# Multivibrators

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# Multivibrator

A **multivibrator** is an electronic circuit used to implement a variety of simple two-state devices such as relaxation oscillators, timers and flip-flops. It consists of two amplifying devices (transistors, OP-AMP or other devices) cross-coupled by resistors and/ or capacitors. The first multivibrator circuit, the astable multivibrator oscillator, was invented by Henri Abraham and Eugene Bloch during 1$^{st}$ World War I. They called their circuit a "multivibrator" because its output waveform was rich in harmonics.

The three types of multivibrator circuits are:

- **Bistable multivibrator**, in which the circuit is stable in either state. It can be flipped from one state to the other by an external trigger pulse. This circuit is also known as a flip-flop. It can store one bit of information, and is widely used in digital logic and computer memory.
- **Monostable multivibrator**, in which one of the states is stable, but the other state is unstable (transient). A trigger pulse causes the circuit to enter the unstable state. After entering the unstable state, the circuit will return to the stable state after a set time. Such a circuit is useful for creating a timing period of fixed duration in response to some external event. This circuit is also known as a **one shot**.
- **Astable multivibrator**s are the most commonly used type of **multivibrator** circuit. It is a free running oscillator that have no permanent “meta” or “steady” state but are continually changing there output from one state (LOW) to the other state (HIGH) and then back again.

# Bistable multivibrator

## Applications

The bistable circuit is a function of a flip flop. It’s able to store bits indefinitely unless desired to be changed. It can be used basically for any circuit in where you want manual control over outputs which you want to keep stable states unless you trigger it to change. Bistable circuits are used as memory storage devices, timing circuits, frequency dividers, electronic toggle switch, counting circuits, shift registers, clock pulse generators, relay controllers and even in the field of radar and communications.

# Bistable multivibrator

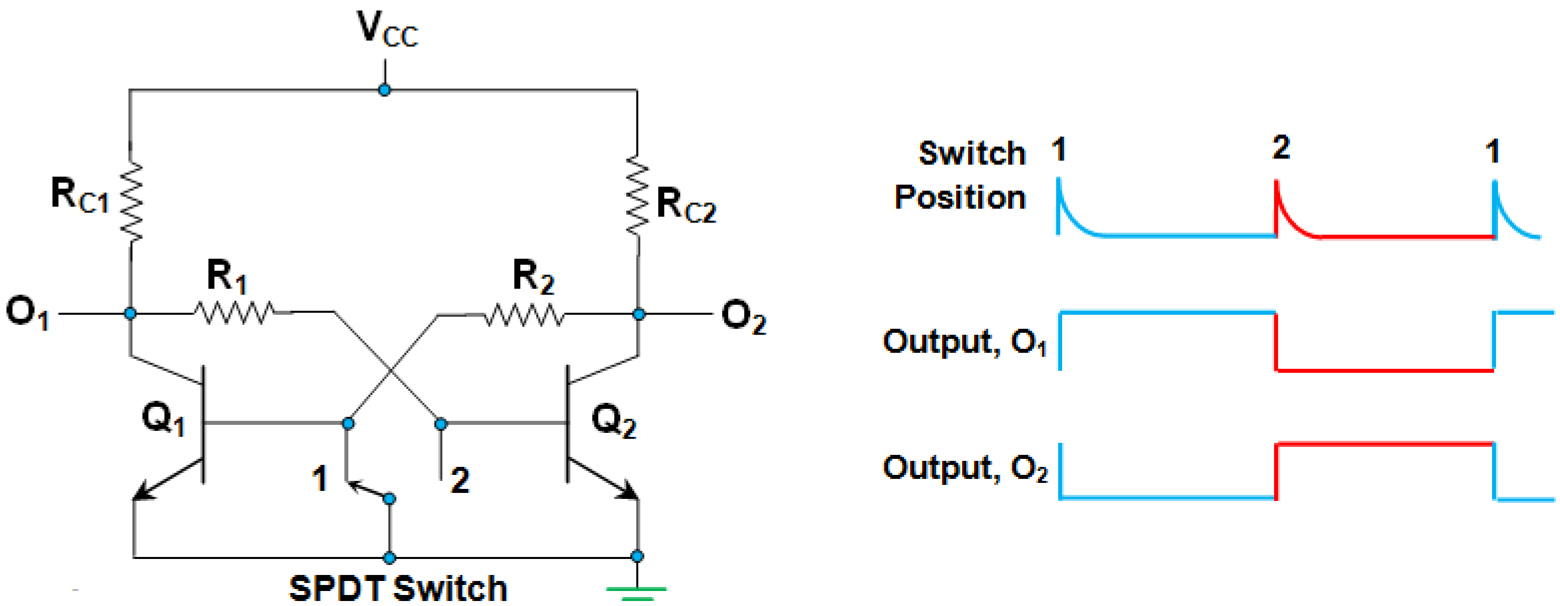



Figure 1 Bistable Multivibrator Designed Using BJTs

SPDT-Single pole double throw

# Operation

- **Bistable Multivibrators** depend upon the external triggers so as to switch between their two permissible stable states. These circuits are also referred to as Trigger Circuits or Eccles Jordon Circuits or Scale-of-2 Toggle Circuits or Binary or more popularly as Flip-Flops, forming the basic building blocks of sequential digital systems. These circuits can be designed in different ways, say for example, they can compose of transistors or Op-Amps or 555 timer ICs along with passive components, the resistors. Figure 1 shows such a circuit designed using two NPN bipolar junction transistors (BJTs) $Q_1$ and $Q_2$ and four resistors $R_{C1}$, $R_{C2}$, $R_1$ and $R_2$.

- Initially, let us consider that the SPDT switch is position 1 which in turn grounds the base of the transistor $Q_1$. As a result, $Q_1$ will be OFF (cutoff region) while its collector will be held at $V_{CC}$, due to which the output at $O_1$ will go high. This in turn forward biases the BE junction of transistor $Q_2$, switching it ON (into saturation mode of operation). Due to this, the collector current flows through the collector resistor $R_{C2}$, shorting the collector terminal of $Q_2$ to ground. Thus, for this case, the output at $O_2$ terminal goes low.

- This state of the circuit remains unchanged for an indefinite period of time, unless triggered externally. In this case, the act of changing the switch position from 1 to 2 acts like an external trigger for the circuit. When done so, the base of transistor $Q_2$ will be grounded, switching it OFF (cutoff region). This also causes the $V_{CC}$ to appear at the collector terminal of $Q_2$, which in turn results in a high output at $O_2$ terminal. Further, at this state, $Q_1$ will switch ON (gets into saturation mode of operation) as it has its base connected to the collector terminal of $Q_2$ via $R_2$. Due to this, the collector terminal of $Q_1$ will be shorted to ground, causing the output at the terminal $O_1$ to go low. This state of the circuit is again maintained until triggered once again. Thus output wave-forms obtained at the terminal O1 and O2 are complementary to each other, always.

# Drawback of fig. 1 and solution

- A very high voltage ~ $V_{CC}$ at base emitter junction of Q1/Q2 in Fig. 1 may destroy the transistors.

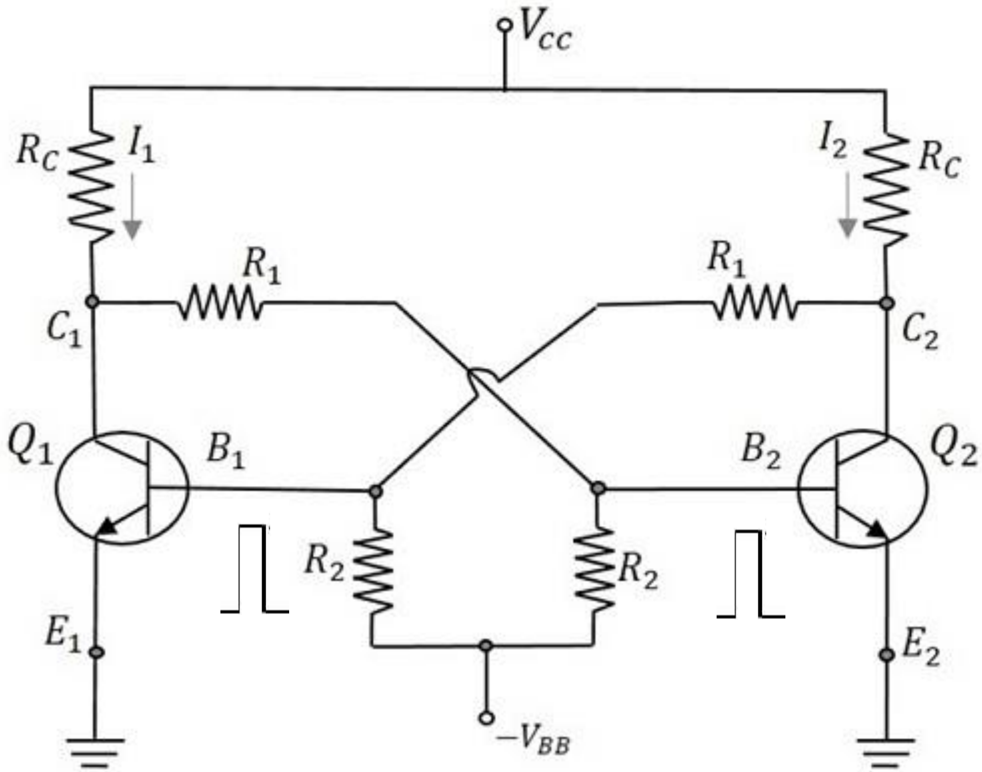
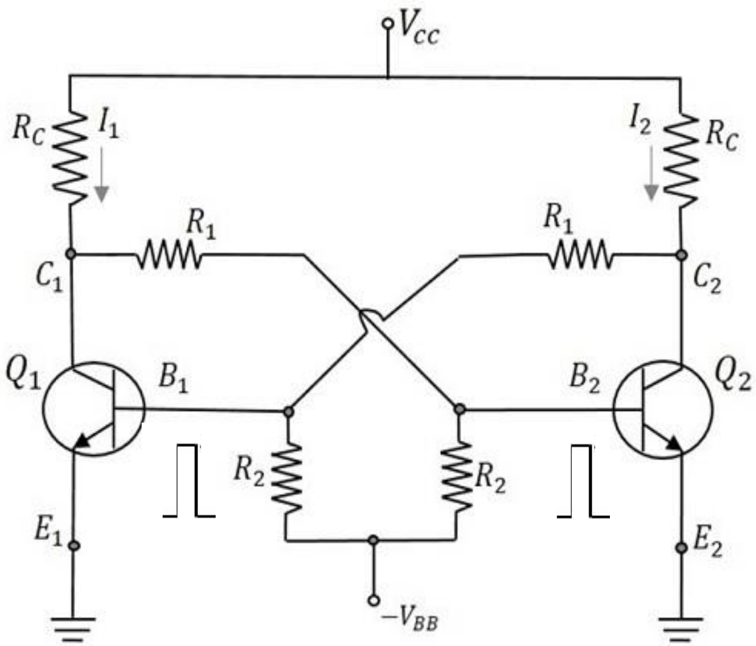
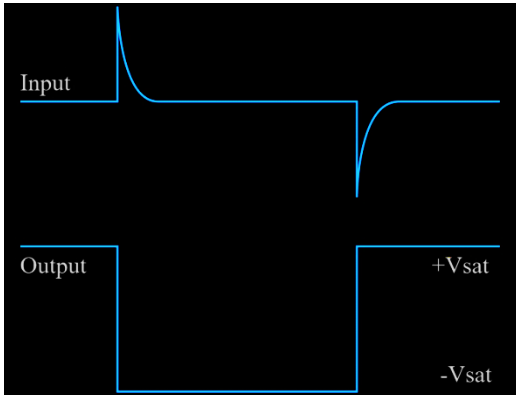



Fig. 2 : Bistable MV with $-V_{BB}$ and $R_2$

Negative power supply $-V_{BB}$ and $R_2$ will keep $V_{BE}$ to a limited reasonable value

# Explanation of Fig. 2




- Because of the symmetry of the circuit we may expect equal amount of quiescent current through two transistors .($I_1$=$I_2$)
- Characteristics of no two transistors are exactly matched to each other.
- Assume $I_1$ is slightly greater than $I_2$. Then $V_{C1}$ will be smaller than $V_{C2}$. Smaller $V_{C1}$ will decrease $V_{B2}$ . Smaller $V_{B2}$ will decrease the collector current in $Q_2$ thereby increasing $V_{C2}$ so does $V_{B1}$. Consequently, $I_1$ will be increased lowering $V_{C1}$. Thus there is positive feedback path and the process cumulates rapidly, driving $Q_1$ to saturation and $Q_2$ to cut-off.
- This state remains indefinitely until a positive trigger pulse is applied at base of $Q_2$.
- If a positive trigger pulse of sufficient magnitude is applied at $B_2$, $Q_2$ starts conducting and goes to saturation driving $Q_1$ to cut-off .
- This state remains indefinitely even after the removal of trigger pulse and the state is changed when a positive trigger pulse of sufficient magnitude is applied at base of $Q_1$.
- Two states are i) $Q_1$ ON, $Q_2$ OFF and ii) $Q_2$ ON, $Q_1$ OFF
- The spacing of the trigger pulse determines the duration of the output voltages,$V_{C1}$ and $V_{C2}$

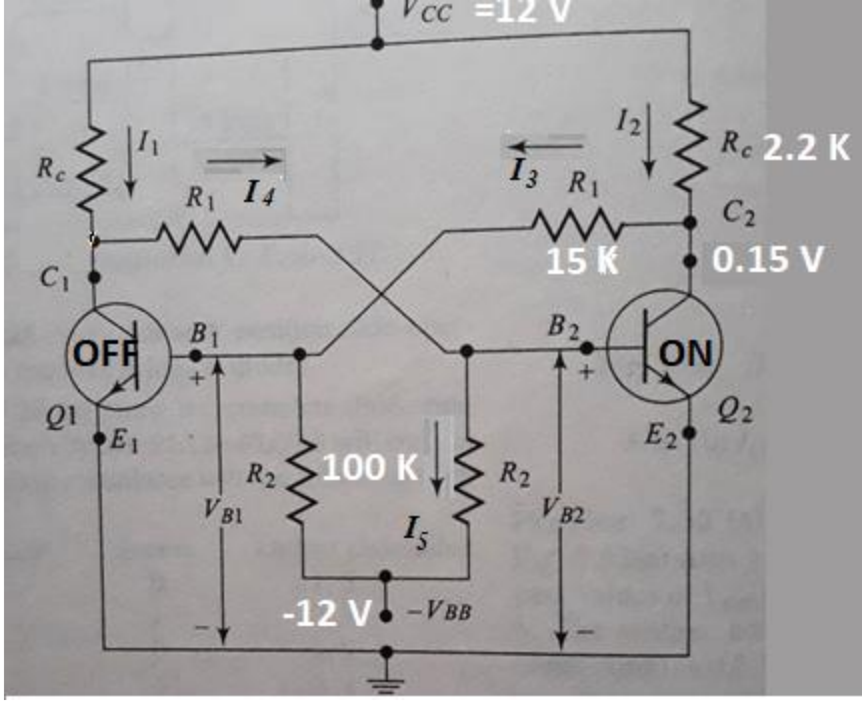
# Example

$V_{cc}$=12 V , -$V_{BB}$=-12V, Rc=22k, $R_1$=15K, $R_2$=100K, $h_{femin}$=20.

Calculate the stable state currents and voltages of the two transistors of Fig .2

Assume Initially $Q_1$ is OFF and $Q_2$ is ON

Assume Vsat=0.15 V and VBE=0.7 V




- $V_{B1} = -12\left(\frac{15}{15+100}\right) + 0.15\left(\frac{100}{15+100}\right) = -1.43$ V
- $I_2 = \frac{12-0.15}{2.2}$ mA= 5.386 mA
- $I_3 = \frac{12+0.15}{15+100}$ mA= 0.106 mA
- $I_{C2} = I_2 - I_3 = 5.28$ mA
- $I_{B2min} = 5.28/20 = 0.264$ mA, to ensure saturation , $I_{B2}$ should be greater its than minimum value
- $V_{B2} = 0.7$V (Given for Q2 ON)
- $I_4 = \frac{12-0.7}{2.2+15}$ mA= 0.657 mA ($Q_1$ OFF, $I_{C1}=0$ mA, $I_1 = I_4$)
- $I_5 = \frac{12+0.7}{100}$ mA= 0.127 mA
- $I_{B2} = I_4 - I_5 = 0.657 - 0.127 = 0.55$ mA > $I_{B2min}$ ensuring saturation of $Q_2$.
- $V_{C1} = V_{CC} - I_1 R_C = 12-(0.657\text{x}2.2) = 10.55$ V and $V_{C2} = V_{CEsat} = 0.15$ V. Voltage swing (VC1-VC2=10.55-0.15=10.4V

Current and voltages are

$V_{C1}$=10.55 V, $V_{B1}$=-1.43 V, $I_{C1}$=0 mA, $I_{B1}$=0 mA

$V_{C2}$=0.15 V, $V_{B2}$=0.7V, $I_{C2}$=5.28 mA, $I_{B2}$=0.55 mA

N.B.: Care should be taken during design so that $V_{BE}$ of the OFF transistor ((here $Q_1$) remains less than its breakdown voltage value.

# A self biased Transistor bistable Multivibrator

- The need of negative power supply may be eliminated

Ex. The self-bias transistor bistable multivibrator shown in Fig. 3 uses *n–p–n* Si transistors. Given that $V_{CC}$ = 15 V, $V_{CE(sat)}$ = 0.2V, $V_{BE(ON)}$ = 0.7 V, $R_C$ = 3 kΩ, $R_1$ = 20 kΩ, $R_2$ = 10 kΩ, $R_E$ = 500 Ω. Find: Stable-state currents and voltages

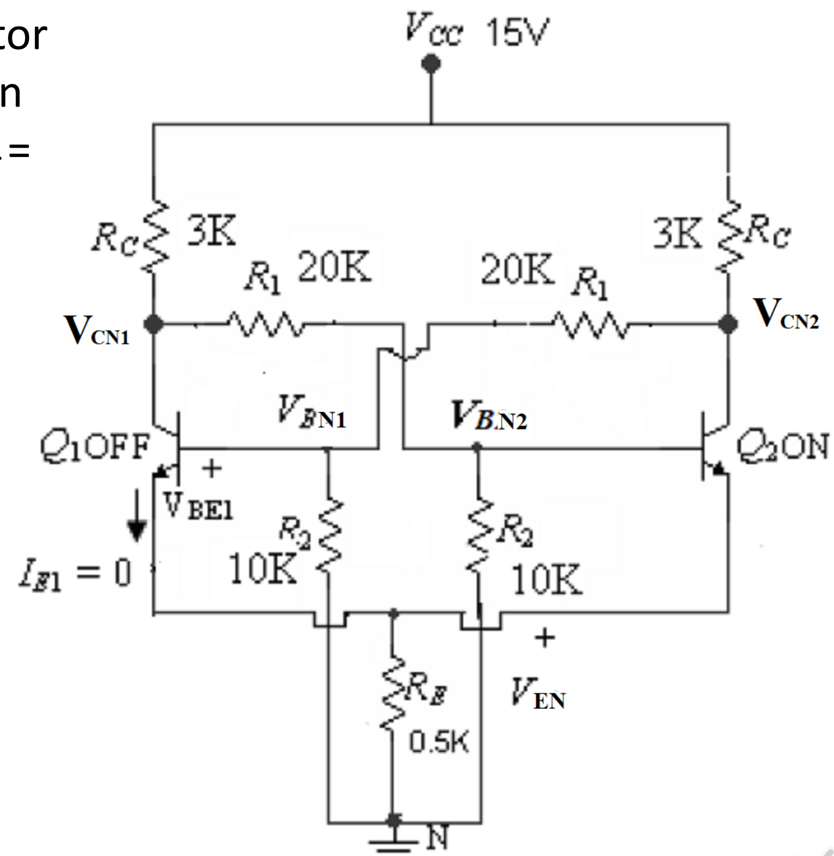
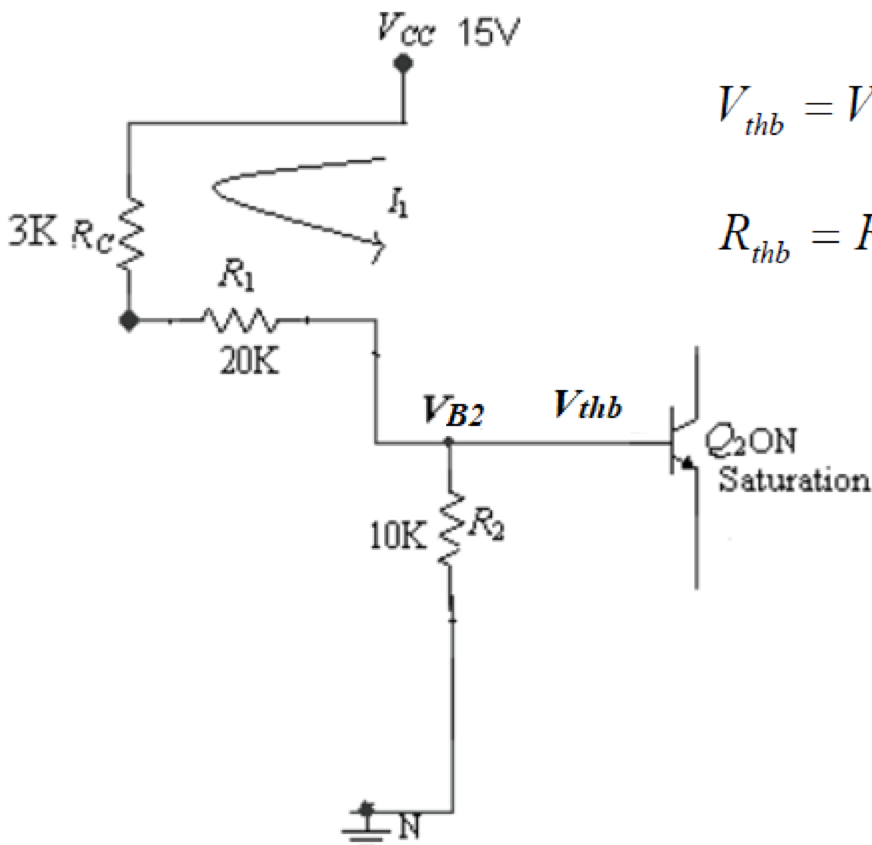



Fig.3

# Calculation

Assume the initial state as $Q_1$ OFF and $Q_2$ ON

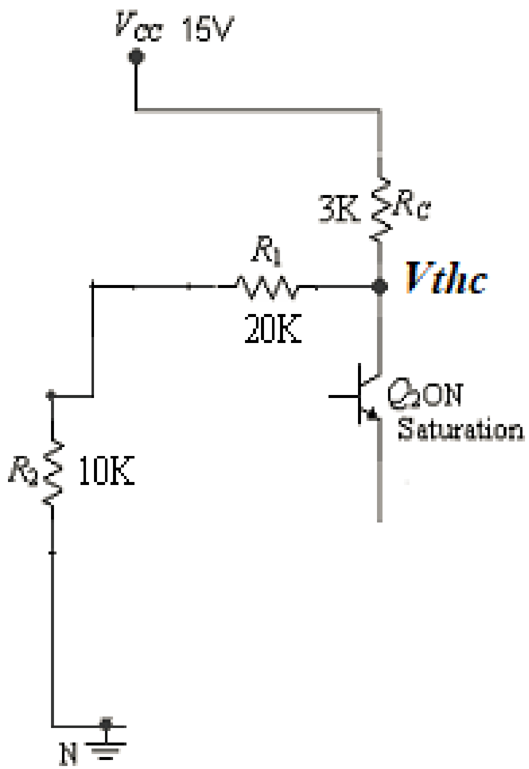



$$V_{thb} = V_{CC} \times \frac{R_2}{R_C + R_1 + R_2} = \frac{15 \times 10}{3 + 20 + 10} = \frac{150}{33} = 4.54\text{ V}$$

$$R_{thb} = R_2 \| (R_C + R_1) = \frac{10 \times (3 + 20)}{3 + 20 + 10} = \frac{230}{33} = 6.96\,\text{k}\Omega$$

Circuit to calculate $V_{\text{thb}}$ and $R_{\text{thb}}$ of $Q_2$.

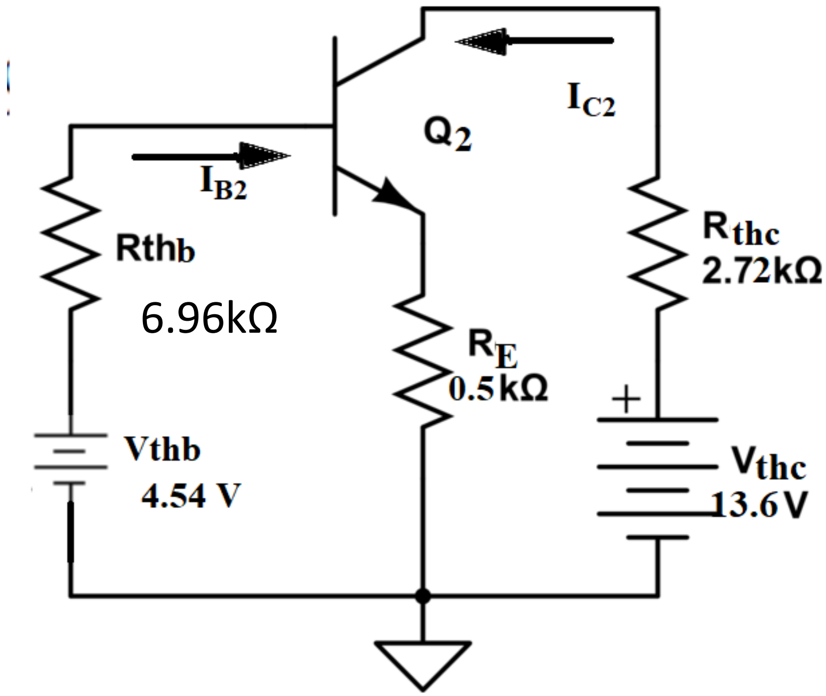
# Calculation




Circuit to calculate $V_{thc}$ and $R_{thc}$ of $Q_2$

$$V_{thc} = V_{CC} \times \frac{R_1 + R_2}{R_C + R_1 + R_2} = \frac{15 \times (20+10)}{3+20+10} = \frac{450}{33} = 13.6\ \text{V}$$

$$R_{thc} = R_C \| (R_1 + R_2) = \frac{3 \times 30}{33} = \frac{90}{33} = 2.72\ \text{k}\Omega$$




Equivalent circuit to calculate $I_{B2}$ and $I_{C2}$

Writing the KVL equations of the input and output loops

$$4.54 - 0.7 = (6.96+0.5)\, I_{B2} + 0.5 I_{C2} \quad (1)$$

$$13.6 - 0.2 = 0.5 I_{B2} + (2.72+0.5)\, I_{C2} \quad (2)$$

Solving equation (1) and (2) $I_{B2} = 0.263$ mA

$$I_{C2} = 3.75 \text{ mA}$$

$$h_{femin} = \frac{I_C}{I_B} = \frac{3.75}{0.263} = 14.25$$

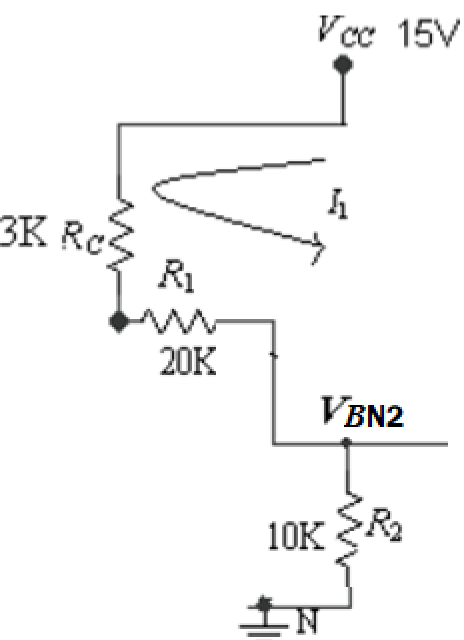
# Calculation

$$V_{EN} = (I_{B2} + I_{C2})R_E = (0.263+3.75)0.5 = 2\text{ V}$$

$$V_{CN2} = V_{EN} + V_{CE(sat)} = 2 + 0.2 = 2.2\text{ V}$$

$$V_{BN2} = V_{EN} + V_{BE2} = 2 + 0.7 = 2.7\text{ V.}$$

Since $I_{B1}=0$ $V_{BN1} = V_{CN2} \times \dfrac{R_2}{R_1+R_2} = \dfrac{2.2\times 10}{20+10} = \dfrac{22}{30} = 0.733\text{ V}$

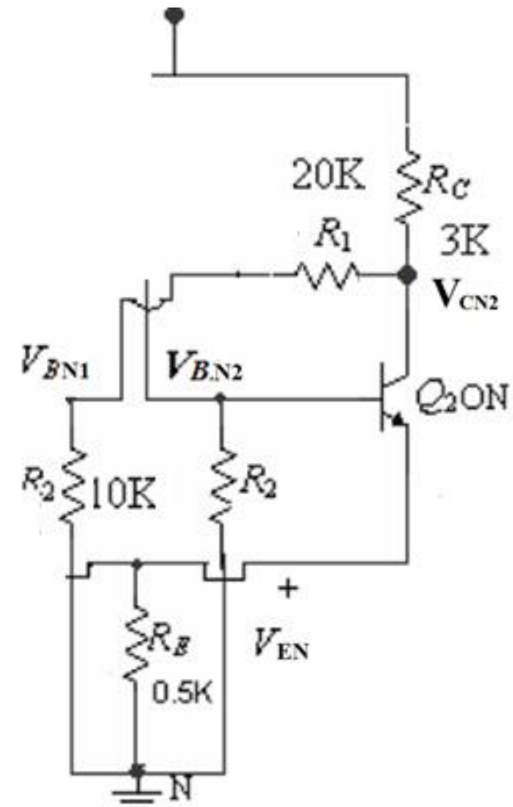



$$V_{BE1} = V_{BN1} - V_{EN} = 0.733 - 2 = -1.26\text{ V}$$

Hence $Q_1$ is OFF

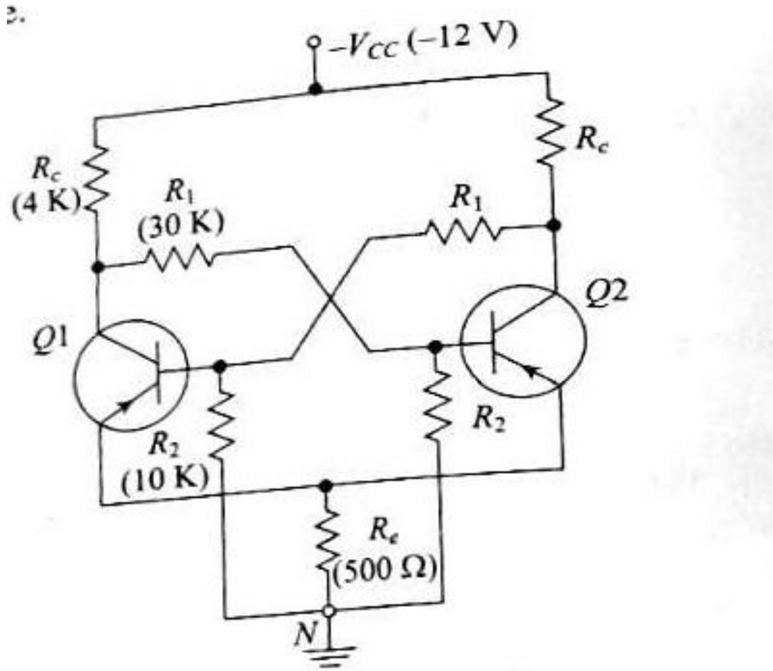
$$I_1 = \frac{V_{CC} - V_{BN2}}{R_C + R_1} = \frac{15-2.7}{3+20} = 0.534\text{ mA.}$$

$$V_{CN1} = V_{CC} - I_1 R_C = 15 - (0.534)(3) = 13.4\text{ V.}$$

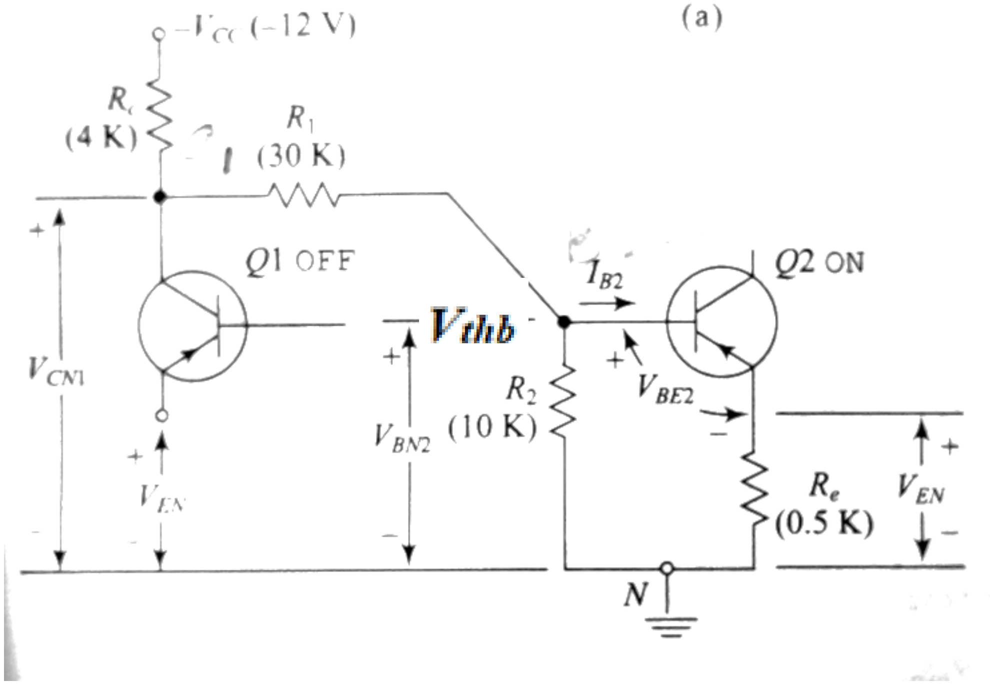



The output swing is 13.4-2.2=11.2 V

Ex. The self-bias transistor bistable multivibrator shown in Fig. 4 uses *p–n–p Ge* transistors. Given that $V_{CC}$ = -12 V, $V_{CE(sat)}$ = 0.2V, $V_{BE(ON)}$ = 0.7 V, $R_C$ = 3 kΩ, $R_1$ = 20 kΩ, $R_2$ = 10 kΩ, $R_E$ = 500 Ω. Find: minimum value of hfe and stable-state currents and voltages . Assume Q1 is in cut-off and Q2 in saturation.



(a)

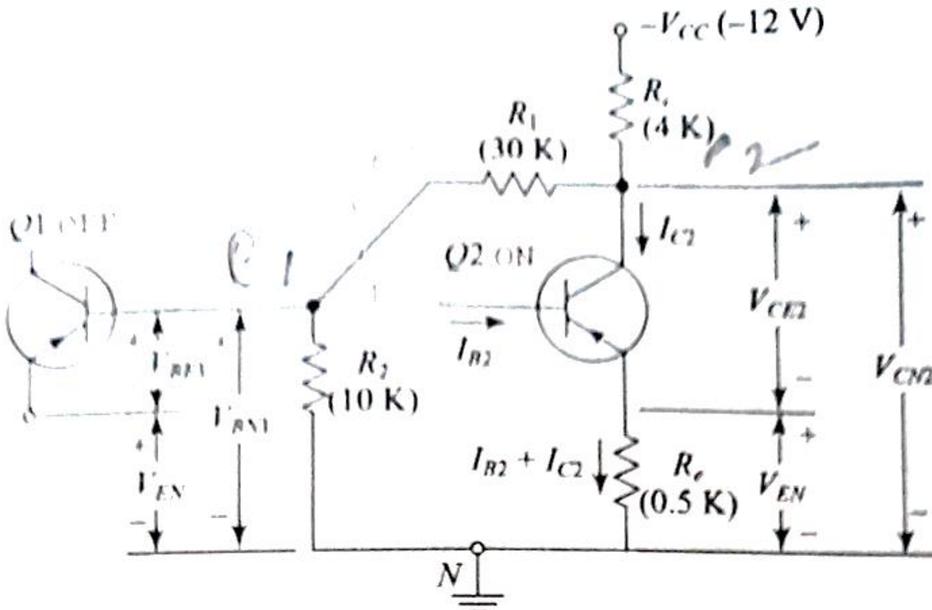



Thévenin's equivalent of the base circuit of $Q2$ :

$$\frac{-V_{CC}R_2}{R_1+R_2+R_c} = \frac{(-12)(10)}{44} = -2.73\text{ V}$$

a resistance

$$\frac{R_2(R_1+R_c)}{R_1+R_2+R_c} = \frac{(10)(34)}{44} = 7.73\text{ K}$$

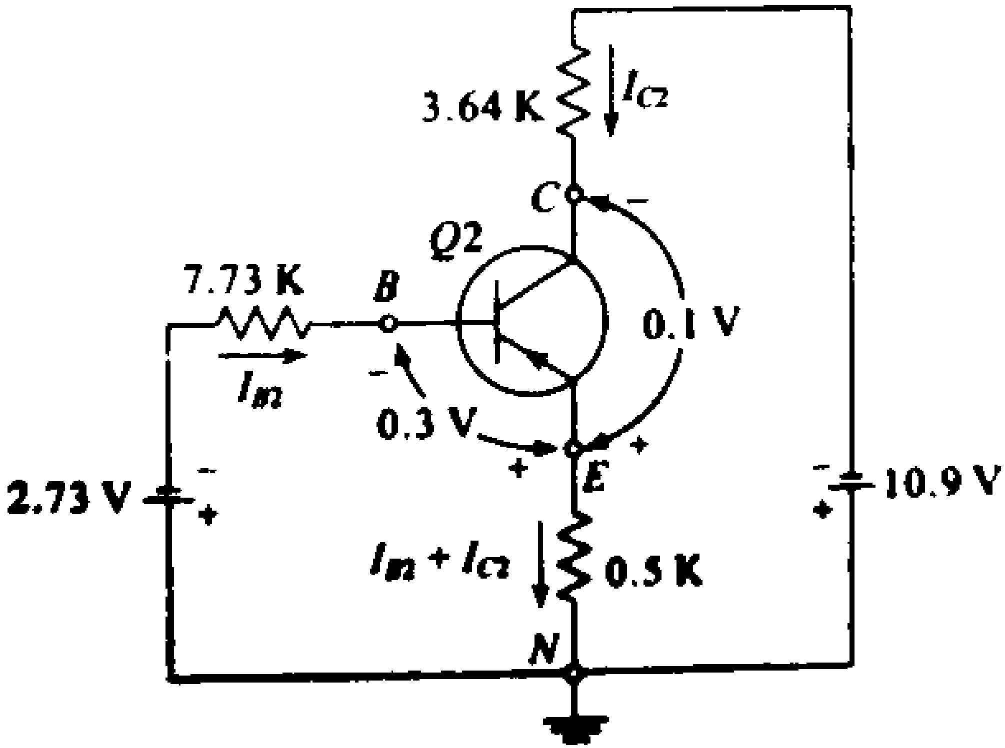


(a)

$$\frac{-V_{CC}(R_1 + R_2)}{R_1 + R_2 + R_c} = \frac{(-12)(30+10)}{30+10+4} = -10.9 \text{ V}$$

Thévenin's resistance

$$\frac{R_c(R_1 + R_2)}{R_1 + R_2 + R_c} = \frac{(4)(40)}{44} = 3.64 \text{ K}$$




$$2.63 - 0.3 + I_{B2}(7.73 + 0.5) + I_{C2}(0.5) = 0$$

$$10.9 - 0.1 + I_{B2}(0.5) + I_{C2}(3.64 + 0.5) = 0$$

ing, we find $I_{B2} = -0.138$ mA and $I_{C2} = -2.59$ mA. Hence

$$h_{FE}(\text{min}) = \frac{I_{C2}}{I_{B2}} = \frac{-2.59}{-0.138} = 18.8$$

higher hfe will maintain saturation $I_c = h_{fe} I_b$

voltages in the circuit are now found from Fig. 8.6(a) and (b)

$$V_{EN} = (I_{B2} + I_{C2})R_e = (-0.138 - 2.59)(0.5) = -1.36 \text{ V}$$

$$V_{CN2} = V_{CE2} + V_{EN} = -0.1 - 1.36 = -1.46 \text{ V}$$

$$V_{BN2} = V_{BE2} + V_{EN} = -0.3 - 1.36 = -1.66 \text{ V}$$

$$V_{BN1} = V_{CN2}\left(\frac{R_2}{R_1 + R_2}\right) = -1.46 = \left(\frac{10}{40}\right) - 0.37 \text{ V}$$

$$V_{BE1} = V_{BN1} - V_{EN} = -0.37 + 1.36 = +0.99 \text{ V}$$

ositive value of $V_{BE}$ of only about 0.1 V is required to cut-off a *p-n-p* transistor. Hence *Q*1 is certainly '.

n Fig. 8.6(b),

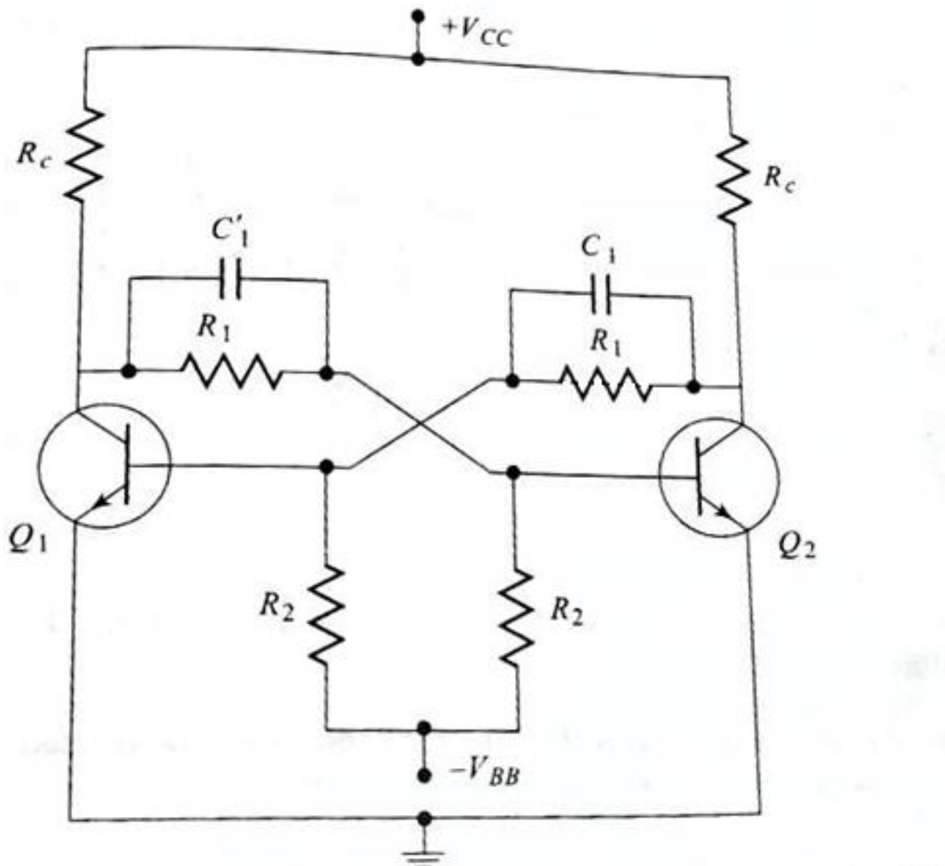
$$V_{CN1} = \frac{-V_{CC} R_1}{R_c + R_1} + \frac{V_{BN2} R_c}{R_c + R_1} = \frac{(-12)(30)}{34} + \frac{(-1.66)(4)}{34} = -10.8 \text{ V} \quad (8.14)$$

ummary, the stable state has the following values:

| | | | |
|---|---|---|---|
| $I_{C1} = 0$ mA | $I_{C2} = -2.59$ mA | $I_{B1} = 0$ mA | $I_{B2} = -0.14$ mA |
| $V_{CN1} = -10.8$ V | $V_{CN2} = -1.46$ V | $V_{BN1} = -0.37$ V | |
| $V_{BN2} = -1.66$ V | $V_{EN} = -1.36$ V | | |

: output swing is $V_w = V_{CN2} - V_{CN1} = -1.46 + 10.8 = 9.3$ V.

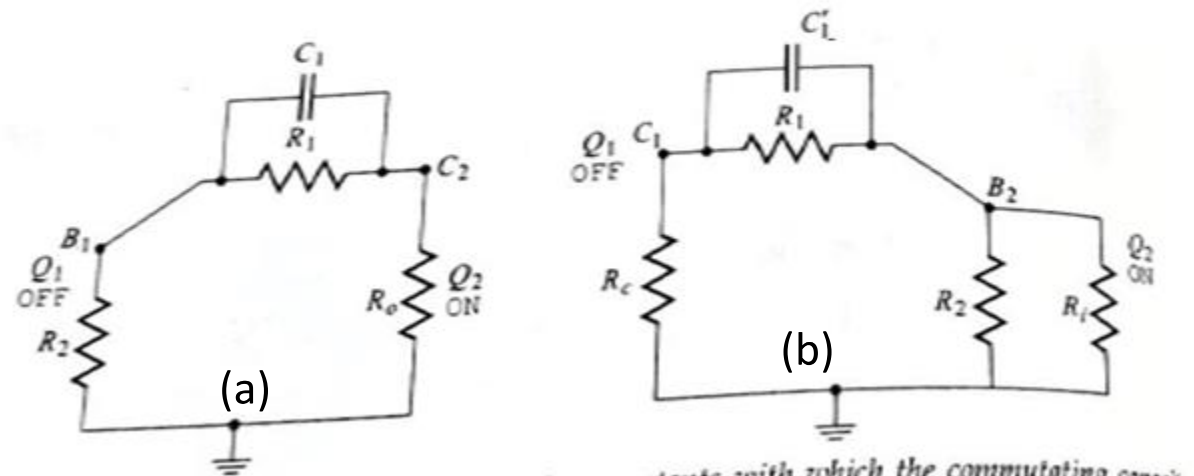
# Commutating capacitors




*A bistable multivibrator including speed-up capacitors* ($C_1' = C_1$).

A flip-flop will remain in one of its stable states indefinitely until it is induced to make a transition as the result of a "triggering" signal, such as a pulse, applied from some external source. There are many applications of flip-flops in which it is desired to have a change of state take place as soon after the application of an abrupt triggering signal as possible. The *transition time* is defined as the interval during which conduction transfers from one transistor (or tube) to the other. The transition time may be reduced by introducing small capacitances in shunt with the coupling resistors $R_1$ of the binary. A flip-flop with such capacitors included is shown in Fig. 10-10. Because these capacitors assist the binary in making abrupt transitions between states, they are known as *commutating*, *transpose*, or *speed-up capacitors*. The usefulness of these capacitors will be seen in the following discussion. To be specific let us assume that the active devices are tubes or *n-p-n* transistors.

$C_1 = \frac{R_2 C_i}{R_1}$




*Equivalent circuits for computing the time constants with which the commutating capacitors recharge.*

- ✓ Time constant in (a) is $\tau = C_1(R_1 || (R_2 + R_0))$. $R_0$ is the output resistance of transistor in saturation which is very small $\tau = C_1(R_1 || R_2)$.
- ✓ Time constant in (b) is $\tau' = C_1'(R1 || (R_c + (R_2 || Ri))) = C_1'(R_1 || (R_c + Ri))$. $R_i$ is the input resistance of transistor in saturation which is very small , $R_i << R_2$

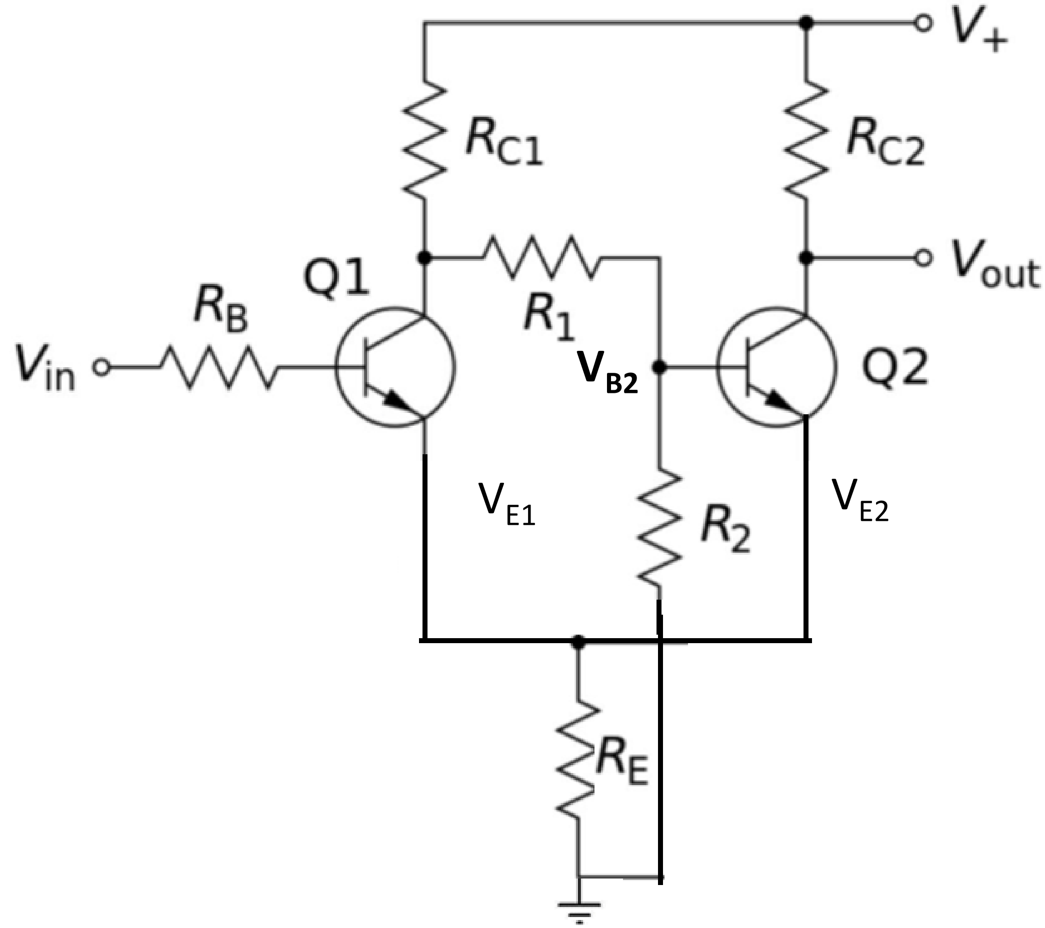
$\tau' = C_1'(R_1 || Rc) << \tau$ since $Rc < R_1$ and $R_2$

- ✓ Thus transition time is $\tau$.
- ✓ If Ci is the input capacitance of Q1, for perfect compensation $C_1$ should be equal to $R_2 Ci / R_1$

# Resolving time

- The smallest allowable interval between triggers is called the resolving time of bistable MV, and it is reciprocal of maximum frequency at which the MV will respond.
- It seems reasonable to assume a time $2\tau$ between triggers so that all the transients will died down reasonably.
- $f_{max}=1/2\tau$

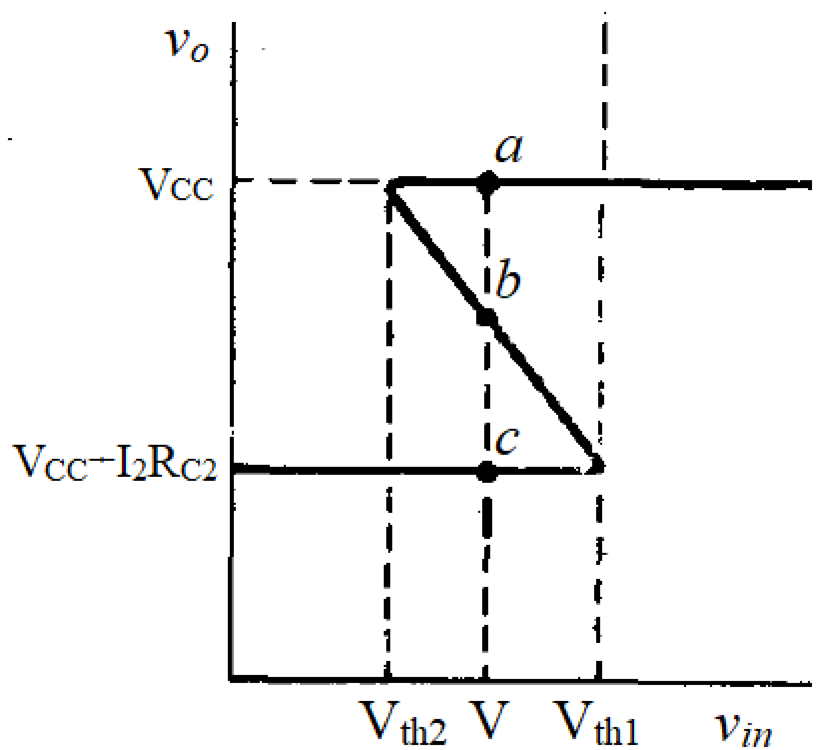
# Schmitt Trigger Circuit

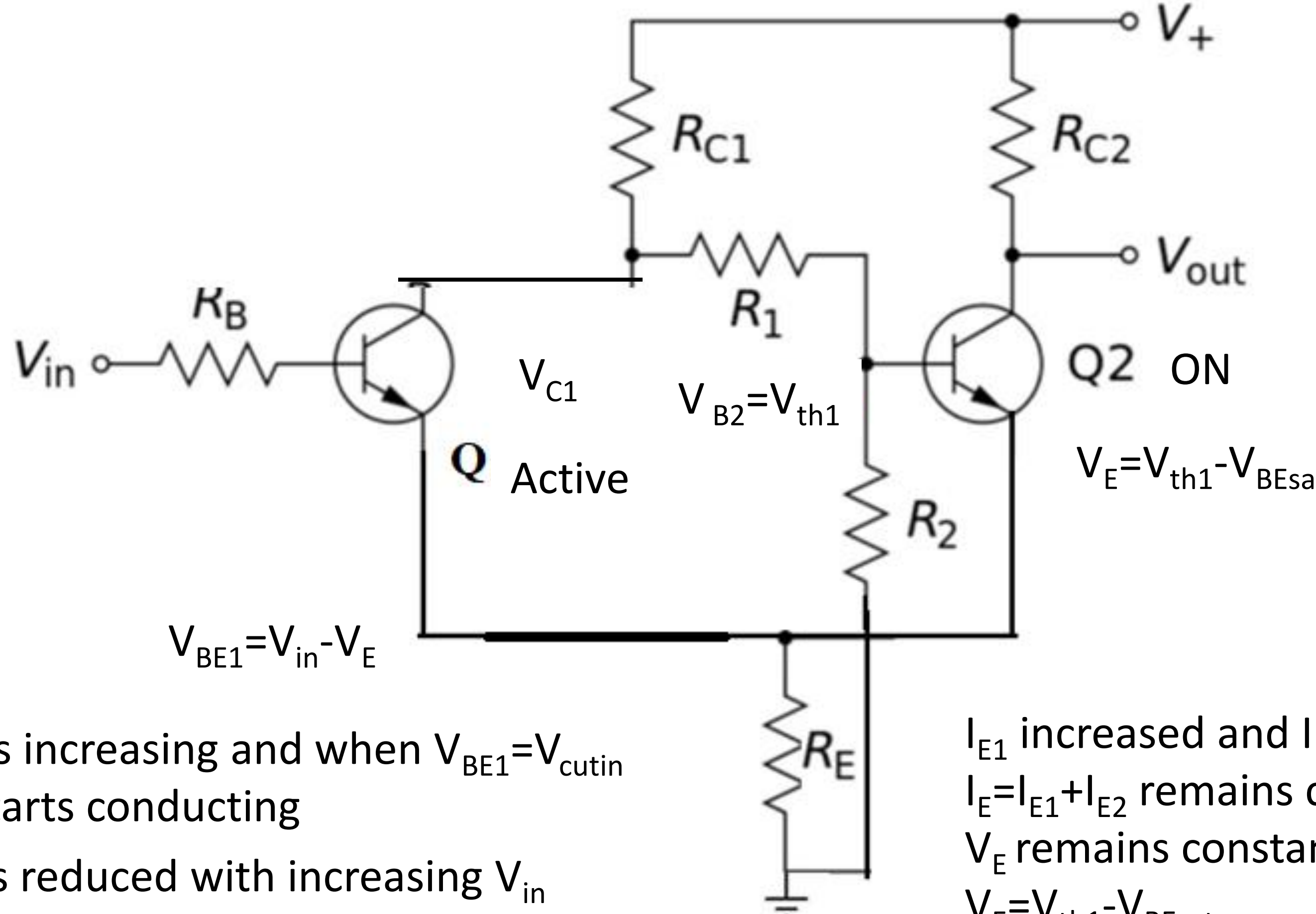


## APPLICATIONS of SCHMITT TRIGGER

- Used in signal conditioning applications to remove noise from signals in digital circuits.
- Used as squaring circuit.
- Used as amplitude comparator or level detector.
- Used as flip-flop circuit.
- Used for reshaping the worn out pulses.
- Used to implement relaxation oscillator in function generators and switching power supplies.

The existence of only two stable states results from the fact that positive feedback is incorporated into the circuit and the loop gain of the circuit is greater than unity.






Vin is increasing and when $V_{BE1}=V_{cutin}$
$Q_1$ starts conducting

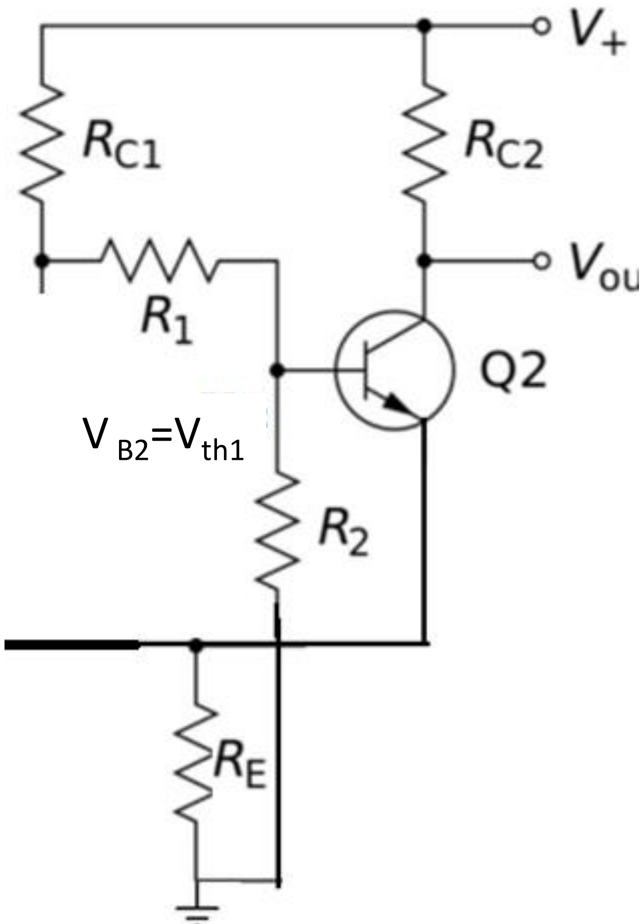
$V_{C1}$ is reduced with increasing $V_{in}$

$V_{B2}$ reduced, current through $Q_2$ is reduced.

$I_{E1}$ increased and $I_{E2}$ reduced
$I_E=I_{E1}+I_{E2}$ remains constant
$V_E$ remains constant
$V_E=V_{th1}-V_{BEsat}$

$Q_1$ will enter into saturation when $V_{BE1}=V_{BEsat}=V_{in}-(V_{th1}-V_{BEsat})$
➔ $V_{in}=V_{th1}$ ➔ transition takes place.

# Calculation of $V_{th1}$




$$V_{th1} = \frac{R_2'}{R_2' + R_{C1} + R_1} V_+$$

$$R_2' = R_2 \,||(1+\beta)R_E$$
$$\cong R_2$$
$$(1+\beta)R_E \gg R_2$$

# Working principle

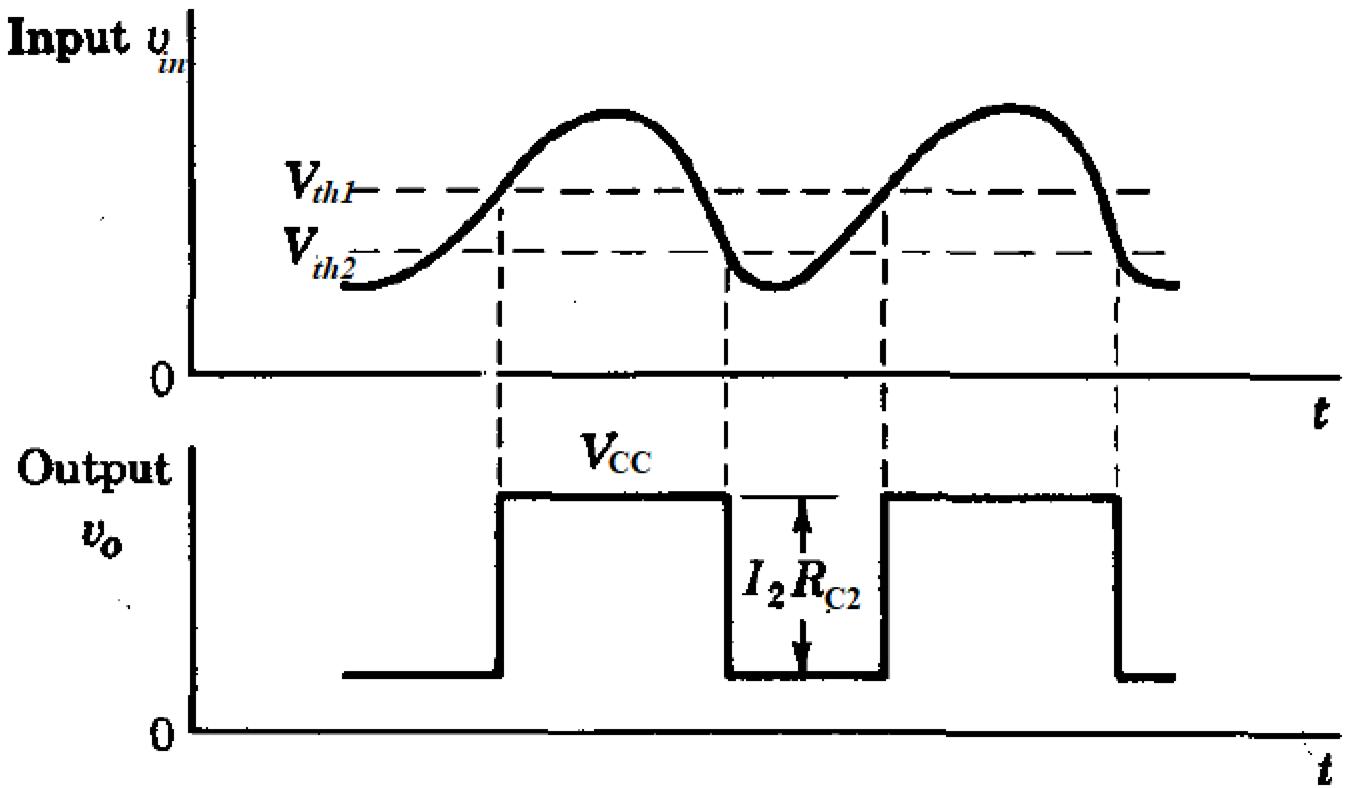
1. Assume $V_{in}$=0V, $Q_1$→ OFF
   i. $V_{B2}= V_{th1}=(V_+ \times R_2)/(R_{C1} + R_1 + R_2)$ → $Q_2$ is in saturation
   ii. $V_{E2}= V_{th1}- V_{BE2sat}= V_{E1}$
   iii. $V_{BE1}= V_{B1}- V_{E1}= V_{in}-(V_{th1}- V_{BE2sat})=V_{in}-V_{th1}+ V_{BEsat}$ → $v_{in}= V_{th1}-V_{BEsat} + V_{BE1}$
   iv. Assume that the two transistors are identical, then as long as the input voltage remains less than the base voltage of $Q_2$, $Q_1$ will remain off and the circuit operation will not change. (Initial state $V_{out}=V_{sat}$)

2. Now, suppose $V_{in}$ rise continuously and it approaches voltage at the base of $Q_2$.
   i. $Q_1$ begins to conduct.
   ii. current through $R_{C1}$ increases →the voltage at the collector of $Q_1$, $V_{C1}$ decreases
   iii. Base voltage on $Q_2$, decreases→ $I_{C2}$ decreases→ $v_{out}= V_{C2}$ increases
   iv. Current flowing through $R_E$ is $I_{E1+} I_{E2}$ . As $I_{E1}$ increase, $I_{E2}$ decreases, and the voltage across $R_E$ doesn't change rapidly.
   v. $V_{BE2}$ decreases and $Q_2$ turns off and the output voltage rises to $V_+$ volts. The circuit has just changed states.

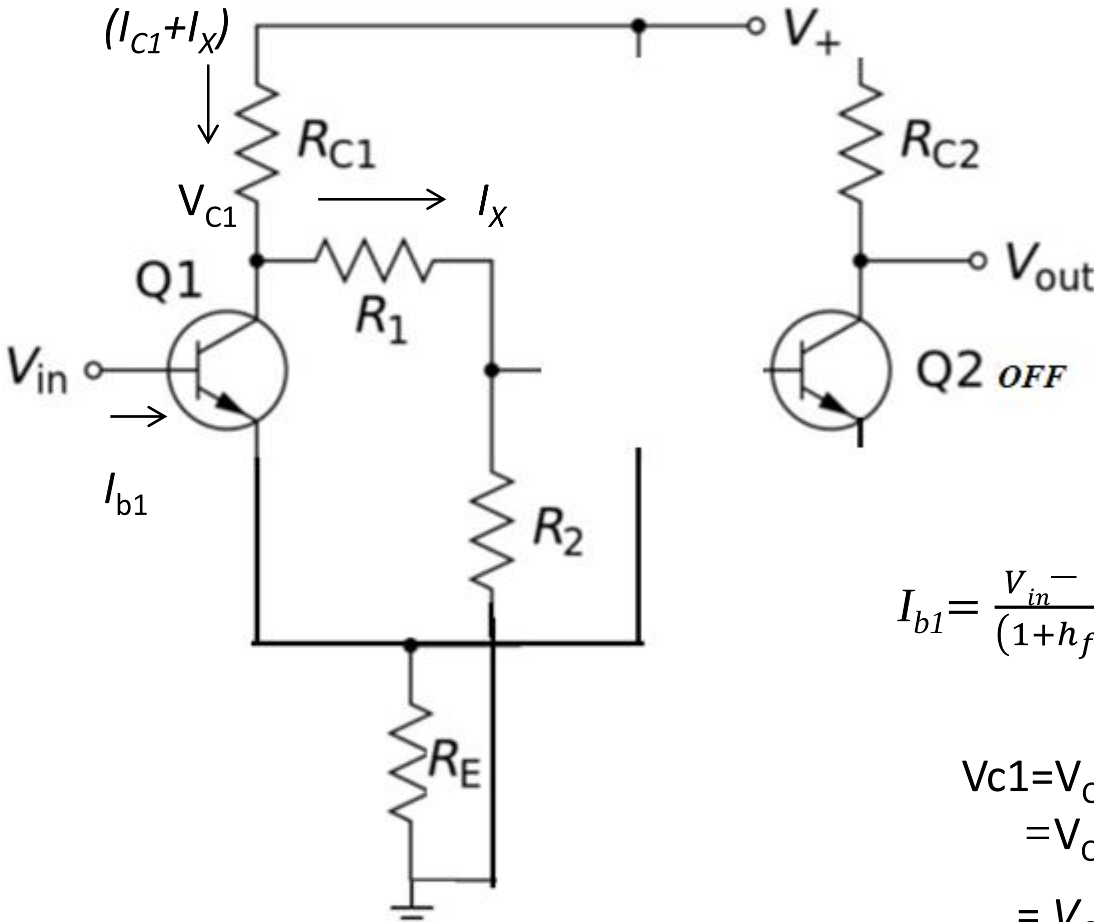
3. If the $V_{in}$ rises further, it will simply keep $Q_1$ turned on and $Q_2$ turned off.

4. However, if the input voltage starts to fall back towards zero
   i. there must clearly be a point at which this circuit will reset itself. The question is, What is the falling threshold voltage? It will be the voltage at which $Q_1$'s base becomes more negative than $Q_2$'s base, so that $Q_2$ will begin conducting again.
   ii. Now, $V_{B2}= (V_{C1}\times R_2)/(R_1 + R_2)$. As $V_{in}$ reduces, $V_{C1}$ increases, $V_{B2}$ increases. As soon as $V_{B2}= V_{E2}+V_{cutin2}=V_{in}-V_{BE1}+V_{cutin2}$, means $V_{in}=V_{B2}+(V_{BE1}-V_{cutin2})=V_{th2}$, $Q_2$ turns ON, $I_{C2}$ increases, $V_{C2}$ decreases.
   iii. $V_{BE1}-V_{cutin2}=\sim 0.1V$
   iv. As $V_{in}$ falls below $V_{th2}$, $Q_1$ turns off.
   v. Thus 1st change of state at $V_{in}=V_{th1}=(V_+\times R_2)/(R_{C1} + R_1 + R_2)$ and 2nd change of state takes place at $V_{in}= V_{th2}$

# Waveform



# Calculation of $V_{th2}$




$$I_{b1} = \frac{V_{in} - V_{be1}}{(1+h_{fe})R_E} \text{ and } I_x = \frac{V_{c1}}{R_1+R_2}$$

$$V_{c1} = V_{CC} - (I_{c1}+I_x)R_{c1}$$

$$= V_{CC} - (h_{fe} I_{b1} + I_x) R_{c1}$$

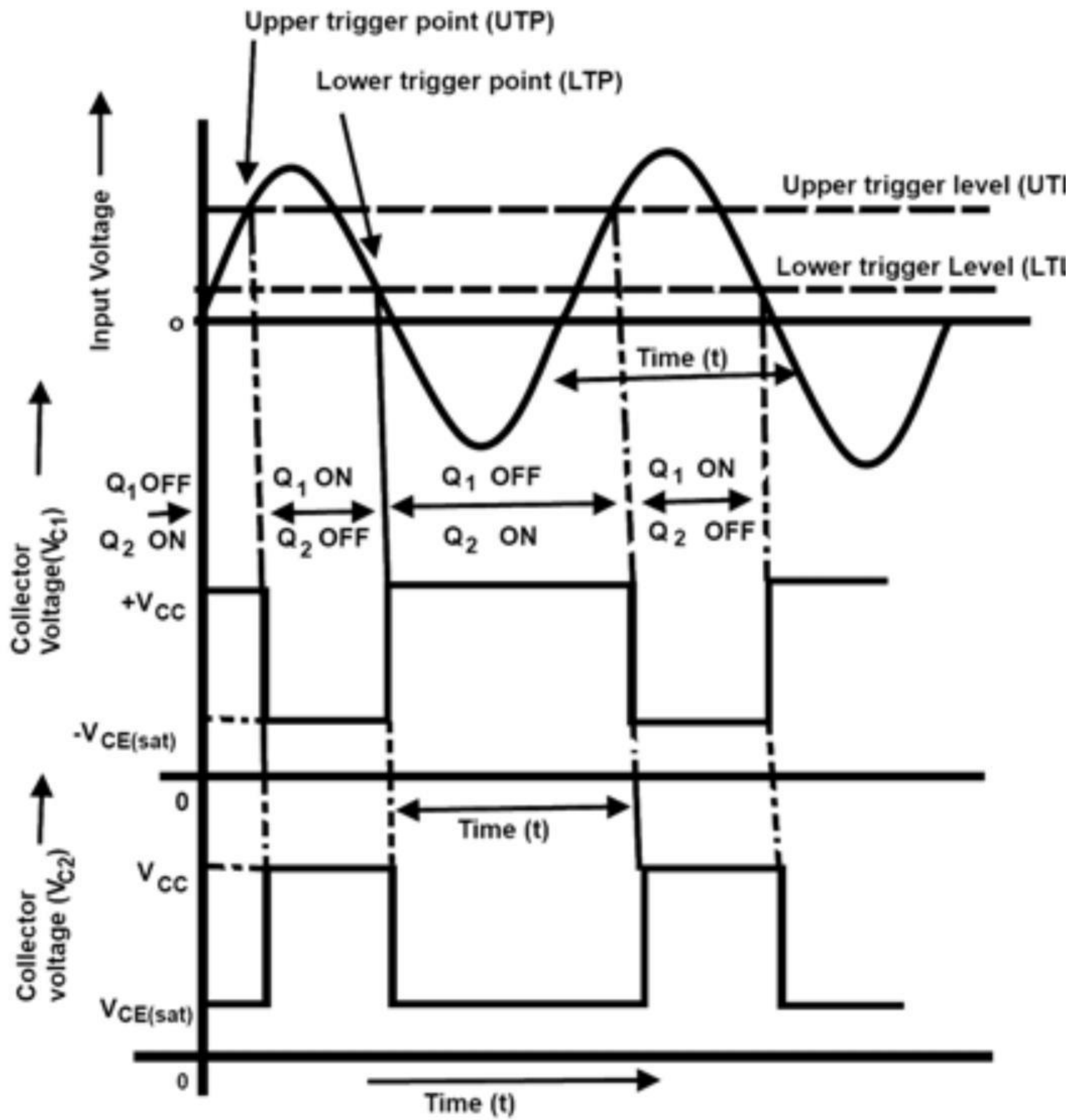
$$= V_{CC} - \left(h_{fe} \frac{V_{in} - V_{be1}}{(1+h_{fe})R_E} + \frac{V_{c1}}{R_1+R_2}\right) R_{c1}$$

$VB2 = V_{c1} \times R2/(R1+R2)$

- When VB2-VE2=VBE2cutin, Q2 turns ON
- VE2=Vin-VBE1sat
- VB2=Vin+VBE2cuitin-VBE1sat
- Vin≅VB2

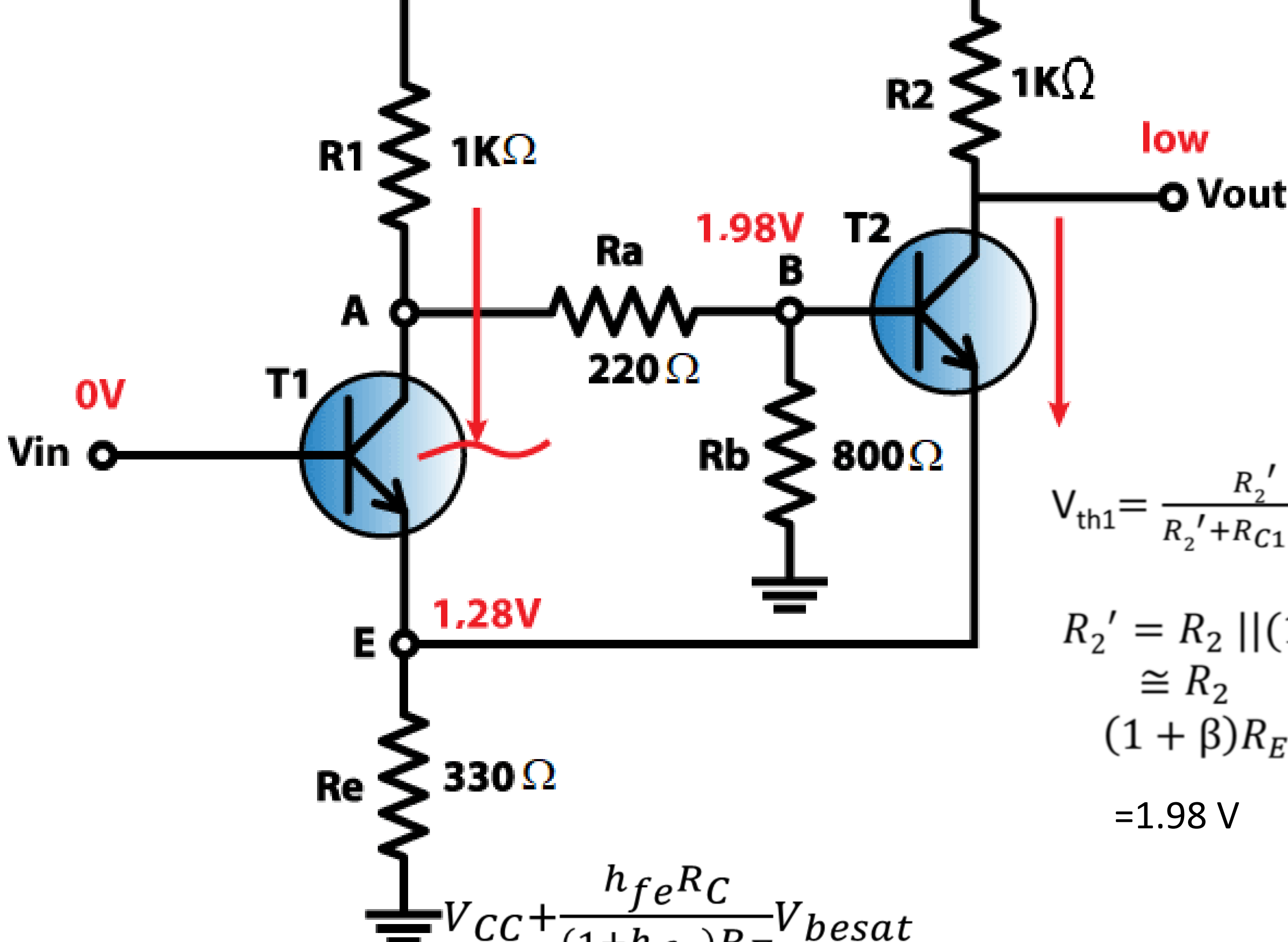
$V_{B2}=V_{c1}xR_2/(R_1+R_2)=V_{th2}$

$$V_{th2}=\frac{V_{CC}+\frac{h_{fe}R_C}{(1+h_{fe})R_E}V_{besat}}{\left(1+\frac{R_1}{R_2}\right)+\left(1+\frac{R_{C1}}{R_1+R_2}\right)+\left(\frac{h_{fe}R_C}{(1+h_{fe})R_E}\right)}$$



Upper trigger point (UTP)
Lower trigger point (LTP)
Upper trigger level (UTL)
Lower trigger Level (LTL)
Input Voltage
o
Time (t)
Q1 OFF
Q2 ON
Q1 ON
Q2 OFF
Q1 OFF
Q2 ON
Q1 ON
Q2 OFF
Collector Voltage(VC1)
+VCC
-VCE(sat)
0
Time (t)
Collector voltage (VC2)
VCC
VCE(sat)
0
Time (t)

- The Schmitt Trigger is a logic input type that provides hysteresis or two different threshold voltage levels for rising and falling edge. This is useful because it can avoid the errors when we have noisy input signals from which we want to get square wave signals.




$$V_{th1} = \frac{R_2'}{R_2' + R_{C1} + R_1} V \quad +$$

$$R_2' = R_2 \,||\, (1+\beta)R_E$$
$$\cong R_2$$
$$(1+\beta)R_E \gg R_2$$

=1.98 V

$$V_{th2} = \frac{V_{CC} + \frac{h_{fe}R_C}{(1+h_{fe})R_E} V_{besat}}{\left(1+\frac{R_1}{R_2}\right) + \left(1+\frac{R_{C1}}{R_1+R_2}\right) + \left(\frac{h_{fe}R_C}{(1+h_{fe})R_E}\right)}$$

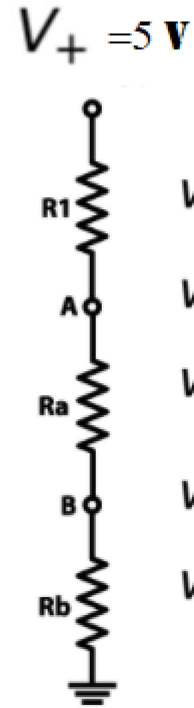
=1.28 V

- We will start like this. Let's suppose that the Vin input is 0 V. That means that transistor T1 is cut off and not conducting. On the other hand the Transistor $T_2$ is conducting because we have a voltage of about 1.98 V at the B node as we can consider this part of the circuit as a voltage divider and calculate the voltage using this expressions.

- So because the Transistor $T_2$ is conducting the output voltage will be low and the voltage at the emitter will be VE=VB-VBE= (1.98-0.7)=1.28 V.

- The emitter of the transistor $T_1$ is connected with the emitter of the transistor $T_2$ so they are at the same voltage level of 1.28 V which means that the transistor T1 will turn on when the voltage Vin at its base will be 0.7 V above this value of 1.28 V, or about 1.98 V.

- So as we increase the $V_{in}$ input and we cross this value of 1.98 the transistor $T_1$ will start conducting. This will cause the voltage at the base of the transistor $T_2$ to drop and will cut the transistor off. As the transistor $T_2$ is no longer conducting the output voltage will go high.




$$Vin = 0\,V;$$

$$Va = \frac{Ra + Rb}{Ra + Rb + R1} V_+$$

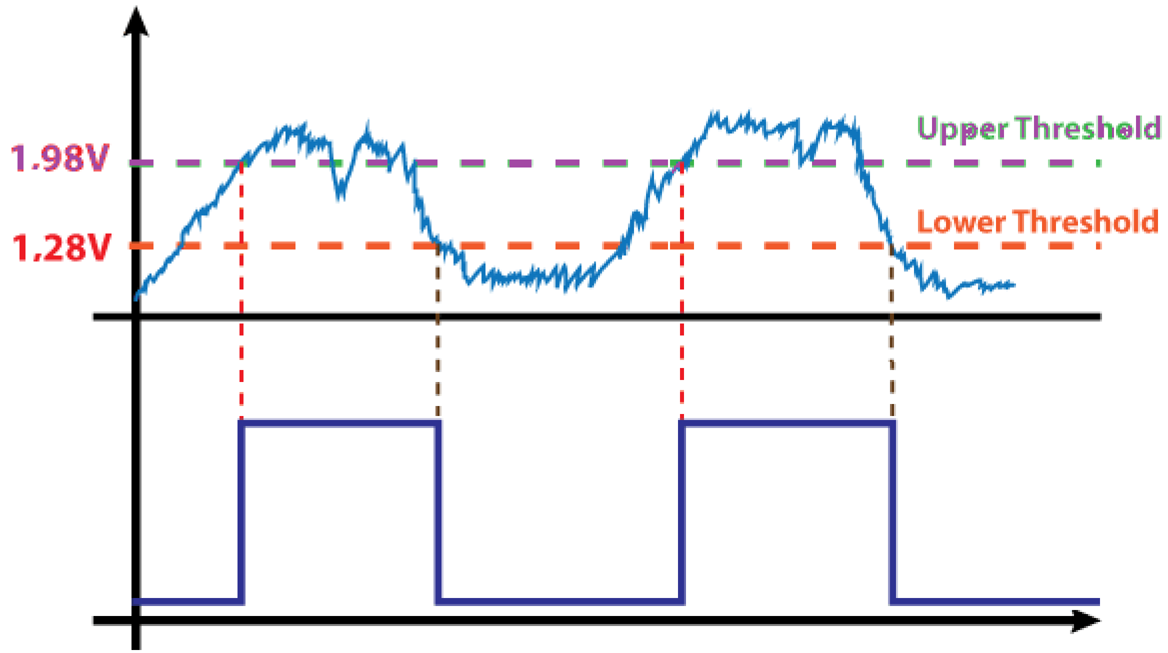
$$Va = 2.52\,V$$

$$Vb = \frac{Rb}{Rb + R1 + Ra} V_+$$

$$Vb = 1.98\,V$$

- Next, the voltage $V_{in}$ at the base of the transistor $T_1$ will start declining and it will turn the transistor off when the base voltage will be 0.7 V above the voltage of its emitter. This will happen as the current in the emitter will decline to a point where the transistor will get into forward-active mode. In this mode the collector voltage will increase, which will also increase the voltage at the base of the transistor $T_2$. This will cause small amount of current to flow through the transistor $T_2$ which will further drop the voltage at the emitters and will cause the transistor $T_1$ to turn off. In our case the $V_{in}$ input needs to drop to about 1.3 V to turn off the transistor $T_1$.

- That's it. Now the cycle repeats over and over again. So we got two thresholds, the high threshold at about 1.98 V and the low threshold at about 1.28 V.



# Monostable MV

- Multivibrator is a non linear oscillator or function generator which can generate square, rectangular and pulse waves.
- **Monostable Multivibrator** or **One Shot Multivibrator** has only one stable state. By default monostable multivibrator will be in its stable state, but when triggered it will switch to unstable state (quasi-stable state) for a time period determined by the RC time constant in the circuit.

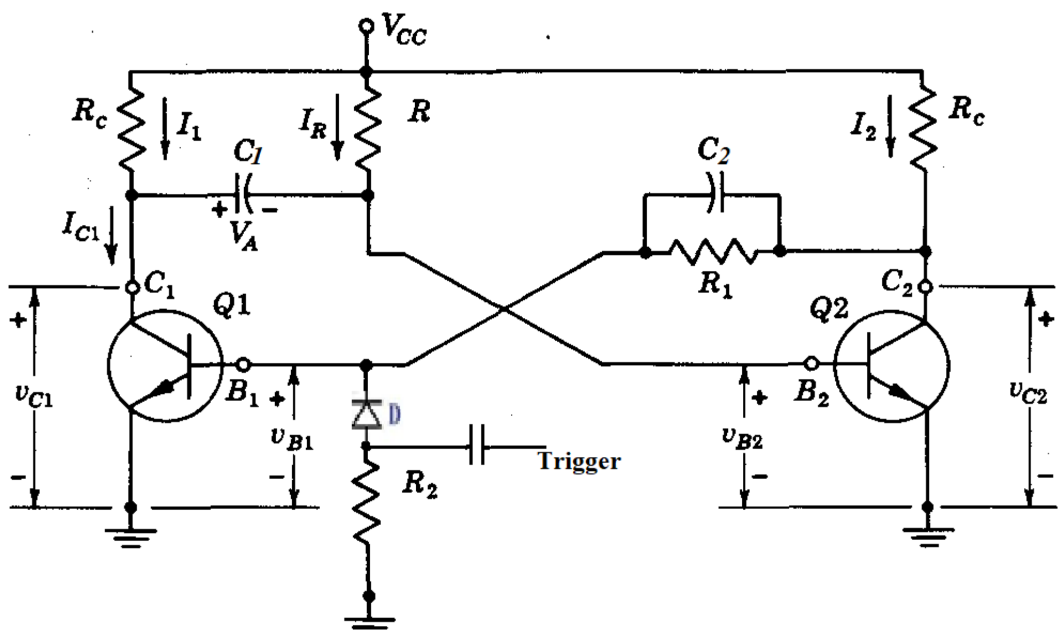



Fig: M1

# Explanation of the circuit

- In the above circuit diagram we can find two transistors which are connected as switches. When a transistor is ON, it works in saturation region and its collector – emitter acts as a short circuit. When a transistor is OFF, it works in cut off region and its collector – emitter acts as an open circuit. So in the above circuit, when a transistor is in OFF state, its collector will have voltage equal to supply voltage Vcc and when the a transistor is in ON state, its collector will be grounded (ideally). Practically $V_{CEsat} > 0 \sim 0.2$ V. $V_{CEcut\text{-}off} < V_{CC}$ due non-zero collector current (reverse saturation current).

- Function of resistor $R_C$ is to limit collector current of both transistors $Q_1$ and $Q_2$. Resistors R & $R_1$ will provide base current for transistors $Q_2$ & $Q_1$ respectively during ON condition.

- A Capacitor and Resistor is designed as a differentiator circuit to provide sharp trigger pulses to the base of the transistor $Q_1$. The diode D allows only positive pulses to the base.

- Capacitor $C_2$ is optional, which is called as speed-up capacitor. It is used for speedy bypassing of signal transitions (LOW to HIGH and HIGH to LOW) at the collector of $Q_2$ to the base of $Q_1$.

# Operation

- When the circuit is switched ON, the circuit parameters are so chosen that transistor $Q_1$ will be OFF and $Q_2$ will be ON.
- Capacitor $C_1$ gets charged during this state.
- When a positive trigger is applied to the base of transistor $Q_1$ it turns ON, which turns OFF the transistor $Q_2$ due the negative voltage from the capacitor $C_1$.(Capacitor can not discharge instantaneously)
- Capacitor $C_1$ starts discharging during this state with td=RC1.
- Transistor $Q_1$ remains in ON state due the positive voltage from the collector of transistor $Q_2$ which is in OFF state.
- Transistor $Q_2$ remains in OFF state until the capacitor $C_1$ discharges completely.

# Design

- $R_C$ **– Collector Resistor** should be calculated depending upon the collector current requirement.

  **Rc = ( Vcc – Vce (sat) ) / Ic**
- **R – Base Resistor** should be chosen such that it will provide enough collector current during saturation to the transistor $Q_2$.
- Min. base current required, $Ib_{min} = Ic / \beta$. Safe base current, $Ib = 3\ Ib_{min} = 3Ic / \beta$
- **R = ( Vcc – Vbe ) / Ib**
- $R_1$ **– Base Resistor Q1** should be chosen such that it should provide enough saturation collector current to the transistor $Q_1$.
- $R_1 = ((Vcc - Vbe) / Ib) - R_C$
- **T – Pulse Time Period,** $T = 0.693RC_1$
- From this we can find the value of capacitor $C_1$ for a given pulse-width, T.
- **Speed-Up Capacitor** is designed by considering a Compensated Attenuator composed of Base Emitter resistance of $Q_1$, Resistor $R_1$, Base Emitter capacitance of $Q_1$ and seed up capacitor C2. $RC2 = r_\pi Ceb$
- Base Emitter resistance of a transistor, $r_\pi = V_T / Ib$, where $V_T$ is the thermal voltage which is approximately equal to 25mV at room temperature and Ib is the base current.
- Base Emitter capacitance , Ceb or Input capacitance will be specified in the datasheet of the transistor. For example, the Input capacitance of BC547 transistor is 9pF.
- **Note 1:** It may be very difficult to get the proper matching speed up capacitor but you can still use a near available one to improve the performance.

  **Note 2:** Actually the speed up capacitor should be sized to remove the charge stored at the base of the transistor during saturation period. Unfortunately the value of $r_\pi$ and Ceb are not constant. It varies depending up on many factors. So the optimum capacitor value can be find out only through a series of experiments.
- **– Differentiator** should be designed depending up on the frequency of the trigger pulse.

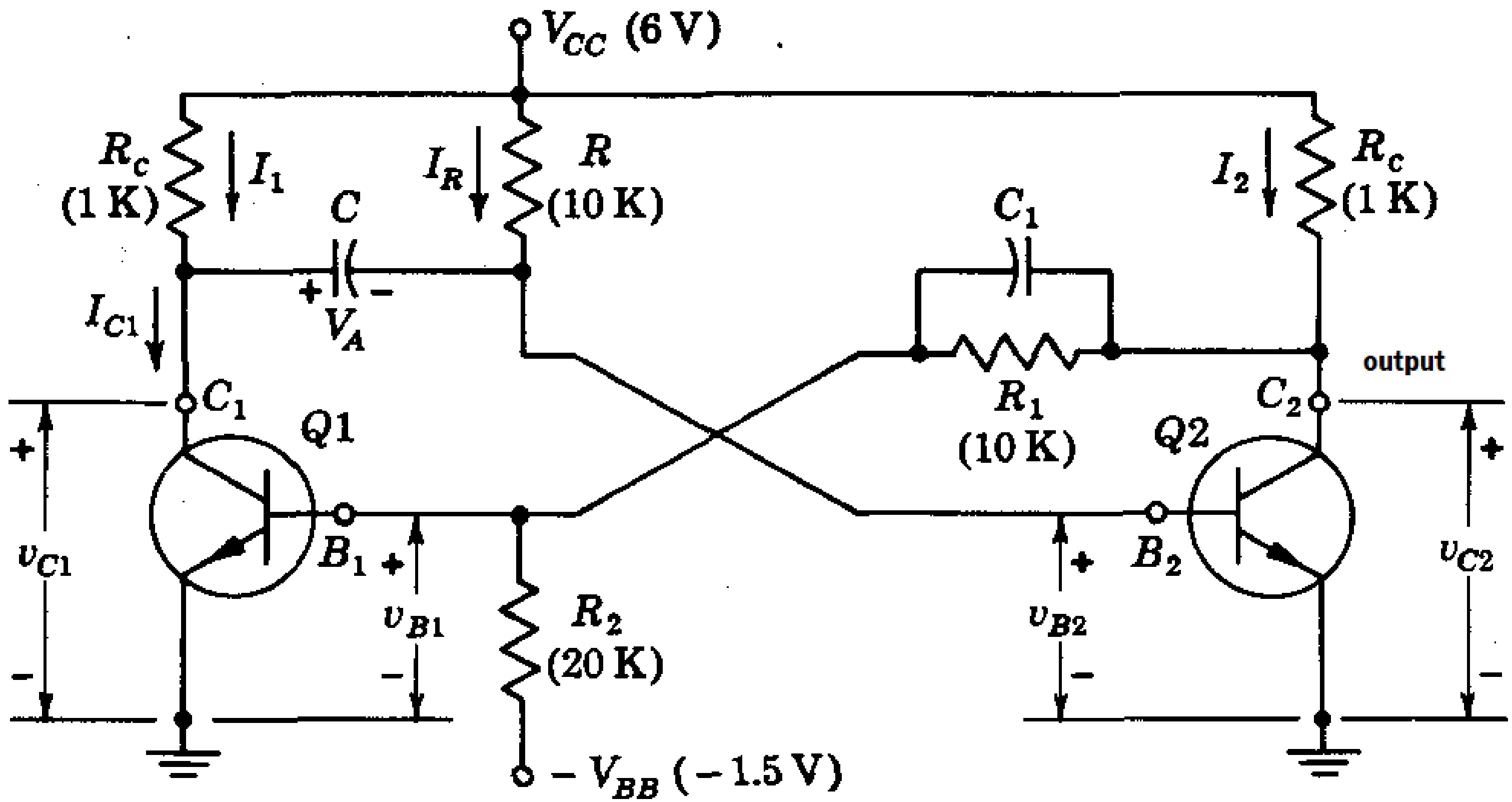



Fig: M3 The collector-coupled *n-p-n* transistor monostable multivibrator

The multi may be induced to make a transition out of its stable state by an application of a positive trigger at B1 or *C2* or by the application of a negative trigger at *B2* or *C1*

A negative trigger will be given to the base of $Q_2$ to move the state to quasi stable state

- Assume a negative trigger will be given to the base of $Q_2$ so that regenerative action takes place driving *Q2* completely below cutoff.
- The voltage at *C2* is now approximately *Vcc* and $I_1$ current is flowing through Rc and the voltage at C1 abruptly decreases by an amount of $I_1R_c$.
- Since voltage across capacitor can not be changed abruptly, the voltage at B2 drops by the same amount as that at C1. Q2→off
- The MV is now in quasi-stable state. C1 starts discharging.
- $B_2$ will rise in voltage, and when passes the cutoff voltage of $Q_2$, a regenerative action will take place turning $Q_1$ off and eventually returning the multi to its initial stable state.
- We look now into the matter of determining the time duration of the quasi-stable state.

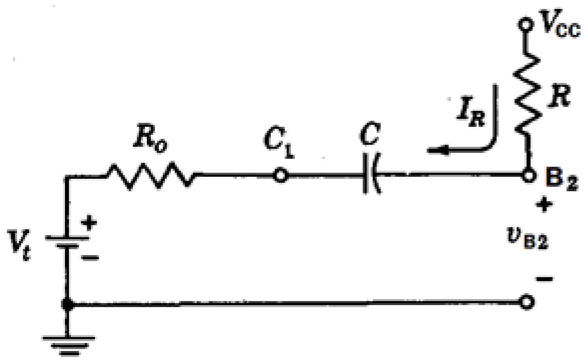



Fig: M4

**Simplified circuit for computing the voltage $v_{B2}$ at the input to Q2, during the quasi-stable state. The Thévenin's voltage $V_t$ is the voltage at $C_1$ if the capacitor $C$ is disconnected from $C_1$. $R_o$ is the output resistance of amplifier induding Rc during saturation. $R_o << R_c$**

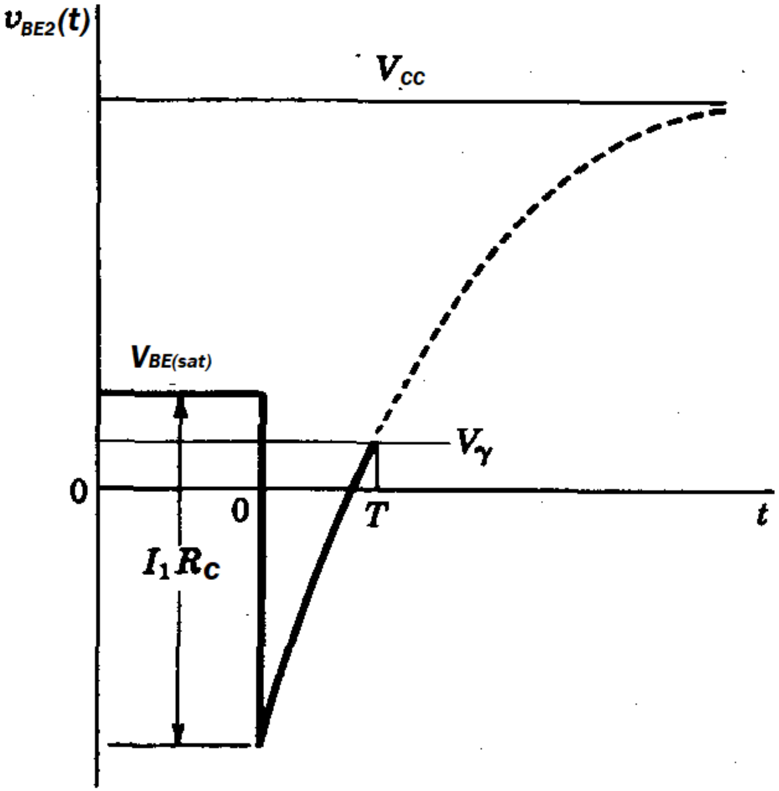



**Fig. M5 Voltage variation of $B_2$ during the quasi-stable state.**

$V\gamma$ is the cut-in voltage of BE diode $V_{BE(sat)}$ at saturation

$V\gamma$=0.5 V for Si and 0.1 V for Ge . $V_{BE(sat)}$ =0.7 V for Si and 0.3 V for Ge .

- During the quasi-stable state (0<t<T) *Q2* is *off* and the voltage changes at *B2* may then be computed from the circuit of Fig. *M4* in which Q1 has been replaced by a Thevenin equivalent voltage *Vt* and a resistance *Ro,* the output impedance, being given by the parallel combination of *Rc* and the dynamic output resistance of Q1, *ro.* The voltage variation at *B2* during the quasi-stable state is shown in Fig. M5.
- At t=0+, the voltage at $C_1$ drops by $I_1Rc$.
- Since B2 is capacitively coupled to C1, $v_{BE2}$ will be changed by $I_1Rc$
- Thus at t=0+, $v_{BE2}= v_{BE(sat)}- I_1Rc$
- Thereafter , $v_{BE2}$ will rise exponentially toward Vcc with a time constant $\tau=(R+Ro)C\cong RC$

- Since at $t=\propto$, $v_{BE2}=Vcc, v_{BE2}(t)=[Vcc-\{Vcc-(v_{BE(sat)}- I_1Rc)\}]e^{(-t/\tau)}$
- As soon as $v_{BE2}(t)$ rises to $V\gamma$, at t=T, Q2 turns on making Q1 off
- MV returns to its initial stable state
- $T=\tau \ln[\frac{(Vcc+ I_1Rc-v_{BE(sat)})}{(Vcc- V\gamma)}=\tau \ln[\frac{(Vcc+(Vcc-v_{CE(sat)})- v_{BE(sat)})}{(Vcc- V\gamma)}]=\tau \ln[\frac{(2Vcc-v_{CE(sat)}- v_{BE(sat)})}{(Vcc- V\gamma)}]$

$=\tau \ln 2+\tau \ln[\frac{(Vcc-(\frac{v_{CE(sat)}-v_{BE(sat)}}{2}))}{(Vcc- V\gamma)}]\cong\tau \ln 2\cong 0.69RC$. $V_{CC}$should be much more greater than junction voltages

# Collector coupled MV

**EXAMPLE** Compute the voltage levels for the waveforms of Fig. M6 for a collector-coupled multi whose components and supply voltages are as given in Fig. M7 . Silicon transistors are used with $r_{bb'} = 200\ \Omega$ and $h_{FE} = 30$.

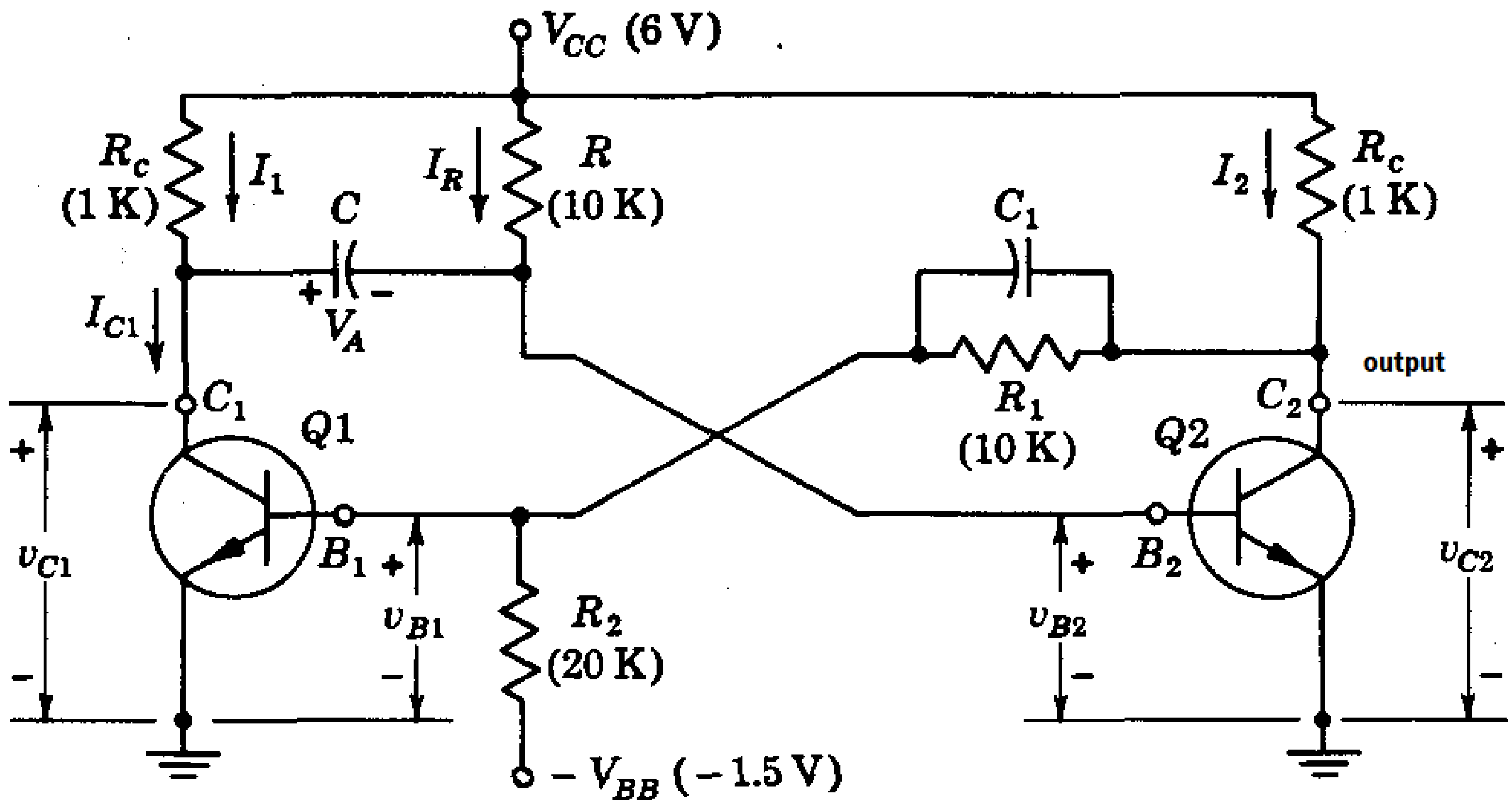



Fig. M6 The collector-coupled *n-p-n* transistor monostable multivibrator

# Waveforms

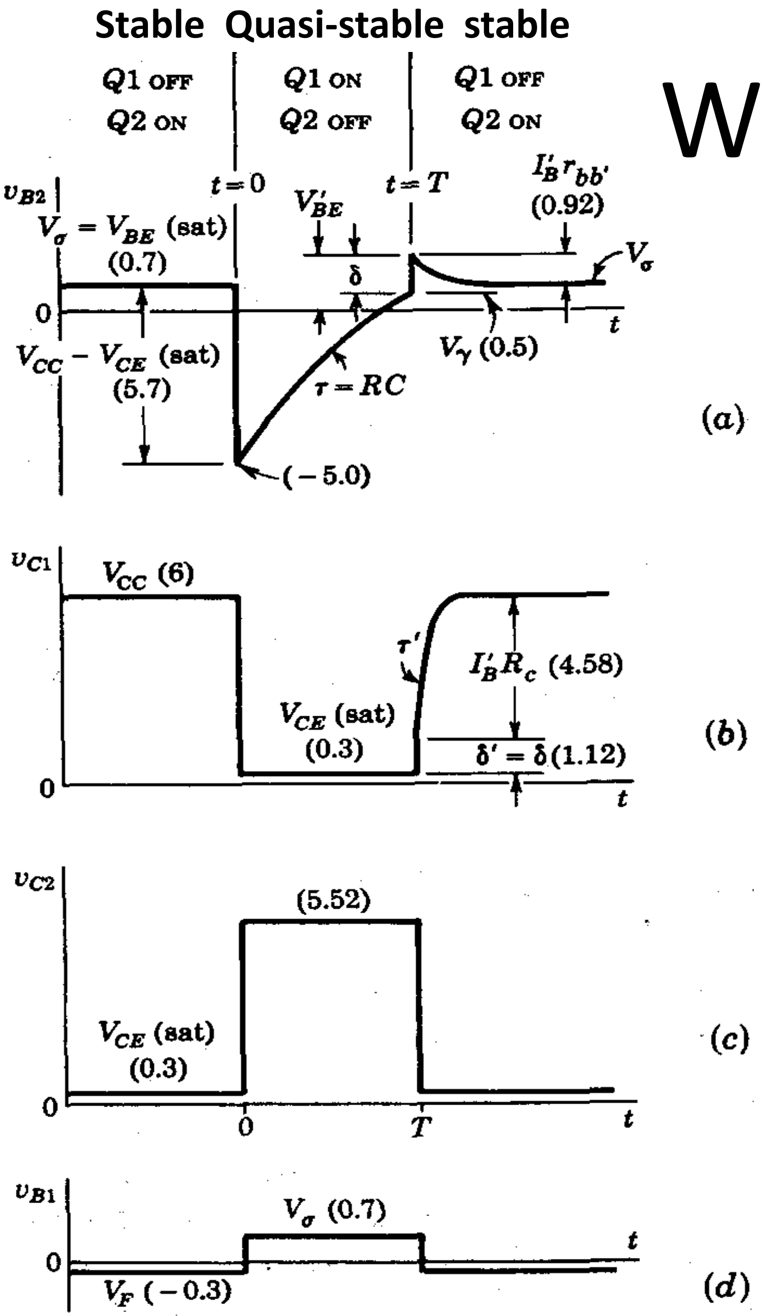



FIG. M7

**Waveforms of the collector-coupled monostable multi. The exponential portions beginning at $t = T$ have a time constant $\tau' = (R_c + r_{bb'})C$. The numerical values (in volts) refer to the circuit M6. rbb' is the base spreading resistance**

- To justify $Q_2$ is in saturation it need to prove $I_{B2} \geq Ic2/hfe$.

$$I_{C2} = \frac{V_{CC} - V_{CE}(\text{sat})}{R_c} = \frac{6 - 0.3}{1} = 5.7 \text{ mA}$$

- Minimum base current to drive Q2 in saturation Ic2/hfe. Assume hfe=30
- $I_{B2min}$= 5.7/30 mA=0.19 mA

$$I_{B2} = \frac{V_{CC} - V_{BE}(\text{sat})}{R} = \frac{6 - 0.7}{10} = 0.53 \text{ mA} > I_{B2min}.$$

- *Thus $Q_2$ is in saturation*

## The Stable State

The base voltage of $Q2$ is $v_{B2} = V_{BE}(\text{sat}) \equiv V_\sigma$.

The collector voltage of $Q2$ is $v_{C2} = V_{CE}(\text{sat})$.

The collector of $Q1$ is at $v_{C1} = V_{CC}$. The base voltage of $Q1$ is calculated by superposition to be

$$v_{BE1} = -\frac{V_{BB}R_1}{R_1 + R_2} + \frac{V_{CE}(\text{sat})R_2}{R_1 + R_2}$$

$$= -1.5\frac{10}{10 + 20} + \frac{(0.3)(20)}{10 + 20} = -0.30 \text{ V}$$

which is certainly enough to cut off $Q1$. Hence $v_{C1} = V_{CC} = 6$ V.

If a negative trigger is applied to the collector of $Q1$ or to the base of $Q2$, the multi will make a transition to its quasi-stable state. From the equivalent circuit of Fig. 11-11 we have, assuming $Q1$ is driven into saturation,

$$I_3 = \frac{6 - 0.7}{10 + 1} = 0.48 \text{ mA} \qquad I_4 = \frac{1.5 + 0.7}{20} = 0.11 \text{ mA}$$

$$I_{B1} = I_3 - I_4 = 0.48 - 0.11 = 0.37 \text{ mA}$$

From Fig. 11-9 we now calculate $I_{c1} = I_1 + I_R$. Since

$$I_1 R_c = V_{CC} - V_{CE}(\text{sat}) = 5.7 \text{ V}$$

$I_1 = 5.7$ mA. Since $I_R R = I_1 R_c + V_A = 5.7 + 5.3 = 11.0$ V, $I_R = 1.1$ mA. Hence $I_{C1} = 5.7 + 1.1 = 6.8$ mA and $(I_{B1})_{\min} = 6.8/30 = 0.23$ mA. Since $I_{B1} > (I_{B1})_{\min}$, then $Q1$ is indeed in saturation during the quasi-stable state. Hence, $v_{B1} = 0.7$ V, $v_{C1} = 0.3$ V, and at $t = 0+$

$$v_{B2} = V_\sigma - I_1 R_c = V_\sigma - [V_{CC} - V_{CE}(\text{sat})] = 0.7 - 6 + 0.3 = -5.0 \text{ V}$$

From Eq. (11-13)

$$v_{C2} = \frac{(6)(10)}{10 + 1} + \frac{(0.7)(1)}{10 + 1} = 5.52 \text{ V}$$

From Eq. (11-16)

$$I'_B = \frac{6 - 0.3 - 0.7 + 0.5}{1 + 0.2} = 4.58 \text{ mA}$$

Since $I'_B r_{bb'} = (4.58)(0.2) = 0.92$ V and $I'_B R_c = 4.58$ V, then all voltages in Fig.M7are known.

# THE EMITTER–COUPLED MONOSTABLE MULTI

A negative supply is not required.

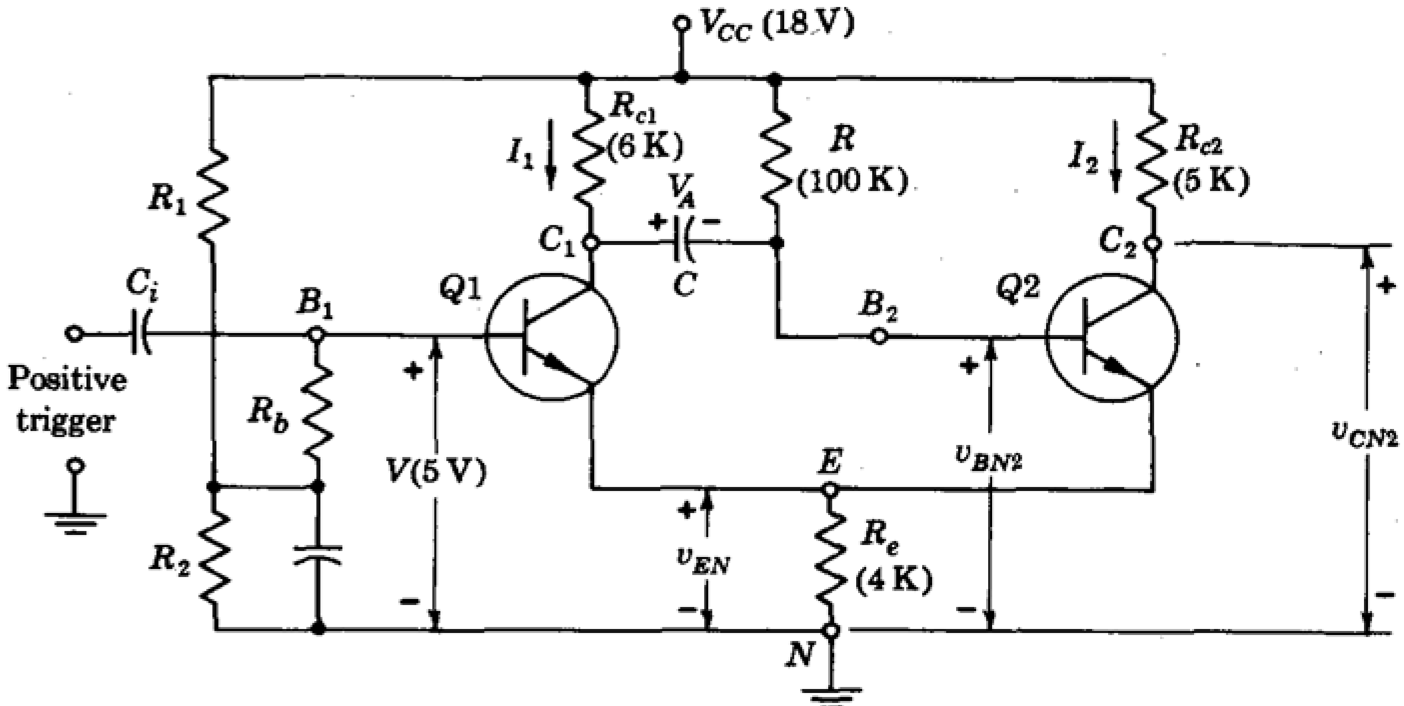



Fig. M8 An emitter-coupled monostable multi. Components and supply voltage refer to the illustrative example. The bias voltage $V$ is obtained from the divider $R_1R_2$ across the supply $V_{CC}$.

# Explanation

- Observe that the coupling from C2 to B1 is lacking and that in its place has been substituted a common emitter resistor *RE*
- No negative supply is required. The signal at *C2* is not directly involved in the regenerative loop. Hence, C2 makes an ideal point from which to obtain an output voltage. B1 is an ideal point at which to inject the triggering signal, since this base is coupled to no other point in the circuit. Hence, the trigger source cannot load the circuit. It also turns out that the pulse width is an accurately linear function of the d-c bias voltage *V* on Q1 , and hence this circuit makes an excellent gate generator whose width is easily and linearly controllable .

**TABLE 1** Typical $n$-$p$-$n$ transistor-junction voltages at 25°C

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

**EXAMPLE** Consider the emitter-coupled multi whose components and supply voltages are indicated in Fig. M8 . If $V = 5.0$ V, calculate the voltage levels of the waveforms in Fig. M9 Assume germanium transistors having $h_{FE} = 50$ and $r_{bb'} = 200\ \Omega$.

*Solution* *Stable-state calculations,* $t < 0$ In the stable state $Q1$ is OFF while $Q2$ is in saturation. The equivalent circuit from which to find the currents in $Q2$ is shown in Fig. M 10 We have again taken advantage of the simplification afforded by using the approximations of Table 1. Applying Kirchhoff's voltage law to the two meshes we have

$$104I_{B2} + 4I_2 = 17.7 \tag{11-18a}$$

$$4I_{B2} + 9I_2 = 17.9 \tag{11-18b}$$

Solving, we obtain $I_2 = 1.95$ mA and $I_{B2} = 0.095$ mA. We note that the base current is more than adequate to ensure that $Q2$ is in saturation, since the minimum current required for saturation is $I_2/h_{FE} = 1.95/50 = 0.039$ mA.

We now find that, in the stable state,

$$v_{CN1} = V_{CC} = 18 \text{ V} \tag{11-19}$$

$$v_{EN} \equiv V_{EN2} = (I_2 + I_{B2})R_e = (1.95 + 0.095)(4) = 8.2 \text{ V} \tag{11-20}$$

Since

$$v_{BE1} = v_{BN1} - V_{EN2} = 5.0 - 8.2 = -3.2 \text{ V}$$

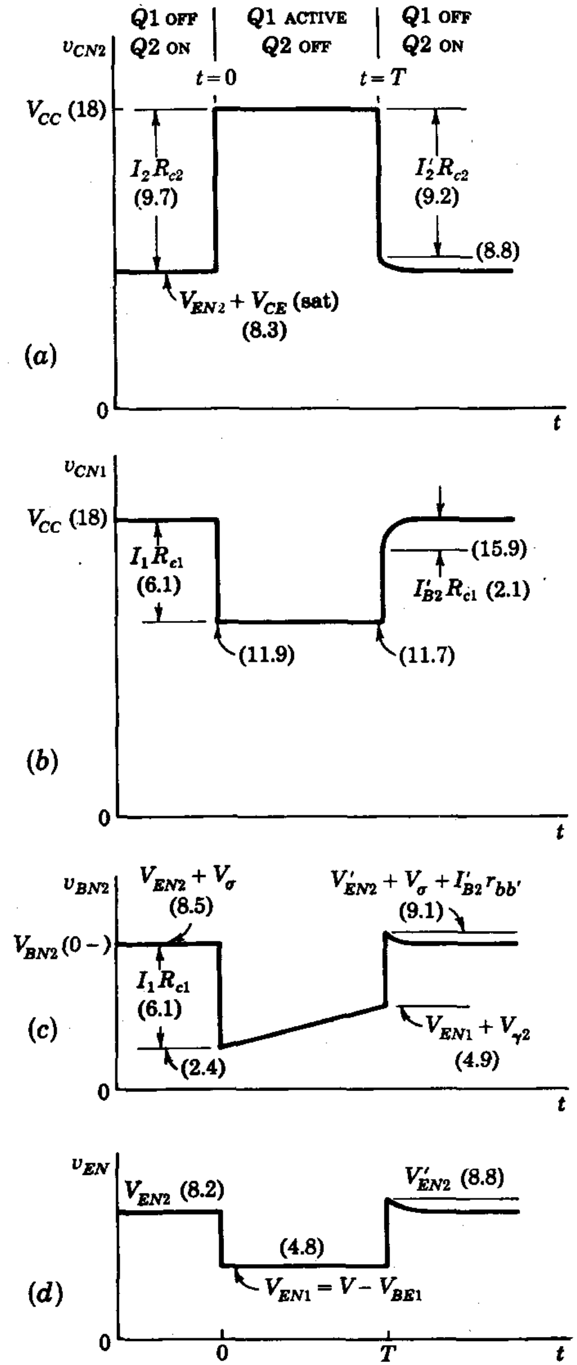
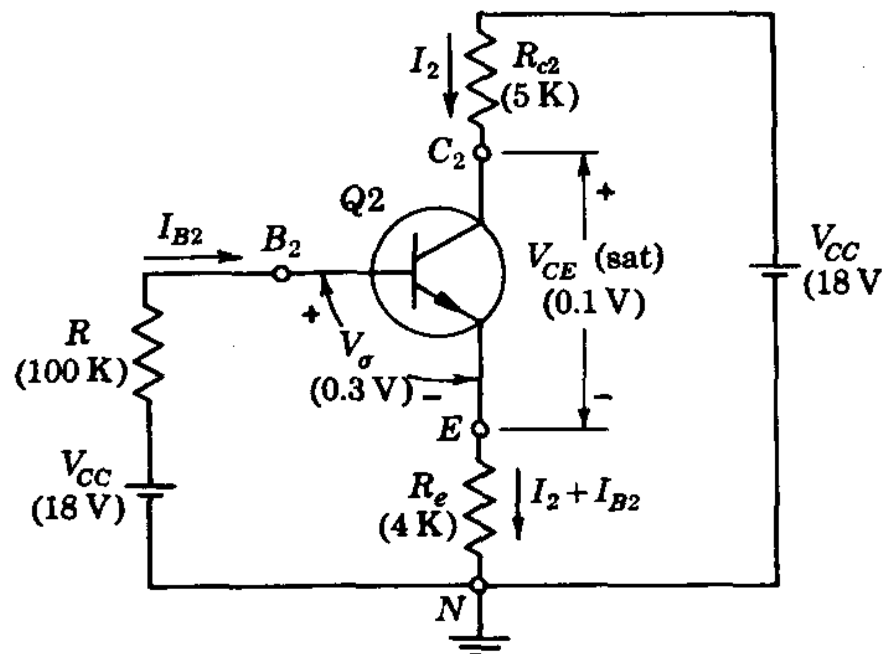



**Fig. M9** Waveforms of emitter-coupled monostable multivibrator. Numerical values (in volts) in parentheses correspond to the illustrative example.

**Fig. M10** *The equivalent circuit of $Q2$ in Fig. M8 in the stable state. The saturation voltages are given in Table 1.*



then $Q1$ is indeed OFF.

$$v_{CN2} = V_{EN2} + V_{CE}(\text{sat}) = 8.2 + 0.1 = 8.3 \text{ V} \tag{11-21}$$

$$v_{BN2} = V_{EN2} + V_{\sigma} = 8.2 + 0.3 = 8.5 \text{ V} \tag{11-22}$$

We also note from Fig. 11-13 that the capacitor voltage is

$$V_A = v_{CN1} - v_{BN2} = 18 - 8.5 = 9.5 \text{ V} \tag{11-23}$$

*Calculations at $t = 0+$* When a triggering signal, applied at $t = 0$, causes a transition from the stable to the quasi-stable state, the current in $Q2$ becomes zero and a current $I_{C1}$ flows in the collector circuit of $Q1$. This current may be determined by the bias voltage $V$. The equivalent circuit for this calculation is shown in Fig. 11-16. *Assuming that $Q1$ is operating in its active region* with $V_{BE1} = 0.2$ V (Table 6-1, page 219),

$$v_{EN} \equiv V_{EN1} = V - V_{BE1} = 5 - 0.2 = 4.8 \text{ V} \tag{11-24}$$

$$I_{C1} + I_{B1} = \frac{V_{EN1}}{R_e} = \frac{4.8}{4} = 1.2 \text{ mA} \tag{11-25}$$

Since

$$I_{C1} + I_{B1} = I_{C1}\left(1 + \frac{1}{h_{FE}}\right) \tag{11-26}$$

then

$$I_{C1} = \frac{h_{FE}}{1 + h_{FE}} \frac{V_{EN1}}{R_e} = \frac{(50)(1.2)}{51} = 1.18 \text{ mA} \tag{11-27}$$

In order to solve for $I_1$, the current in $R_{c1}$ at $t = 0+$, note that

$$I_1 + I_R = I_{C1} = 1.18 \text{ mA} \tag{11-28}$$

and from Kirchhoff's voltage law applied to the mesh containing $R_{c1}$, $C$, and $R$,

$$R_{c1}I_1 + V_A - RI_R = 0 \tag{11-29}$$

Since the capacitor voltage does not change at the transition, then, from Eq. (11-23), $V_A = 9.5$ V and Eq. (11-29) becomes

$$6I_1 - 100I_R = -9.5 \text{ V} \tag{11-30}$$

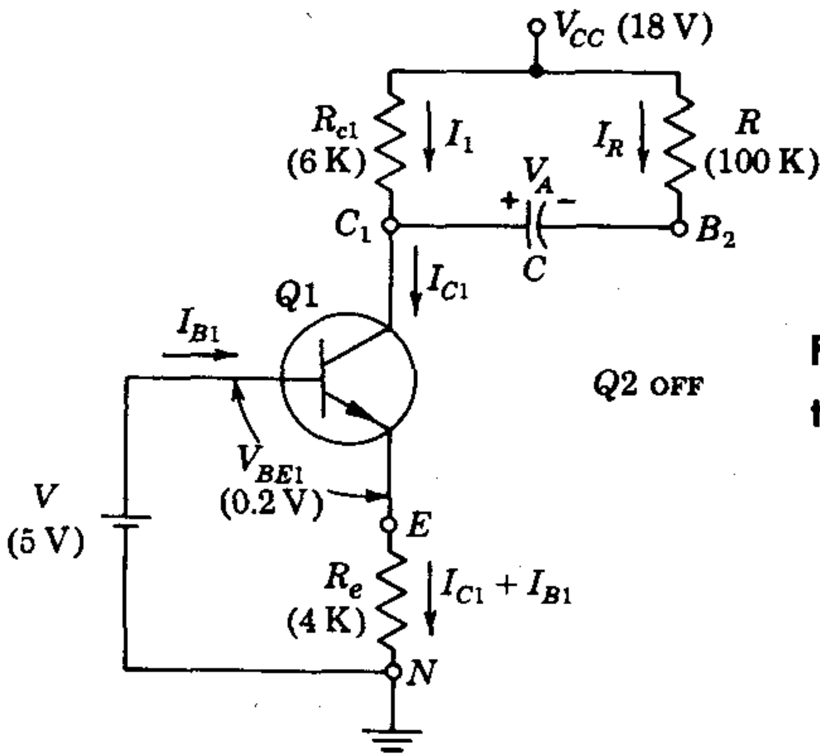



Fig. M11 The circuit of Fig. M8 during the quasi-stable state when $Q2$ is OFF.

Solving Eqs. (11-28) and (11-30) we obtain

$$I_R = 0.156 \text{ mA} \qquad \text{and} \qquad I_1 = 1.02 \text{ mA} \tag{11-31}$$

Accordingly, at $t = 0+$ the multi voltages are

$$v_{CN2} = V_{CC} = 18 \text{ V} \tag{11-32}$$

$$v_{CN1} = V_{CC} - I_1R_{c1} = 18 - (1.02)(6) = 11.9 \text{ V} \tag{11-33}$$

$$v_{EN} = V_{EN1} = V - V_{BE1} = 4.8 \text{ V} \tag{11-34}$$

$$v_{BN2} = V_{BN2}(0-) - I_1R_{c1} = 8.5 - 6.1 = 2.4 \text{ V} \tag{11-35}$$

Equation (11-35) follows from the fact that the abrupt change $I_1R_{c1}$ in the voltage at $C_1$ is transmitted unattenuated to $B_2$ at $t = 0$. Alternatively, we see from Fig. 11-16 that $v_{BN2}$ may be calculated from

$$v_{BN2} = v_{CN1} - V_A = 11.9 - 9.5 = 2.4 \text{ V} \tag{11-36a}$$

or, also, from

$$v_{BN2} = V_{CC} - I_RR = 18 - 15.6 = 2.4 \text{ V} \tag{11-36b}$$

In the above calculation for $v_{CN1}$ we have explicitly taken into account the effect of the output impedance of the transistor stage $Q1$. The current $I_R$ flows back into this impedance and increases the voltage $v_{CN1}$. Neglecting this effect we would have obtained

$$v_{CN1} = V_{CC} - I_{C1}R_{c1} = 18 - (1.18)(6) = 10.9 \text{ V}$$

which is too low by 1.0 V. The output impedance is $R_o \approx R_{c1} = 6$ K, since the impedance seen looking into the collector of $Q_1$ operating in its active region is much larger than 6 K. Note that $I_RR_o \approx (0.16)(6) \approx 1.0$ V.

The loading of $R$ does not influence the waveforms in the collector-coupled one-shot if $Q1$ is driven into saturation, since $R_o$ is then very small. The loading only affects the value of $(h_{FE})_{min}$ required to keep $Q1$ in saturation

We must check our assumption that $Q1$ is in its active region. Since

$$v_{CB1} = v_{CN1} - v_{BN1} = 11.9 - 5.0 = +6.9 \text{ V} \tag{11-37}$$

is positive, then the collector junction is reverse-biased (for an *n-p-n* transistor). Hence, $Q1$ is indeed in its active region.

*Calculations at* $t = T-$ During the quasi-stable state $v_{CN2}$, $v_{EN1}$ and $i_{C1}$ remain constant, and $v_{BN2}$ increases exponentially with a time constant $(R + R_o)C$. This voltage starts at 2.4 V at $t = 0+$ and rises to the point at which $Q2$ attains its cutin value at $t = T-$ when

$$v_{BN2} = V_{EN1} + V_{\gamma 2} = 4.8 + 0.1 = 4.9 \text{ V} \tag{11-38}$$

The voltage $v_{CN1}$ at $t = T-$ is calculated as follows. At this time

$$I_R = \frac{V_{CC} - v_{BN2}}{R} = \frac{18 - 4.9}{100} = 0.13 \text{ mA} \tag{11-39}$$

$$I_1 = I_{C1} - I_R = 1.18 - 0.13 = 1.05 \text{ mA} \tag{11-40}$$

and hence

$$v_{CN1} = V_{CC} - I_1 R_{c1} = 18 - (1.05)(6) = 11.7 \text{ V} \tag{11-41}$$

Thus, while at $t = 0+$, $v_{CN1} = 11.9$ V, this voltage falls to 11.7 V at $t = T-$, as indicated in Fig. 11-14*b*.

The capacitor voltage at $t = T-$ is

$$V_A = v_{CN1} - v_{BN2} = 11.7 - 4.9 = 6.8 \text{ V} \tag{11-42}$$

*Calculations at* $t = T+$ The voltage levels attained at $t = T+$, immediately after the reverse transition. at which time the overshoots occur, may be calculated from the circuit of Fig. M12. Here, $I'_{B2}$ and $I'_2$ are the base and collector currents at $t = T+$. To perform the calculation we shall use the fact that the voltage $V_A = 6.8$ V across $C$ is the same immediately before and after the reverse transition. Since the overshoots are not required with great precision, then for $R \gg R_{c1}$

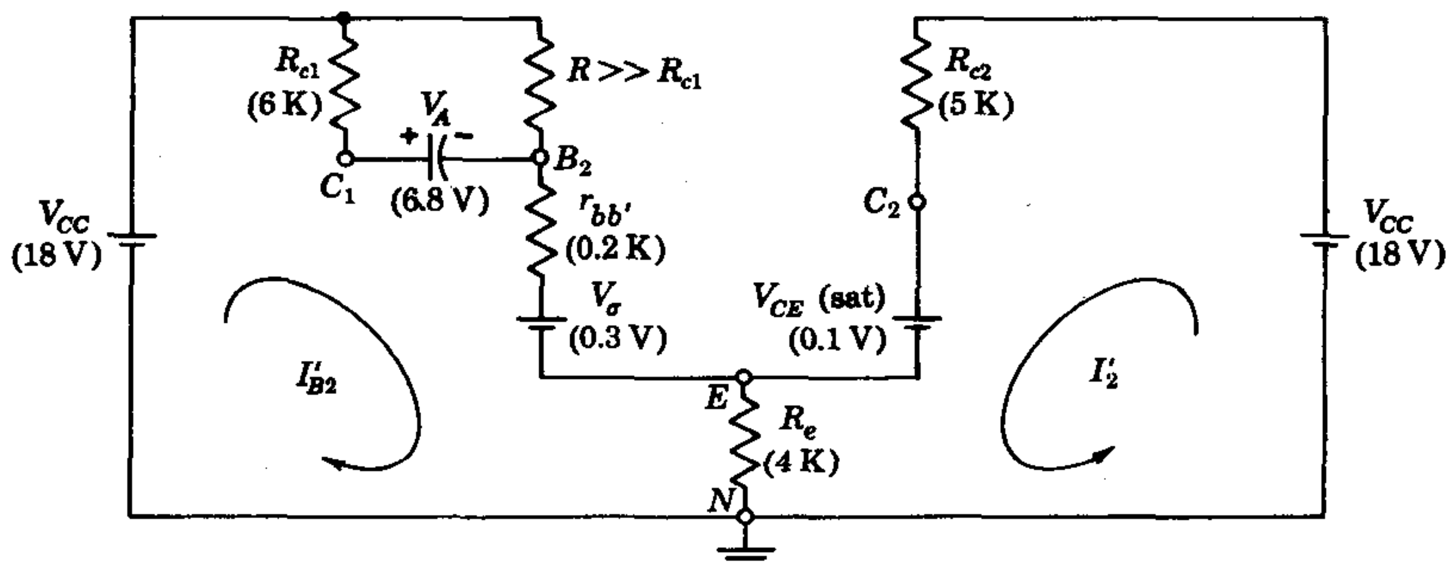



**Fig. M12** The equivalent circuit at $t = T+$ from which to calculate the overshoots. The emitter voltage is $V_{EN} = (I'_2 + I'_{B2})R_E$. $Q1$ is OFF, and $Q2$ is driven heavily into saturation.

the current in $R$ may be neglected compared with $I'_{B2}$. We may write the mesh equations

$$18 - 6.8 - 0.3 = 10.2I'_{B2} + 4I'_2 = 10.9 \tag{11-43a}$$

$$18 - 0.1 = 4I'_{B2} + 9I'_2 = 17.9 \tag{11-43b}$$

Solving, we obtain

$$I'_{B2} = 0.35 \text{ mA} \qquad \text{and} \qquad I'_2 = 1.84 \text{ mA}$$

Hence at $t = T+$

$$v_{EN} \equiv V'_{EN2} = (I'_2 + I'_{B2})R_e = (1.84 + 0.35)(4) = 8.76 \text{ V} \tag{11-44}$$

$$v_{CN2} = V'_{EN2} + V_{CE}(\text{sat}) = 8.76 + 0.1 = 8.86 \text{ V} \tag{11-45}$$

or, alternatively,

$$v_{CN2} = V_{CC} - I'_2R_{c2} = 18 - (1.84)(5) = 8.80 \text{ V} \tag{11-46}$$

$$v_{CN1} = V_{CC} - I'_{B2}R_{c1} = 18 - (0.35)(6) = 15.9 \text{ V} \tag{11-47}$$

$$v_{BN2} = V'_{EN2} + V_\sigma + I'_{B2}r_{bb'}$$

$$= 8.76 + 0.3 + (0.35)(0.2) = 9.13 \text{ V} \tag{11-48}$$

From Fig. **M12** we see that the overshoots decay with a time constant

$$\tau' = C\left(R_{c1} + r_{bb'} + \frac{R_e R_{c2}}{R_e + R_{c2}}\right) \tag{11-49}$$

**Extreme Limits of $V$** In the stable state $Q1$ must be OFF. This condition establishes a maximum allowable value $V_{max}$ for the bias voltage $V$ applied to the base of $Q1$. When $Q1$ goes ON the current $I_{C1}$ must be large enough to drive $Q2$ to cutoff. This condition establishes a minimum allowable bias voltage $V_{min}$. *For proper operation of the circuit the bias voltage $V$ must lie between these two values $V_{max}$ and $V_{min}$.* These extreme voltages will now be calculated.

Since in the stable state $Q1$ is OFF, $V$ must not be greater than $V_{EN2}$ by more than the cutin voltage $V_{\gamma 1}$. The maximum allowable value of $V$ is therefore

$$V_{max} = V_{EN2} + V_{\gamma 1} \tag{11-50}$$

If there is to be a quasi-stable state, then the current $I_1$ must be large enough to drive $Q2$ OFF. Let us call this minimum current $(I_1)_{min}$ and the corresponding value of emitter voltage $(V_{EN1})_{min}$. Then the minimum input voltage is

$$V_{min} = (V_{EN1})_{min} + V_{BE1} \tag{11-51}$$

where $V_{BE1}$ is the base-to-emitter voltage corresponding to an emitter current $V_{EN1}/R_e$. To find $(V_{EN1})_{min}$ proceed as follows. The voltage at the base of

$Q2$ at $t = 0+$ must be less than the emitter voltage $V_{EN1}$ by at least the cutin voltage $V_{\gamma 2}$. Using Eq. (11-35),

$$V_{BN2}(0+) = V_{BN2}(0-) - I_1 R_{c1} \leq V_{EN1} + V_{\gamma 2} \tag{11-52}$$

The relationship between $I_1$ and $V_{EN1}$ is found from Eqs. (11-27) and (11-28):

$$I_{C1} = \frac{h_{FE}}{1 + h_{FE}} \frac{V_{EN1}}{R_e} = I_1 + I_R \tag{11-53}$$

We also have from Fig. 11-16 that

$$RI_R = V_{CC} - V_{BN2}(0+) = V_{CC} - V_{BN2}(0-) + I_1 R_{c1} \tag{11-54}$$

If the equals sign applies in Eq. (11-52), then this equation together with Eqs. (11-53) and (11-54) determines the three minimum values $I_1$, $V_{EN1}$, and $I_R$. If we solve for $(V_{EN1})_{\min}$ and substitute this value into Eq. (11-51) we obtain

$$V_{\min} = V_{BE1} + \frac{V_{BN2}(0-) - V_{\gamma 2} + (R_{c1}/R)(V_{CC} - V_{\gamma 2})}{1 + (R_{c1}/R) + (R_{c1}/R_e)[h_{FE}/(1 + h_{FE})]} \tag{11-55}$$

For the circuit of Fig. 11-13, Eq. (11-50) gives

$$V_{\max} = 8.2 + 0.1 = 8.3 \text{ V}$$

and Eq. (11-55) yields

$$V_{\min} = 0.2 + \frac{8.5 - 0.1 + (0.06)(18 - 0.1)}{1 + 0.06 + (6/4)(50/51)} = 3.94 \text{ V}$$

## GATE WIDTH OF AN EMITTER–COUPLED ONE–SHOT

$v_{BN2}$ at $t = 0+$ is, from Fig. M9c

$$V_{BN2}(0+) = V_{BN2}(0-) - I_1 R_{c1} \tag{11-56}$$

If $Q2$ did not conduct, then as $t \to \infty$, $v_{BN2}$ would approach $V_{CC}$. Hence, the instantaneous voltage at the base of $Q2$ is given by

$$v_{BN2} = V_{CC} - [V_{CC} - V_{BN2}(0-) + I_1 R_{c1}]\epsilon^{-t/\tau} \tag{11-57}$$

where $\tau = C(R + R_{c1})$. At $t = T-$, we see from Fig. M9c that

$$v_{BN2} = V_{EN1} + V_{\gamma 2}$$

Substituting this value into Eq. (11-57) and solving for $T$, we obtain

$$T = \tau \ln \frac{V_{CC} - V_{BN2}(0-) + I_1 R_{c1}}{V_{CC} - V_{EN1} - V_{\gamma 2}} \tag{11-58}$$

The delay time is found to vary fairly linearly with the bias voltage $V$, as will be demonstrated in the following illustrative problem.

**EXAMPLE** (*a*) Find the delay time $T$ as a function of the input voltage $V$ for the one-shot of Fig. M8 (*b*) For what value of $V$ is $T = 0$?

*Solution* *a*. Equation (11-58) gives

$$T = \tau \ln \frac{18 - 8.5 + 6I_1}{18 - V_{EN1} - 0.1} = \tau \ln \frac{9.5 + 6I_1}{17.9 - V_{EN1}} \tag{11-59}$$

From Eq. (11-24)

$$V_{EN1} = V - V_{BE1} = V - 0.2 \tag{11-60}$$

We may obtain $I_1$ in terms of $V_{EN1}$ and hence $V$ from Eqs. (11-53) and (11-54). For the circuit under consideration these equations are

$$\frac{50}{51}\frac{V_{EN1}}{4} = I_1 + I_R = 0.245V_{EN1}$$

$$100I_R = 18 - 8.5 + 6I_1 = 9.5 + 6I_1$$

Eliminating $I_R$, we obtain

$$I_1 = 0.232V_{EN1} - 0.090 = 0.232V - 0.136$$

where we have made use of Eq. (11-60). Substituting for $I_1$ and $V_{EN1}$ in Eq. (11-59) yields

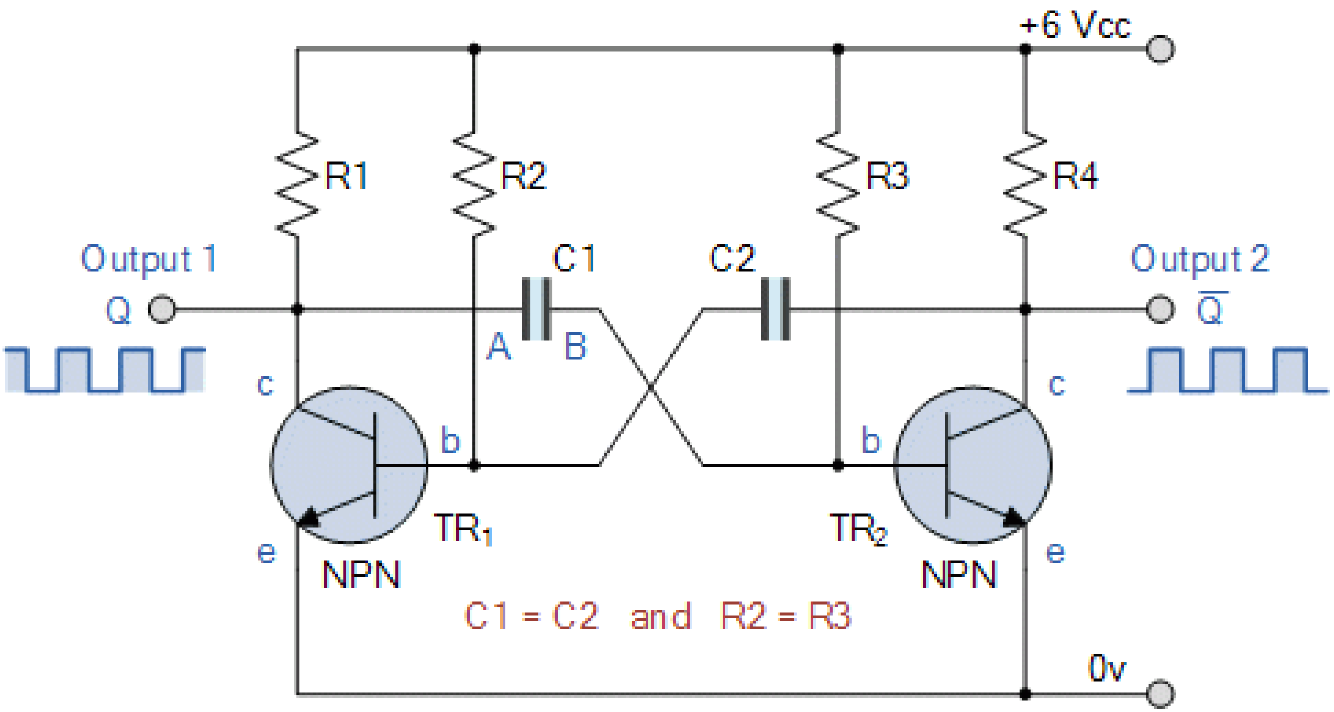
$$T = \tau \ln \frac{8.68 + 1.39V}{18.1 - V} \tag{11-61}$$

*b*. If the numerator is equated to the denominator, then $T = 0$:

$$8.68 + 1.39V = 18.1 - V \qquad \text{or} \qquad V = 3.94 \text{ V}$$

Zero delay occurs at $V = V_{min}$, as might have been anticipated.

# Astable Multivibrator

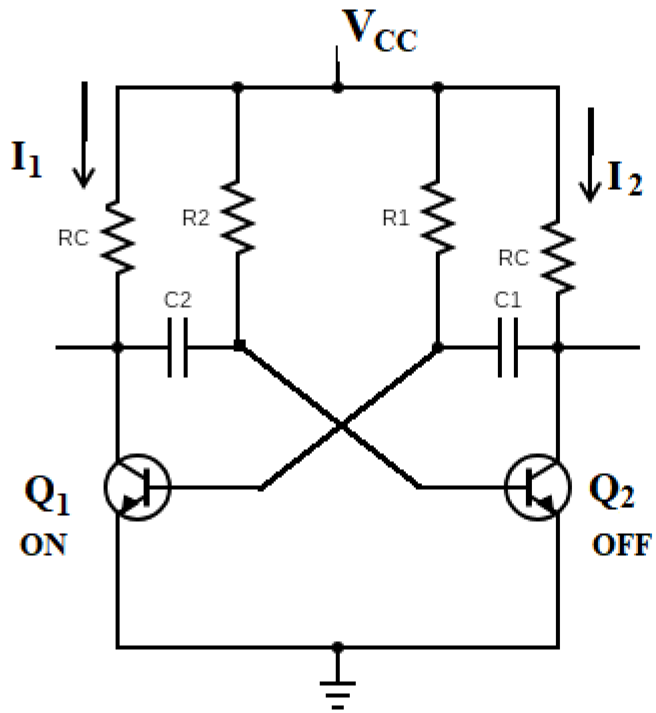


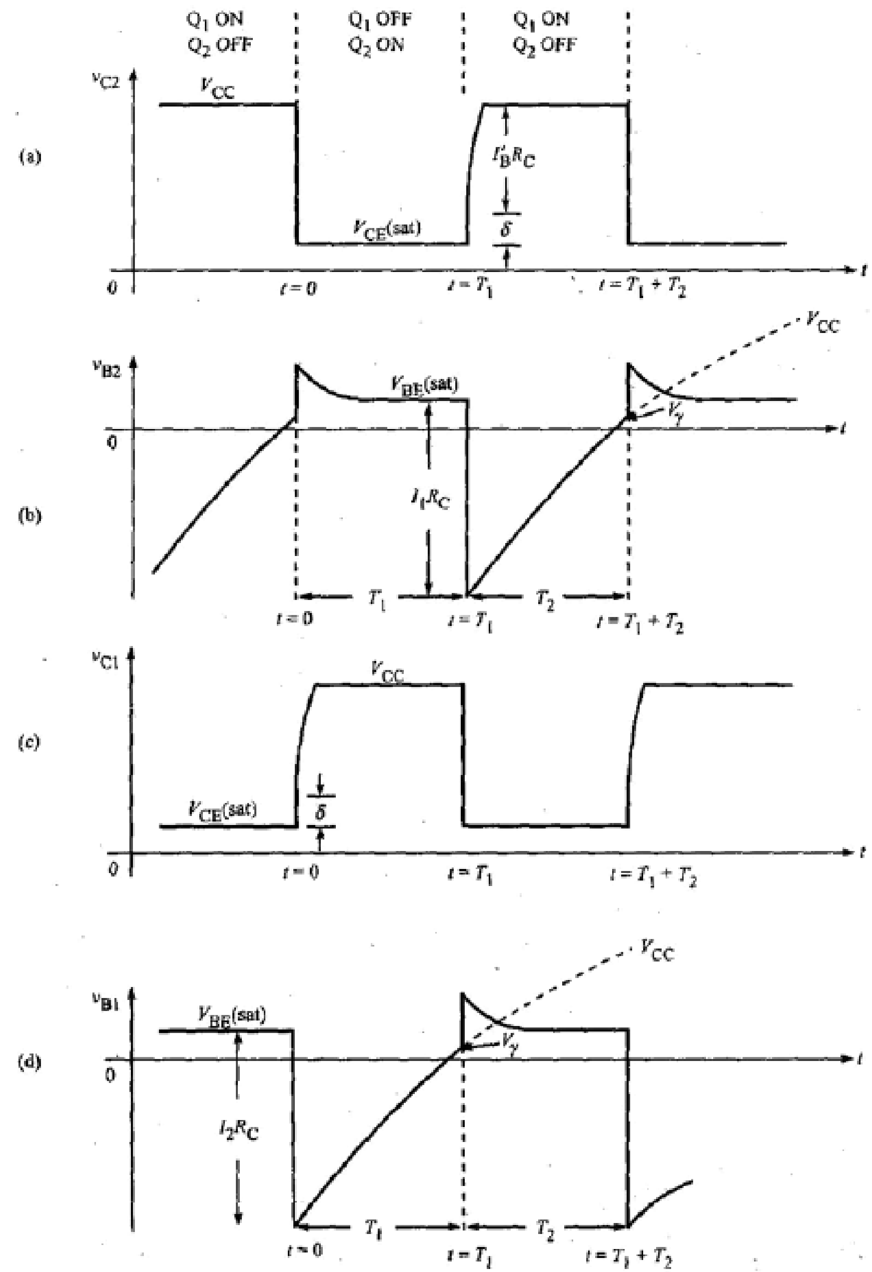
- Regenerative switching circuits such as **Astable Multivibrators** are the most commonly used type of relaxation oscillator because not only are they simple, reliable and ease of construction they also produce a constant square wave output waveform.
- **Astable Multivibrator** has automatic built in triggering which switches it continuously between its two unstable states both set and reset.
- Two outputs that repeatedly change state at a rate determined by the time constants of its feedback network. Although largely superseded by its equivalent op amp or timer IC versions in many applications, it is still a useful and flexible design for square wave and pulse generation.
- The circuit produces two anti-phase square wave signals, with an amplitude almost equal to its supply voltage.

- two cross-coupled inverters, i.e. the output of the first stage is coupled to the input of
- the second stage and the output of the second stage is coupled to the input of the first stage. In
- bistable circuits both the coupling elements are resistors (i.e. both are dc couplings). In
- monostable circuits, one coupling element is a capacitor (ac coupling) and the other coupling
- element is a resistor (dc coupling) In astable multivibrators both the coupling elements are capacitors (i.e. both are ac couplings).

- It is a free running multivibrator.
- It generates square waves.
- It is used as a master oscillator.
- There are two types of astable multivibrators:
    1. Collector-coupled astable multivibrator
    2. Emitter-coupled astable multivibrator

# Collector-coupled astable multivibrator





Q1 ON
Q2 OFF
Q1 OFF
Q2 ON
Q1 ON
Q2 OFF
(a)
vC2
VCC
I'B RC
δ
VCE(sat)
0
t = 0
t = T1
t = T1 + T2
t
(b)
vB2
VCC
VBE(sat)
Vγ
I1RC
T1
T2
(c)
vC1
VCC
VCE(sat)
δ
(d)
vB1
VBE(sat)
I2RC

Consider the waveform at the base of $Q_1$ shown in Figure (d). At $t = 0$,

$$v_{B1} = V_{BE}(\text{sat}) - I_2 R_C$$

But

$$I_2 R_C = V_{CC} - V_{CE}(\text{sat})$$

$\therefore$

$$\text{At } t = 0,\ v_{B1} = V_{BE}(\text{sat}) - V_{CC} + V_{CE}(\text{sat})$$

For $0 < t < T_1$, $v_{B1}$ rises exponentially towards $V_{CC}$ given by the equation,

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

$\therefore$

$$v_{B1} = V_{CC} - [V_{CC} - (V_{BE}(\text{sat}) - V_{CC} + V_{CE}(\text{sat}))]e^{-t/\tau_1}, \text{ where } \tau_1 = R_1 C_1$$

At $t = T_1$, when $v_{B1}$ rises to $V_\gamma$, $Q_1$ conducts

$\therefore$

$$V_\gamma = V_{CC} - [2V_{CC} - (V_{BE}(\text{sat}) + V_{CE}(\text{sat}))]e^{-T_1/R_1C_1}$$

or

$$e^{T_1/R_1C_1} = \frac{2\left[V_{CC} - \dfrac{V_{BE}(\text{sat}) + V_{CE}(\text{sat})}{2}\right]}{V_{CC} - V_\gamma}$$

$$T_1 = R_1C_1 \ln \frac{2\left[V_{CC} - \dfrac{V_{CE}(\text{sat}) + V_{BE}(\text{sat})}{2}\right]}{V_{CC} - V_\gamma}$$

$$T_1 = R_1C_1 \ln 2 + R_1C_1 \ln \frac{\left[V_{CC} - \dfrac{V_{CE}(\text{sat}) + V_{BE}(\text{sat})}{2}\right]}{V_{CC} - V_\gamma}$$

At room temperature for a transistor,

$$V_\gamma = \frac{V_{CE}(\text{sat}) + V_{BE}(\text{sat})}{2}$$

$\therefore$

$$T_1 = R_1C_1 \ln 2 = 0.693 R_1C_1$$

- On similar lines considering the waveform of Figure (b), we can show that the time $T_2$ for which $Q_2$ is OFF and $Q_1$ is ON is given by

$$T_2 = R_2C_2 \ln 2 = 0.693R_2C_2$$

- The period of the waveform, $T = 2 \times 0.693RC = 1.386RC$

$R_1 = R_2 = R$, and $C_1 = C_2 = C$, then $T_1 = T_2 = T/2$

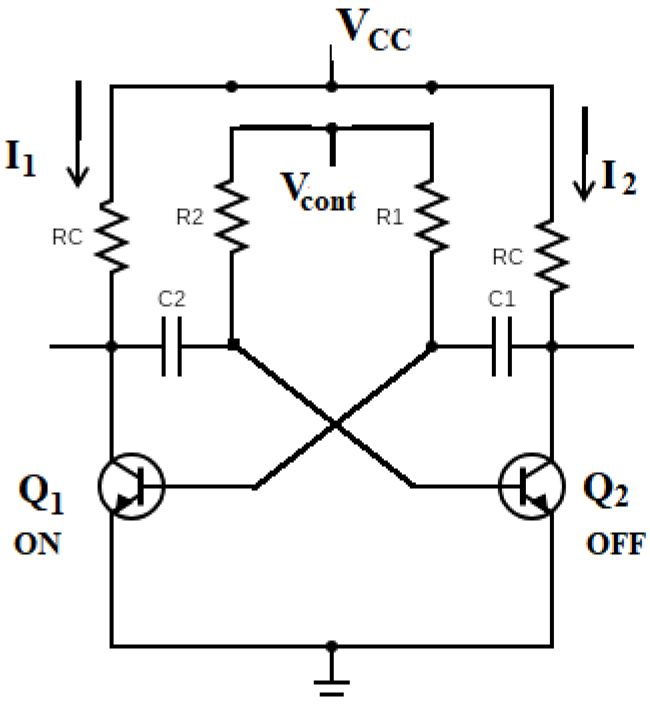
- The frequency of oscillation,

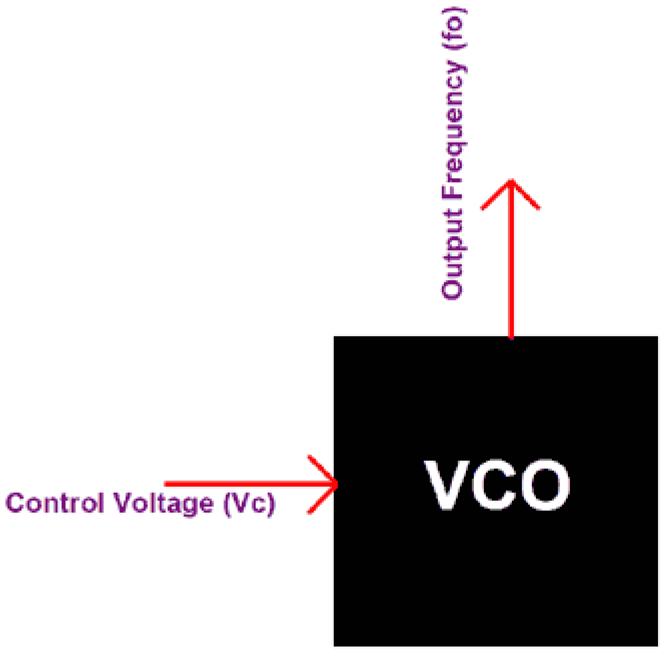
$$f = \frac{1}{1.386RC}$$

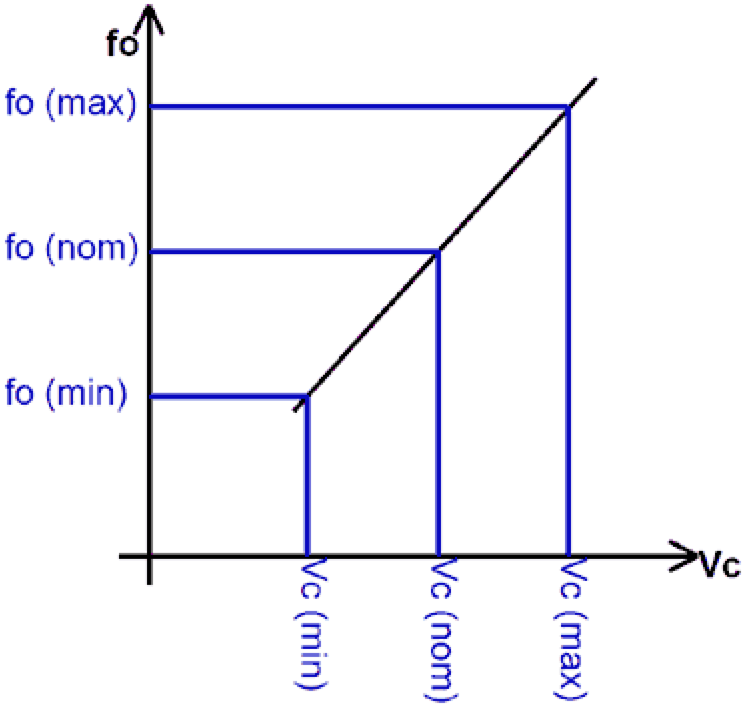
# VCO

- **Applications of VCO**
- Tone Generators
- Function generators
- Phase-Locked Loops
- In synthesizers to generate variable tones for the production of electronic music
- In communication equipment these are used as frequency synthesizers
- Clock generators
- Frequency Shift Keying

# Circuit diagram of VCO




Output Frequency (fo)
VCO
Control Voltage (Vc)


fo
fo (max)
fo (nom)
fo (min)
Vc
Vc (min)
Vc (nom)
Vc (max)

- In this circuit a time constant resistors R1 and R2  are brought out to an external control line $V_{control}$.
- The voltage to which C1 and C2 discharges through R1 and R2 changes with change in $V_{control}$ voltage.
- Hence the discharging speed is increased with increase of $V_{control}$.

# Calculation of frequency

* H.W.