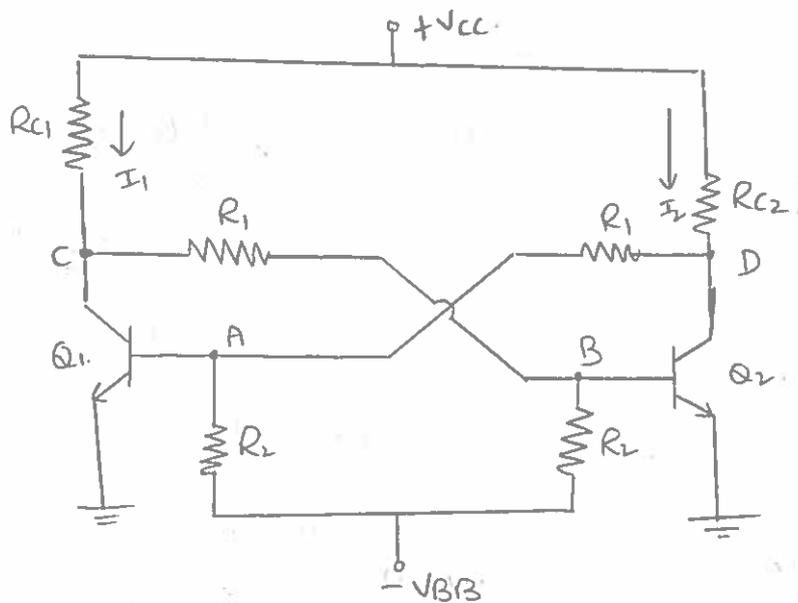


## Classification:-

Multivibrators (or multis, as they are generally abbreviated) are broadly classified into three categories, based upon their output states.

- \* Bistable multivibrators.
- \* Monostable multivibrators.
- \* Astable multivibrators.

## Principle of operation of bistable multivibrator:-



$Q_1, Q_2$  - identical n-p-n transistors

$R_1, R_2, R_{C1}$  &  $R_{C2}$  are resistors. ( $R_{C1} = R_{C2}$ )

$V_{CC}$  = supply voltage.

$V_{BB}$  = Fixed bias voltage.

Let it be assumed that  $Q_1$  &  $Q_2$  are n-p-n transistors. They are coupled to each other as shown. It is evident that the output of each transistor is coupled to the input of the transistor [The collector current of  $Q_1$  supplies the base-drive for  $Q_2$ , and the collector current of  $Q_2$  provides the base-drive for  $Q_1$ ]. Since the transistors are identical, their quiescent currents would be the same, unless the loop gain is greater than unity.

Let the loop gain greater than unity, and let  $I_1 > I_2$  by a small margin.

When  $I_1$  increases slightly, the voltage drop across the collector resistance  $R_{C1}$  increases. Since  $V_{CC}$  is fixed the voltage of point C decreases [Because  $V_C = V_{CC} - I_1 R_{C1}$ . All voltages are w.r.t ground]. The effect of decreasing the base current  $Q_2$ . This, in turn, decreases the collector current  $Q_2$  viz.  $I_2$  [Because  $I_C = h_{fe}$

$I_B$  where  $h_{fe}$  = forward current gain]. If  $I_2$  decreases, the voltage drop  $I_2 R_{C2}$  decreases. Hence the voltage of point D increases [ $\because V_D = V_{CC} - I_2 R_{C2}$ ].

Due to increase of  $V_D$  the base current of  $Q_1$  increases. This increases the collector current of  $Q_1$ , viz.  $I_1$

Thus  $I_1$  further increases.  $I_{RC1}$  drop further decreases,  $V_c$  further decreases, the base current of  $Q_2$  further decreases, with the result that  $I_2$  further decreases.

Standard specifications:-

The following values may be assumed for the junction voltages.

i) In the cut-off region i.e. for the OFF state

$V_{BE}(\text{cutoff}) : \leq 0$  volt for silicon transistor.

$\leq -0.1$  volt for germanium transistor.

$V_{BE}$  is base-emitter voltage.

For the base-emitter junction to be forward biased, it is necessary that  $V_{BE} > V_{BE}(\text{cutoff})$ .

ii) In the saturation region i.e. for the ON state

$V_{BE}(\text{sat}) : 0.7$  volt for silicon transistor.

$0.3$  volt for germanium transistor.

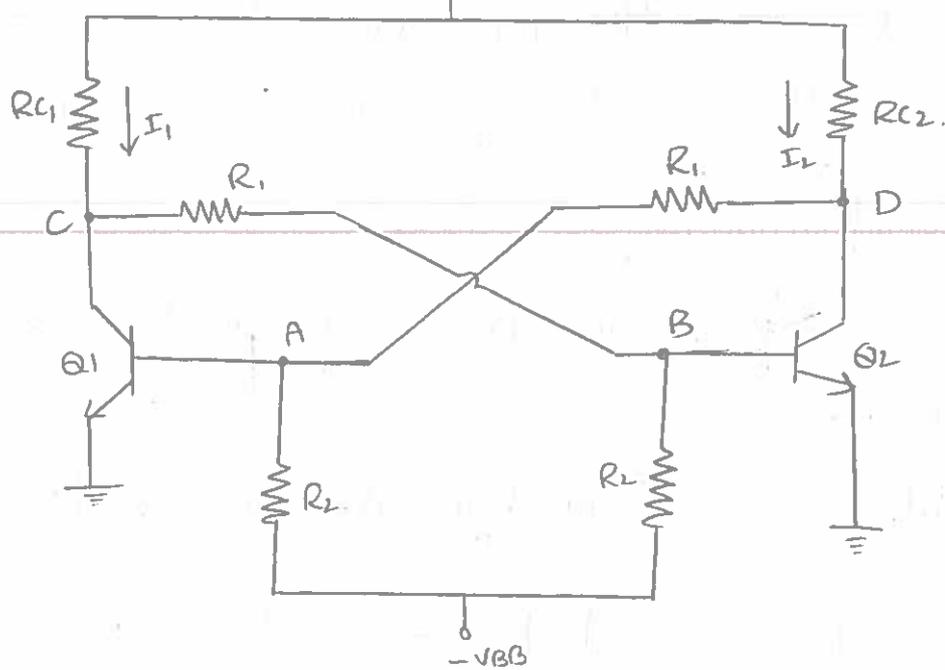
$V_{CE}(\text{sat}) : 0.3$  volt for silicon transistor.

$0.1$  volt for germanium transistor.

These values hold good for n-p-n transistor.

Note:-  $V_{CE}$  is collector emitter voltage.  $V_{CE}(\text{sat})$  is the voltage measurable across emitter and collector terminals, when the diode is operating in saturation region.

## Fixed bias transistor binary:-



The fixed bias bistable multivibrator of figure is reproduced here for ready reference.

The binary under consideration is a fixed bias binary. It may be observed that there is no emitter resistor. Instead there is biasing voltage  $V_{BB}$ . Since the transistors used are n-p-n transistors,  $V_{CC}$  is positive and  $V_{BB}$  is negative. In practice  $V_{CC}$  and  $V_{BB}$  are only of a few volts.

This is due to the fact that when a transistor ( $Q_1$  or  $Q_2$ ) is OFF, its collector current is practically zero, and hence its collector voltage is almost equal to  $V_{CC}$ . Thus the voltage across the transistor i.e.  $V_{CE} \approx V_{CC}$ .

Since  $V_{ce}$  should be smaller than the collector break-down voltage  $BV_{ce}$ ,  $V_{cc}$  should be limited to a few volts only (usually 3-20V).

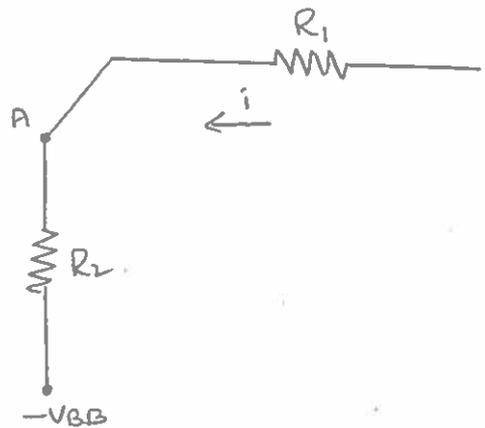
Let it be assumed that, during a switching transient, transistor  $Q_2$  becomes ON and transistor  $Q_1$  becomes OFF.

Since  $Q_2$  is in saturation, the potential of point D is practically zero.

For  $V_D = V_{ce(sat)} \approx 0$ .

potential of point A =  $V_D - IR_1$

$$\begin{aligned} \text{we have } i &= \frac{V_D - (-V_{BB})}{R_1 + R_2} \\ &= \frac{V_D + V_{BB}}{R_1 + R_2} \end{aligned}$$



$$\therefore V_A = V_D - \left[ \frac{V_D + V_{BB}}{R_1 + R_2} \right] R_1$$

$$= 0 - \frac{(0 + V_{BB})R_1}{R_1 + R_2}, \text{ putting } V_D = 0.$$

$V_{BB}$  is positive for n-p-n transistor.

$$\therefore V_A \text{ is } -ve \text{ i.e. } V_A < 0.$$

But  $V_A = V_{BE}$  of  $Q_1$   $\therefore V_{BE}(Q_1)$  becomes negative.

When  $V_{BE}$  becomes negative, it is  $< V_{BE}(\text{cut-off})$ .

$\therefore$  Transistor  $Q_1$  is indeed OFF

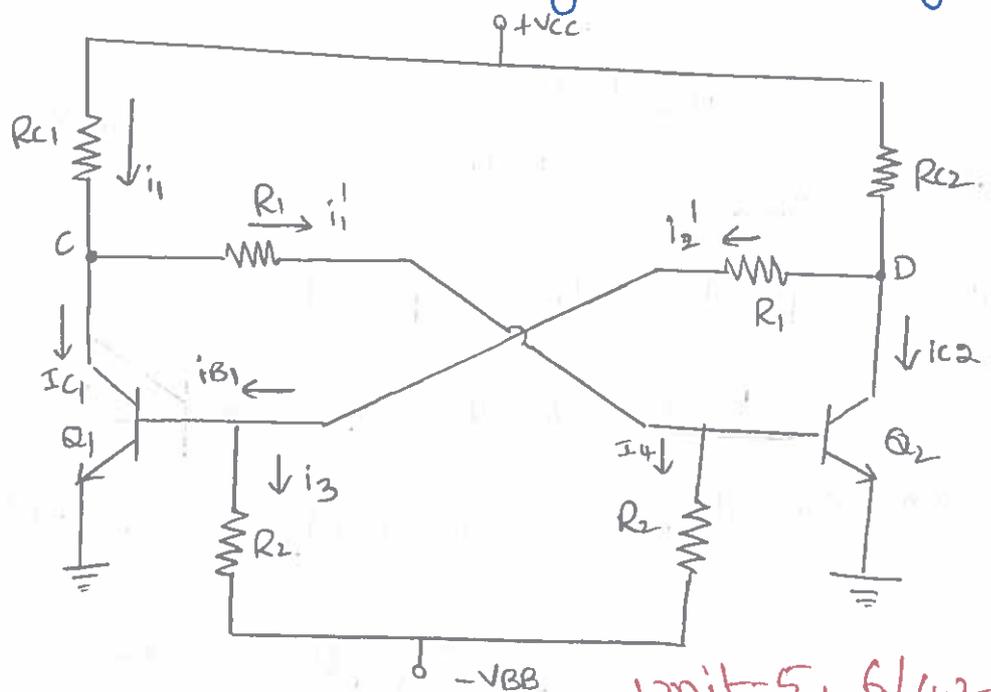
Since  $Q_1$  is OFF, there is no collector current and hence  $V_C = V_{CC}$ . This large positive potential  $V_C$  is sufficient to overcome the negative bias of  $-V_{BB}$  and force a current through the base of transistor  $Q_2$ , with the result that  $I_{B2} > I_{B2(\text{min})}$ , where  $I_{B2(\text{min})}$  is the minimum base current needed to keep  $Q_2$  in saturation.

Hence transistor  $Q_2$  is indeed ON.

Thus it is seen that when  $Q_2$  is ON,  $Q_1$  is OFF. It can similarly be shown that when  $Q_1$  is ON,  $Q_2$  is OFF. Thus the binary has two stable states.

To evaluate the state state currents and voltages

consider the fixed bias binary shown in figure.



mit-5, 6/42

$Q_1, Q_2 \dots n-p-n$  transistors  $R_1, R_2, R_{c1}, R_{c2} \dots$

Resistors [ $R_{c1} = R_{c2}$ ].

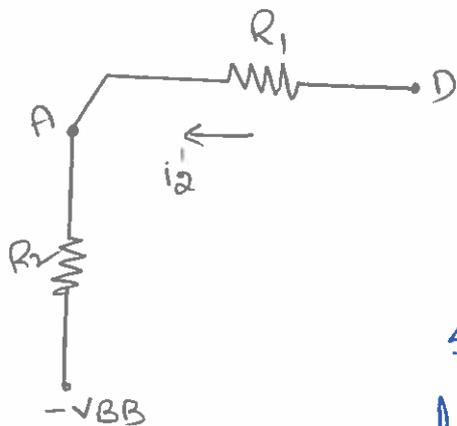
Let it be required to determine all stable state currents and voltages. Let, during a switching transient, transistor  $Q_2$  is ON, and transistor  $Q_1$  be OFF.

Since  $Q_2$  is ON,  $V_D = V_{CE(sat)}$  and

$$V_B = V_{BE(sat)}.$$

[All voltages are w.r.t. ground].

To find  $i_2, i_2'$  and  $i_{c2}$  :-



we have

$$i_2 = \frac{V_{CC} - V_D}{R_{c2}}$$

Since  $Q_1$  is OFF,  $I_{B1} = 0$  as  $Q_1$  forms an open circuit and

$$i_3 = i_2'.$$

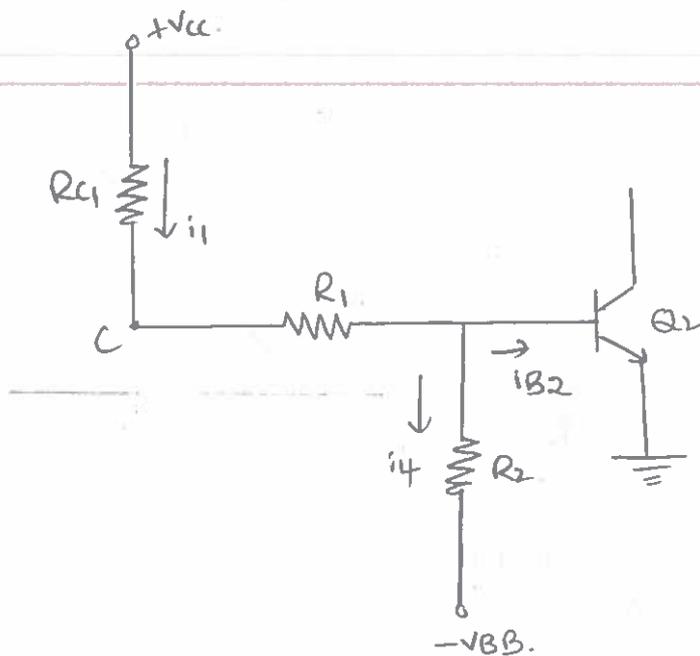
From the circuit of figure.

$$i_2' = \frac{V_D - (-V_{BB})}{R_1 + R_2}$$

$$= \frac{V_D + V_{BB}}{R_1 + R_2}$$

Also  $i_{C2} = i_2 - i_2'$ .

Thus currents  $i_2$ ,  $i_2'$  and  $i_{C2}$  are evaluated.



Since  $Q_1$  is OFF,  $i_{C1} = 0$

$$\therefore i_1 = i_1'$$

$$i_{B2} = i_1 - i_4.$$

We have 
$$i_1 = \frac{V_{CC} - V_B}{R_{C1} + R_1}$$

$$i_4 = \frac{V_B - V_{BB}}{R_2} = \frac{V_B + V_{BB}}{R_2}$$

$$\therefore i_{B2} = \left( \frac{V_{CC} - V_B}{R_{C1} + R_1} \right) - \left( \frac{V_B + V_{BB}}{R_2} \right)$$

We have

$$i_{B2(\min)} = \frac{i_{C2}}{h_{fe(\min)}}$$

If the calculated value of  $I_{B2} > I_{B2(\min)}$ , then it implies that  $Q_2$  is indeed ON. We shall next verify whether  $Q_1$  is indeed OFF.

From the circuit of figure  $V_A = V_D - I_2 R_1$ .

But  $V_A = V_{BE}$  of  $Q_1$ .

It would be found that  $V_A < 0$ . i.e.  $V_{BE}(Q_1) < 0$ .

This proves that  $Q_1$  is indeed OFF.

To find the stable stage voltage  $V_C$ :

We have  $V_C = V_{CC} - I_1 R_{C1}$  from the circuit

The other voltages  $V_A$ ,  $V_B$  and  $V_D$  have already been evaluated.

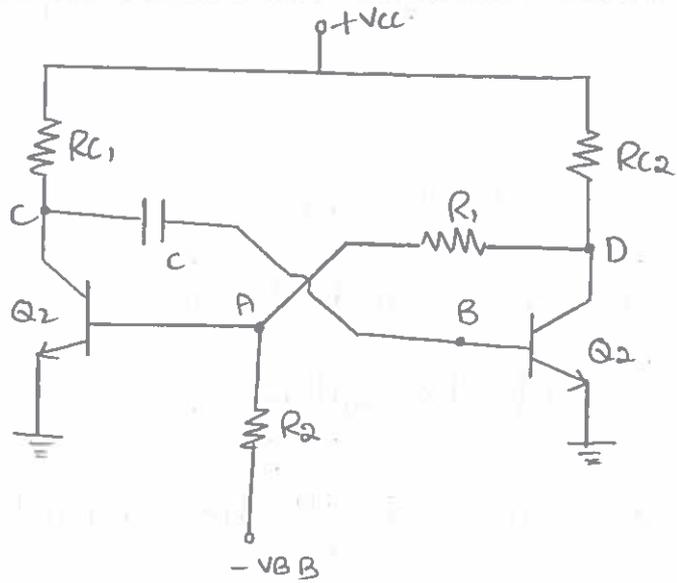
\* Principle of operation of Monostable multivibrator:

In figure is shown a collector-coupled monostable multivibrator. Of the two transistors  $Q_1$  and  $Q_2$ ,  $Q_1$  is normally OFF and  $Q_2$  is normally ON. Resistors  $R_1$  &  $R_2$  are connected to the normally-OFF transistor, and the capacitor 'C' is connected to the normally-ON transistor.

The principle of working is as follows:

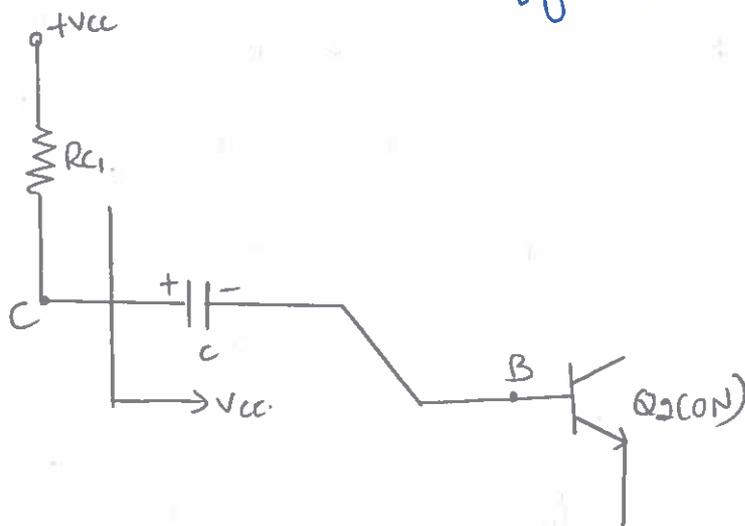
It is seen from the circuit of the monostable multivibrator that, under normal conditions, the supply voltage  $V_{CC}$  provides enough base drive to the

transistor  $Q_2$  through resistance  $R$ , with the result that  $Q_2$  goes into saturation. with  $Q_2$  on,  $Q_1$  goes off as already studied in the context of binary operation.



$Q_1, Q_2$  n-p-n transistors  $R_1, R_{c1}, R_{c2}, R, \dots$  Resistors  
 $[R_{c1} = R_{c2}]$   $V_{cc}$ ... Supply voltage  $V_{BB}$ ... Bias voltage.

with  $Q_2$  ON and  $Q_1$  OFF, the capacitor finds a charging path as shown in figure.

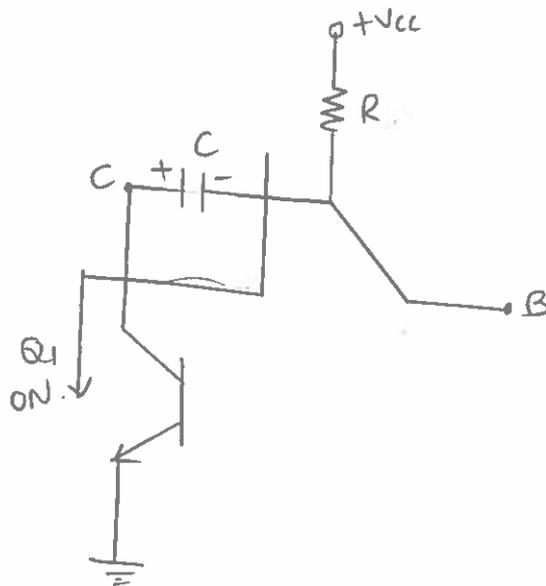


The voltage across the capacitor is  $V_{CC}$  with polarity as shown [positive on the left, negative on the right].

It is obvious that in the stable state of the multi,  $Q_2$  is ON and  $Q_1$  is OFF.

If a negative triggering pulse is applied to the collector of  $Q_1$ , it is transmitted to the base of  $Q_2$  through the capacitor, and hence makes the base of  $Q_2$  negative. Immediately  $Q_2$  goes OFF and  $Q_1$  becomes ON. However, this is only a quasi-stable state, as is obvious from the following observation.

With  $Q_1$  ON and  $Q_2$  OFF, the capacitor  $C$  finds a discharging path as shown in figure.



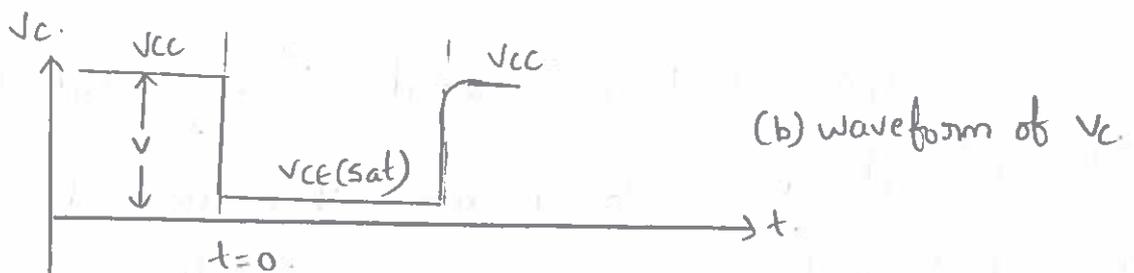
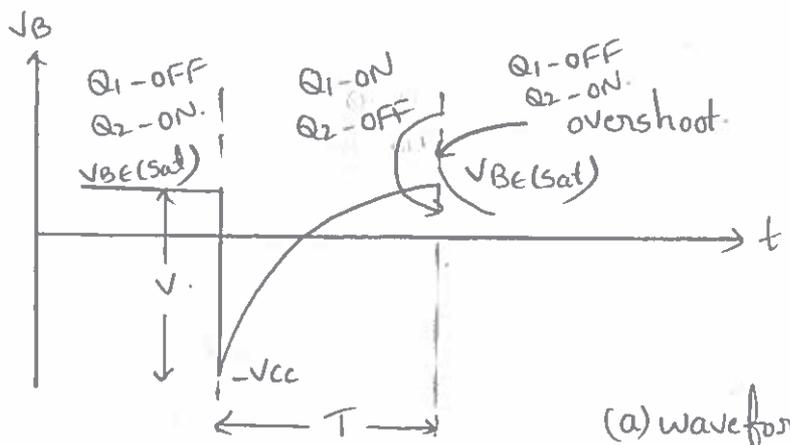
As the capacitor discharges, it is seen that the potential of point  $B$  becomes less and less negative and after a time, we have  $V_B = V_r$ , the cut in voltage of  $Q_2$ .

As soon as  $V_B$  crosses the level of  $V_r$ ,  $Q_2$  starts conducting and gets saturated. When  $Q_2$  becomes ON,  $Q_1$  becomes OFF. Thus the original stable state of the multi is restored.

[In quasi-stable state:  $Q_1$  is ON and  $Q_2$  is OFF].

The interval during which the quasi-stable state of the multi persists i.e.  $Q_2$  remains OFF is dependent upon the rate at which the capacitor  $C$  discharges. This duration of the quasi-stable state is termed as delay time or pulse width or gate time. It is denoted by  $T$ .

The waveforms of the voltages at point B (Base of  $Q_2$ ) and C (Collector of  $Q_1$ ) are as shown.



(Instant of application of triggering signal)

stable state | Quasi-stable state | stable state

Expression for pulse width (or delay time),  $T$  :-

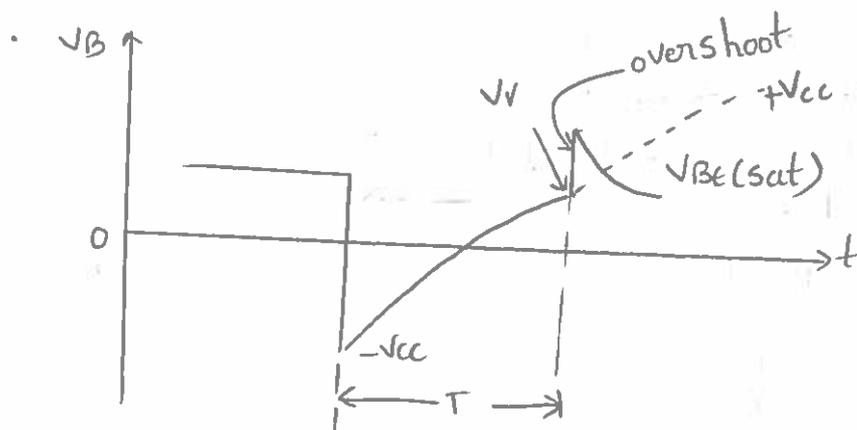
Referring to the wave form of figure, we have initial value of  $v_B$  (at  $t=0$ ),  $v_{in} = -V_{CC}$

As the capacitor discharges, the voltage  $v_B$  rises exponentially and would attain the value  $+V_{CC}$ , but for the fact that at  $t=T$ ,  $Q_2$  becomes ON and, as a result,  $v_B$  takes the value  $v_B = V_V$ , the cut in voltage of  $Q_2$ , which may be taken as zero.

$\therefore$  Final value of  $v_B$  (at  $t = \infty$ ),  $v_{final} = +V_{CC}$  and

At  $t=T$ ,  $v_B = V_V = 0$ .

The exponentially increasing voltage  $v_B$  is mathematically expressed as.



$$v_B = v_{final} - (v_{final} - v_{initial}) \cdot e^{-t/Rc}$$

$$\text{or } v_B = V_{CC} - [V_{CC} - (-V_{CC})] e^{-t/Rc}$$

we have at  $t=T$ ,

$$V_B = V_r = 0$$

$$\therefore 0 = V_{CC} - 2V_{CC} \cdot e^{-T/RC}$$

$$\text{or } 0 = 1 - 2e^{-T/RC}$$

$$\text{or } 2e^{-T/RC} = 1, \text{ whence } e^{T/RC} = 2.$$

$$\text{i.e. } \frac{T}{RC} = \log_e 2 = 0.69.$$

$\therefore$  Gate width (or pulse width),  $T = 0.69RC$ .

Note:- In obtaining the above expression for the pulse width the reverse current  $I_{CBO}$  of transistor  $Q_1$  has been ignored. If however,  $I_{CBO}$  is taken into consideration, the expression for  $T$  modifies as:

$$T = RC \log_e \left[ \frac{2V_{CC} + I_{CBO}R}{V_{CC} + I_{CBO}R} \right]$$

If we put  $I_{CBO} = 0$ , we get

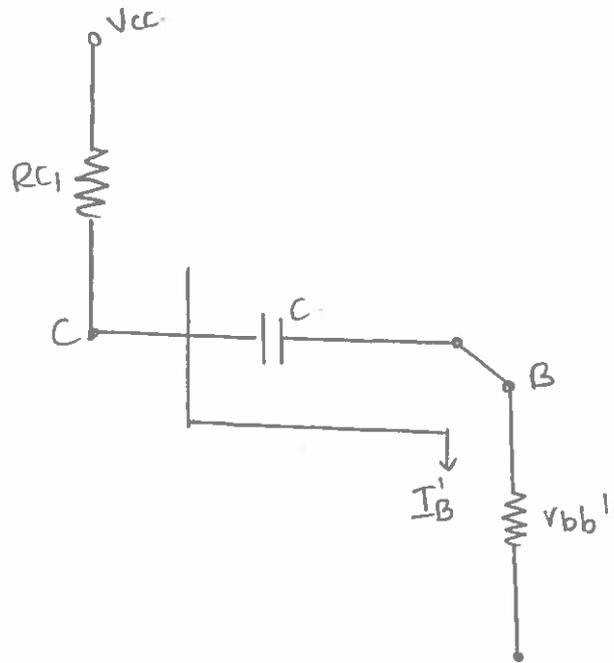
$$T = RC \log_e 2 \text{ or } T = 0.69RC, \text{ as before.}$$

Expression for overshoot ( $\delta$ ):-

Referring to figure. It is observed that there is an overshoot of the base voltage of  $Q_2$  at the end of the quasi-stable state i.e. at  $t = T$ . Let it be denoted as  $\delta$ .

$\delta$  can be evaluated as follows:-

Let  $V_B(T^-) = V_B$  just prior to the end of the quasi-stable state and  $V_B(T^+) = V_B$  immediately after the end of the quasi-stable state.



It is evident that  $V_B(T^-) = V_r$  the cut-in voltage, and  $V_B(T^+) = V_{BE(sat)} + I_B' r_{bb'}$  where  $r_{bb'}$  = base spread resistance of  $Q_2$ , and  $I_B'$  = discharge current.

The path of  $I_B'$  is shown in figure.

Let the instantaneous change of voltage at point B (base of  $Q_2$ ) at  $t = T$  be denoted as  $\delta_1$ .

we have  $\delta_1 = V_B(T^+) - V_B(T^-)$

$$= V_{BE(sat)} + I_B' r_{bb'} - V_r \rightarrow \textcircled{1}$$

At point c:—

let  $v_c(T^-)$  = value of  $v_c$  just prior to the end of the quasi-stable state, and

$v_c(T^+)$  = value of  $v_c$  immediately after the end of the quasi-stable state.

Let us instantaneous change of voltage at point 'c' (collector of  $Q_1$ ) at  $t = T$  be denoted as  $\delta_2$ .

$$\begin{aligned} \text{We have } \delta_2 &= v_c(T^+) - v_c(T^-) \\ &= V_{CC} + I_B' R_{C1} - V_{CE(sat)} \rightarrow \textcircled{2} \end{aligned}$$

since, from the waveform of figure

$$v_c(T^-) = V_{CE(sat)} \text{ and}$$

$$v_c(T^+) = V_{CC} - I_B' R_{C1} \text{ from figure.}$$

$$\text{But } \delta_1 = \delta_2 = \delta.$$

$$\therefore V_{BE(sat)} + I_B' r_{bb'} - V_r = V_{CC} - I_B' R_{C1} - V_{CE(sat)}$$

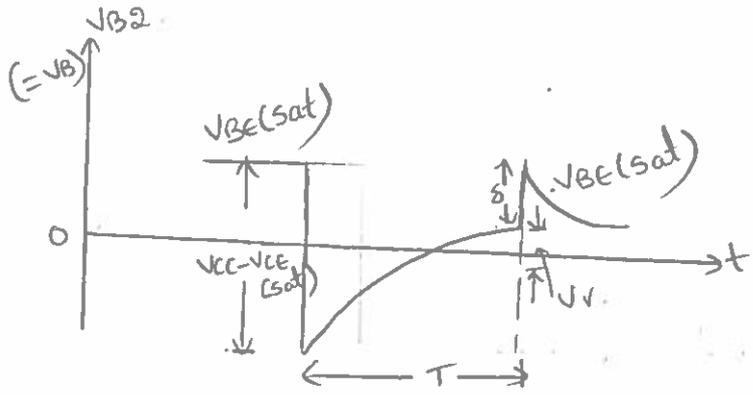
Rearranging we have

$$I_B' = \frac{V_{CC} - V_{CE(sat)} - V_{BE(sat)} + V_r}{r_{bb'} + R_{C1}}$$

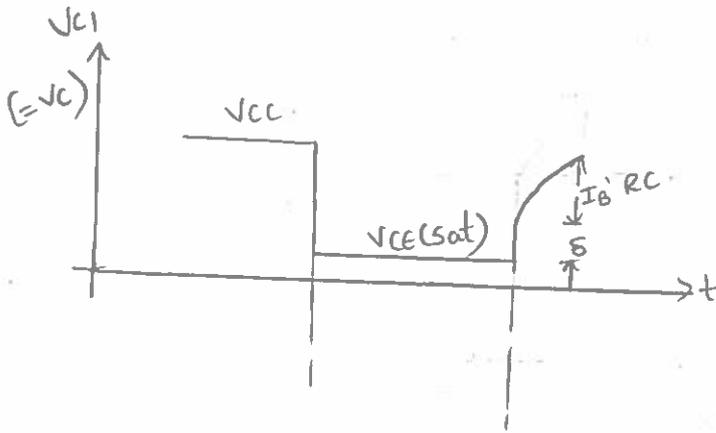
$$\text{Also } \delta = \delta_1 = V_{BE(sat)} + I_B' r_{bb'} - V_r.$$

$$\delta = \delta_2 = V_{CC} - I_B' R_{C1} - V_{CE(sat)}.$$

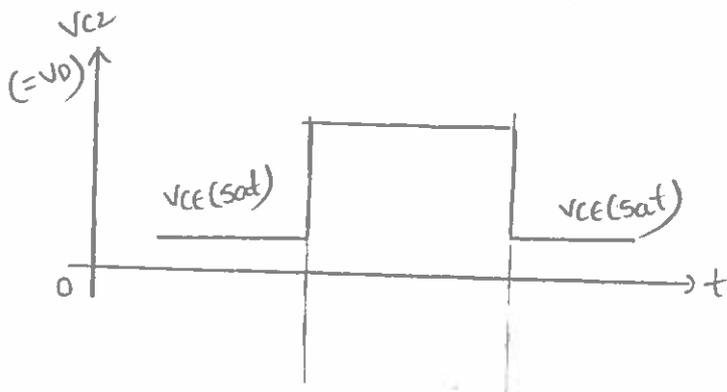
With this knowledge of  $\delta$ , the waveforms of  $v_B$  and  $v_C$  are drawn. The waveforms of the voltages at point A (base of  $Q_1$ ) and D (collector of  $Q_2$ ) also can be plotted for the duration of the quasi-stable state. These are as shown.



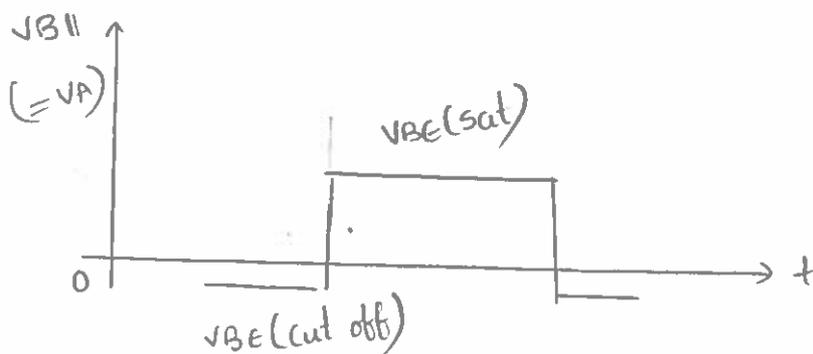
(a)  $v_{B2}(=v_B)$  - voltage at the base of  $Q_2$



(b)  $v_{C1}(=v_C)$  - voltage at collector of  $Q_1$ .

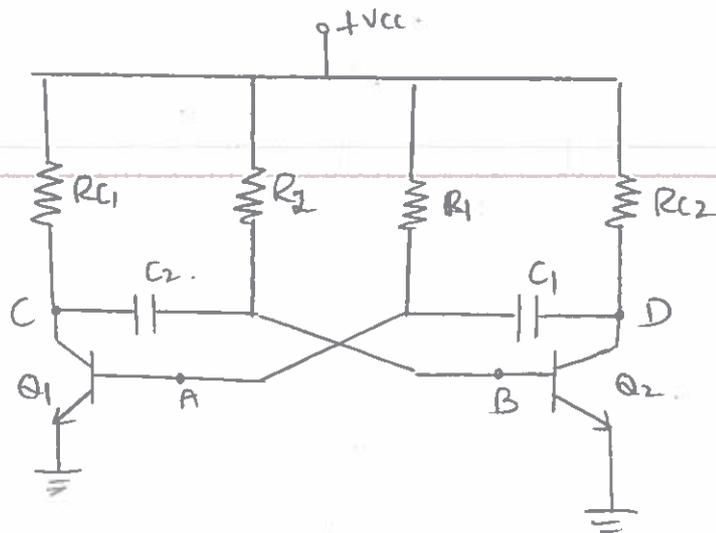


(c)  $v_{C2}(=v_D)$  - voltage at the collector of  $Q_2$ .



(d)  $v_{B1}(=v_A)$  - voltage at the base of  $Q_1$ .

## Principle of operation of Astable multivibrator:-



A collector-coupled astable multivibrator using n-p-n transistor is shown in figure.

$R_1, R_2, R_{C1}, R_{C2}$  --- resistors.

$C_1, C_2$  - capacitors

$Q_1, Q_2$  --- n-p-n transistors

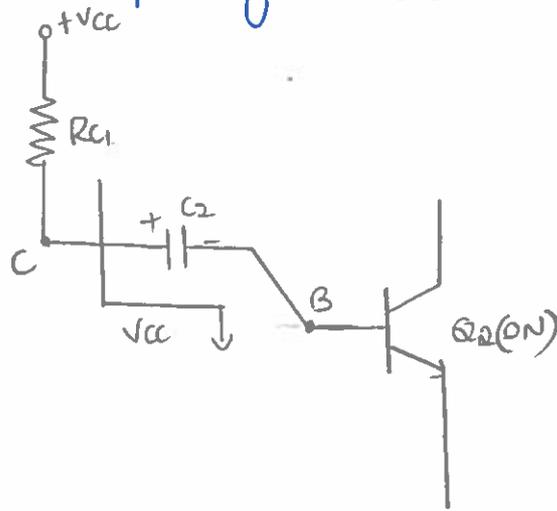
(silicon or germanium)

$V_{CC}$  - supply voltage.

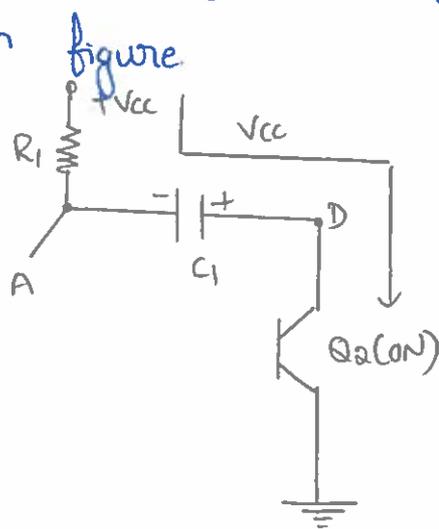
The working of an astable multivibrator can be studied w.r.t the above circuit.

Let it be assumed that the multi is already in action and is oscillating i.e. switching between the two states. Let it be further assumed that at the instant considered,  $Q_2$  is ON and  $Q_1$  is OFF.

(i) Since  $Q_2$  is ON, capacitor  $C_2$  charges through resistor  $R_{C1}$ , as is evident from figure. The voltage across  $C_2$  is  $V_{CC}$  with polarity as shown.



(ii) Capacitor  $C_1$  discharges through resistor  $R_1$ , as is obvious from figure.



The voltage across  $C_1$  when it is about to start discharging is  $V_{CC}$  with polarity as shown.

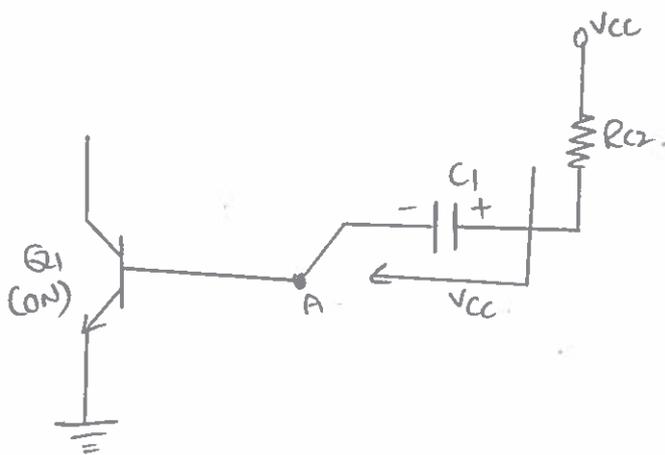
[Capacitor  $C_1$  get charged to  $V_{CC}$  with polarity as shown, when  $Q_1$  is ON].

As capacitor  $C_1$  discharges more and more, the potential of point A becomes more and more positive

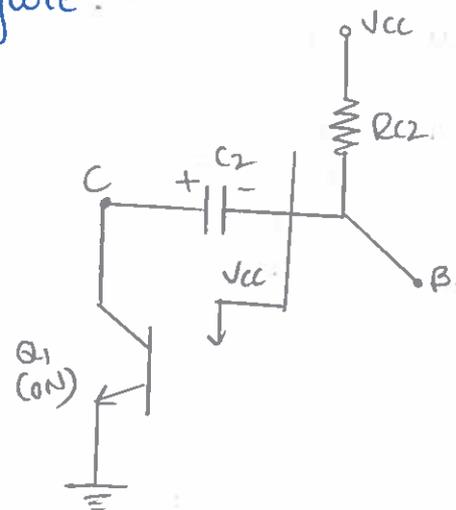
(or less and less negative), and eventually  $V_A$  becomes equal to  $V_r$ , the cut-in voltage of  $Q_1$ . For  $V_A > V_r$ , transistor  $Q_1$  starts conducting. when  $Q_1$  is ON,  $Q_2$  becomes OFF.

Similar operations repeat when  $Q_1$  becomes ON and  $Q_2$  becomes OFF.

Thus with  $Q_1$  ON and  $Q_2$  OFF, capacitor  $C_1$  charges through resistor  $R_{C2}$  and capacitor  $C_2$  discharges through resistor  $R_2$ . The charging and discharging paths are as shown in figure.



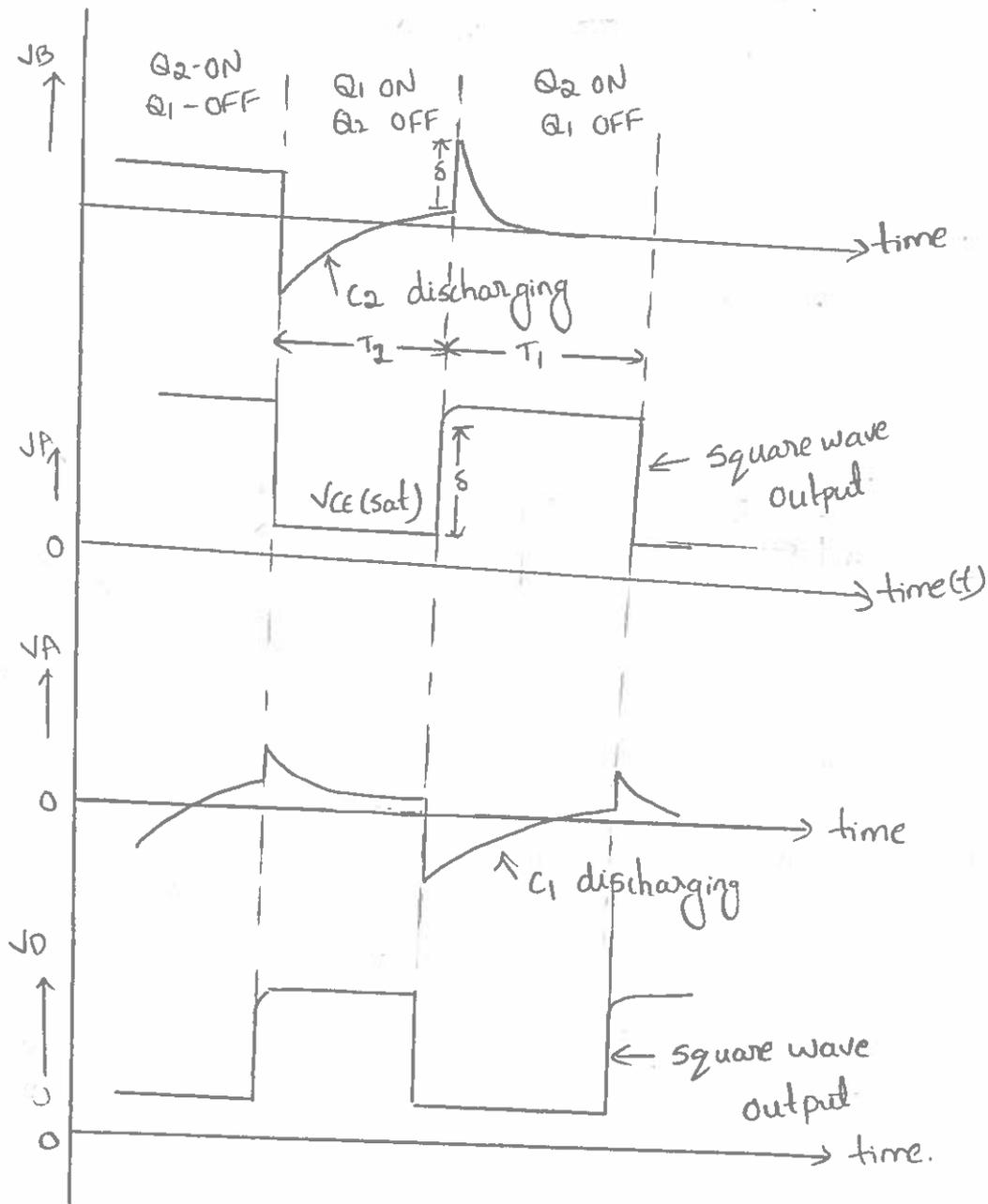
$C_1$  charging



$C_2$  charging

As capacitor  $C_2$  discharges more and more it is seen that the potential of point B becomes less and less negative (or more and more positive), and eventually  $V_B$  becomes equal to  $V_r$ , the cut-in voltage of  $Q_2$  when  $V_B > V_r$ , transistor  $Q_2$  starts conducting. when  $Q_2$  becomes ON,  $Q_1$  becomes OFF.

It is thus seen that the circuit keeps on switching continuously between the two quasi-stable states and once in operation, no external triggering is needed. Square wave voltages are generated at the collector terminals of  $Q_1$  &  $Q_2$  i.e. at points C and D. The input and output waveforms are as shown.

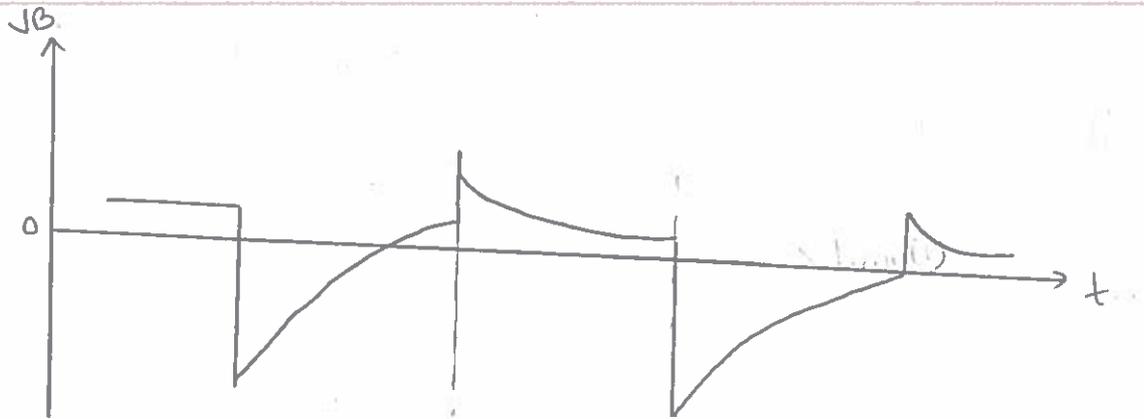


- $V_B$  - Input voltage at the base of  $Q_2$ .
- $V_C$  - output voltage at the collector of  $Q_1$  (square wave)
- $V_A$  - Input voltage at the base of  $Q_1$ .

$v_o$  = output voltage at the collector of  $Q_2$  (square wave).

Expression for  $T$ , the period of oscillations:-

Consider the wave form of figure.



The input waveform at the base of  $Q_2$  for the period  $T = T_1 + T_2$  is shown.

When the capacitor  $C_2$  discharges progressively, the voltage  $v_B$  rises exponentially from  $-V_{CC}$  upto  $v_B = V_f$ . Let  $v_{in}$  denote the initial value and  $v_f$  denote the final value of  $v_B$ .

The mathematical expression for the exponentially rising voltage is

$$v_B = v_f - (v_f - v_{in}) e^{-t/R_2 C_2}$$

We have  $v_{in} = -V_{CC}$  and  $v_f = V_{CC}$ .

$$\therefore v_B = V_{CC} - [V_{CC} - (-V_{CC})] e^{-t/R_2 C_2}$$

$$= V_{CC} - 2V_{CC} \cdot e^{-t/R_2 C_2}$$

At  $t = T_2$ , we have  $v_B = v_V = 0$ .

$$\therefore V_{CC} - 2V_{CC} \cdot e^{-T_2/R_2C_2} = 0$$

$$\text{i.e. } 1 - 2e^{-T_2/R_2C_2} = 0$$

$$\text{or } e^{T_2/R_2C_2} = 2$$

$$\text{i.e. } \frac{T_2}{R_2C_2} = \log_e 2 = 0.69$$

$$\text{or } T_2 = 0.69 R_2 C_2 \rightarrow \textcircled{1}$$

Similarly it can be proved that  $T_1 = 0.69 R_1 C_1 \rightarrow \textcircled{2}$

$\therefore$  period of oscillations  $T = T_1 + T_2$

$$= 0.69 R_1 C_1 + 0.69 R_2 C_2$$

$$= 0.69 (R_1 C_1 + R_2 C_2)$$

$$\text{or } T = 0.69 \times 2RC, \text{ if } R_1 = R_2 = R$$

$$\text{and } C_1 = C_2 = C$$

$$\text{i.e. } \boxed{T = 1.38RC}$$

Also frequency of oscillations  $f = 1/T$ .

Note (i) :- For symmetrical square wave.

$$T = 0.69 (2RC) \text{ or } T = 1.38RC$$

ii) For unsymmetrical square wave.

$$T = 0.69 (R_1 C_1 + R_2 C_2)$$

## Principle of operation of Schmitt trigger:-

Consider the circuit of figure.

Let the input to the transistor  $Q_1$  be a sinusoidal voltage  
 $v_i = v_m \sin \omega t$ .

When  $v_i = 0$ , obviously  $Q_1$  is OFF. However  $Q_2$  gets adequate base drive from the supply voltage  $V_{CC}$  and hence it conducts, but it would be conducting in the active region. Let  $i_{c2}$  denote the collector current and  $i_{e2}$  the base current of  $Q_2$ .

Due to flow of current  $i_{e2}$  through  $R_e$ , there is a voltage drop  $i_{e2} R_e$ .

$\therefore$  potential of point E w.r.t ground =  $i_{e2} R_e$ .

Assuming

$i_{e2} = i_{c2}$ , we have  $V_E = i_{c2} R_e$ .

From the circuit, it is evident that for  $Q_1$  to conduct  $v_i$  should rise to the voltages  $(V_E + V_r)$ , where  $V_r$  is the cut in voltage of  $Q_1$ .

Hence as long as  $v_i < V_E + V_r$ ,  $Q_1$  cannot conduct, and  $Q_1$  starts conducting when  $v_i = V_E + V_r$

i.e  $v_i = i_{c2} R_e + V_r$ .

As  $Q_1$  conducts, the potential of its collector terminal <sup>5th year</sup> C progressively decreases (due to increasing  $i_{C1}R_{C1}$  drop), and because of feedback, the potential of point B progressively diminishes, with the result that  $Q_2$  becomes OFF. This value of the input voltage  $v_i$  which makes  $Q_1$  conduct is termed as "upper trigger point" or "upper trip point", abbreviated as UTP, it is denoted as  $V_1$ .

$$\therefore V_1 = V_{E2} + V_r \text{ or}$$

$$V_1 = i_{C2} R_2 + V_r$$

Now  $Q_2$  is OFF and  $Q_1$  is ON, for  $v_i > \text{UTP}$ .

If the original state viz  $Q_2$  ON and  $Q_1$  OFF is to be restored, it is essential that  $v_i$  decreases.

As  $v_i$  decreases more and more,  $V_A$  decreases, the potential of point C progressively increases [As  $v_i$  decreases,  $i_{C1}$  decreases, volt drop  $i_{C1}R_{C1}$  decreases and hence potential of point C, and hence that of B would increase], and eventually when  $V_B$  become equal to  $V_A = V_{BE}(\text{act}) + V_E$ ,  $Q_2$  starts conducting, and  $Q_1$  goes OFF.

This value of the input voltage which makes  $Q_2$  conduct again is termed as Lower trigger point (or lower trip point), abbreviated as LTP. It is denoted as  $V_2$ .

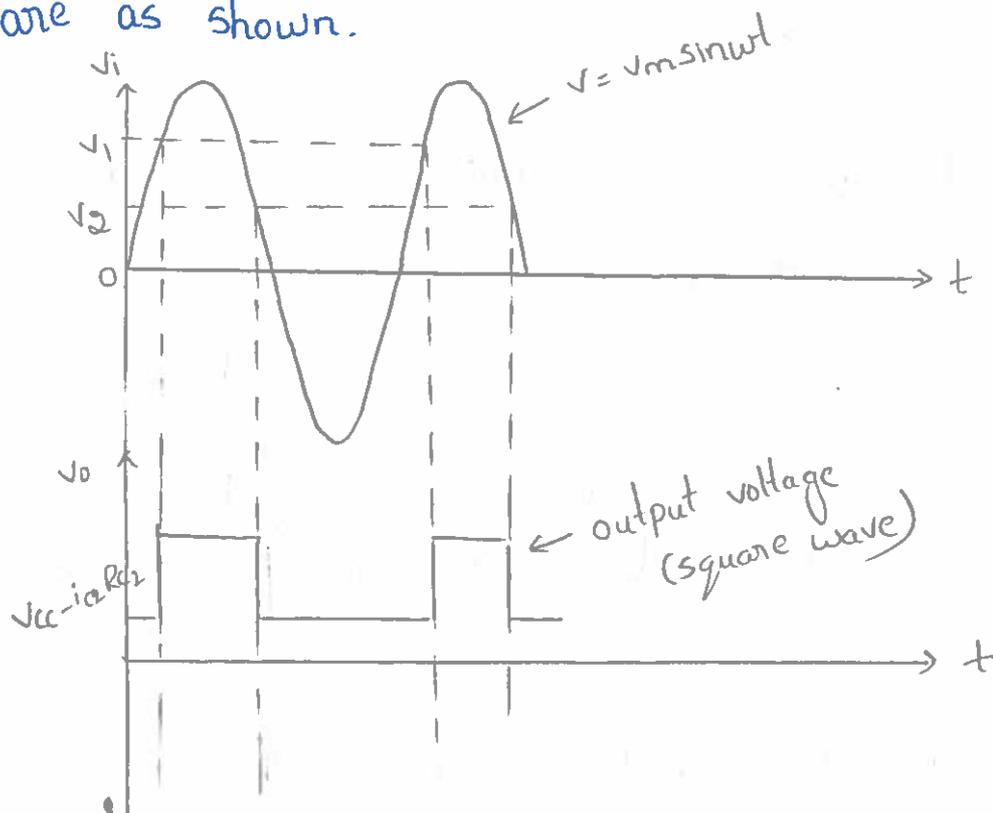
$$\therefore \text{we have } V_2 = V_{BE(\text{act})} + V_E$$

$$\text{or } \boxed{V_2 = V_{BE(\text{act})} + i_{C1} R_e}$$

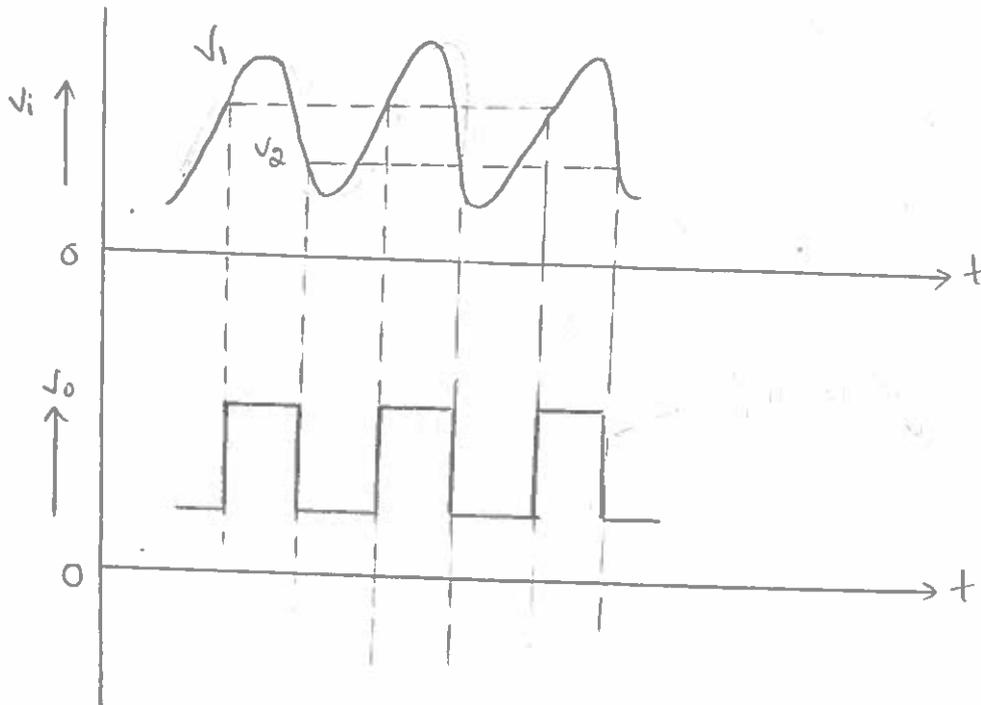
$$\text{Since } V_E = i_{E1} R_e = i_{C1} R_e$$

Note:  $\boxed{\begin{aligned} \text{UTP} = V_1 &= V_T + i_{C2} R_{e1} \text{ and} \\ \text{LTP} = V_2 &= V_{BE(\text{act})} + i_{C1} R_{e1} \end{aligned}}$

It should be clearly understood that UTP and LTP are two specific values of the input voltage  $V_i$ . The waveforms of input signal  $V_i$  and output voltage  $V_o$  are as shown.



It is observed that the output is an unsymmetrical square wave. Thus the Schmitt trigger readily converts sinusoidal wave to square wave. It is, therefore, termed as "sine to square converter" or "squaring circuit".



It should be noted, however, that the output of a Schmitt trigger is a square wave, whatever the wave form of the input signal (not necessary sine wave; it may be of any arbitrary form). This is evident from figure.

The following values may be assumed for n-p-n silicon and germanium transistors.

(i) Silicon transistor

$$V_r = 0.5V$$

$$V_{BE(Act)} = 0.6V$$

$$V_{BE(Sat)} = 0.7V$$

(ii) Germanium transistor

$$V_r = 0.1V$$

$$V_{BE(Act)} = 0.2V$$

$$V_{BE(Sat)} = 0.3V$$

For p-n-p transistor all values are negative.

Hysteresis:-

Consider the circuit of figure.

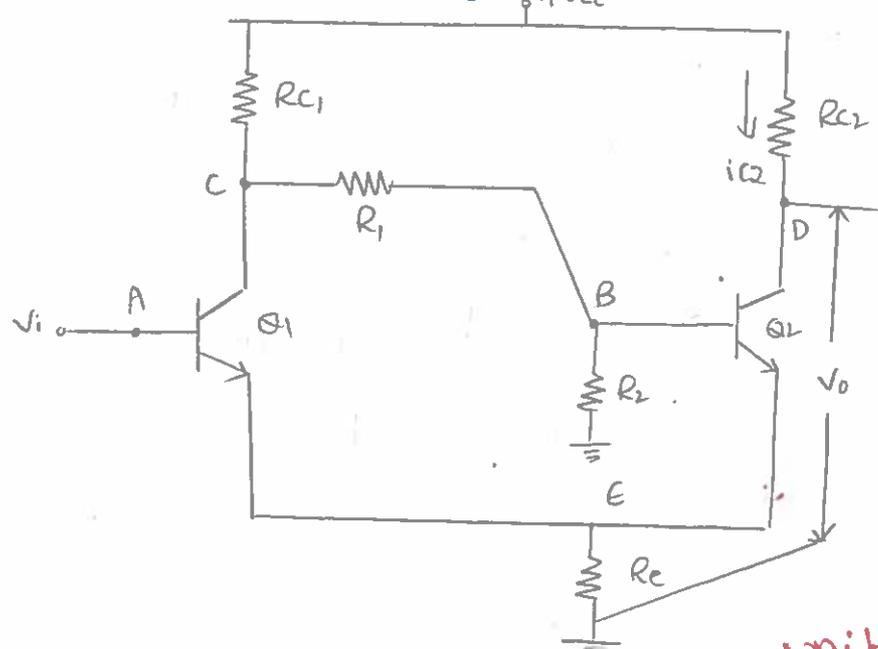
In order to study the hysteresis loop on the transfer characteristic of the schmitt trigger, it is necessary to obtain values of  $v_o$ , the output, for varying input  $v_i$ .

Let us consider an example.

A schmitt trigger has the following specifications.

$V_{CC} = 10V$ , U.T.P = 5V, L.T.P = 3V and  $i_C = 2mA$ , and

$$R_{C1} = R_{C2} = 3k\Omega$$



mit-5, 29/6/21

(i) When  $Q_2$  conducts and  $Q_1$  is OFF

We have

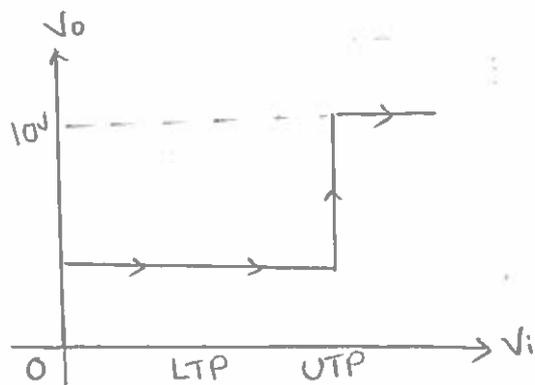
$$\text{Output } V_o = V_{cc} - i_{c2} R_{c2} \rightarrow (1)$$

$$10 - 2 \times 3 = 4 \text{ volts}$$

When  $Q_2$  is OFF,  $i_{c2} = 0$  and hence potential of point D =  $V_{cc}$

$$\therefore V_o = V_{cc} = 10 \text{ volts} \rightarrow (2)$$

When the input voltage is increasing from zero, the transfer characteristics is as shown in figure.



Between  $V_i = 0$  and  $V_i = UTP$ , the output remains constant at 4V. During this period,  $Q_2$  is ON and relation (1) holds good.

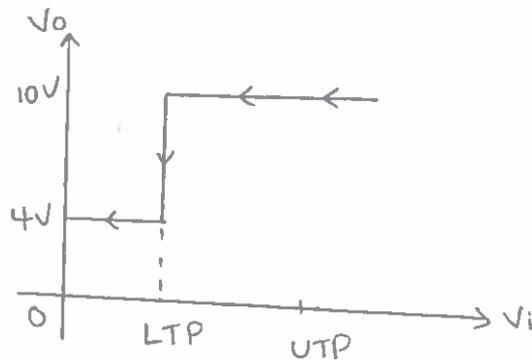
For  $V_i = UTP$ ,  $Q_1$  begins to conduct. As long as  $Q_1$  is ON,  $Q_2$  is OFF and relation (2) holds good.

(ii) As  $V_i$  increases beyond UTP, reaches its positive peak and falls to LTP, output remains at 10V.

When  $v_i = LTP$ ,  $Q_2$  again begins to conduct and  $Q_1$  goes OFF.

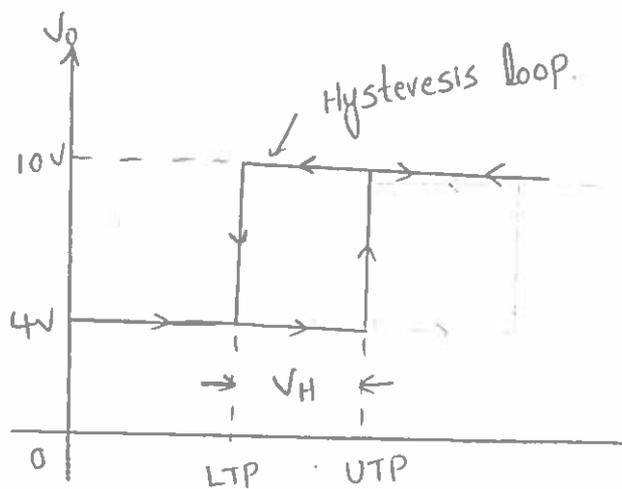
At  $v_i = LTP$ , output instantly falls to 4V and remains at that level until  $v_i$  again becomes zero.

The transfer characteristic when  $v_i$  decreases from its positive peak is as shown in figure.



The curves of figure may be clubbed together and the total transfer characteristic (graph of  $v_o$  vs  $v_i$ ) can be plotted. It is as shown in figure.

$$\begin{aligned} V_H &= \text{Hysteresis voltage} \\ &= UTP - LTP \end{aligned}$$

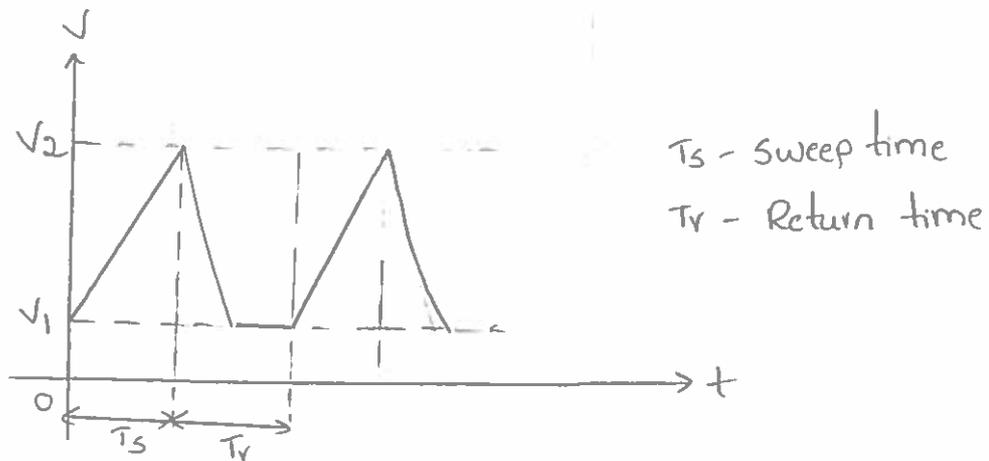


The closed loop shown in the figure is termed as "hysteresis loop". The voltage difference between UTP and LTP, represented by the loop width, is called "hysteresis voltage".

$$\begin{aligned} \text{we have hysteresis voltage} &= \text{UTP} - \text{LTP} \\ &= v_1 - v_2. \end{aligned}$$

General wave form of time-base (or sweep) voltage:-

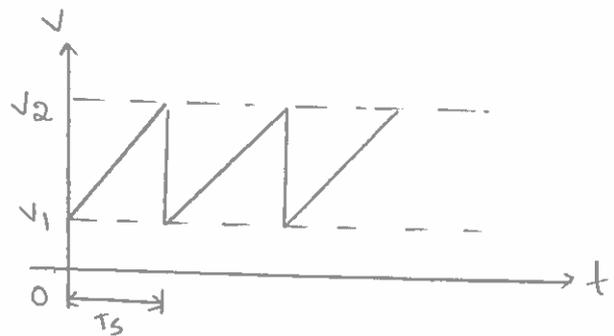
A sweep voltage which, ideally, varies linearly with time has the general wave form shown in figure.



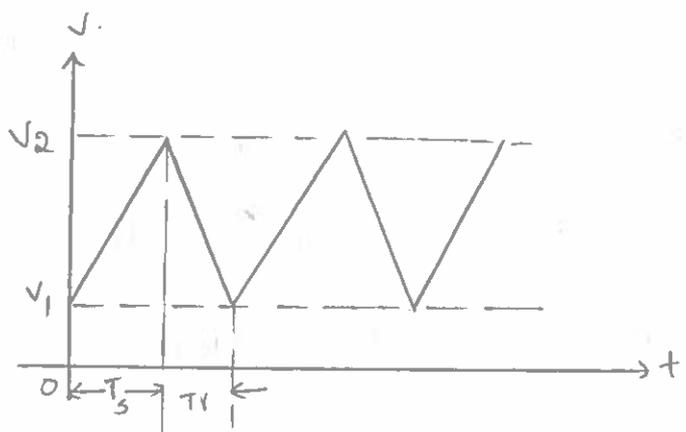
From the figure, it is evident that the voltage, starting from an initial value,  $v_1$  rises linearly to a peak value  $v_2$ , and falls to the initial value  $v_1$  over a short period of time. The time taken by the wave to reach the maximum value, starting from the initial value, is termed as "sweep time",

and the time during which it returns to the initial value is termed as "return time" or restoration time or fly-back time. Sweep time is denoted by  $T_s$  and return time is denoted as  $T_r$ .

The return time  $T_r$  and the wave shape during the return time normally do not matter, but in some special applications, the restoration time  $T_r$  should be quite small in comparison with the sweep time  $T_s$ . If the voltage returns to the initial value instantaneously, the wave form would be as shown in figure. It is termed as "saw-tooth wave".



A sweep voltage may also be a triangular wave as shown.



## Deviation from linearity:-

In practice, the signals generated by time-base circuits are not perfectly linear. Additionally, even if the signals are linear, they suffer distortion when transmitted through coupling networks.

The deviation of a signal from linearity can be fully described in terms of three types of errors

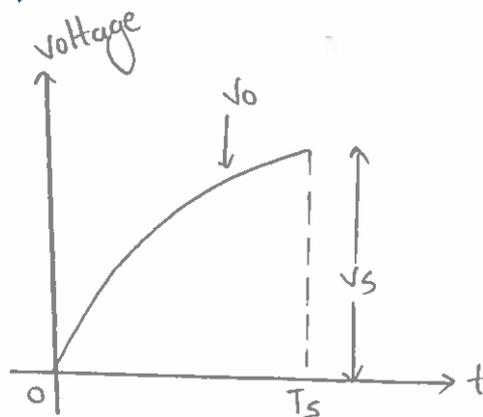
- \* Sweep error.
- \* transmission error.
- \* Displacement error.

These errors and their inter-relationship are studied in the following paragraphs.

Sweep error (or sweep-speed error or slope error or slope-speed error ( $e_s$ ))

Consider the waveform shown in figure.

The waveform is not perfectly linear, and as such, the slopes of the line at different points are different.



Slope error is defined as the difference between initial slope (i.e. slope at  $t=0$ ) and final slope (i.e. slope at  $t=T_s$ ), expressed as the fraction of the initial slope.

It is denoted as  $e_s$ .

We have, slope error  $e_s = \frac{\text{Initial slope} - \text{final slope}}{\text{Initial slope}}$

$$\text{or } e_s = \frac{\left(\frac{dv_s}{dt}\right)_{t=0} - \left(\frac{dv_s}{dt}\right)_{t=T_s}}{\left(\frac{dv_s}{dt}\right)_{t=0}}$$

The exponentially increasing voltage is mathematically expressed as  $v_s = V(1 - e^{-t/RC})$  where  $RC =$  time constant of the circuit.

Differentiating w.r.t. we get

$$\begin{aligned}\frac{dv_s}{dt} &= V(0 - e^{-t/RC}) \cdot \left(-\frac{1}{RC}\right) \\ &= \frac{V}{RC} \cdot e^{-t/RC}\end{aligned}$$

$$\therefore \left(\frac{dv_s}{dt}\right)_{t=0} = \frac{V}{RC} \text{ and } \left(\frac{dv_s}{dt}\right)_{t=T_s} = \frac{V}{RC} \cdot e^{-T_s/RC}$$

$$\therefore \text{Slope error } e_s = \frac{\frac{V}{R_c} - \frac{V}{R_c} \cdot e^{-T_s/R_c}}{V/R_c}$$

$$\text{or } e_s = 1 - e^{-T_s/R_c}$$

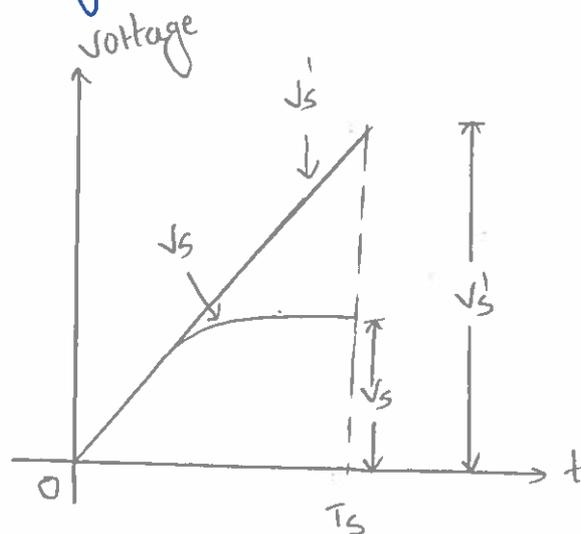
If  $T_s \ll R_c$ , we have  $e^{-T_s/R_c} \approx 1 - \frac{T_s}{R_c}$

$$\therefore e_s = 1 - \left(1 - \frac{T_s}{R_c}\right) = \frac{T_s}{R_c}$$

$$\therefore \text{Slope speed error } \boxed{e_s = \frac{T_s}{R_c}}$$

Transmission error ( $e_t$ ):-

In the chapter on linear wave-shaping, we studied that when a ramp voltage is transmitted through  $R_c$  high pass network, it gets distorted as shown in figure.



At the end of time  $T_s$ , the value of the voltage  $V_s < V_s'$ , due to the fact that it has deviated from linearity.

Transmission error is defined as the difference between the input and the output, expressed as a fraction of the input.

Let at  $t = T_s$ , input =  $V_s'$  and output  $V_s$ .

We have transmission error =  $\frac{\text{Input} - \text{output}}{\text{Input}}$

$$\text{or } e_1 = \frac{V_s' - V_s}{V_s'}$$

As already studied, since the output voltage  $V_s$  exponentially increases, we have

$$V_s = V(1 - e^{-t/RC})$$

$$1 - e^{-t/RC} = 1 - \left[ 1 - \frac{t}{RC} + \frac{(t/RC)^2}{2} - \frac{(t/RC)^3}{6} + \dots \right]$$

$$= 1 - 1 + \frac{t}{RC} - \frac{(t/RC)^2}{2}, \text{ neglecting the higher order terms}$$

$$= \frac{t}{RC} \left( 1 - \frac{t}{2RC} \right)$$

$$\therefore V_s = \frac{Vt}{Rc} \left( 1 - \frac{t}{2Rc} \right)$$

At  $t = T_s$ ,  $V_s = V_s$

$$\therefore V_s = \frac{VT_s}{Rc} \left( 1 - \frac{T_s}{2Rc} \right) \rightarrow \textcircled{1}$$

The input voltage is a ramp and hence it can be expressed as

$$v_s' = \alpha t, \text{ where } \alpha = \text{slope.}$$

$$\text{But Slope} = \frac{V}{Rc}$$

$$\therefore v_s' = \frac{Vt}{Rc}$$

At  $t = T_s$ ,  $v_s' = V_s$ .

$$\therefore V_s = \frac{VT_s}{Rc} \rightarrow \textcircled{2}$$

$$\text{Transmission error } e_t = \frac{\frac{VT_s}{Rc} - \frac{VT_s}{Rc} \left( 1 - \frac{T_s}{2Rc} \right)}{\frac{VT_s}{Rc}}$$

$$= 1 - 1 \left( 1 - \frac{T_s}{2Rc} \right)$$

$$= \frac{T_s}{2Rc}$$

$$\therefore e_t = \frac{T_s}{2Rc}$$

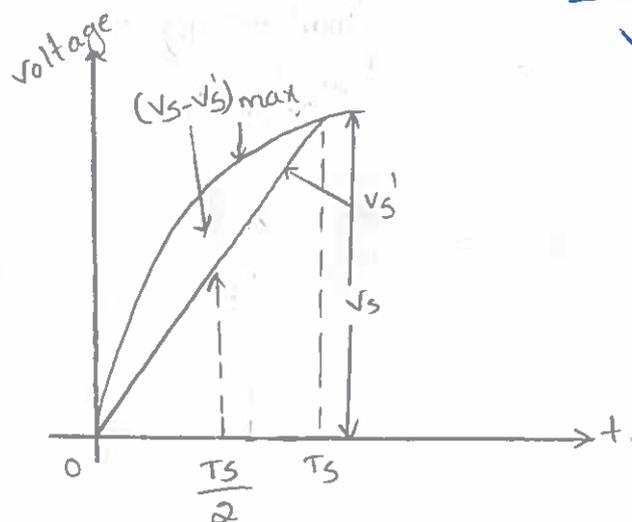
unit-5, 37/42

Displacement Error ( $e_d$ ):-

Another important criterion of linearity is the maximum difference between the actual sweep voltage and the linear sweep which passes through the beginning and end points of the actual sweep.

Displacement error may be defined as the ratio of maximum deviation of the actual sweep from the linear sweep to the amplitude of the sweep voltage. Thus, referring to figure we have.

$$\text{Displacement error, } e_d = \frac{\text{Maximum deviation}}{\text{Sweep amplitude}}$$
$$= \frac{|v_s - v_s'|_{\text{max}}}{v_s}$$



since the actual sweep  $v_s$  is exponential, we

have  $v_s = v(1 - e^{-t/Rc})$

$$= \frac{v_i}{Rc} \left(1 - \frac{1}{2Rc}\right), \text{ as already seen.}$$

The ideal (linear) ramp is given as  $v_s' = \alpha t = \left(\frac{v}{RC}\right)t$ ,

Since slope  $\alpha = \frac{v}{RC}$

$$\begin{aligned}\therefore \text{Deviation} &= |v_s - v_s'| = \frac{vt}{RC} \left(1 - \frac{t}{2RC}\right) - \frac{vt}{RC} \\ &= \frac{vt}{RC} \cdot \frac{t}{2RC} \rightarrow \textcircled{1}\end{aligned}$$

taking only the amplitude.

The deviation is maximum at  $t = \frac{T_s}{2}$

$\therefore$  Maximum deviation at  $t = \frac{T_s}{2}$  is  $|v_s - v_s'|_{\max}$

$\therefore$  putting  $t = \frac{T_s}{2}$ , we have

$$|v_s - v_s'|_{\max} = \frac{v(T_s/2)}{RC} \cdot \frac{(T_s/2)}{2RC} \rightarrow \textcircled{2}$$

$$\text{we have } v_s' = \alpha t = \frac{vt}{RC} \rightarrow \textcircled{3}$$

At  $t = T_s$ ,  $v_s' = v_s$  from figure

$$\therefore v_s = \frac{vT_s}{RC}, \text{ putting } t = T_s \text{ in (3)} \rightarrow \textcircled{4}$$

Displacement error  $e_d = \frac{\text{Maximum deviation}}{\text{Sweep amplitude}}$

$$= \frac{\sqrt{\left(\frac{T_s}{2R_c} \cdot \frac{T_s}{4R_c}\right)}}{\frac{\sqrt{T_s}}{R_c}} \quad \text{from } \textcircled{2} \text{ \& } \textcircled{4}$$

or  $\boxed{e_d = \frac{T_s}{8R_c}}$  on simplification.

Inter-relationship of  $e_d$ ,  $e_s$  and  $e_t$ :-

When

$$e_d = \frac{T_s}{8R_c} \quad \text{and} \quad e_s = \frac{T_s}{R_c} \quad \therefore e_d = \frac{1}{8} e_s$$

$$\text{Also } e_t = \frac{T_s}{2R_c} \quad \text{and} \quad e_d = \frac{T_s}{8R_c} \quad \therefore e_d = \frac{1}{4} e_t$$

Combining we get,

$$\boxed{e_d = \frac{1}{8} e_s = \frac{1}{4} e_t.}$$

If one of the errors is known, the other errors can easily be computed on the basis of the above relationship.

## Methods of generating time-base waveform:-

There are many practical methods by which sweep voltages, which are practically linear, can be generated. The more important amongst these are

### \* Exponential charging:-

In this method, a capacitor is charged through a resistor to a voltage which is quite small as compared to the charging voltage.

### \* Constant current charging:-

In this method, a capacitor is charged from a constant current source. As will be seen later, the voltage across the capacitor is a ramp voltage.

### \* Miller circuit:-

In this method, a step voltage is converted to a ramp, using an integrating circuit like the miller integrator.

### \* Bootstrap circuit:-

In this method, a constant current is passed through a capacitor and the voltage across the capacitor is a ramp. Constant current is obtained by maintaining a constant voltage across a fixed

resistor in series with the capacitor.

\* Compensating networks:-

To improve the linearity of miller and boot strap time-base generators, several compensating circuits are introduced.

These methods are studied in brief.

\* Comparison between miller and Bootstrap Sweep Circuits

Although both the bootstrap sweep circuit and miller sweep circuit generate ramp voltage using the same basic principle, they differ in some aspects so far as their functioning is concerned. The following are the dissimilarities.

Bootstrap Sweep circuit	Miller sweep circuit
* The circuit employs positive feedback.	* The circuit employs negative feedback.
* The circuit generates positive going ramp.	* The circuit generates -ve going ramp.
* The circuit employs an emitter follower whose gain is nearly unity.	* The circuit requires an amplifier whose gain is very very large (ideally infinite).
* The amplifier must have high input resistance.	* Amplifier with high input resistance is not very essential.