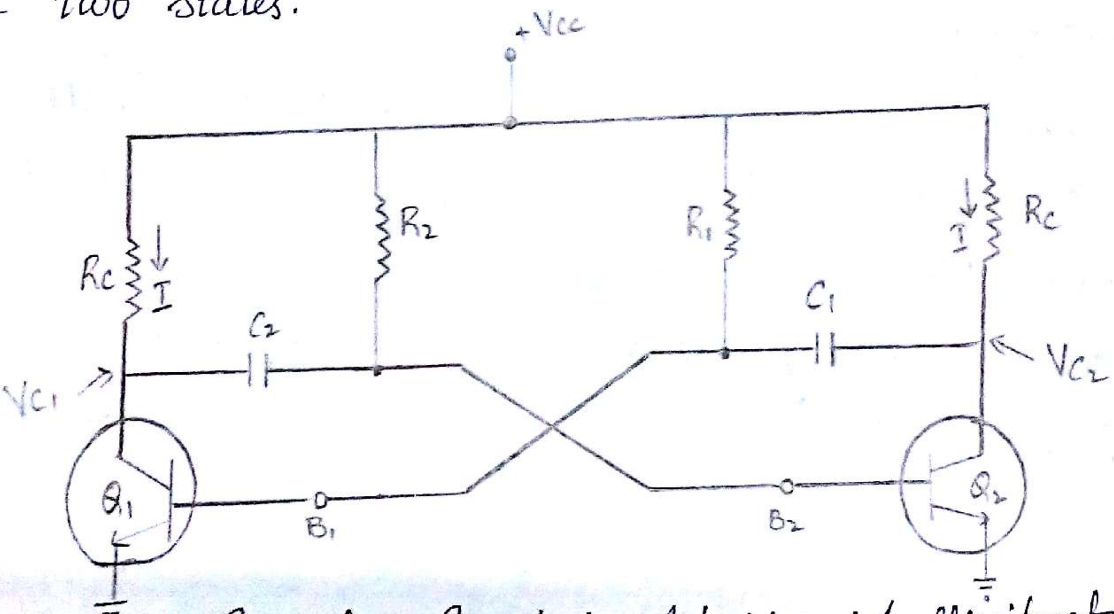
Astable Multivibrator

* The astable multivibrator has two states, both are quarsi-stable. None of the states is stable.
* Without external trigger, multivibrator keeps on alternating the states.
* It cannot remain indefinitely (i.e) stable in any of these two states.



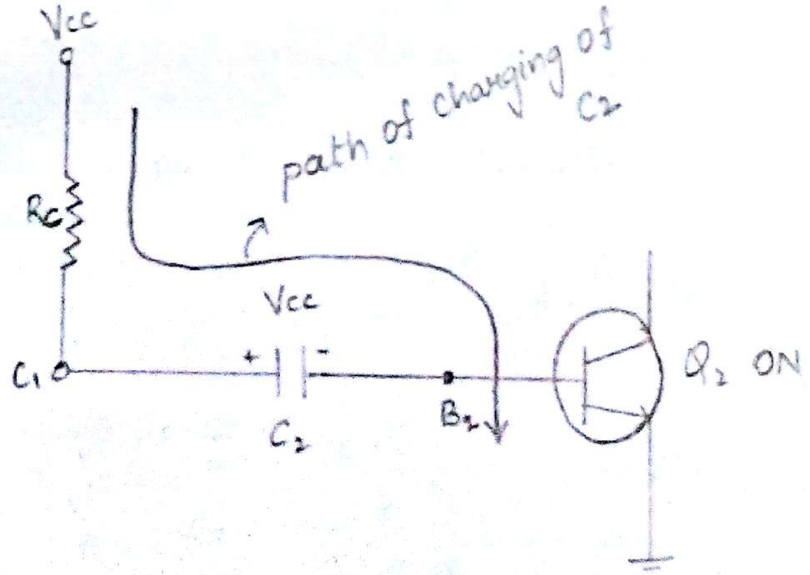
Collector Coupled Astable Multivibrator.

Construction:- The Q_1 and Q_2 are identical npn transistors.

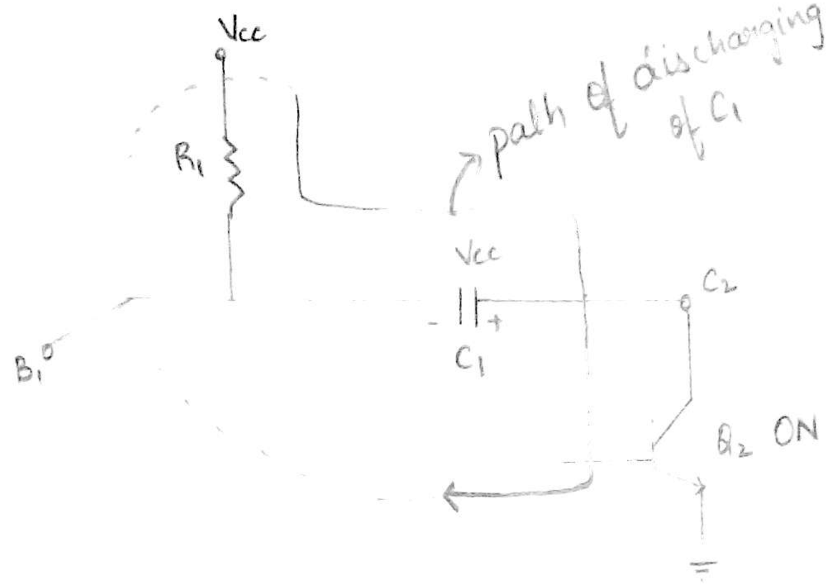
* The two collector resistances are equal to R_C
* The collector Q_1 is coupled to the base of Q_2 through capacitor C_2 while the collector of Q_2 is coupled to the base of Q_1 through capacitor C_1.

* The capacitive coupling is used between the due to which neither transistor (both the trans can remain permanently cut-off.
* The circuit has two quasi-stable states and it make periodic switching between these states, without any external trigger signal.

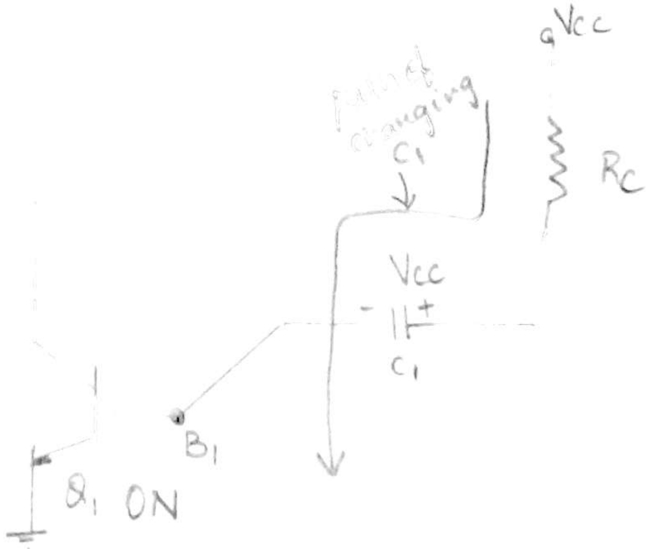
Working:-
* Assume that the state is Q_2 ON and Q_1 OFF.
* The capacitor C_2 starts charging towards Vcc through path Rc, C_2 and ON Q_2.
* Finally voltage across C_2 becomes equal to Vcc with proper polarity.



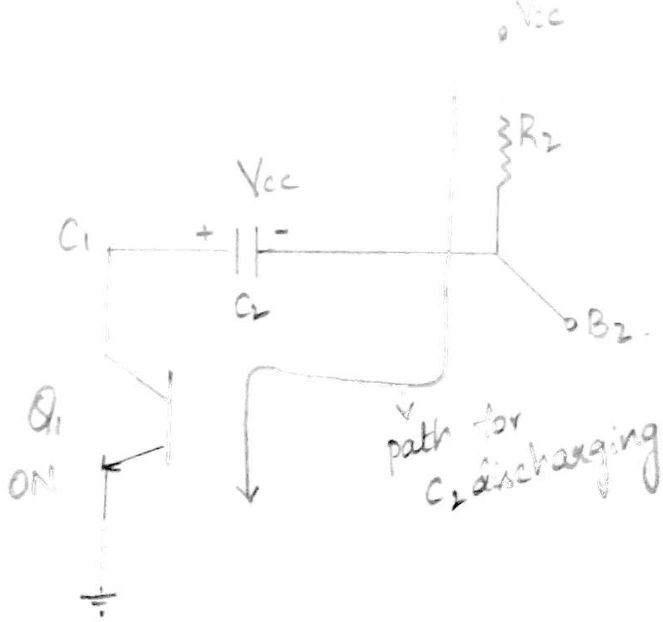
* At the same time capacitor C_1 which is charged to Vcc in earlier state, starts discharging through path Q_2, Vcc, R_1, C_1.



* The base of Q_1 is at $-V_{cc}$ at the begining.
* As C_1 starts discharging, it becomes less and less negative (i.e) becomes more positive.
* It becomes equal to V_γ, the cut in voltage of transistor Q_1
* When it becomes greater than V_γ, the transistor Q_1 starts conducting, so Q_1 becomes ON at the same time Q_2 becomes OFF
* The negative potential applied at B_2, due to charged C_2 ensures that Q_2 becomes indeed OFF
* When this happens, the capacitor C_1 starts charging through R_C, C_1 and ON transistor Q_1



* C_2 starts discharging, through the path V_{cc}, R_2, C_2 and ON transistor Q_1

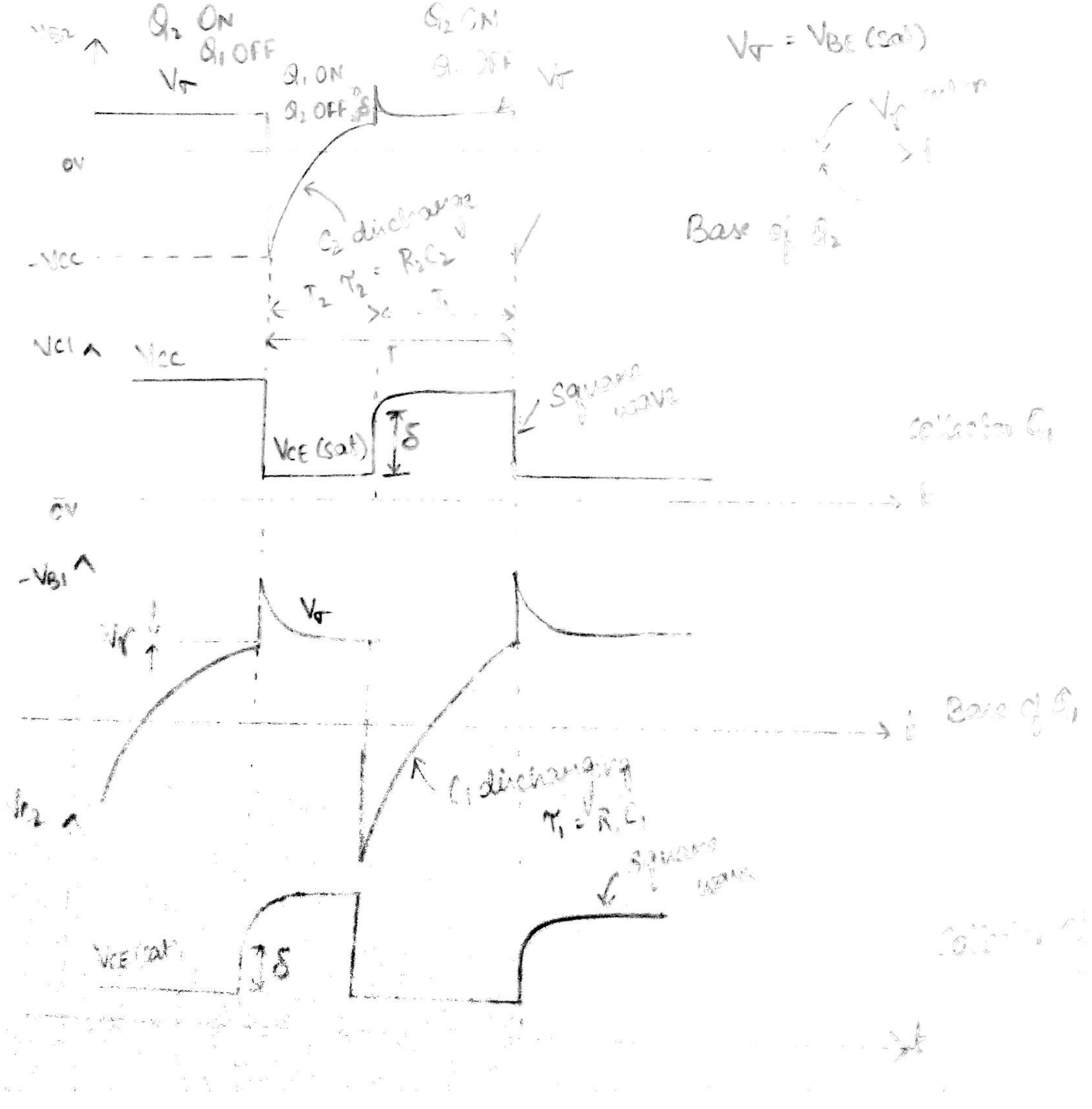


* As C_2 discharges, B_2 potential becomes less negative (i.e) it increases towards positive.
* When V_{B2} becomes just greater than cut in voltage V_γ of Q_2, the Q_2 starts conducting.
* Thus now Q_2 becomes ON and Q_1 OFF.
* Thus changes in two states is automatic and without any external triggering signal.

Waveforms of Astable Multivibrator.

When Q_1 is OFF and Q_2 is ON, the C_1 discharges and voltage at B_1 (i.e) V_{B1} increases. This increases exponentially with the time constant R_1C_1. This valtage was initially at $-V_{cc}$. When this voltage increases beyond cut in voltage of Q_1, Q_1 starts conducting when Q_1 is in staturation

$V_{B1} = V_{BE(sat)}$ $\quad V_{C1} = V_{CE(sat)}$ $\quad V_{C2} = V_{CC}$



When Q_1 is ON and Q_2 is OFF, the C_2 discharges and the capacitor voltage at B_2 i.e V_{B2} increases. This increases exponentially with time constant R_2C_2. This voltage was initially at $-V_{CC}$ in previous state. When this voltage increases beyond cut in voltage of Q_2, Q_2 conducts and goes in saturation.

$$V_{B2} = V_{BE}(sat), \quad V_{C2} = V_{CE}(sat)$$

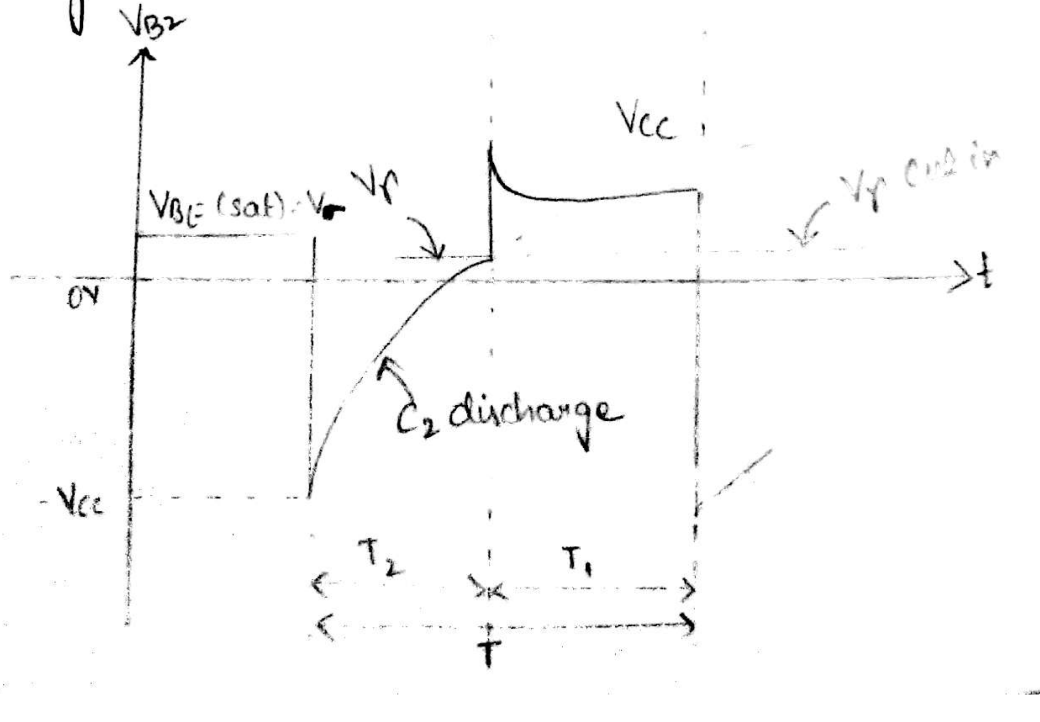
$$V_{C1} = V_{CC}$$

The analytical expression for δ is same as derived earlier for monostable multivibrator.

While the base current I_B at overshoot is given by

$$I_B' = \frac{V_{CC} - V_{CE}(sat) - V_\delta + V_\gamma}{R_C + r_{bb}'}$$

<u>Expression for Time period T</u> :- For deriving the expression for time period T which is $T_1 + T_2$, consider the wave form at base of any transistor.



The capacitor C_2 discharges exponentially and the voltage V_{B2} increases exponentially.

V_i = Initial value of V_{B2} = $-V_{cc}$ — ①

V_f = Final value of V_{B2} = $+V_{cc}$ — ②

Though it stops increasing beyond V_r (cut in) voltage, its rise is towards $+V_{cc}$, which is its final steady state value, with the time constant $\tau_2 = R_2 C_2$.

For the capacitor we can write the basic equation as,

$$\boxed{V_0 = V_f - (V_f - V_i)\, e^{-t/\tau}}$$

Here V_0 means the base voltage,

$$V_{B2} = V_{cc} - (V_{cc} - (-V_{cc}))\, e^{-t/R_2C_2} \quad — ③$$

$$\therefore V_{B2} = V_{cc} - 2V_{cc}\, e^{-t/R_2C_2} \quad — ④$$

(or)

$$V_{B2} = V_{cc}\left(1 - 2e^{-t/R_2C_2}\right) \quad — ④.$$

We know that at switching time,

$$\boxed{t = T_2 \text{ and } V_{B2} = V_r} \quad — ⑤$$

Substituting in equation ④

$$V_r = V_{cc}\left(1 - 2e^{-T_2/R_2C_2}\right)$$

The best approximation to obtain T_2 is, $V_r = 0V$.

$$\therefore 0 = V_{cc}\left(1 - 2e^{-T_2/R_2C_2}\right)$$

$$\therefore 1 - 2e^{-T_2/R_2C_2} = 0$$

$$e^{-T_2/R_2C_2} = 0.5$$

$\therefore \ln\left(e^{-T_2/R_2C_2}\right) = \ln(0.5)$

$$-\frac{T_2}{R_2C_2} = -0.69$$

$$\boxed{T_2 = 0.69\, R_2C_2} \quad — (7)$$

Similarly we can write the eqn at $t = T_1$ and find out expression for T_1 which same as for T_2.

$$\boxed{T_1 = 0.69\, R_1C_1} \quad — (8)$$

$$T = T_1 + T_2$$

$$T = 0.69(R_1C_1 + R_2C_2) \quad — (9)$$

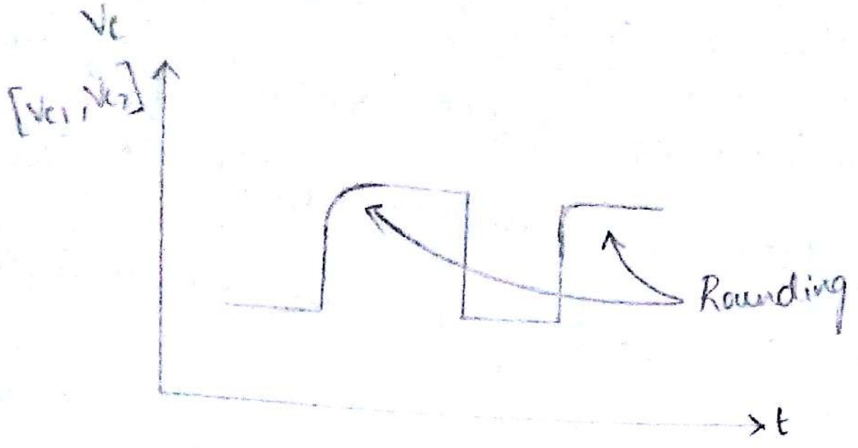
if $R_1 = R_2 = R$; $C_1 = C_2 = C$ then

$$\boxed{T = 0.69\,(2RC) = 1.38\, RC} \quad — (10)$$

For this case we get square wave at the output which is symmetrical.

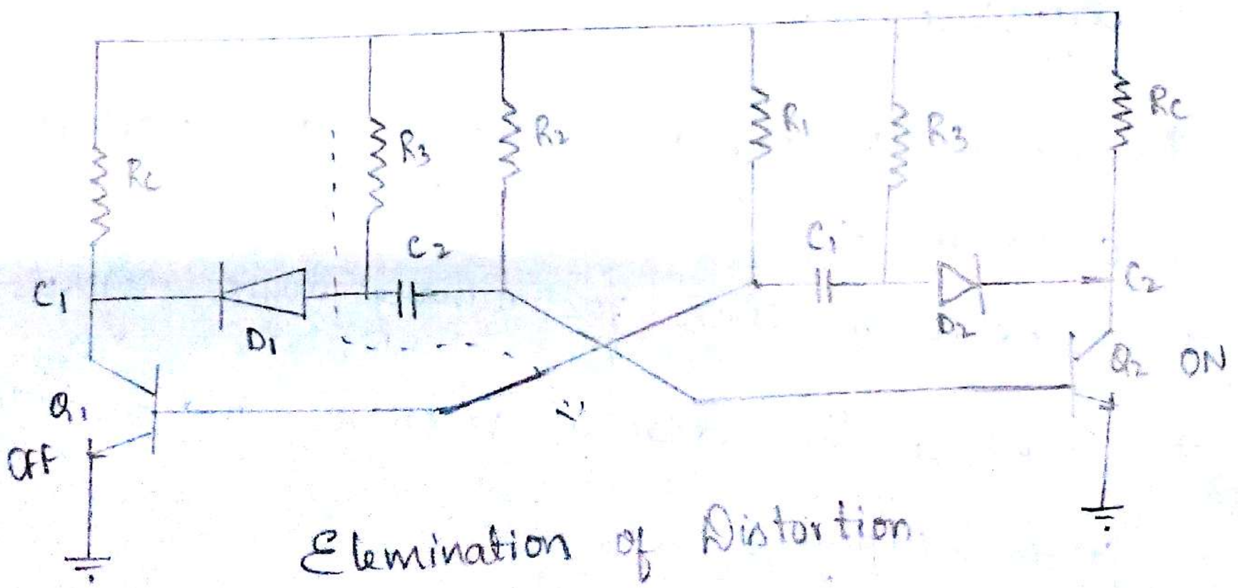
Distortion and its Elimination

It can be seen that, in the collector wave-forms, there is certain distortion present. Instead of exact square wave we get vertical rising edges little but rounded. This is called rounding. For square wave output such a rounding is undesirable and must be eliminated.



Rounding in collector waveforms

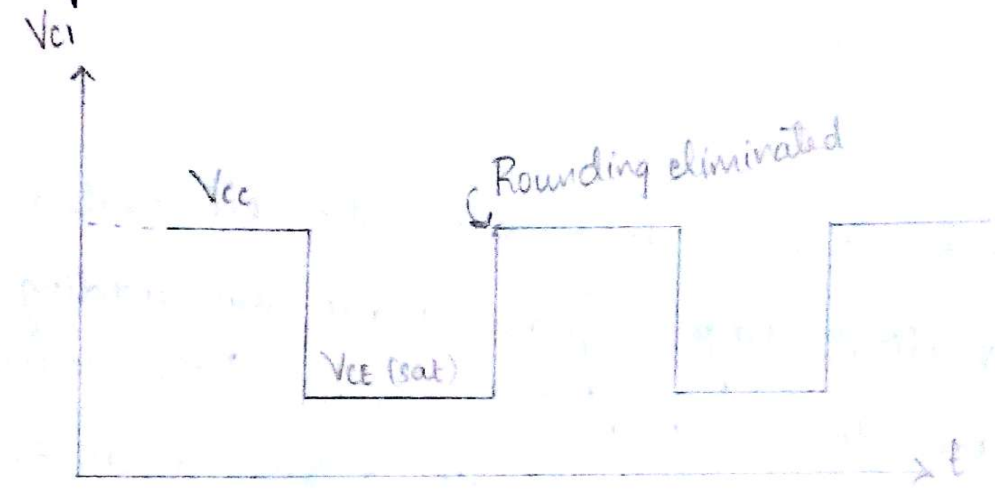
Such a rounding can be eliminated to obtain the vertical edges square wave by adding two collector diodes and two resistors. The collector coupled astable multi vibrator with collector diodes and auxilary resistors.



Elemination of Distortion

If Q_1 is OFF, then its collector voltage increases suddenly to V_{cc} thus making D_1 reverse biased.

Thus the charging of C_2 now takes place through R_3 rather than R_C. As current does not flow through R_C the collector voltage can rise suddenly to V_{CC} and the rounding at collector completely gets eliminated.



Elimination of Rounding.

Applications :-

* Used as square wave generator, voltage to frequency converter etc.
* Used as a clock for binary logic signals.
* Used in the digital voltmeter and switched mode power supply.
* As an oscillator to generator wide range of audio and radio frequencies.

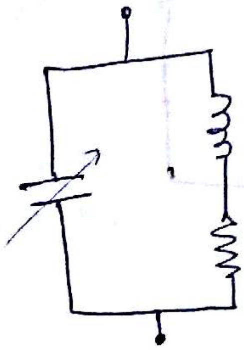
UNIT - 3

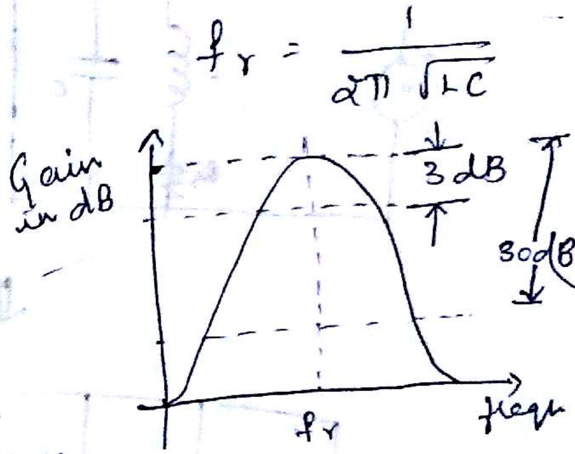
TUNED AMPLIFIER

High i/p } current
High o/p }
Hi i/p } voltage
low o/p }
Ladder - RC

Practice

* Tuned amplifier (or) Narrow band amplifier:





Ratio of 30dB Bw to 3dB Bw is called skirt selectivity

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

* Loaded and unloaded Q:

unloaded:

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

Loaded:

$$Q = \frac{2\pi \times \text{Maximum energy stored per cycle}}{\text{Energy dissipated per cycle + Energy dissipated due to presence of load}}$$

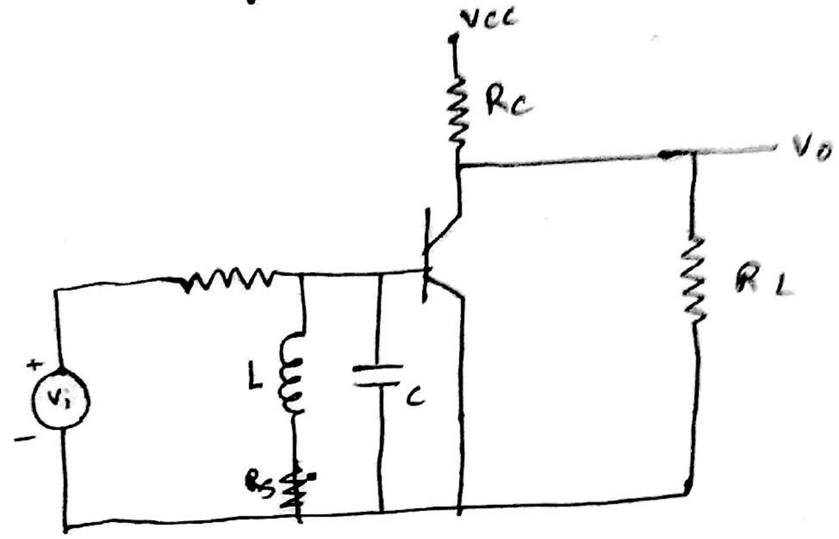
Quality factor determines the 3dB Bandwidth for the reference circuit

$$B.W = \frac{f_o}{Q}$$

* Small Signal Tuned Amplifier

$Q_L =$

$Y_1 =$



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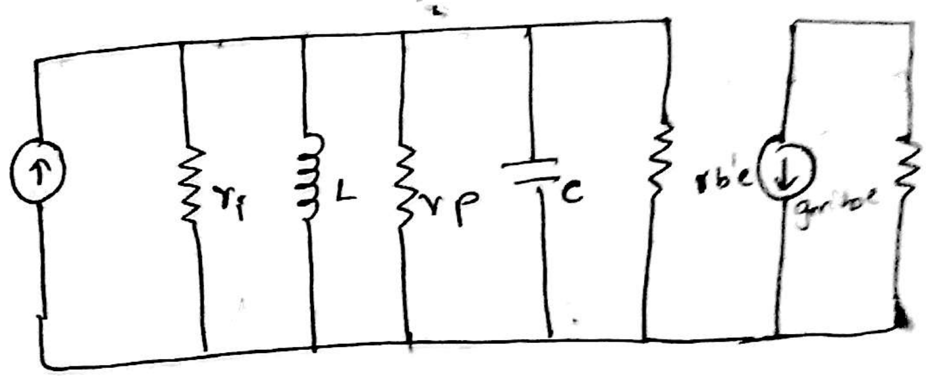
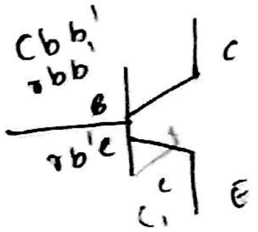




where $C = C' + C_{b'e} + (1 + g_m R_L) \cdot C_{b'c}$

Assumption :

1) $R_1 << R_C$
2) $r_{bb'} = 0$.
3) $C = C' + C_{b'e} + (1 + g_m R_L) \cdot C_{b'c}$

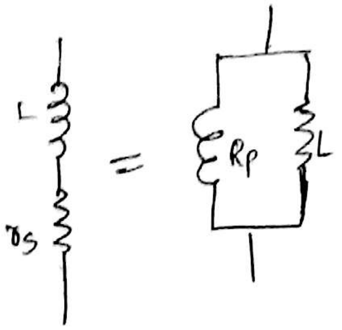
C' = external capacitance used to tune the ~~capacitance~~ tank circuit

$(1 + g_m R_L) C_{b'c}$ = Miller capacitance

R_s = loss in the coil

4) Assume coil losses are low, Q is very high

$$Q_L = \frac{\omega L}{r_s} >> 1$$



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

$$= \frac{1}{r_s + j\omega L} \times \frac{r_s - j\omega L}{r_s - j\omega L}$$

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

$\omega L >> r_s$

for series circuit $$Y_1 = \frac{r_s}{r_s^2 + \omega^2 L^2} + \frac{1}{j\omega L} \quad \text{——— (1)}$$

for parallel circuit $$Y_2 = \frac{1}{R_p} + \frac{1}{j\omega L} \quad \text{——— (2)}.$$

Equate Y_1 and Y_2

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

$$\frac{1}{R_p} = \frac{r_s}{r_s^2 + \omega^2 L^2} \qquad \left[\frac{\omega L}{r_s} = Q\right]$$

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

$$1/R_p = \frac{r_s^2}{r_s \cdot \omega^2 L^2}$$

$$\frac{1}{R_p} = \frac{1}{r_s Q^2}$$

$$R_p = r_s \cdot Q^2$$

from the figure calculate $R = R_i \| R_p \| r_{b'e}$

<u>Current gain of the amplifier:</u>

$$A_i = \frac{-g_m R}{1 + j\left(\omega RC - \frac{R}{\omega L}\right)}$$

$$= \frac{-g_m R}{1 + j\omega_0 RC\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}$$, where $\omega_0^2 = \frac{1}{LC}$

At resonant frequency :

$$Q_c = \frac{R}{\omega_0 L} = \omega_0 RC$$

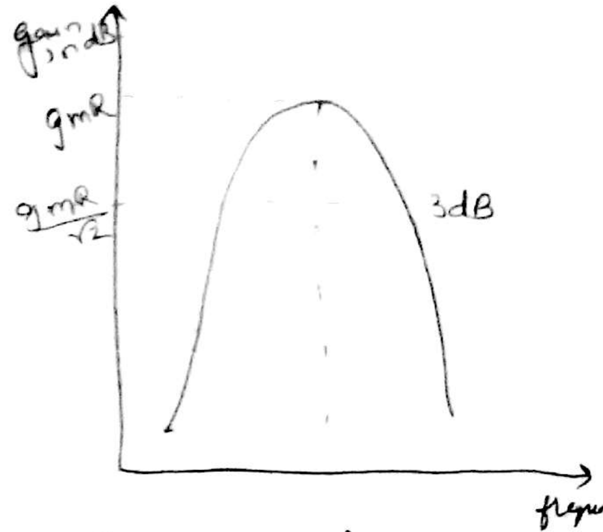
$$A_i = \frac{-g_m R}{1 + jQ_1\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right)}$$

At $\omega = \omega_0$ (maximum gain)

$$A_i(max) = -g_m R$$

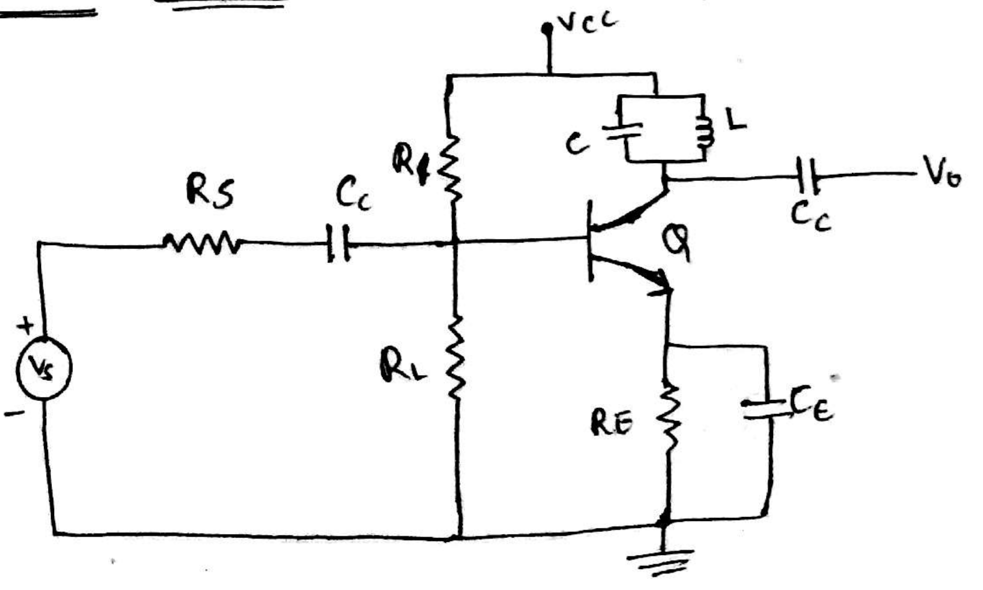
At 3 dB condition ,

$$|A_i| = \frac{g_m R}{\sqrt{2}}$$



100%

* SINGLE TUNED CAPACITIVE COUPLED AMPLIFIER :



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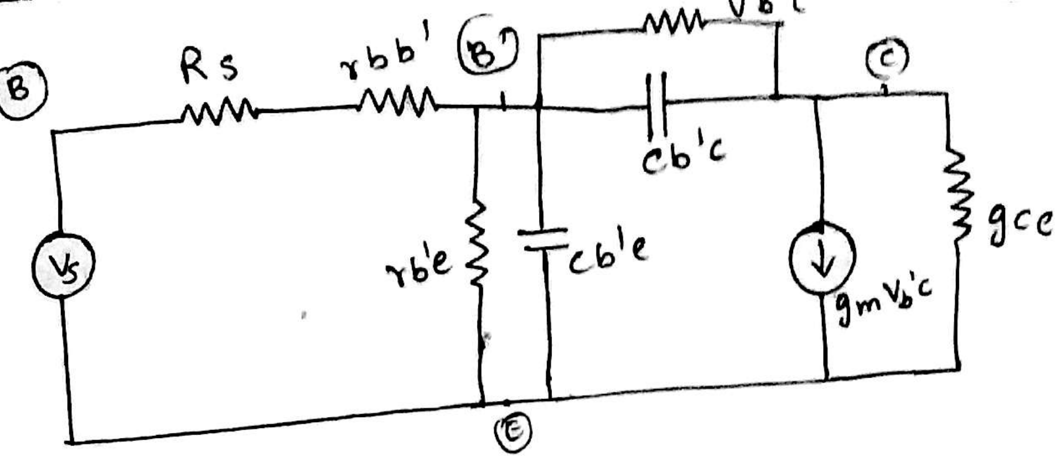
Where



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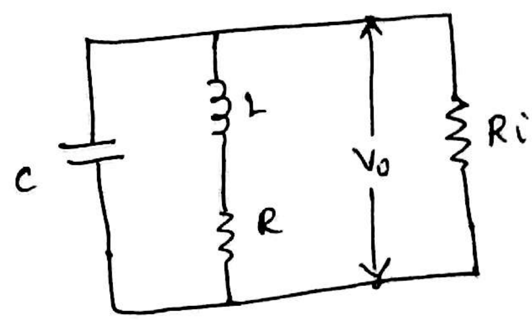
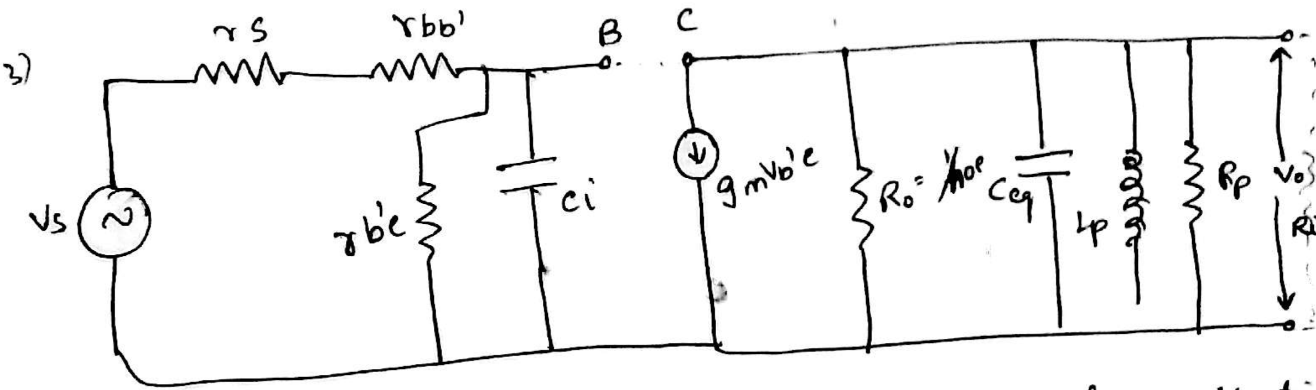


3)



By Applying Miller theorem we have that equivalent diagram

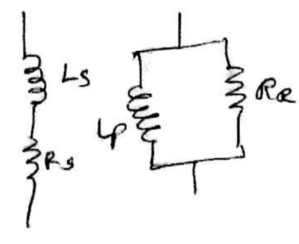
Here, $C_i = C_{b'e} + C_{b'c}(1-A)$

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

$$G_{ce} = \frac{1}{R_{ce}} = h_{oe}$$

The series RL circuit represented at a parallel circuit in the 3rd diagram

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



multiply $(R - j\omega L)$ on numerator and denominator

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

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

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

$$= \frac{1}{R_p} + \frac{1}{j\omega L_p}$$

where,

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

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

→ Centre frequency :

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

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

$$C_{eq} = C_{b'e} \left(\frac{1-A}{A}\right) + C$$

→ Quality factor (Q)

Quality factor 'Q' of the coil is given by :

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

The tuned amplifier Q value is always very high. $Q \ggg 1$, so $\omega L \ggg R$

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

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

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

Similary, $L_p = \dfrac{R^2 + \omega^2 L^2}{\omega^2 L}$

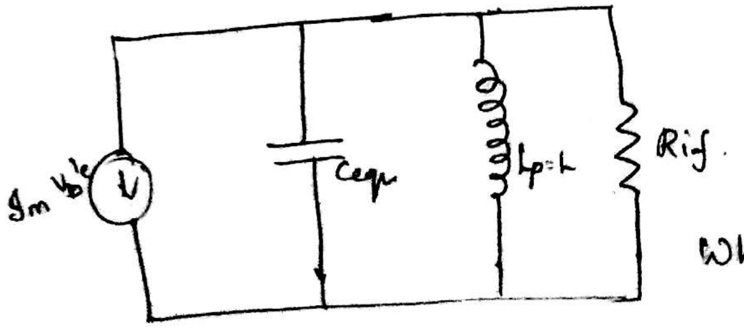
$$\frac{R^2}{\omega^2 L} + \frac{\omega^2 L^2}{\omega^2 L}$$

$$L_p = L.$$

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

$$= \omega_0 \cdot Q_r \cdot L \quad \left[Q_r = \frac{\omega_r L}{R}\right]$$

$$\boxed{Q_r = \frac{R_p}{\omega_r L}}$$



$$Q_{eff} = \frac{R_t}{\omega_r L} \quad (or)$$

$$\omega_r \cdot C_{eq} R_t$$

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

Effective Quality factor,

$$Q_{eff} = \frac{\text{Suceptance of Inductance (or) Capacitance}}{\text{Conductance of Shunt resistance } R_t}$$

→ * Voltage gain (A_V):

A_V for tuned amplifier is given by

$$A_V = -g_m \frac{r_{b'e}}{r_{b'b} + r_{b'e}} \times \frac{R_t}{1 + 2j\, Q_{eff} \cdot \delta}$$

where,

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

δ = fraction variant in the resonant frequency

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

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

→ * 3dB Bandwidth:

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

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

$$= \frac{\omega_r}{2\pi Q_{eff}} \qquad \because Q_{eff} = \omega_r R_t C_{eq}$$

$$= \frac{f_r}{Q_{eff}} \qquad \because \omega_r = 2\pi f_r$$

$$\boxed{B_W = \frac{f_r}{Q_{eff}}}$$

A_V / $A_V($

Effect of Cascaded Single tuned amplifier on Bandwidth

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

n - number of stages.

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

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

At 3dB condition,

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

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

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

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

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

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

$$\boxed{(2\cdot\delta Q_{eff}) = \sqrt{2^{1/n}-1}}$$

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

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

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

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

Let us assume f_1 and f_2 are the lower and upper cut off frequency

Higher cut-off frequency:

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

Lower cut-off frequency:

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

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

$$= f_2 - \cancel{f_r} + \cancel{f_r} - f_1$$

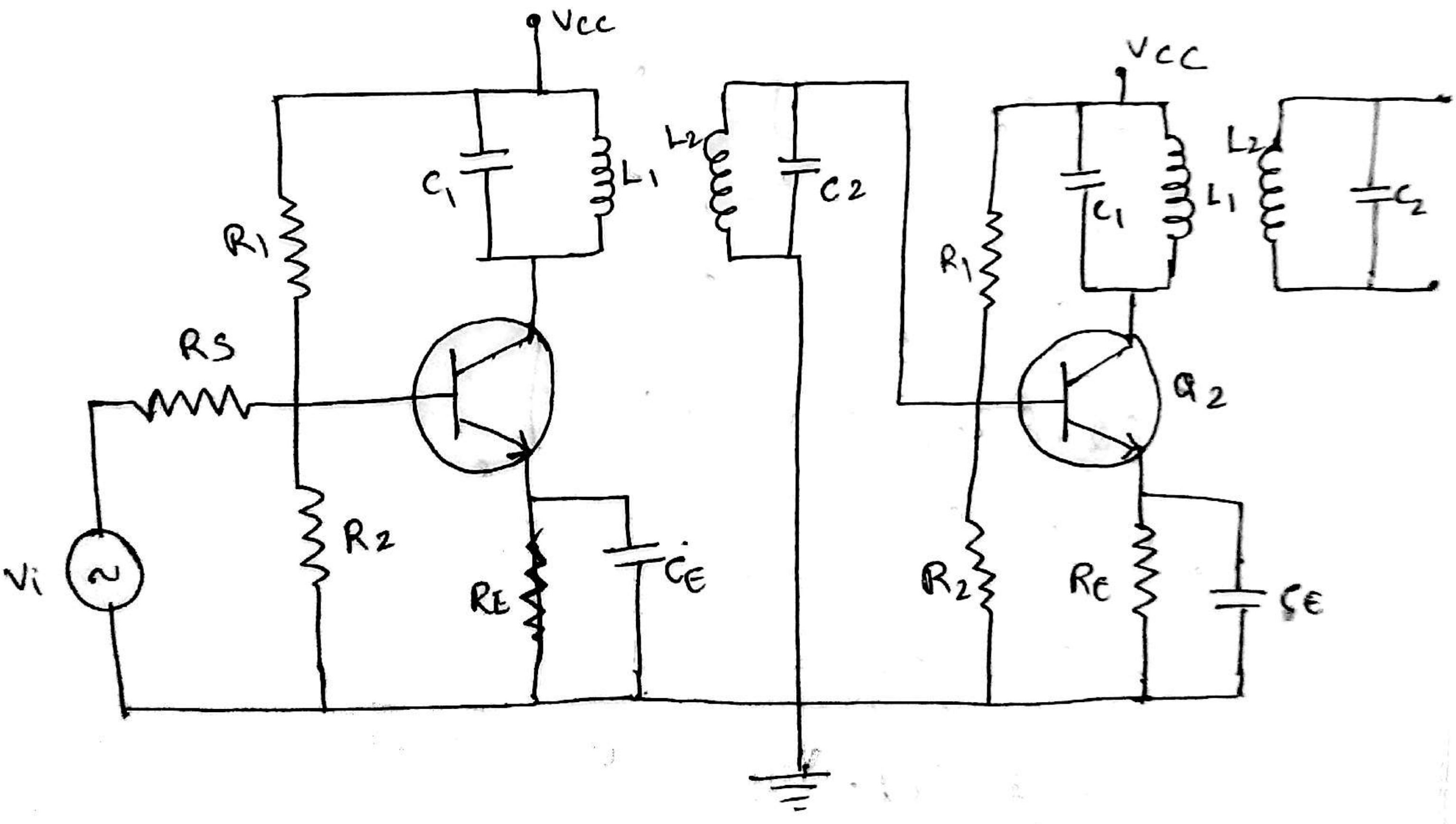
$$= f_2 - f_1$$

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

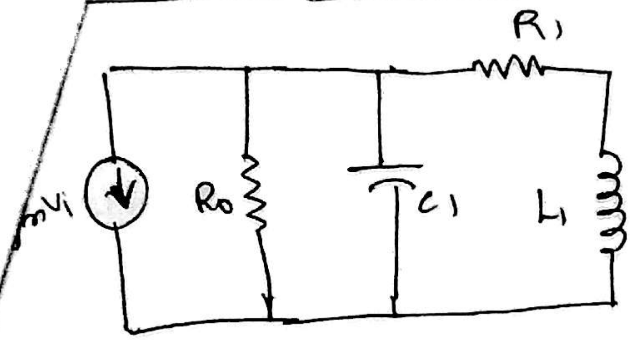
$$\boxed{BW_n = BW_1 \sqrt{2^{1/n} - 1}}$$

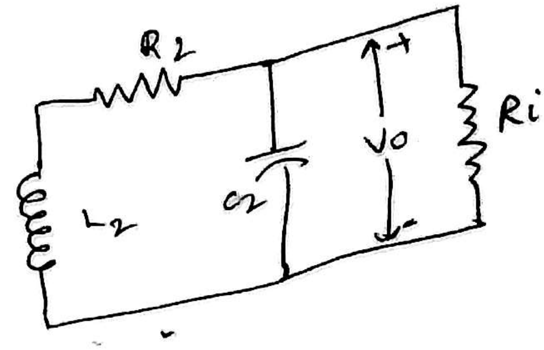
losses
1) copper loss
2) Eddy current loss
3) Hysteresis loss

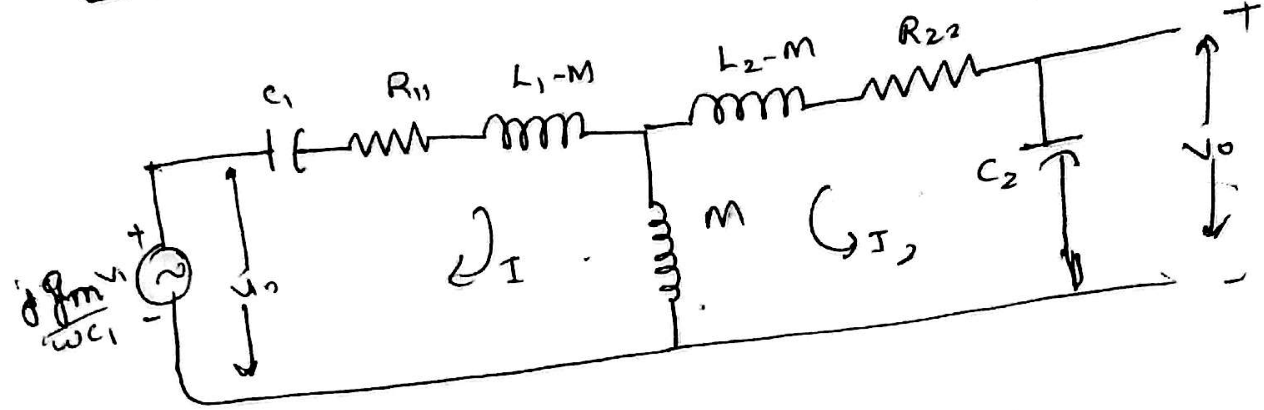
* DOUBLE TUNED AMPLIFIER:



ivalent circuits :







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

From equi circuit,

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

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

W.K.T, $$Q = \frac{\omega_r L}{R}$$

$$\therefore Q_1 = \frac{\omega_r L}{R_{11}} \quad \text{and} \quad Q_2 = \frac{\omega_r L_2}{R_{22}}$$

Q factor both circuits are kept same,

$\therefore Q_1 = Q_2 = Q$ and

resonant frequency,

$$\omega_r^2 = 1/L_1 C_1 = 1/L_2 C_2$$

The o/p voltage,

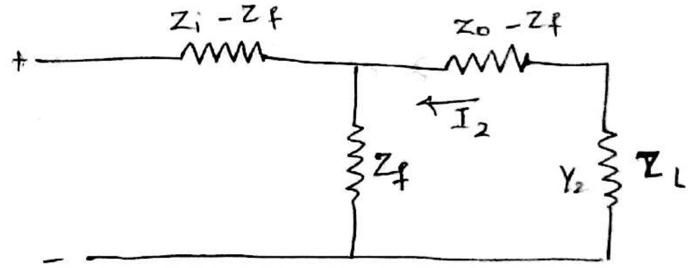
$$V_o = \frac{-j}{\omega_r C_2} I_2$$

Y_T (Transfer admittance)

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

where,
$$Z_{11} = \frac{V_1}{I_1} = Z_i - \frac{Z_f^2}{Z_o + Z_L}$$

and
$$A_i = \frac{I_2}{I_1} = \frac{-Z_f}{Z_o + Z_L}$$



$$Z_f = j\omega_r M$$
$$Z_i = R_{11} + j\left(\omega L_1 - \frac{1}{\omega C_1}\right)$$
$$Z_o + Z_L = R_{11} + j\left(\omega L_2 - \frac{1}{\omega C_2}\right)$$

By simplifying,
$$Z_f = j\omega_r M = j\omega_r K\sqrt{L_1 L_2}$$

Multiply Numerator and denominator by $\omega_r L_1$ for Z_i,
we get
$$Z_i = \frac{R_{11}\omega_r L_1}{\omega_r L_1} + j\omega_r L_1\left(\frac{\omega L_1}{\omega_r L_1} - \frac{1}{\omega C_1 \omega_r L_1}\right)$$

$$= \frac{\omega_r L_1}{Q} + j\omega_r L_1\left(\frac{\omega}{\omega_r} - \frac{\omega_r}{\omega}\right)$$

$$\left(\because Q = \frac{\omega_r L}{R_{11}} \text{ and } \frac{1}{\omega_r L} = \omega_r C\right)$$

$$= \frac{\omega_r L_1}{Q} + j\omega_o L_1 (2\delta)$$

$$\left(\because \frac{\omega}{\omega_r} - \frac{\omega_r}{\omega} = 1 + \delta - (1 - \delta) = 2\delta\right)$$

$$= \frac{\omega_r L_1}{Q_{11}} + (1 + j2Q\delta)$$

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

By Analysing, Z_i can be written as,

$$Z_o + Z_L = \frac{\omega_r L_2}{Q} + (1 + j\,2Q\delta)$$

Then,

$$Y_T = \frac{Z_f}{Z_f^2 - Z_i(Z_o+Z_L)}$$

$$= \frac{1}{Z_f - Z_i(Z_o+Z_L)/Z_f}$$

$$Y_T = \frac{1}{j\omega_r K\sqrt{L_1 L_2} - \left[\dfrac{\frac{\omega_r L_1}{Q}(1+j\,2Q\delta)\left\{\frac{\omega_r L_2}{Q}(1+j\,2Q\delta)\right\}}{j\omega_r K\sqrt{L_1 L_2}}\right]}$$

$$Y_T = \frac{KQ^2}{\omega_r\sqrt{L_1L_2}\,[4Q\delta - j(1+K^2Q^2 - 4Q^2\delta^2)]}$$

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

we get

$$V_o = \frac{-j}{\omega_r C_2}\,\frac{j g_m v_i}{\omega_r C_1}\left[\frac{KQ^2}{\omega_r\sqrt{L_1L_2}\,[4Q\delta - j(1-KQ^2 - 4Q^2\delta^2)]}\right]$$

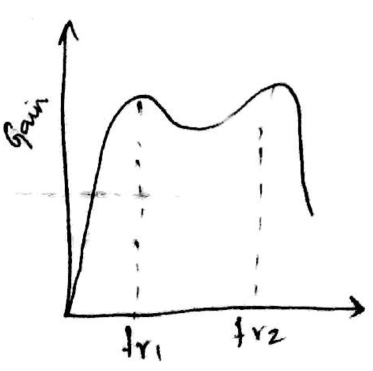
$$\because V_i = \frac{j g_m}{\omega C_1} v_i$$

$$A_v = \frac{V_o}{V_i}$$

$$= g_m \omega_r^2 L_1 L_2 \left[\frac{KQ^2}{\omega_r\sqrt{L_1L_2}\,[4Q\delta - j(1+K^2Q^2 - 4Q^2\delta^2)]}\right]$$

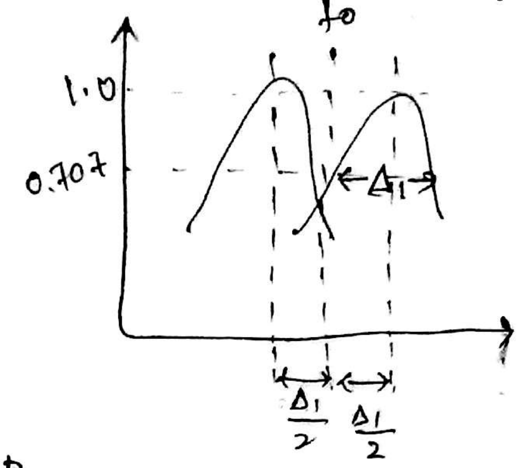
$$\because \left(\frac{1}{\omega_r C} = \omega_o L\right)$$

$$= \left[\frac{g_m \omega_r \sqrt{L_1L_2}\,KQ^2}{4Q\delta - j(1+K^2Q^2 - 4Q^2\delta^2)}\right]$$

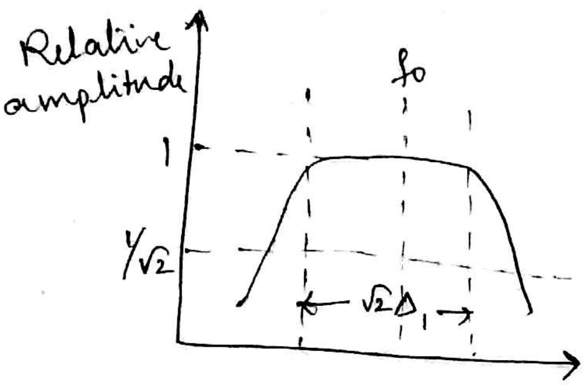


* Staggered tuned amplifier.

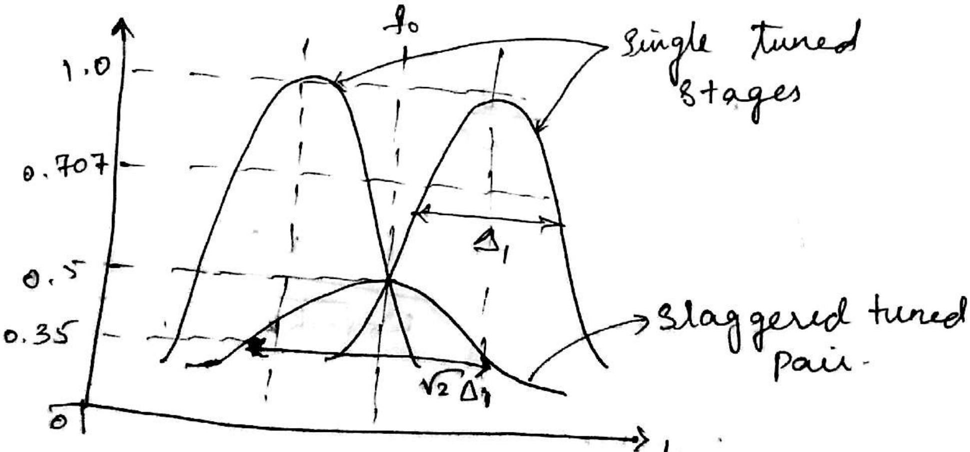
(a) Response of individual stages



(b) Overall Response of staggered pair



Response of individually tuned and staggered tuned pair:



$$\frac{A_v}{A_v\,(\text{resonance})} = \frac{1}{1 + 2j\,Q_{eff}\,\delta}$$

$$= \frac{1}{jx + 1}, \quad \text{where } x = 2\,Q_{eff}\,\delta$$

$f_{r_1} = f_r + \delta$ and $f_{r_2} = f_1 - \delta$

According to tuned frequencies the selectivity functions can be,

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

$$\frac{A_v}{A_v(\text{resonance})} = \frac{1}{1+j(x-1)}$$

Overall gain of these 2 stages is product of individual stages

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

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

$$= \frac{1}{2+2jx-x^2} = \frac{1}{(2-x^2)+(2jx)}$$

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

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

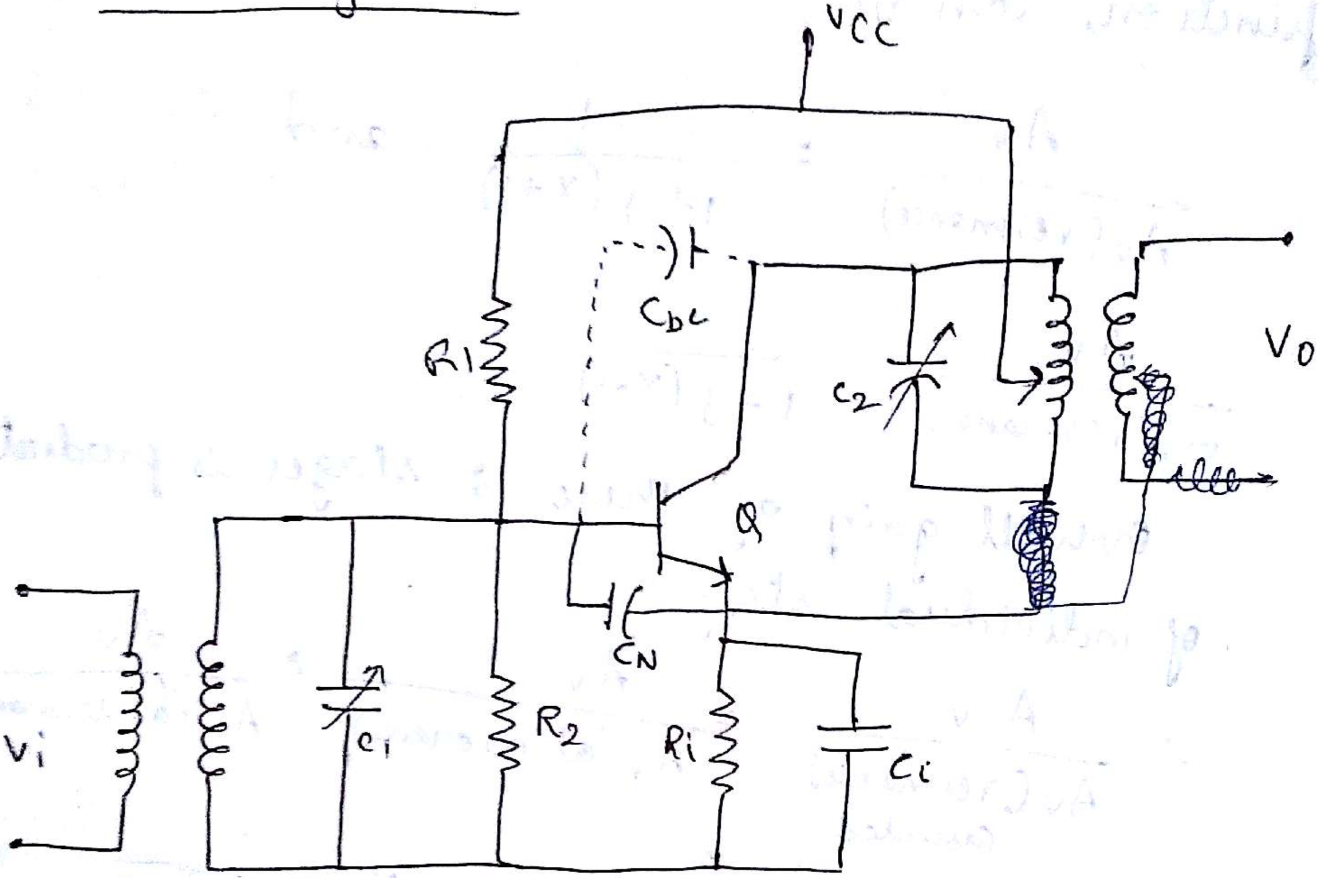
Substituting the value of x, we get.

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

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

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

* Neutralization



Types :

(1) Hazelthine Neutralization

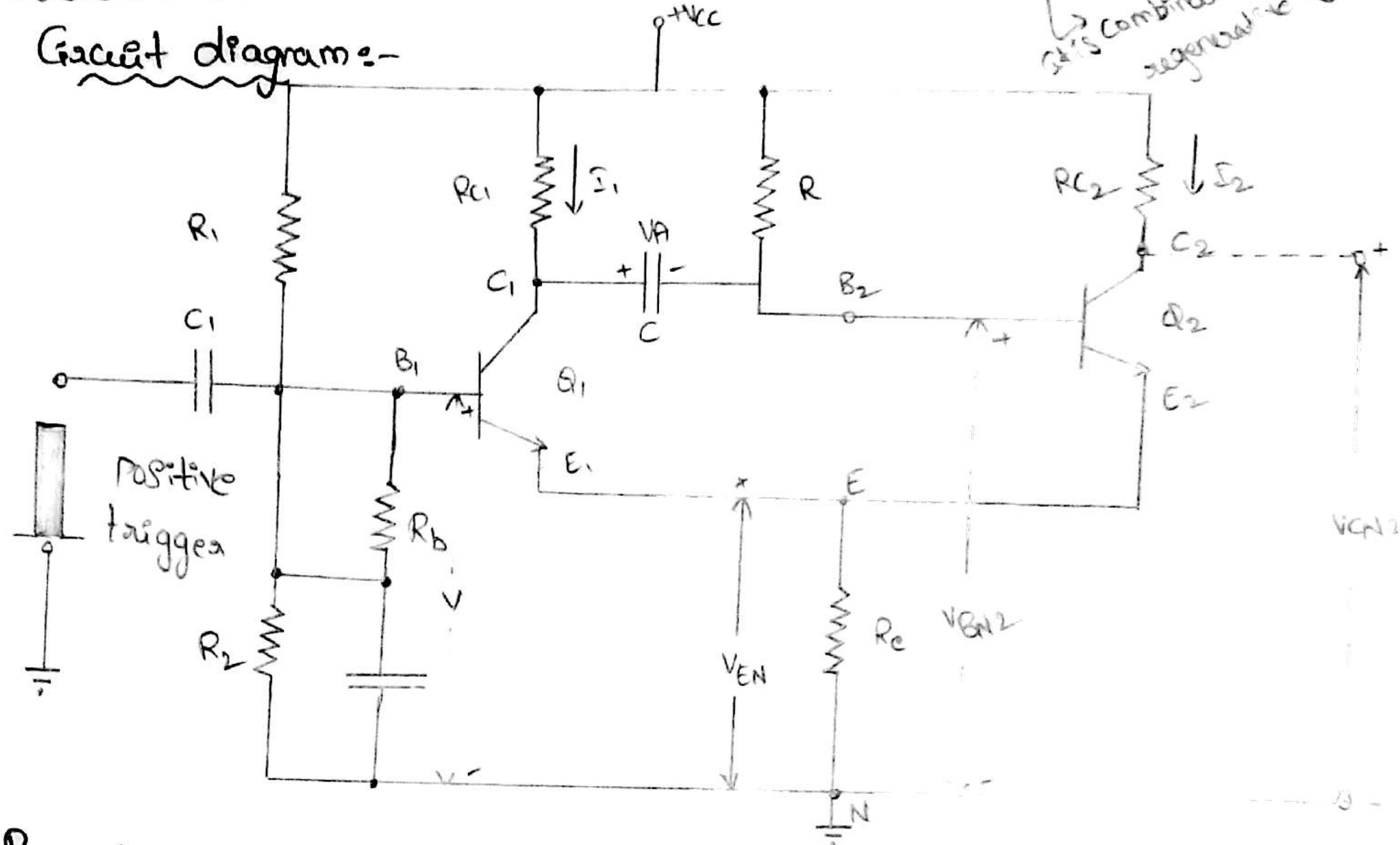
(2) Modified Hazeltine Neutralization.

(3) Neutralization using coil.

EMITTER COUPLED MONOSTABLE MULTIVIBRATOR :-

↳ It is combination of regenerative circuits

Circuit diagram :-



Procedure :-

The emitter terminals of both the transistors are coupled together hence called emitter coupled. The connection from collector C_2 to the base B_1 is absent. The feedback is provided through a common emitter resistance Re. A negative supply is not required.

Key point :-

+ve F/B o/p will be improving

→ As collector C_2 is not involved in the regenerative loop, it is ideal to take output voltage from this terminal.

The input trigger is connected to terminal B_1, which is also not connected to any other point. Thus trigger input source cannot load the circuit.

In collector coupled monoshot, I_1 controls the gate i.e
only one o/p
it is not possible to maintain I_1 stable. But in the emitter monoshot, I_1 can be stabilised with emitter resistance R_e. Hence, the control of T with I_1 can be achieved. This is possible because when Q_2 goes off, Q_1 goes ON and operates with sufficient emitter resistance.

The I_1 in turn can be controlled by biasing voltage V and it is observed that T varies linearly with V.

this, the applied

Key point:-

Thus emitter coupled monashot is used as a perfect gate wave generator whose gate width can be controlled easily and linearly with the help of an electric signal.

Wave forms:-

The waveform at the emitters is more important in this monoshot. Consider the mode of operation with Q_1 cut-off and Q_2 in saturation in the stable state.

In the stable state Q_1 is off and Q_2 is ON. The Q_2 is in saturation. It derives base drive from V_{cc} and resistance R. Due to its emitter current, it produces a voltage V_{EN2} across the resistance R_e. This voltage is more than the base potential of Q_1 which ensures that Q_1 remains off.

When the positive trigger is applied to the base of Q_1, then V_{B1} becomes more than V_{EN2} which drives Q_1 into conduction. Due to this, the collector voltage of Q_1 drops by $I_1 R_{C1}$. This negative step is applied to the base of Q_2 which makes Q_2 off. The capacitor C now charges through R from Vcc through ON transistor Q_1. The ON Q_1 developes potential drop of V_{EN1} across Re. This is the quasistable state with Q_1 ON and Q_2 off.

As the capacitor charges, the base B_2 of Q_2 becomes more positive. when the potential V_{BN2} becomes more than $V_{EN1} + V_\gamma$ then the transistor Q_2 starts conducting. Due to the regenerative feedback, Q_2 goes into saturation while Q_1 is cut-off. Thus the stable state with Q_1 off and Q_2 ON is acheived.

The waveforms are shown in the figure.

Extreme limits of V :-

In stable state Q_1 must be off. Hence there is a limit V_{max} for bias voltage V to be applied to base of Q_1.

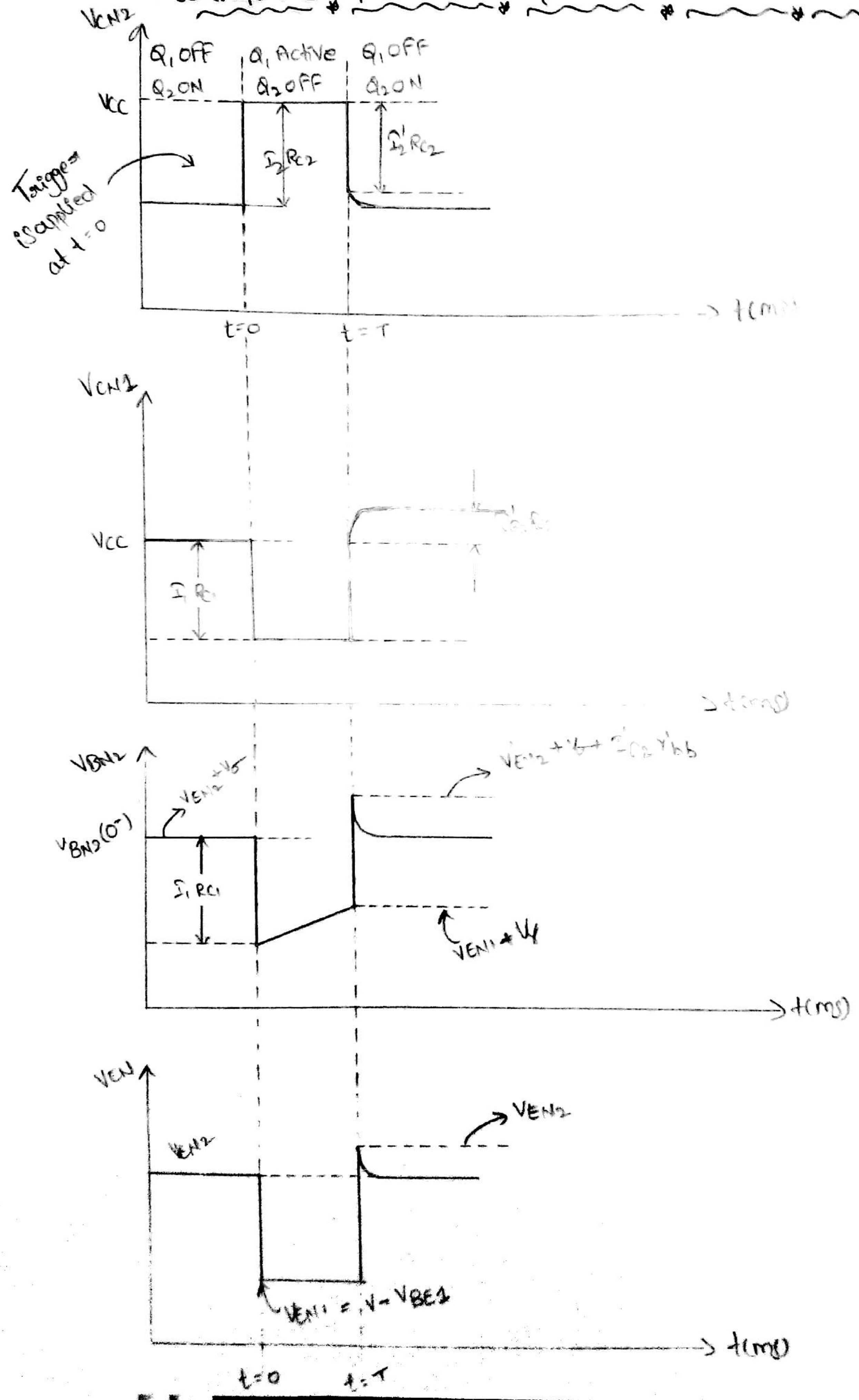
While when Q_1 is ON, then $I_1 R_{E1}$ drop must be large so as to cut-off Q_2. This puts a minimum limit V_{min} for bias voltage V.

Key point:-
The bias voltage V must be between V_{max} and V_{min}, for proper operation of the circuit

for Q_1 off in stable state, $V_{EN2} > V_{\gamma 1}$

$$\therefore \boxed{V_{max} = V_{EN2} + V_{\gamma 1}} \quad \text{—— (1)}$$

* Waveforms of emitter coupled monostable multivibrator *



Astable while for having minimum I_1 to cut-off Q_2 is quasi-stable state, V_{min} is given by,

$$V_{min} = (V_{EN1})_{min} + V_{BE1} \longrightarrow (2)$$

This equation (2), further can be rearranged as,

$$\boxed{V_{min} = V_{BE1} + \frac{V_{BN2}(0^-) - V_{\gamma2} + (R_{C1}/R)(V_{CC} - V_{\gamma2})}{1 + \left(\frac{R_{C1}}{R}\right) + \left(\frac{R_{C1}}{R_e}\right)[h_{FE}/(1+h_{FE})]}}$$

Gate width of emitter coupled monoshot :-

The expression of gate width can be derived using the basic relation,

$$V_C = V_f + (V_i - V_f)\, e^{-t/\tau} \quad \text{——— (1)}$$

Initial voltage   final voltage

The voltage V_{BN2} just after trigger is applied is given by,

$$V_{BN2}(0^+) = V_{BN2}(0^-) - I_1 R_{C1} \quad \text{——— (2)}$$

If Q_2 did not conduct, V_{BN2} would approach V_{CC}.

$\therefore$ V_{BN2} = Instaneous voltage at B_2

$$\therefore V_{BN2} = V_{CC} - [V_{CC} - V_{BN2}(0^-) + I_1 R_{C1}]\, e^{-t/\tau} \quad \text{——— (3)}$$

where $\tau = C(R + R_{C1})$

At $t = T^-$, just before pulse ends,

$$V_{BN2} = V_{EN1} + V_{\gamma2} \quad \text{——— (4)}$$

Equating equations (3) and (4),

$$V_{EN1} + V_{\gamma2} = V_{CC} - [V_{CC} - V_{BN2}(0^-) + I_1 R_{C1}]\, e^{-t/\tau} \quad \ldots \{t = T\}$$

Solving for T, the gate width is obtained as,

$$T = \tau \ln \left\{ \frac{V_{CC} - V_{BN12}(0^-) + I_1 R_{C1}}{V_{CC} - V_{EN1} - V_{\gamma 2}} \right\} \qquad (5)$$

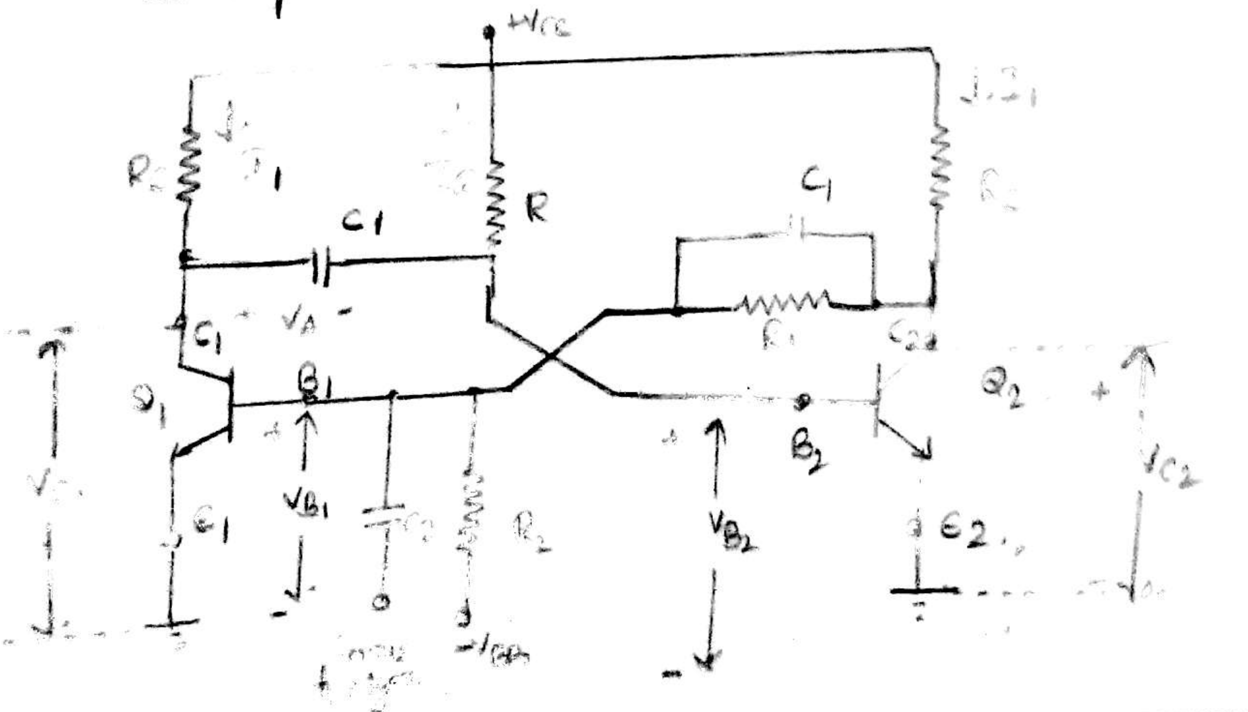
Key point :-

This time T Varies linearily with the bias voltage V. Due to this feature, it can be used as voltage to time Converter.

Collector Coupled Monostable Multivibrator.

The monostable multivibrator has one stable state. When an external trigger is applied, the circuit changes its state from stable to quasi-stable state. And then automatically, after some time interval T, the circuit returns back to the original normal the stable state. The time T is dependent on the circuit components.

The below figure shows the collector coupled monostable multivibrator circuit, which uses npn transistors. This circuit is also called collector to base coupled monostable



→ The Q_1 & Q_2 are identical npn transistors. The the two collector resistances are equal to R_c.

→ The output of Q_2 ie., collector of Q_2 is coupled to the base of Q_1 through a resistance R_1, which is shunted by a same capacitor C_1.

→ The capacitor C_1 is a speed-up capacitor required to make the transition fast & reduce the transition time. The collector of Q_1 is coupled to the base of Q_2 through a capacitor C.

→ Thus a d.c. coupling in bistable multivibrator is replaced by a capacitive coupling.

→ The resistance R at the input of Q_2 is returned to the supply voltage V_{cc}.

→ The value of R_2 and $-V_{BB}$ are chosen such that the transistor Q_1 is off, by reverse biasing it.

→ The transistor Q_2 is ON ie, in saturation. This is possible by forward biasing Q_2 with the help of V_{CC} and resistance R. Thus Q_2 ON and Q_1 off is the normal stable state of the circuit.

The positive triggering pulse is to be applied to the base of Q_1 through Capacitor C_2.

It must be noted that the triggering is unsymmetrical and is to be applied to one transistor only and not to both simultaneously.

→ When a positive trigger of sufficient magnitude and duration is applied to the base of Q_1, the transistor Q_1 starts Conducting.

→ Due to this, voltage at its Collector V_{C_1} decreases.

→ This is Coupled to base of Q_2 through C.

→ But voltage across Capacitor Cannot change instantaneously.

→ Hence decrease in V_{C_1} directly Cause a decrease in the base voltage of Q_2. ie., V_{B_2}.

→ The drop in voltage is about $I_1 R_C$. This decreases the forward bias of Q_2 and hence Collector Current I_2 decreases.

→ Thus the Collector voltage of Q_2 increases which is applied to the base of Q_1, through R_1.

→ This further increases the base potential of Q_1 & Q_1 is quickly driven in Saturation and at the same time, the transistor Q_2 gets ~~drawn~~ driven to

Cutoff.

→ This is quasi-stable of the circuit.

→ The circuit will remain in this quasi-stable sta
for only a finite time T.

→ In the quasi-stable state, the capacitor C starts charging through the path Vcc, R and ON transistor Q_1.

→ As it starts charging towards Vcc, the base of Q_2 experiences rise in voltage.

→ When this voltage becomes more than Cut in Voltage V_γ of Q_2, the Q_2 starts Conducting And due to the regenerative action, Q_1 is turned OFF.

→ Thus the circuit returns back to its stable state.

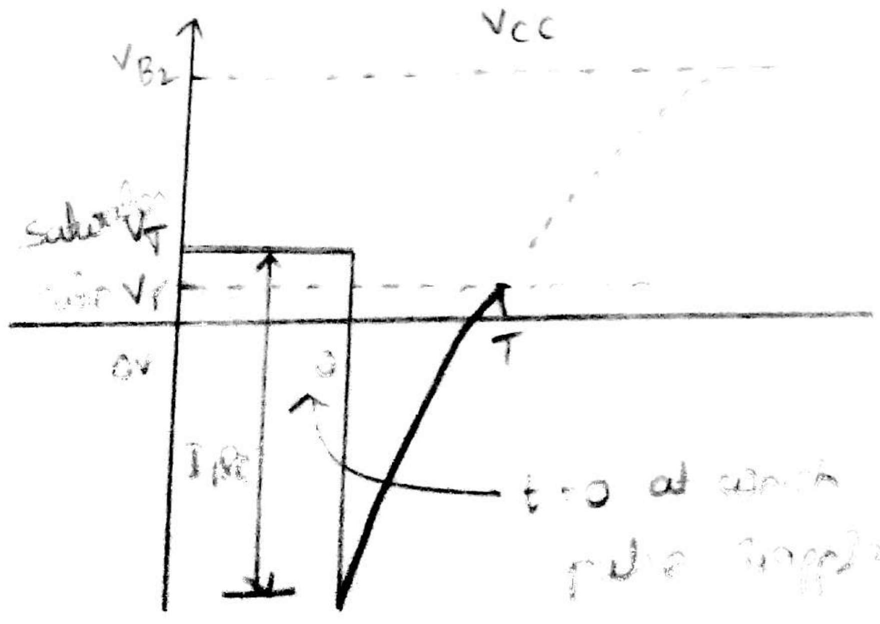
Pulse width of collector Coupled Monostable Multivibrator

The pulse width ~~of~~ the time for which the circuit remains in the quasi-stable state. It is also called gate width and denoted as T.

Derivation of Pulse width

To derive its expression, Consider the voltage variation at base of Q_2.

Initially Q_2 is in saturation and hence $V_{B2} = V_{BE_2}(sat) = V_T$ which is about 0.8 V for silicon transistor. When the pulse is applied at $t=0$, then at $t=0^+$, as capacitor voltage cannot change instantaneously, the voltage V_{B2} decrease by $I_1 R_C$. Then. Capacitor charges exponentially hence V_{B2} also increases exponentially, whose final value at $t \to \infty$ is V_{CC}. But when V_{B2} becomes equal to V_γ then Q_2 starts conducting and circuit comes back to stable state. This is the time T at which transition occurs. The graph of V_{B2} against time.



V_{B2} Vs t graph

To write equation for exponential charging of this capacitor we can write,

$t = 0^+$, $v_i = V_\gamma - I_1 R_C$.

$t = \infty$, $v_f = V_{CC}$

$$\boxed{v_C = v_f - (v_i - v_f)e^{-t/\tau}}$$

where τ = Time constant

$\therefore$ $$v_{B2} = V_{CC} - (V_{CC} - V_\gamma + I_1 R_C)e^{-t/\tau} \rightarrow ①$$

at $t = T$, $$v_{B2} = V_\gamma \rightarrow ②$$

Substituting eq ② into eq ① & solving for T we get

$$T = \tau \ln\left[\frac{V_{CC} + I_1 R_C - V_\gamma}{V_{CC} - V_\gamma}\right] \rightarrow ③$$

where $V_\gamma = 0.3v$ for germanium

$\Rightarrow 0.8v$ for silicon.

When Q_1 is in saturation under quasi-stable state we can write,

$$V_{C1} = V_{CE(sat)} \rightarrow ④$$

$\therefore$ $$I_1 R_C = V_{CC} - V_{CE(sat)} \rightarrow ⑤$$

Substituting in eq ③.

$$T = \tau \ln\left[\frac{V_{CC} + V_{CC} - V_{CE(sat)} - V_{BE(sat)}}{V_{CC} - V_\gamma}\right] \rightarrow ⑥$$

This is because, $V_T = V_{BE}(sat)$

$$\therefore T = \tau \ln\left[\frac{2V_{CC} - V_{CE}(sat) - V_{BE}(sat)}{V_{CC} - V_\gamma}\right] \rightarrow (7)$$

$$\therefore T = \tau \ln\left\{\frac{2\left[V_{CC} - \left(\frac{V_{CE}(sat) + V_{BE}(sat)}{2}\right)\right]}{V_{CC} - V_\gamma}\right\} \rightarrow 8.$$

$$\therefore T = \tau \ln(2) + \tau \ln\left\{\left[\frac{V_{CC} - \left(\frac{V_{CE}(sat) + V_{BE}(sat)}{2}\right)}{V_{CC} - V_\gamma}\right]\right\} \rightarrow (9)$$

At room temperature,

$V_{CE}(sat) + V_{BE}(sat) \cong 2V_\gamma$.

Substituting in eqn (9)

$$T = \tau \ln(2) + \tau \ln(1)$$

$$\therefore \tau \Rightarrow \tau \ln(2) \rightarrow (11)$$

The time constant τ for the charging path of capacitor C is RC.

$$\tau = RC \rightarrow (12)$$

$$\tau = 0.69RC \rightarrow (13)$$

Thus the gate width is ~~depart ment~~ depen on temperature, as $V_{BE}(sat)$, $V_{CE}(sat)$ and V_γ depend on the temperature.

One remedy for this temperature effect is to connect R not to V_{CC} but to source V whose value decreases as the temperature decreases, thus compensating the effect of temperature.

Wave shapes of Monostable Multivibrator.

To obtain the waveforms, let us assume that the external pulse is applied at $t=0$ & reverse transition from quasi-stable state occurs at $t=T$.

The stable state:-

The transistor Q_1 is OFF and Q_2 is ON in the stable state. The stable state currents and voltages can be obtained as shown earlier for bistable multivibrator circuit. The Q_2 is in saturation hence for Q_2,

$$\therefore V_{B_2} = V_{BE}(sat) = V_\sigma \quad \rightarrow (14)$$

$$V_{C_2} = V_{CE}(sat) \quad \rightarrow (15)$$

The Q_1 is OFF hence as current is zero we can write,

$$V_{C_1} = V_{CC} \rightarrow (16)$$

The base voltage of Q_1 can be obtained by superposition principle as shown by equation (2) in case (ii) ~~of~~ €

$$V_{B_1} = \left.\frac{-V_{BB}R_1}{R_1+R_2}\right|_{V_{C_2}=0} + \left.\frac{V_{C_2}R_2}{R_1+R_2}\right|_{V_{BB}=0} \cong V_f \rightarrow (17)$$

To have Q_1 off, $|V_f| \leq 0.1V$ for germanium transistor.

The quasi-stable state:- When pulse is applied at $t=0$ Q_2 becomes OFF and Q_1 becomes ON. The voltages at V_{C_1} & V_{B2} drop instantaneously by the amount I_1R_C where I_1 is current drawn by Q_1 when it starts conducting. Then Q_1 gets driven into saturation hence,

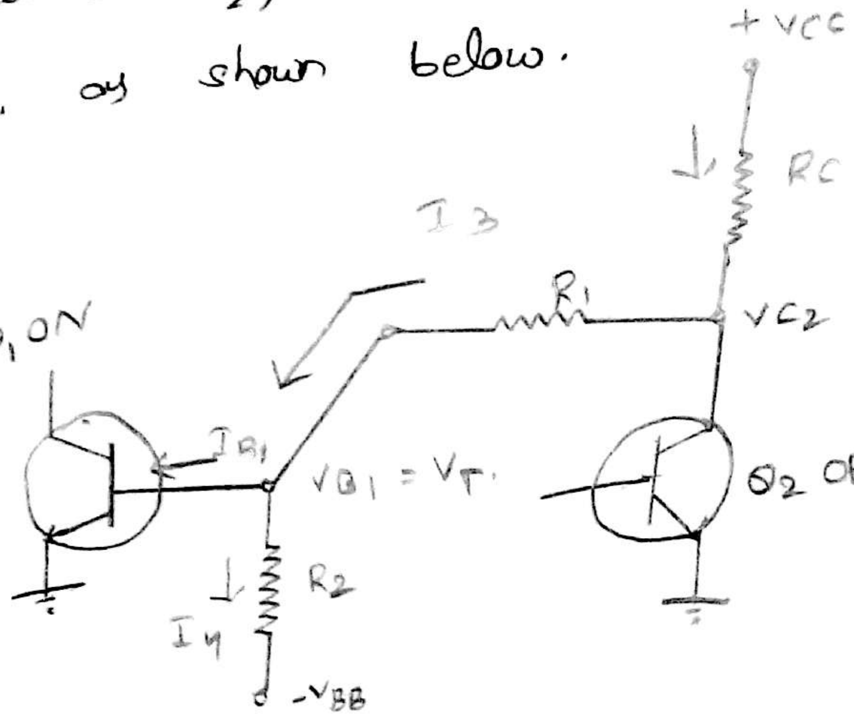
$$V_{B_1} = V_{\gamma} \rightarrow (18)$$

$$V_{C_1} = V_{ce(sat)} \rightarrow (19)$$

&

$$I_1R_C = V_{CC} - V_{ce(sat)} \rightarrow (20)$$

To Calculate V_{C_2}, Consider the Equivalent ckt to V_{CC} this 0 < t < T, as shown below.



By using the Superposition principle we can write,

$$V_{C_2} = \left.\frac{V_{CC}R_1}{R_1+R_C}\right|_{V_B=0} + \left.\frac{V_r R_C}{R_1+R_C}\right|_{V_{CC}=0} \rightarrow (21)$$

As Capacitor starts charging, the voltage at base of Q_2 rises exponentially as discussed earlier, towards V_{CC}. This Continues till V_{B_2} becomes equal to cut in voltage V_r at $t = T$.

Waveforms for $t \geq T$: At $t = T$, reverse transition occurs. Thus Q_1 is cutoff and Q_2 starts conducting at $t = T^+$. Thus V_{C_2} drops instantly to $V_{CE(sat)}$ & V_{B_1} returns to V_F. The voltage V_{C_1} rises abruptly as Q_1 becomes off. This rise is such that V_{C_1} becomes almost equal of V_{CC}.

this sudden increase in V_{C1} is applied to base of Q_2 & Q_2 suddenly gets driven into saturation. Hence at $t = T^+$ an overshoot occurs at base of Q_2 in V_{B2}. This overshoot decreases exponentially as the capacitor C recharges bcoz of base current.

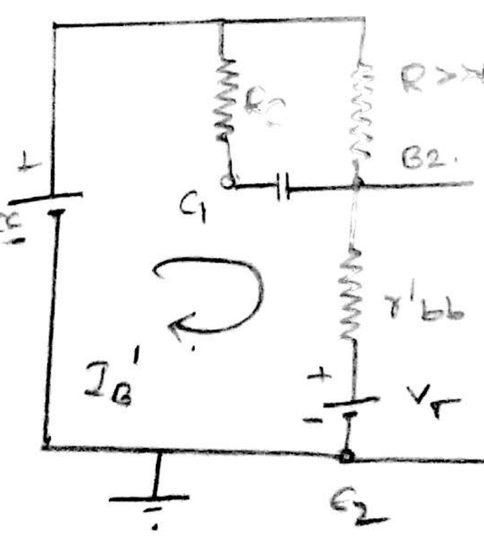
Magnitude of the overshoot in V_{B2}:-

The equivalent circuit at $t = T^+$ is shown in below figure. In this circuit, the input circuit of Q_2 is replaced by base spreading resistance r'_{bb} in series with base saturation voltage V_σ.

The base current at $t = T^+$ is denoted as I_B' while current through R is neglected as $R >> R_C$

$$V_{B'E} = I_B' r'_{bb} + V_\sigma \rightarrow (22)$$

$$V_{C1} = V_{CC} - I_B' R_C \rightarrow (23)$$



Thus the overshoot at $t = T^+$ in the voltage V_{B2} is now can be denoted as δ and is given by,

$$\delta = V'_{BE} - V_\gamma = \text{Overshoot}$$

$$\therefore \delta = I_B' r'_{bb} + V_\sigma - V_\gamma \rightarrow (24)$$

while the overshoot in the collector voltage of Q_1 ie, in V_{C1}

is denoted as δ' and given by,

$$\delta' = V_{C_1} - V_{CE(sat)}$$

$$\delta' = V_{CC} - I_B' R_C - V_{CE(sat)} \longrightarrow (25)$$

But the voltage. V_{C_1} is applied to base of Q_2.

Equating eq (24) & (25) we get

$$I_B' r'_{bb} + V_\sigma - V_\gamma = V_{CC} - I_B' R_C - V_{CE(sat)}$$

$$\therefore I_B'(R_C + r'_{bb}) = V_{CC} - V_{CE(sat)} - V_\sigma + V_\gamma$$

$$\boxed{I_B' = \frac{V_{CC} - V_{CE(sat)} - V_\sigma + V_\gamma}{R_C + r'_{bb}}} \rightarrow (26)$$

After $t = T^+$, the V_{B_2} decreases exponentially to its steady state value V_σ. The time constant with which it decays is given by.

$$\tau' = (R_C + r'_{bb}) C \longrightarrow (27)$$

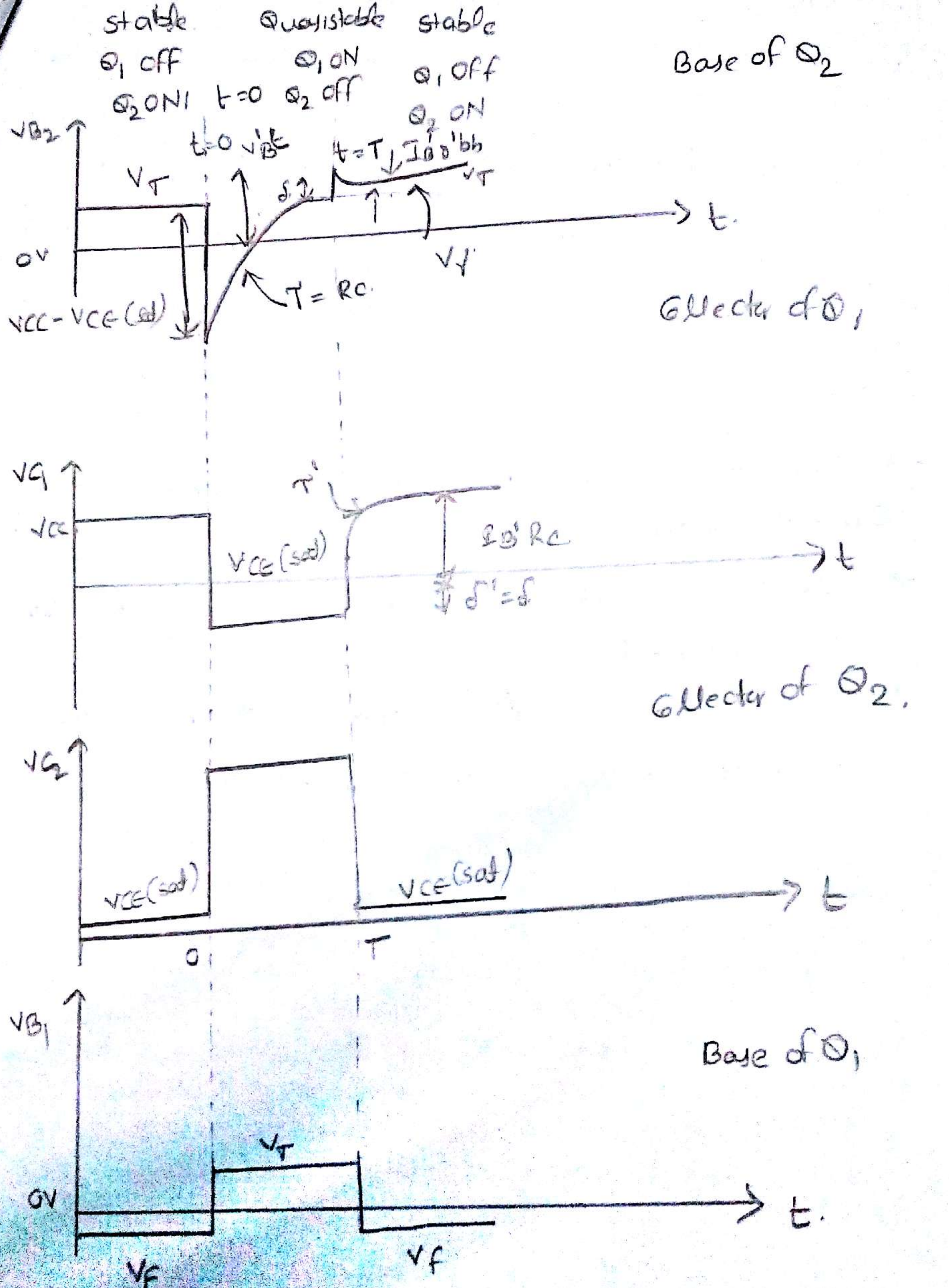
Note that if V_{CC} is large compared with the junction voltages, then.

$$\therefore \boxed{I_B' \approx \frac{V_{CC}}{R_C}} \text{ if } V_{CC} \text{ is large.} \rightarrow (28)$$

Similar to overshoot in V_{C_1} & V_{B_2}, there exists an undershoot in V_{C_2} and V_{B_1} at $t = T^+$ but these undershoots are very small & hence usually neglected & not shown in wave forms.

The Complete waveforms Considering above discussion are shown in.

waveforms of Collector Coupled monostable multivibrator



The typical n-p-n transistor junction voltages at 25°C are given here. The table may be reffoured if the voltages are not given in problem. All the voltages are in volts.

	$V_{CE}(sat)$	$V_{BE}(sat) = V_{\sigma}$	$V_{BE}(active)$	$V_{BE}(cutin) = V_{\gamma}$	$V_{BE}(cutoff)$
Si	0.3	0.7	0.6	0.5	0.0
Ge.	0.1	0.3	0.2	0.1	-0.1

Applications

→ Used to produce rectangular waveform & hence can be used as gating circuit.

→ Used to introduce time delays as gate width is adjustable

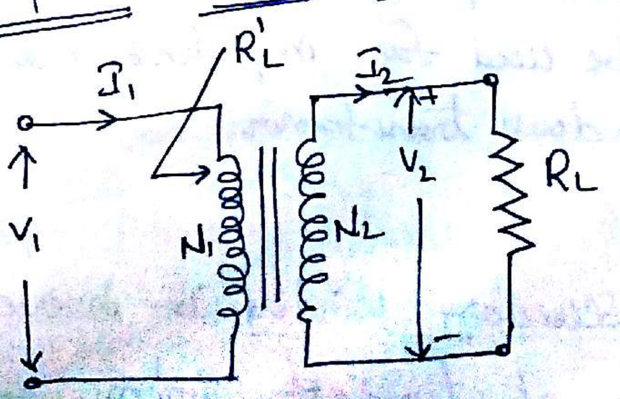
→ Used to generate uniform width pulses from a variable width input pulse train.

AEC

Transformer Coupled class A Amplifier:-

* For maximum power transfer to the load, the impedence matching is necessary.
* For loads like loud speaker, having low impedence values, impedence matching is difficult using directly coupled amplifiers circuit.
* This is because loud speaker resistance is in the range of 3 to 4 ohms to 16 ohms while the output impedence of series fed directly coupled class A amplifier is very much high.
* This problem can be eliminated by using transformer to deliver power to the load. The transformer is called an output transformer. and amplifier is called transformed coupled class A amplifier.

Properties of transformer:-



* Consider a transformer as shown in the fig. which is connected to a load resistance R_L.
* while analysing transformer, is is assumed that the transformer is ideal and there are no loses in transformer. Similarly the winding resistances are assumed to be zero.

N_1 = Number of turns on primary
N_2 = Number of turns on secondary
V_1 = Voltage applied to primary
V_2 = Voltage on secondary

I_2 = primary Current.

1) Turns ratio:-

The ratio of no. of turns on secondary to the no. of turns on primary is called turns ratio and is denoted by n.

$$\boxed{n = \frac{N_2}{N_1}}$$

* Sometimes it is specified as $\frac{N_2}{N_1} : 1$ (or) $\frac{N_1}{N_2} : 1$

2) Voltage transformation:-

The transformer transforms the voltage applied on one side to other side proportional to turns ratio. The transformer can be step up (or) step down transformer.

$$\boxed{\frac{V_2}{V_1} = \frac{N_2}{N_1} = n}$$

In amplifier analysis, load impedance is going to be small. And the transformer is to be used for impedance matching. Hence it has to be a step down transformer.

3) Current transformation:-

The current in the secondary winding is inversely proportional to the number of turns of windings.

$$\boxed{\frac{I_2}{I_1} = \frac{N_1}{N_2} = \frac{1}{n}}$$

(10) (11)

Impedence transformation:-

As current and voltage get transformed from primary to secondary, an impedence seen from either side also changes. The impedence of the load on secondary is R_L.

$$R_L = \frac{V_2}{I_2} \quad \text{and} \quad R_L = \frac{V_1}{I_1}$$

$$V_1 = \frac{N_1}{N_2} V_2 \quad \text{and} \quad \frac{I_1}{I_2} = \frac{N_2}{N_1}$$

$$I_1 = \frac{N_2}{N_1} I_2$$

$$R_L' = \frac{\frac{N_1}{N_2} V_2}{\frac{N_2}{N_1} I_2} = \left[\frac{N_1}{N_2}\right]^2 \times \frac{V_2}{I_2}$$

$$\boxed{R_L' = \frac{R_L}{n^2}}$$

The R_L' is reflected impedence.

Problem

The load of 4Ω is connected to the secondary of a transformer having primary turns of 200 and secondary turns of 20. Calculate the reflected load impedence on primary.

Solution Given that

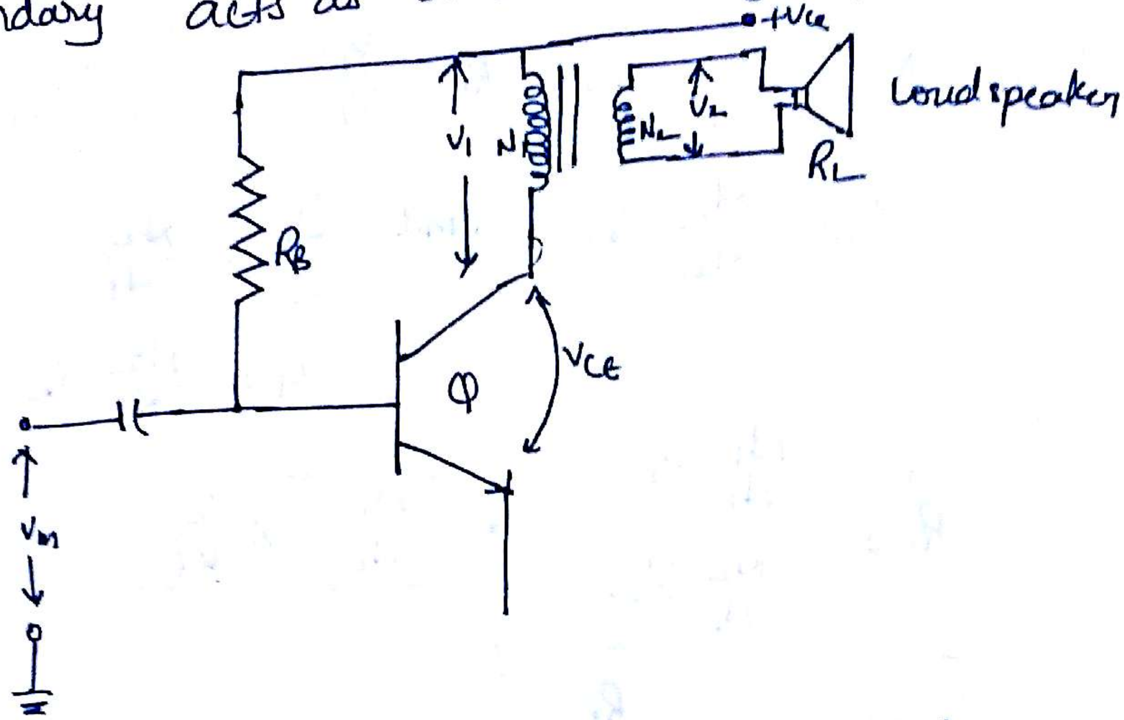
$$R_L = 4\Omega \quad N_1 = 200 \quad N_2 = 20$$

$$n = \frac{N_2}{N_1} = \frac{20}{200} = 0.1$$

$$R_L' = \frac{R_L}{n^2} = \frac{4}{(0.1)^2} = 400\Omega$$

Circuit diagram of transformer Coupled Amplifier:

The basic circuit of a transformer Coupled amplifier shown in below fig. The loudspeaker Connected to the secondary acts as a load having impedance of $R_L \Omega$.



The transformer used is a step down transformer with turns ratio as $n = \frac{N_2}{N_1}$.

DC operation:

* It is assumed that winding resistances are 0Ω. Hence for d.c. purposes the resistance is 0Ω.
* There is no voltage drop across the primary winding of the transformer.
* The slope of the d.c load line is reciprocal of the d.c resistance in collector circuit, which is zero in this case.
* Hence slope of d.c load line is ideally infinite. The d.c load line is vertically straight line

Apply KVL at Collector side

$$V_{CC} - V_{CE} = 0$$

$$V_{CC} = V_{CE}$$

This is dc bias voltage V_{CEQ} for the transistor

$$\therefore \quad V_{CC} = V_{CEQ}$$

(*) The intersection of d-c load line and the base current set by the circuit is the Quiescent operating point. The d-c load line as shown in fig (1).

DC power input:

The d-c power input is provided by supply voltage with no signal input. the dc current drawn is the Collector bias current I_{CQ}

$$\boxed{P_{DC} = V_{CC} I_{CQ}}$$

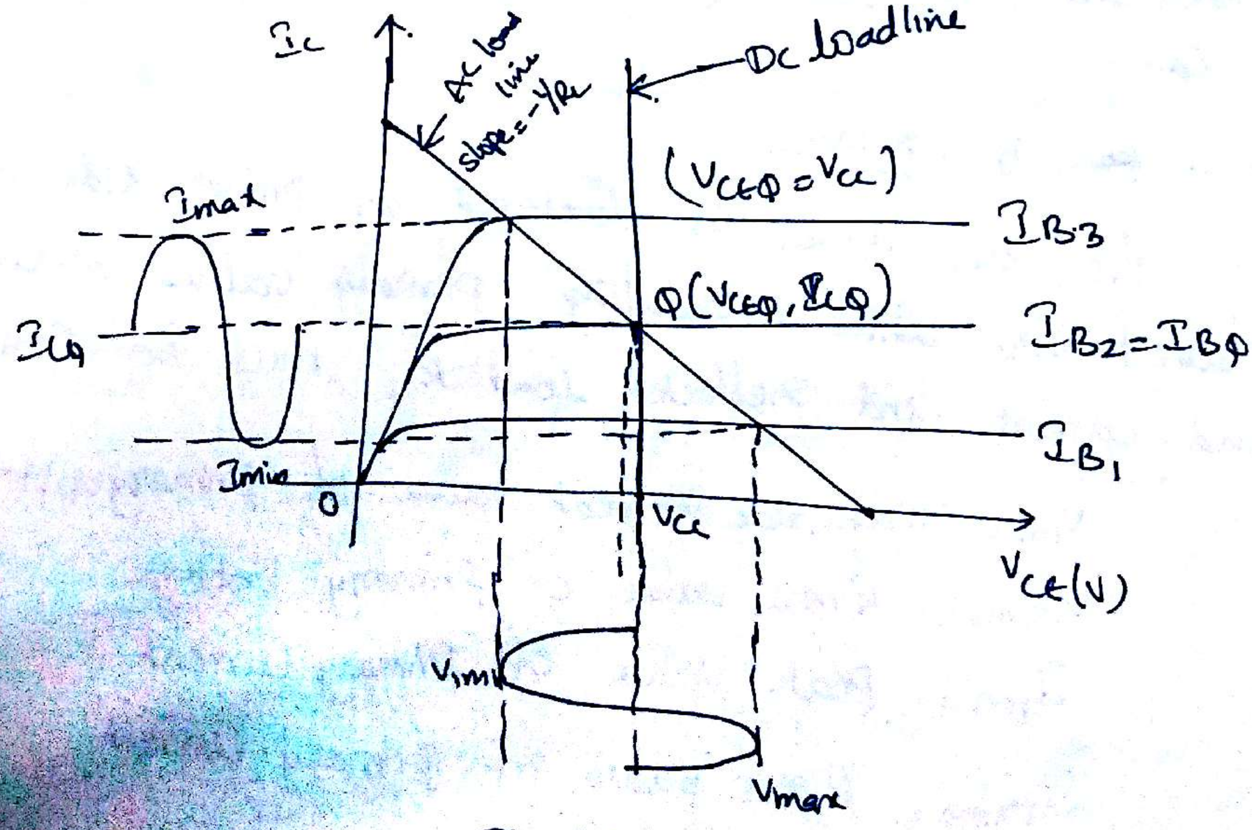


Fig (1).

(14) (15)

a.c power developed on primary

$$P_{ac} = V_{1rms} I_{1rms}$$

$$P_{ac} = I^2_{rms} R'_L \quad \left[\because \begin{matrix} V = IR \\ V_{rms} = I_{rms} R_L \end{matrix} \right]$$

$$= \frac{V^2_{1rms}}{R'_L}$$

$$\boxed{P_{ac} = \frac{V_m I_m}{2} = \frac{I^2_{1m} R'_L}{2} = \frac{V^2_{1m}}{2R'_L}}$$

a.c power developed on secondary

$$P_{ac} = V_{2rms} I_{2rms} = I^2_{2rms} R_L = \frac{V^2_{2rms}}{R_L}$$

$$\boxed{P_{ac} = \frac{V_{2m} I_{2m}}{2} = \frac{I^2_{2m} R_L}{2} = \frac{V^2_{2m}}{2R_L}}$$

* power delivered on primary is same as power delivered to the load on secondary assuming ideal transformer.
* For practical circuit, the transformer can not be ideal. Hence power delivered to load is slightly less than primary.

In class A amplifier

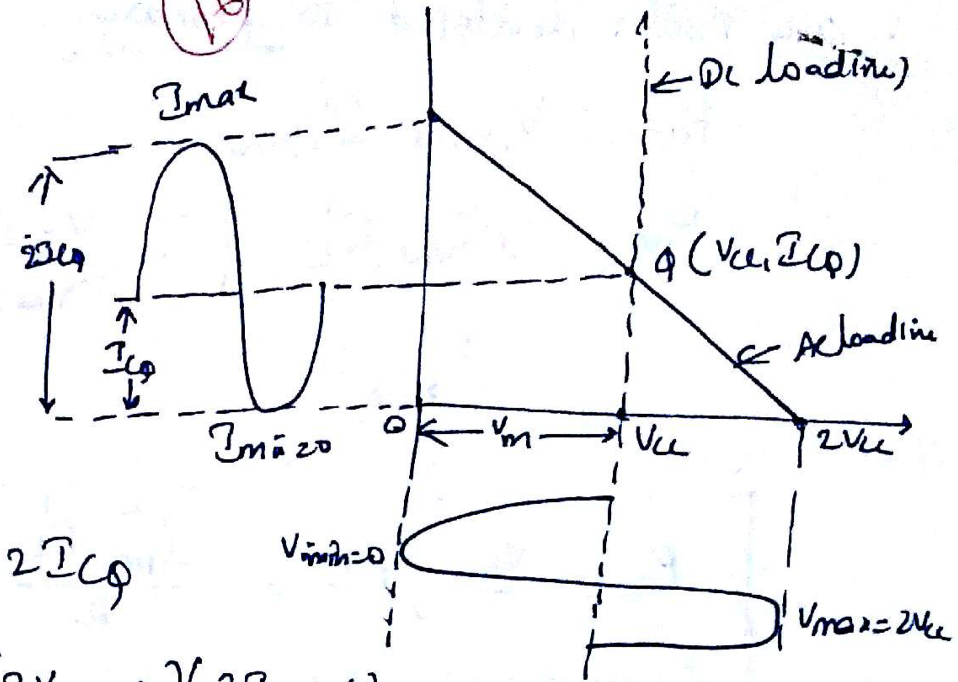
$$P_{ac} = \frac{(V_{max} - V_{min})(I_{max} - I_{min})}{8}$$

Efficiency!

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

(16)

Maximum Efficiency



$V_{min} = 0$

$V_{max} = 2V_{cc}$

$I_{min} = 0 \quad I_{max} = 2I_{CQ}$

$$\% \eta_{max} = \frac{(2V_{cc} - 0)(2I_{CQ} + 0)}{8 V_{cc} I_{CQ}} \times 100$$

$$= \frac{4 V_{cc} I_{CQ}}{8 V_{cc} I_{CQ}} \times 100$$

$$= 50 \%$$

Power dissipation:

Power dissipation is the difference between a.c power output and the d.c power input

$$\boxed{P_d = P_{DC} - P_{ac}}$$

Advantages

1) The Efficiency of operation is higher than directly Coupled amplifier.
2) The impedence matching required for maximum power transfer is possible

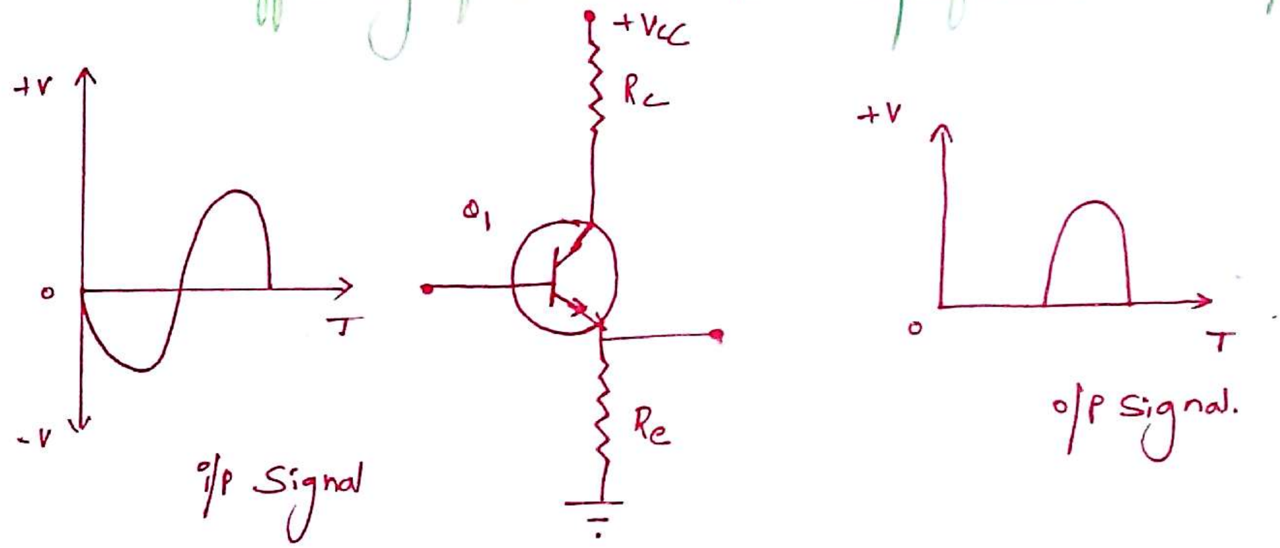
Disadvantages

1) Due to transformer, the circuit becomes bulkier, heavier and costlier Compared to directly Coupled circuit
2) the response frequency of circuit is poor.

Class B Amplifier:

Class 'B' Amplifier:

Class B amplifier is a type of amplifier where the active device transistor can conducts only for half cycle of the input signal. The conduction angle is 180° for a class 'B' Amplifier. The Quiescent point is fixed at x-axis (cut off), hence it dissipates less power and effeciency is improved. The maximum effeciency of a class B Amplifier is 78.5%



Advantages of class 'B' over class A

- In class B output has great power
- Effeciency is high
- negligible power loss at no i/p Signal

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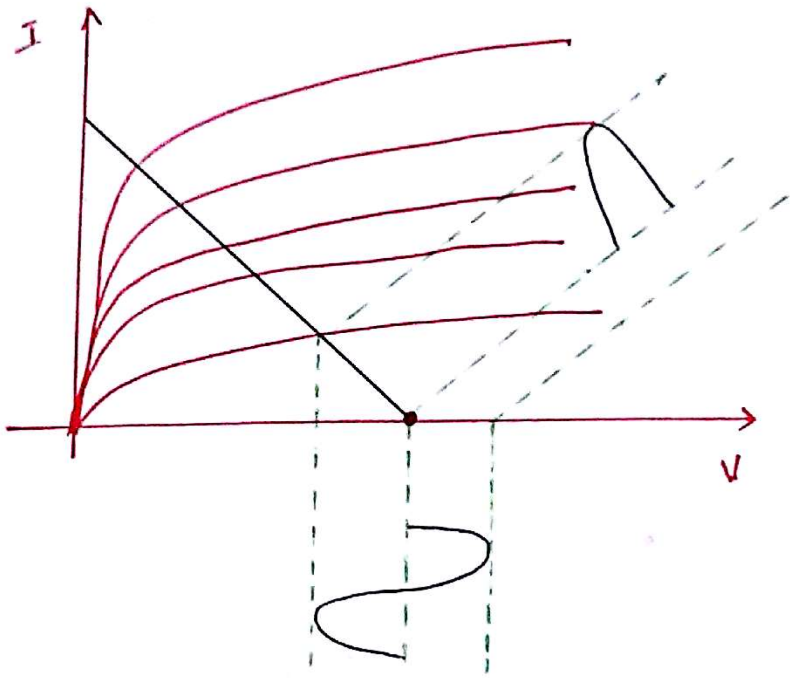
Effeciency:



The Effective Load Resistance is $R_L' = \left(\frac{N_1}{N_2}\right)^2 R_L$

$N_1 \rightarrow$ No. of primary turns

$N_2 \rightarrow$ No. of secondary turns

for Class B the dissipation at collector is zero in Quiescent State and increases with excitation.

Advantages of class B Amplifier:

→ High Effeciency when compared to the class A configurations

→ push - pull mechanism avoids even harmonics

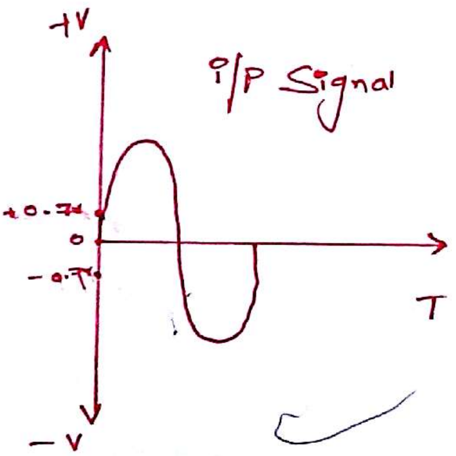
→ No DC components in the o/p

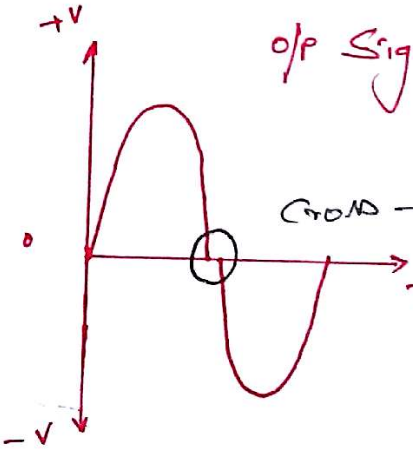
Disadvantages of Class 'B'-Amplifiers:

→ The major disadvantage is cross-over distortion
→ Coupling transformers increases cost & size
→ It is difficult to find ideal transformers
→ Transformer coupling is not practical in case of huge loads

Cross-over distortion:

Since the active elements starts conduction only after the input signal amplitude has risen above 0.7V the regions of the input signal where the amplitude is less than 0.7V will be missing in o/p signal & it is called as cross over distortion. The schematic representation is shown below.

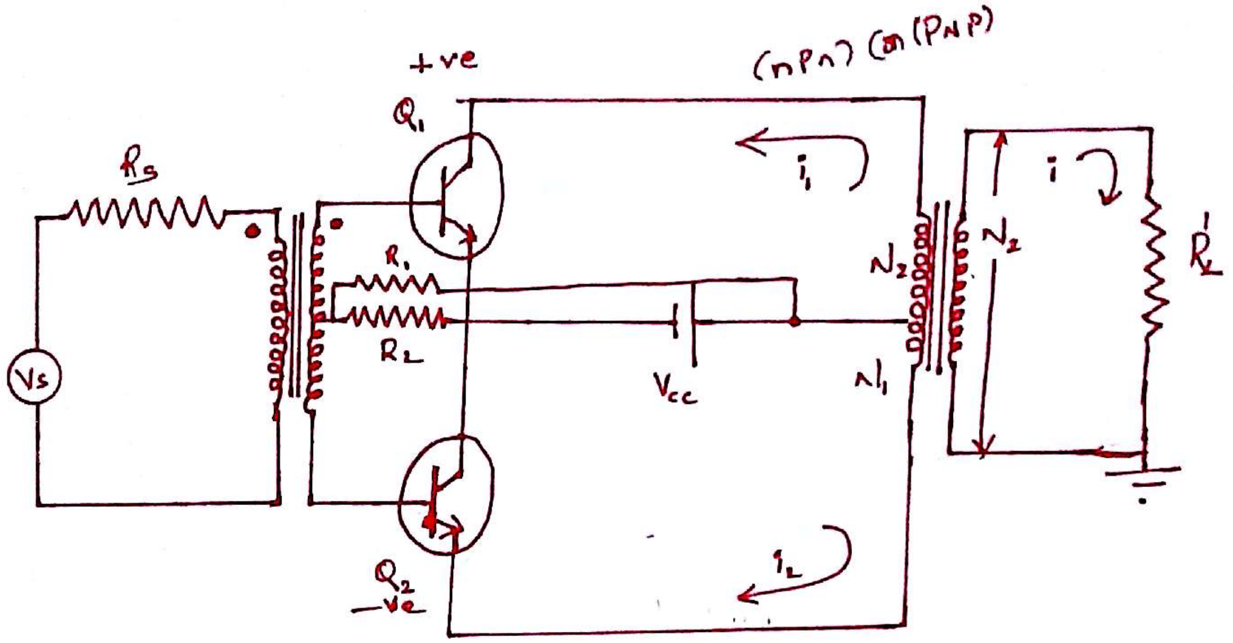




20

lock dia

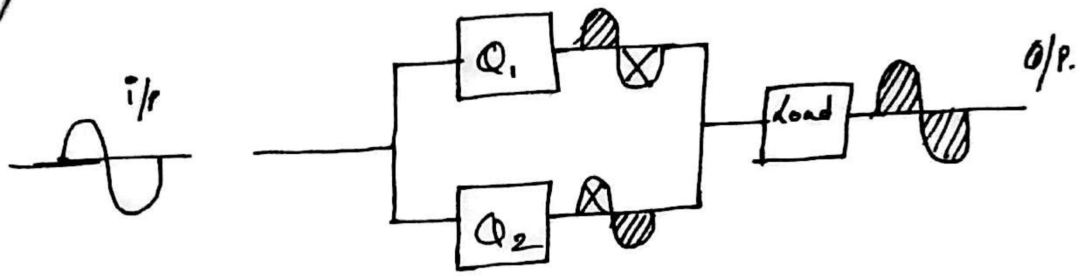
Push - Pull class 'B' - Amplifier:



In class B-Amplifier the o/p will be half of the input signal To obtain full wave form we use push-pull class B Amplifier. It consists of two transistors. Q1 & Q2 and two center tapped transistor one as i/p transformer known as driver transformer one to Resistance load at the o/p. An Vcc source is attached to the transistor in order to avoid the mismatching of transistors.

The transistors may be NPN (or) PNP transistor. The polarities in the transformer will be equal hence there is no phase shift. If the polarities are changed then a phase shift will occur.

Block diagram:



dc operation:

Each transistor act as a half wave rectifier.

If I_m is the peak current of the o/p then average value is I_m/π

$$\therefore I_{dc} = \frac{I_m}{\pi} + \frac{I_m}{\pi} = 2I_m/\pi$$

The dc power is

$$P_{dc} = V_{cc} \times I_{dc}$$
$$= 2/\pi \, V_{cc} \times I_m$$

Ac operation:

When Ac signal is applied to driver transformer to Q_1

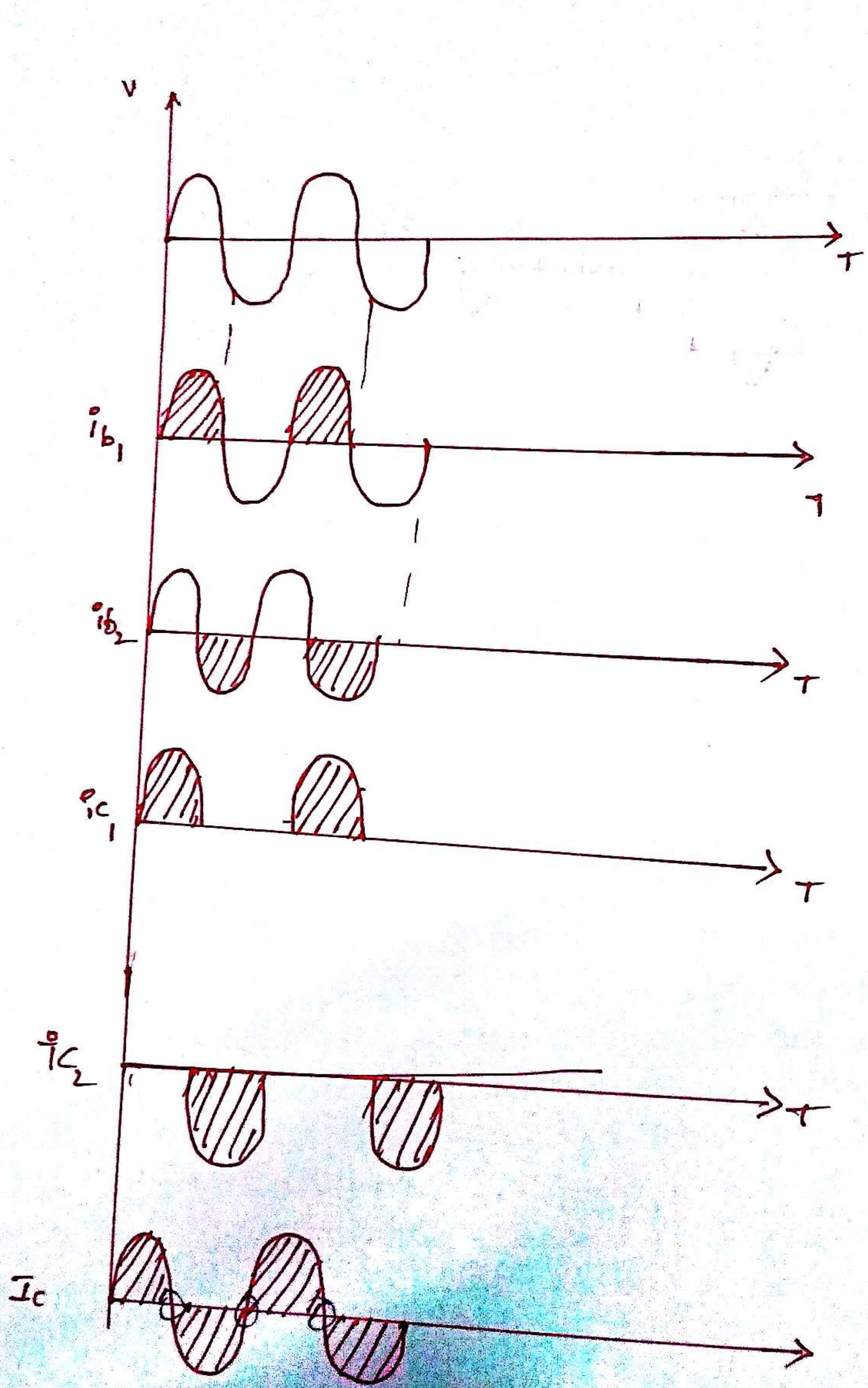
& connecting Q_1 & Q_2 to the load

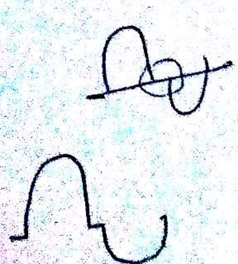
$$R_L' = R_L/n^2$$

where n N_2/N_1

The slope of the graph will be $-1/R_L'$

O/P wave form:





Accerding to ohms law

$V = IR$.

Slope in term of R_L'

$$\frac{1}{R_L'} = \frac{\frac{1}{V_m}}{I_m}$$

Ac power output:

we know that $V_{rms} = V_m/\sqrt{2}$ (V)

$I_{rms} = I_m/\sqrt{2}$ (I).

$$P_{rms} \text{ (or) } P_{ac} = V_{rms} \cdot I_{rms}$$

$$= V_m I_m / 2$$

$$P_{ac} = V_m^2 / 2R_L$$

Efficiency: %

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

$$= \frac{\frac{V_m I_m}{2}}{\frac{2}{\pi} V_{cc} I_m} \times 100$$

$$= \pi/4 \times 100 \times \frac{V_m}{V_{cc}}$$

$(V_m = V_{cc})$

$$= \pi/4 \times 100$$

$$= 78.5\%$$

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Power dissipation:

It is equal to difference b/w output to dc power

$P_d = P_{dc} - P_{ac}$

$= \frac{2}{\pi} V_{cc} I_m - V_m I_m / 2$

$= \frac{2}{\pi} V_{cc} \cdot V_m / R_L' - \frac{V_m^2}{2 R_L'}$ — ①

by differentiating w.r.to V_m

we get

$V_m = \frac{2}{\pi} V_{cc}$

Put V_m in ①

$P_{d(max)} = \frac{2}{\pi^2} \cdot \frac{V_{cc}^2}{R_L'}$

for maximum efficiency $V_m = V_{cc}$

If η% P_d(max) cannot be occured simultaneously

Power dissipated per transistor:

$P_{ac} = V_m \times V_m / 2R_L' = V_m^2 / 2R_L'$

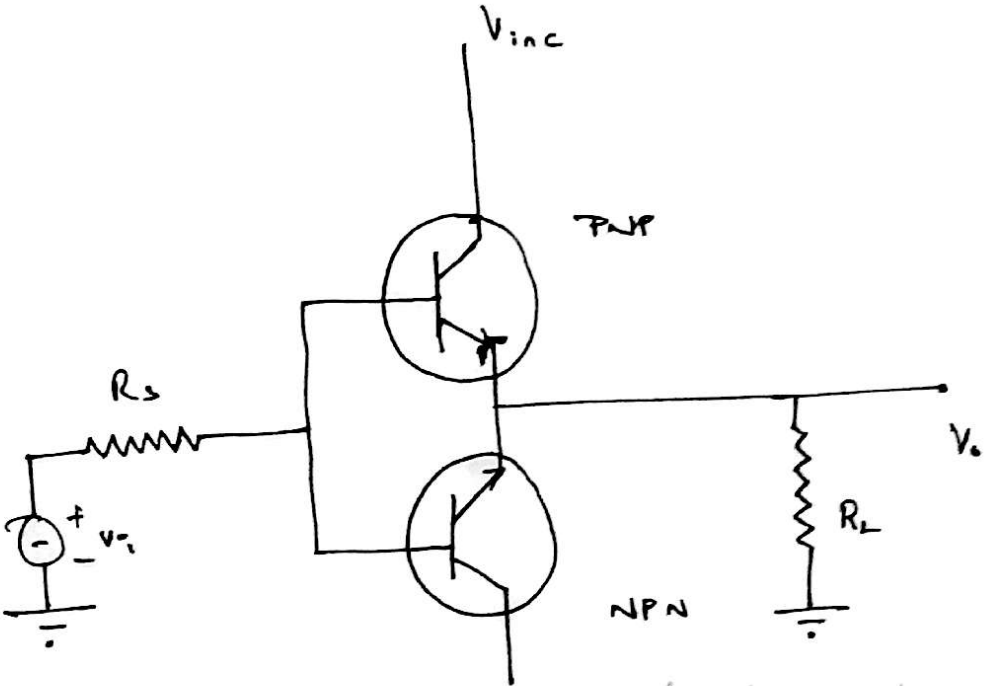
$(P_{ac})_{max} = \frac{V_{cc}^2}{2R_L'}$

Now $P_d max = \frac{2}{\pi^2} \frac{V_{cc}^2}{R_L'}$

$= \frac{4}{\pi^2} (P_{ac})_{max}$

$\frac{P_d (max)}{T} = \frac{2 (P_{ac})_{max}}{\pi^2}$

Complementary Symmetry Amplifier:



In Complementary Symmetry Amplifier there are two type of transistors are connected one is PNP & NPN transistors. Here No Centre tapped transformers are connected when voltage is supplied Base Emitter of NPN transistor is forward biased & PNP transistor is reverse biased for the +ve half cycle. This is vice versa for negative half cycle

Power Calculation:

$$P_{ac} = \frac{1}{2}\frac{V_{cc}^2}{R_L}$$

$$V_m = V_{cc} \Rightarrow R_L = V_m / I_m$$

$$I_m = \frac{V_m}{R_L} = V_{cc}/R_L$$

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$P_{dc} = \frac{2}{\pi} V_{cc} I_m$

$= \frac{V_{cc}^2}{R_L} \times \frac{2}{\pi}$

$P_{dc} - P_{ac} = 0.1366\ V_{cc}^2/R_L$

Advantages:

→ Less cost

→ Impedance matching is possible

Disadvantages:

→ Two Seperate Voltage Supplies

→ O/P is distorted to cross over distortion

Push - Pull

Each

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Push - Pull Class B	Complementary Symmetry class B
Both should be PNP (or) NPN	→ One is PNP & other is NPN
→ Transformers are used	→ No transformer is used.
→ Impedance matching is due to o/p transformer	→ impedance matching is due to common collector configuration
→ frequency response is poor	→ frequency response is improved
→ Bulky, costly	→ Bulky not costly
→ Single supply	→ dubble supply is needed
→ Efficiency is higher than Class A	→ Efficiency is higher than Push pull Amplifiers.

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CLASS-B : PUSH PULL AMPLIFIER

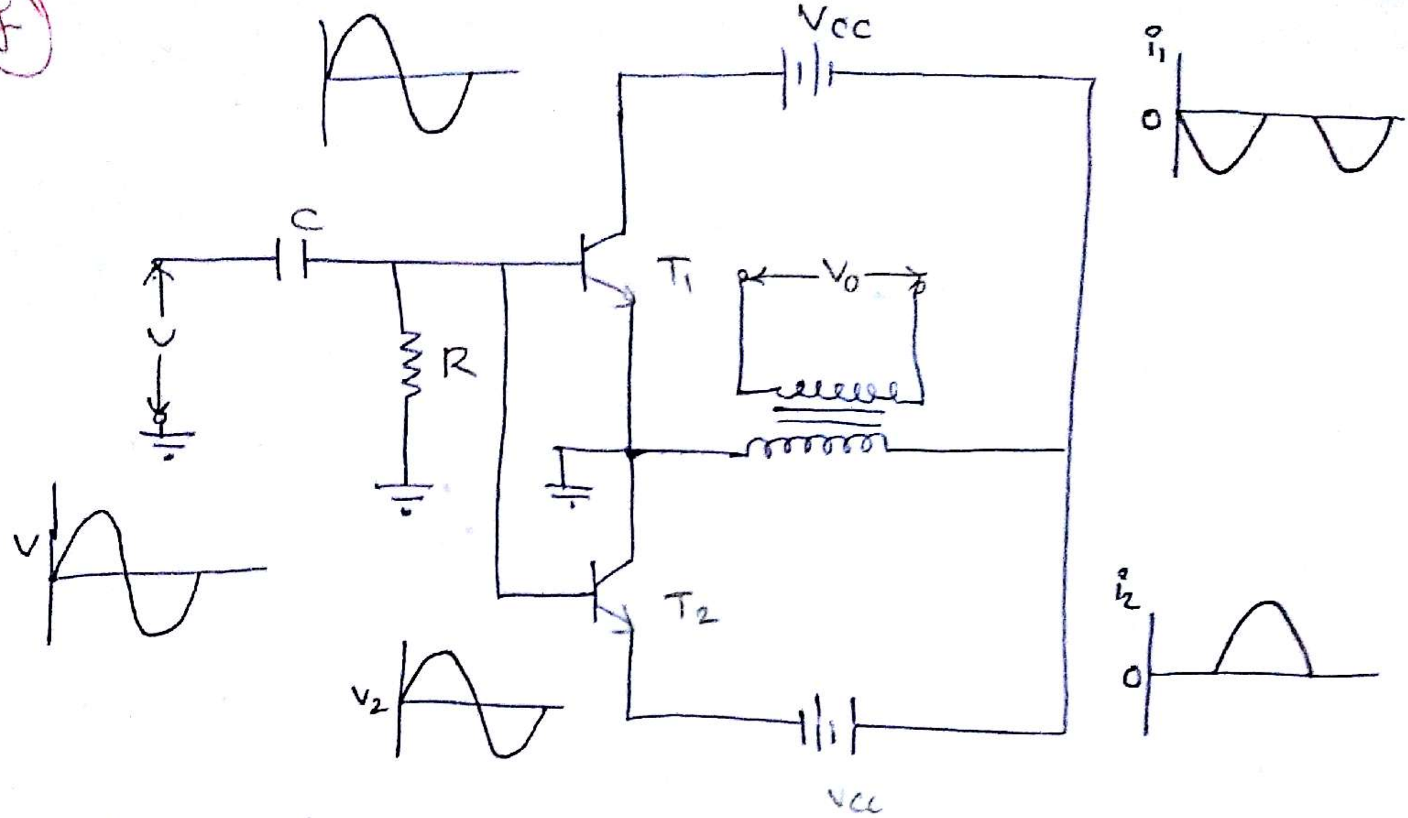
B. Bhanusree
II yr. ECE - A sec.

* Class B:-
(Introduction :)
To Improve the full power efficiency of the previous class A amplifier by reducing the wasted power in the form of heat, it is possible to design the power amplifier circuit with two transistors in its output stage producing what is commonly termed as a class B Amplifier also known as a push pull amplifier configuration.

* Push-pull amplifiers use two "complementary" or matching transistors, one being an NPN-type and the other being a PNP-type with both power transistors receiving the same input signal together that is equal in magnitude, but in opposite phase to each other. This results in one transistor only amplifying one half or 180° of the input waveform cycle while the other transistor amplifies the other half or remaing 180° of the input waveform cycle with the resulting "two-halves" being put back together again at the output terminal.
* Then the conduction angle for this type of amplifier circuit is only 180° or 50% of the i/p signal. This pushing and pulling effect of the alternating half cycles by the transistors gives this type

of ckt, its amusing "push-pull" name, but B
generally known as the "class B Amplifier".

Class B push pull Transformer ampr ckt :-



construction :-

* It employs one NPN transistor (T_1)
* It employs one PNP transistor (T_2)
* Hence it is called as complementary push pull amplifier
* It requires no centre tapped transformers.

Working :-

During positive half-cycle,

* the NPN transistor (T_1) conducts and the PNP transistors (T_2) cut-off.
* Thus collector current I_{C_1} flows through the primary side of the transformer.

ng Negative half cycle,

NPN transistor (T_1) cuts off and the PNP transistor (T_2) conducts.

Thus the collector current I_{C2} flows through the primary side of the transformer.

⇒ In this way NPN transistor amplifies the positive half cycle of the input signal.

⇒ The PNP transistor amplifies the negative half cycle of the input signal.

⇒ An output transformer is used for impedance matching.

Advantages :-

* High AC output power is obtained
* This circuit does not require center tapped transformers
* Less weight.
* Low cost.

Disadvantages :-

* It is difficult to get a pair of transistors that have similar characteristics.
* This ckt requires two power supplies
  +Vcc for NPN
  −Vcc for PNP
* Cross-over distortion occurs in the output wave form.

Output characterstics:

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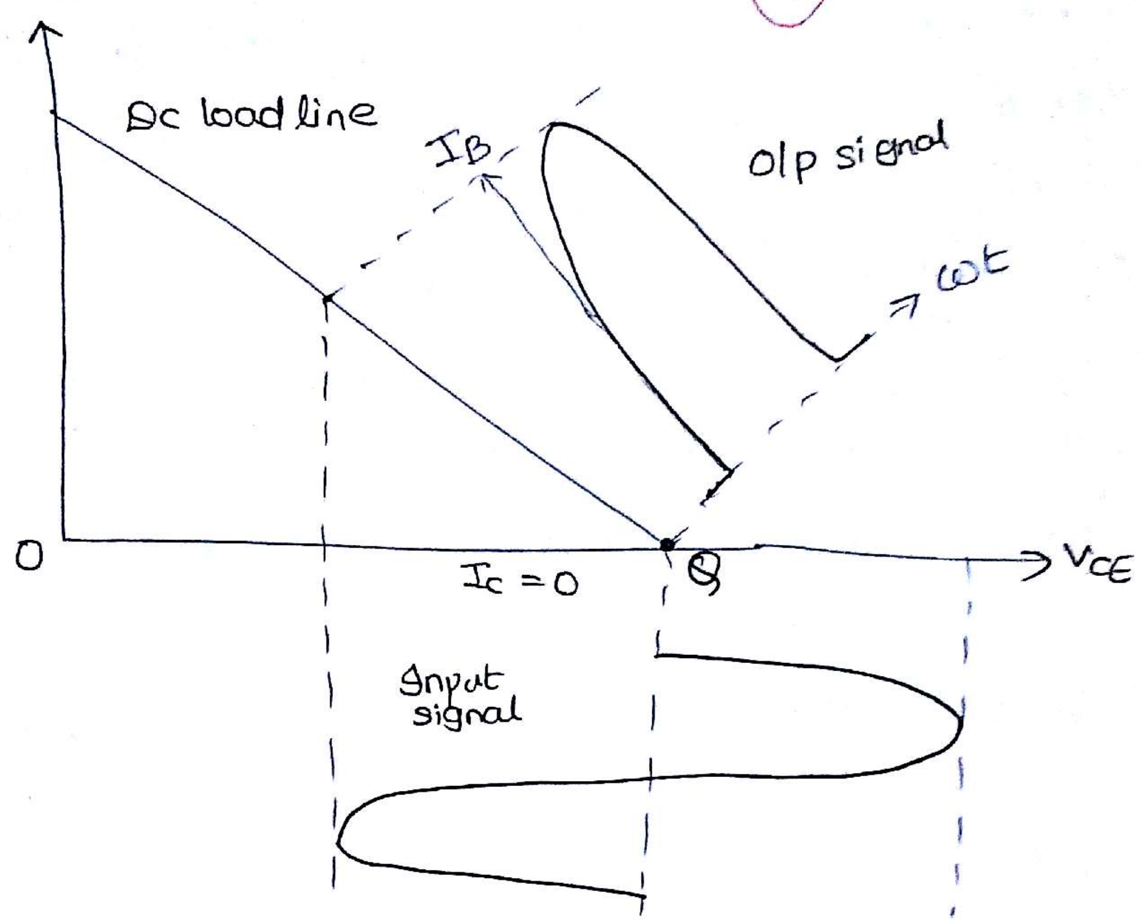


class B Ampr operation has zero Dc bias as the transistors are biased at the cut off, so each transistor only conducts when the i/p signal is greater than the base emitter voltage.

∴ At zero i/p ter is zero o/p and no power is being consumed. This then means that the actual Q-point of a class B ampr is on the Vce part of the load line as shown in graph.

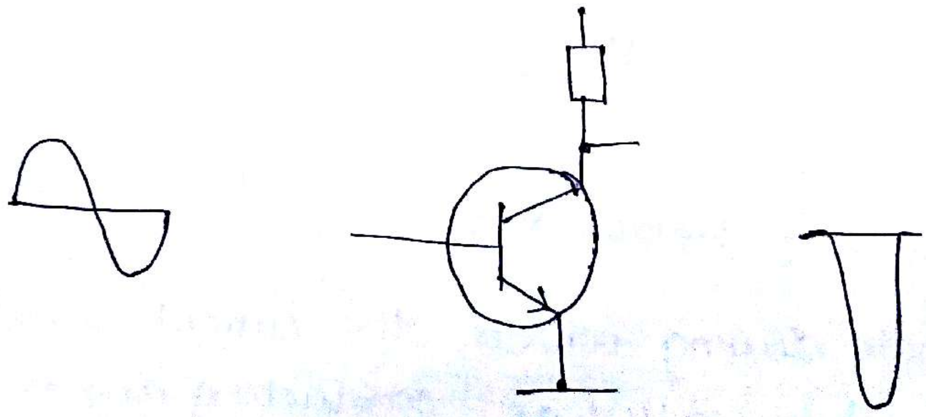
Class C Amplifiers:

* Only half of the i/p signal is given. ie class C amplifiers conduct less than 50% of i/p signal.
* Distortion at o/p is high
* It has higher efficiencies ie 90%. when compared to Class A and Class B.
* Applications:

→ RF transmitters.

→ Mega phones.

* I/p signal is used to roughly switch the amplifying device on and off, which causes pulses of current to flow through a tuned circuit.



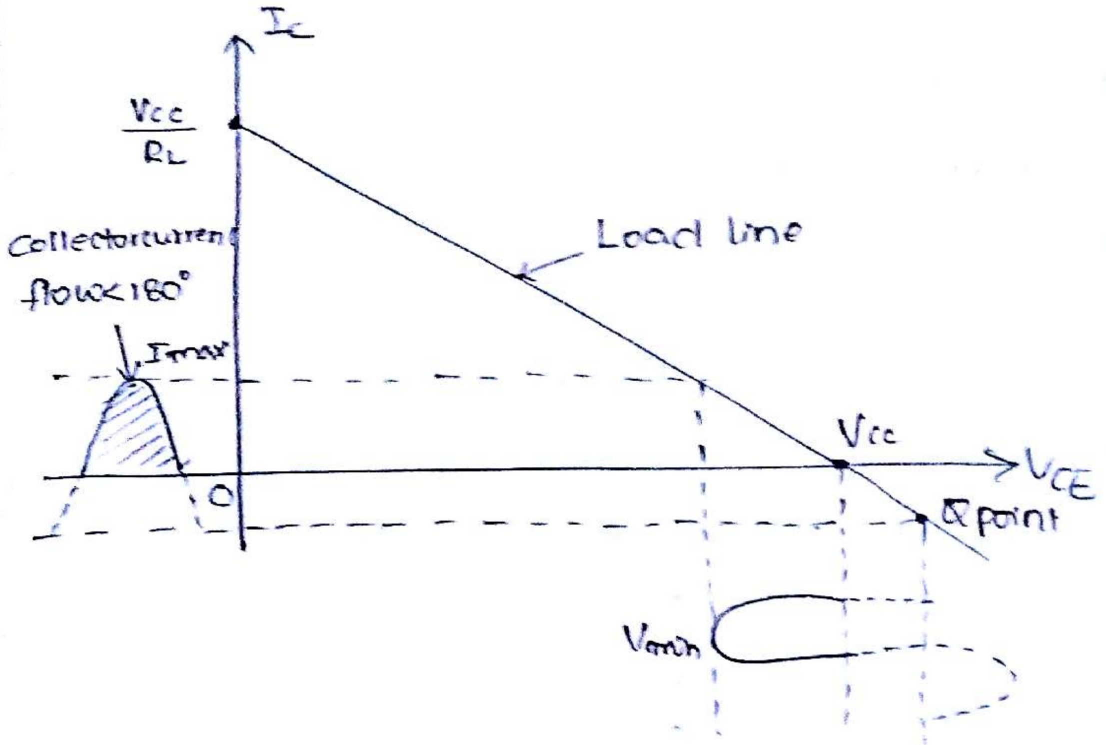
Definition:

A amplifier is said to be class C; if the Q point and i/p signal selected such that o/p signal is obtained for less than half cycle, for a full i/p cycle.

* Due to selection of Q point, only transistor re[…] active, for less than half cycle, so only that […] part is reproduced at o/p.

* for remaining part of i/p signal transistor is cut-[…] and no signal is produced at o/p.

Wave form of class C



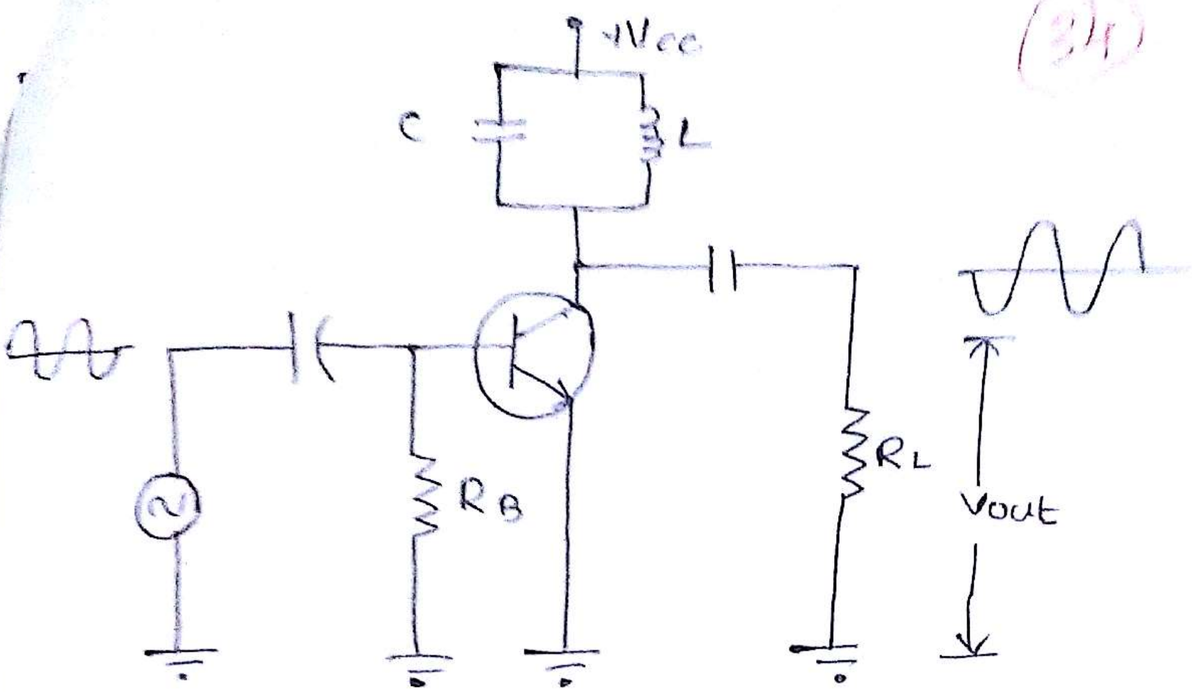
* Here Q point lies below X-axis.

* The total angle during which the current flows is less than 180°. So it called as conducting angle. (θ_c)

→ In class-C, ||el resonant ckt acts as a load impedance

→ Since collector current flows for less than half cycle, it consists of series of pulses with harmonics of i/p signal.

→ A ||el tuned circuit acting as a load impedance is tuned to i/p frequency. So it filters the harmonic frequencies and produces sinewave o/p with fundament i/p signal component


+Vcc
C
L
RB
RL
Vout

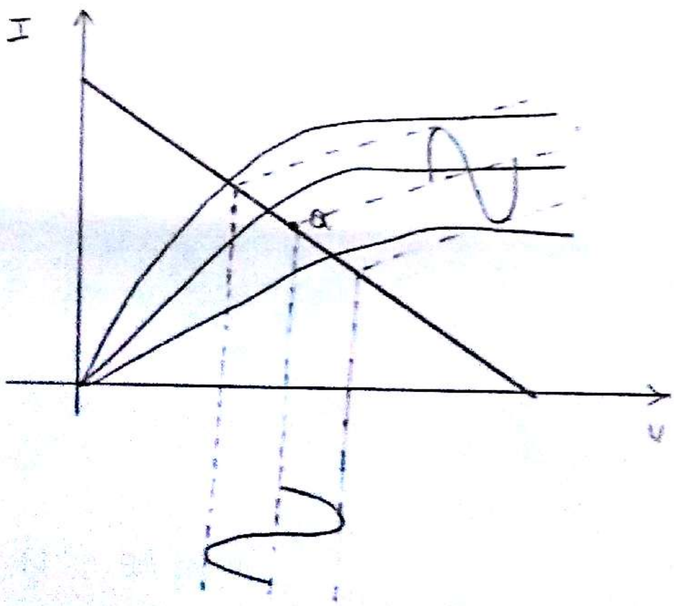
CLASS-A POWER AMPLIFIER:

In these power amplifiers if we plot a plot graph between I and V, as I on y axis and V on x axis

we locate Q point at the midpoint of active region

The output phase in the class A amplifier is 360°

we have.



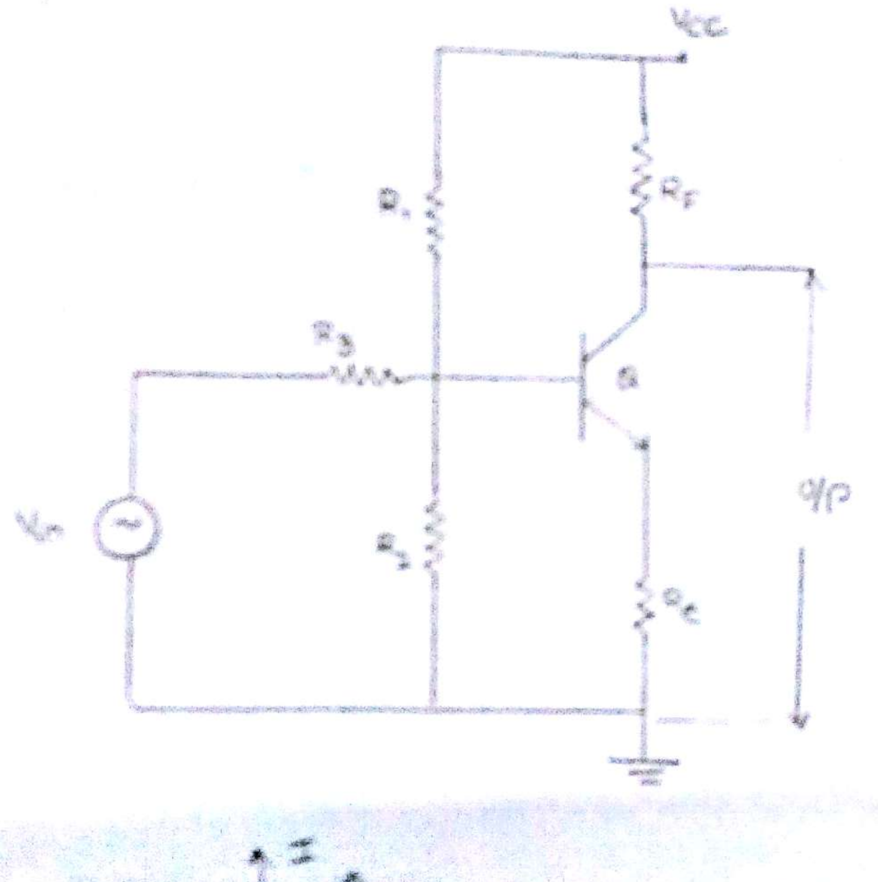
Q is at mid point on load line

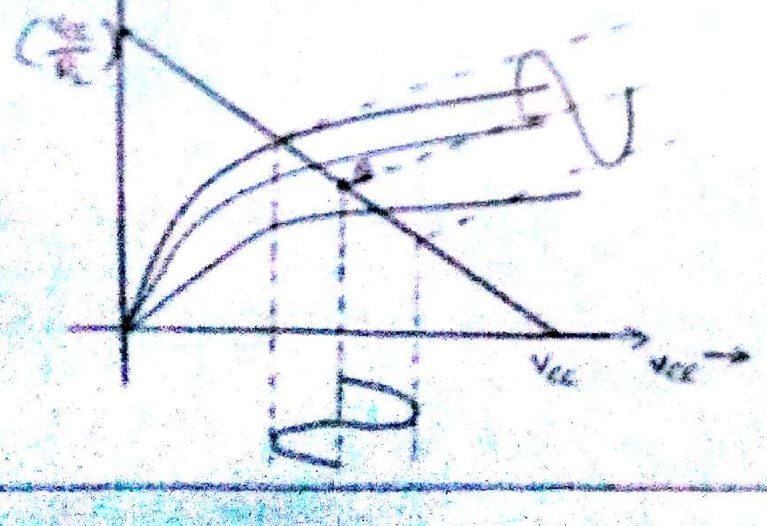
CLASS A POWER AMPLIFIERS

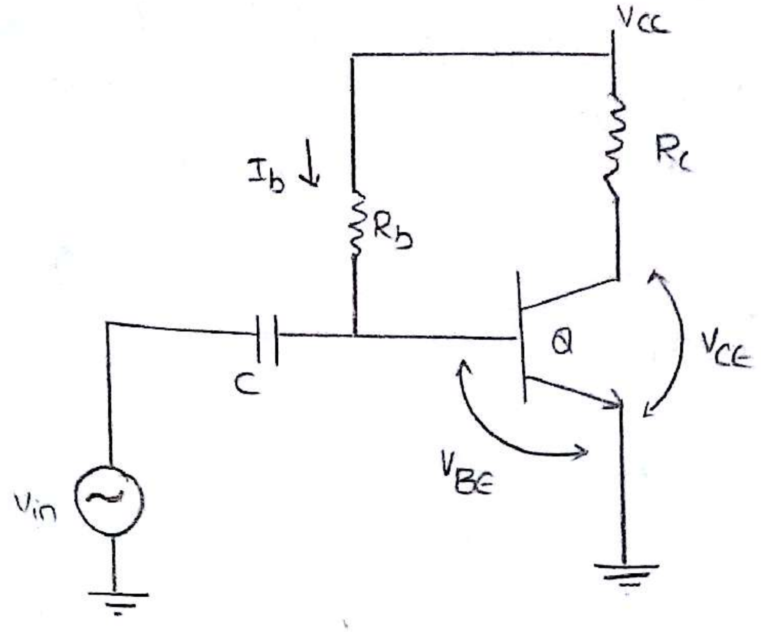
The efficiency of class A power amplifier is 25-50%.

There will be no distortion in the output waveform of class A power amplifier

Analysis of class A power Amplifier







Dc Analysis:

Apply KVL, we have

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

$$I_C R_L = V_{CC} - V_{CE}$$

$$I_C = \frac{V_{CC} - V_{CE}}{R_L}$$

$$I_C = \frac{V_{CC}}{R_L} - \frac{V_{CE}}{R_L}$$

$$I_C = \left(\frac{-1}{R_L}\right) \cdot V_{CE} - \frac{V_{CC}}{R_L}$$

we consider $y = mx + c$ the m is slope

Here slope is $\left(\frac{+1}{R_L}\right)$

DC operation:

$$I_{BQ} = \frac{V_{CE} - V_{BE}}{R_b}$$

$$I_{BQ} = \frac{V_{CE} - 0.7}{R_b}$$

$$I_{CQ} = \beta I_{BQ} \Rightarrow \beta = \frac{I_{CQ}}{I_{BQ}}$$

$$\beta = I_{CQ}\left(\frac{R_b}{V_{CE} - 0.7}\right)$$

$$I_{CQ} = \beta\left(\frac{V_{CE} - 0.7}{R_b}\right)$$

$$V_{CEQ} = V_{CC} + I_C R_C$$

Hence Q is defined at point $Q(V_{CEQ}, I_{CQ})$

DC power input:

when AC input signal is applied, the base current varies sinosudially.

I_C current varies around its Quiscent point will the output voltage V_{CC} get varies. Varying current and output voltage delivers AC power to the load.