\,min}$ is minimum value of d.c current gain of Q_1 and Q_2.

Application

1. It is used as square wave generator.
2. It is source of production of harmonic frequencies
3. It is used in construction of digital voltmeter and SMPS
4. It is operated as oscillator over a wide range of audio and radio frequencies.

Emitter Coupled astable multivibrator.



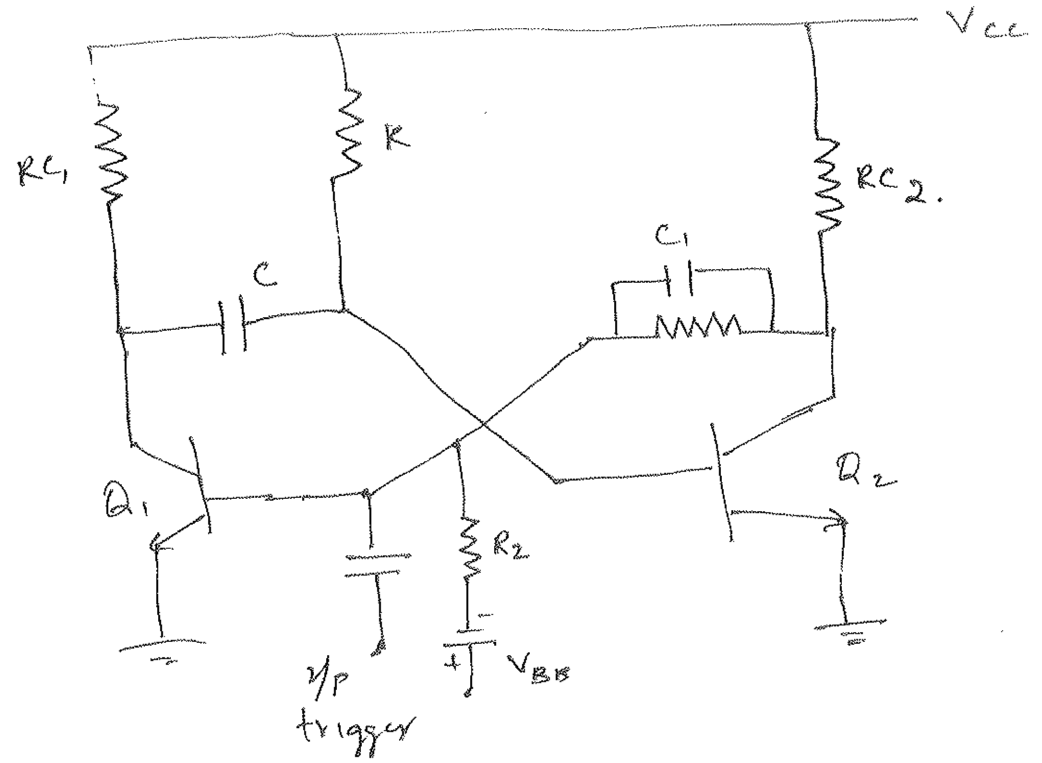
* The circuit is symmetry
* Both the transistors Q_1 and Q_2 remain OFF or ON
* So the circuit may not oscillate.
* By shorting base & emitter of one of the transistor for short time, oscillation is started
* When one Transistor starts conducting V_E of other increases and V_B decreases.
* R_E eliminates both transistor remain ON at the same time

$$R_E = V_{cc}/2I_c$$

Monostable Multivibrator

* It has one stable state and one quasi stable state.
* It acts as one shot multivibrator
* It remains in its stable state until an input pulse triggers it into quasi stable state for time determined by RC
* It cannot generate square wave of its own.
* only external trigger pulse cause to generate rectangular wave.

Collector Coupled Monostable Multivibrator.



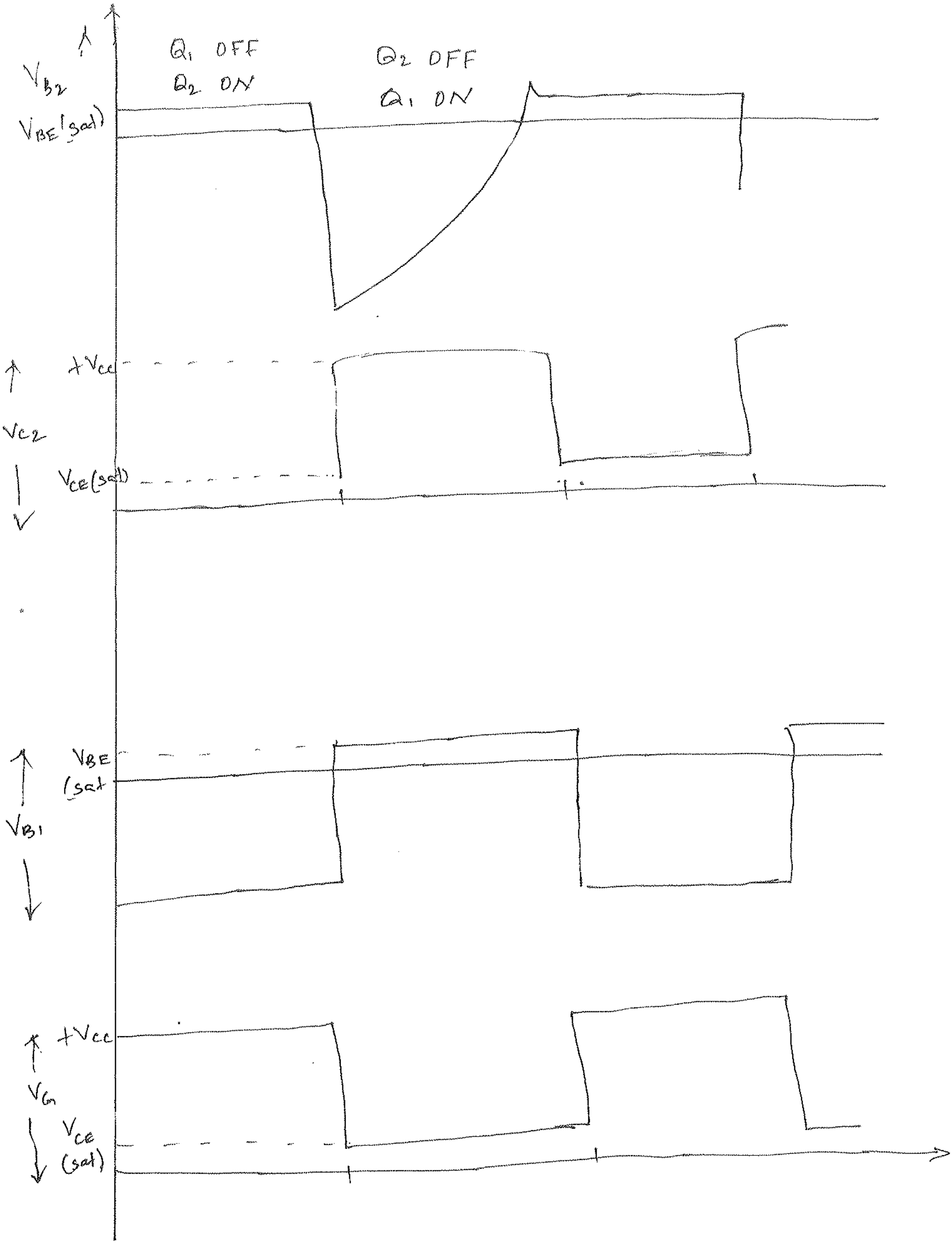
* It has two identical transistor Q_1 and Q_2 and $R_{C1} = R_{C2}$
* The output of Q_2 coupled to input of Q_1
* C_1 is the speed up capacitor to speed up transition.
* R_2 and $-V_{BB}$ are chosen to reverse bias Q_1 to keep it in OFF state.
* Vcc and R forward bias Q_2 and keep it in ON. This is stable state
* When +ve pulse of short duration is applied to base of Q_1 through C_2, Q_1 conducts.
* V_{C1} decreases which is coupled to Q_2 through 'C'

* This decreases forward bias of Q_2 and its collector current decreases.
* This decreases forward bias on Q_2 and IC_2 decreases
* 'c' is charged approximately to +Vce through Vcc, R and Q.
* As 'c' discharges base of Q_2 is forward biais and collector current starts to flow to Q_2.
* Q_2 is driven to saturation and Q_1 to cutoff.
* It remains in this State until another pulse Causes to switch over the states.
* duration of output pulse of monostable multivibrator in $T = 0.693\ RC$

Application

1. Used as adjustable pulse width generator
2. generate Sharp pulse from distorted pulse

Output wave for for Monostable



Expression for T for monostable.

* The pulse width T is the time for which the circuit remains in quasi stable state.

The exponential charging of capacitor is

When $t = 0^+$ $\quad V_i = V_\sigma - I_1 R_c$

$t = \infty$ $\quad V_f = V_{cc}$

Here
$$V_c = V_f - (V_f - V_i)\, e^{-t/\tau}.$$

τ = Time constant.

$$\therefore V_{B2} = V_{cc} - (V_{cc} - V_\sigma + I_1 R_c)\, e^{-t/\tau}$$

At $t = T$ $\quad V_{B2} = V_\gamma$

$$\therefore V_\gamma' = V_{cc} - (V_{cc} - V_\sigma + I_1 R_c)\, e^{-t/\tau}.$$

$$T = \tau \ln \left(\frac{V_{cc} + I_1 R_c - V_\sigma}{V_{cc} - V_\gamma} \right)$$

$V_\sigma = 0.2$ for Ge

0.7 for Si

When Q_1 is in saturation

$V_{C_1} = V_{CE\,(sat)}$ and $I_1 R_c = V_{cc} - V_{CE\,(sat)}$

Since $V_\sigma = V_{BE\,(sat)}$.

$$T = \tau \ln \left(\frac{V_{cc} + V_{cc} - V_{CE\,(sat)} - V_{BE\,(sat)}}{V_{cc} - V_\gamma} \right)$$

$$= \tau \ln \left(\frac{2V_{cc} - V_{CE\,(sat)} + V_{BE\,(sat)}}{V_{cc} - V_\gamma} \right)$$

$$= \tau \ln 2 \left\{ \frac{V_{CC} - \left(\frac{V_{CE(sat)} + V_{BE(sat)}}{2} \right)}{V_{CC} - V_\gamma} \right\}$$

$$= \tau \ln(2) + \tau \ln \left\{ \frac{V_{CC} - \frac{V_{CE(sat)} + V_{BE(sat)}}{2}}{V_{CC} - V_\gamma} \right\}$$

At room temperature $V_{CE(sat)} + V_{BE(sat)} = 2V_\gamma$

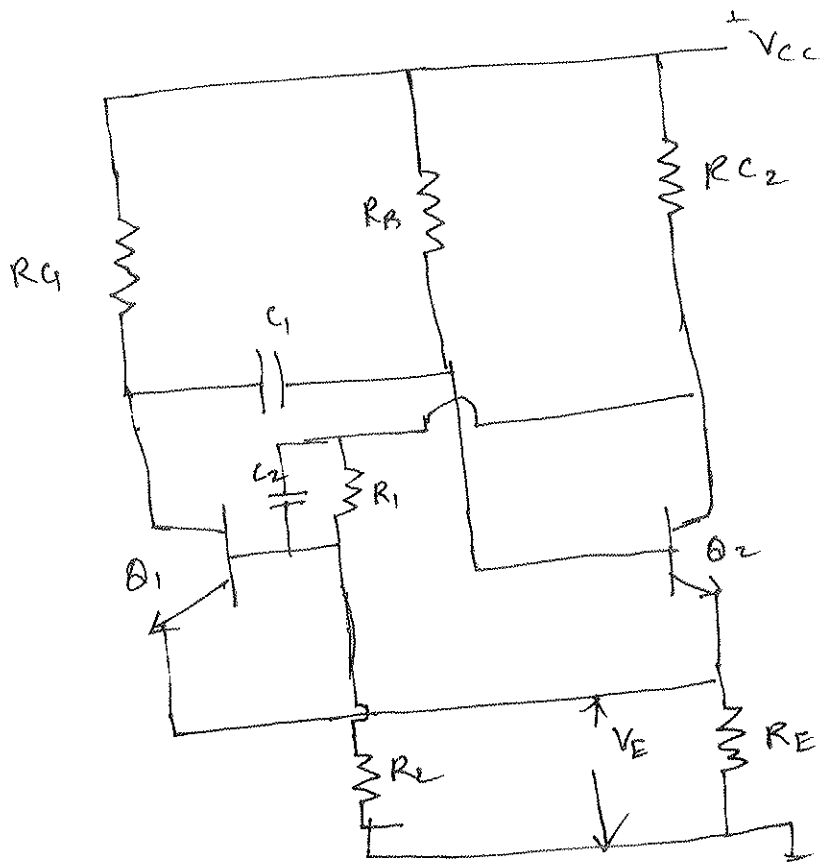
$\therefore \quad T = \tau \ln(2) + \tau \ln(1) = \tau \ln(2)$

Time constant $\tau = RC$

$$\boxed{T = 0.693RC}$$

The duration of output pulse of monostable multivibrator is given by $T = 0.693RC$

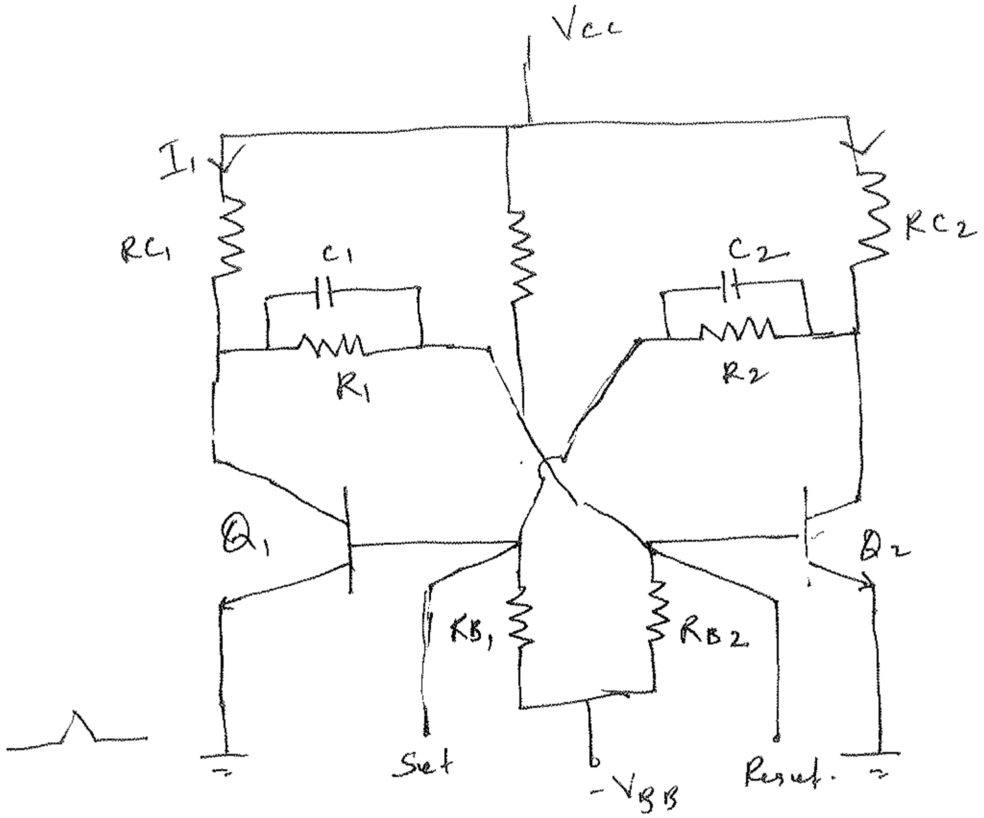
<u>Emitter Coupled monostable multivibrator.</u>



* Both emitter of Q_1 and Q_2 connected to ground through R_E
* R_2 is connected to ground (instead of $-V_e$ supply hence uses single supply)
* The transistors are unsaturated due to presence of R_E (hence switch faster than collector coupled).
* Q_2 is supplied with I_B through R_B so normally ON with V_E across R_E
* V_{C2} becomes lower than V_{CC}
* Q_1 is biased from V_{C2} through R_1 and R_2 selected $V_{B1} < V_E$ to keep Q_1 OFF & Q_2 ON
* $V_{C1} = V_{CC}$ and voltage across C_1 is $(V_{CC} - V_{B2})$
* Q_1 is ON, V_{C1} falls, the charge on C_1 causes V_{B2} to drop $\rightarrow$ Q_2 OFF.
* As V_{C2} increases, increases V_{B1} and $V_E = V_{B1} - V_{BE}$, to allow Q_2 remain OFF until C_1 discharged to V_{B2} which increases to increase V_E.

Bistable multivibrator

* it is also called as Flip flop, Eccles jordan circuit or binary two stable state circuit
* it has two stable states.
* When trigger pulse is given it switches from one state to other
* When another trigger pulse is applied it switch back to original state.



* it consists of two npn transistor
* The output of Q_2 is coupled to base of Q_1 through R_2
* Similarly output of Q_1 is coupled to base of Q_2 through R_1
* When abruptly changing pulse is applied to circuit transition takes place instantaneously

* The transition time must be kept small.
* Capacitors C_1 and C_2 improve switching characteristics of circuit by passing high frequency component Square wave pulse.
* When the circuit is switched ON, one transistor starts conducting more than the other
* This transistor is driven to saturation and the other transistor to cutoff.
* If Q_1 is ON and Q_2 OFF, it remains in this state until trigger is applied.
* If positive going trigger is applied to reset input (ie base of Q_2) I_{C2} increases and V_{C2} decreases which is applied to base of Q_1
* So Q_1 is turned to cutoff.

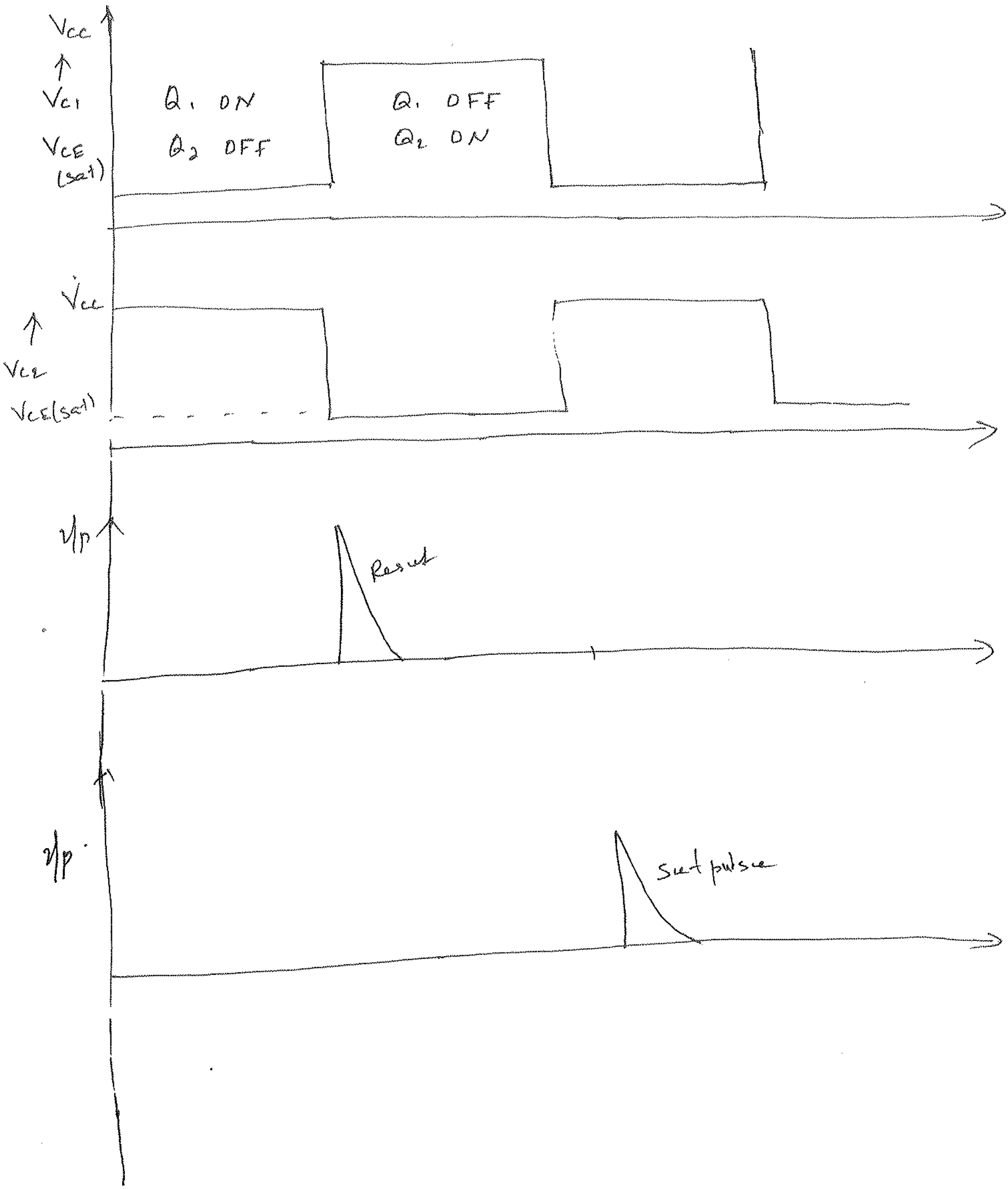
Application

(1) it is used as memory element in shift register

(2) it is used to generate square wave

(3) it is used as frequency divider

output wave form for bistable multivibrator.

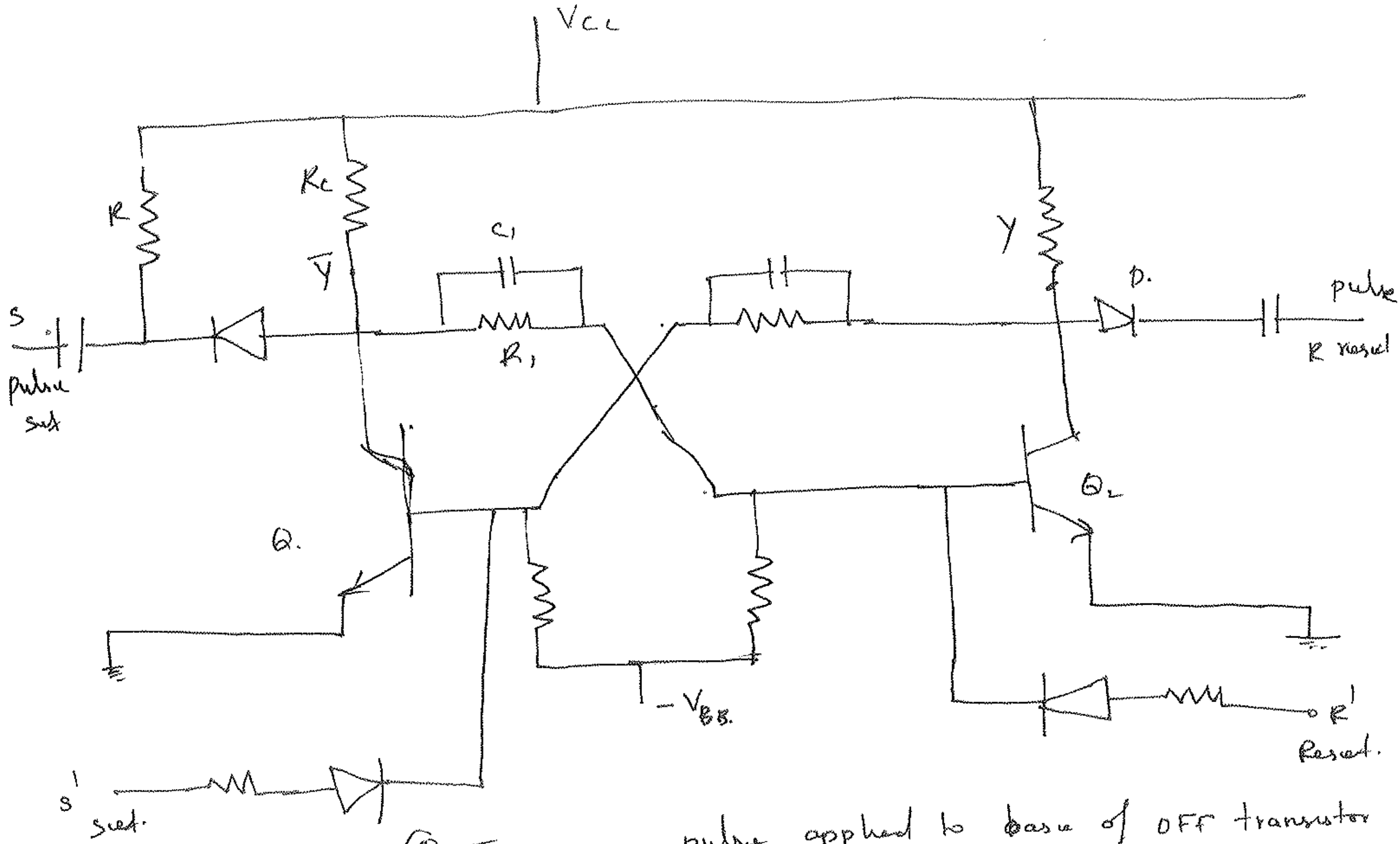


Triggering methods of bistable multivibrator.

* Triggering induce transition of flip flop from one state to other

Two types of triggering
(i) unsymmetrical triggering
(ii) symmetrical triggering

Unsymmetrical triggering



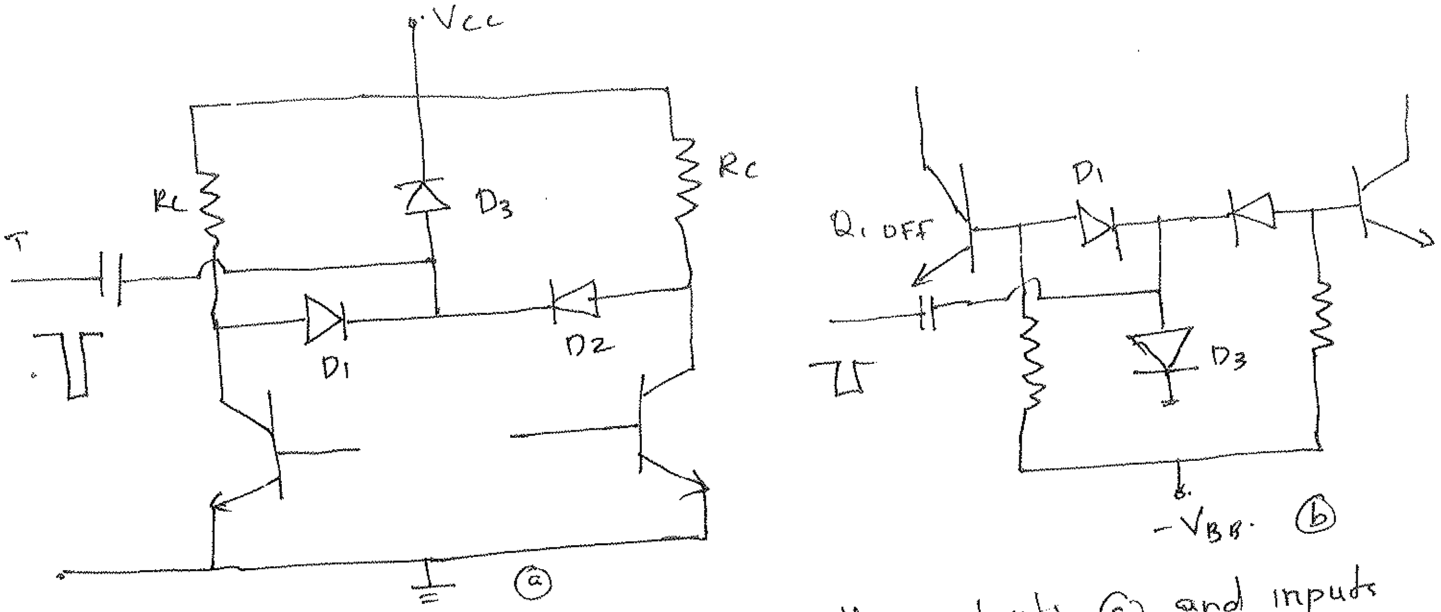
(a) Triggering pulse applied to base of OFF transistor

* Triggering signal from first input is applied to set circuit in one particular state.

* Triggering signal from second input is applied to reset circuit.

Symmetrical Triggering

* In symmetrical triggering, each successive triggering changes the state of flip flop.
* It is used in binary counting circuits
* It is symmetrical form of unsymmetrical except resistor R is replaced by D_3.
* When Q_2 is ON the drop ($= V_{cc}$) across R_c reverse biases D_2



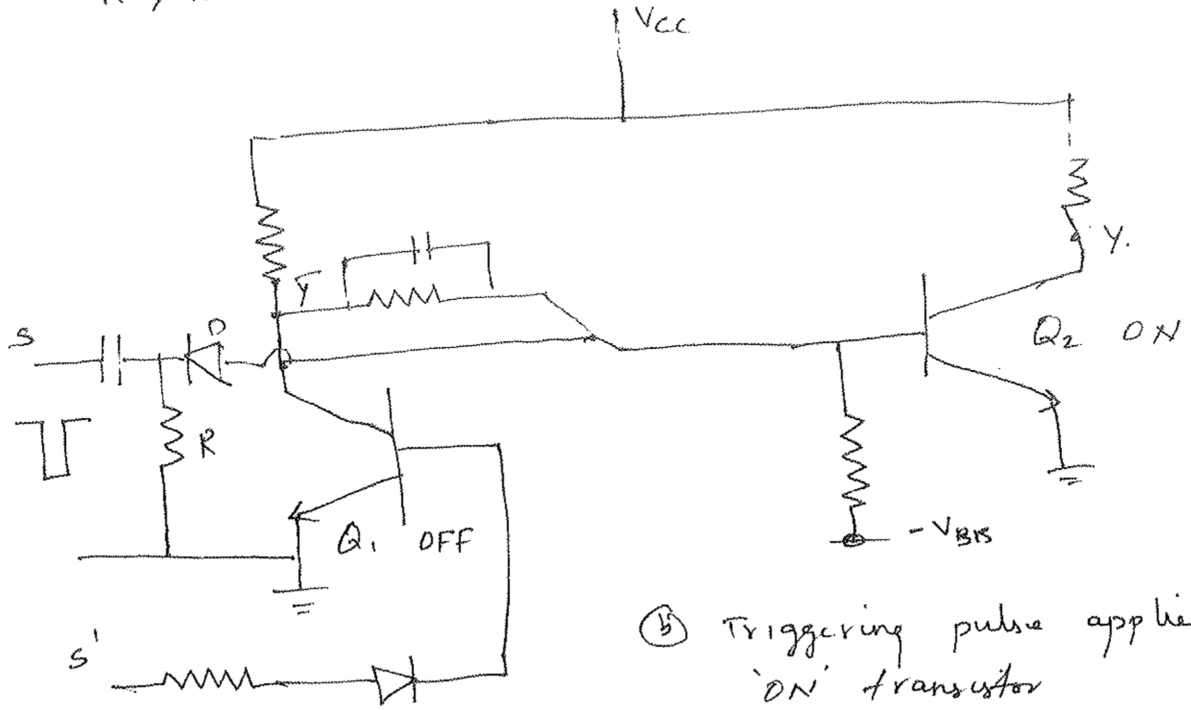
Symmetrical triggering at the outputs (a) and inputs (b) of amplifiers

* At the same time Q_1 is OFF and hence there is zero voltage across R_c of Q_1 and D_1 is at zero bias

→ When a negative going triggering pulse is applied D_1 conducts.

* Hence this triggering signal reaches collector.

* if transistor Q_1 OFF diode 'D' will transmit
* If transistor Q_1 is ON diode D is reverse biased
* When -Ve going pulse is applied to set terminal S then Q_1 is ON and Q_2 is OFF and output $Y=1, \bar{Y}=0$
* When -Ve going pulse is applied to reset terminal R, then Q_1 is OFF and Q_2 is ON and output $Y=0 ; \bar{Y}=1$



(b) Triggering pulse applied to base of 'ON' transistor

* Here R is connected to ground instead of supply voltage.
* The negative triggering pulse is applied to set terminal S that is applied to base of ON stage Q_2 through diode 'D'
* A consists of two identical transistors Q_1 and Q_2 coupled through emitter resistor R_E

It consists Resistors R_1 and R_2 form a voltage divider across V_{C1} and ground

* This provides a forward bias to base emitter junction of transistor Q_2.

* When supply is switched ON, with no input signal transistor Q_2 starts conducting.

* The rise in current I_E of Q_2 causes a voltage across R_E (ie) $V_{RE} = I_E R_E$.

* This voltage provides a reverse bias across the emitter base junction of Q_1 and it is driven into cut off state.

* Since collector of Q_1 is coupled to base of Q_2 through resistor R_1 the forward bias for transistor Q_2 is increased.

* Q_2 is driven into saturation and

$$V_{C1} = V_{CC} \quad \text{and} \quad V_{C2} = V_{CE(sat)} + V_{RE}$$

* Consider an a.c input signal is applied to base of Q_1

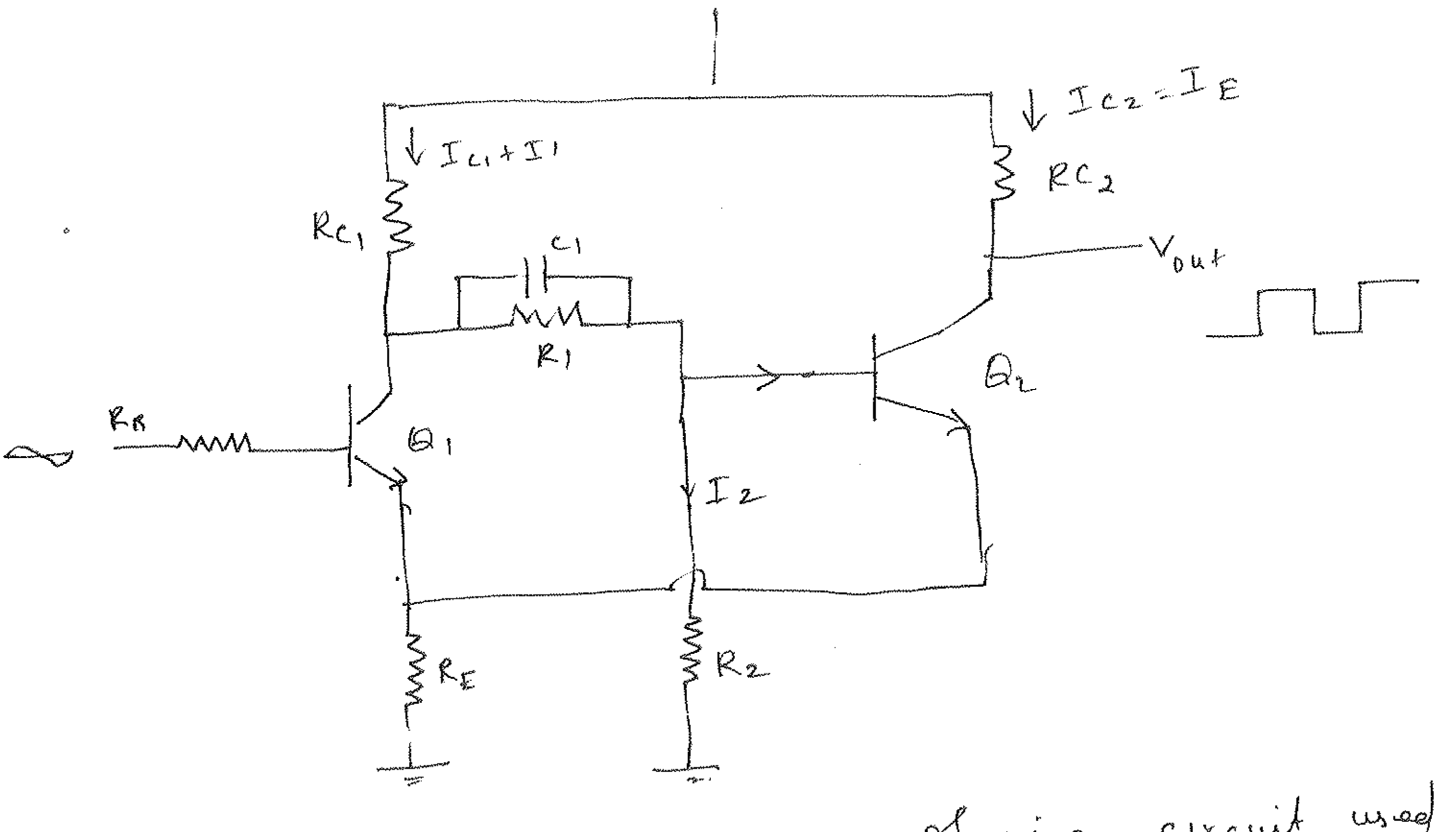
* When input increases above UTP (upper trigger point) ie $V_{in} < V_{RE} + V_{BE1}$ Q_1 conducts.

* As Q_1 conducts its collector voltage falls below V_{CC}, as collector Q_1 is coupled to base of Q_2.

of ON stage Q_2 via R_1C_1 connecting o/p of Q_1 to input of Q_2 which turns OFF

* After transition is over D_1 will be reverse biased and $D_2 \rightarrow$ zero bias
* The next triggering pulse will pass through D_2 instead of D_1
* D_1 and D_2 are called 'steering diodes'

Schmitt Trigger



* Schmitt trigger is a wave shaping circuit used for generation of square wave from sine wave
* It is a bistable circuit in which two transistor switches are connected regeneratively.

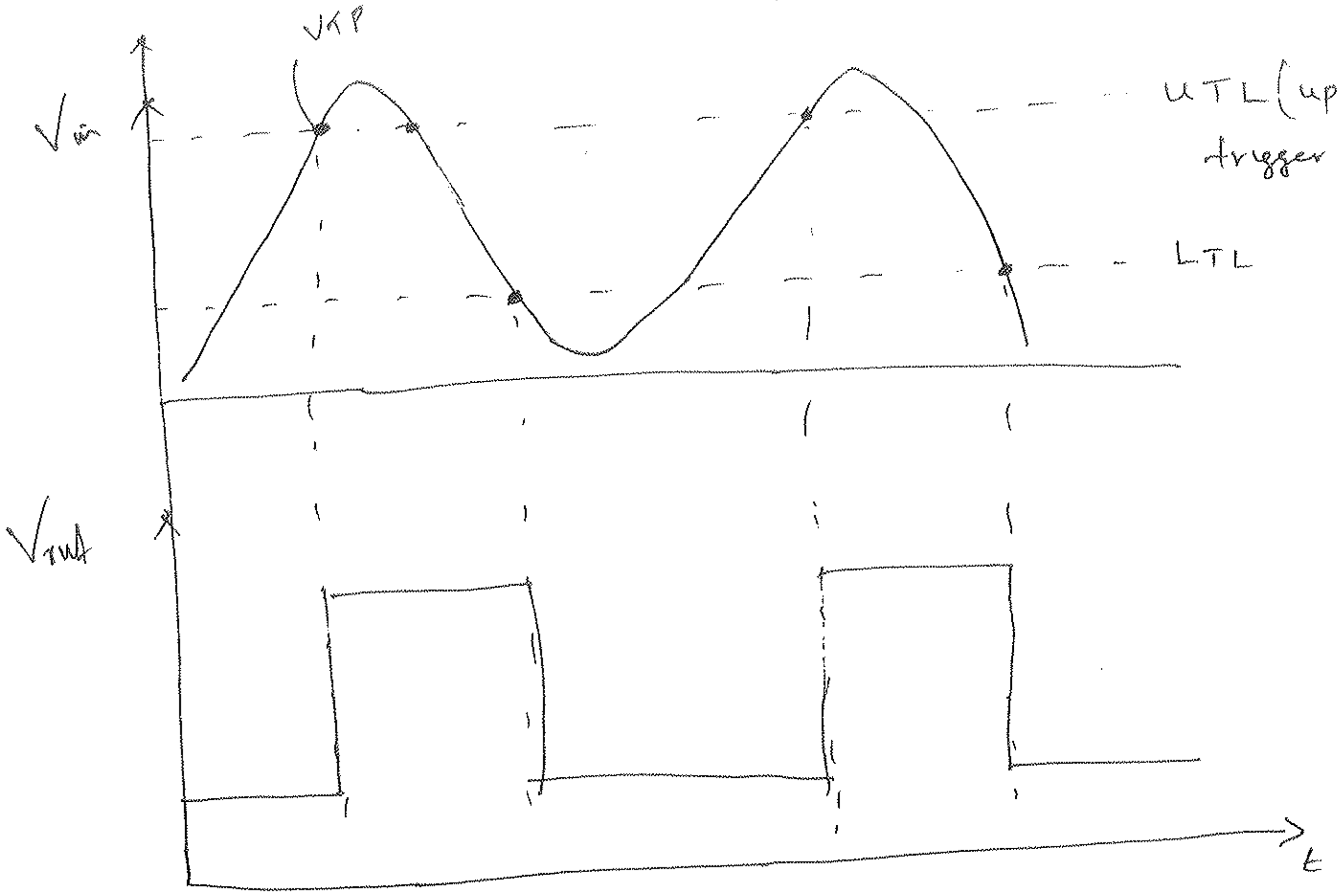
The forward bias of Q_2 is reduced

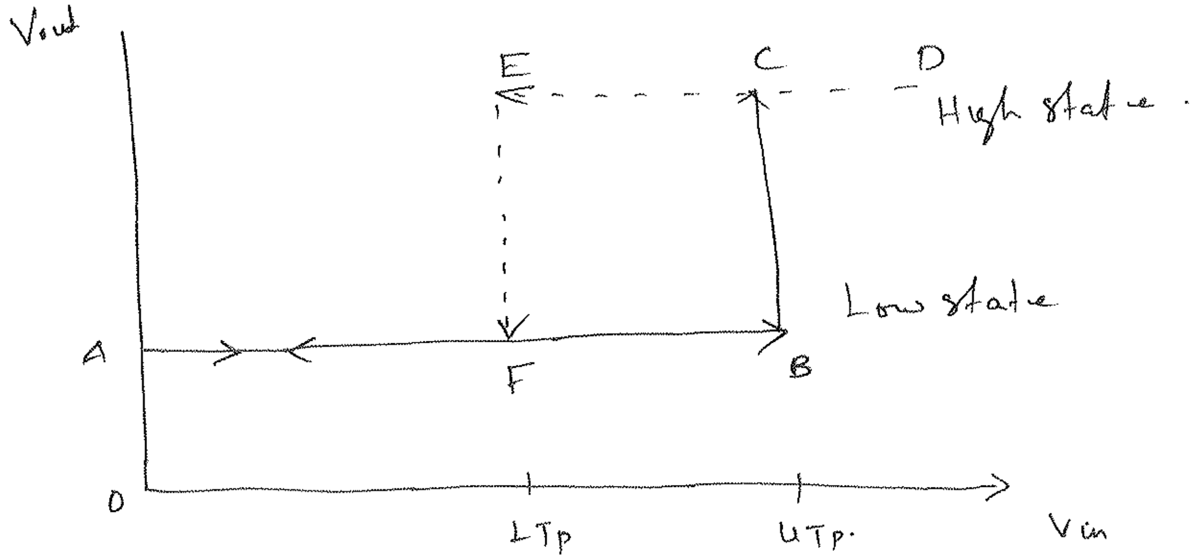
* This reduces the current of transistor Q_2 and voltage drop across R_E
* The reverse bias of Q_1 is reduced and it conducts more and drives Q_2 to cutoff.
* Q_1 continue to conduct till the input voltage crosses the lower Trigger level (LTL)
* At LTL $V_{in} < V_{RE} + V_{BE1}$ and V_{BE1} of Q_1 is reverse biased.
* Hence its collector voltage starts rising towards V_{cc}
* This forward bias Q_2. The point at which Q_2 starts conducting is lower trigger point (LTP)
* At this instant $V_{c1} = V_{cc}$, $V_{c2} = V_{CE(sat)} + V_{BE}$
* The difference between UTP & LTP is known as hysteresis voltage V_H.
* V_H is called "dead zone" of schmitt trigger.
* The lagging of lower threshold voltage from upper threshold is known as hysteresis.

Applications

1. it is used for wave shaping
2. it is used for generation of rectangular waveform with sharp edges from sine wave.

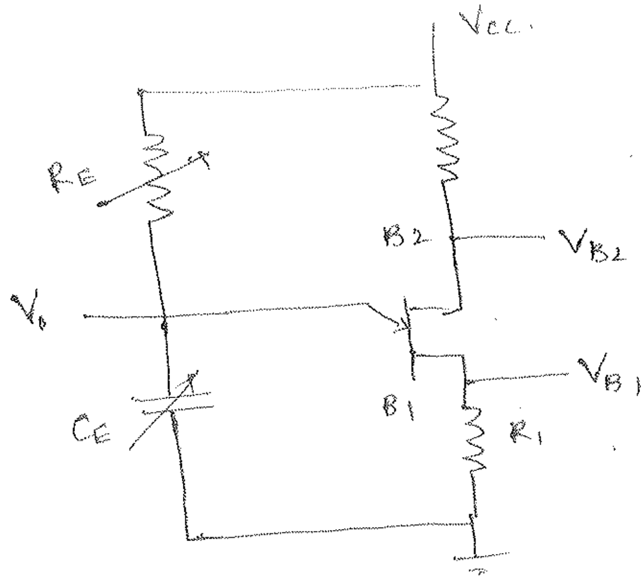
Input waveform





Hysteresis of Schmitt trigger

UJT as relaxation oscillator

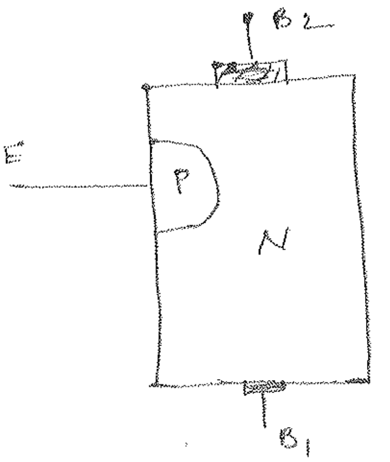


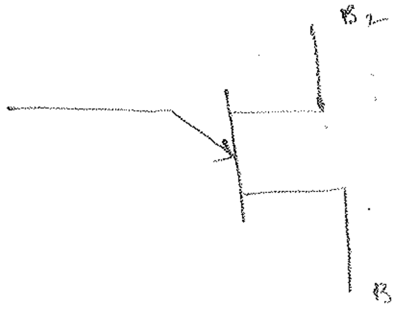
* It generates sawtooth waveform.
* it consists of UJT, with C_E charged through R_E
* When voltage across the capacitor increases exponentially and when voltage reaches pinch off voltage (V_p), UJT starts conducting
* The capacitor volt is discharged rapidly through Emitter, Base 1 and Resistor (R_1)
* After V_p it provides negative resistance to discharge path useful as relaxation oscillator.
* As capacitor voltage reaches zero, device is cutoff and C_E starts to charge again
* The cycle is repeated and sawtooth waveform is produced
* Resistors R_1 & R_2 series with B_1 & B_2 provides spikes.

UJT

* It is used to generate single or train of pulses using regenerative feedback characteristic

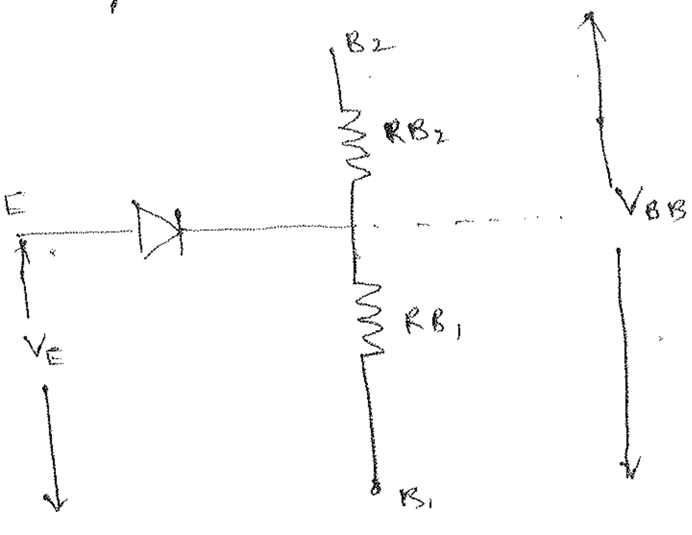
* It is a three terminal device





Symbol.

Equivalent circuit



* The Interbase resistance between B_2 and B_1 is

$$R_{BB} = R_{B1} + R_{B2}$$

* The voltage drop across $R_{B1} = V_1 = \eta V_{BB}$

$$\eta = \frac{R_{B1}}{R_{B1} + R_{B2}}$$

η is called intrinsic stand off ratio its value ranges from 0.56 to 0.75.

* When UJT fires sudden change in surge current through B_1 provides drop across R_1 which provides positive spikes.
* The fall of V_{EB_1} causes I_2 to decrease which gives negative spikes.
* By changing R_E and C_E the frequency is changed

<u>Frequency of oscillation</u>

Assuming 'C' is initially charged

$$V_C = V_{BB}\left(1 - e^{-t/R_E C_E}\right)$$

$R_E C_E$ is charging time of RC circuit.

* discharge of capacitor occurs when V_C is equal to peak point.

$$V_P = \eta V_{BB} = V_{BB}\left(1 - e^{-t/R_E C_E}\right)$$

$$\eta = \left(1 - e^{-t/R_E C_E}\right)$$

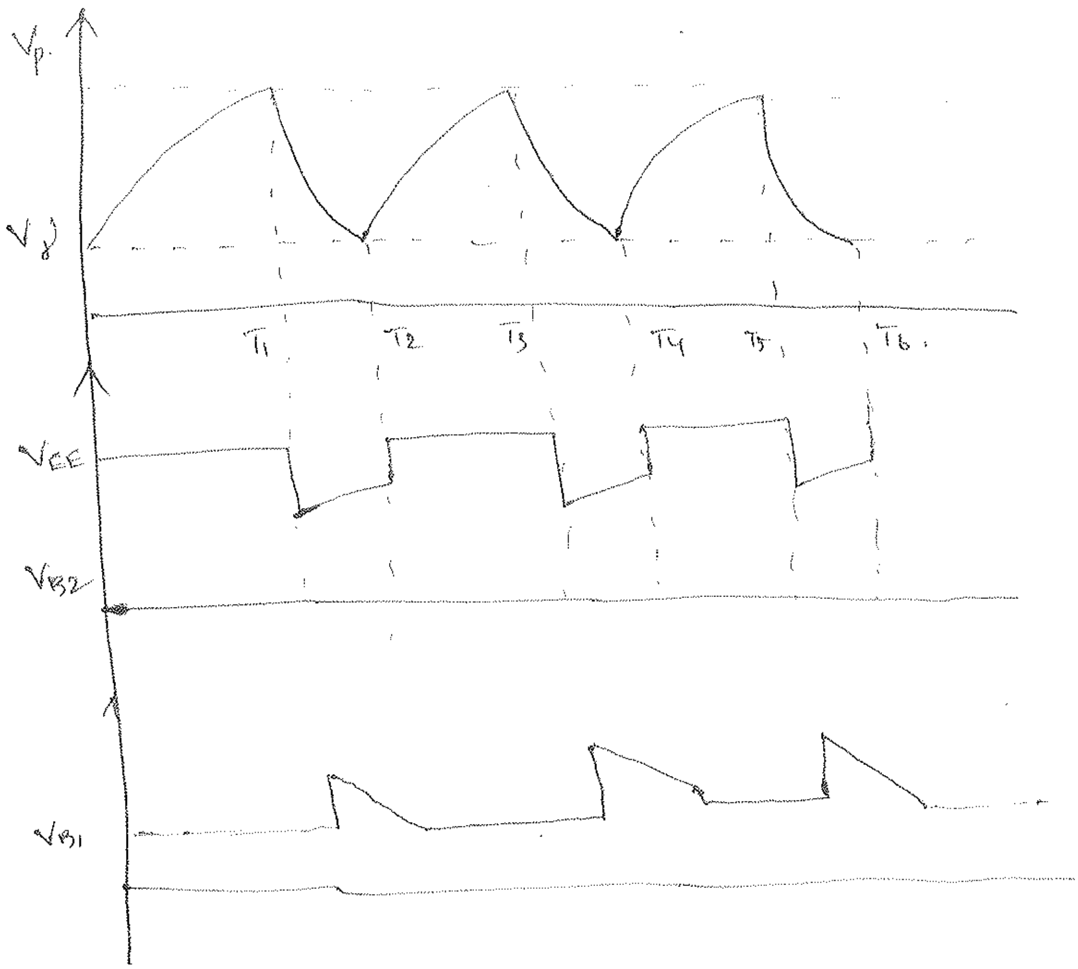
$$e^{-t/R_E C_E} = 1 - \eta$$

$$t = R_E C_E \times \log_e \frac{1}{1-\eta}$$

$$t = 2.303\, R_E C_E \log_{10} \frac{1}{1-\eta}$$

If discharge time is neglected, $t = T$, period of wave is

$$f_0 = \frac{1}{T} = \frac{1}{2.303\, R_E C_E \log_{10}\left(\frac{1}{1-\eta}\right)}$$

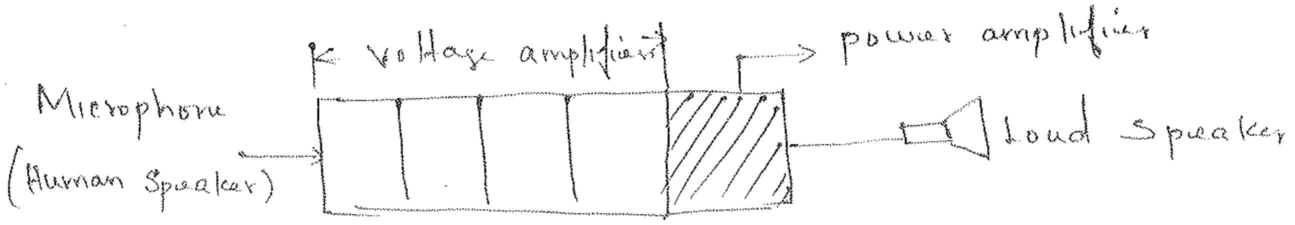


Vp.
Vγ
T1
T2
T3
T4
T5
T6
VEE
VB2
VB1

Unit - 5

power amplifiers and DC convertors.

power amplifiers



* In a public address system, there are many stages connected in cascade.
* The input and intermediate stages are small signal amplifiers.
* The last stage gives output to load like loud speaker.
* A stage which develops sufficient power to the load like loud speaker is called large signal amplifier or power amplifier.

power amplifiers finds application in public address systems, radio receivers, industrial control system tape players, T.V receivers and CRT.

Types

* The position of Q point on load line decides the class of operation of power amplifier
* The various classes of power amplifier are (i) class A (ii) class B (iii) class C and class AB.

class A amplifier

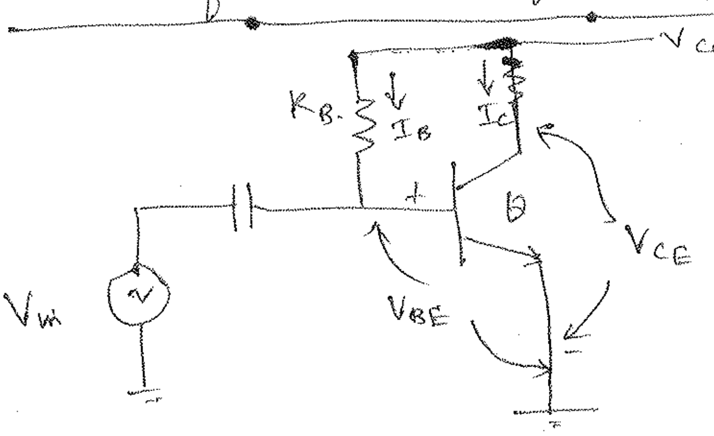
* In class A amplifier Q point is selected at the centre of load line.
* The output flows for full input cycle.
* The transistor remains in active region for full cycle.
* The collector current flows for 360°

Types of class A.

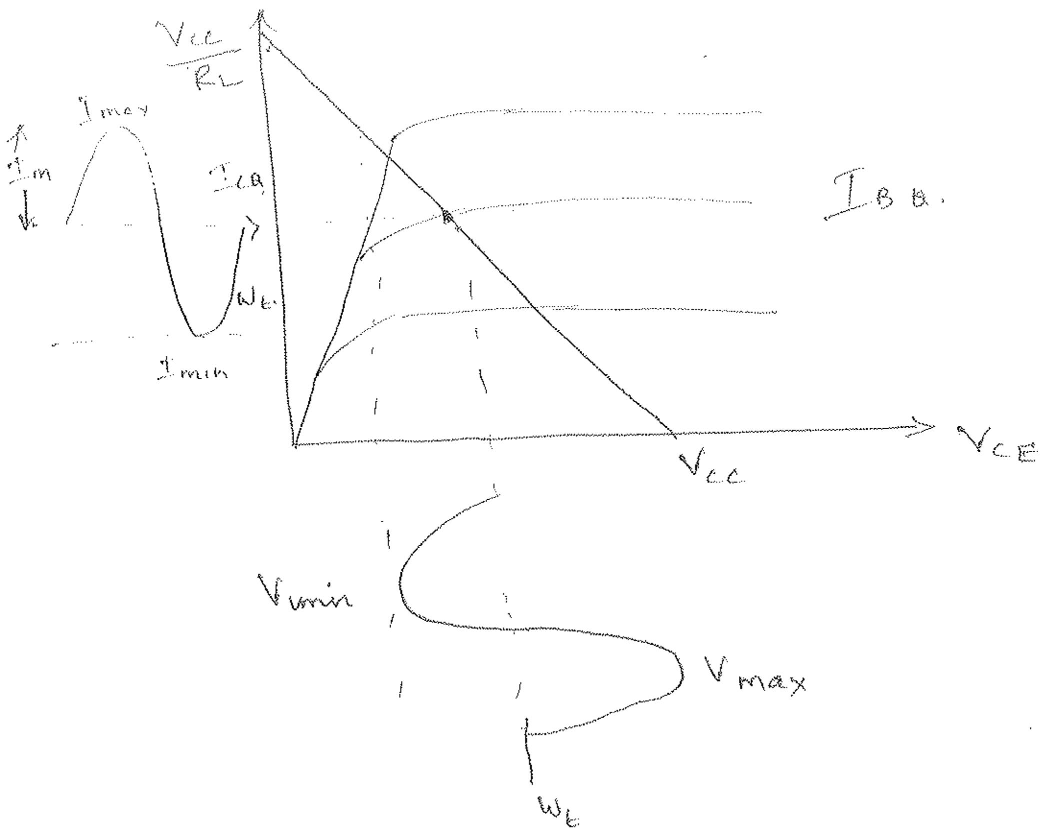
(i) Directly coupled class A amplifier (series fed)

(ii) Transformer coupled class A.

Series fed, Directly coupled class A.



* In directly coupled the load is directly connected in collector circuit.
* The transistor is used is power transistor
* The value of RB is selected such a way that Q pt lies in centre of d.c. load line.



Apply KVL to the circuit

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

$$I_C R_L = -V_{CE} + V_{CC}$$

$$I_C = \left[-\frac{1}{R_L}\right] V_{CE} + \frac{V_{CC}}{R_L}$$

The slope of load line is $-\frac{1}{R_L}$ and y intercept is $\frac{V_{CC}}{R_L}$

<u>DC operation</u>

* The collector supply voltage V_{CC} and resistance R_B divides dc bias current I_{BQ}.

Applying KVL to base emitter loop.

$$I_{BQ} = \frac{V_{CC} - 0.7}{R_B}$$

$$I_{CQ} = \beta I_{BQ}.$$

$V_{CEQ} = V_{CC} - I_{CQ} R_L$

Dc power input.

Dc power input is

$$P_{dc} = V_{CC} \cdot I_{CQ}.$$

Ac power output

$$V_m = \frac{V_{PP}}{2} = \frac{V_{max} - V_{min}}{2}$$

$V_m \rightarrow$ amplitude of ac output voltage.

$V_{max} \rightarrow$ maximum instantaneous value of collector voltage.

$V_{min} \rightarrow$ minimum "

$V_{PP} \rightarrow$ peak to peak value of a.c output voltage

$$P_{ac} = \frac{V_m I_m}{2} \cdot \frac{\frac{V_{PP}}{2} \times \frac{I_{PP}}{2}}{2}$$

$$P_{ac} = \frac{V_{PP} \times I_{PP}}{8}$$

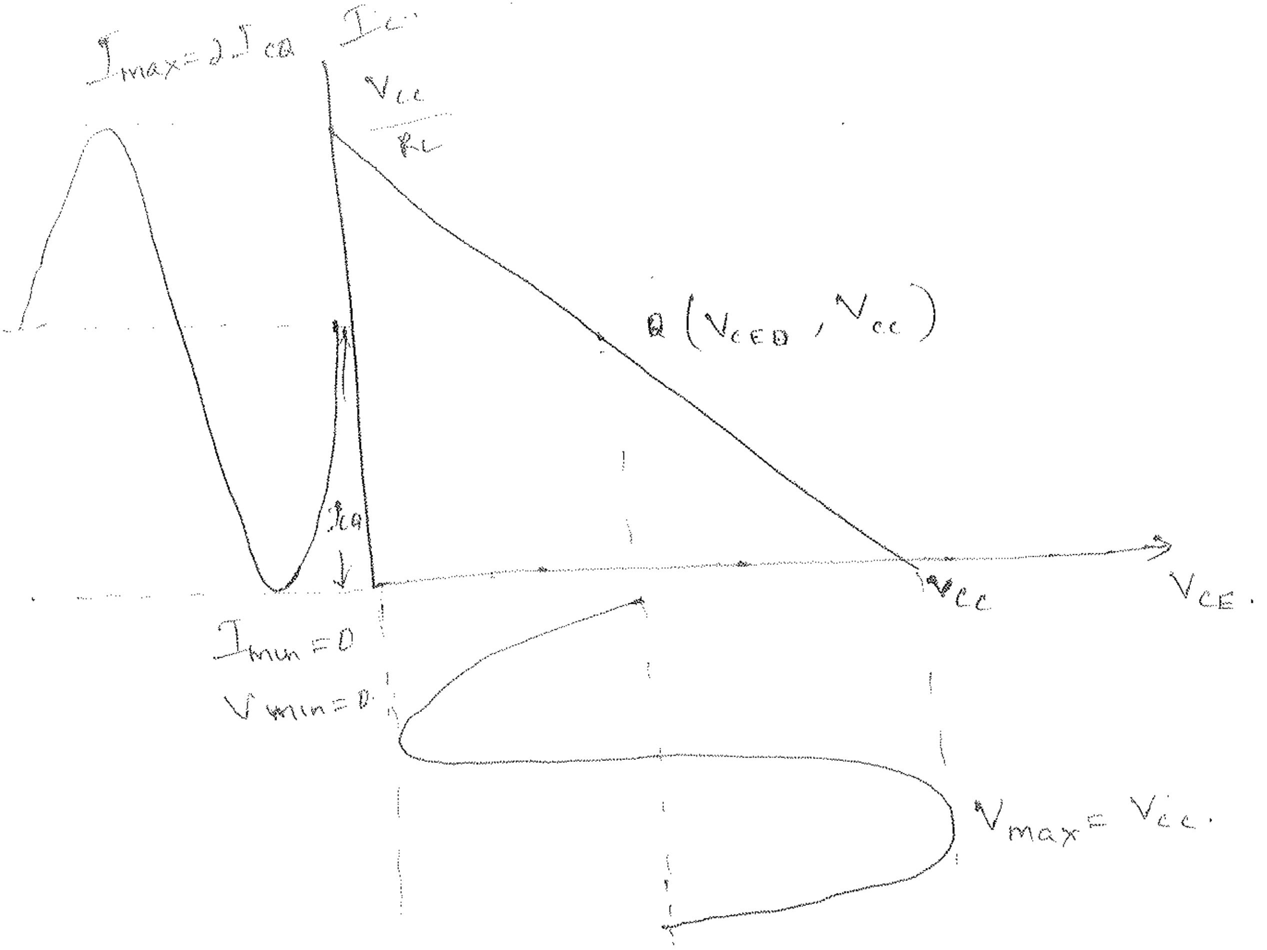
$$P_{ac} = \frac{I_{PP}^2 R_L}{8}$$

$$P_{ac} = \frac{V_{PP}^2}{8 R_L}$$

$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$$

$V_{max} = V_{CC}$ & $V_{min} = 0$

$I_{max} = 2 I_{CQ}$ & $I_{min} = 0$



$$\% \eta = \frac{(V_{CC} - 0)(2 I_{CQ} - 0)}{8 V_{CC} I_{CQ}} \times 100 = \frac{2 V_{CC} I_{CQ}}{8 V_{CC} I_{CQ}} \times 100$$

$= 25\%$

power dissipation

$P_d = P_{DC} - P_{ac}$

* maximum power dissipation occurs when there is no a.c input signal

$P_{d\,max} = V_{CC} \times I_{CQ}$.

Advantages:

① Circuit is simple to design

② load is connected directly no transformer is necessary

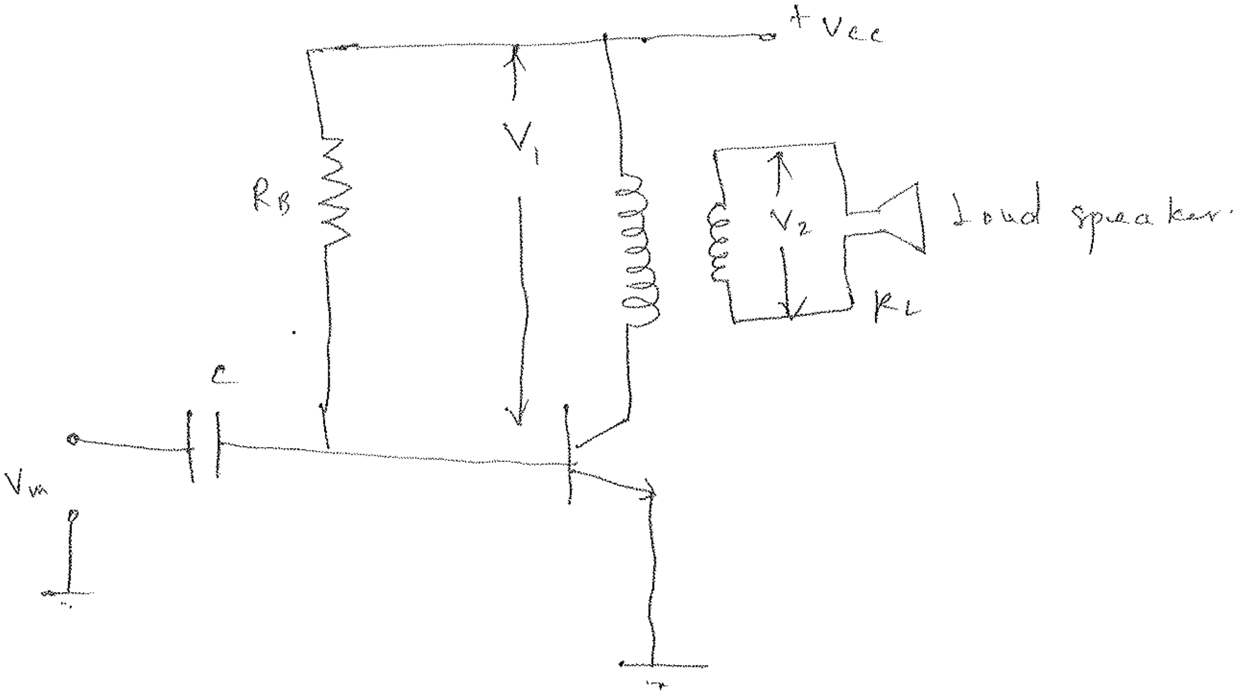
③ Less no. of components is required.

Disadvantages.

1. Since load resistance is directly connected in collector there is wastage of power.
2. power dissipation is more
3. The output impedance is high.
4. The efficiency is very poor.

Transformer coupled class A amplifier

* For maximum power transfer to load, impedance matching is necessary
* The loud speaker has low output impedance which has to be matched with class A amplifier of high output impedance.
* This is eliminated by using transformer to deliver power to load.



Dc operation

Apply KVL to collector circuit

$$V_{cc} - V_{CE} = 0$$

$$V_{cc} = V_{CE}$$

$$\boxed{V_{CE\,Q} = V_{cc}}$$

Dc operation power input

$$P_{DC} = V_{cc} I_{CQ}.$$

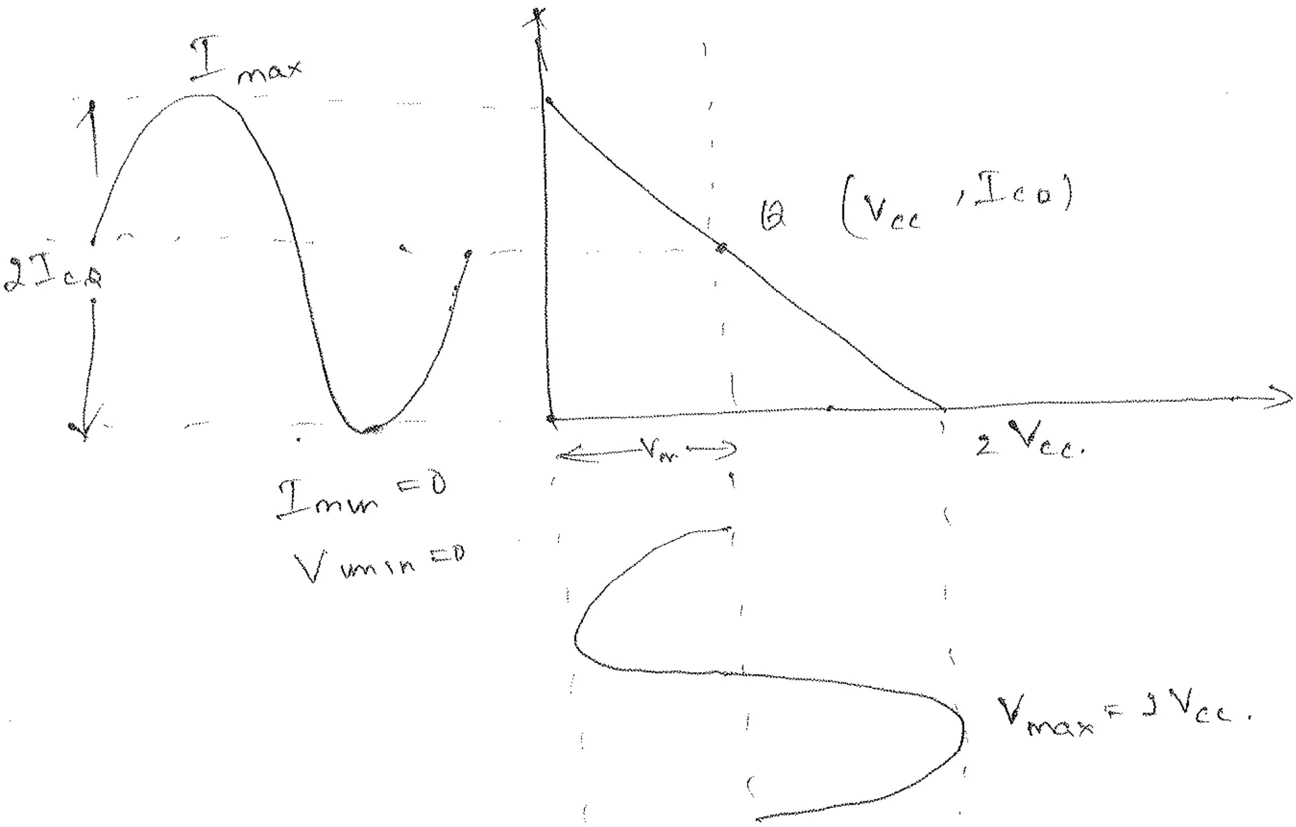
Ac output power

$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$$

Efficiency

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

Maximum Efficiency



$V_{min} = 0$ and $V_{max} = 2 V_{cc}$

$I_{min} = 0$ and $I_{max} = 2 I_{CQ}$.

$$\% \eta_{max} = \frac{(2 V_{cc} - 0)(2 I_{CQ} - 0)}{8 V_{cc} I_{CQ}} \times 100$$

$$= \frac{4 V_{cc} I_{CQ}}{8 V_{cc} I_{CQ}} \times 100 = 50\%$$

power dissipation

$$P_d = P_{DC} - P_{ac}$$

* When input signal is larger more power is delivered to load and less is power dissipation

* When there is no input signal the entire dc i/p power gets dissipated in the form of heat.

$$P_{d\,max} = V_{cc} \times I_{CQ}.$$

Advantages

1. The efficiency is higher
2. Impedance matching is possible

Disadvantages

1. Due to transformer, the circuit is bulkier
2. The circuit is complicated
3. Frequency response is poor.

Analysis of class B.

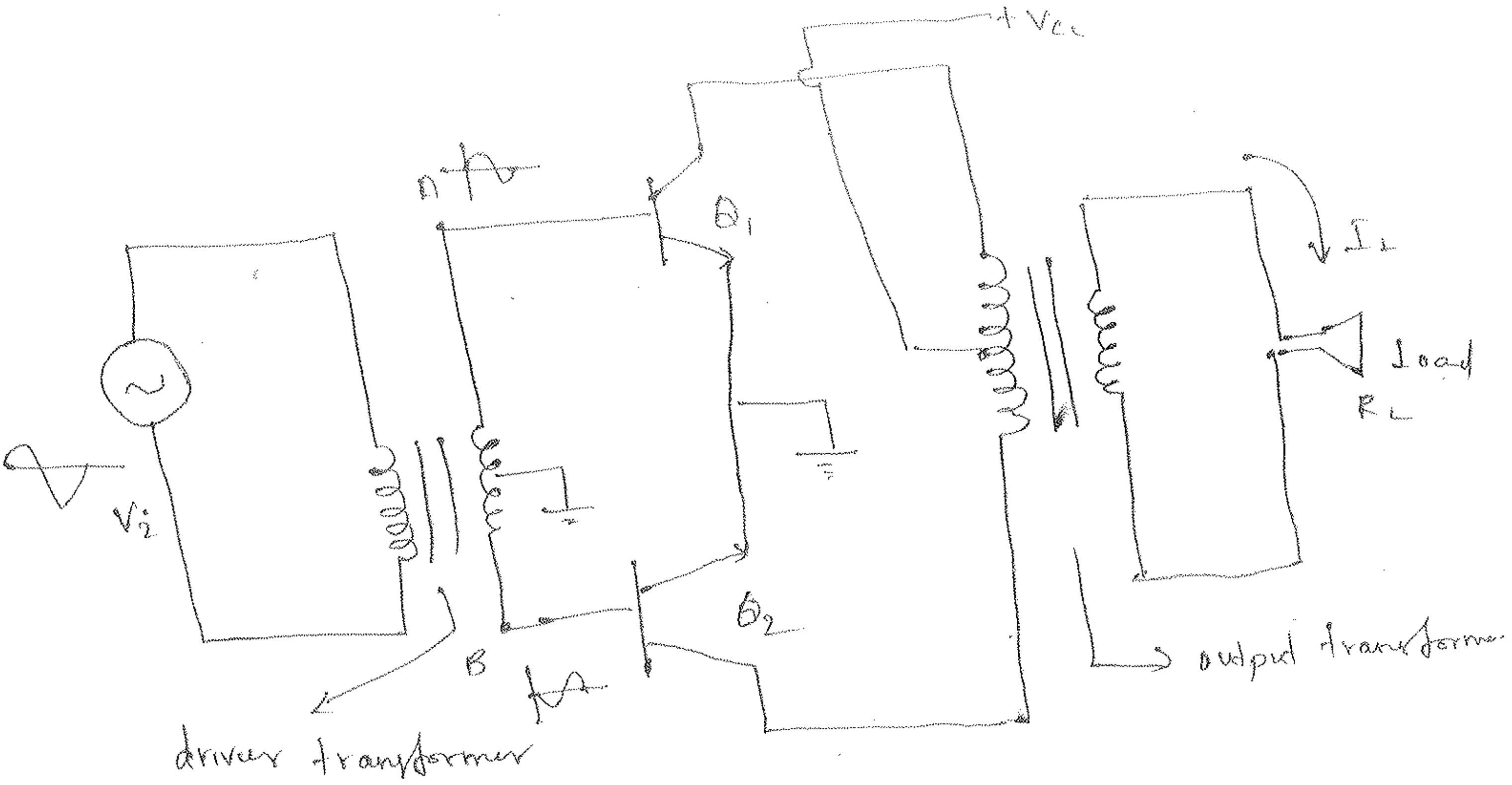
* The Q point is located on x-axis itself
* The collector current flows only for half cycle
* To get full cycle across load, two transistors conduct in alternate half cycle.

Two types of class B

1. class B push pull
2. complementary symmetry class AB.

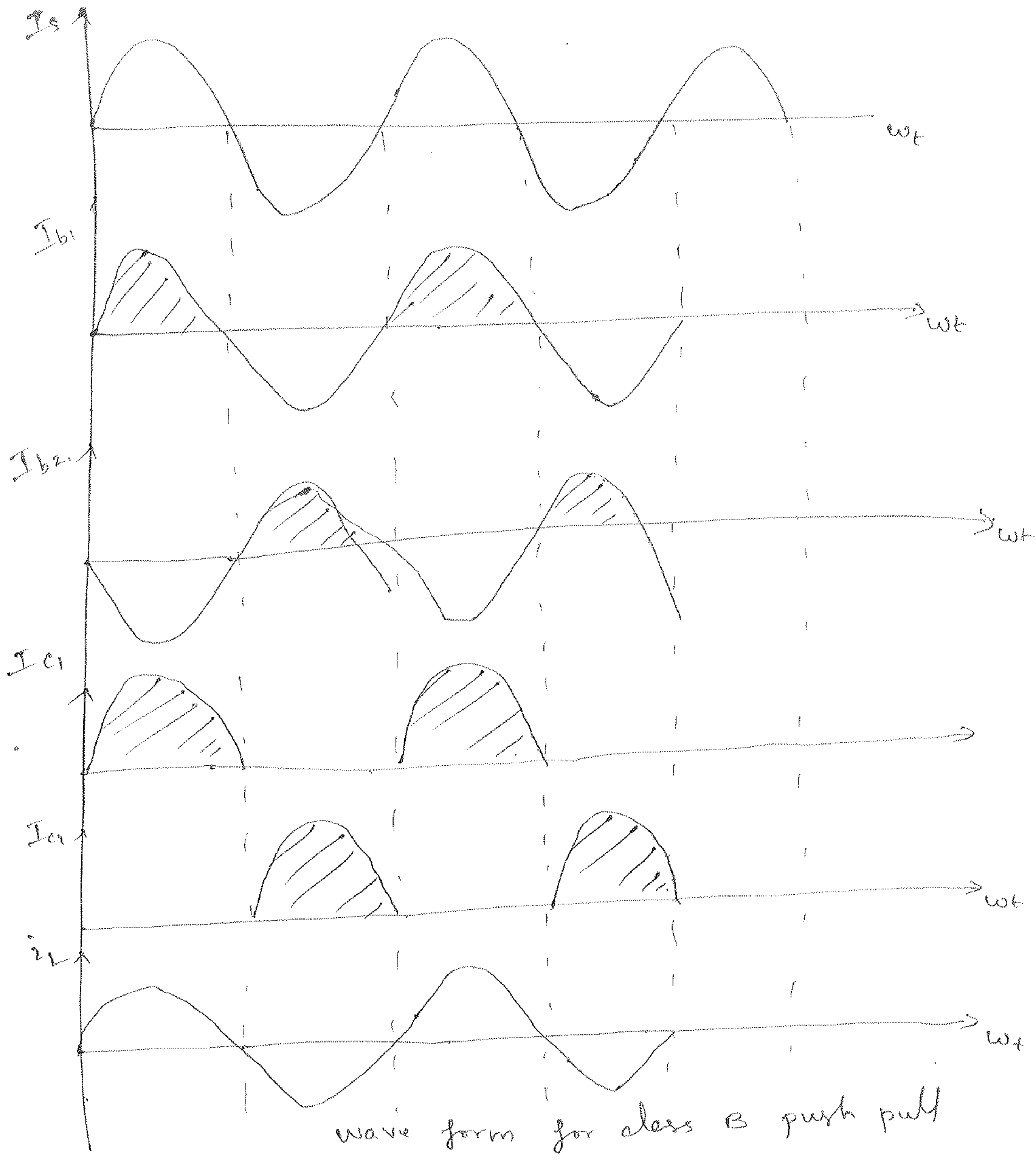
class B push pull

* It has two transformers, input and output transformer.
* It has two transistors both of same type either npn or pnp.
* The input signal is applied to primary of driver transformer.
* The centre tap on secondary of driver transformer is grounded.
* The voltage of secondary of driver transformer is equal with opposite polarity
* The input to base of Q_1 and Q_2 will be 180° out of phase.



<u>Operation</u>

* For positive half cycle of input A is positive and B is negative
* Transistor Q1 conducts and Q2 is cut off.
* So i_{b1} flows and $i_{b2}=0$ and i_1 flows for upper part of primary.
* For negative half cycle, B is positive and A is negative.
* Transistor Q2 conducts and Q1 is cut off.
* So i_{b2} flows and $V_{b1}=0$ and i_{22} flows through lower part of primary.
* Thus full cycle is obtained across the load.



wave form for class B push pull

<u>Dc power input.</u>

$$P_{dc} = V_{cc} \times I_{dc}$$

$$I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi} = \frac{2 I_m}{\pi} \quad (\because \text{two transistor})$$

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

Ac power output

$$P_{ac} = \frac{V_m I_m}{2}$$

Efficiency

$$\% \eta = \frac{P_{ac}}{P_{dc}} \times 100 = \frac{\left(\frac{V_m I_m}{2}\right)}{\frac{2}{\pi} V_{cc} \times I_m} \times 100$$

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

Max efficiency

Maximum value of V_m is possible if

$$V_m = V_{cc}$$

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

$$\% \eta = 78.5\%$$

power dissipation

$V_m = \frac{2}{\pi} V_{cc}$ - for max. power dissipation

$$P_d = P_{dc} - P_{ac} = \frac{4}{\pi^2} \cdot \frac{V_{cc}^2}{R_L'} - \frac{2}{\pi^2} \cdot \frac{V_{cc}^2}{R_L'}$$

$$P_{d\,max} = \frac{2}{\pi^2} \cdot \frac{V_{cc}^2}{R_L'}$$

$$\boxed{P_{d\,max} = \frac{2}{\pi^2} \cdot \frac{V_{cc}^2}{R_L'}}$$

Advantages:

① Efficiency is much higher than class A

② Due to transformer impedance matching is possible.

Disadv

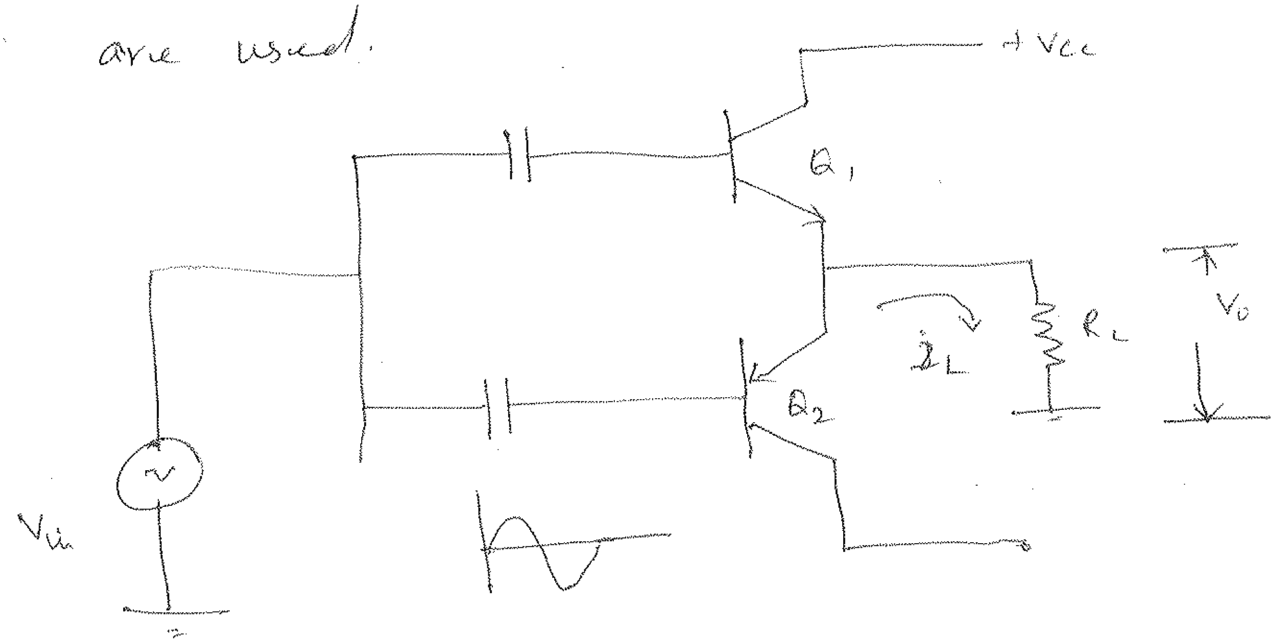
① Centre tap transformer are necessary.

② Transformer make the circuit bulky and costlier

③ Frequency response is poor.

Complementary Symmetry class B. Amplifier

* Instead of same type of transistor (npn (or) pnp) one npn and one pnp is used.
* So this circuit is transformer less circuit
* It is difficult to match the output impedance
* Hence matched pair of complementry transistors are used.



* The circuit is driven from dual supply of $+V_{cc}$
* Q_1 is npn & Q_2 is pnp.
* During positive half cycle Q_1 is driven to active and Q_2 is cut off.
* So current flows through R_L
* During negative half cycle Q_2 conducts and Q is cut off.
* Hence Q_2 conducts during negative half cycle of input and current flows through R_L.

Analysis is same as class B pushpull.

Advantages:

(1) As circuit is transformer less, weight, size and cost are less.

(2) Due to common collector impedance matching is possible.

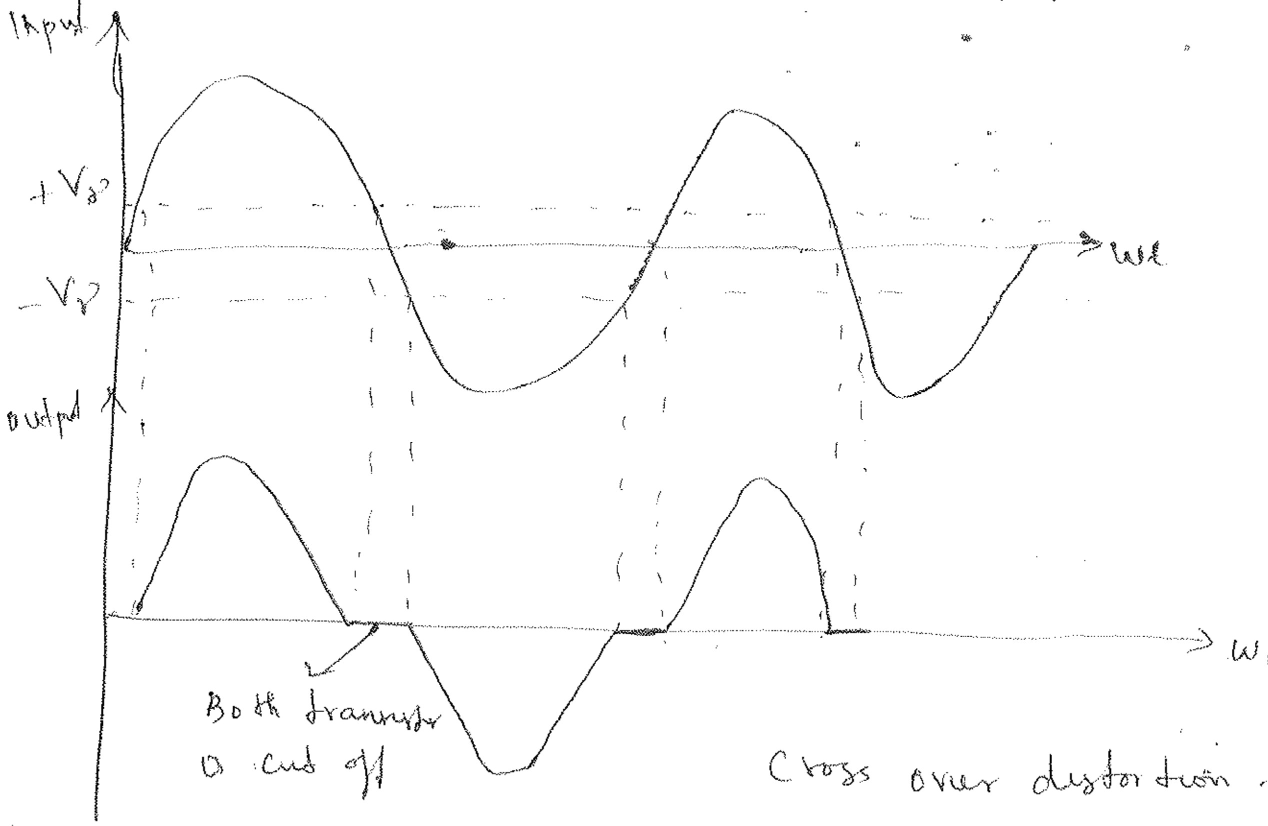
Disadvantages.

(1) It needs two separate voltage supplies

(2) Output is distorted to cross over distortion.

Cross over distortion

* For transistor to be in active region the base emitter junction must be forward bias.
* The junction is made forward bias till the voltage applied becomes greater than cut in voltage (0.7V) for si.
* When magnitude of input is less than 0.7 the collector current remains zero and transistor remains in cutoff.
* There is a period between crossing of half cycles, none of transistor conducts and output is zero.
* Hence the nature of output is distorted and do not remains same.
* Such a distortion is called cross over distortion.



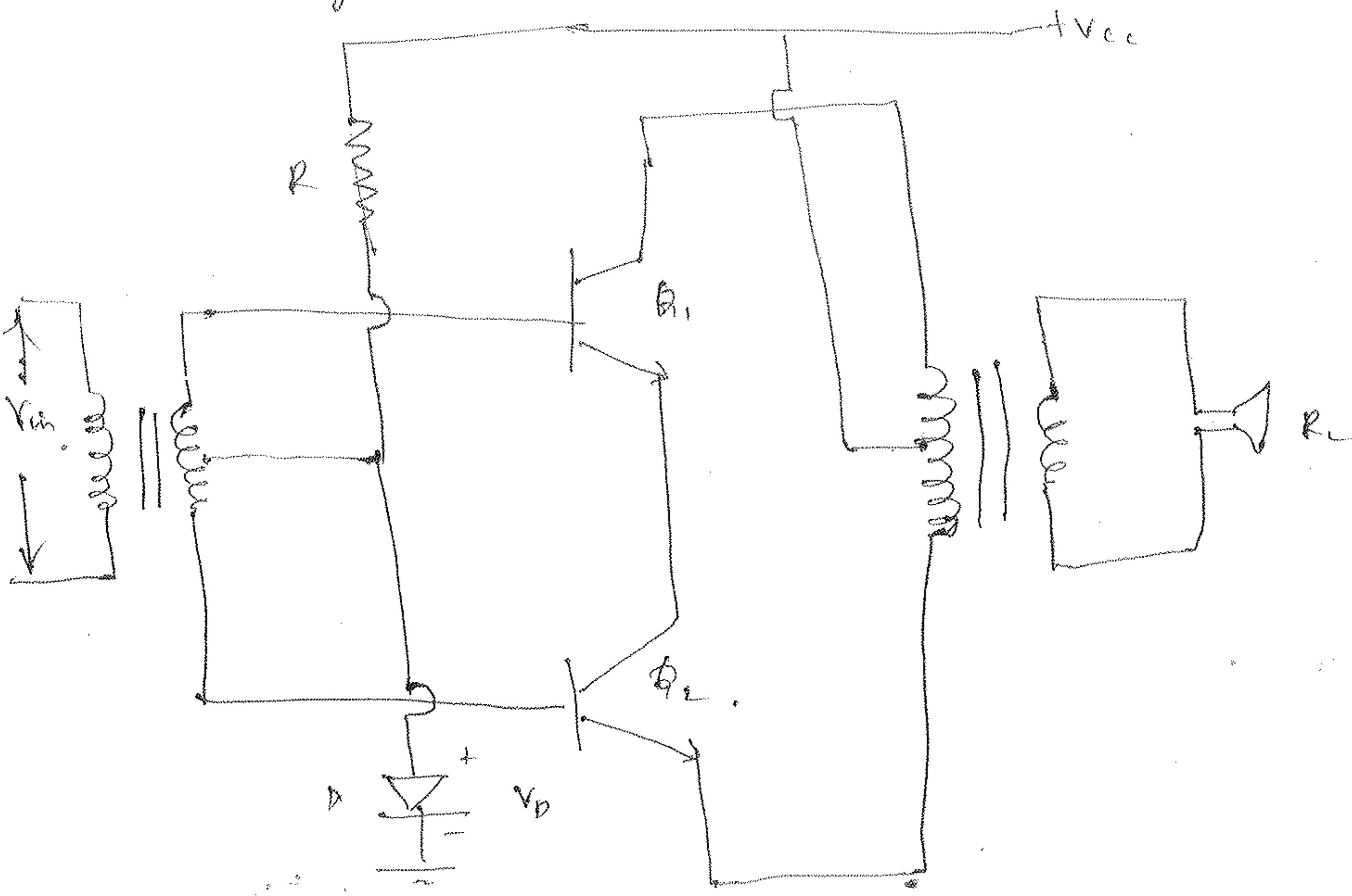
Cross over distortion.

Elimination of cross over distortion

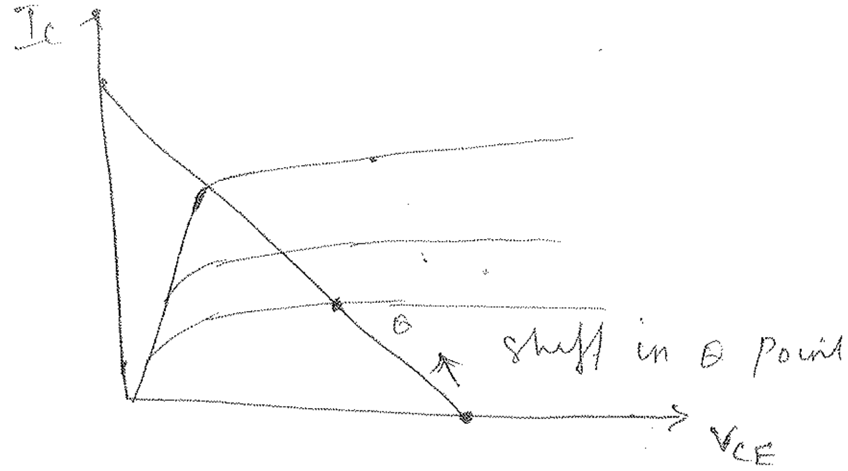
To eliminate cross over distortion a small forward bias is applied to transistor

push pull class AB

* The forward bias across the transistor is provided by using diode.

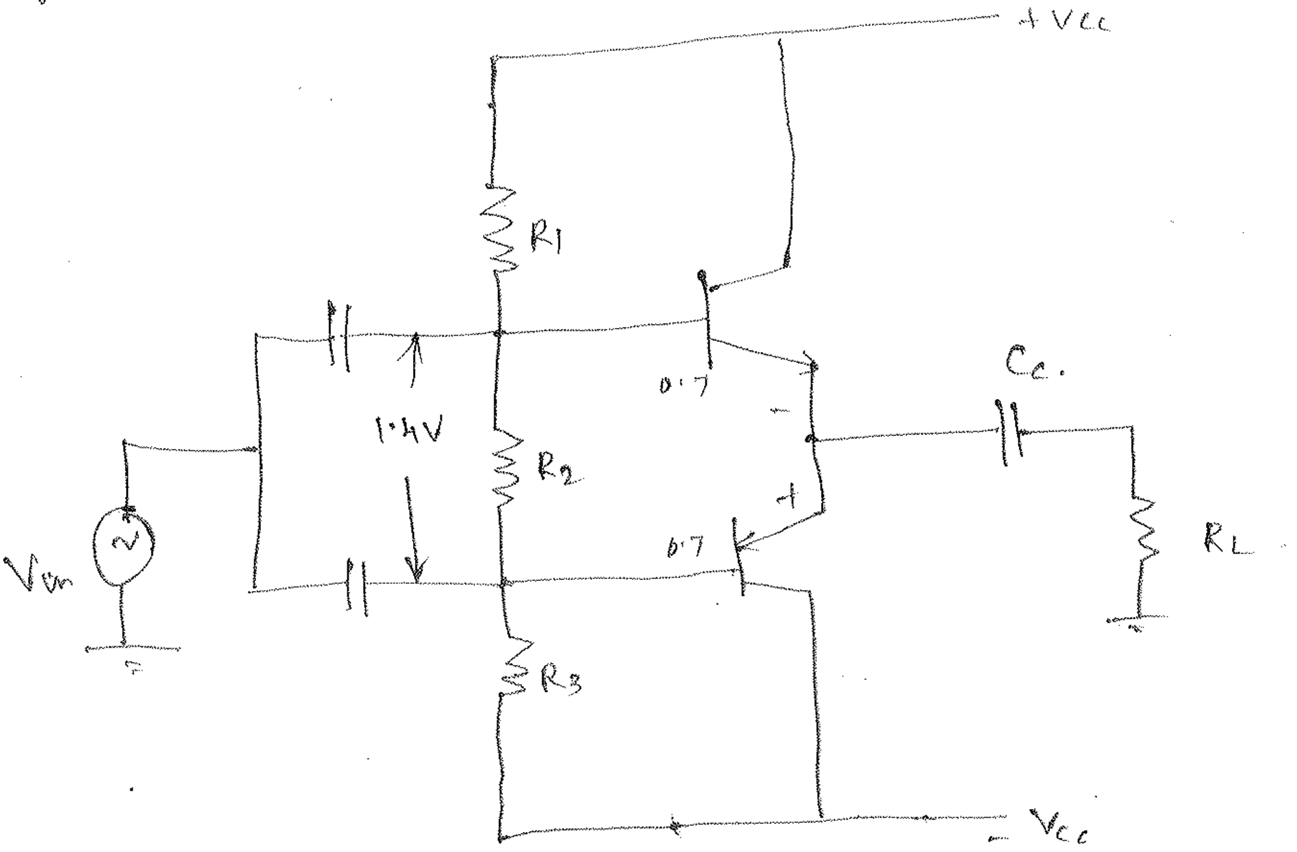


* The drop across diode D is equal to cut in voltage of base emitter junction of transistor.
* Hence both transistor conducts for full cycle
* The Q point shifts upward and the operation becomes class AB.



Complementary symmetry class AB.

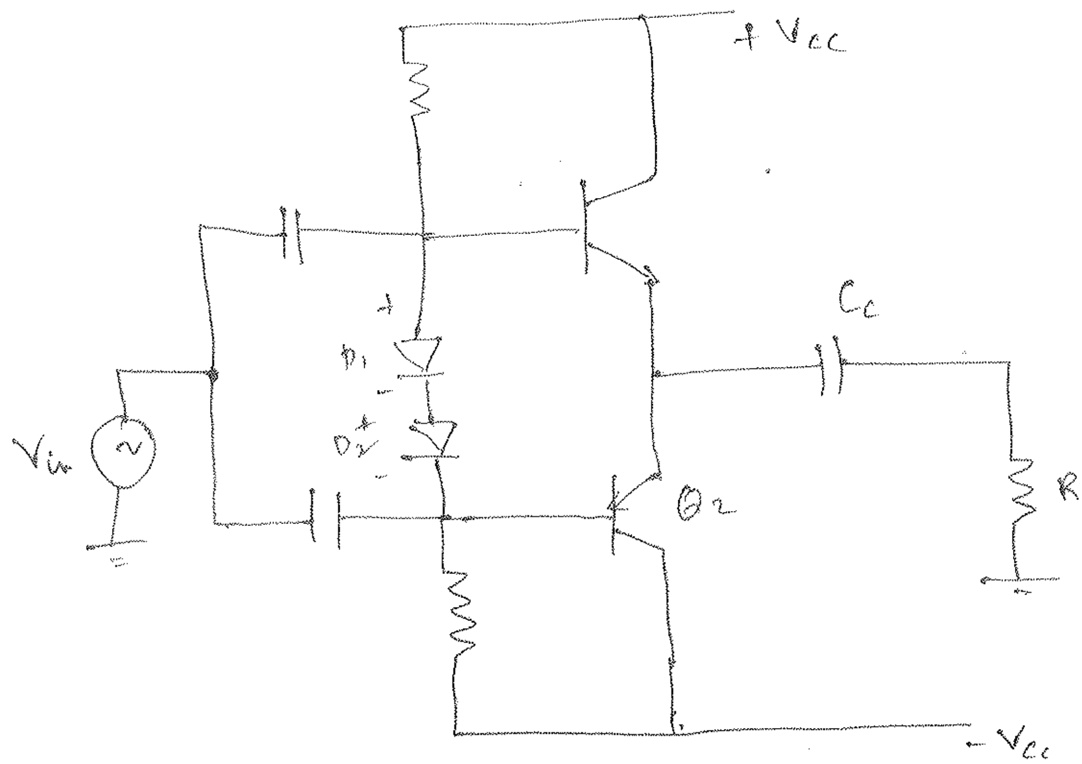
* In complementary symmetry base emitter junction of both Q_1 and Q_2 are required to provide fixed bias



* For silicon transistor fixed bias of $0.7 + 0.7 = 1.4V$ is required
* This is achieved by using potential divider arrangement
* As the junction cut in voltage changes with temperature there is still possibility of

distortion as temperature changes.

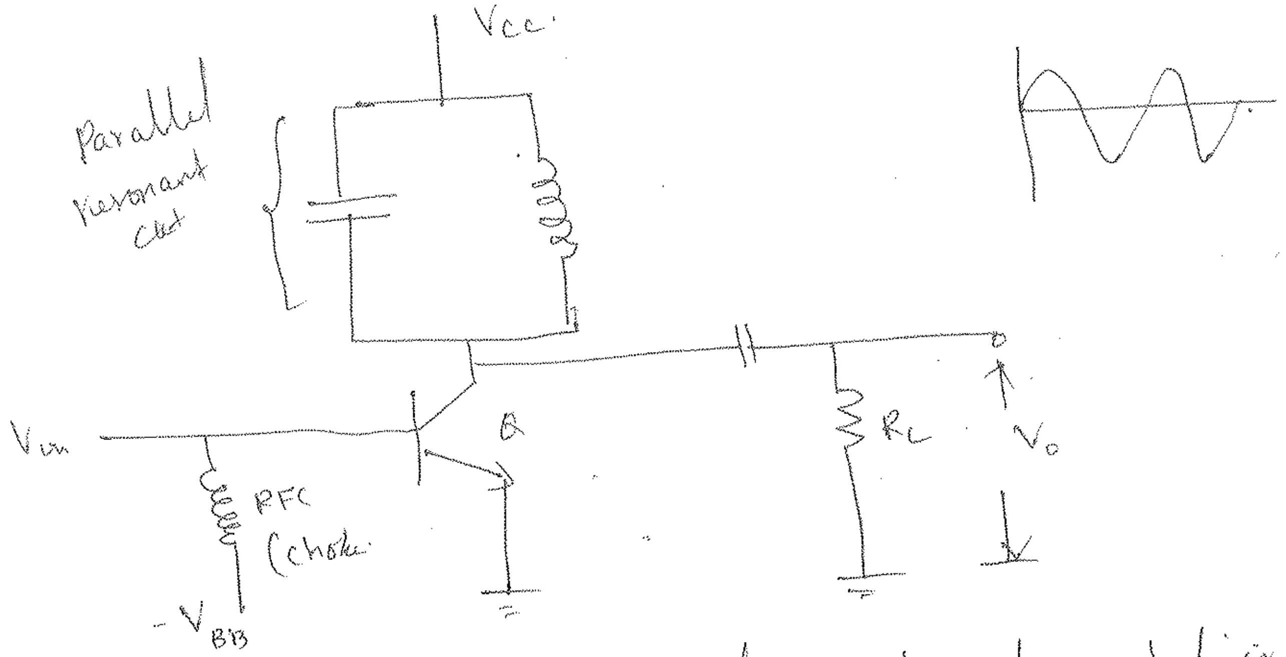
* Hence instead of R_2 two diodes is used to provide required fixed bias.



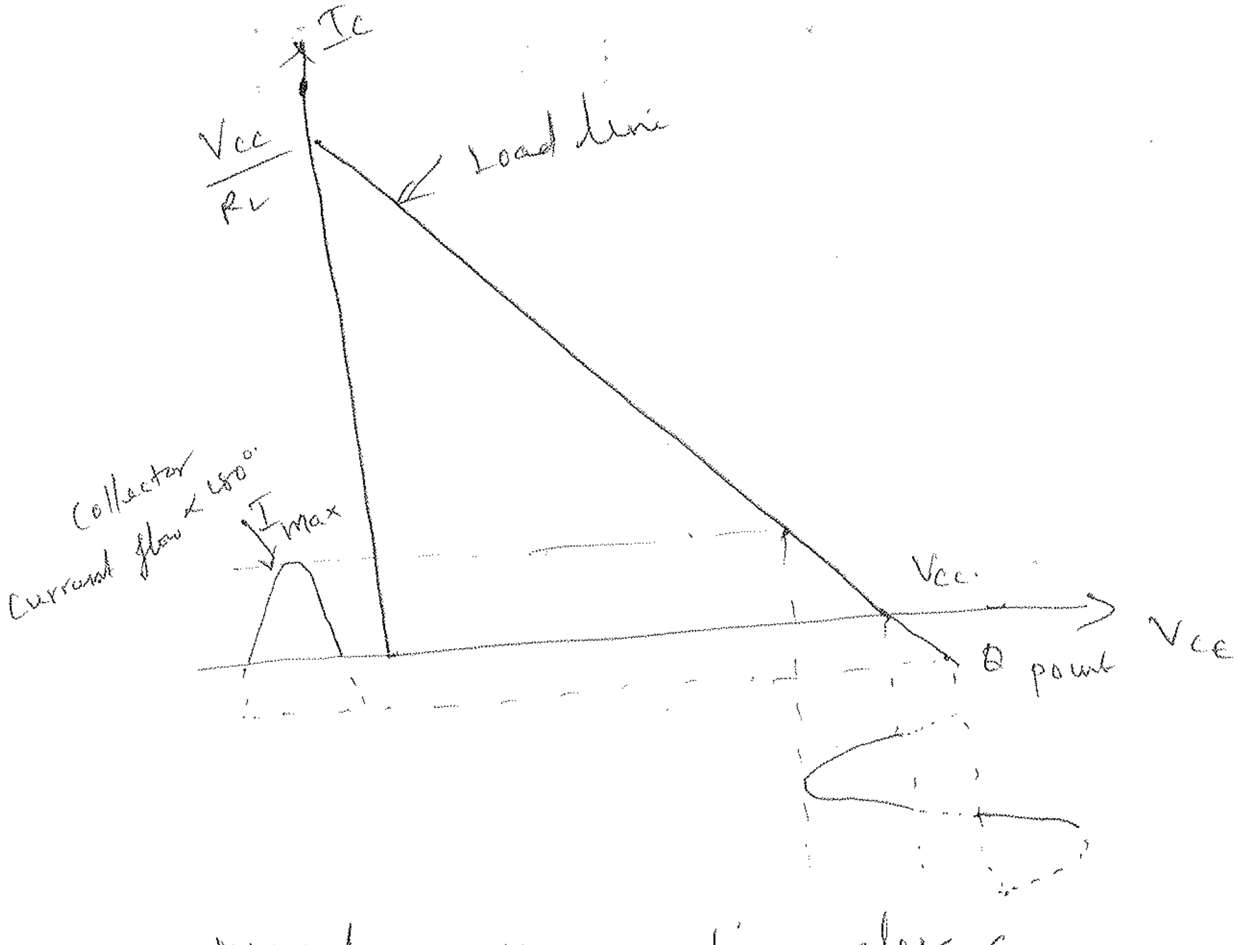
* As the temperature changes with junction characteristics the diode gets changed and maintain necessary biasing to overcome distortion.

class C operation

* In class C, resonating circuit is used as load. So most of class C amplifiers are tuned amplifiers



class C tuned amplifier

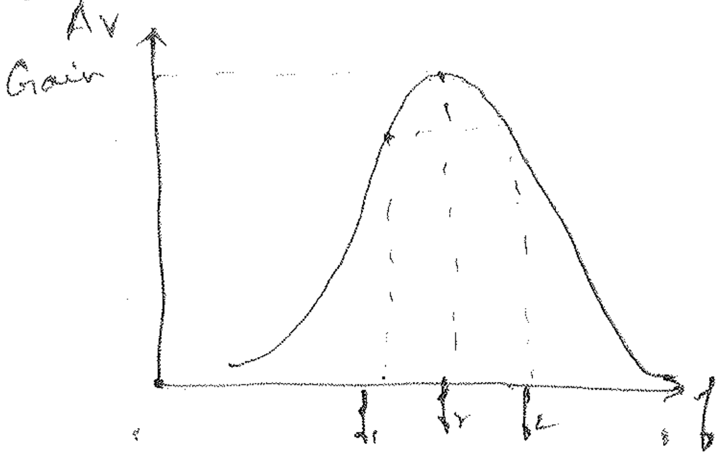


waveform representing class C.

* A parallel resonant circuit acts as load impedance
* The collector current flows for less than half a cycle and consists of series of pulses with harmonics
* The resonant frequency is given by

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

* The output voltage is maximum at resonant frequency. The gain drops on either side



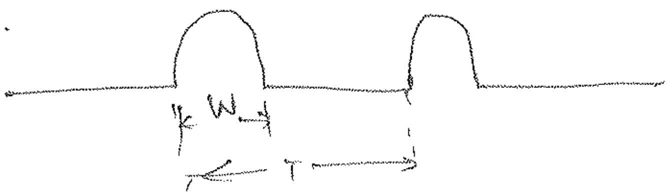
Bandwidth

$$BW = \frac{f_r}{Q}.$$

$$B.W = f_2 - f_1$$

Q - Quality factor.

Duty cycle.

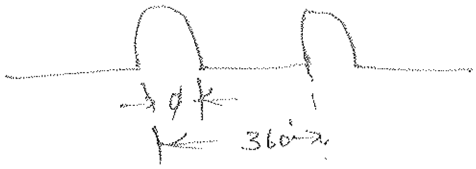


$$D = \frac{W}{T}$$

W - width of pulse, T - period of pulse.

* interms of conduction angle

$$D = \frac{\phi}{360^\circ}$$



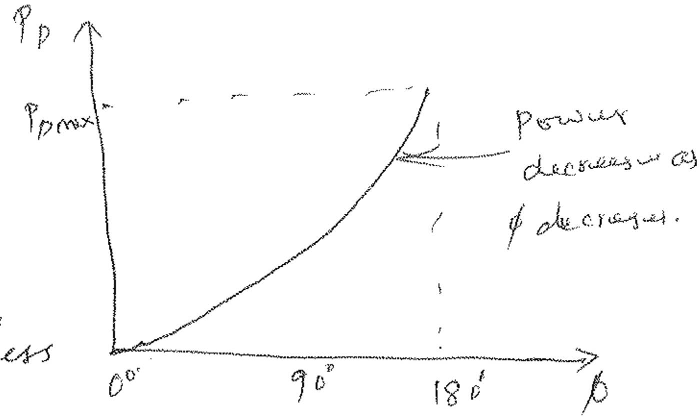
<u>Output power</u>

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

$$V_{pp} = 2V_m = 2\sqrt{2}\,V_{rms}$$

$$P_{out} = \frac{(V_{pp}/2\sqrt{2})^2}{R_L} = \frac{V_{pp}^2}{8R_L}$$

<u>Transistor dissipation</u>

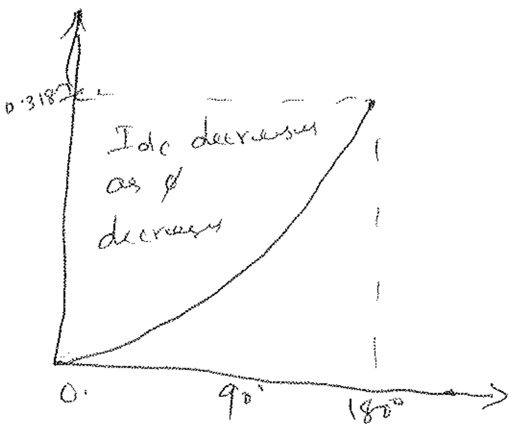


$$P_D\,(max) = \frac{V_{pp\,max}^2}{40\,V_c}$$

* power dissipation depends on
. Conduction angle φ
* less φ, less dc power, and less
is transistor dissipation

<u>Dc input power</u>

$$I_{dc} = \frac{I_c(sat)}{\pi} = 0.318\, I_c\,(sat)$$



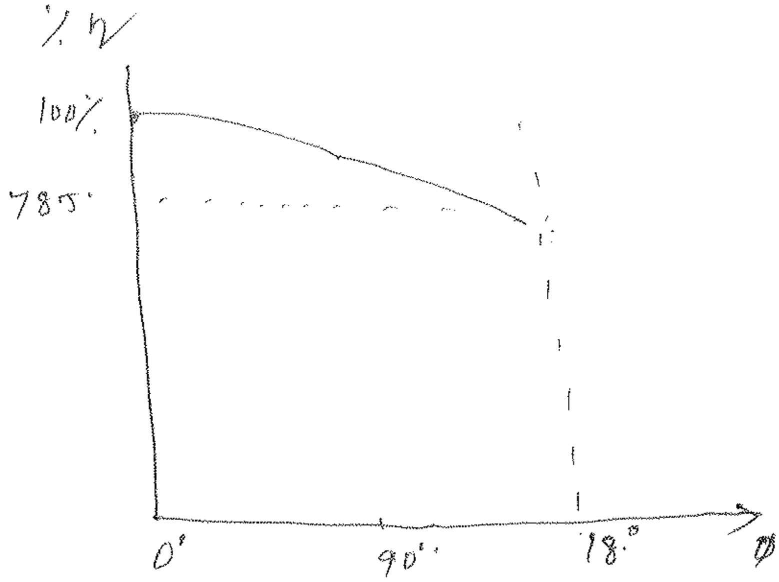
If the conduction angle is 180°,
current is half wave of
rectified waveform.

Efficiency

* Efficiency is given by ratio of a.c power output to dc power input.

$$\%\eta = \frac{P_{out}}{P_{DC}} \times 100 = \frac{P_{out}}{V_{cc} \times I_{dc}} \times 100.$$

100% is efficiency achieved at very small conduction angle.



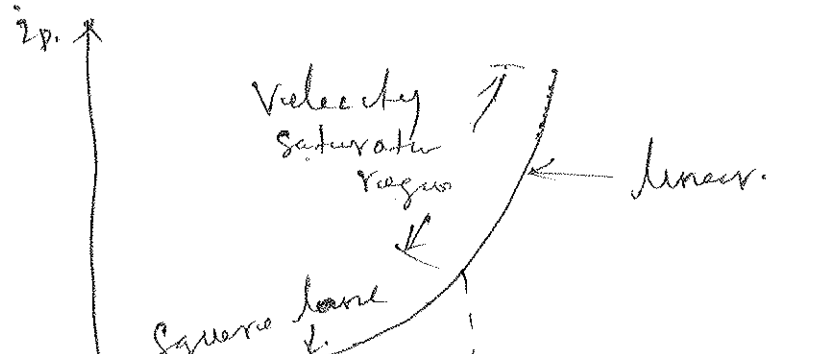
Power MOSFET

* The operation of power MOSFET is same as conventional MOSFET but power handling of conventional MOSFET is less than 1W.

Features of power MOSFET

1. power handling is more than 100 W
2. Current handling is in ampere range.
3. Large forward conductance.
4. As input impedance is very high, large currents can be switched with very small control currents.

Characteristics of power MOSFET



* Power MOSFET have threshold voltage in range of 2V to 4V.
* In saturation the drain current is related to V_{GS} by square law characteristics
* The linear portion of characteristics is as a

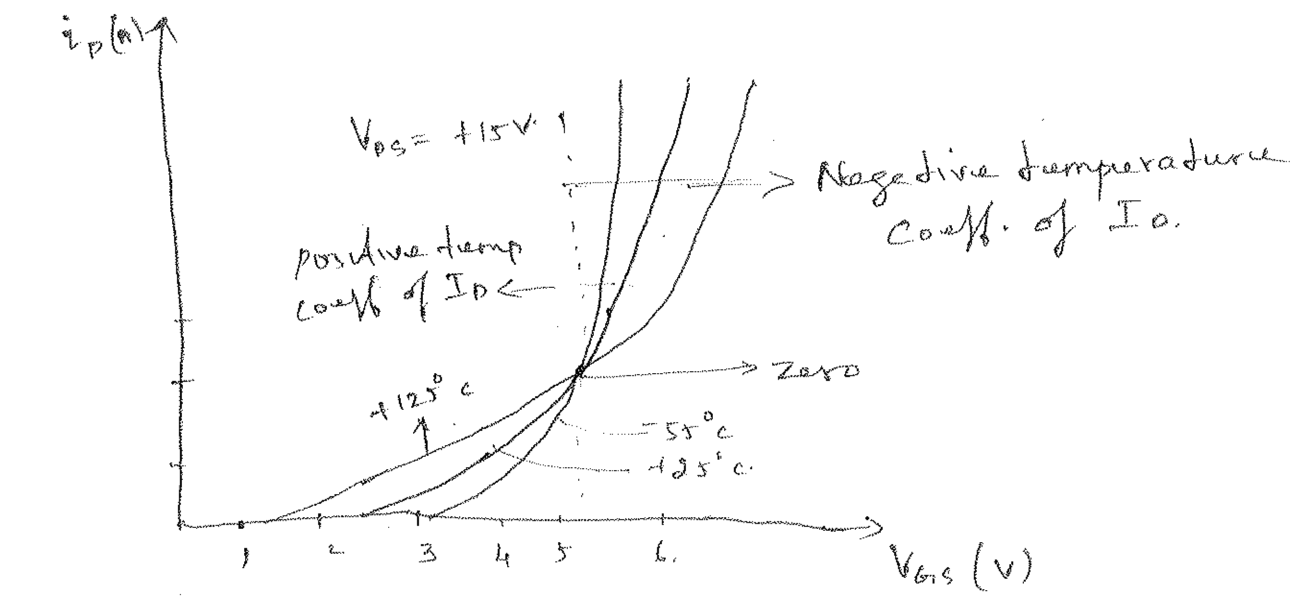
result of high electric field along short channel causing velocity to reach upper limit, known as velocity saturation.

* The linear $i_D - V_{GS}$ relation implies constant g_m
* MOSFET is driven in cut off by applying

$$V_{GS} < V_{GS(th)}$$

* In ohmic region, MOSFET conducts heavily
* In power application MOSFET is never operated in active region

MOSFET Temperature effects



* it is observed that there is value of V_{GS} in the range of 4V to 6V at which temp coeff of i_D is zero.

* At higher values of V_{GS}, i_D exhibits negative temperature coefficient.

* At i_D lower than zero temperature coeff, the temperature coefficient of i_D is positive and MOSFET suffer from thermal runaway.

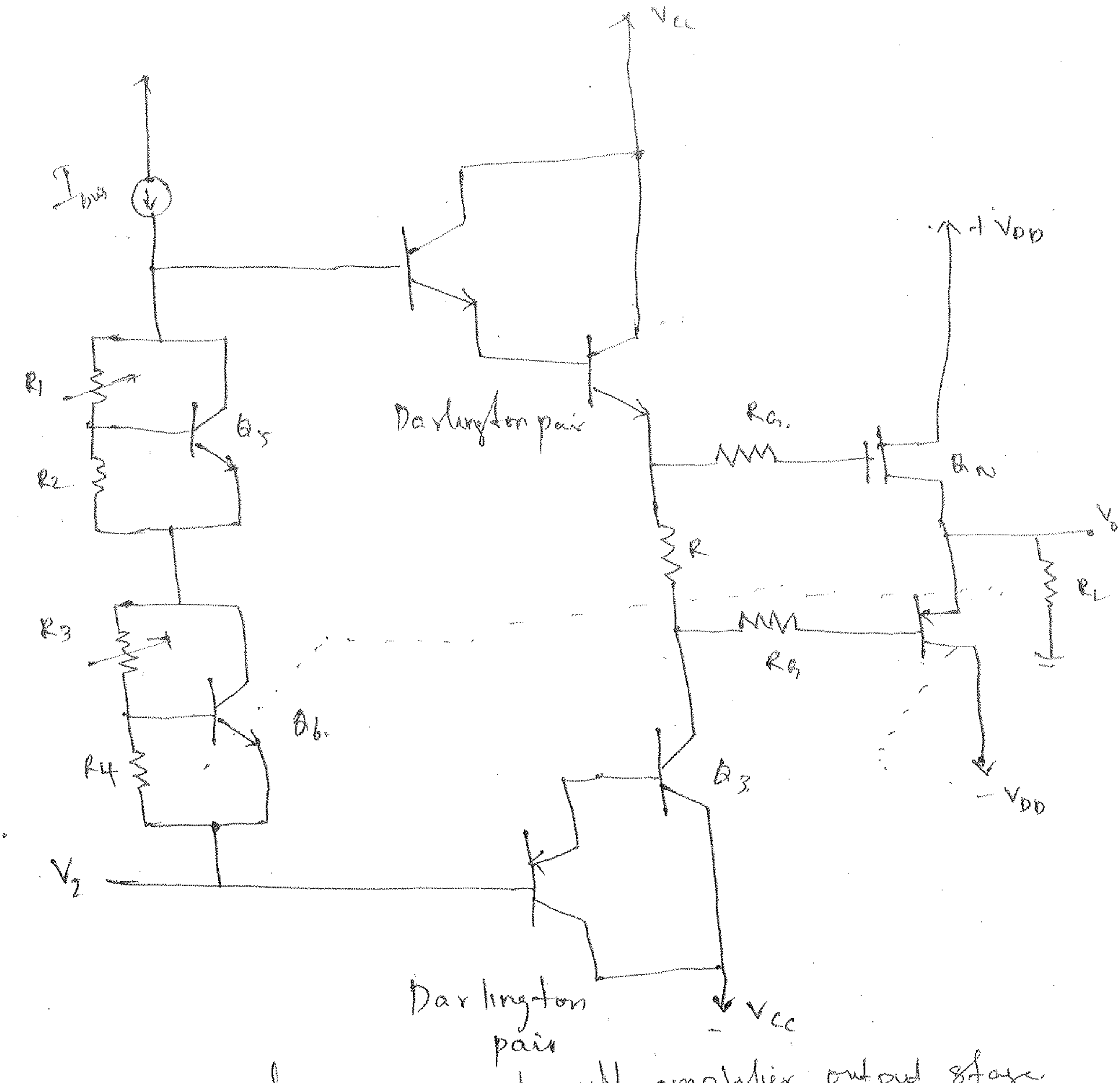
* The reason for positive temperature coeff of i_D at low current is that $V_{OV} = V_{GS} - V_t$ is relatively low.

* Temperature dependence is dominated by negative temperature coefficient of V_t which causes V_{OV} to rise with temperature.

<u>Class AB power amplifier using power MOSFET</u>

* The output stage of class AB has a pair of complementary MOSFET
* BJT are used for biasing and as a driver

* Complementary darlington emitter follower formed by Q_1 through Q_4 provide low output resistance necessary for driving output MOSFET at high speeds.



class AB push pull amplifier output stage.

* The bias circuit uses two VBE multipliers formed by Q_5 and Q_6.
* By adjusting the V_{BE} multiplication factor of Q_6 the bias voltage V_{GG} can be made to decrease with temperature at same rate.

$$V_{GG} = \left(1 + \frac{R_3}{R_4}\right) V_{BE6} + \left(1 + \frac{R_1}{R_2}\right) V_{BE5} - 4 V_{BE}$$

The value of V_{GG} is adjusted to get the desired Q point.

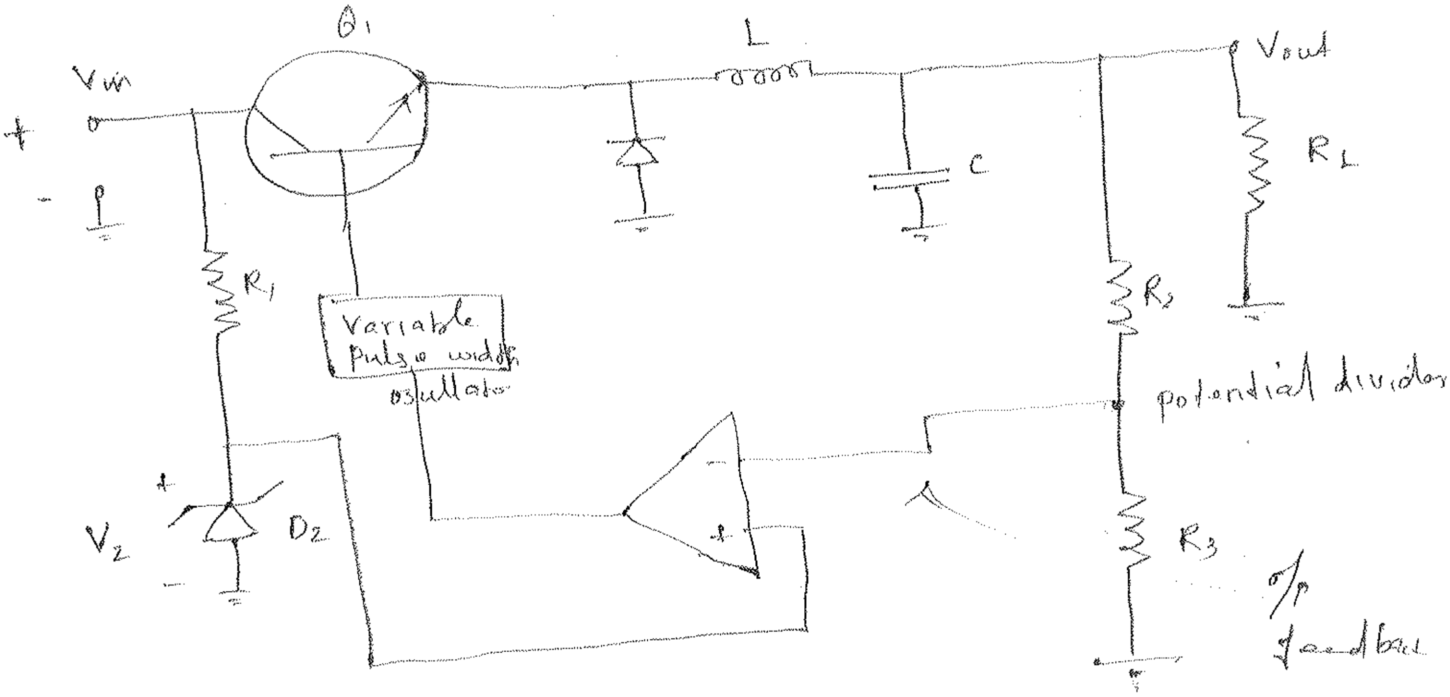
DC/DC Convertors

* The switch mode regulator is used to describe a ckt which takes dc i/p & provides single d.c o/p.

There are three basic configuration of switching regulator

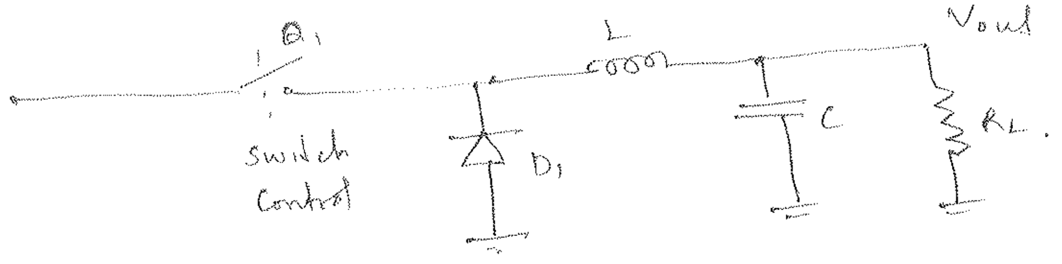
1. Step down or Buck switching regulator
2. Step up or boost switching regulator
3. Inverting type (Buck - Boost switching regulator).

Step down Switching Regulator (Buck)

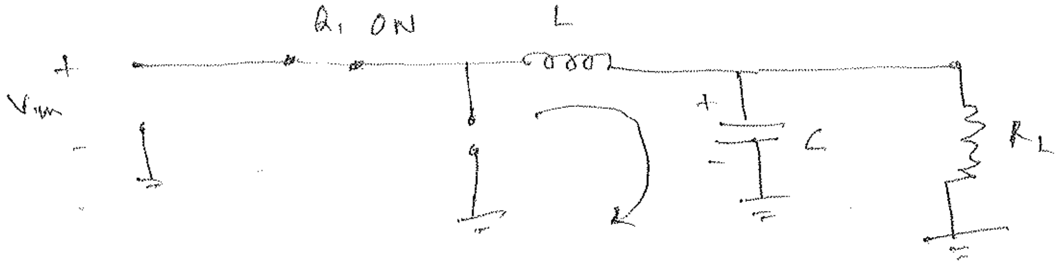


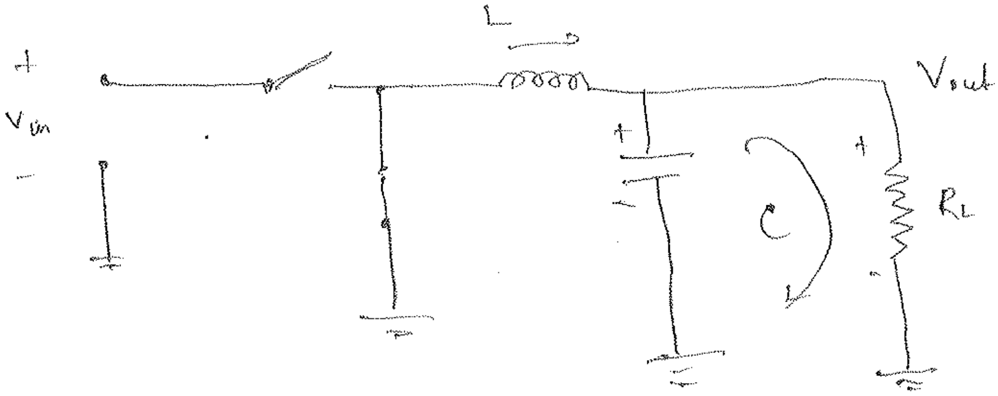
* It uses an inductor L and series transistor Q_1 which acts as a switch.
* The reference for error amplifier is provided by Zener voltage V_Z.

* The o/p is fedback to error amplifier through potential divider.
* The pulse width oscillator controls the operation of Q_1 as on or off, depending on load requirements.
* The equivalent ckt of regulator is shown in fig



* Q_1 is used for switching the i/p voltage for the required period of time
* LC filter averages the switched voltage
* When Q_1 is ON, the capacitor charges through L and when Q_1 is OFF the capacitor discharges through load.





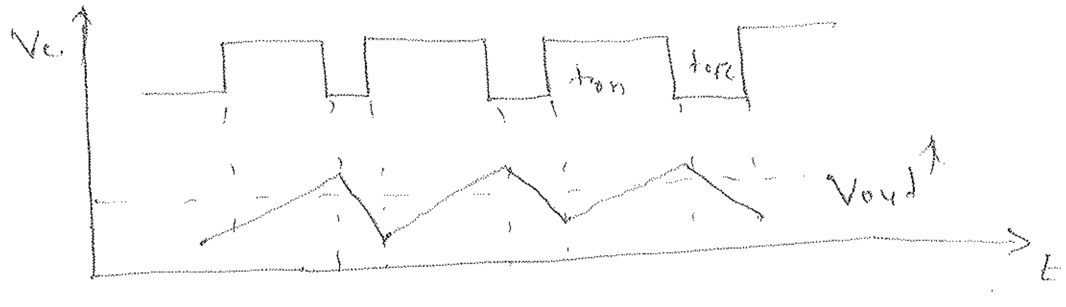
* The variable pulse width oscillator controls ON/OFF periods of Q_1
* When ON time is more than OFF time, the capacitor charges more, increasing o/p voltage.
* When OFF time is more than ON time, the capacitor discharges more reducing o/p voltage.
* By adjusting duty cycle $\delta = \frac{t_{on}}{t_{on}+t_{off}}$ of Q_1, the o/p voltage can be regulated.
* If o/p volt ↓ the voltage across R_3 ↓, the error across error amplifier is more.
* This produces pulse of higher width. This increases the charging of capacitor, producing more o/p voltage thus the decreased voltage gets compensated.
* If o/p volt ↑, the voltage across R_3 ↑, which produces pulse of smaller width, which reduces t_{on} for Q_1
* This makes the capacitor to discharge more which increases the o/p voltage.
* Thus o/p voltage is maintained constant by controlling duty cycle of Q_1.

* the o/p voltage is given by

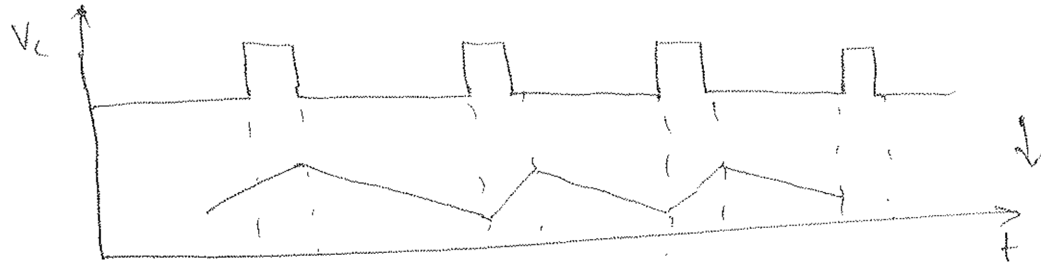
$$V_{out} = \delta V_{in}$$

Where $\delta = \dfrac{t_{on}}{t_{on} + t_{off}} = \dfrac{t_{on}}{T} = t_{on}(f).$

f = frequency

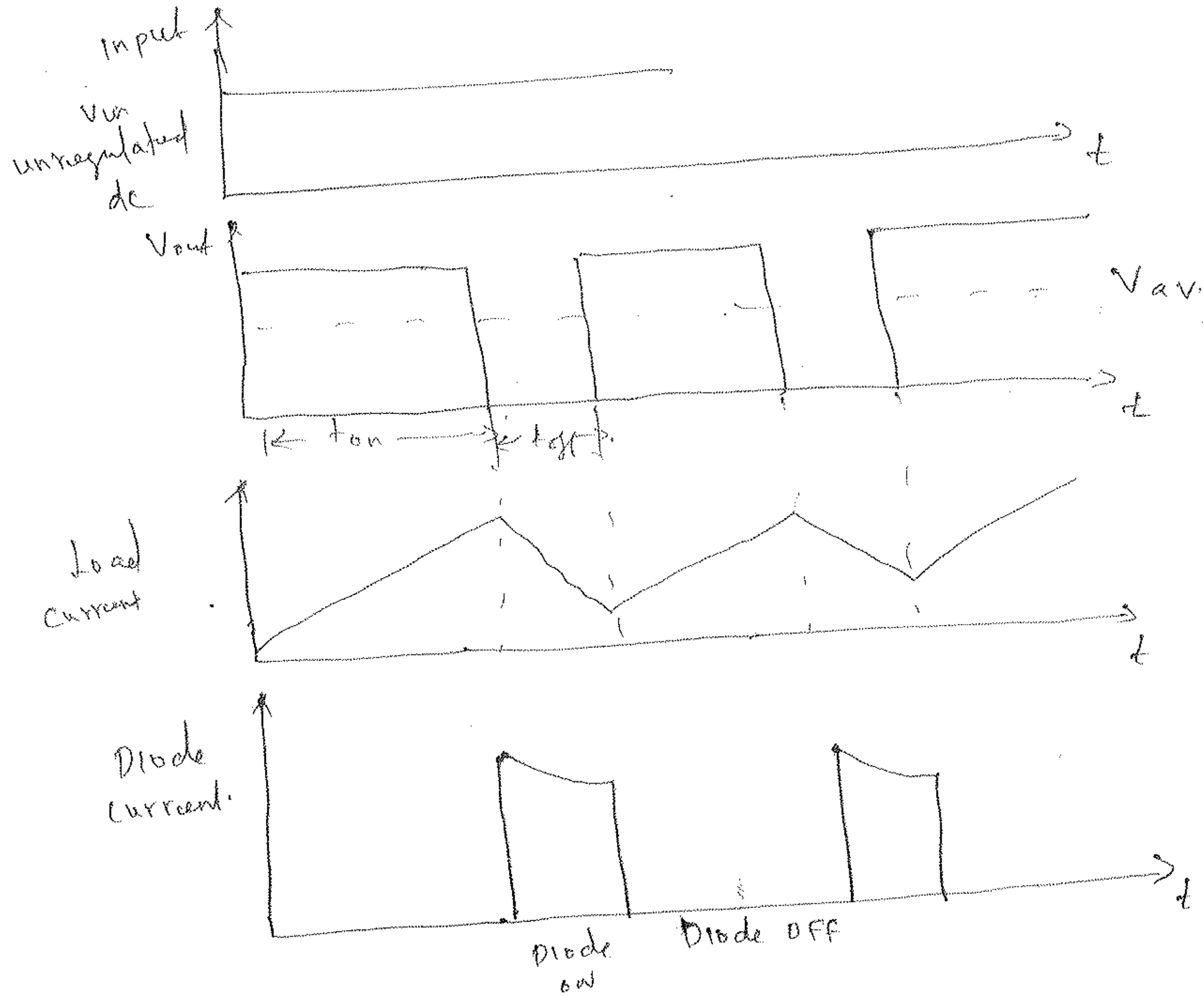


@ ton > toff



↓ Vout

ton < toff



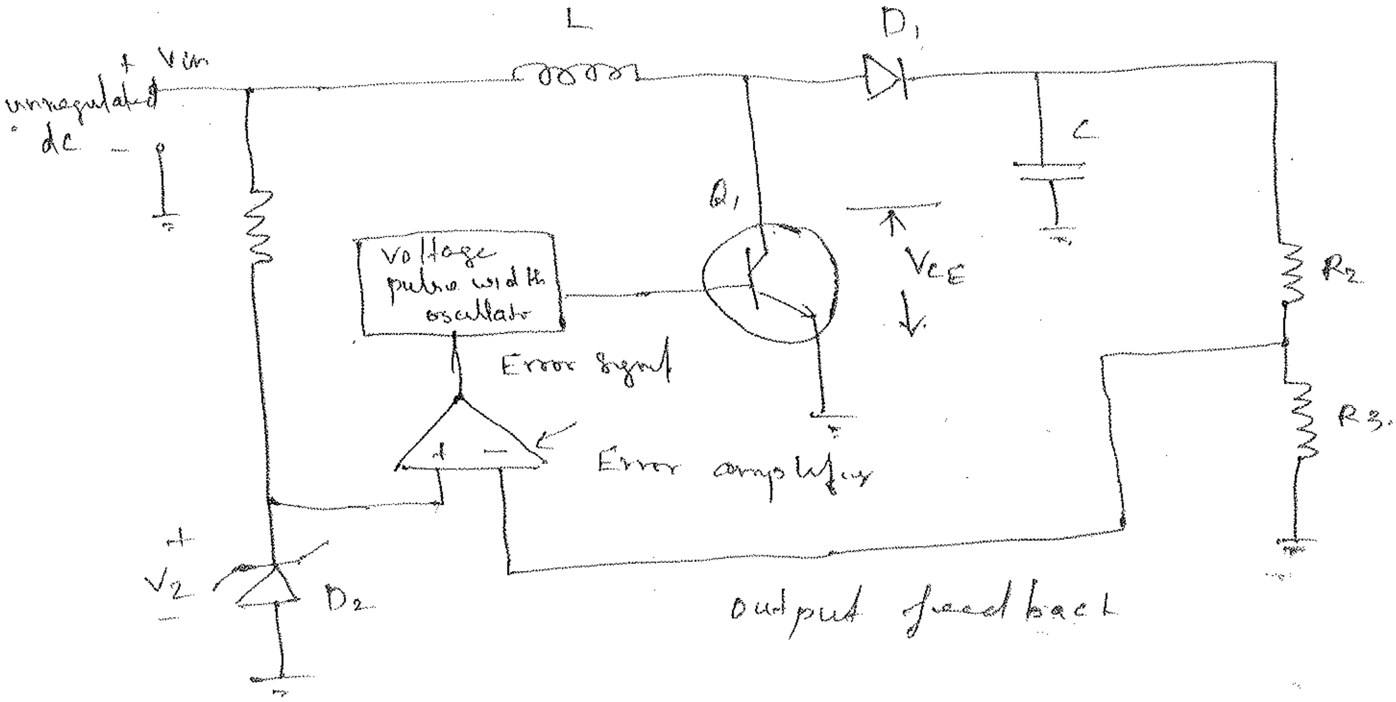
Advantage:

1. High efficiency
2. Simple to design
3. Low ripple condent
4. Small output filter

Disadvantage.

1. Single o/p
2. High i/p ripple
3. No isolation b/w i/p & o/p.
4. Slow transient response.

Step up switching Regulator (Boost)



* Transistor Q_1 acts as ON/OFF switch
* When Q_1 is driven into saturation, V_{CE} is very very small and acts as short circuit.

Case (1) Let Q_1 is ON (saturation)

* When Q_1 is ON, V_{CE} is denoted as V_{CE} (sat)
* The voltage across L becomes $[V_{in} - V_{CE}(sat)]$
This expands the magnetic field around the inductor
* During ON time of Q_1 the voltage across the inductor starts decreasing exponentially from $[V_{in} - V_{CE}(sat)]$

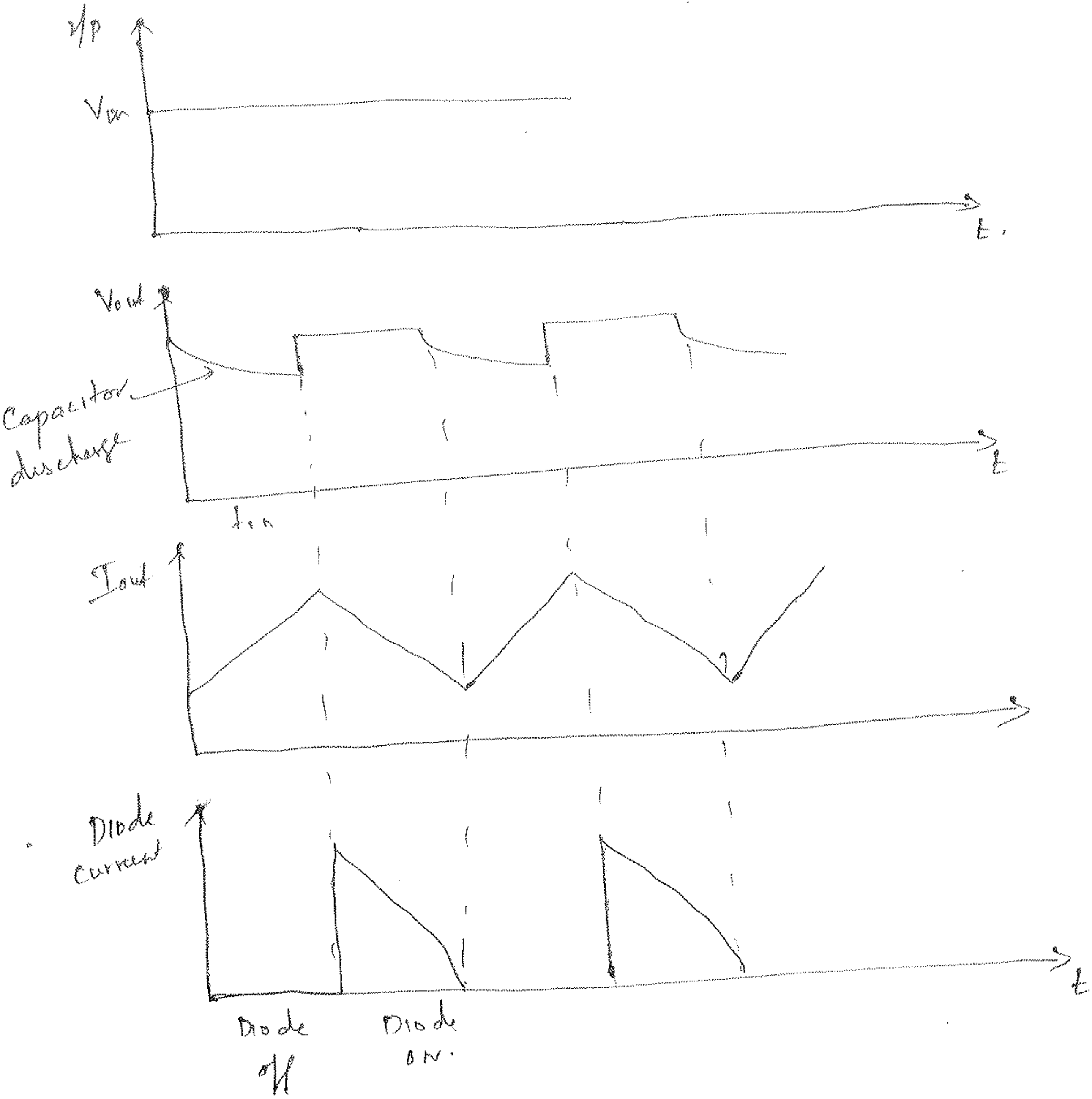
Case 2 : Let Q_1 is OFF (cutoff)
* When Q_1 is OFF the magnetic field of inductor L, collapses and its polarity gets reversed, since inductor current cannot change instantly.
* The shorter the ON period the greater is V_1.
* The longer the ON time, the smaller the inductor voltage V_1 & less voltage get added to V_{in}, decreasing o/p voltage.
* When o/p voltage decrease due to increase in load current then ON time of Q_1 gets reduced.
* When o/p voltage increase the ON time of Q_1 gets increased

The output voltage is given by

$$V_{out} = \frac{V_{in}}{\delta}$$

$$\delta = \frac{t_{on}}{T}$$

Waveform for step up switching regulator.



Advantage

1. O/p voltage is higher than I/p voltage
2. η is high
3. Low I/p ripple
4. Simple in design

Disadvantage

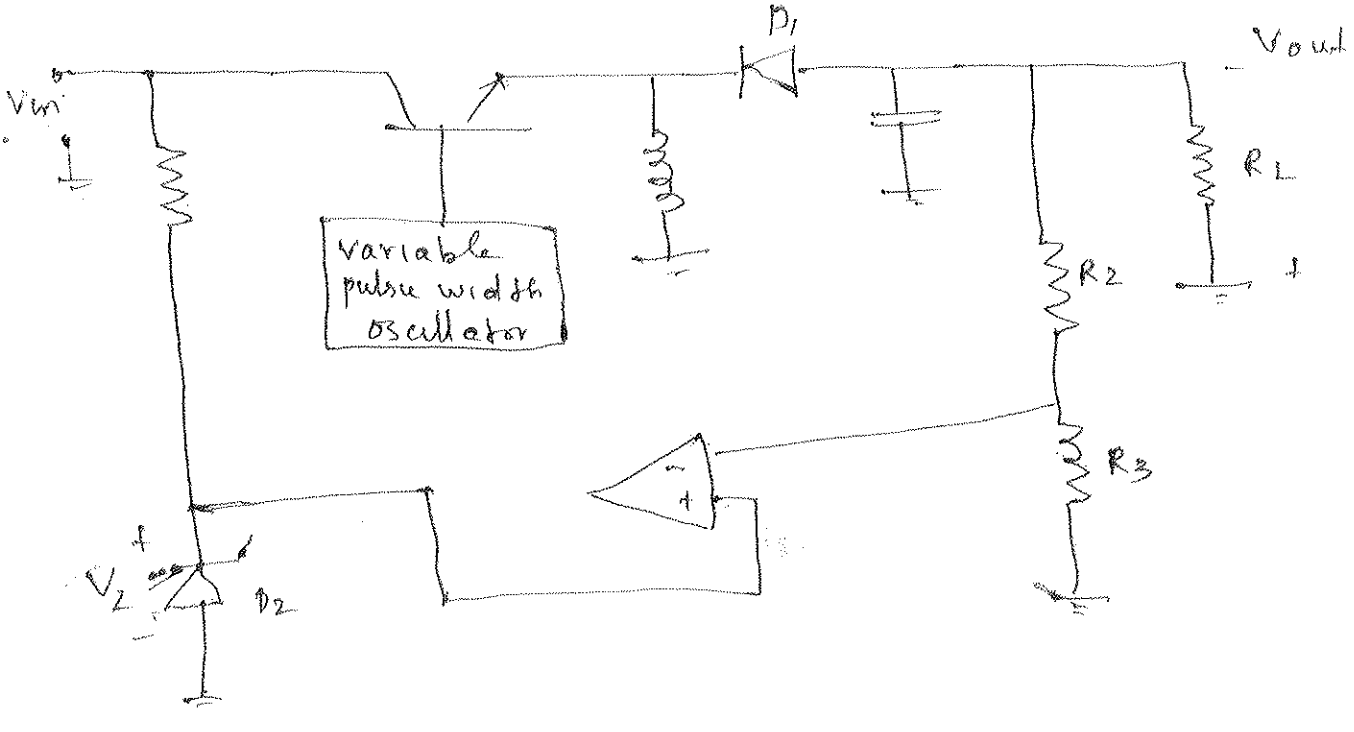
1. It provides single O/p
2. Duty cycle limited to 50%
3. No isolation b/w I/p & O/p.

Buck - Boost (or) Voltage Inverter Type switching regulator

* it produces output voltage having polarity opposite to that of input voltage.

* Any change in output produces error which gets amplified by opamp error amplifier.

* Thus control on/off period of Q_1 to regulate the output, through variable pulse width oscillator.



Case :1 Let Q_1 is switched ON

→ Q_1 goes into saturation and voltage across it drops to $V_{CE(sat)}$ about 0.3 V

→ Due to this voltage across inductor suddenly rises to $[V_{in} - V_{CE(sat)}]$ and magnetic field around it suddenly expands.

* The inductor value starts exponentially decreasing from intial value $[V_{in} - V_{CE}(sat)]$

Case 2 : Let Q_1 is OFF.

* As Q_1 OFF, the magnetic field across L gets collapsed
* Voltage reverses its polarity
* Due to reverse V_1, diode D_1 is now forward biased.
* The capacitor charges through D_1 producing output voltage of opposite polarity to that of V_{in}
* Hence regulator is voltage inverter type.