

$$C = \frac{V_{max}}{f \cdot R_L \cdot V_{rip}}$$

H.W

$$C = \frac{V_{max}}{2f R_L V_{rip}}$$

f.w

ANALOG CIRCUITS

- ① OP-AMP. ✓
- ② Linear wave shaping circuit. ✓
- ③ Diode circuit
 1. Rectifiers and filter ✓
 2. Clippers & clamper circuit. ✓
- ④ Transistor biasing & stabilisation
- ⑤ Amplifiers
 1. Low freq. analysis.
 2. High freq. analysis.
 - Low & high freq analysis of FET
 - multistage amplifier.
 - feedback amplifier.
 - oscillator. (sinusoidal)
 - Power Amplifier
- ⑥ wave form generators: (schmitt trigger) ✓
 - multivibrator ✓
 1. Bistable
 2. monostable (square wave gen)
 3. Astable multi

→ By using opamp ✓

By using transistor

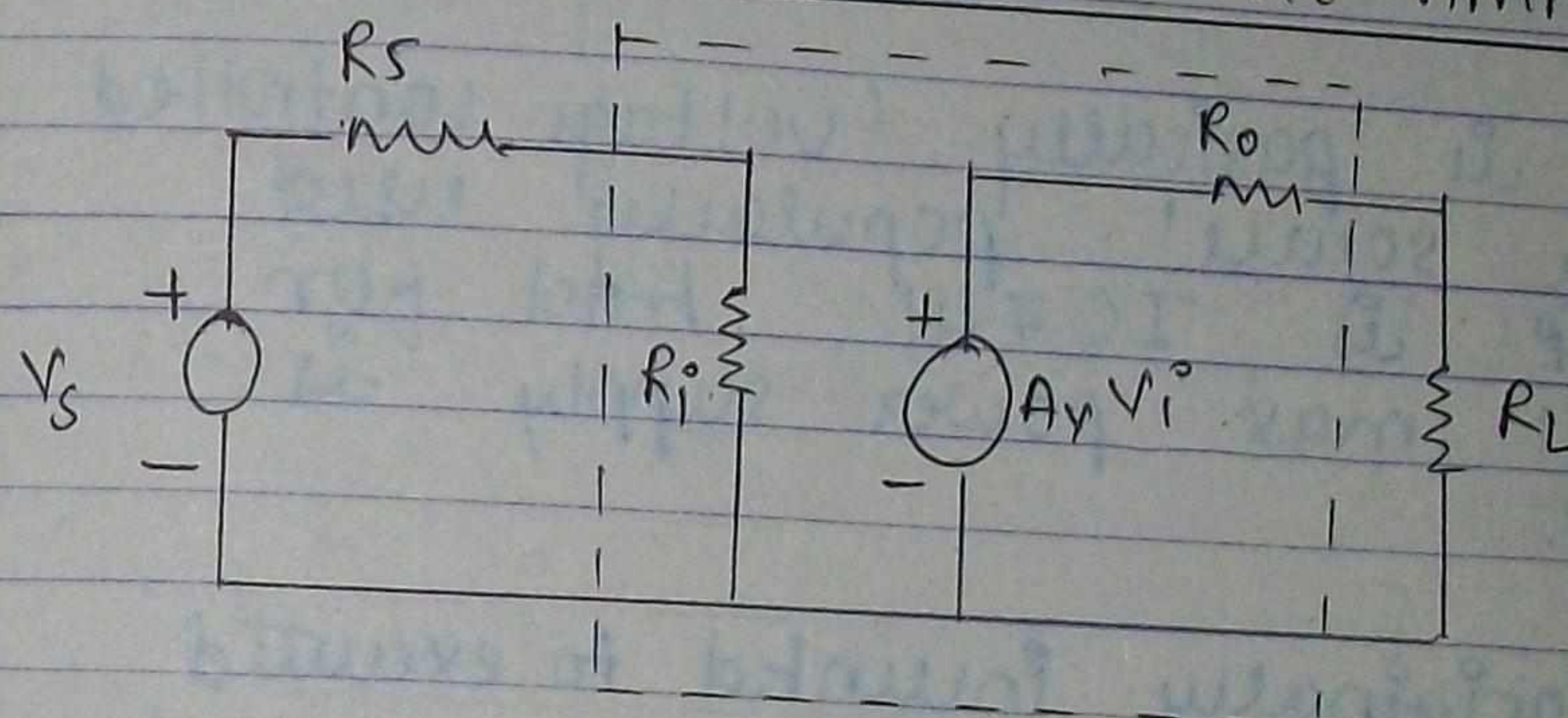
By using 555 timer
- Triangular wave generator. ✓

OPERATIONAL AMPLIFIERS

20/08/11

- OP-amp is basically a "VOLTAGE CONTROLLED VOLTAGE SOURCE" or "voltage amplifier."

⇒ EQUIVALENT CIRCUIT OF VOLTAGE AMPLIFIER -



① $R_i \gg R_s$

$R_i = \infty$ (ideally)

② $A_v = A_{OL} = \text{very high.}$

$A_v = \infty$ (ideally)

$A_v = 10^6$ (practically).

③ $R_o \ll R_L$

$R_o = 0$ (ideally)

$R_o \ll R_L$ (practically)

- OPAMP is monolithic integrated unit or a semicondⁿ chip fabricated with VLSI by using epitaxial method.

- OPAMP is basically 'voltage controlled voltage source' popularly used. OPAMP is IC 741. And for IC 741 max power supply is ± 15 V.

- It is originally invented to execute mathematical operations hence called operational amp.

- It is versatile, predictable and economic system building block. It has small size, high reliability, reduced cost and low offset voltage & current.

- The OPAMP is "Direct coupled High Gain amplifier" i.e. open loop gain is very high. Therefore the freq. stability of the signal is very less and to compensate this, a small amount of negative feedback so that freq. stability increases & gain reduced.

- OPAMP are generally operated under closed loop condition i.e. by applying negative feedback.

- In an OPAMP, Gain bandwidth product is a constant.

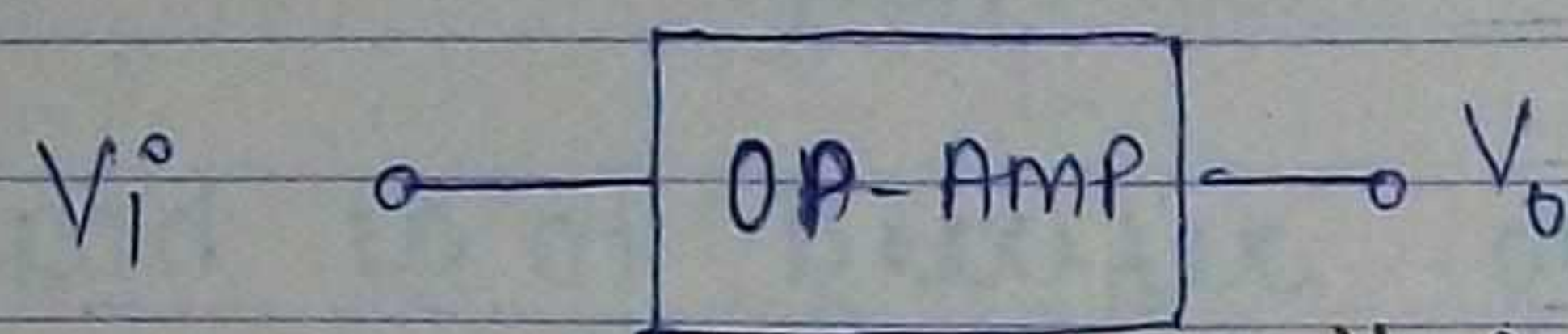
- It is also referred to as basic linear integrated circuit.

⇒

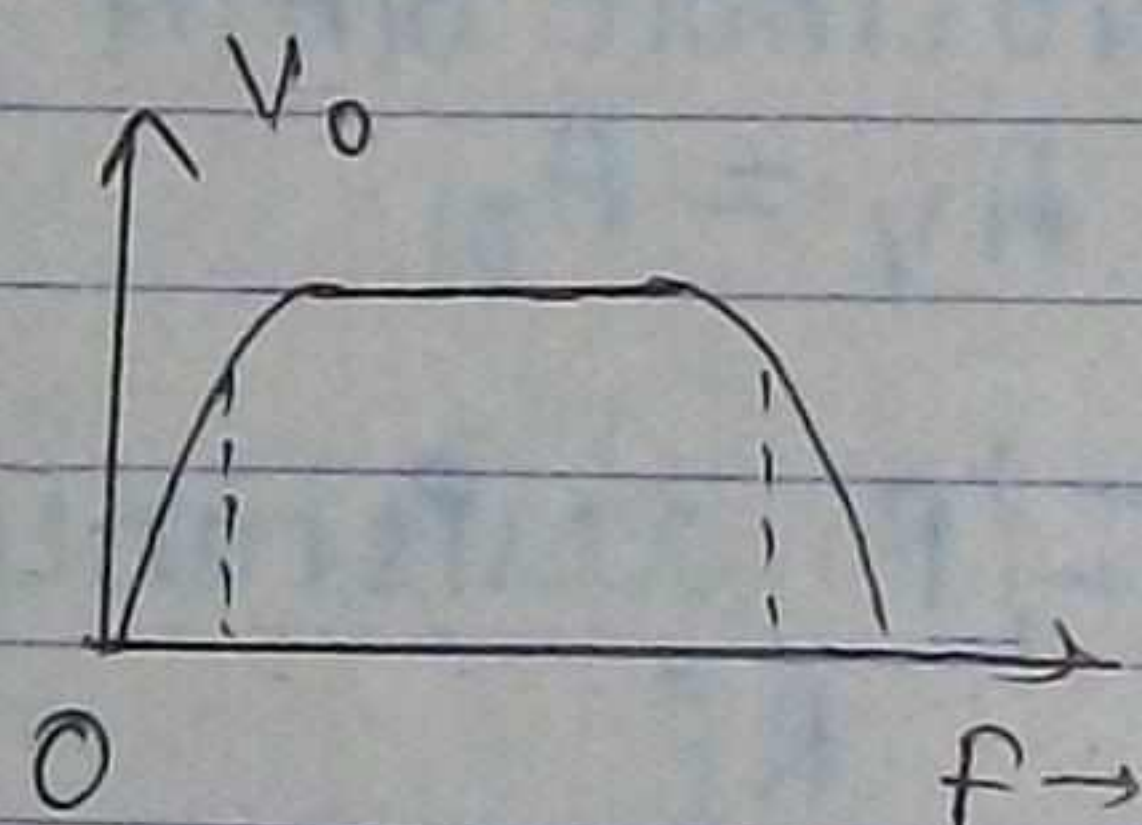
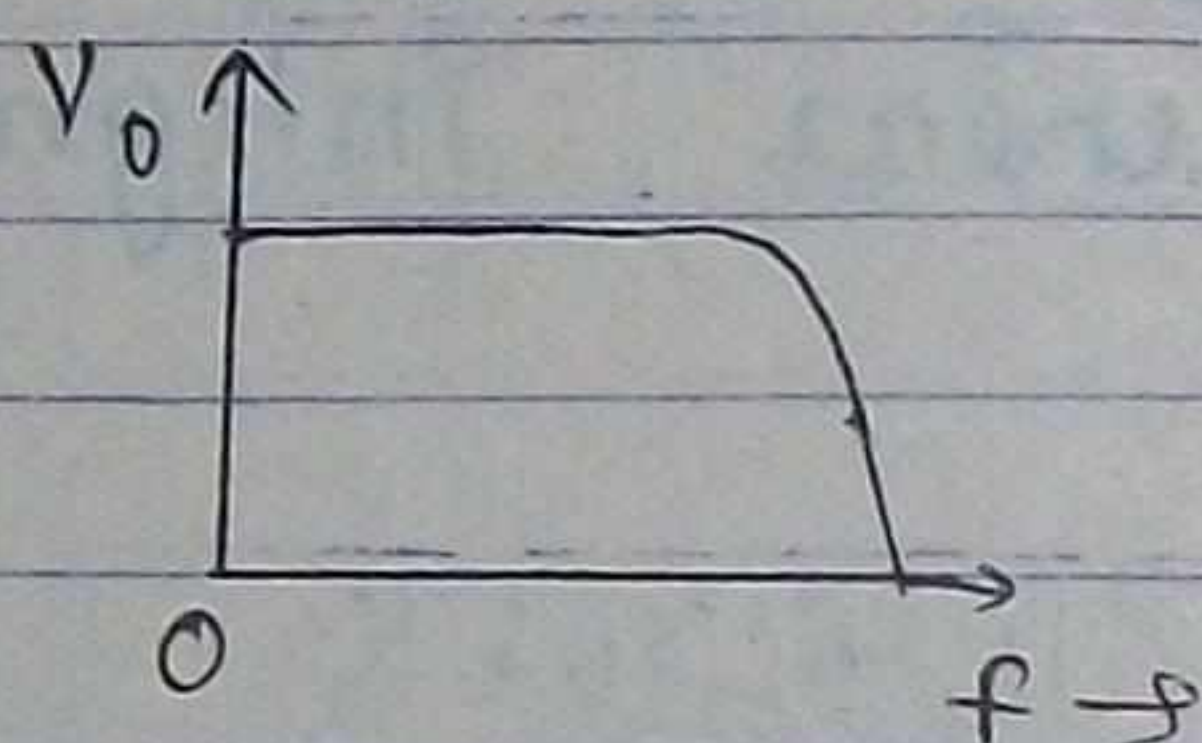
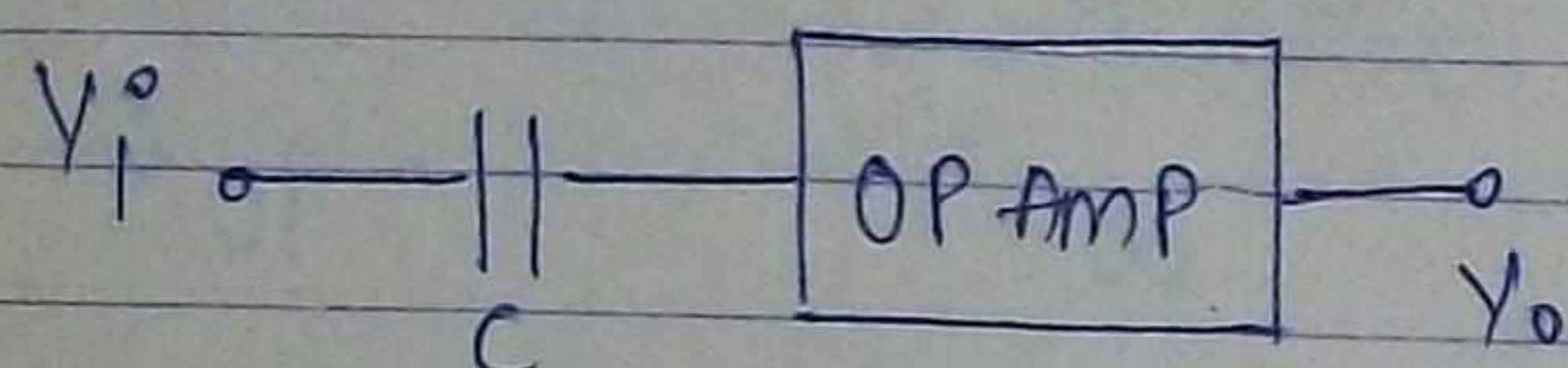
CHARACTERISTIC	IDEAL OPAMP	PRACTICAL OPAMP.
① VOLTAGE GAIN $A_V = A_{OL}$	∞	10^6
② I/P resistance R_i	∞	$10^6 \Omega = 1 \text{ M}\Omega$
③ O/P resistance R_o	0	$10 - 100 \Omega$
④ Gain Bandwidth product	∞	10^6 Hz
⑤ Slew Rate	∞	$80 \text{ V}/\mu\text{sec}$
⑥ CMRR	∞	10^6 or 120 dB

⑦ Bandwidth ∞ 10^6 Hz.

- Direct coupled means amplify both DC as well as AC signals.



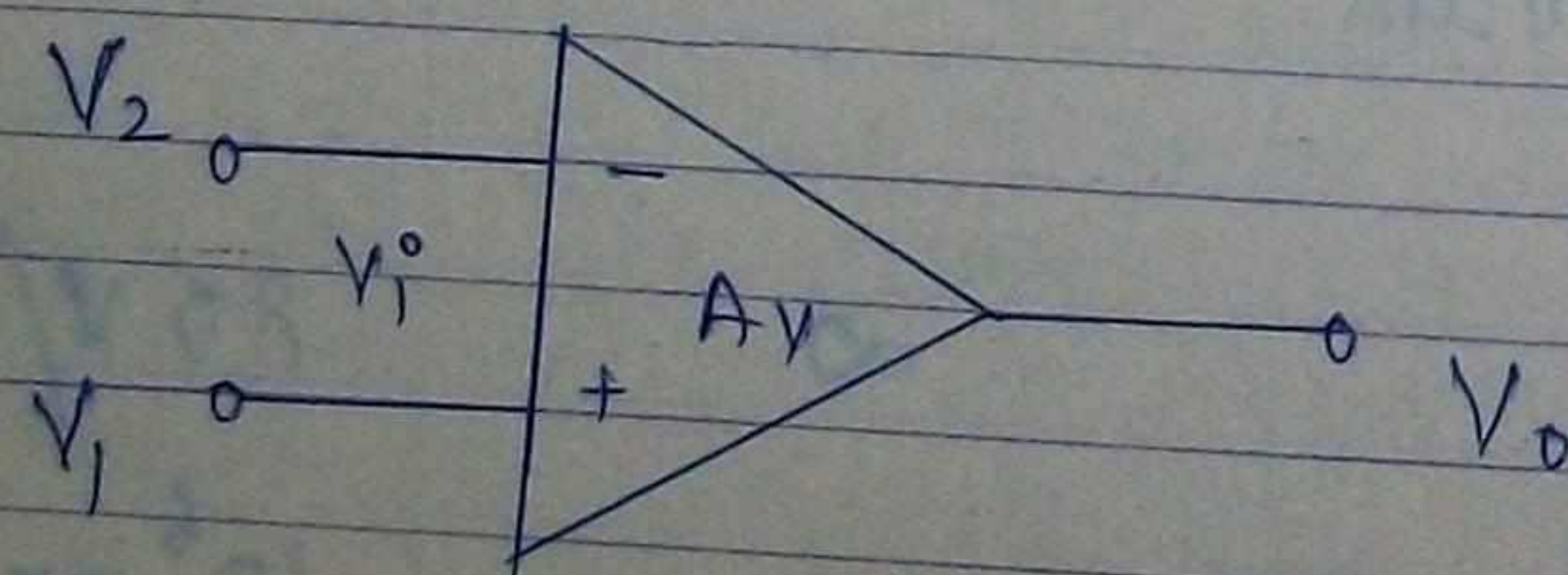
Capacitor coupled -



VIRTUAL GROUND CONDITION
⇒ PROPERTIES OF IDEAL OPAMP -

$$A_{OL} = A_v = \infty$$

①



$$X_C = \frac{1}{2\pi fC}$$

$$V_i = V_1 - V_2$$

$$V_o = A V_i$$

V_o = finite output

$$V_i = \frac{V_o}{A_v} = \frac{V_o}{\infty} = 0$$

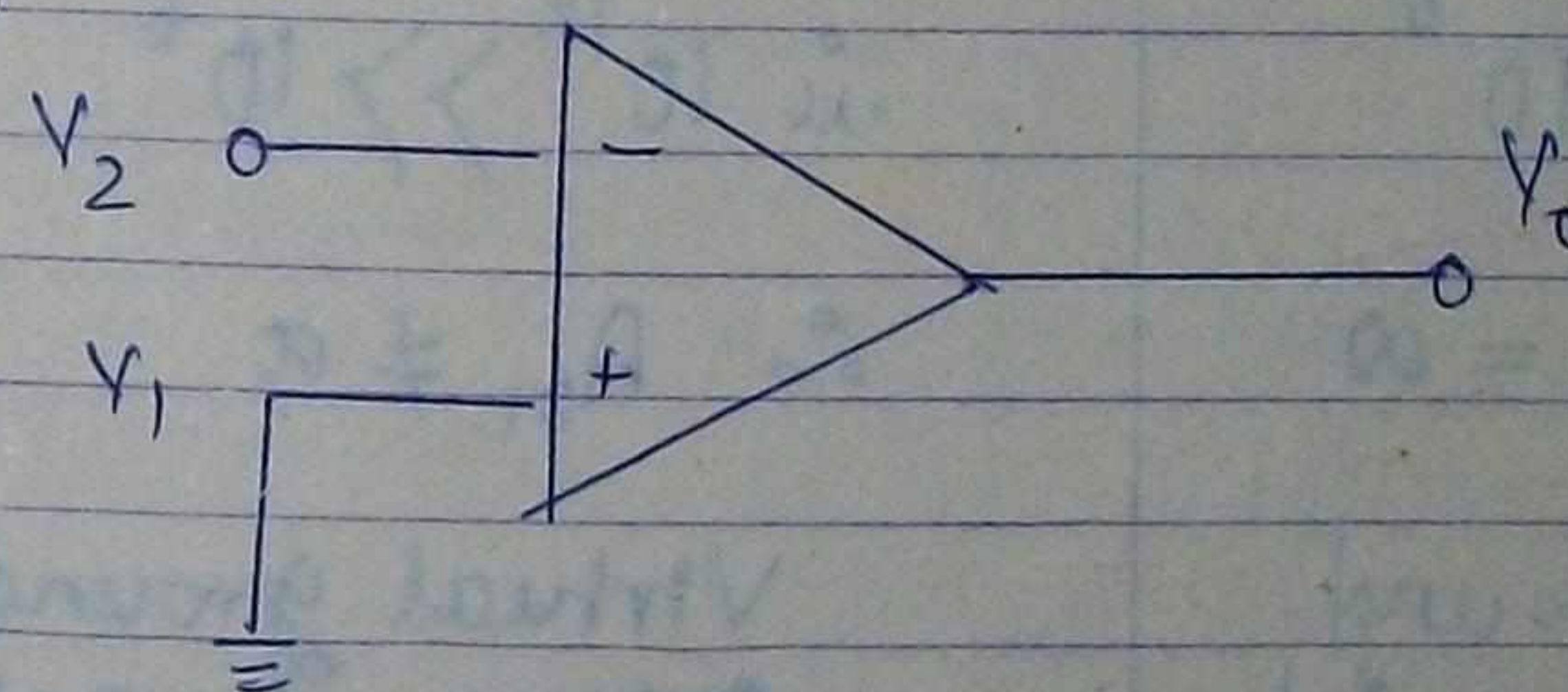
There is finite O/P without any I/P.

$$V_i = V_1 - V_2 = 0$$

∴

$$V_1 = V_2$$

Now,



V_1 = physically ground.

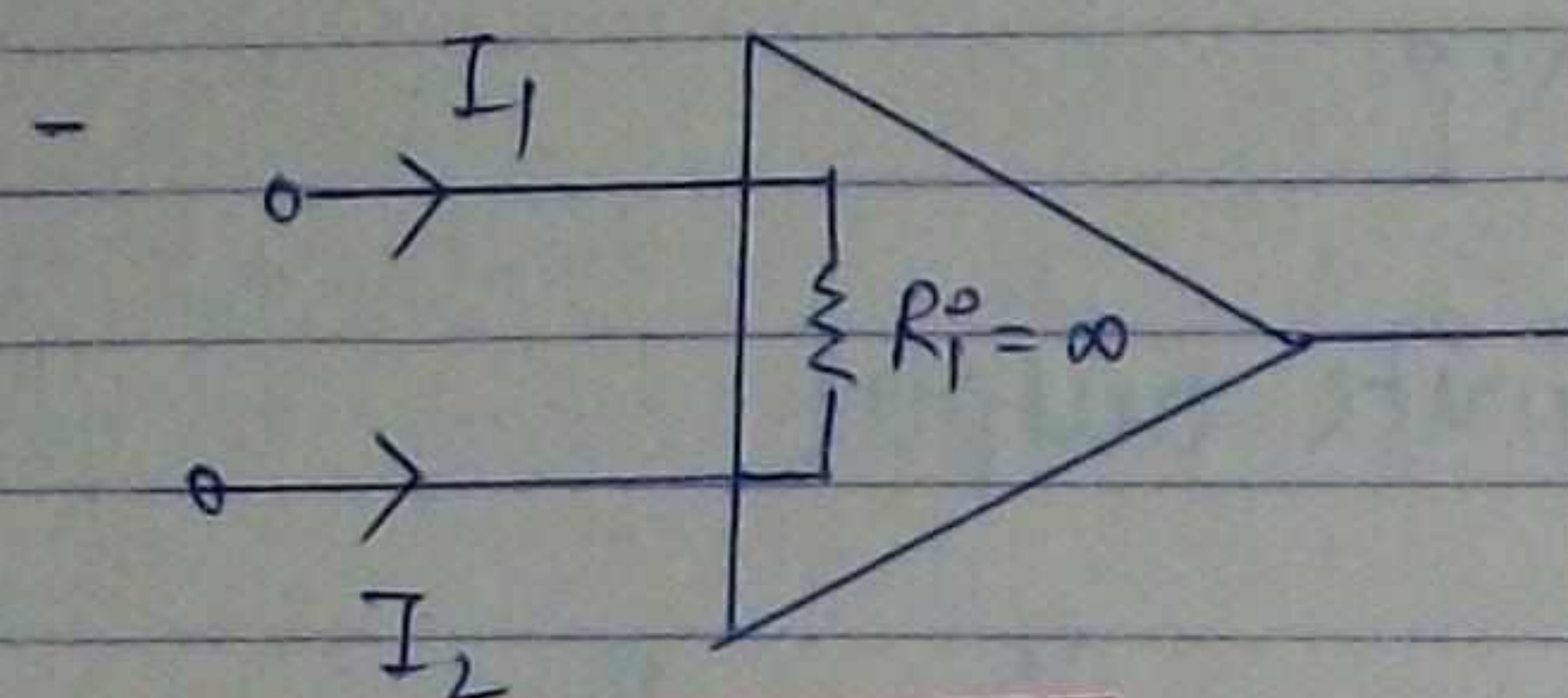
$$V_1 = 0$$

$$V_2 = V_1 = 0$$

[VIRTUAL
GROUND
PROCESS]

V_2 is virtually ground

② $R_1^o = \infty$



$$I_1 = I_2 = 0$$

NEGATIVE
FEEDBACK

$$|A_{cl}| \ll |A_{ol}|$$

$$\approx 10 \ll 10^6$$

$$\therefore \text{let } A_{ol} = \infty$$

Virtual ground
principle is valid

$$V_1 = V_2$$

POSITIVE
FEEDBACK

$$|A_{cl}| \gg |A_{ol}|$$

$$\approx 10^{12} \gg 10^6$$

$$\therefore A_{ol} \neq \infty$$

Virtual ground
principle is not
valid.

$$V_1 \neq V_2$$

$\Rightarrow$ MODE OF OPERATION -

1. INVERTING MODE -

phase shift, $\phi = 180^\circ$

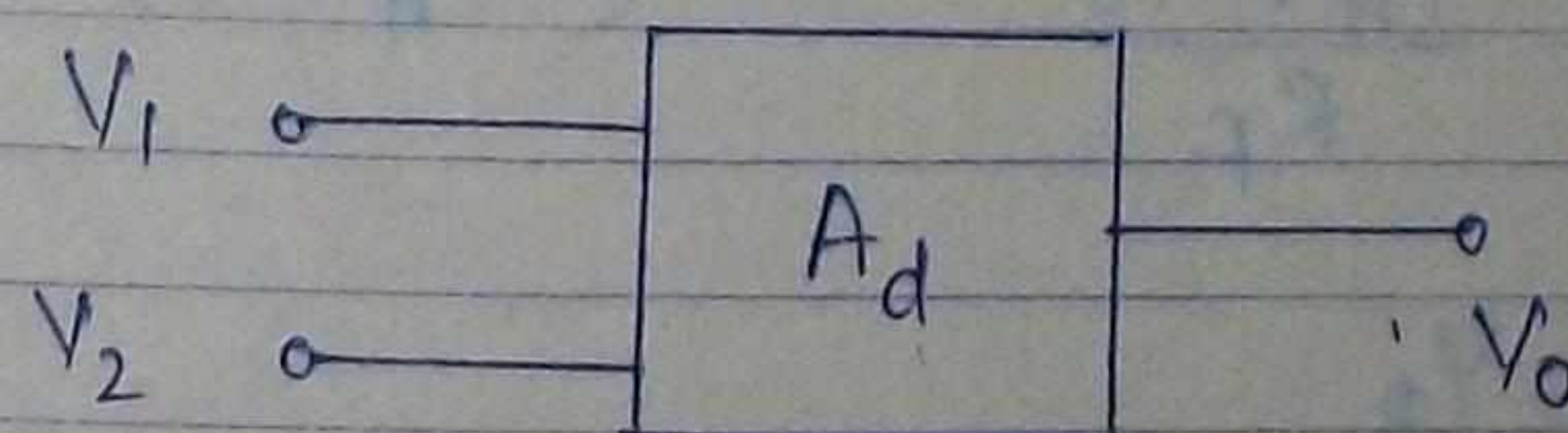
Output is 180° out of phase with i/p.

2. NON - INVERTING MODE -

phase shift, $\phi = 0, 2\pi, \dots, 2n\pi$
 $n = 0, 1, 2, \dots$

Output is in phase with i/p.

3. DIFFERENTIAL MODE -



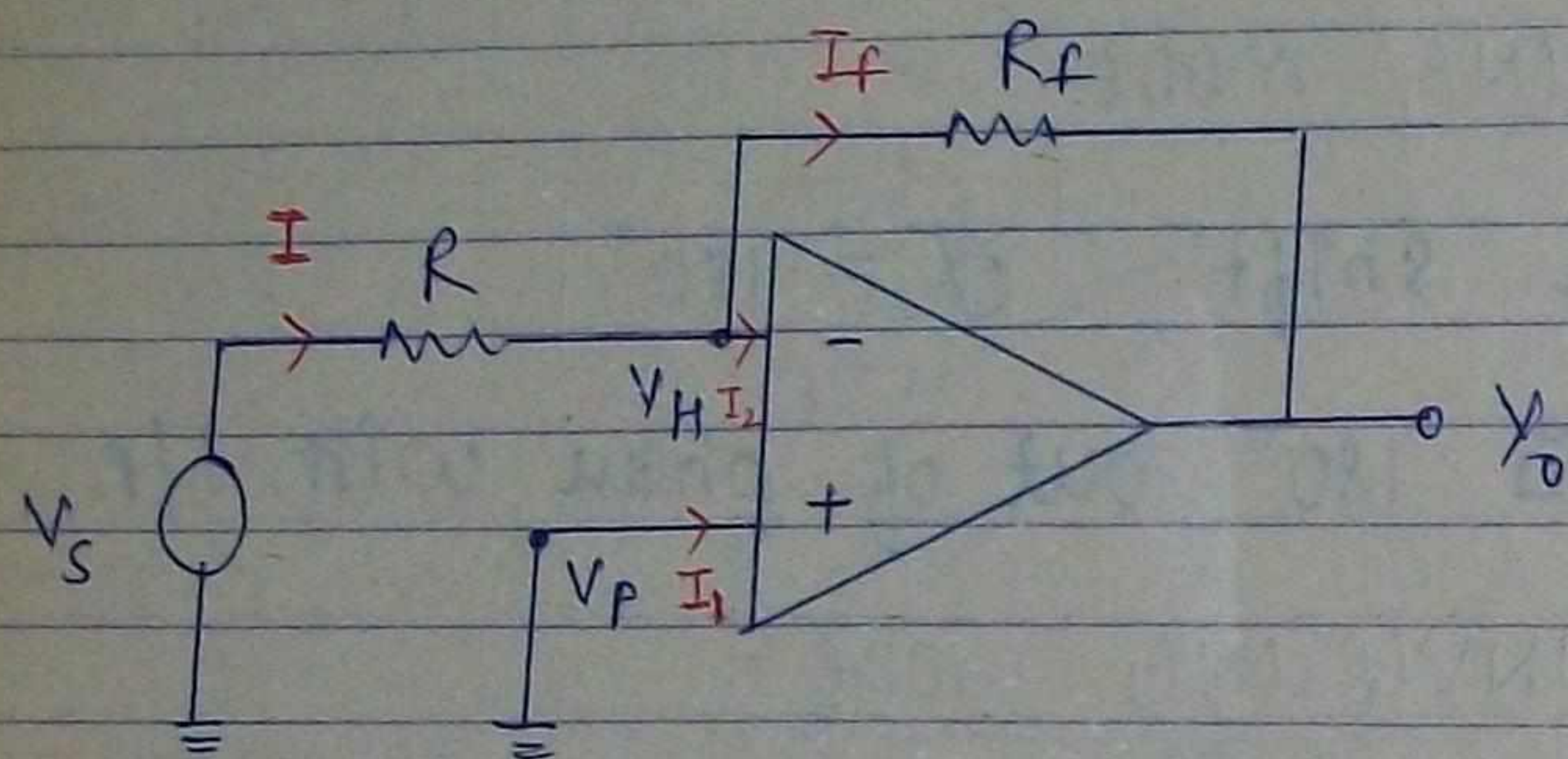
$$V_0 = A_d [V_1 - V_2] = A_d V_d$$

where,

A_d = difference gain

$$V_d = V_1 - V_2$$

① INVERTING OPERATIONAL AMPLIFIER -



$$A_{OL} = \infty, \quad V_P = V_H = 0 \text{ (VGP)}$$

$$I_1 = I_2 = 0 \quad [R_i \approx \infty]$$

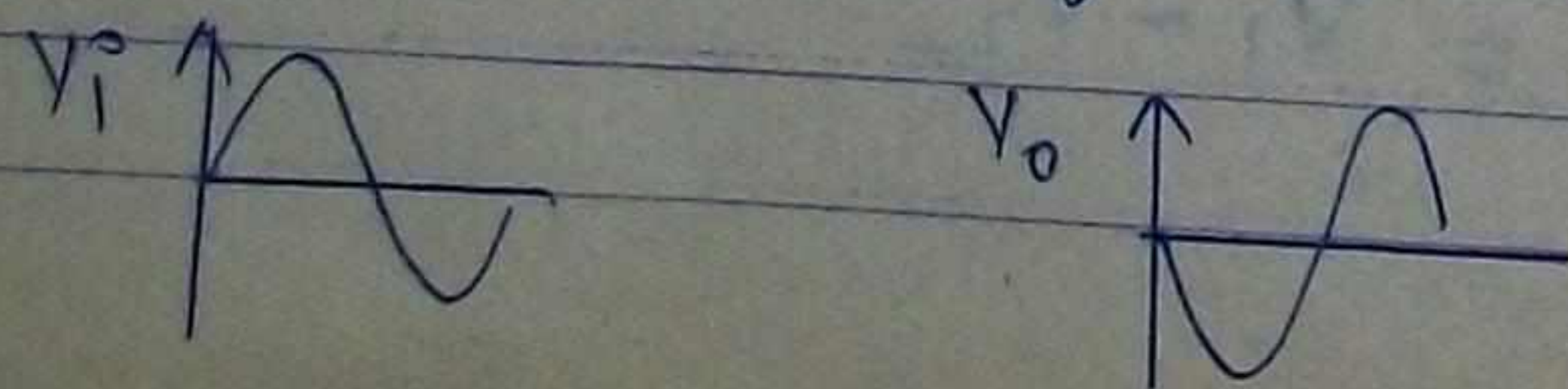
$$I = I_f$$

$$\frac{V_s - V_H}{R} = \frac{V_H - V_o}{R_f} \quad (V_H = 0)$$

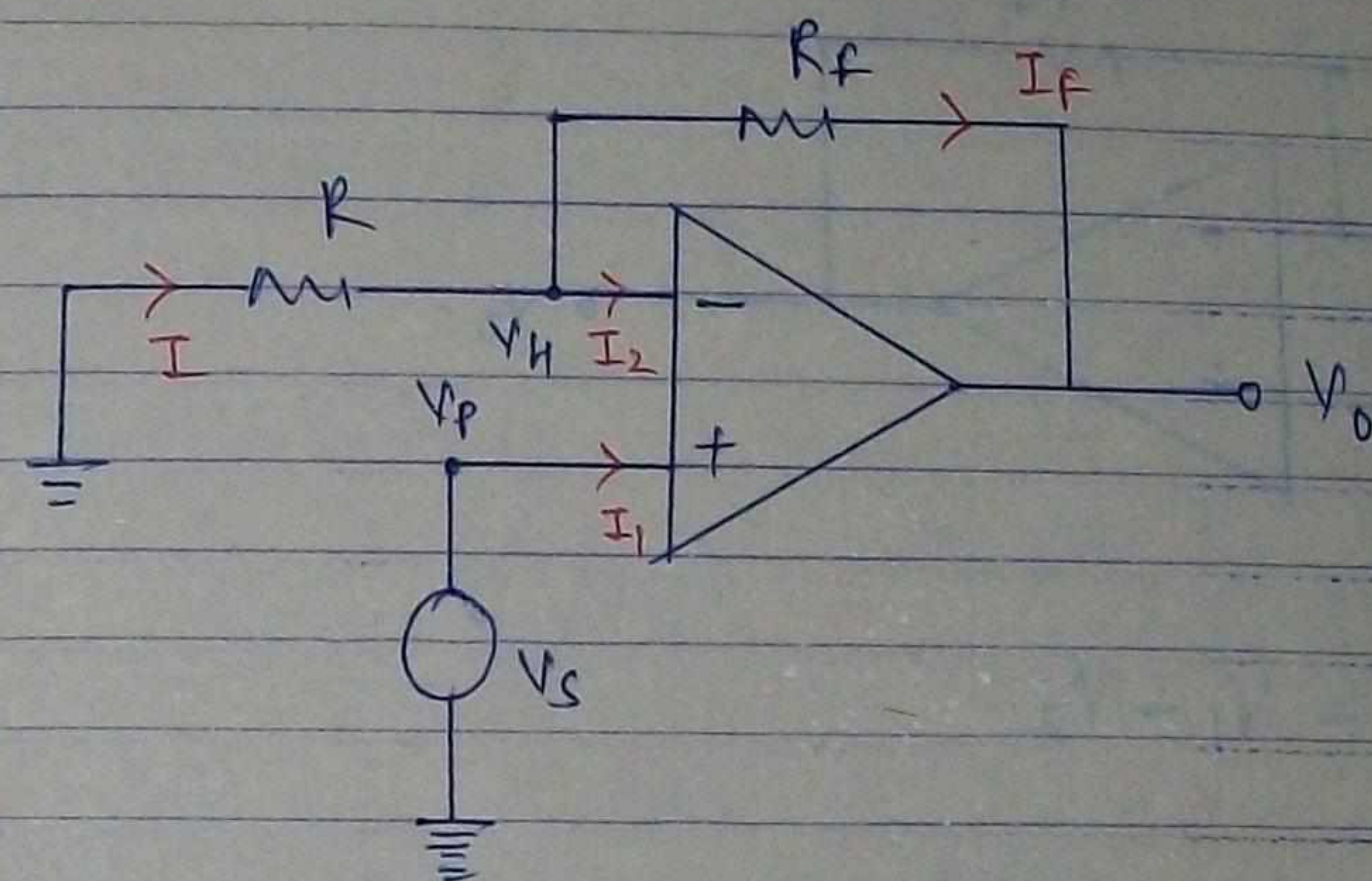
$$\frac{V_s}{R} = -\frac{V_o}{R_f}$$

$$A_{CL} = \frac{V_o}{V_s} = -\frac{R_f}{R}$$

$$[\phi] \text{ phase shift} = 180^\circ$$



② NON-INVERTING OPAMP -



$$V_P = V_H = V_s \text{ (virtual ground)}$$

$$I_1 = I_2 = 0 \quad (R_i \approx \infty)$$

$$I_f = I$$

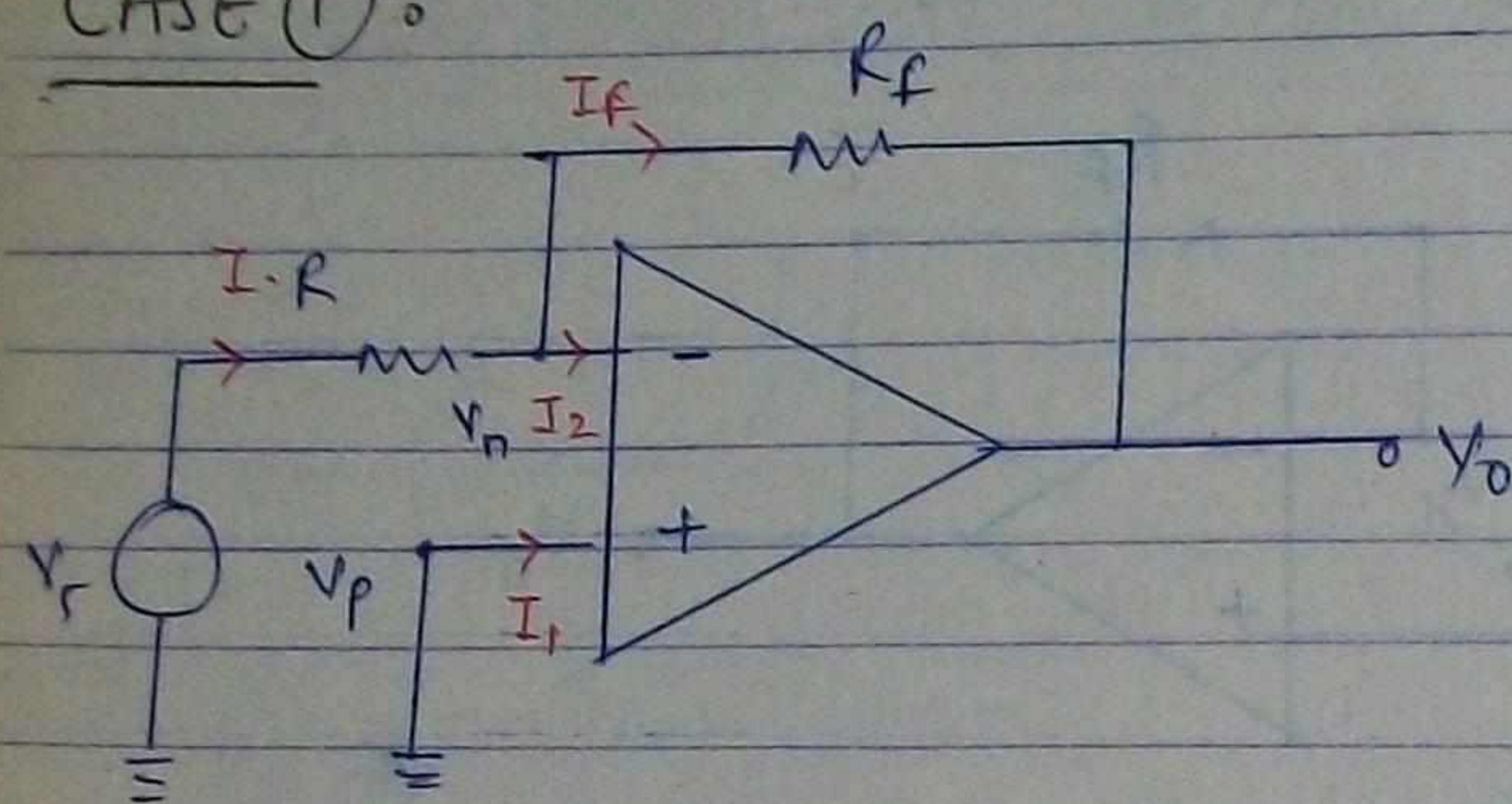
$$\frac{V_H - V_o}{R_f} = \frac{0 - V_H}{R}$$

$$\frac{V_s - V_o}{R_f} = -\frac{V_s}{R}$$

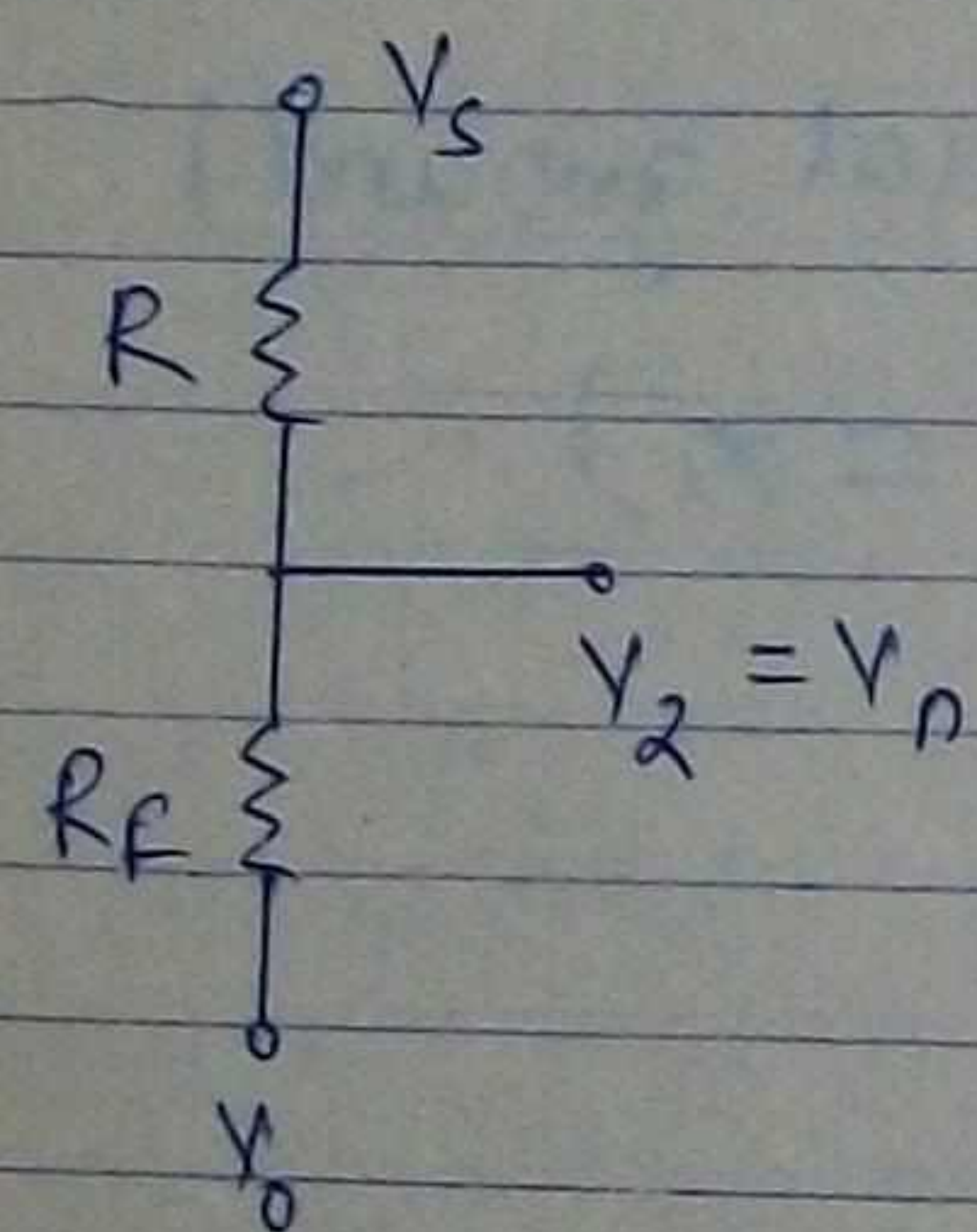
$$A_{CL} = \frac{V_o}{V_s} = \left[1 + \frac{R_f}{R} \right]$$

$$[\phi] \text{ phase shift} = 0^\circ$$

CASE (1) :



$$V_i = V_1 - V_2$$



$$V_2 = \frac{R_f \cdot V_s}{R + R_f} + \frac{V_o R}{R + R_f}$$

$$V_o \uparrow \quad V_2 \uparrow \quad V_i \downarrow$$

$$V_o \downarrow \quad V_2 \downarrow \quad V_i \uparrow$$

∴ -ve feedback.

$$V_o = A_{OL} V_i = A_{OL} [V_1 - V_2]$$

$$V_o = A_{OL} \left[0 - \left(\frac{V_s R_f}{R + R_f} + \frac{V_o R}{R + R_f} \right) \right]$$

$$V_o \left[1 + \frac{A_{OL} R}{R + R_f} \right] = - \frac{A_{OL} V_s R_f}{R + R_f}$$

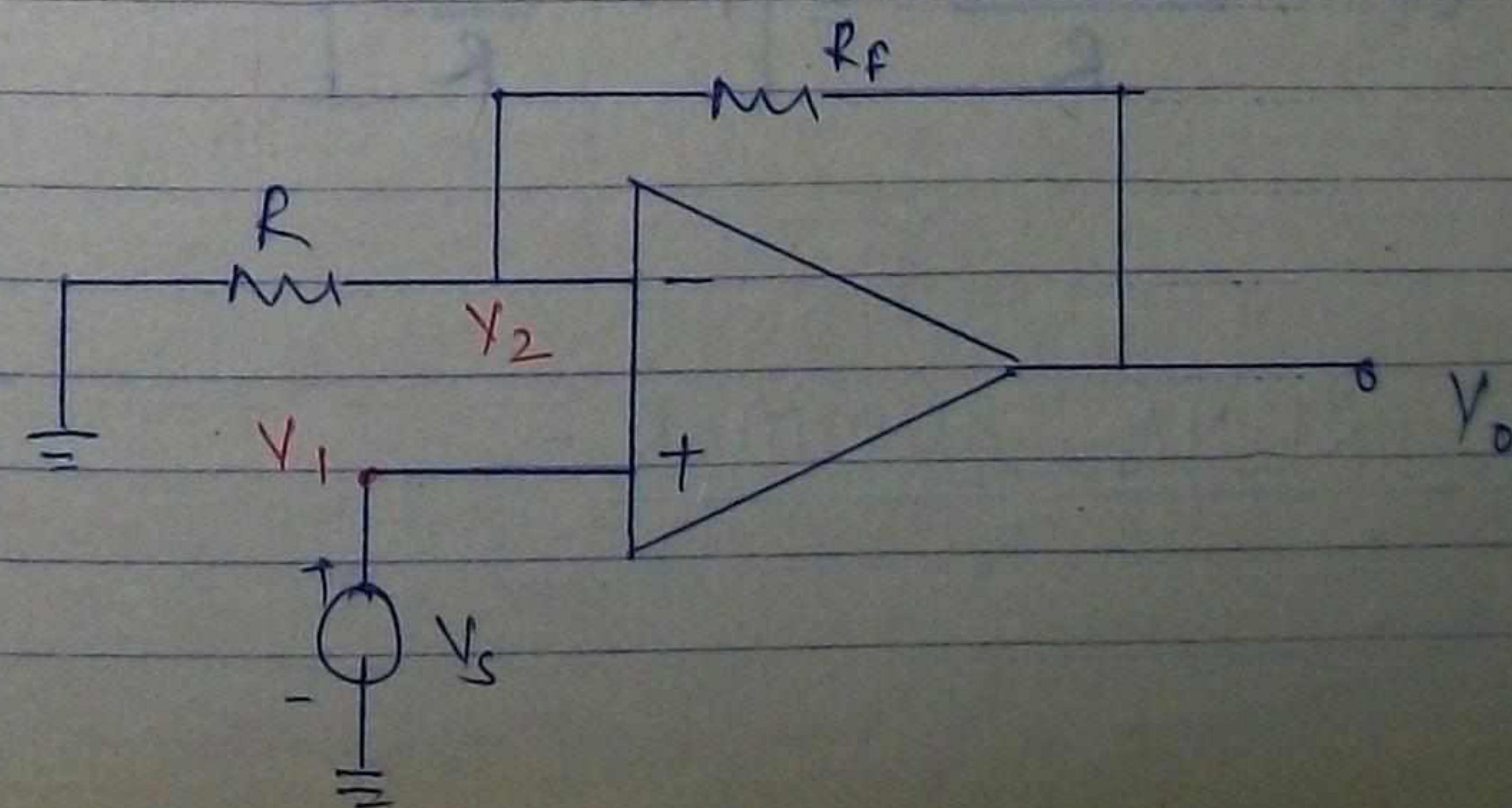
$$\textcircled{A} \quad \frac{V_o}{V_s} = A_{CL} = \frac{-A_{OL} R_f}{R_f + (1 + A_{OL}) R}$$

when, $A_{OL} = \text{very high}$

$$A_{CL} = \lim_{A_{OL} \rightarrow \infty} \frac{-A_{OL} R_f}{R_f + (1 + A_{OL}) R}$$

$$\textcircled{B} \quad A_{CL} = -\frac{R_f}{R}$$

CASE (2) :



$$V_o = A_{OL} = A_{OL} [V_1 - V_2]$$

$$V_1 = V_s$$

$$V_2 = \frac{V_o R}{R + R_f}$$

$$V_o = A_{OL} \left[V_s + \frac{V_o R}{R + R_f} \right]$$

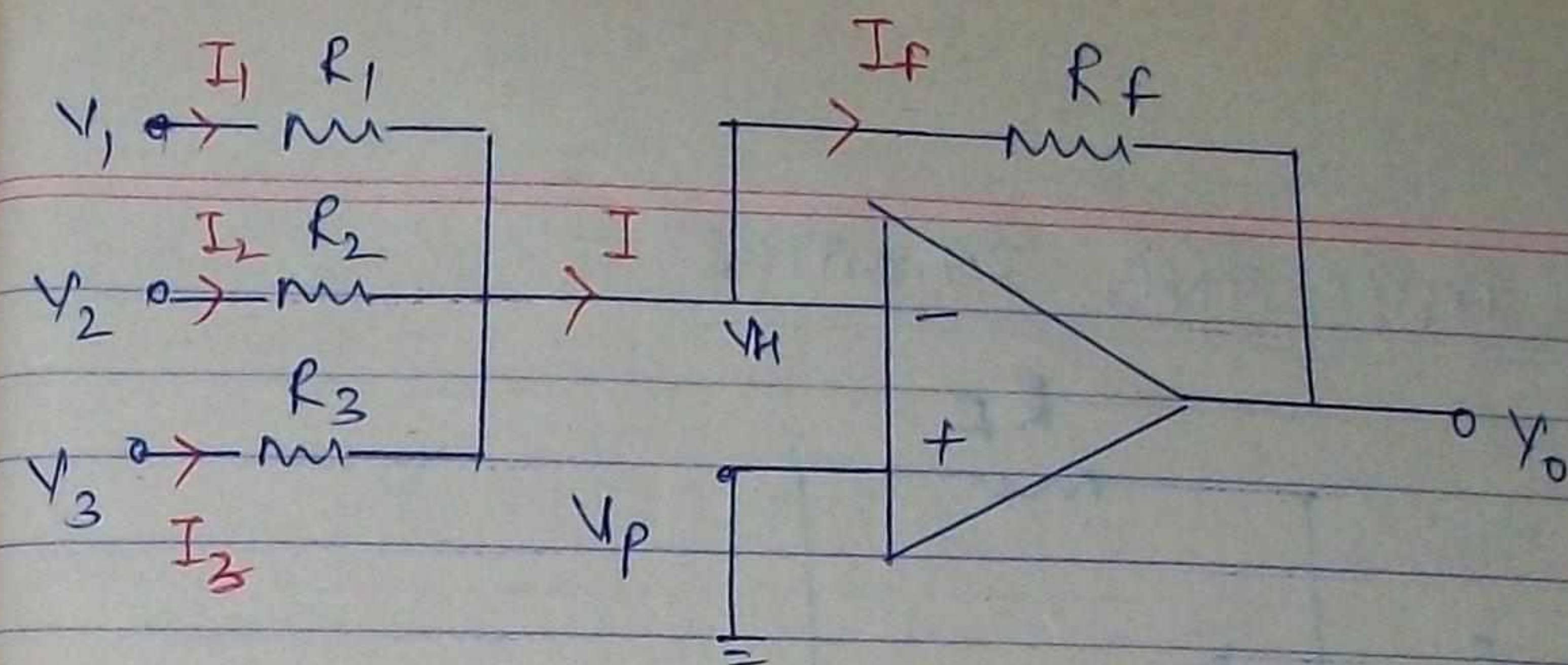
$$V_o \left[1 + \frac{R A_{OL}}{R + R_f} \right] = A_{OL} V_s$$

$$\textcircled{A} \quad \frac{V_o}{V_s} = A_{CL} = \frac{A_{OL} (R + R_f)}{R_f + (1 + A_{OL}) R}$$

$$\text{Since, } A_{OL} \gg 1 \quad A_{OL} R \gg R_f$$

$$\textcircled{A} \quad A_{CL} = \frac{R + R_f}{R} = \left[1 + \frac{R_f}{R} \right]$$

$\Rightarrow$ INVERTING SUMMER -



$$V_p = V_n = 0 \quad (\text{Virtual ground})$$

$$I_f = I = I_1 + I_2 + I_3$$

$$\frac{0 - V_o}{R_f} = \frac{V_1}{R_1} + \frac{V_2}{R_2} + \frac{V_3}{R_3}$$

$$\textcircled{A} \quad V_o = -R_f \left[\frac{V_1}{R_1} + \frac{V_2}{R_2} + \frac{V_3}{R_3} \right]$$

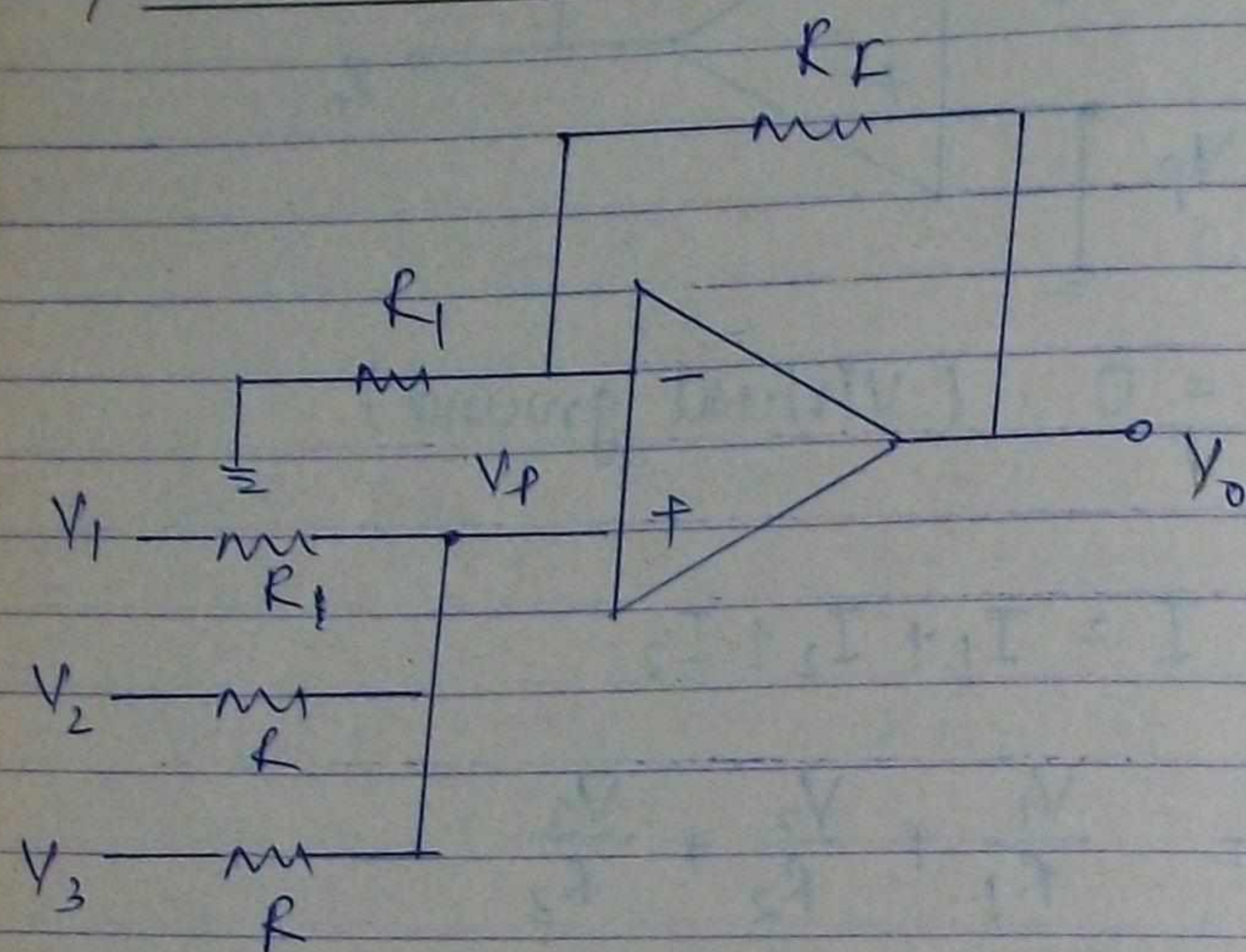
$$\text{for, } R_f = R_1 = R_2 = R_3 = R$$

$$V_o = -R \left[\frac{V_1}{R} + \frac{V_2}{R} + \frac{V_3}{R} \right]$$

$$\textcircled{A} \quad V_o = -[V_1 + V_2 + V_3]$$

$$\phi = 180^\circ$$

⇒ NON INVERTING SUMMER -



$$\frac{V_p - V_1}{R} + \frac{V_p - V_2}{R} + \frac{V_p - V_3}{R} = 0$$

$$3V_p = V_1 + V_2 + V_3$$

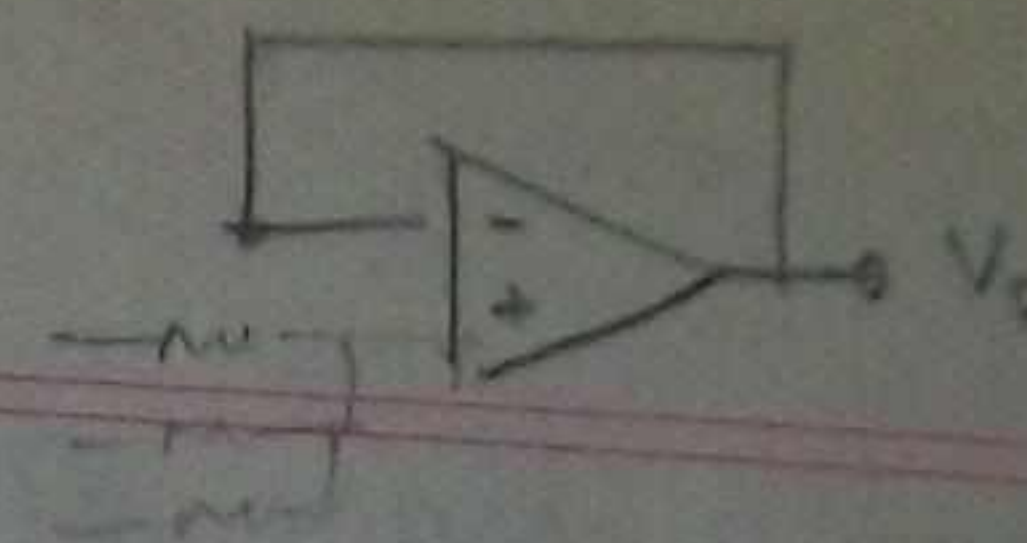
$$V_p = \frac{V_1 + V_2 + V_3}{3}$$

Since $V_o = \left[1 + \frac{R_F}{R_1} \right] V_p$ (NON INV. GAIN)

$$V_o = \left[1 + \frac{R_F}{R_1} \right] \left[\frac{V_1 + V_2 + V_3}{3} \right]$$

$$R_F = 2R_1$$

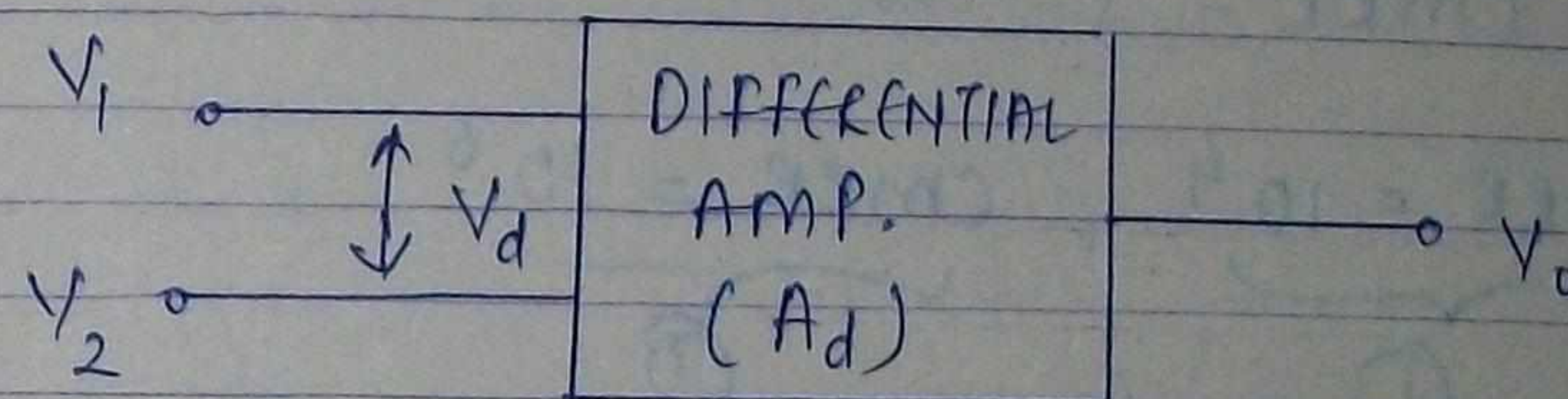
averager



let, $R_F = 2R_1$

$$V_o = V_1 + V_2 + V_3, \quad \phi_{\text{SHIFT}} = 0.$$

⇒ DIFFERENTIAL AMPLIFIER -



A_d = differential gain

$$V_d = V_1 - V_2$$

$$V_o = A_d V_d \quad (\text{ideal cond}^n)$$

$$V_o = A_d V_d + A_c V_c \quad (\text{pratically})$$

where, A_c = common mode gain

$$V_c = \frac{V_1 + V_2}{2} \quad \text{ideally } A_c = 0$$

COMMON MODE REJECTION RATIO -

$$CMRR = e = \frac{|A_d|}{|A_c|}$$

Ideally $A_c = 0$,

$$\therefore CMRR = e = \infty$$

Ex - $CMRR = 10^4$ (1), $CMRR = 10^6$ (2)

Amp (2) is better than (1).

- CMRR is called 'Figure of merit' of practical OPAMP.

- CMRR should be high

$$CMRR = \infty \quad (\text{Ideally})$$

$$CMRR = 10^6 \quad (\text{practically})$$

In db,

$$CMRR = 20 \log \frac{|A_d|}{|A_c|}$$

$$CMRR(\text{db}) = 20 \log 10^6$$

$$\Rightarrow CMRR(\text{db}) = 120 \text{ db.}$$

$\Rightarrow$ TO CALCULATE CMRR -

① CALCULATION FOR A_c -

$$V_1 = V_s, \quad V_2 = V_s$$

$$V_d = 0, \quad V_c = V_s$$

$$V_o = A_c V_s$$

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

② CALCULATION FOR A_d -

$$V_1 = V_s/2, \quad V_2 = -V_s/2$$

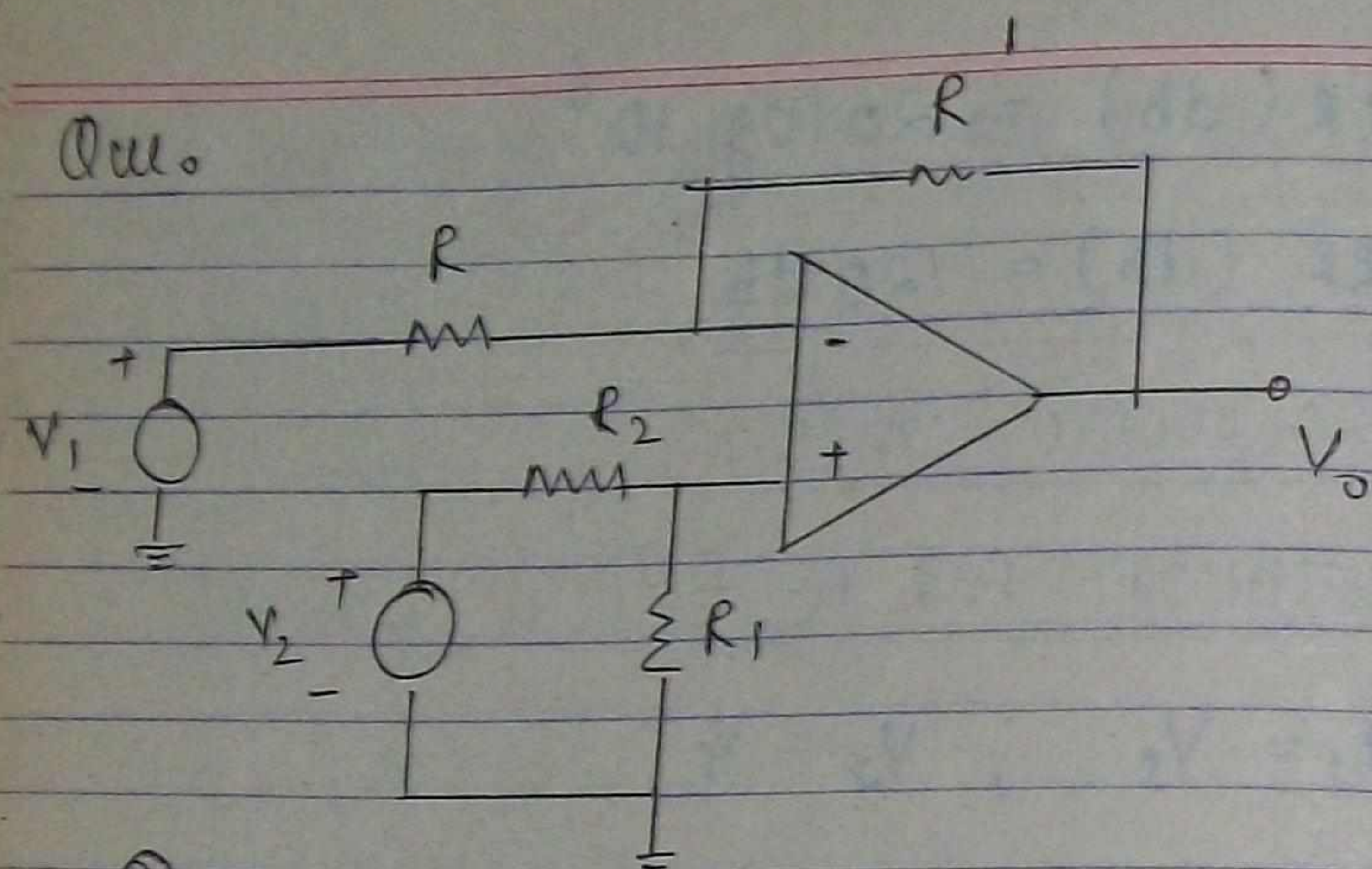
$$V_d = V_s, \quad V_c = 0$$

$$V_o = \frac{V_o}{V_s} = A_d V_s + A_c \times 0$$

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

$$\therefore CMRR = \frac{|A_d|}{|A_c|}$$

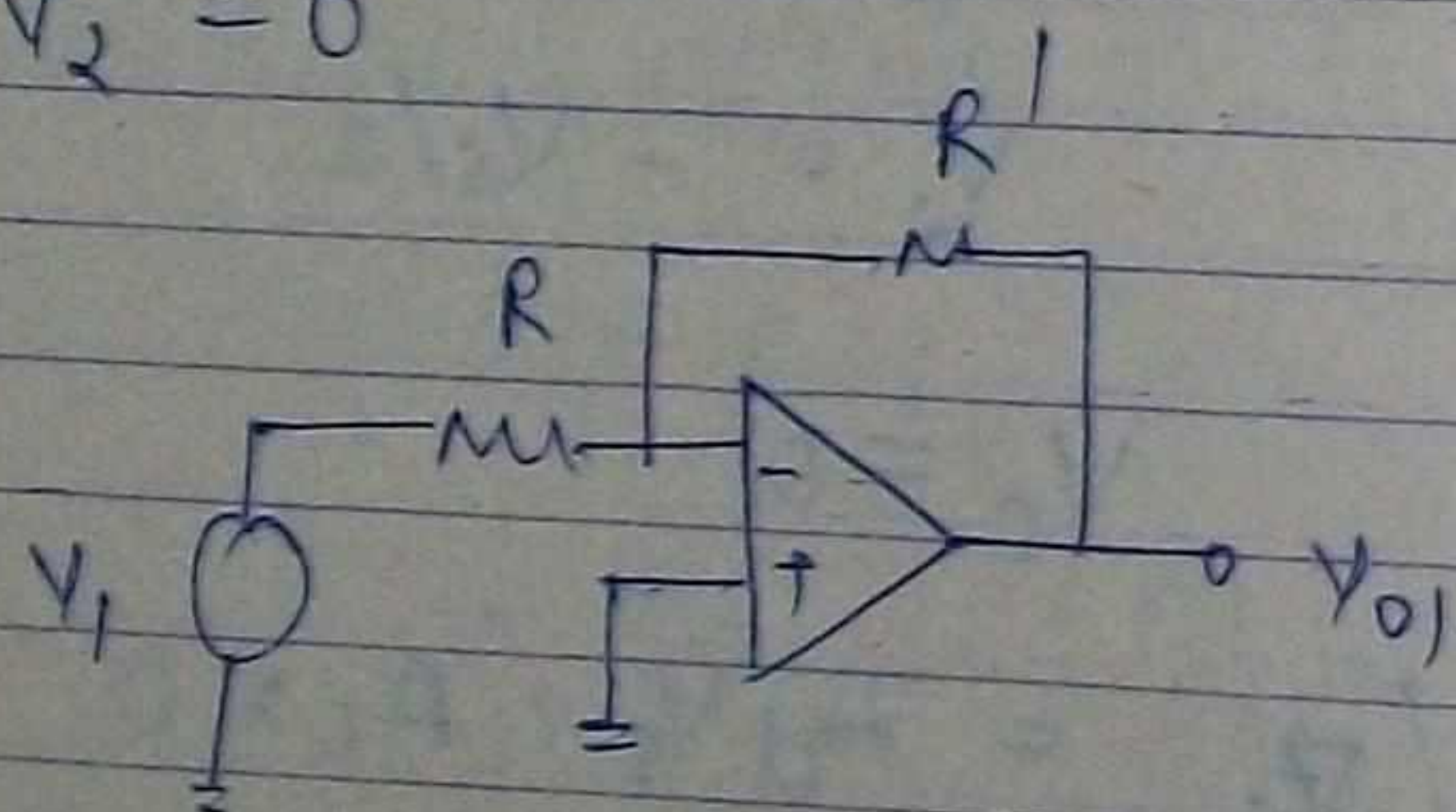
Ques.



- ① find V_0
- ② Calculate CMRR
- ③ Show that if $\frac{R'}{R} = \frac{R_1}{R_2}$ then CMRR = ∞

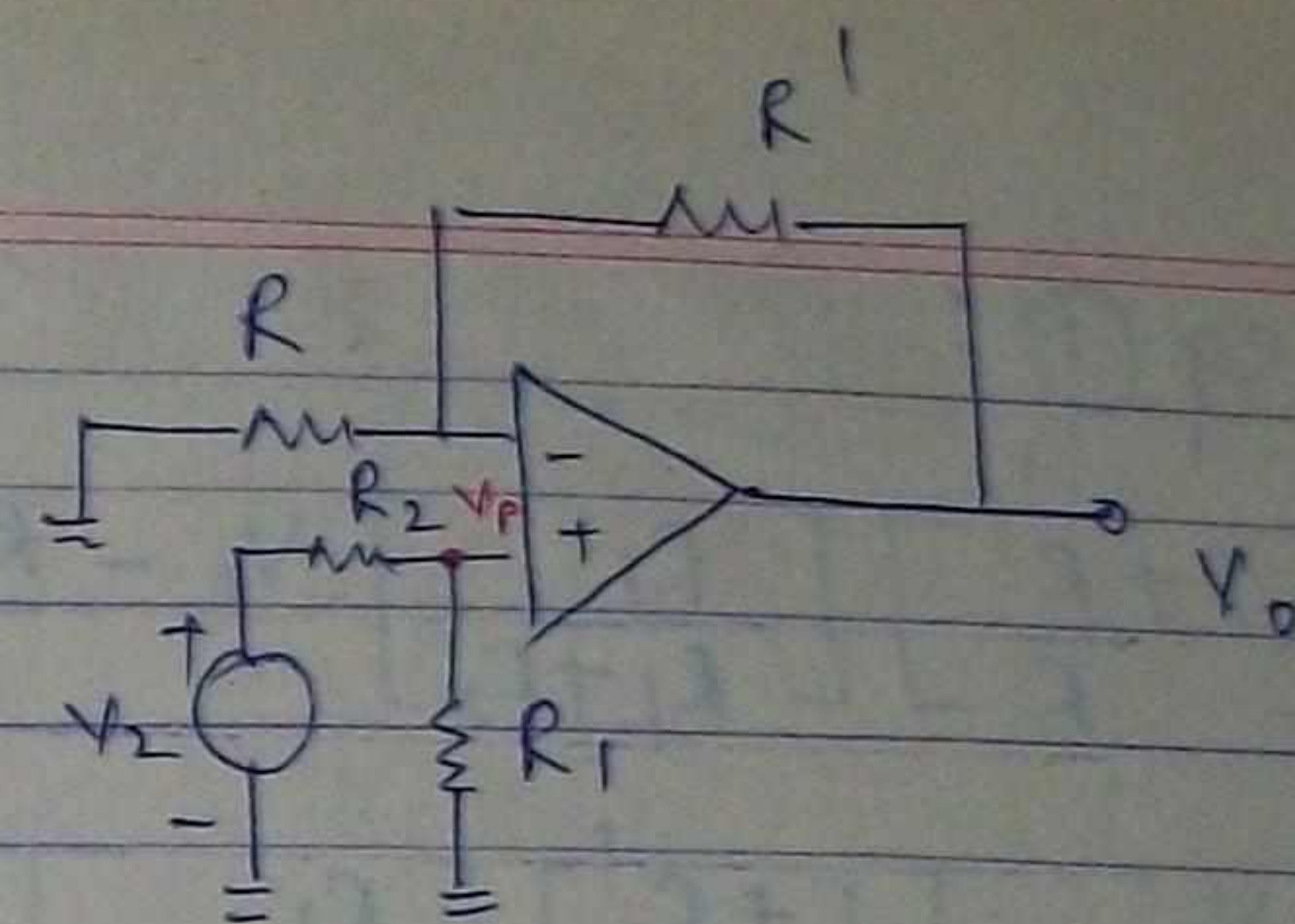
Solⁿ: ① Applying super position.

① $V_2 = 0$



$$V_{01} = -\frac{R'}{R} V_1$$

② $V_1 = 0$



$$V_0 = \left[1 + \frac{R'}{R} \right] V_p$$

where, $V_p = \frac{V_2 R_1}{R_1 + R_2}$

$$V_0 = V_{02} + V_{01}$$

$$V_0 = \left[1 + \frac{R'}{R} \right] \left[\frac{R_1}{R_1 + R_2} \right] V_2 - \frac{R'}{R} V_1 \quad \text{--- ①}$$

$$V_0 = A_d V_d + A_c V_c$$

$$\text{② CMRR} = \frac{|A_d|}{|A_c|}$$

$$V_1 = V_2 = V_s, \quad V_d = 0, \quad V_c = V_s$$

$$V_0 = A_d \times 0 + A_c V_s$$

$$A_c = V_0 / V_s$$

from eq ①:

$$V_o = \left[1 + \frac{R'}{R}\right] \left[\frac{R_1}{R_1 + R_2}\right] V_s - \frac{R'}{R} V_s$$

$$A_c = \frac{V_o}{V_s} = \left[1 + \frac{R'}{R}\right] \left[\frac{R_1}{R_1 + R_2}\right] - \frac{R'}{R}$$

for A_d ,

$$V_1 = V_s/2 \quad V_2 = -V_s/2$$

from eq ②:

$$V_o = - \left[1 + \frac{R'}{R}\right] \left[\frac{R_1}{R_1 + R_2}\right] \frac{V_s}{2} - \frac{R'}{R} \left(\frac{V_s}{2}\right)$$

$$\frac{V_o}{V_s} = A_d = -\frac{1}{2} \left[\left(1 + \frac{R'}{R}\right) \left(\frac{R_1}{R_1 + R_2}\right) + \frac{R'}{R} \right]$$

$$\therefore CMRR = \left| \frac{A_d}{A_c} \right|$$

$$\textcircled{3} \quad \text{If } \frac{R'}{R} = \frac{R_1}{R_2}$$

$$\text{for } CMRR = \infty, \quad A_c = 0$$

$$A_c = \left[1 + \frac{R'}{R}\right] \left[\frac{R_1}{R_1 + R_2}\right] - \frac{R'}{R} = 0$$

$$\left(\frac{R + R'}{R}\right) \left(\frac{R_1}{R_1 + R_2}\right) = \frac{R'}{R}$$

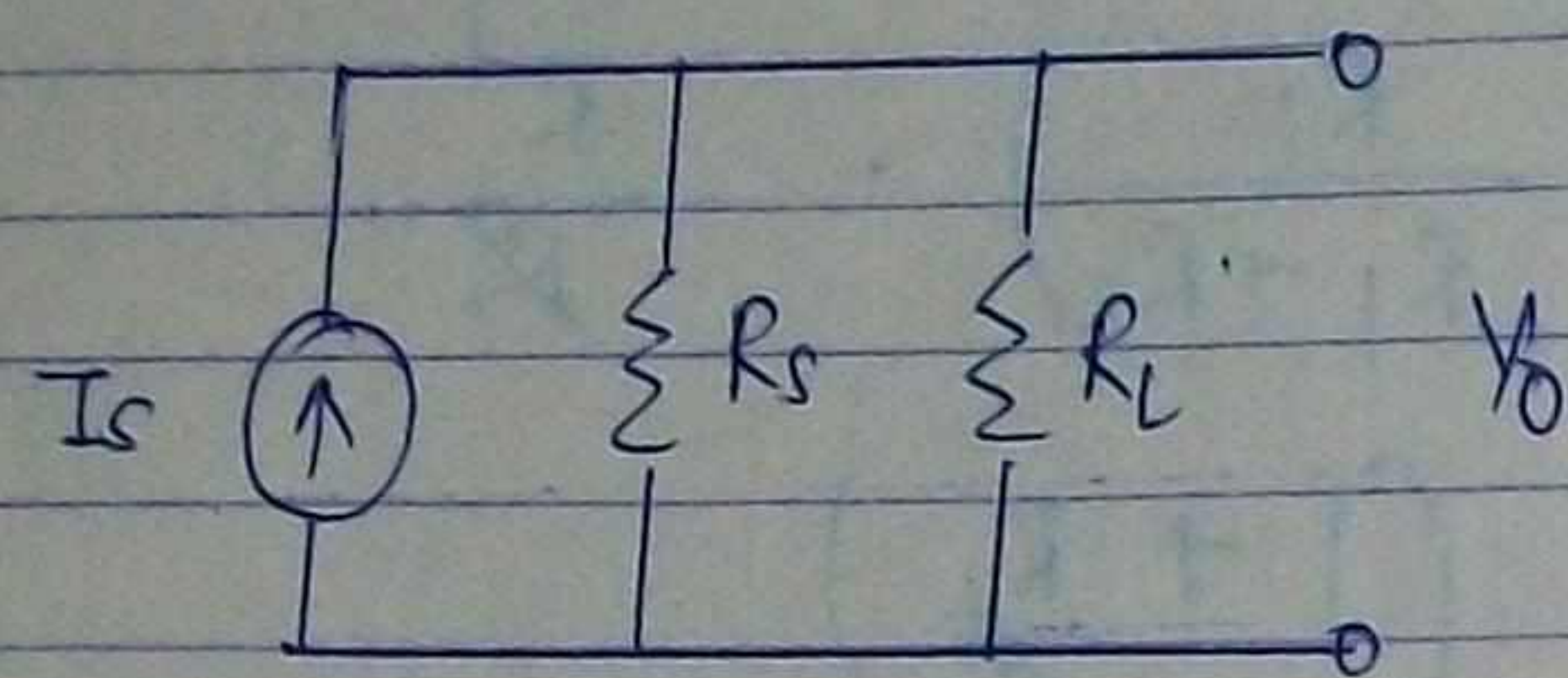
$$\left(\frac{R}{R'} + 1\right) = \left(1 + \frac{R_2}{R_1}\right)$$

$$\frac{R'}{R} = \frac{R_1}{R_2} \quad \text{Hence Proved.}$$

NOTE -

- In inverting OPAMP the feedback mechanism is 'Voltage shunt f.B'
- In non inverting OPAMP the feedback mechanism is 'voltage series f.B'

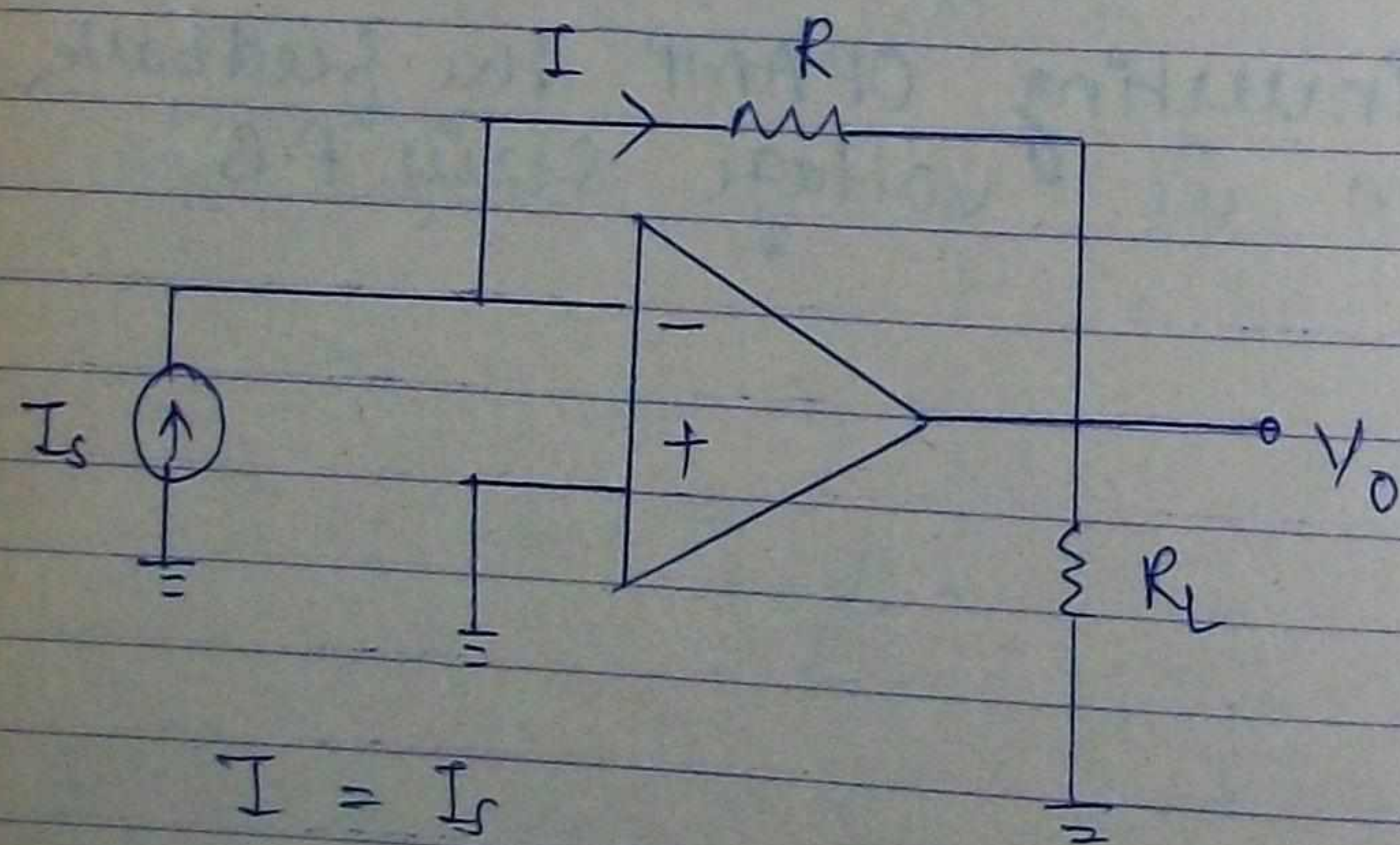
⇒ CURRENT TO VOLTAGE CONVERSION -



$$R_L' = \frac{R_s R_L}{R_s + R_L}$$

$$V_o = R_L' I_s$$

Here output voltage depends on R_L
 ∴ It is not a converter.



$$I = I_s$$

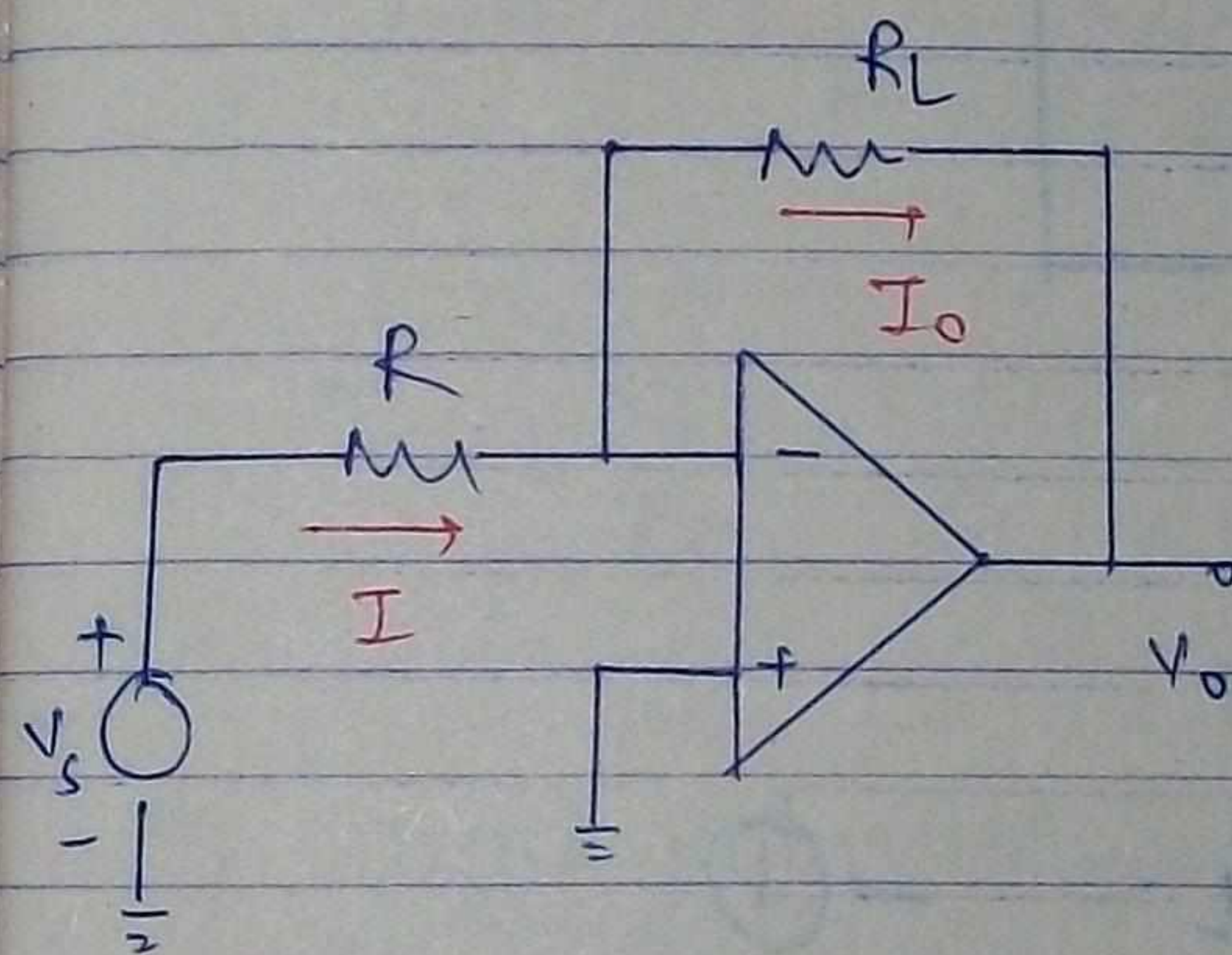
$$\frac{0 - V_o}{R} = I_s$$

$$V_o = -I_s R$$

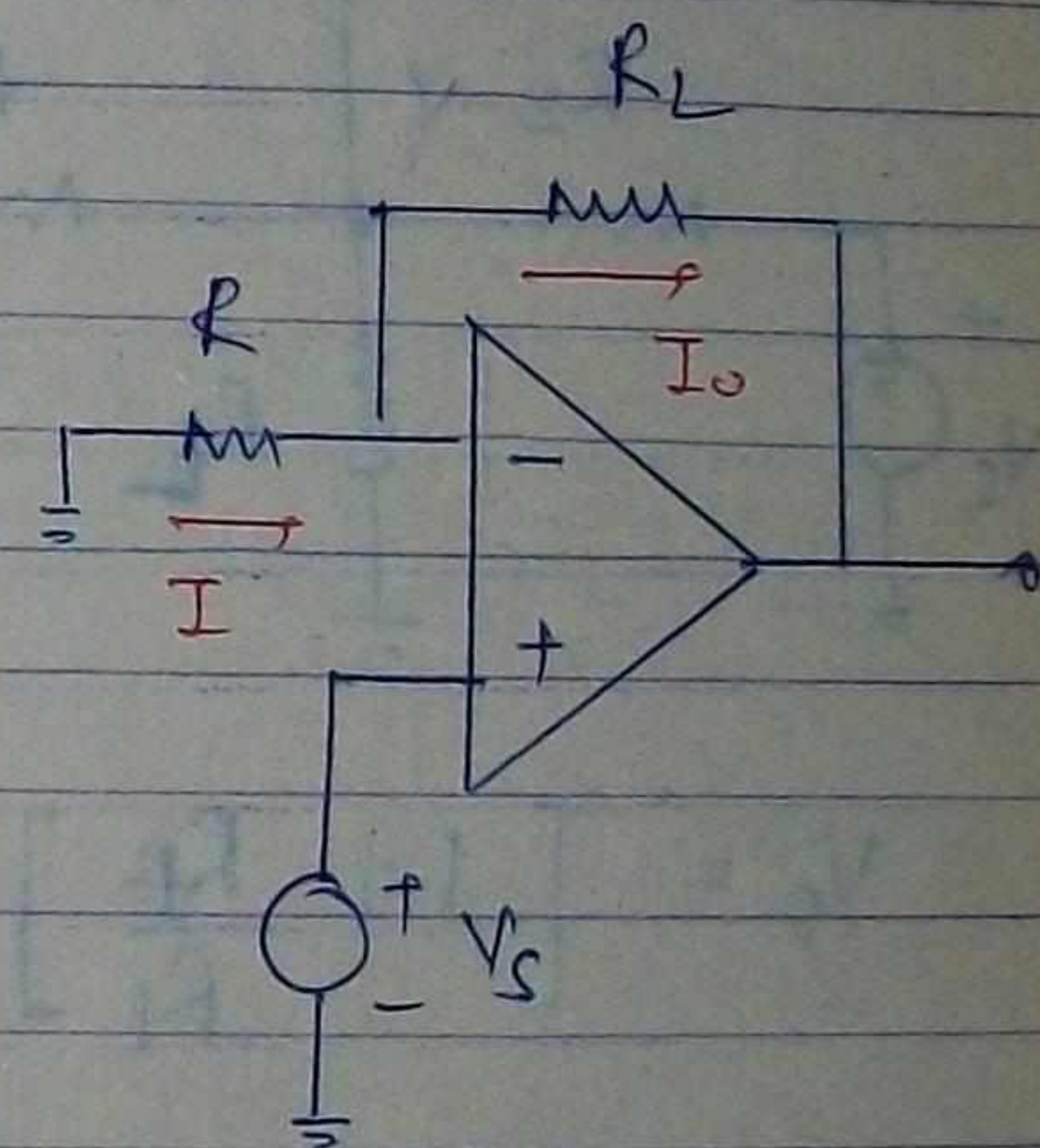
∴ V_o is independent of R_L .

⇒ VOLTAGE TO CURRENT CONVERSION -

① FLOATING LOAD -



$$I_o = I = \frac{V_s}{R}$$

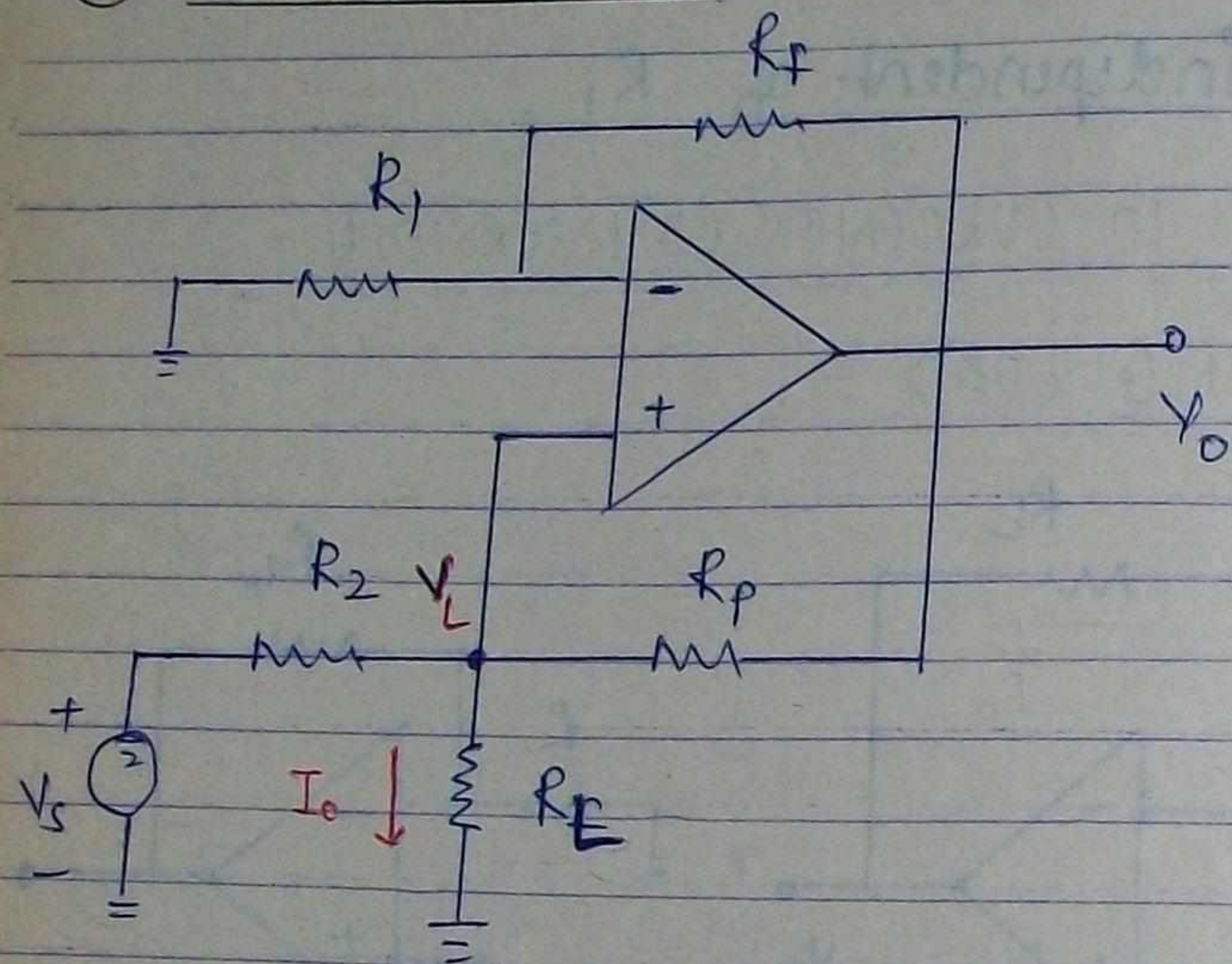


$$I_o = I = \frac{0 - V_s}{R}$$

$$I_o = -\frac{V_s}{R}$$

direction of current is opposite.

② GROUNDING LOAD -



$$V_0 = \left[1 + \frac{R_f}{R_1} \right] V_L \quad \text{--- (1)}$$

for V_L KCL,

$$\frac{V_L - V_s}{R_2} + \frac{V_L}{R_L} + \frac{V_L - V_0}{R_p} = 0$$

$$V_L \left[\frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_p} \right] = \frac{V_s}{R_2} + \frac{V_0}{R_p} \quad \text{--- (2)}$$

from (1) and (2):

$$V_L \left[\frac{1}{R_2} + \frac{1}{R_L} + \frac{1}{R_p} - \frac{1}{R_p} \left(1 + \frac{R_f}{R_1} \right) \right] = \frac{V_s}{R_2}$$

$$V_L \left[\frac{R_1 R_p R_L + R_1 R_2 R_p - R_f R_2 R_L}{R_1 R_2 R_p R_L} \right] = \frac{V_s}{R_2}$$

$$V_L = V_s \left[\frac{R_1 R_p R_L}{R_L (R_1 R_p - R_f R_2) + R_1 R_2 R_p} \right]$$

⊛

$$I_0 = \frac{V_L}{R_L} = \frac{V_s}{R_L} \left[\frac{R_1 R_p R_L}{R_L (R_1 R_p - R_f R_2) + R_1 R_2 R_p} \right]$$

for circuit to be independent of R_L

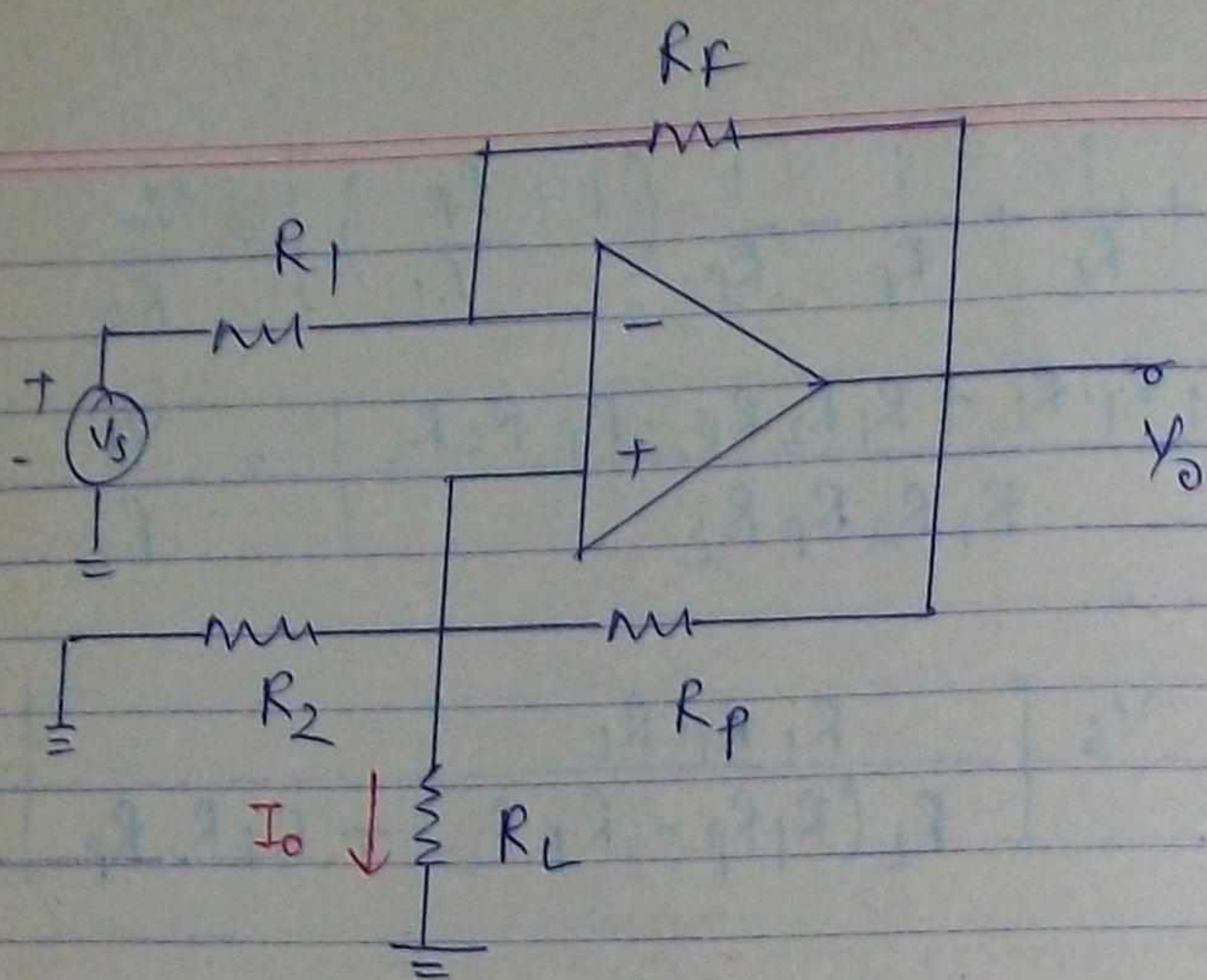
$$(R_1 R_p - R_f R_2) = 0$$

$$\boxed{R_1 R_p = R_f R_2} \quad \text{(balanced bridge condition)}$$

⊛ ∴

$$\boxed{I_0 = \frac{V_s}{R_2}}$$

FOR INVERTING TERMINAL - (DOUBT)



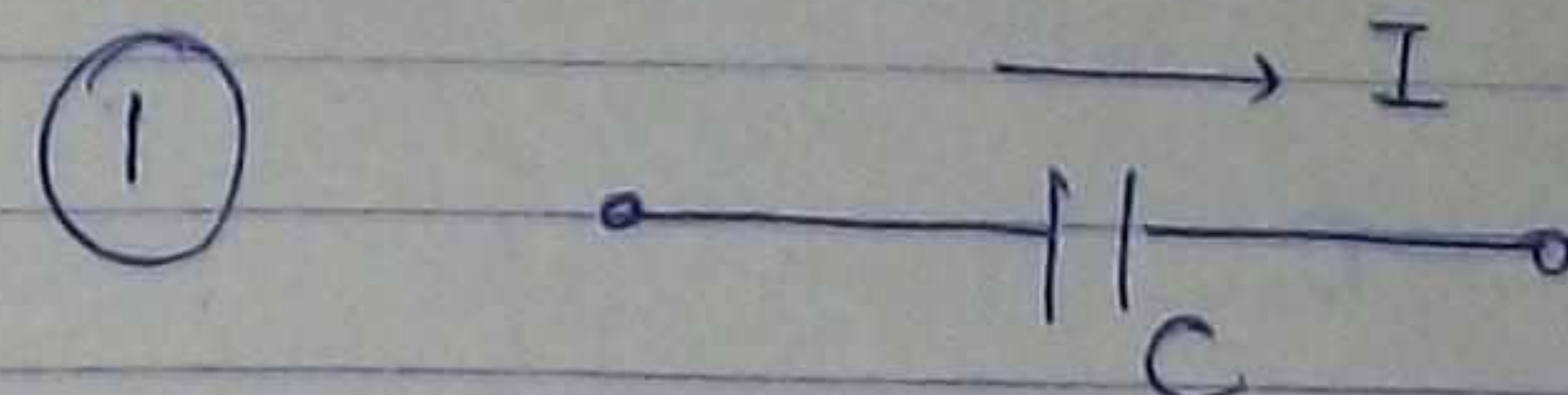
If $\frac{R_1}{R_2} = \frac{R_f}{R_p}$

then,

$$I_0 = -\frac{V_s}{R_2}$$

RC CIRCUITS

⇒ NETWORK BASIC -



$$V_c = \frac{1}{C} \int_0^t I(t) dt, \quad I(t) = \text{finite}$$

$$\int_{0^-}^{0^+} I(t) dt = q = 0$$

∴ Capacitor does not allow sudden change of voltage for finite current.

$$V_c(0^-) = V_c(0^+)$$

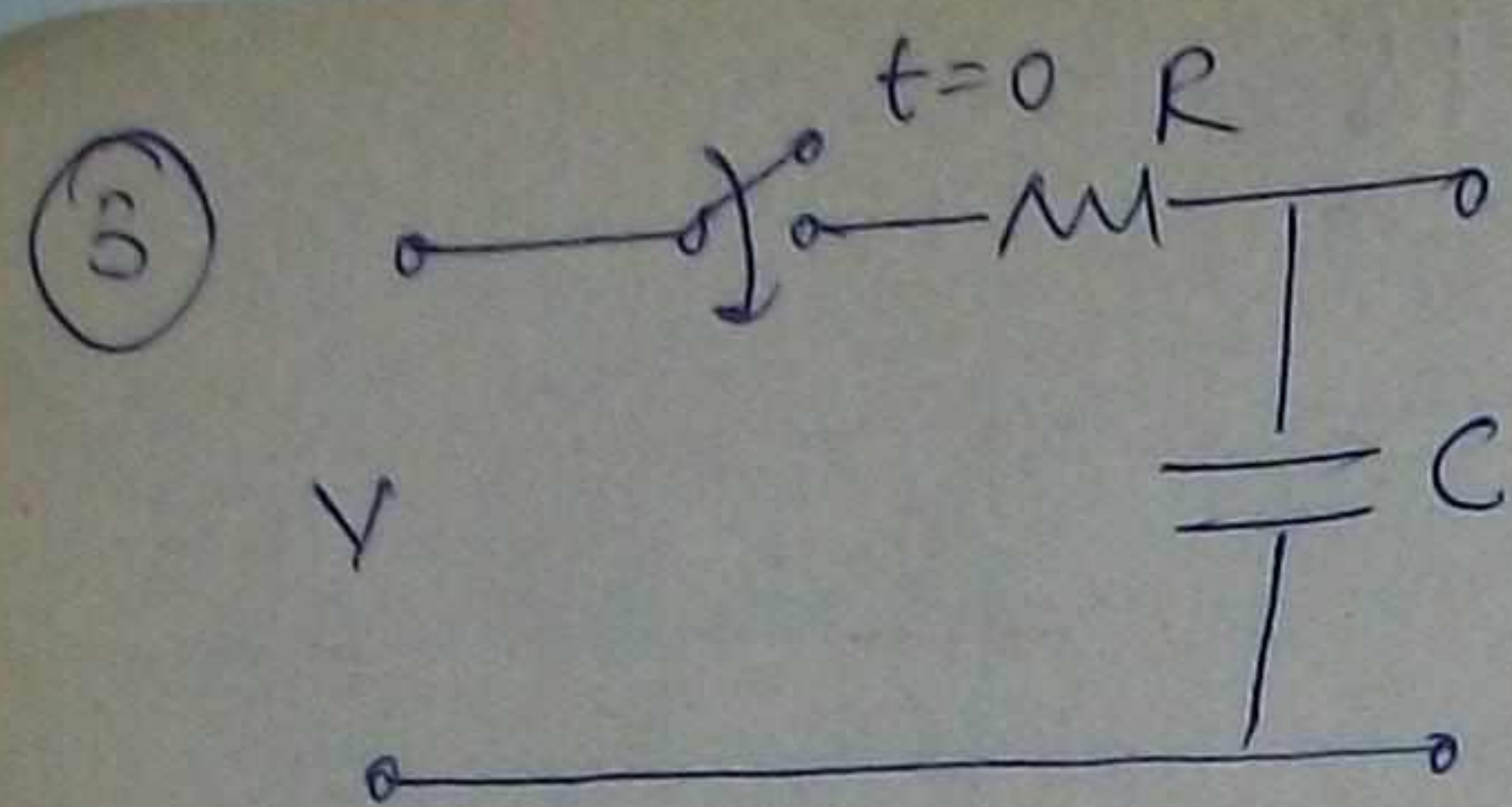
② let $I(t) = \text{infinite or Impulsive current}$

$$\int_{0^-}^{0^+} \delta(t) dt = 1$$

$$\int_{0^-}^{0^+} I(t) dt = q \quad (\text{finite})$$

$$V_c = \frac{q}{C} \quad (\text{finite})$$

used for waveform generation circuit.



Switch is closed at $t = 0$

$$V_C(0^-) = 0 = V_C(0^+)$$

Initially 'C' acts as S.C.

$$V_C = V [1 - e^{-t/RC}]$$

At $T = 0$:

$$V_C = 0 \text{ (S.C.)}$$

At $T = \infty$:

$$V_C = V \text{ (O.C.)}$$

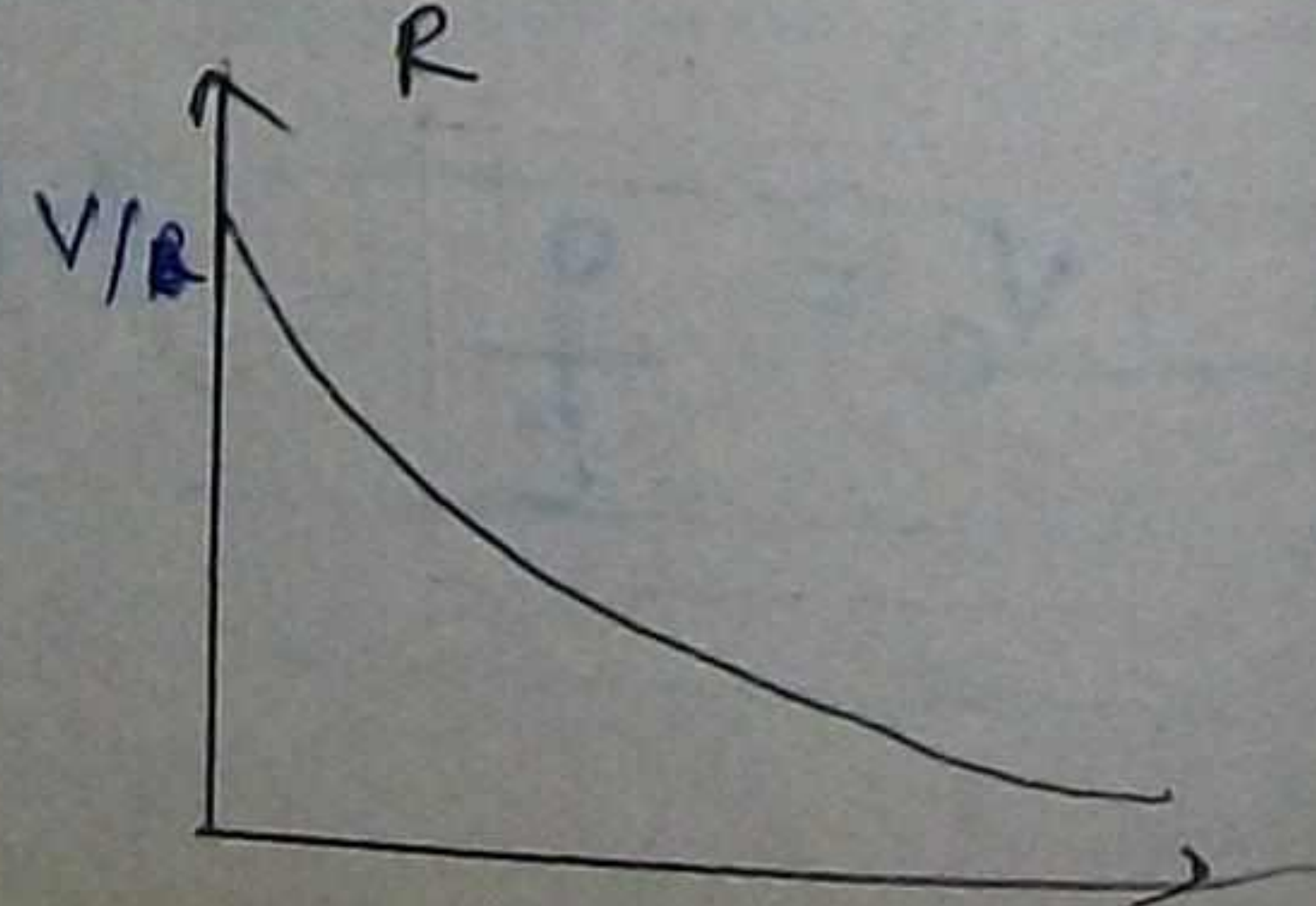
Now,

$$V_R = V - V_C$$

$$V_R = V e^{-t/RC}$$

at $t = 0$ $V_R = V$

$t = \infty$ $V_R = 0$



⊗

$$V_x = V_f - [V_f - V_i] e^{-t/\tau}$$

for capacitor:

$$V_f = V, V_i = 0, \tau = RC$$

↓
GENERAL
EQUATION

$$V_C = V - [V - 0] e^{-t/RC}$$

$$V_C = V [1 - e^{-t/RC}]$$

→ CHARGING
OF
CAPACITOR

for Resistor:

$$V_f = 0, V_i = V, \tau = RC$$

$$V_R = 0 - [0 - V] e^{-t/RC}$$

$$V_R = V e^{-t/RC}$$

→ DISCHARGING

for current:

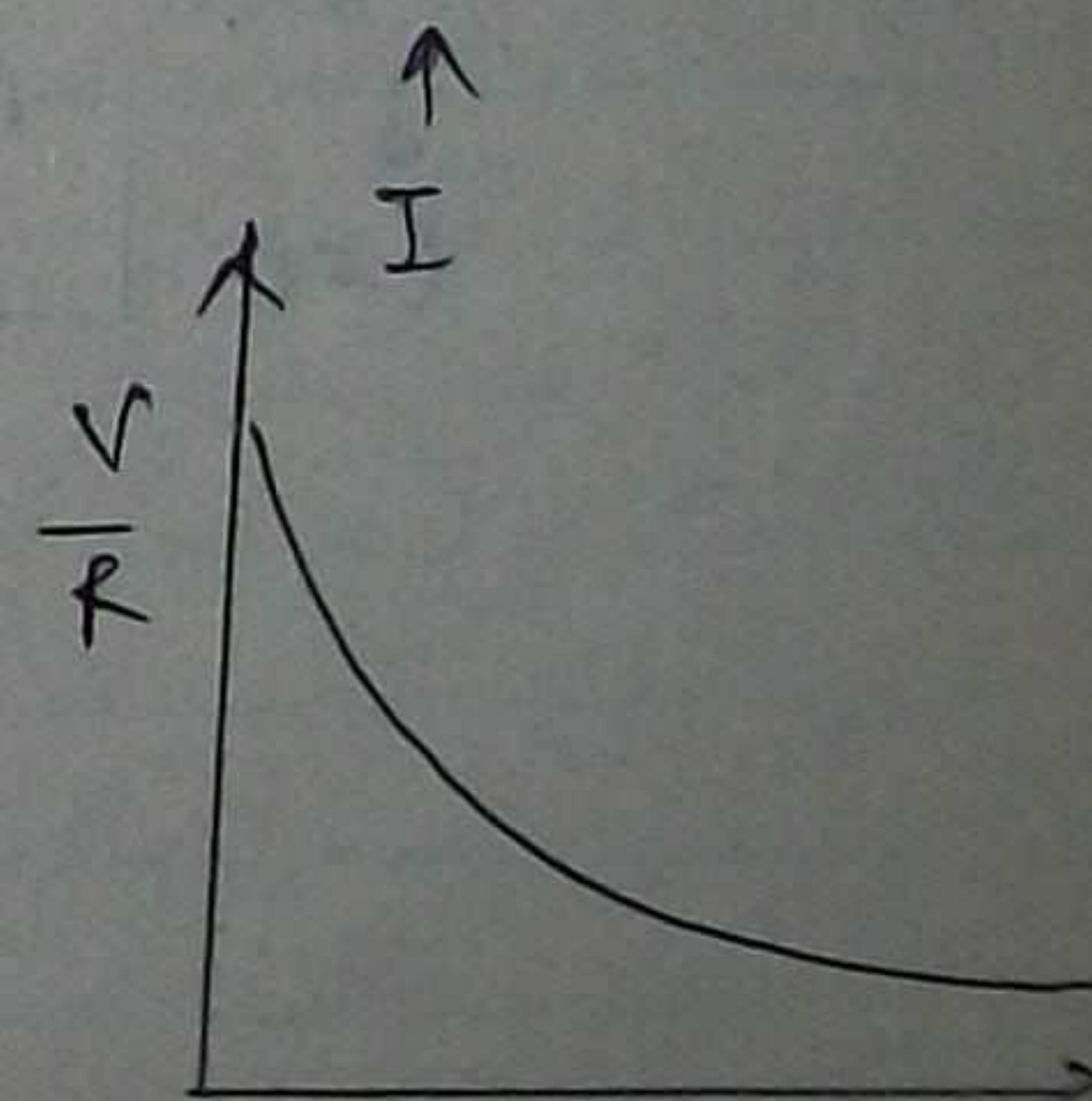
$$I = \frac{V - V_C}{R} = \frac{V - V(1 - e^{-t/RC})}{R}$$

$$I = \frac{V}{R} e^{-t/RC}$$

$$I_i = V/R, I_f = 0, \tau = RC$$

$$I_x = 0 - (0 - V/R) e^{-t/RC}$$

$$I_x = \frac{V}{R} e^{-t/RC}$$

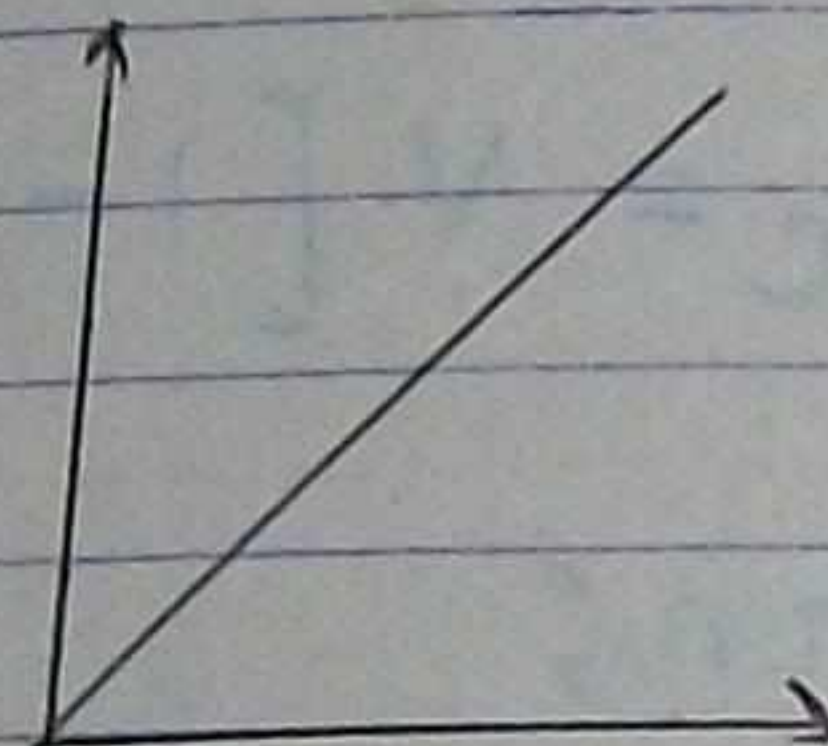


④ for linear charging of capacitor,

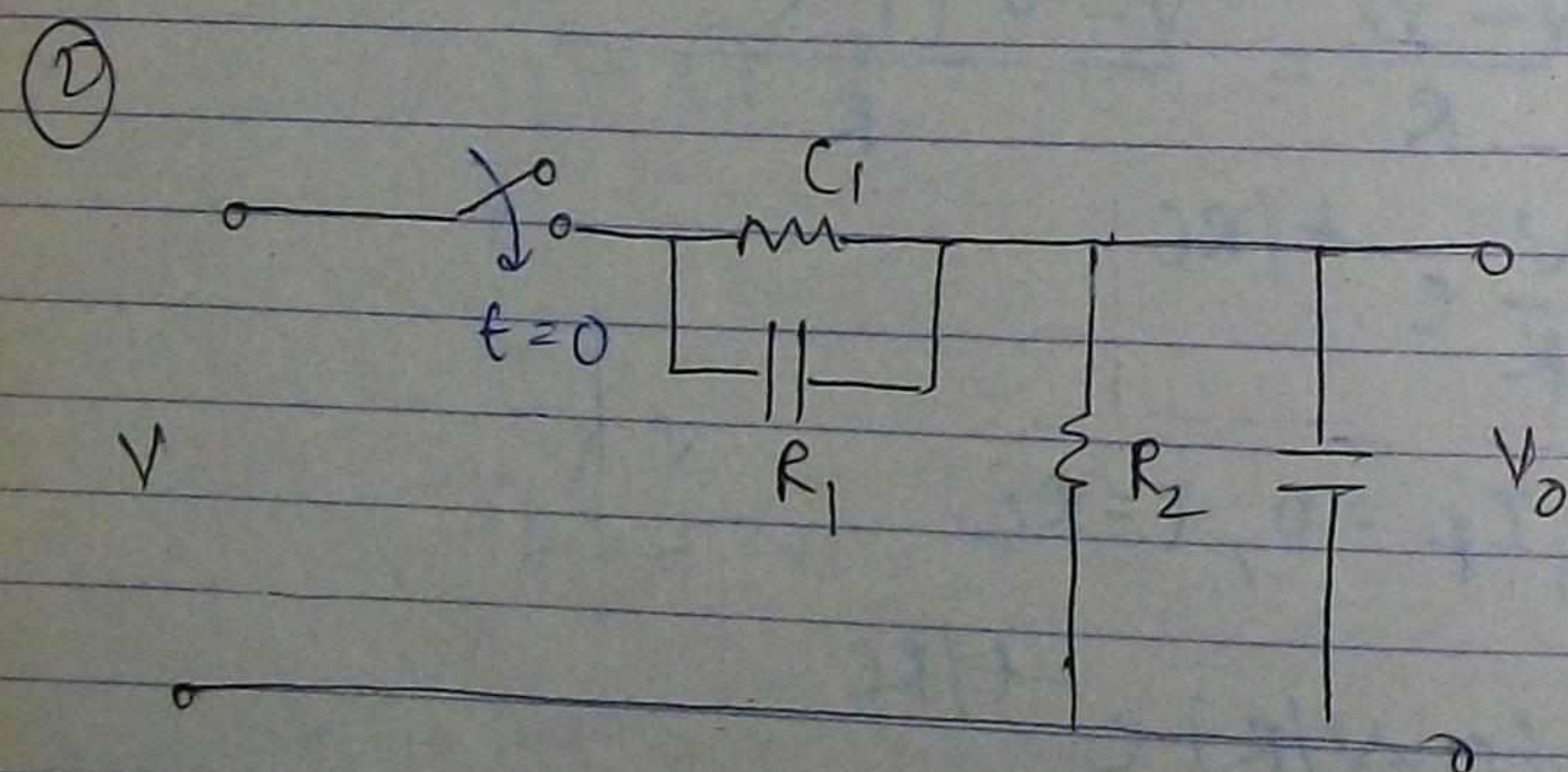
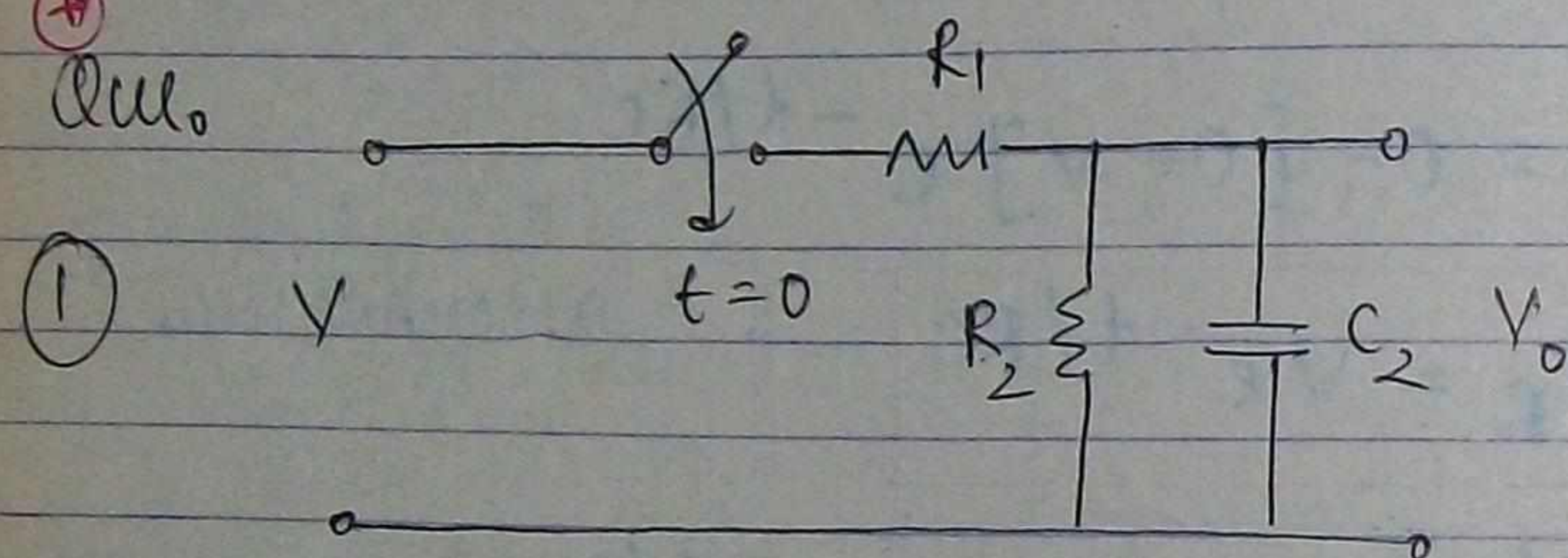
$$I = \text{const}$$

$$V_c = \frac{1}{C} \int_0^t I \cdot dt$$

$$V_c = \frac{It}{C}$$



Ques.



① $V_c(0^-) = 0 = V_c(0^+) = 0$

$$I(0^+) = \frac{V}{R_1} = \text{finite}$$

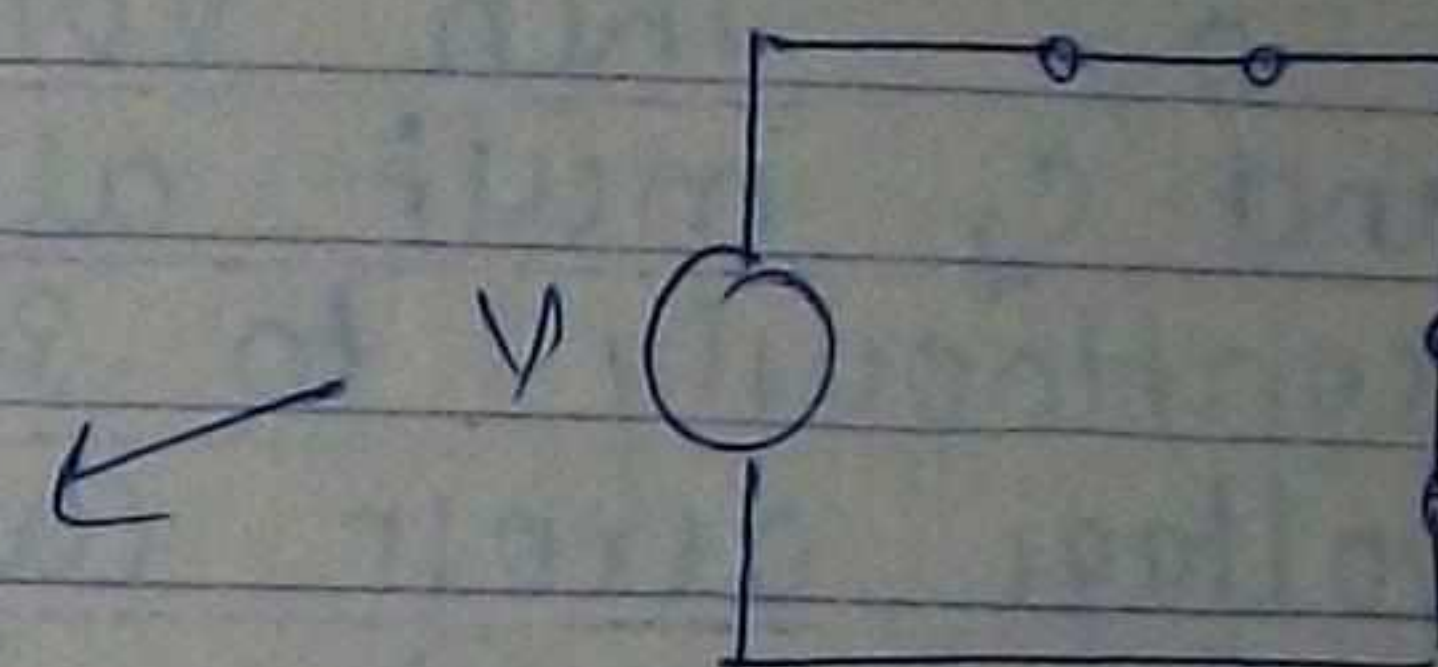
KVL is satisfied

② $V_{C_2}(0^-) = V_{C_2}(0^+) = V_0(0^+) = 0$

$$V_{C_1}(0^-) = V_{C_1}(0^+) = 0$$

$$V = 0$$

KVL is violated



Now, $t = 0^+$

$$I(0^+) = \frac{V}{0} = \infty$$

$$\therefore q(0^+) = \int_{0^-}^{0^+} I(0^+) dt = \text{finite}$$

$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2}$$

$$q(0^+) = V C_{eq} = V \frac{C_1 C_2}{C_1 + C_2}$$

$$V_o(0^+) = V_{C_2}(0^+) = \frac{q(0^+)}{C_2}$$

$$V_o(0^+) = \frac{VC_1}{C_1 + C_2}$$

① As i/p changes abruptly from 0 to V (when switch is closed at $t=0$) then voltage across C_1 and C_2 must also change ~~dis~~ - continuously to satisfy KVL. But voltage across capacitor cannot change instantaneously. If current remains finite hence an impulsive current (infinite current) must flow in the circuit.

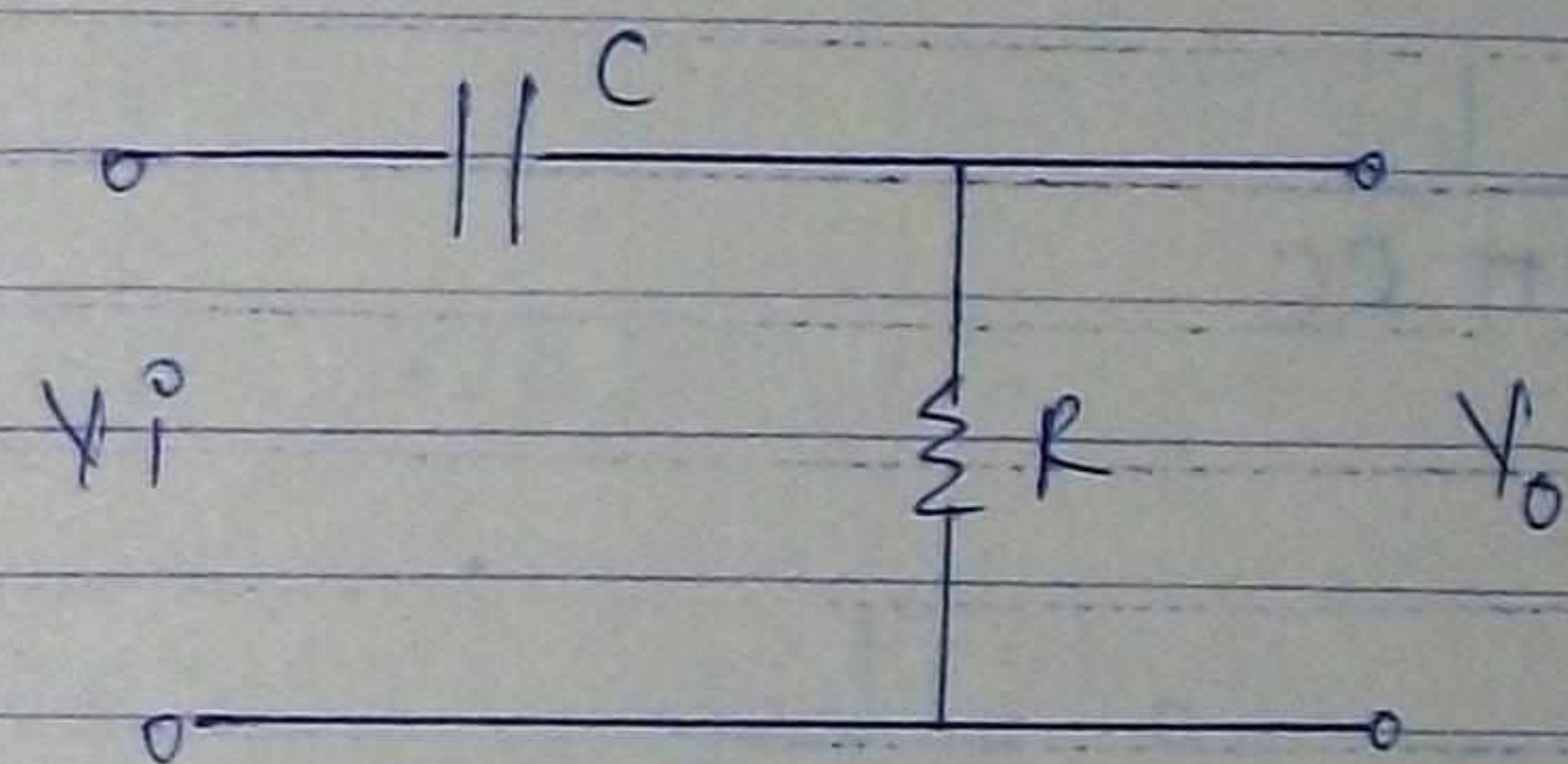
② A infinite current exist at $t=0^+$ so that a finite charge $q(0^+)$ is delivered to each capacitor and cap. allow the sudden change of voltage.

LINEAR WAVE SHAPING CIRCUIT

⇒ LINEAR WAVE SHAPING CIRCUIT -

- The process where by the form of a non-sinusoidal signal is altered by transmission through a linear network is called linear waveshaping.

① HIGH PASS RC CIRCUIT -



$$\frac{V_o}{V_i} = A_v = \frac{R}{R + \frac{1}{j\omega C}} = \frac{R}{R - \frac{j}{\omega C}}$$

$$A_v = \frac{1}{1 - \frac{j}{\omega RC}}$$

$$\phi = \tan^{-1}\left(\frac{1}{\omega RC}\right)$$

$$\phi = +ve \quad (\text{LEAD CIRCUIT})$$

$$|A| = \frac{1}{\sqrt{1 + \frac{1}{\omega^2 R^2 C^2}}}$$

At lower freq.

$$\omega_L = 2\pi f_L = \frac{1}{RC}$$

$$f_L = \frac{1}{2\pi RC}$$

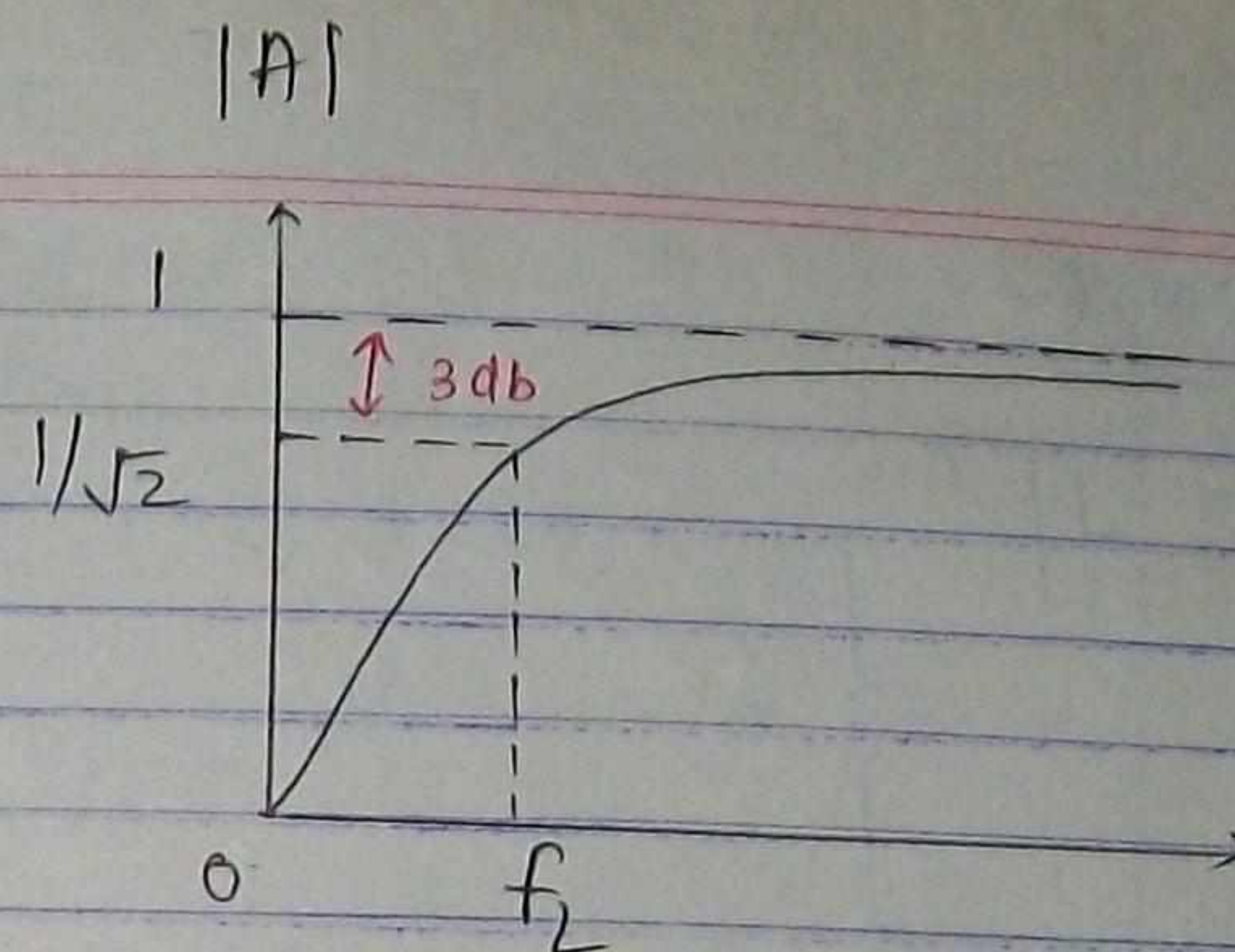
$$A = \frac{1}{1 - j\frac{\omega_L}{\omega}} = \frac{1}{1 - j\frac{f_L}{f}}$$

$$|A| = \frac{1}{\sqrt{1 + \left(\frac{f_L}{f}\right)^2}}$$

at $f = 0$ $|A| = 0$

$f = f_L$ $|A| = 1/\sqrt{2}$

$f = \infty$ $|A| = 1$

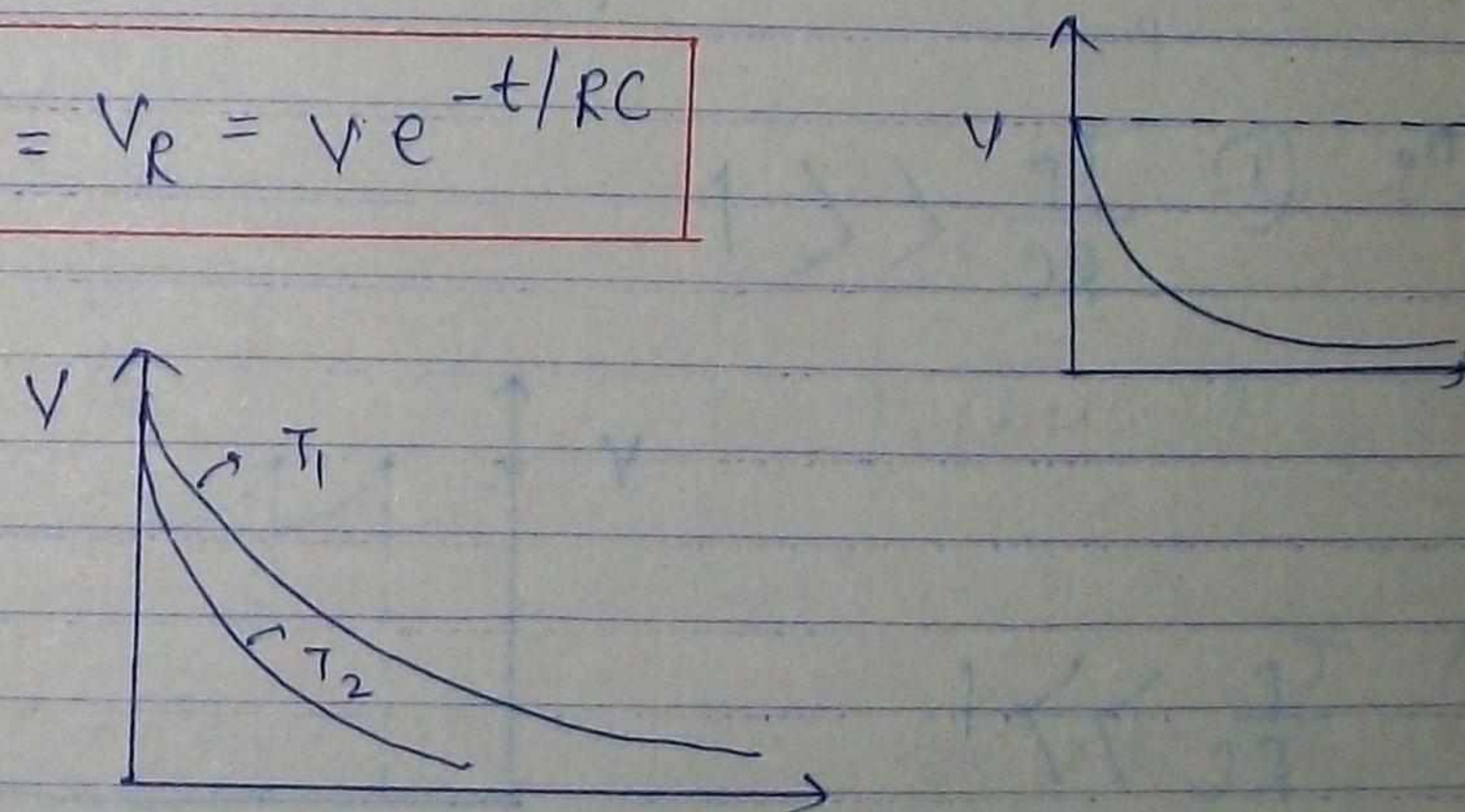


$f \uparrow$ $|A| \downarrow$

$f_L = 3 \text{ db}$ (cutoff frequency)

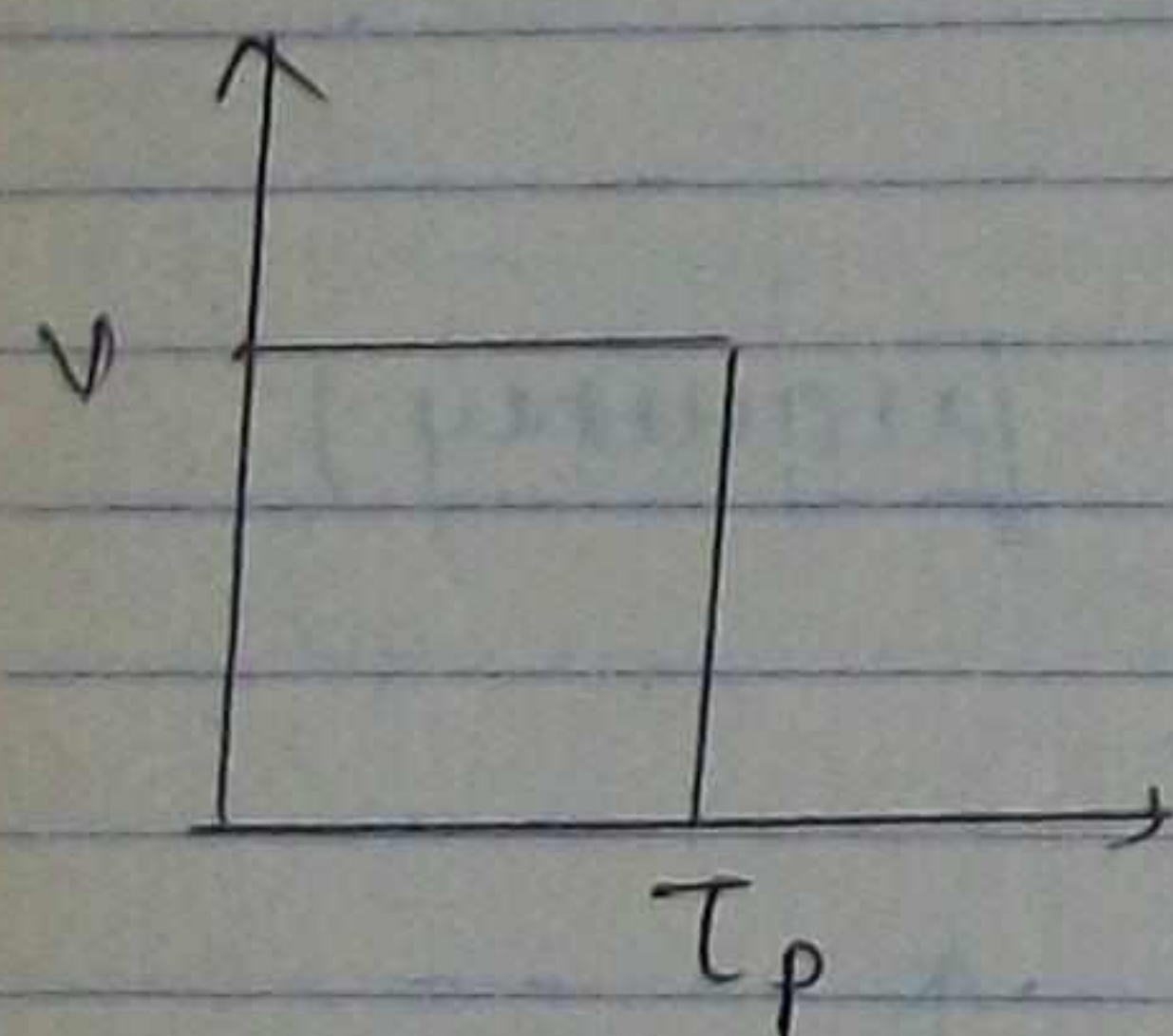
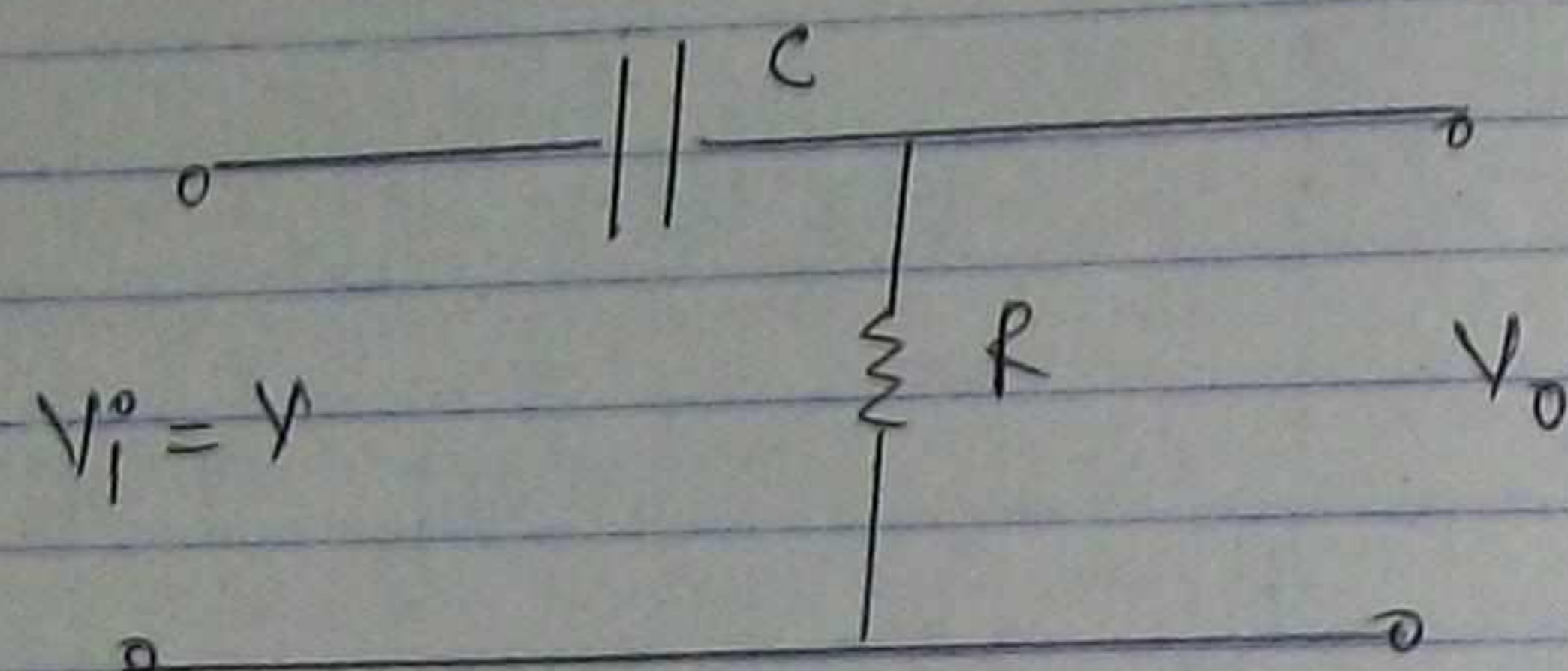
① for step response -

$$V_o = V_R = V e^{-t/RC}$$



Here $T_2 < T_1$

② PULSE RESPONSE-



① $\frac{T_p}{RC} \ll 1$

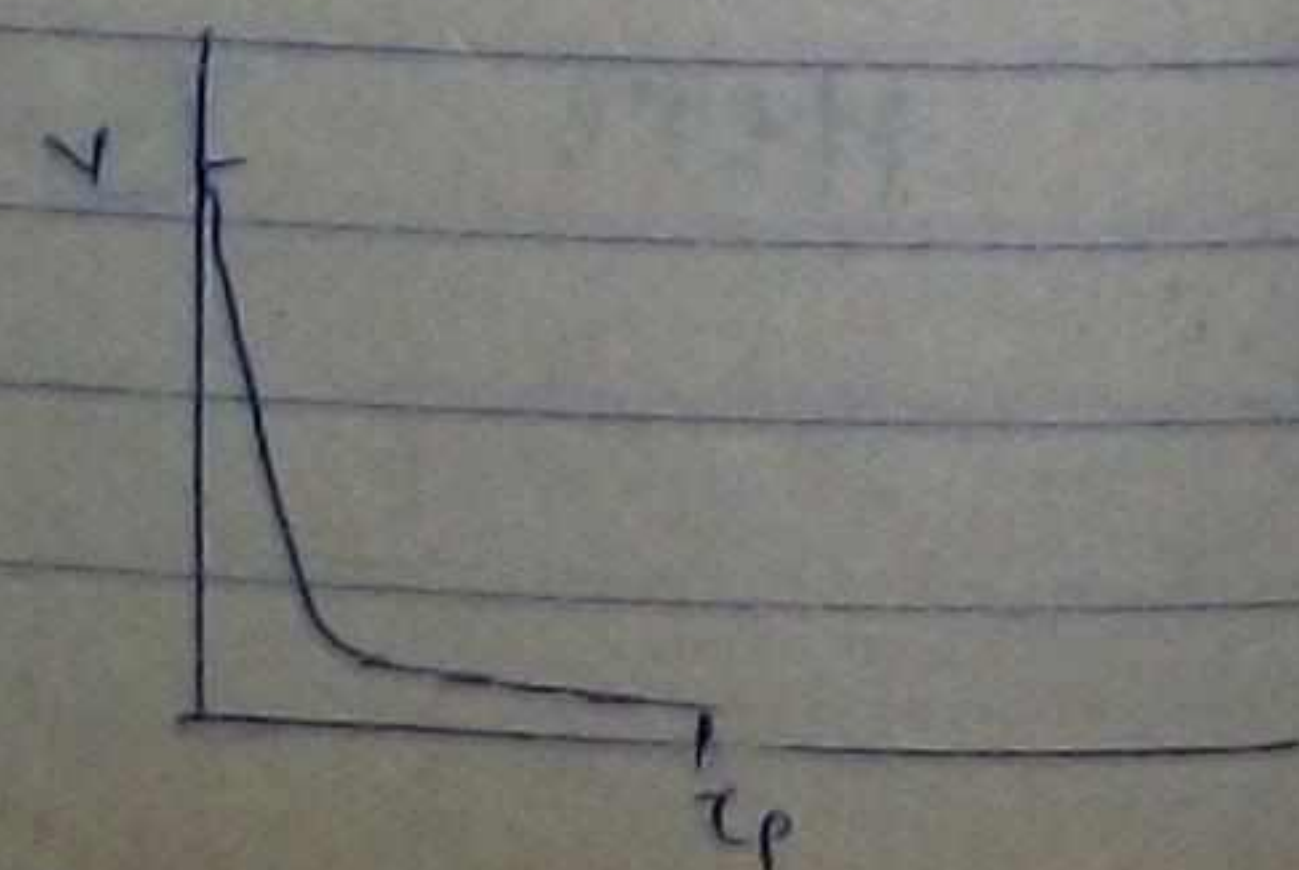
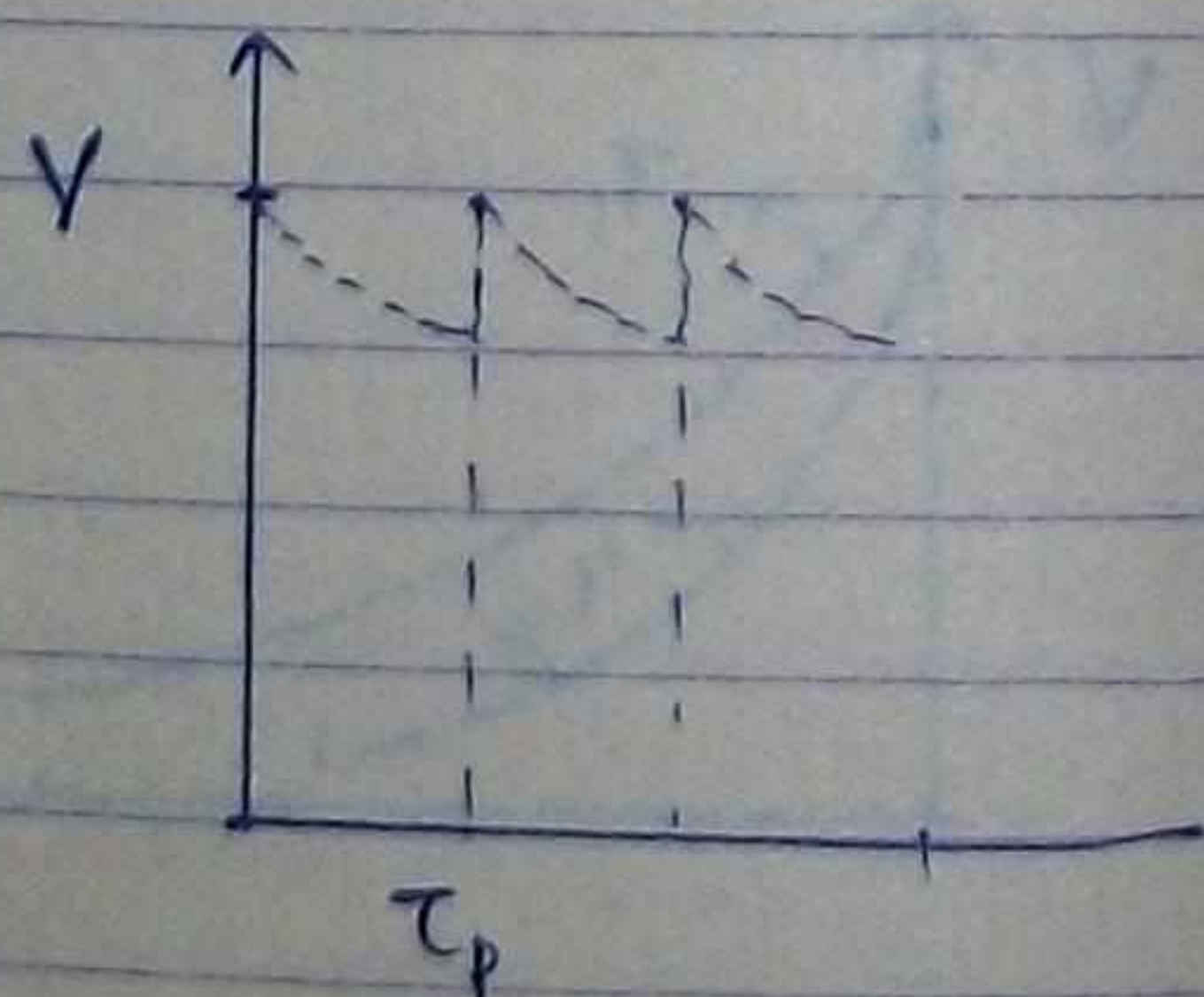
② $\frac{T_p}{RC} \gg 1$

Solⁿ: ① $\frac{T_p}{RC} \ll 1$

(WRONG)

② $\frac{T_p}{RC} \gg 1$

(WRONG)



Solution:

① $\frac{T_p}{RC} \ll 1$

$V_c(0^-) = V_c(0^+) = 0$

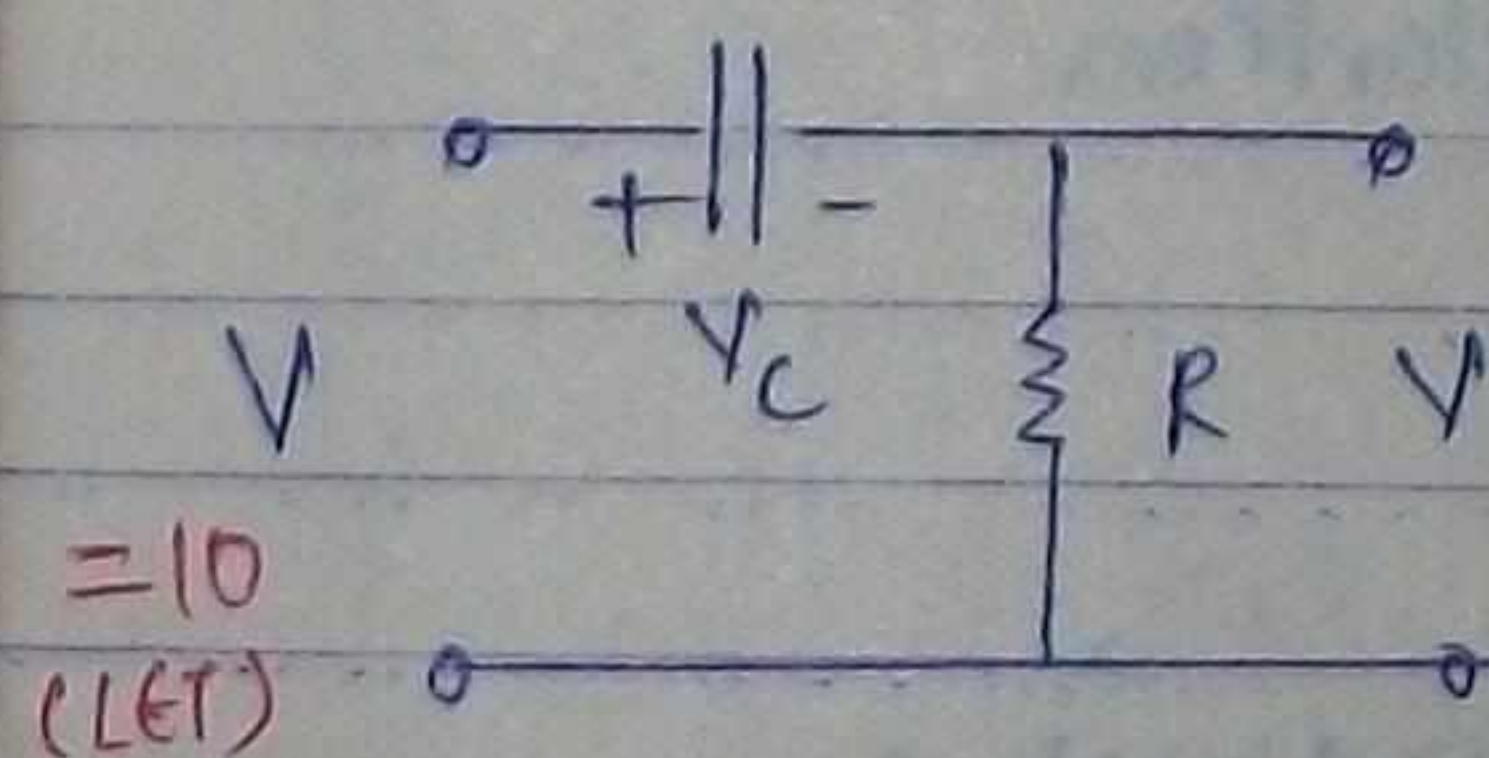
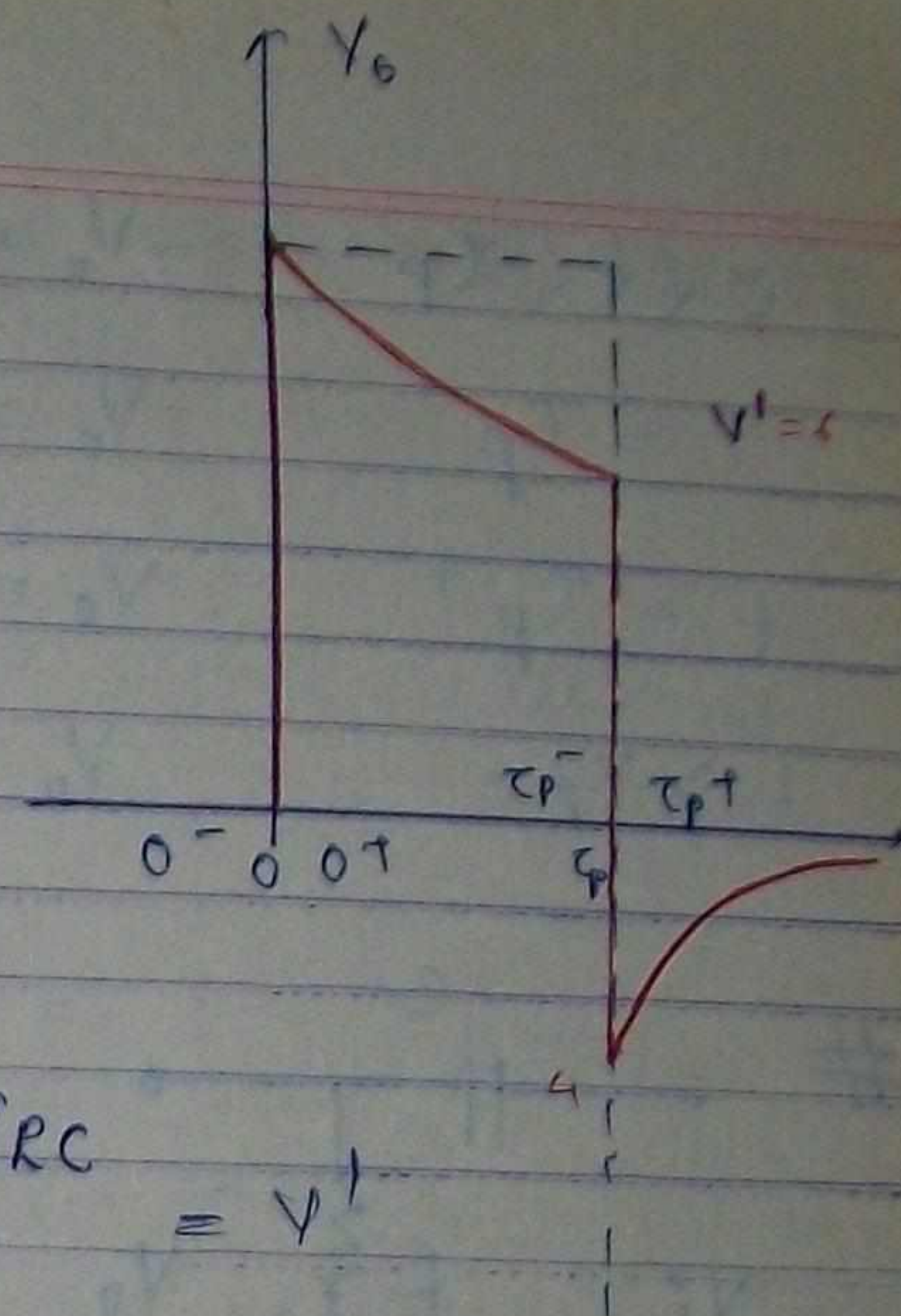
$V_o(0^+) = V$

$V_o = V e^{-t/RC}$

at $t = T_p^-$

$V = V_o(T_p^-) = V e^{-T_p/RC} = V'$

at $t = T_p^-$, $V_i = V$

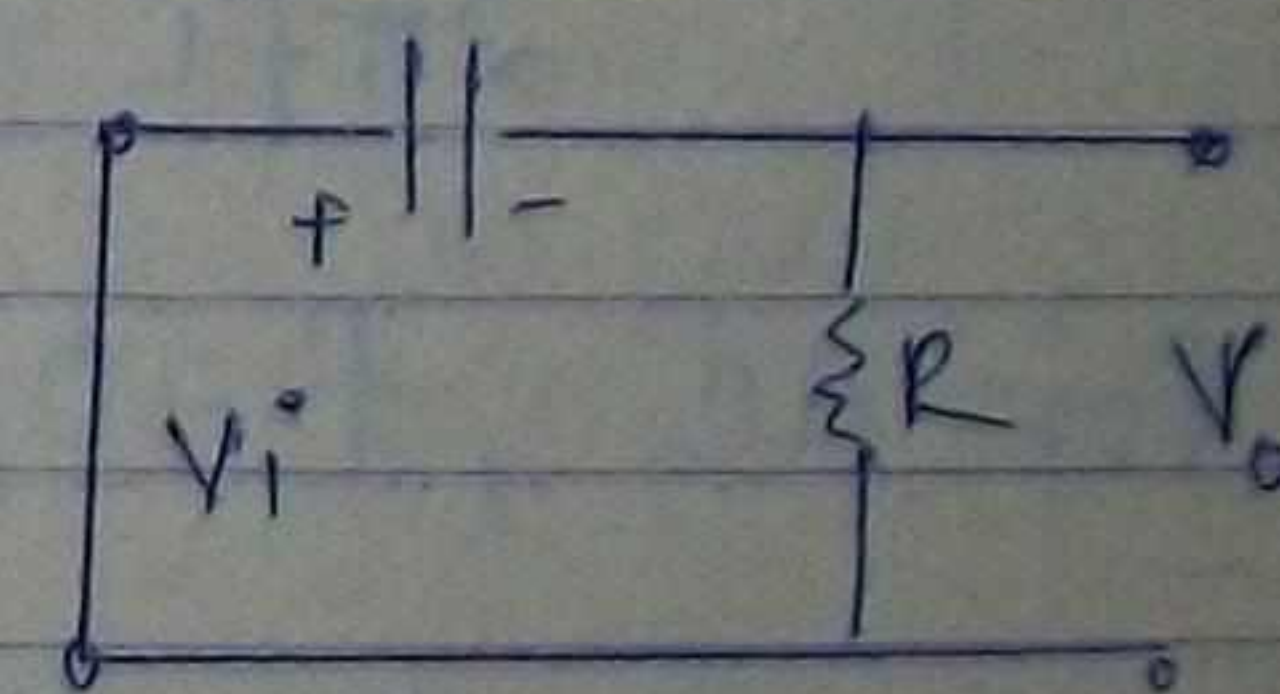


= 10
(LET)

$V_c(T_p^-) = V - V' = 10 - 6 = 4V$

$V_c(T_p^-) = V_c(T_p^+) = 4V$

at $t = T_p^+$, $V_i = 0$



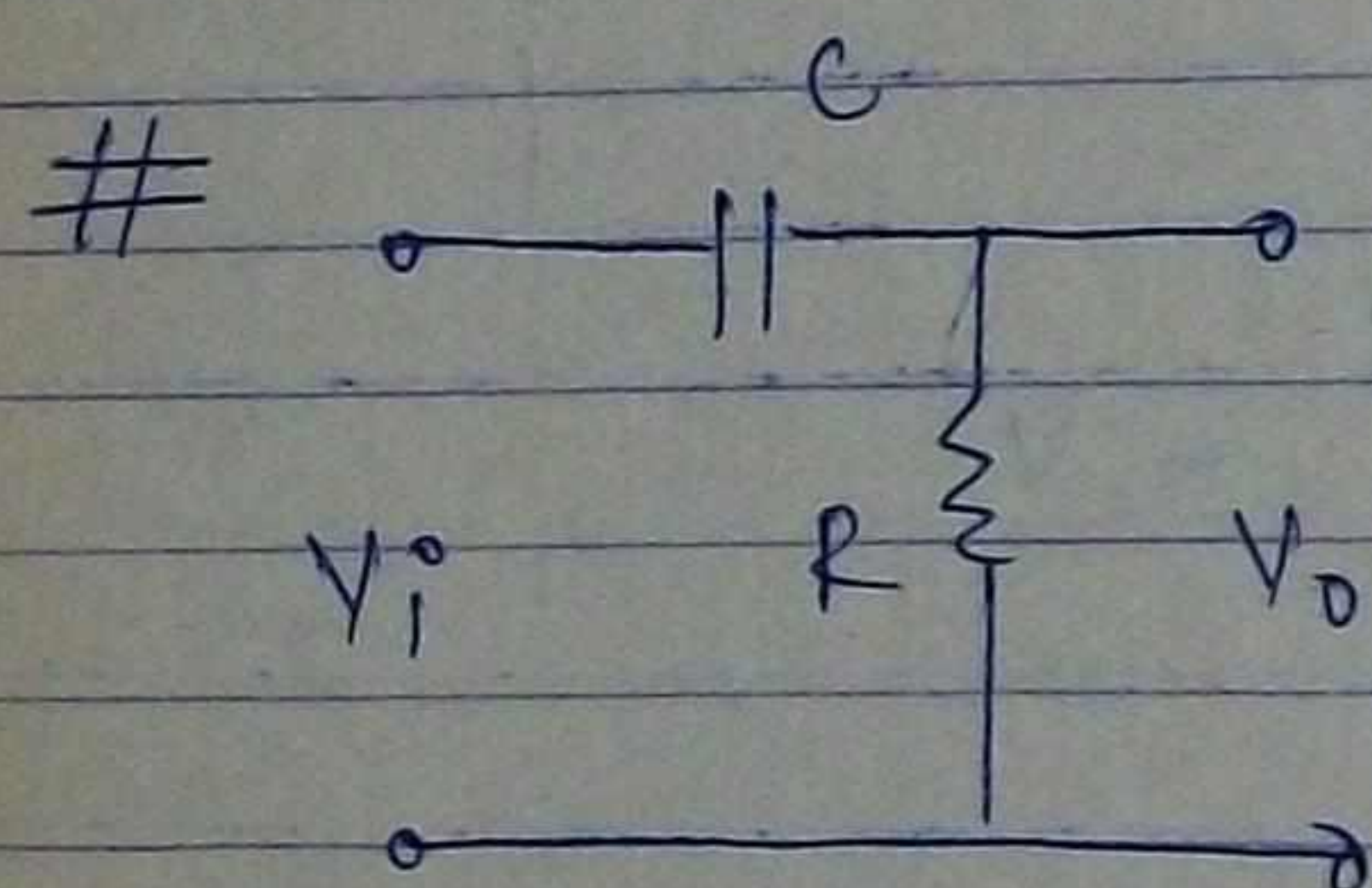
$V_o(T_p^+) = -V_c(T_p^+) = 4$

$V_c = 4 e^{-t/RC}$

$V_o = -4 e^{-t/RC}$

$V_o = -(V - V') e^{-\frac{(t-T_p)}{RC}}$

$$\begin{aligned}
 0 < t < \tau_p & V_o = V e^{-t/RC} \\
 t = \tau_p & V_o = V e^{-\tau_p/RC} = V' \\
 t = \tau_p^+ & V_o = -(V - V') \\
 t > \tau_p & V_o = -(V - V') e^{-(t - \tau_p)/RC}
 \end{aligned}$$



$$Z_C = \frac{1}{2\pi f C}$$

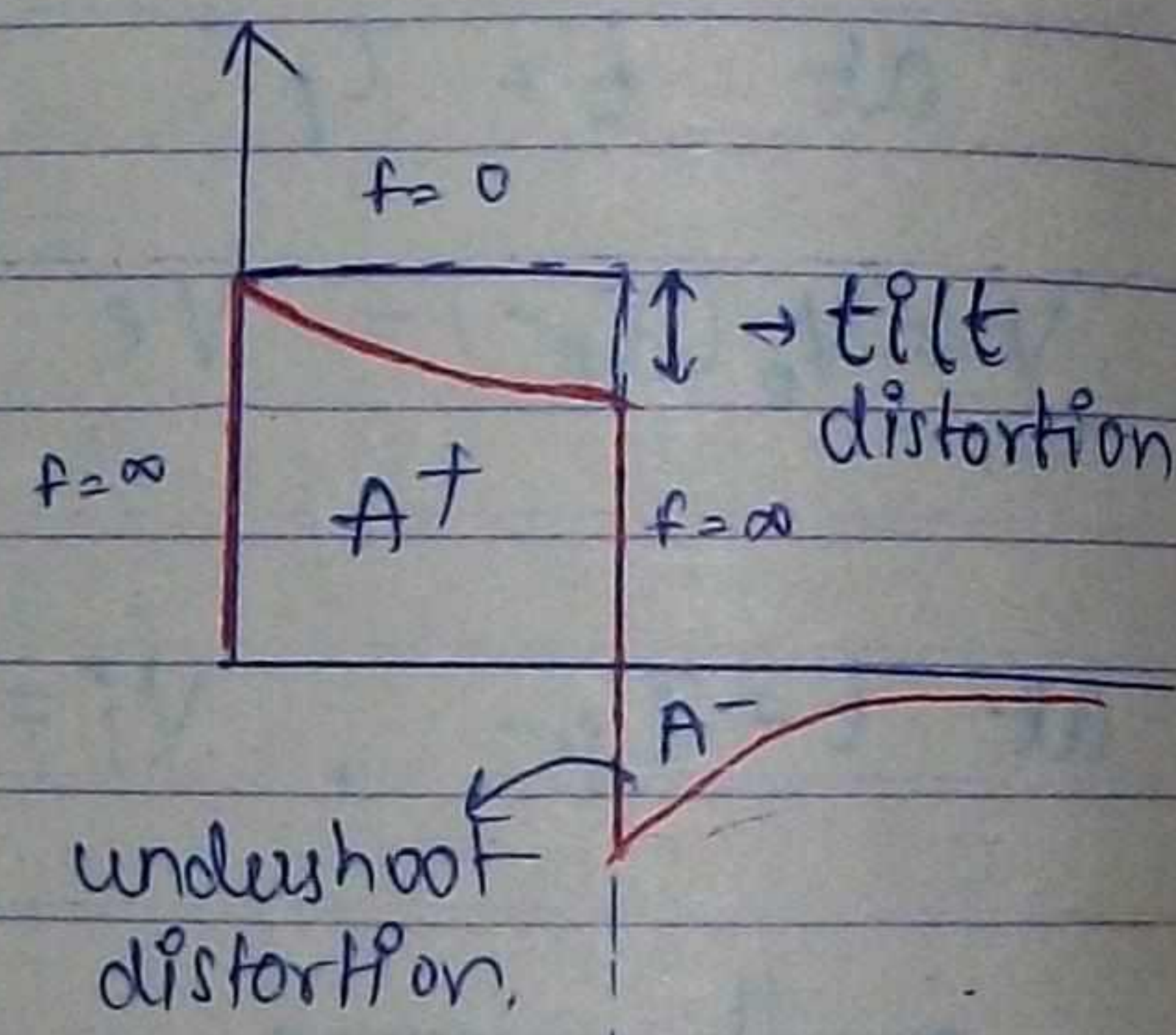
for dc, $f=0$, $Z_C = \infty$

C acts as dc blocking capacitor

$\therefore$ dc value or average value of O/P = 0

ii

$$A^+ = A^- \quad (\text{always})$$



- to reduce Tilt and undershoot RC should be maximum.

FOR HIGH PASS FILTER -

- Average level of O/P signal is always zero, independent of the average level of I/P

ie $A^+ = A^-$ (Area)

- when I/P change discontinuously by an amount V then O/P change discontinuously by an equal amount and in same direction.

- During any finite time interval when I/P maintains a constant level the O/P decays exponentially towards zero voltage.

- for high pass circuit that when the time constant is very large, we are getting 'tilt' at the top of pulse and 'undershoot' at the end of the pulse.

[DIFFERENTIATOR]

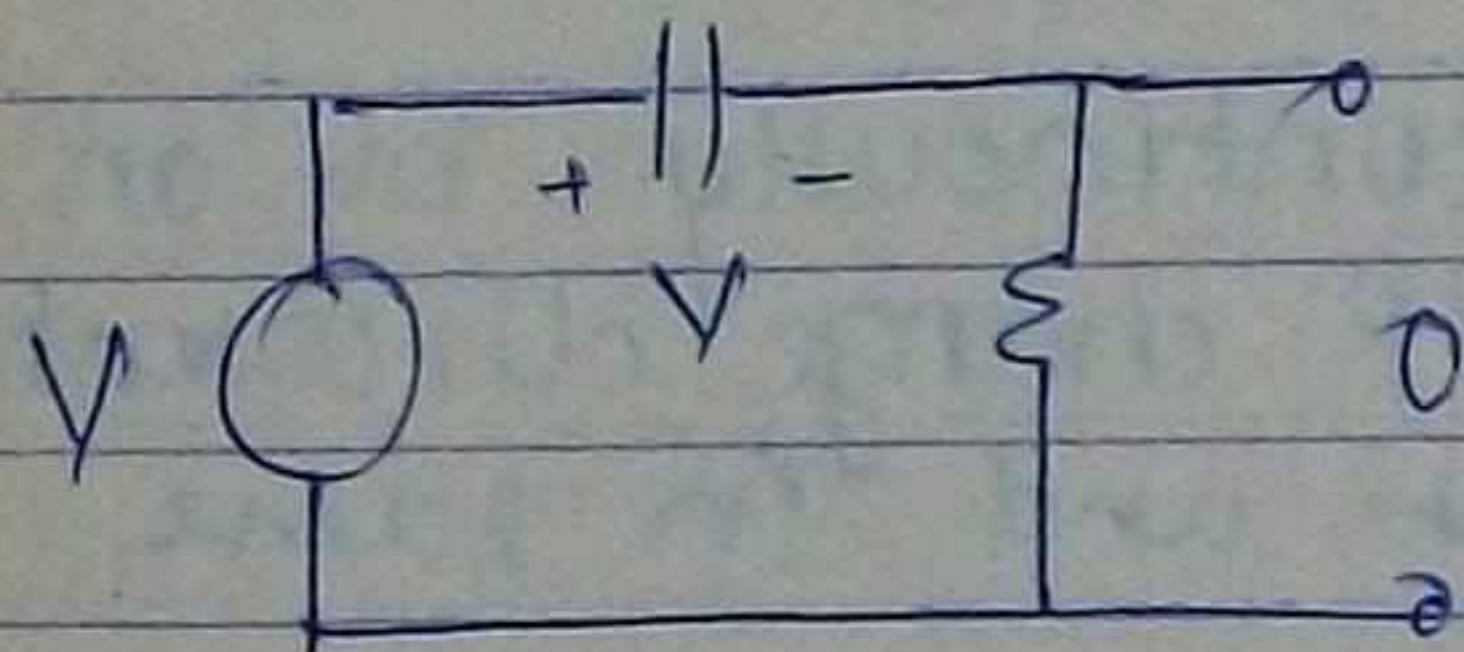
② $RC \ll \tau_p$

$$V_c(0^-) = V_c(0^+) = 0$$

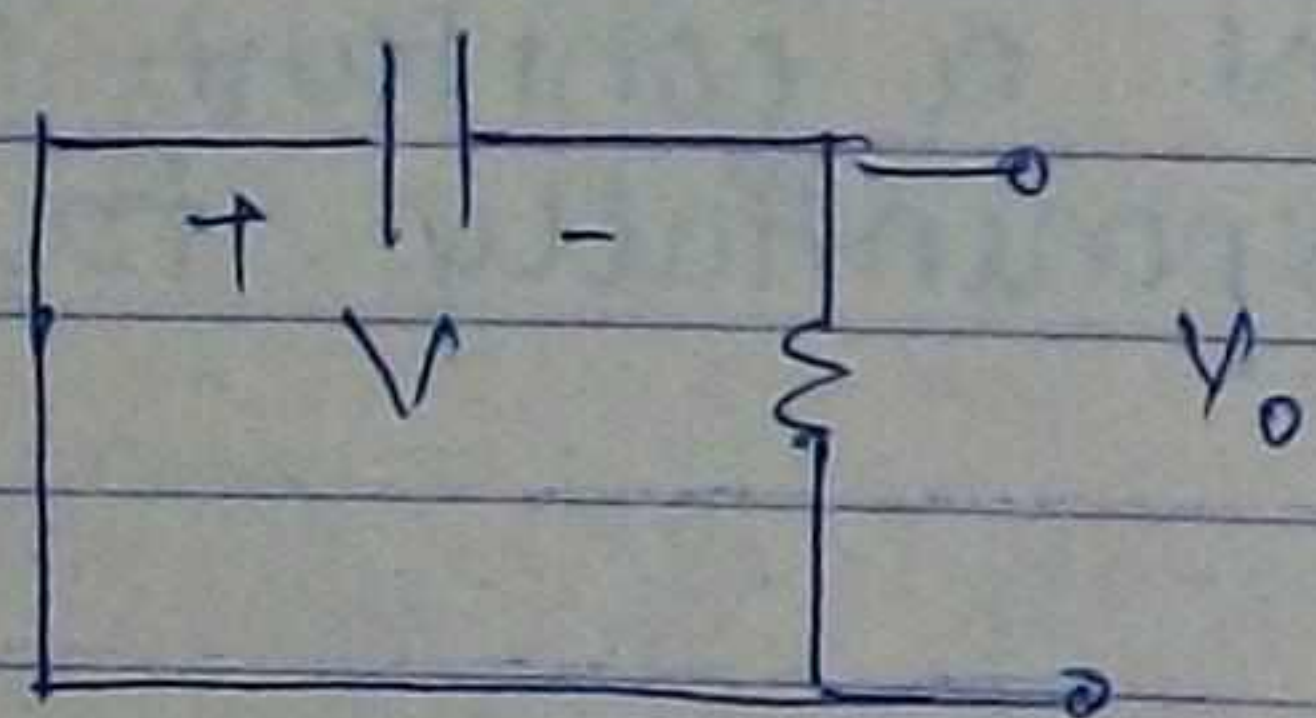
$$V_c(0^+) = V$$

$$\text{at } t = t_1 = V_c = V$$

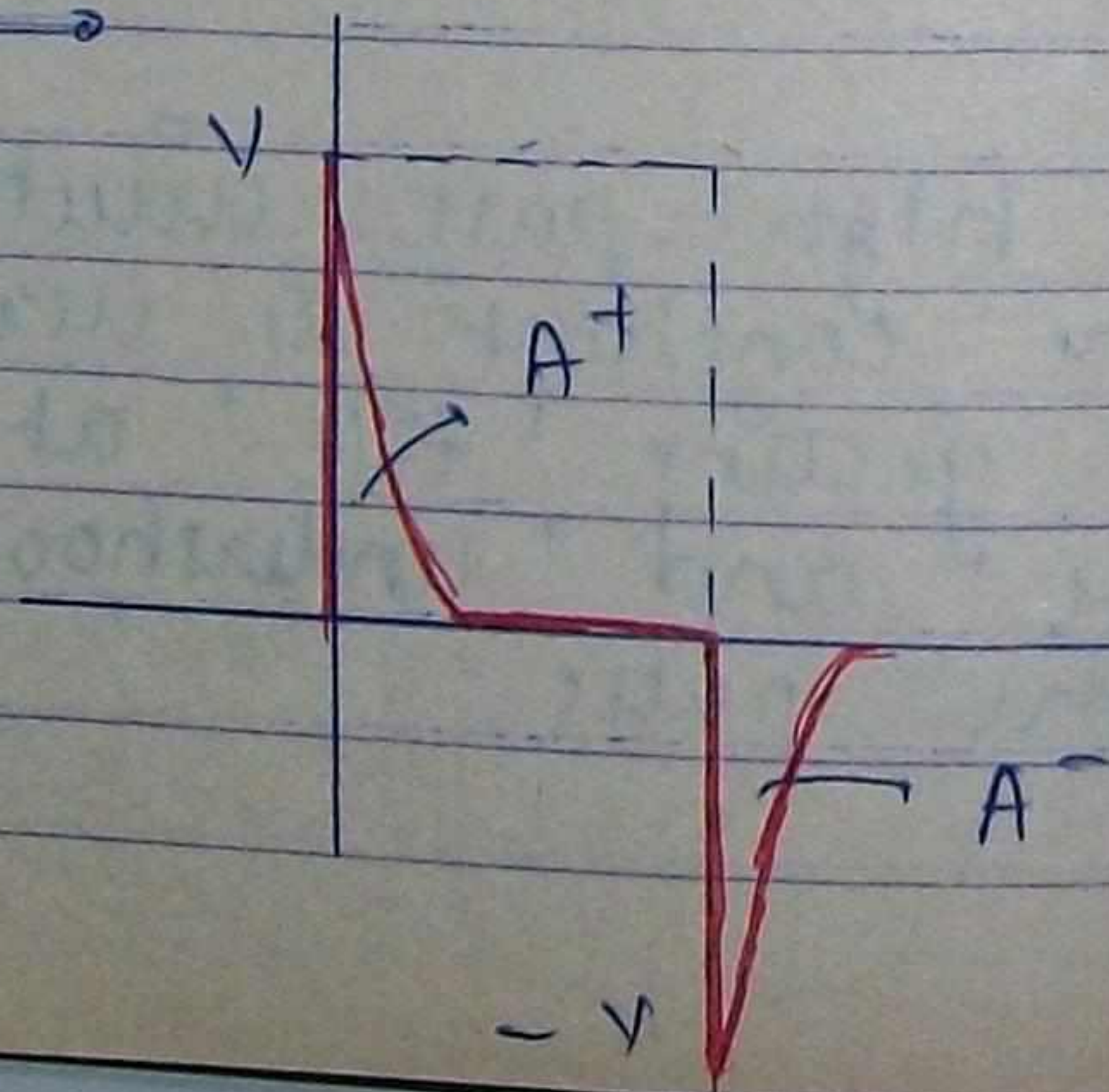
$$V_o = V_R = 0$$



$$\text{at } t = \tau_p + V_i = 0$$



$$A^+ = A^-$$



③ HIGH PASS FILTER = DIFFERENTIATOR

- Positive spike of amplitude V at beginning of pulse and -ve of same size at end of the pulse. This process of converting pulse into spikes or PIPS by means of a high pass RC circuit of short time constant is called PEAKING.

- High pass filter is act as a differentiator when time constant of circuit is smaller than time period of i/p pulse.

21/08/11

⇒ TILT OR SAGIT -

$RC = \text{very high.}$

$$V_o = V e^{-t/RC}$$

$$x = \frac{t}{RC} = \text{very low}$$

$$e^{-x} = [1 - x]$$

$$V_o = V \left[1 - \frac{t}{RC} \right]$$

linear equation.

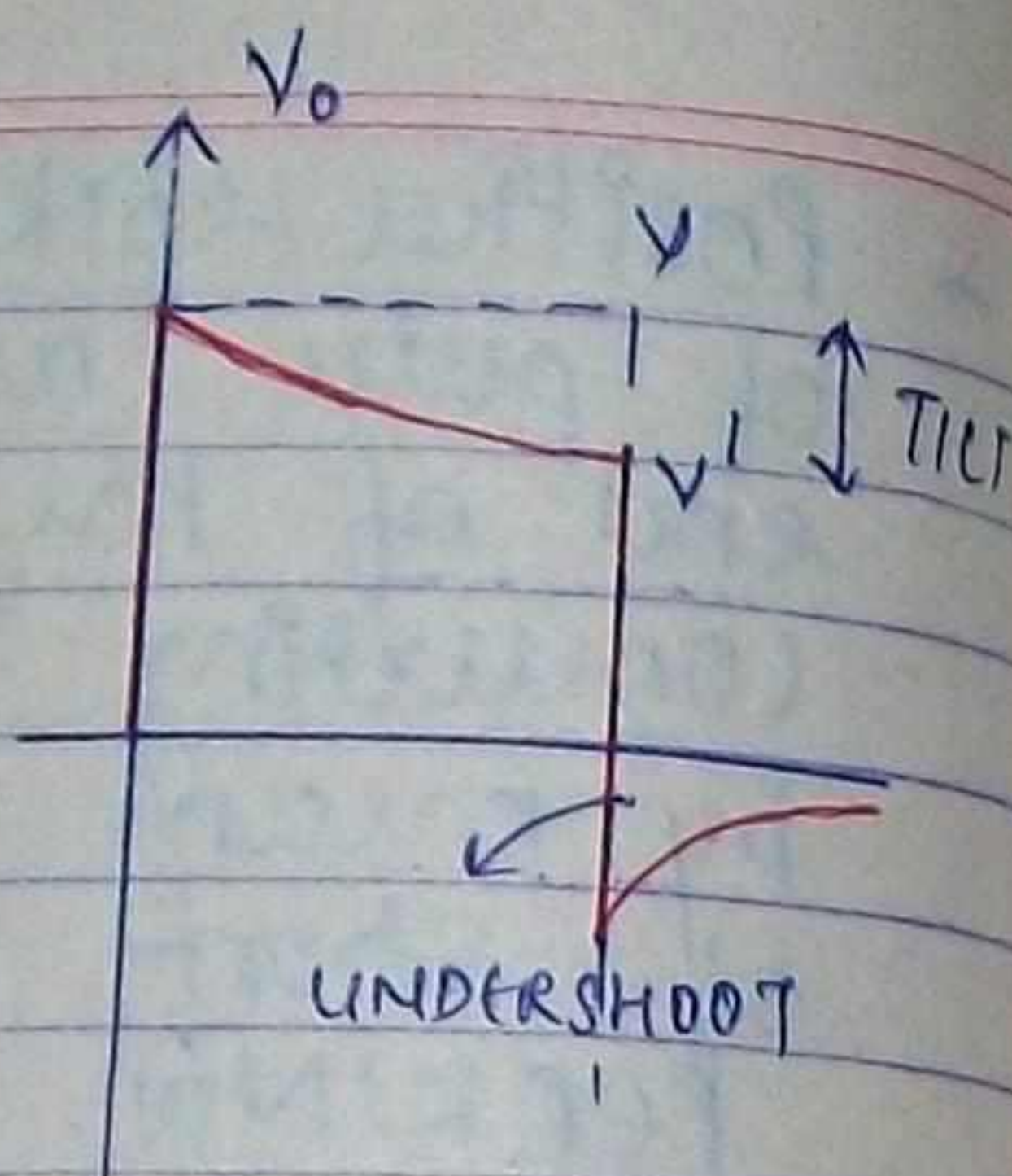
At $t = t_1$

$$\text{Tilt} = V - V'$$

$$\% \text{ TILT} = \frac{V - V'}{V} \times 100\%$$

at $t = t_1$, $V_o = V'$

$$V' = V \left[1 - \frac{t_1}{RC} \right]$$

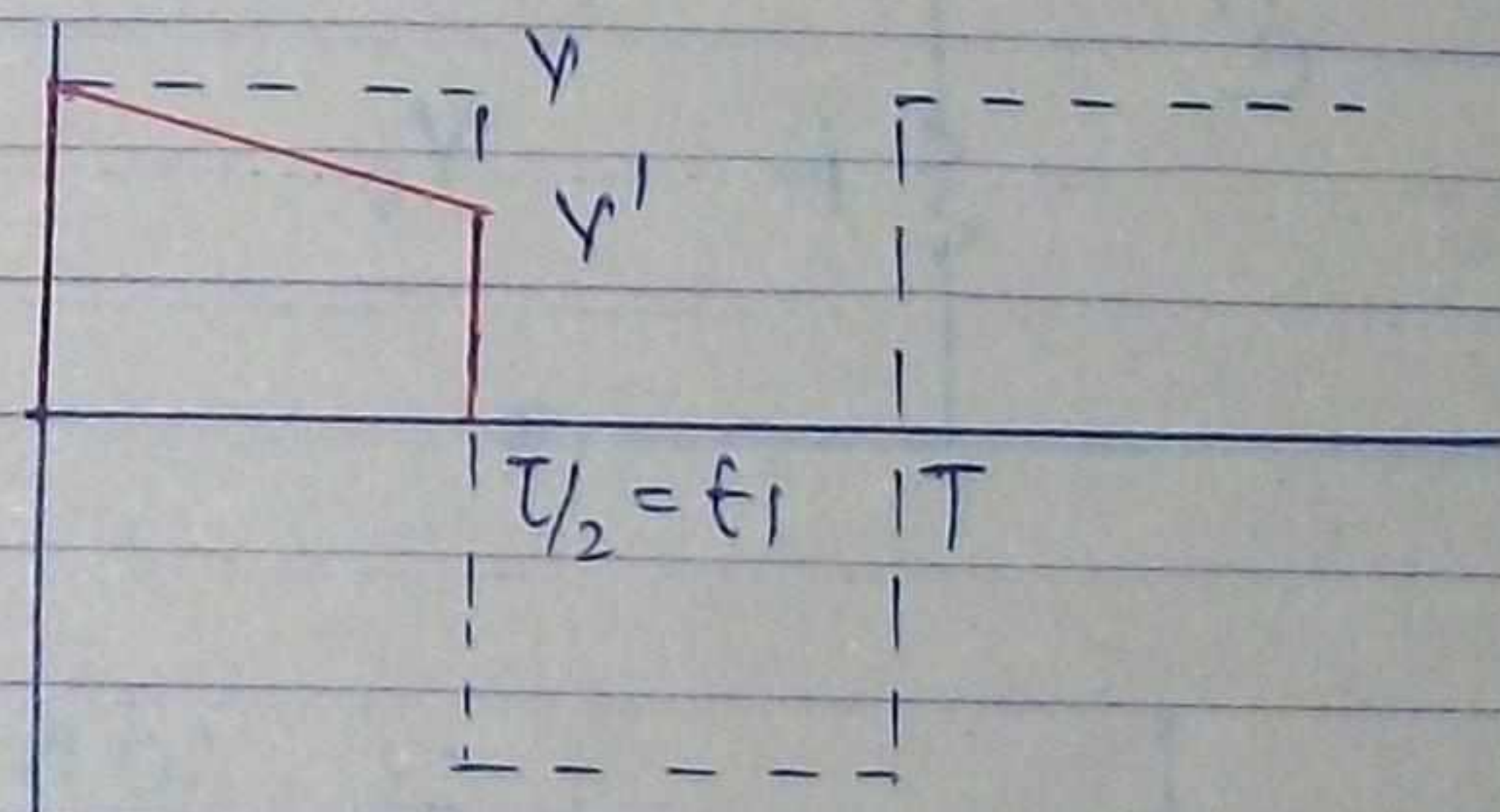


$$\% \text{ TILT} = \frac{t_1}{RC} \times 100\%$$

$$\omega_L = 2\pi f_L = \frac{1}{RC}$$

$$\% \text{ TILT} = 2\pi f_L t_1 \times 100\%$$

Tilt for symmetrical square wave



$$t_1 = T/2 = 1/2f$$

$$\because T = 1/f$$

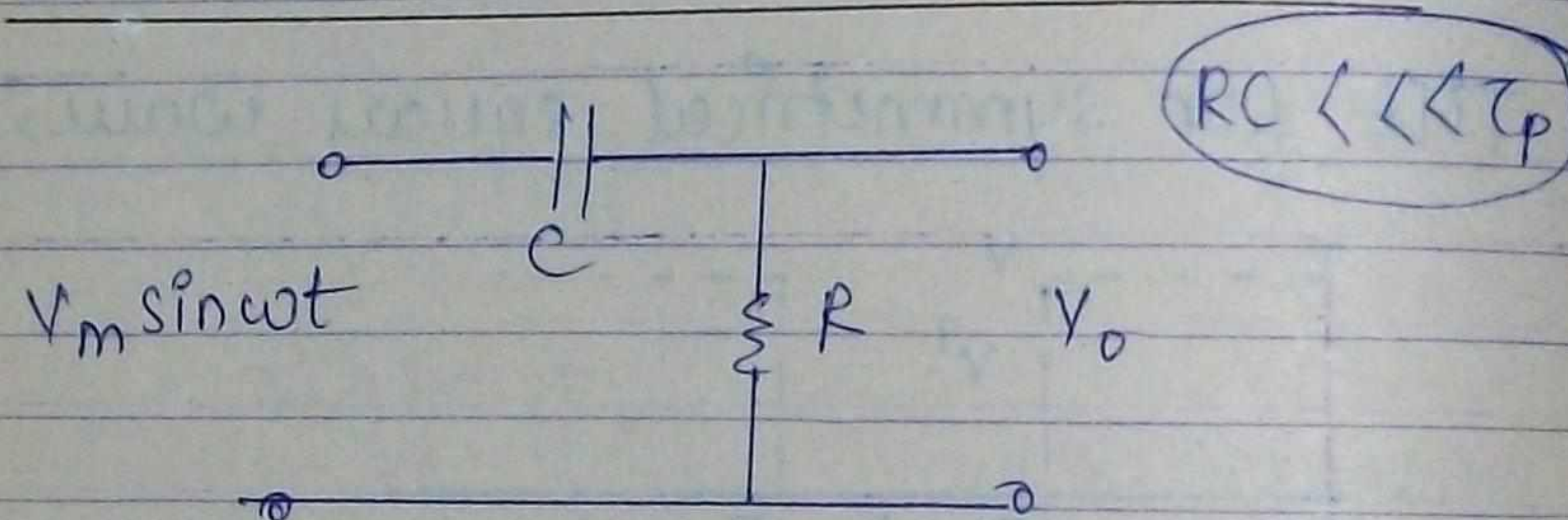
$$\% \text{ TILT} = 2\pi f_L \times \frac{1}{2f} \times 100\%$$

$$\% \text{ TILT} = \frac{\pi f_L}{f} \times 100\%$$

⇒ HIGH PASS RC CIRCUIT AS A DIFFERENTIATOR.

- when the time constant is very small in comparison with time period of I/P signal then the circuit is called differentiator.

⇒ CRITERIA FOR GOOD DIFFERENTIATOR -



$$|A| = \frac{1}{\sqrt{1 + \frac{1}{\omega^2 R^2 C^2}}} \approx \omega RC$$

$$\phi = \tan^{-1} \left(\frac{1}{\omega RC} \right)$$

$$V_o = |A| V_m \sin(\omega t + \phi)$$

$$V_o = \frac{V_m}{\sqrt{1 + \frac{1}{\omega^2 R^2 C^2}}} \sin(\omega t + \phi)$$

φ

$$\omega RC = \frac{2\pi}{T} \times RC$$

① IDEAL DIFFERENTIATOR	90°	0 (practically not possible)
② BEST DIFF.	89.4° ✓✓	0.01 ✓✓
③ GOOD DIFF.	84.3	0.1

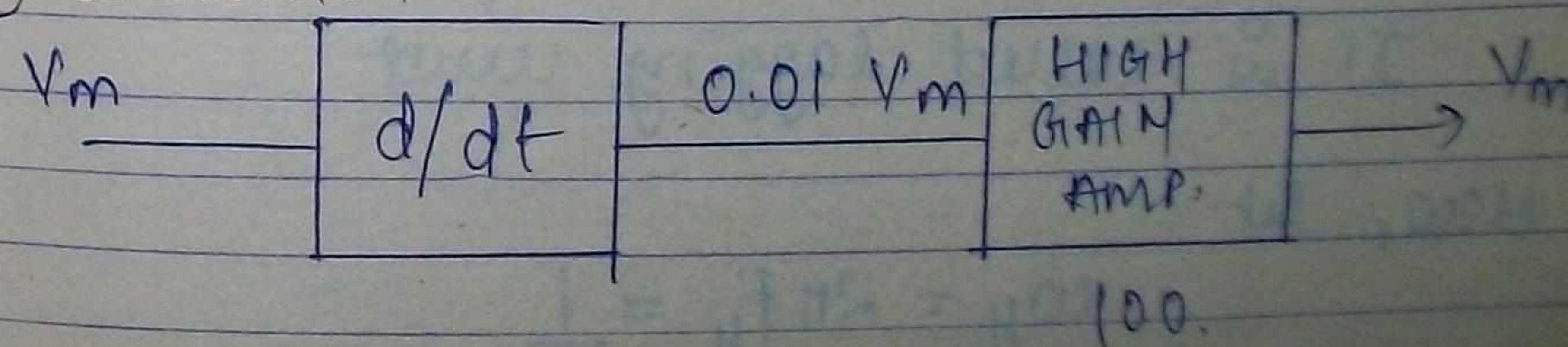
Let, $\phi = 89.4^\circ$ and $\omega RC = 0.01$

$$\frac{1}{\omega^2 R^2 C^2} \gg 1$$

$$V_o = \omega RC V_m \sin(\omega t + \phi)$$

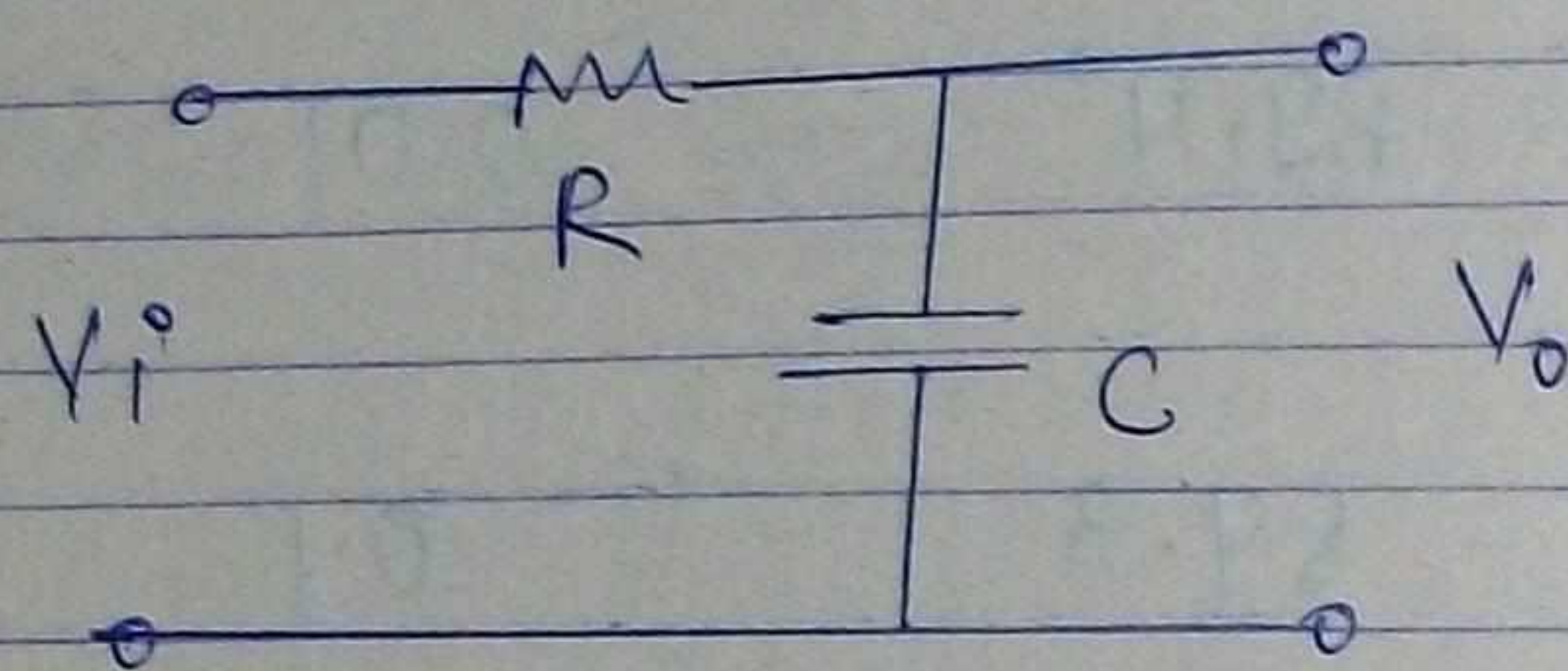
$$V_o = 0.01 V_m \sin(\omega t + 89.4^\circ)$$

* CONCLUSION -



- Thus differentiator is always followed by an high gain amplifier.

② LOW PASS RC CIRCUIT -



$$A = \frac{V_o}{V_i} = \frac{1/j\omega C}{R + 1/j\omega C} = \frac{1}{1 + j\omega RC}$$

$$|A| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}}$$

$$\phi = -\tan^{-1}(\omega RC)$$

since $\phi = -\omega$

It is called lagging unit

Now, let

$$\omega_H = 2\pi f_H = \frac{1}{RC}$$

$$A = \frac{1}{1 + j\frac{\omega}{\omega_H}} = \frac{1}{1 + j\frac{f}{f_H}}$$

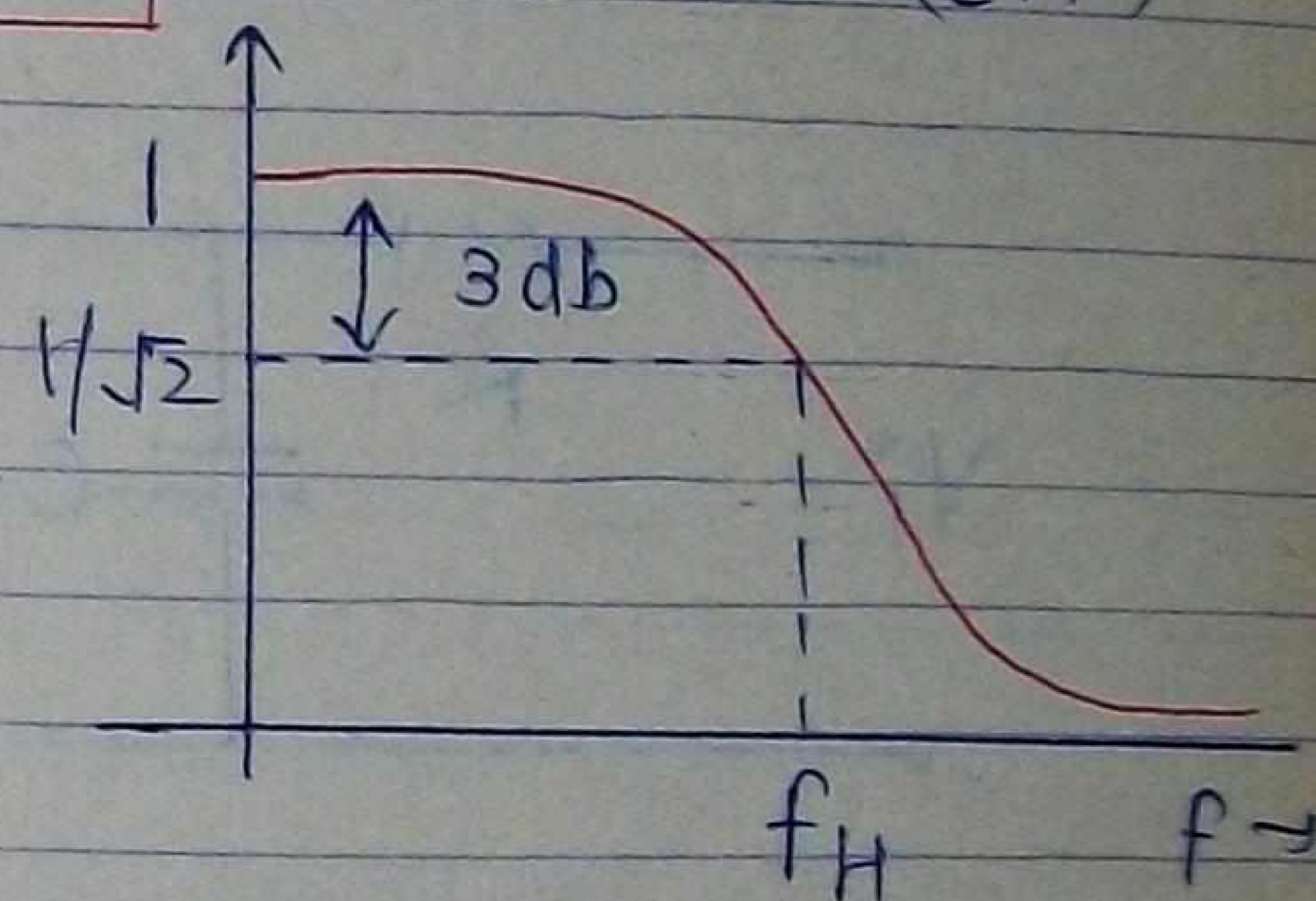
$$|A| = \frac{1}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}}$$

(LPP)

$$f = 0 \quad |A| = 1$$

$$f = f_H \quad |A| = 1/\sqrt{2}$$

$$f = \infty \quad |A| = 0$$



$f_H = 3\text{db}$ (cutoff freq.)

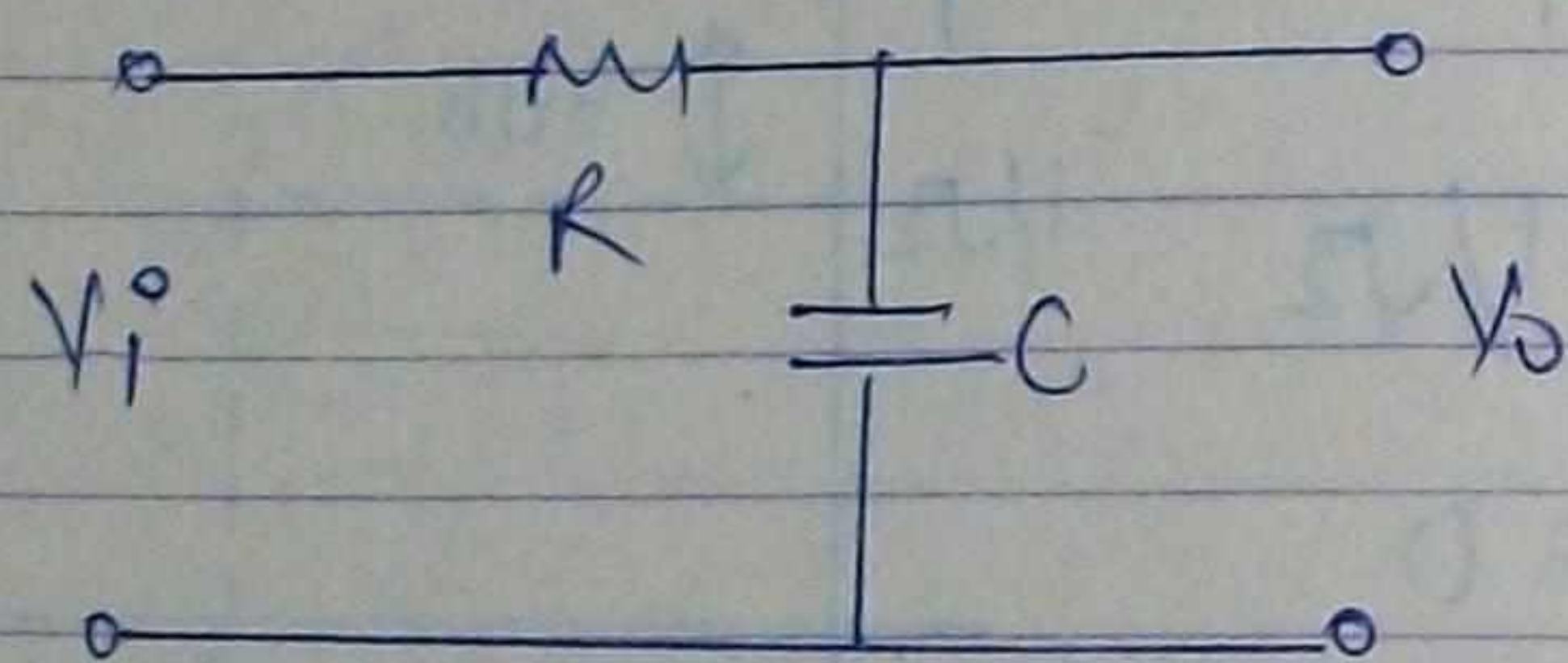
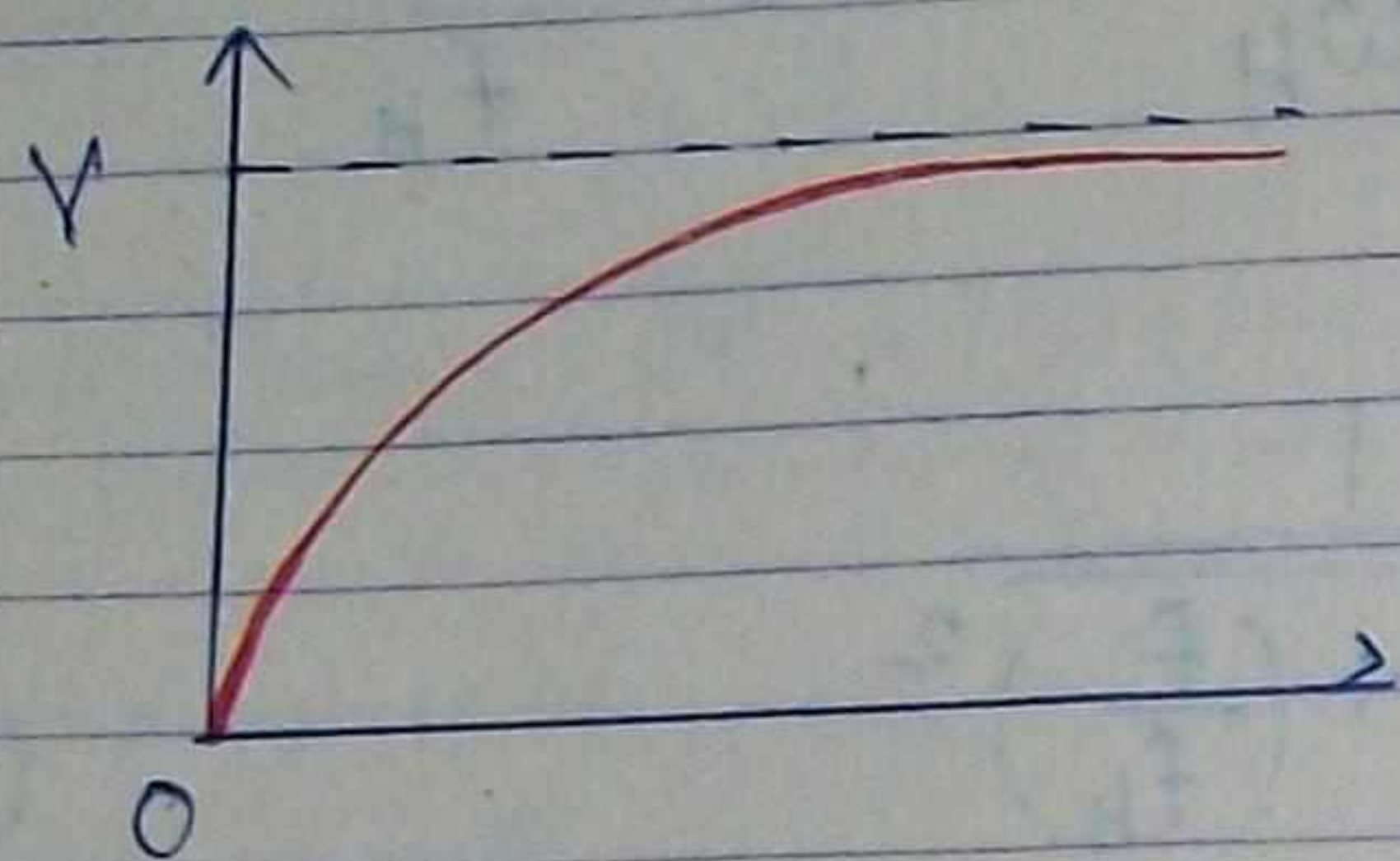
$\therefore f \uparrow \quad |A| \downarrow$

at $f = f_H$ gain = $\frac{\text{max gain}}{\sqrt{2}}$

$$\frac{1}{\sqrt{2}} = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}}$$

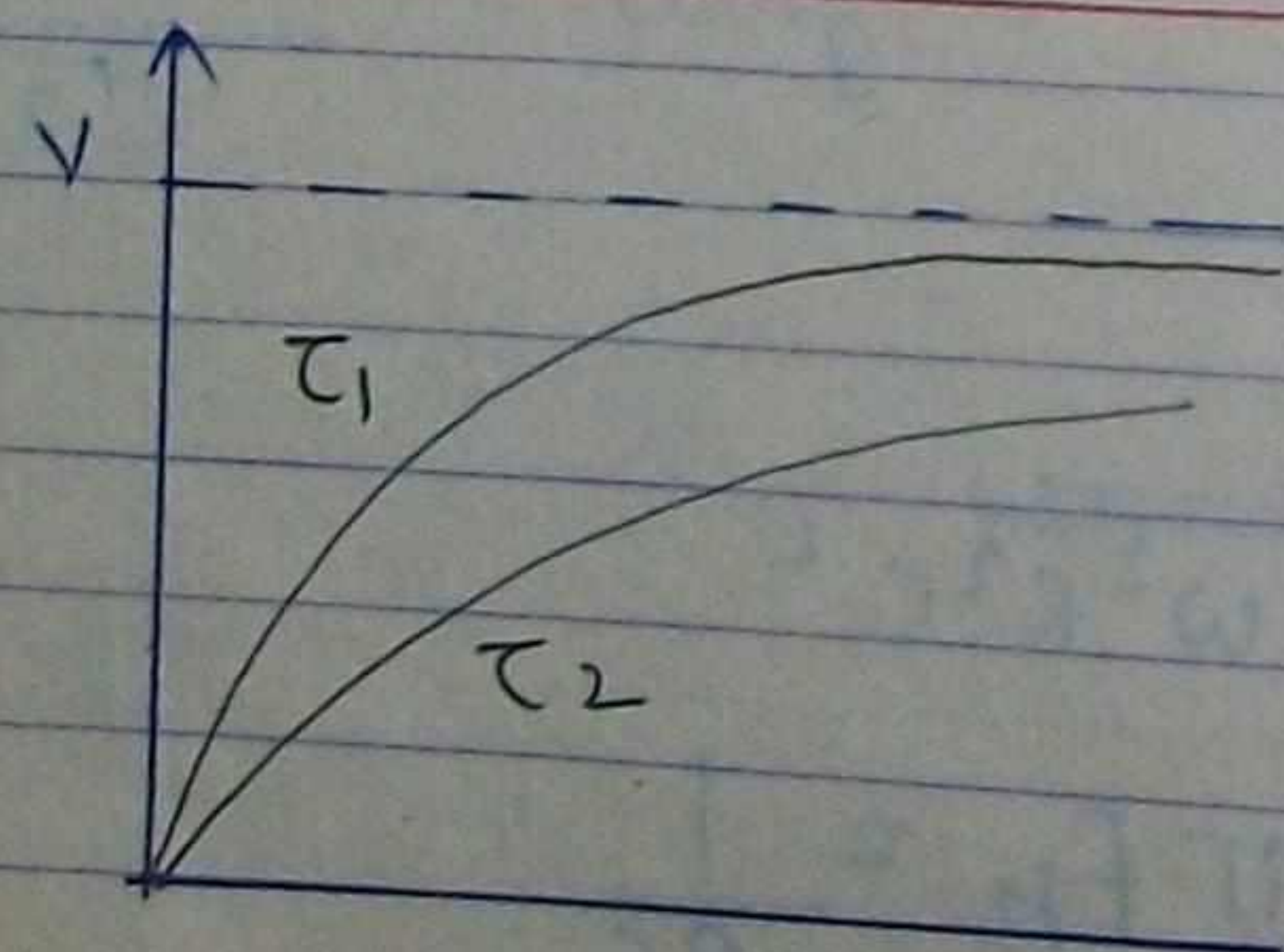
$$\omega_H = 2\pi f_H = \frac{1}{RC}$$

(A) STEP RESPONSE -



$$V_c(0^-) = V_c(0^+) = 0$$

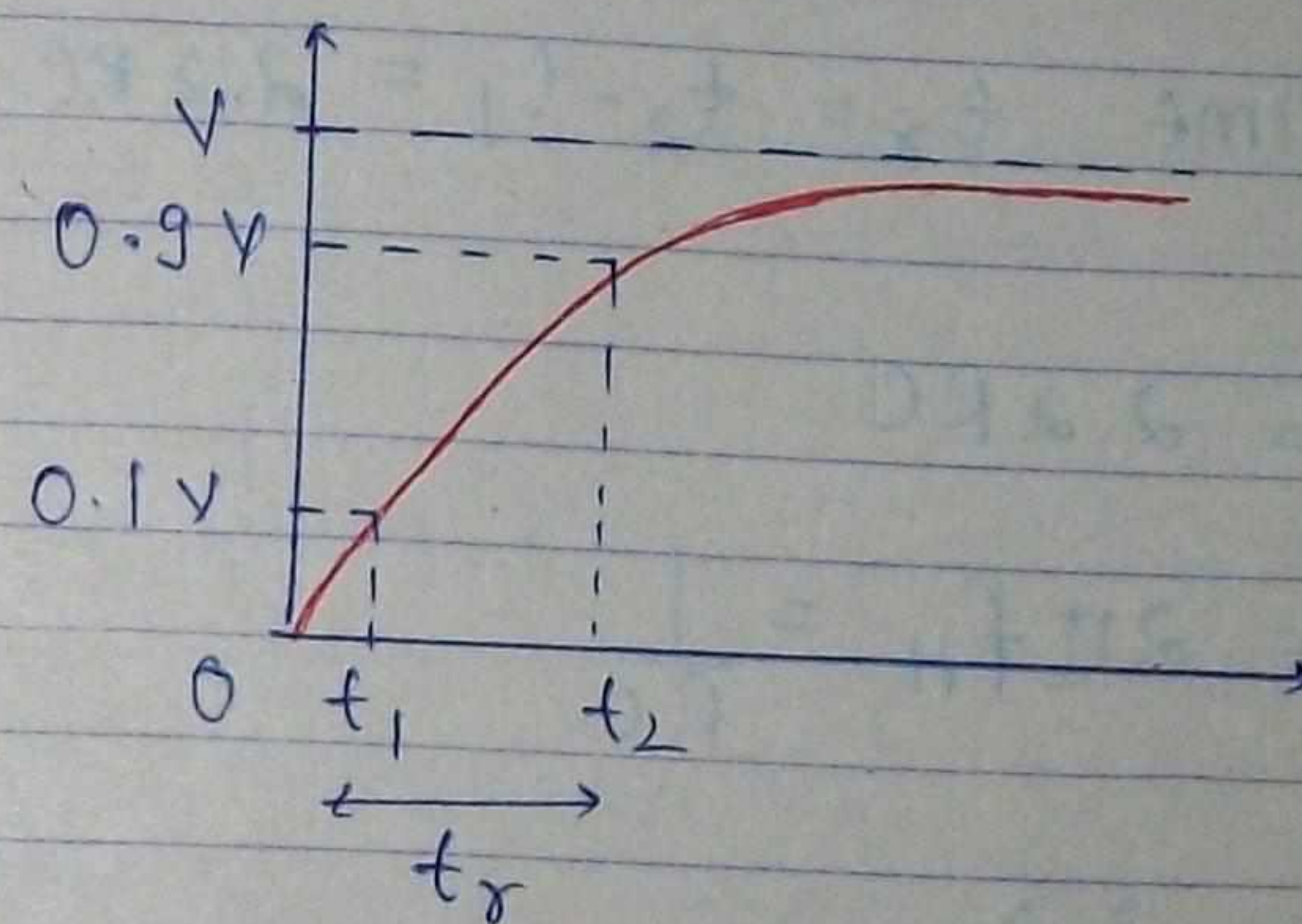
$$V_c = V_o = V [1 - e^{-t/RC}]$$



$$\tau_1 < \tau_2$$

RISE TIME -

Time taken by the signal to rise from 10% to 90% of its final value is called rise time.



* Rise time

$$t_r = t_2 - t_1$$

Since,

$$V_o = V [1 - e^{-t/RC}]$$

at $t = t_1$, $V_o = 0.1V$ (10%)

$$0.1V = V [1 - e^{-t_1/RC}]$$

$$t_1 = 0.1RC$$

at $t = t_2$, $V_o = 0.9V$ (90%)

$$0.9 V = V [1 - e^{-t_2/RC}]$$

$$t_2 = 2.3 RC$$

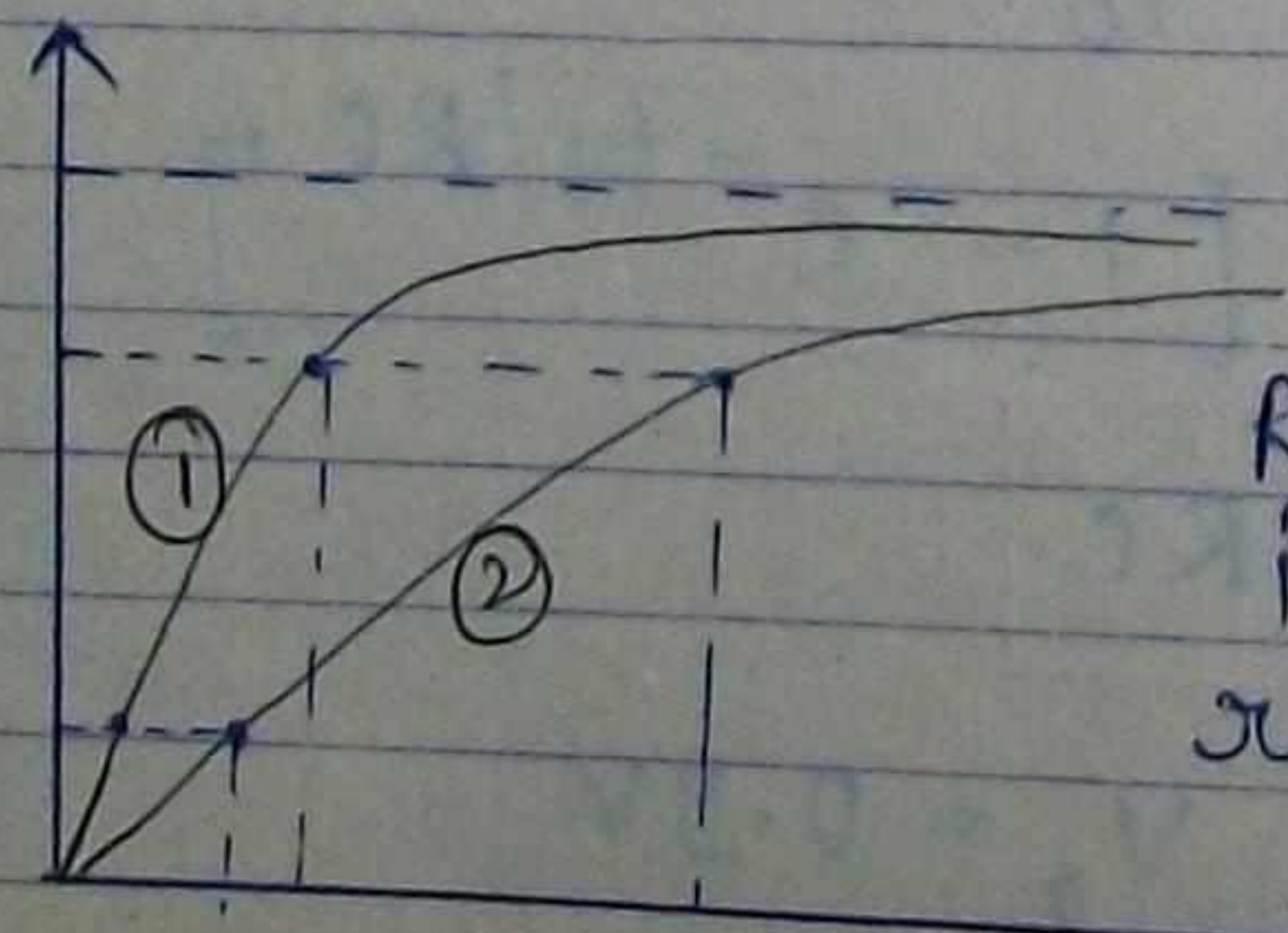
$$\therefore \text{Rise time } t_r = t_2 - t_1 = 2.2 RC$$

$$t_r = 2.2 RC$$

$$\text{Sinc } \omega_H = 2\pi f_H = \frac{1}{RC}$$

$$t_r = \frac{2.2}{2\pi f_H}$$

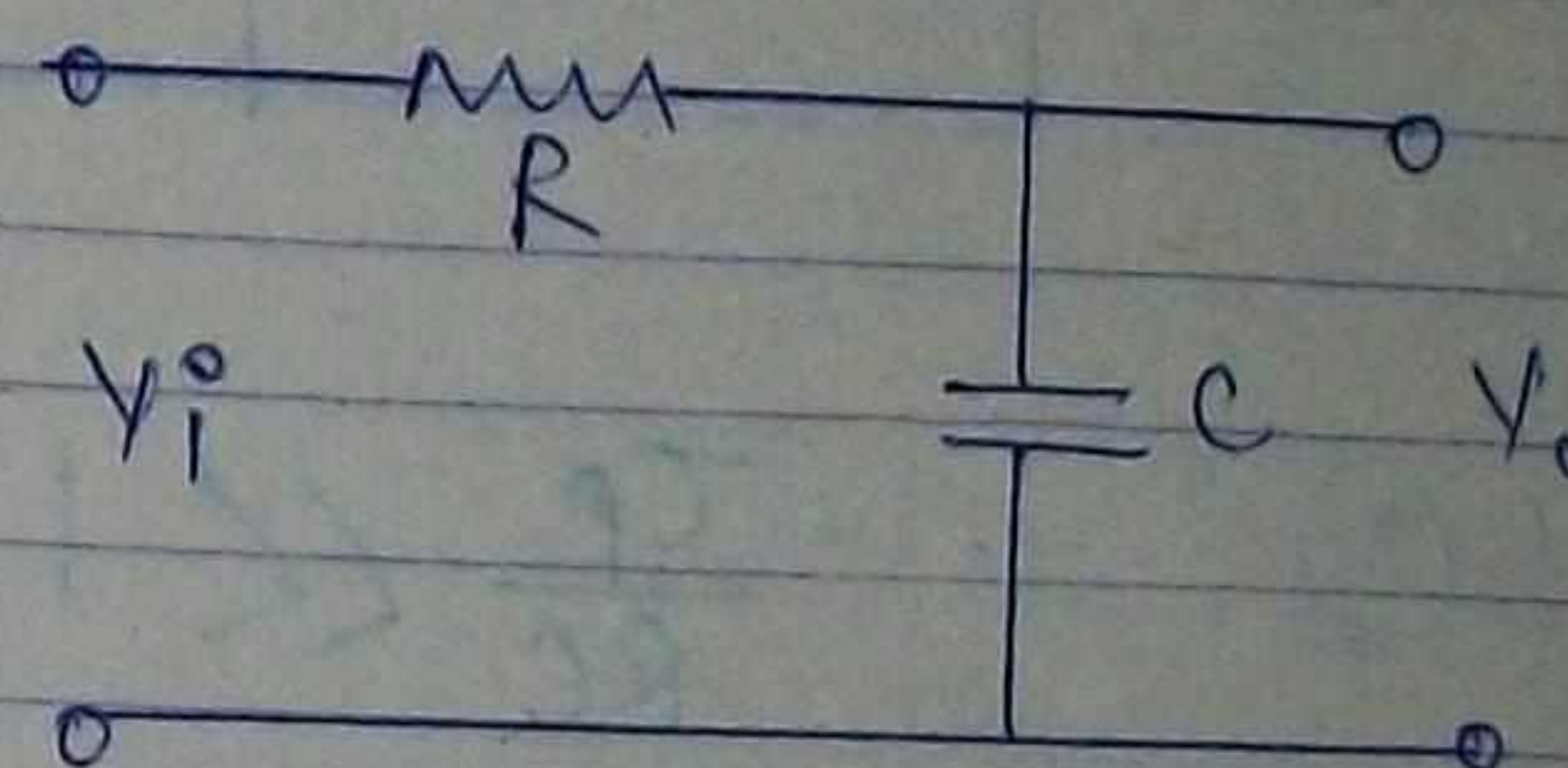
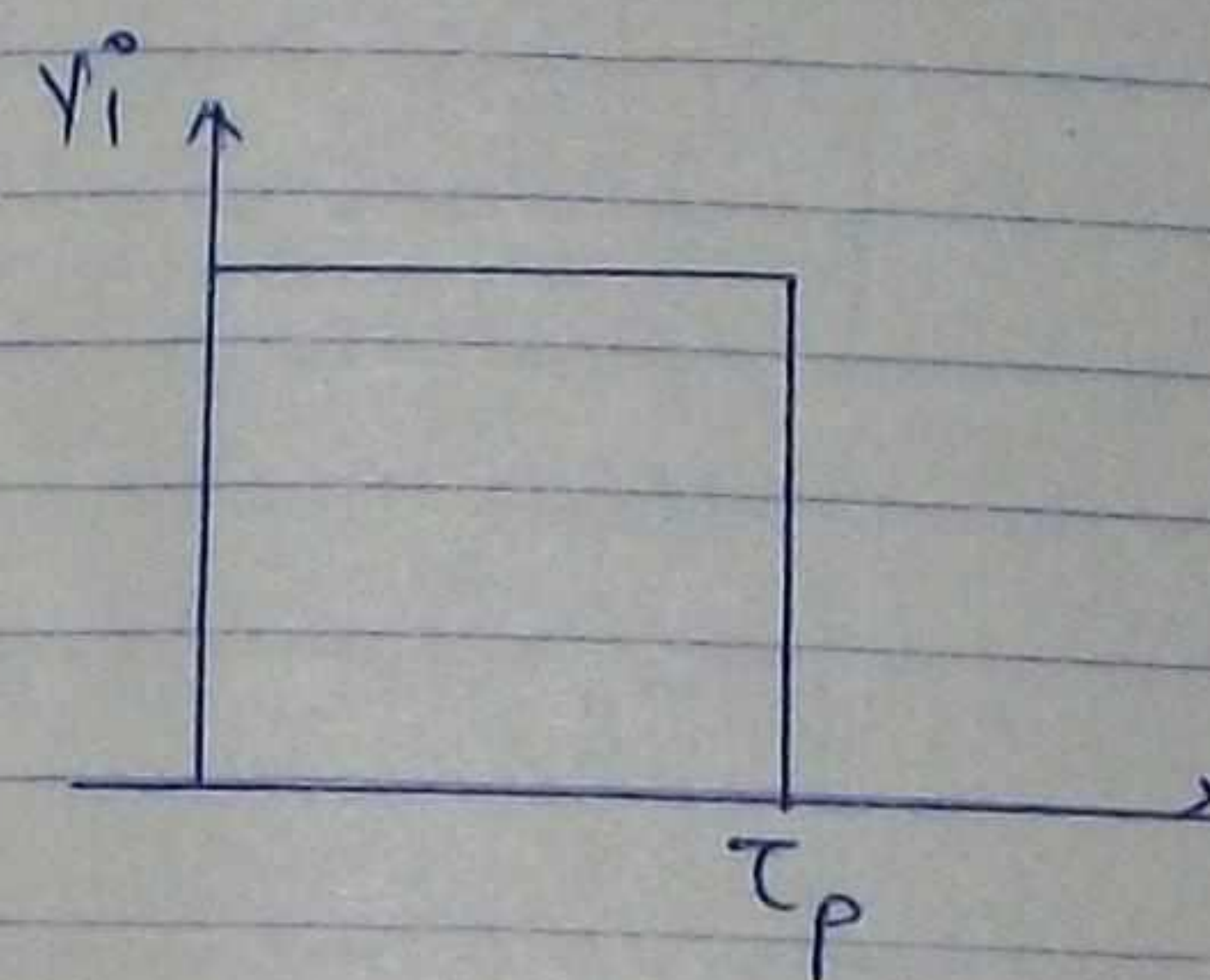
$$t_r = \frac{0.35}{f_H}$$



Response of ① is faster than response of ②

- Output have smaller rise time is faster.

⑧ PULSE RESPONSE -



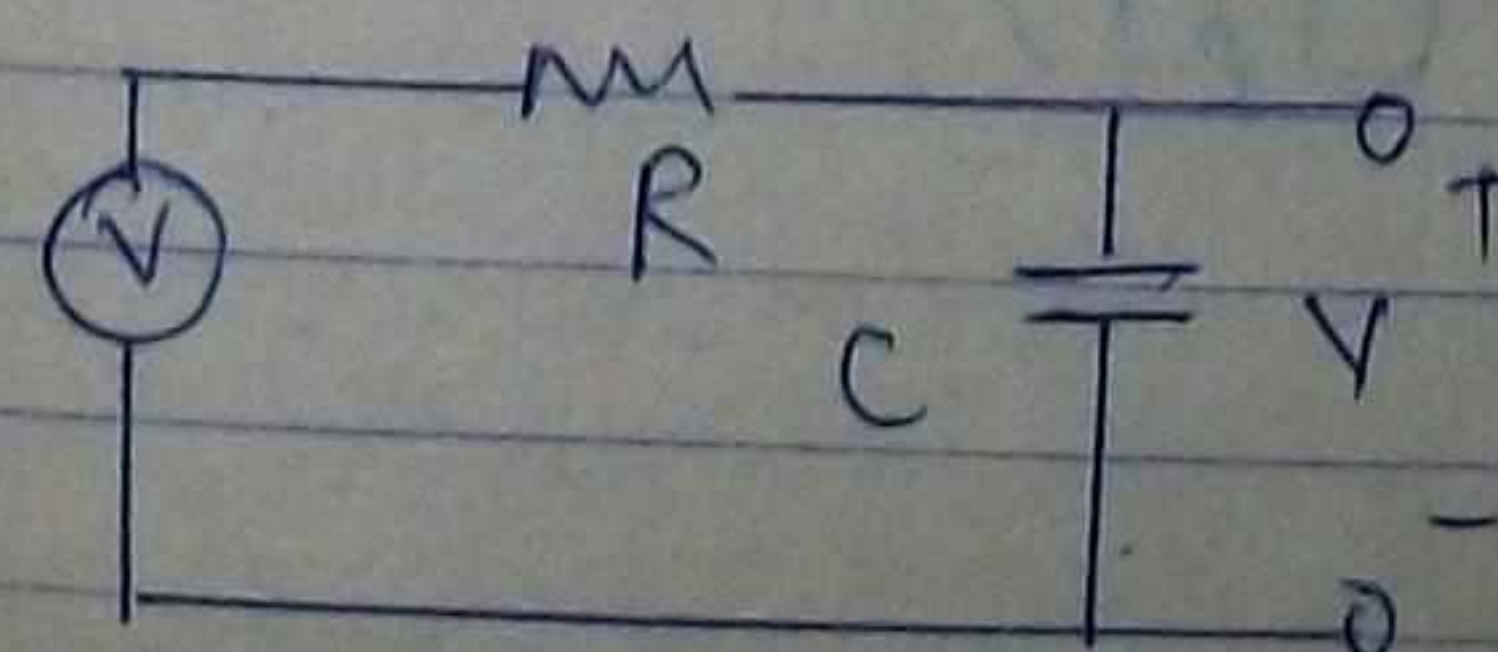
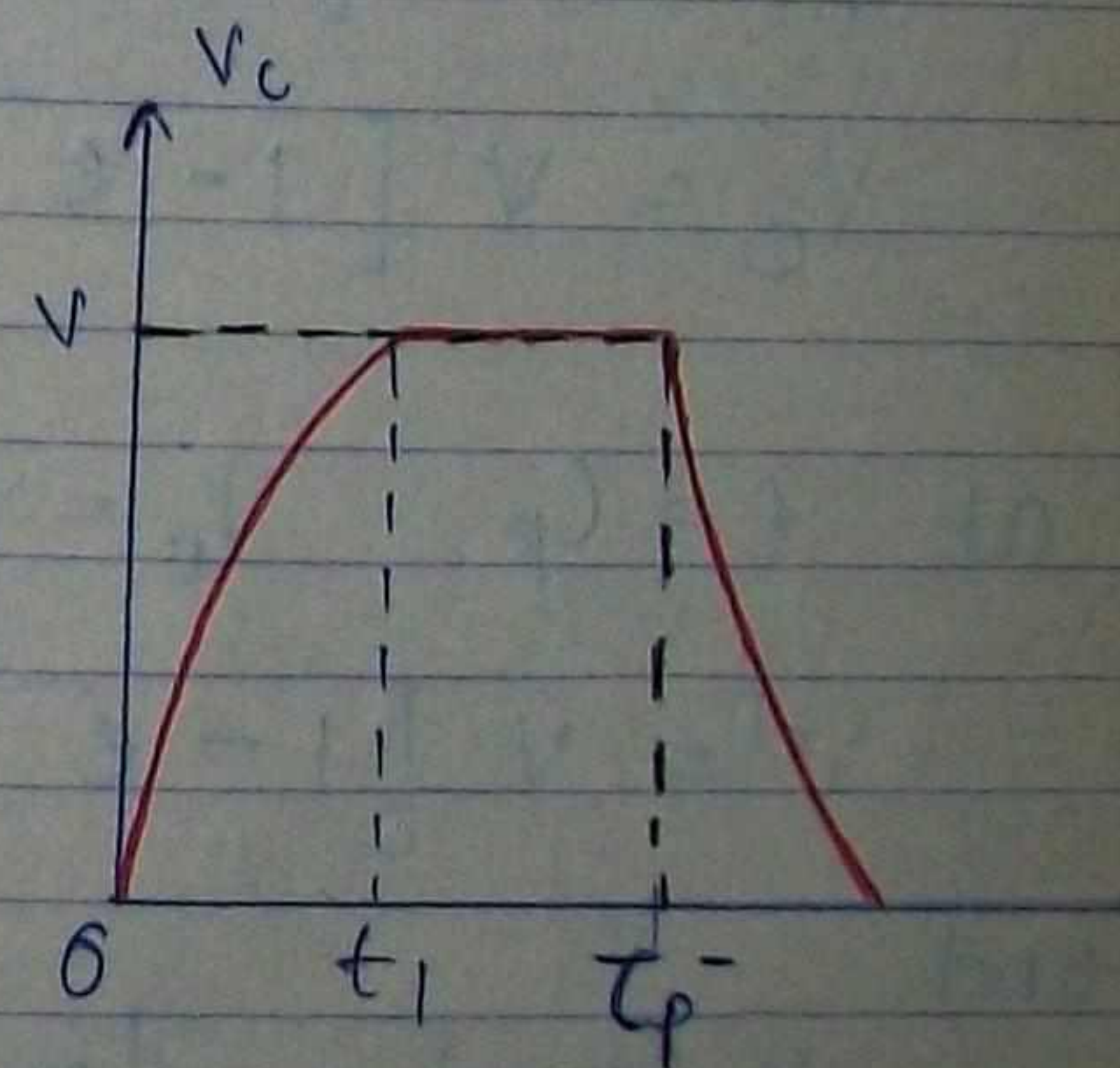
CASE ① : when $\frac{\tau_p}{RC} \gg 1$

$$RC \ll \tau_p$$

$$V_c(10^-) = V_c(10^+) = 0$$

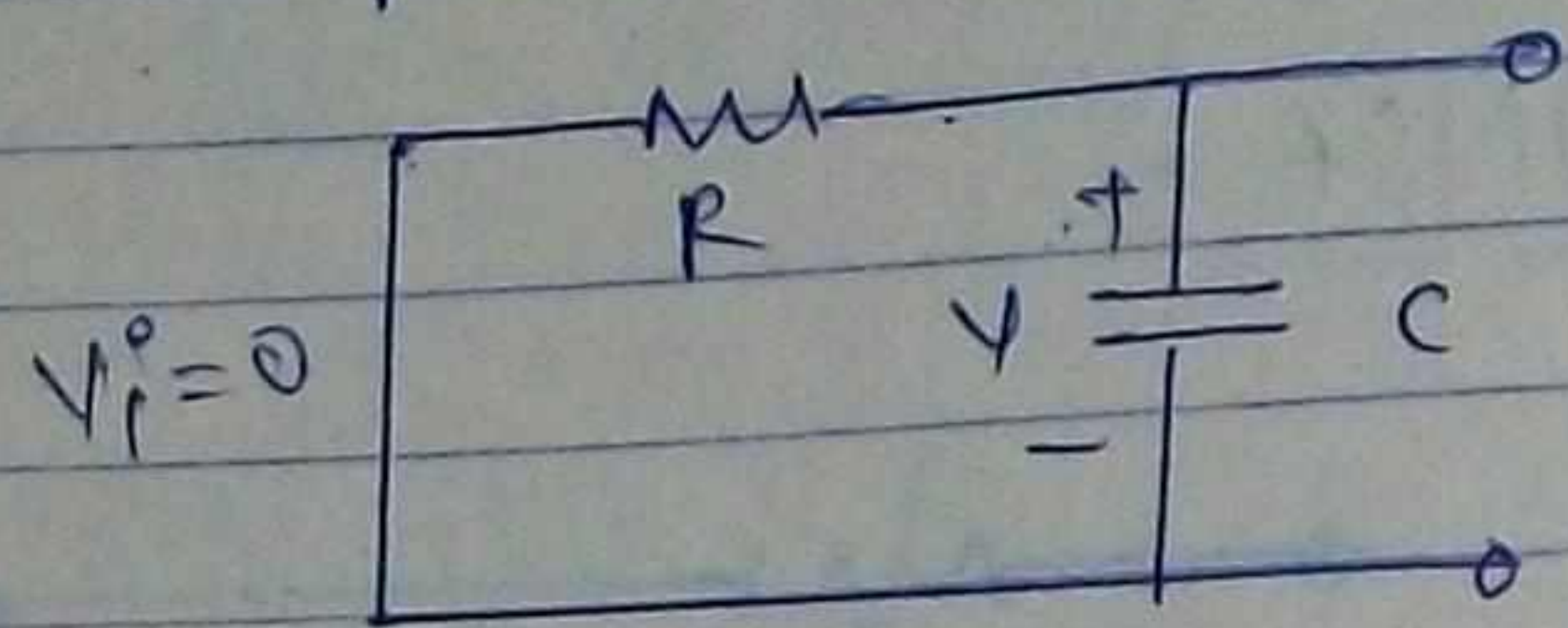
$$\text{at } t = t_1$$

$$V_c(t_1) = V$$



$$V_c(\tau_p^-) = V_c(\tau_p^+) = V$$

at $t = \tau_p^+$ $V_i^o = 0$



CASE ②: $\frac{\tau_p}{RC} \ll 1$

$$\tau_p \ll RC$$

$$V_c(0^-) = V_c(0^+) = 0$$

$$V_c = V [1 - e^{-t/RC}]$$

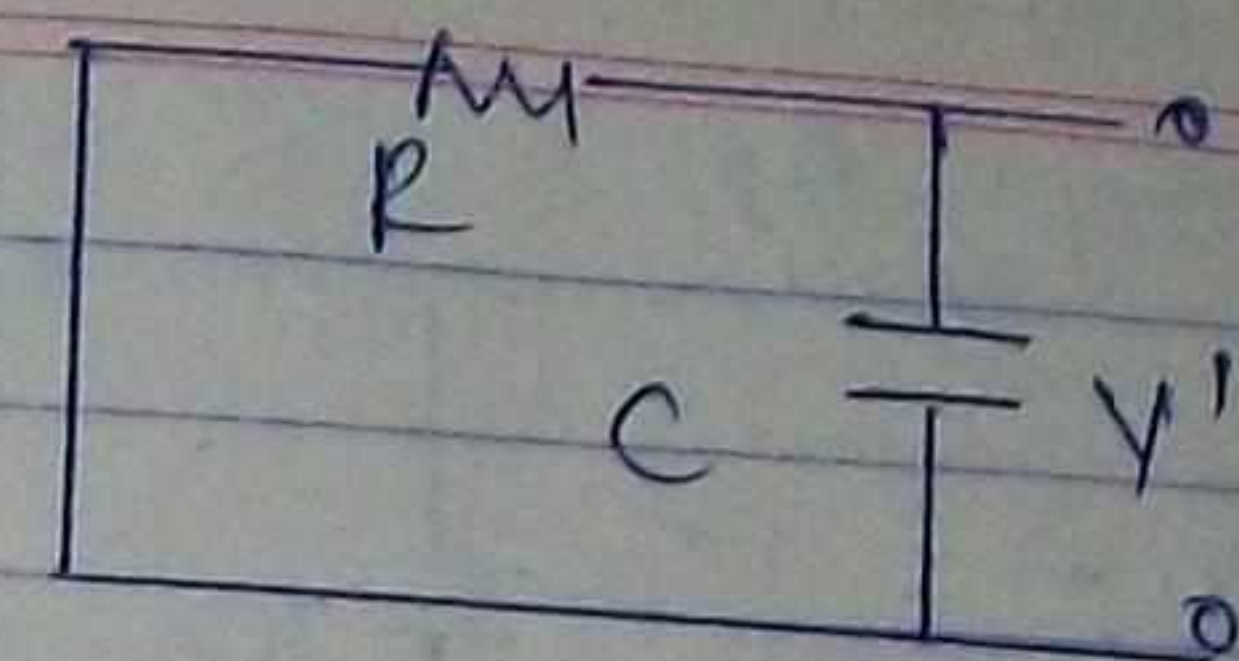
at $t = \tau_p$ $V_c = V'$

$$V' = V [1 - e^{-\tau_p/RC}]$$

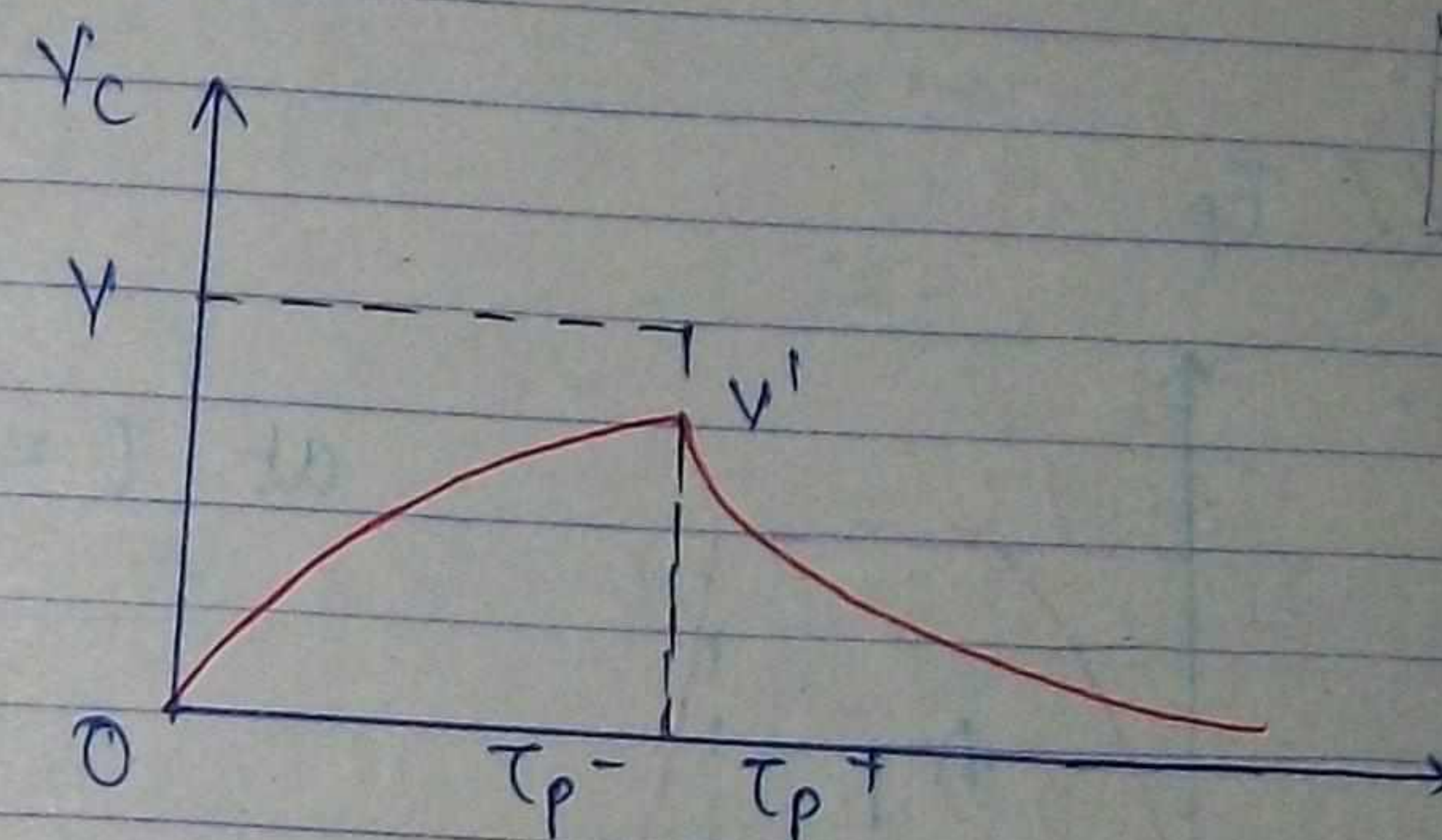
and

$$V_c(\tau_p^-) = V' = V_c(\tau_p^+)$$

$t = \tau_p^+$ $V_i^o = 0$



$$V_c = V' e^{-(t-\tau_p)/RC}$$

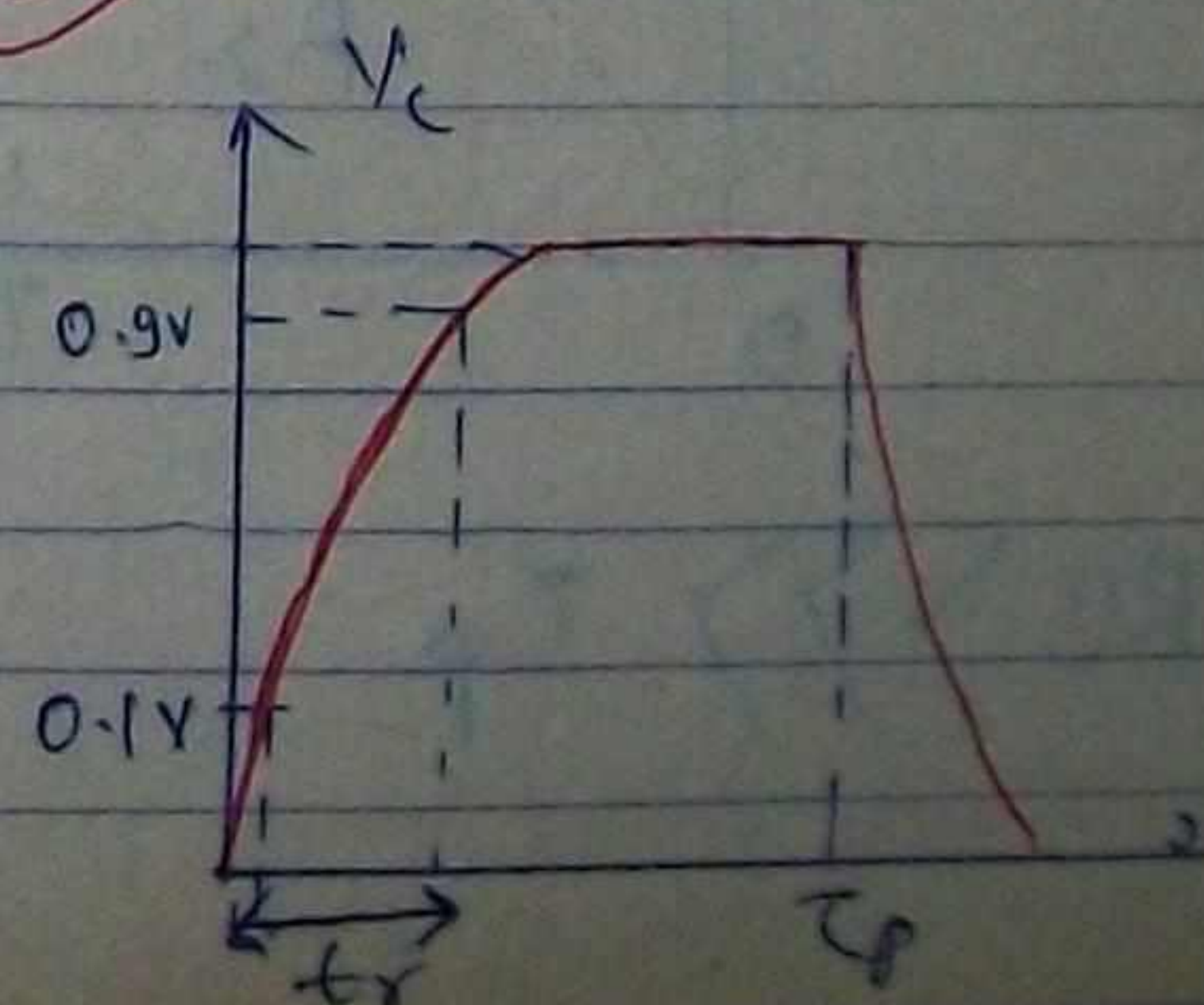


The pulse shape will be preserved if:

$$f_H = \frac{1}{\tau_p}$$

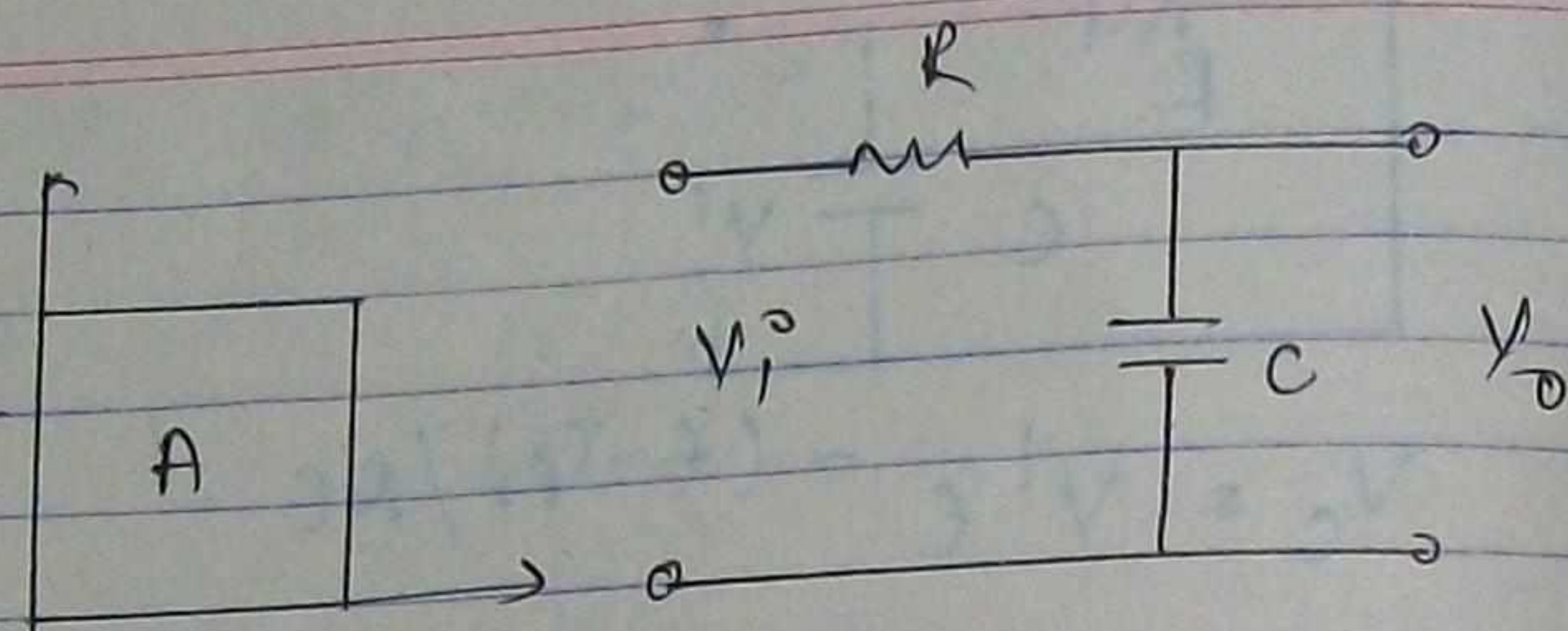
Since, $t_r = \frac{0.35}{f_H}$, $f_H = \frac{0.35}{t_r}$

$$t_r = 0.35 \tau_p$$

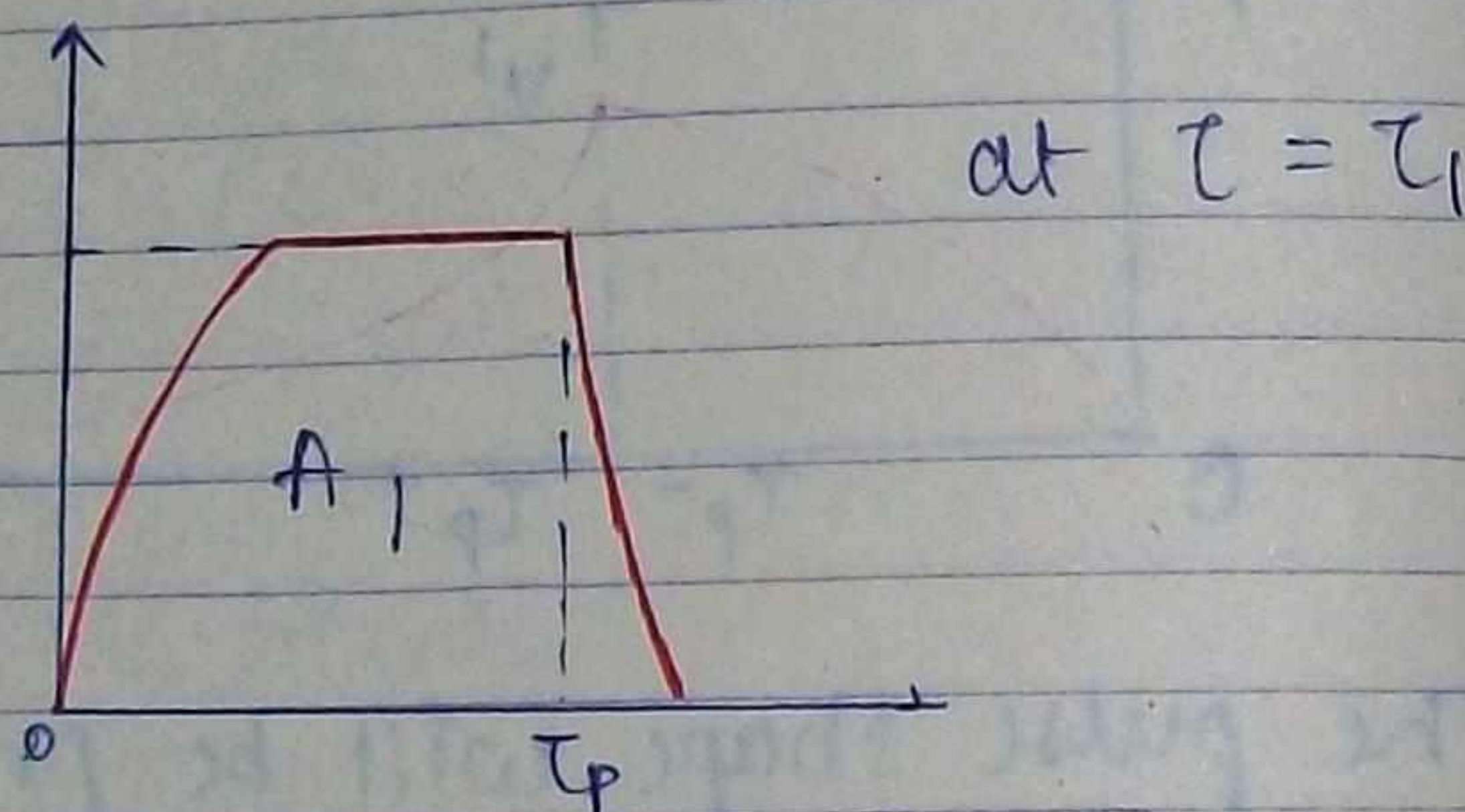


$$t_r \leq 0.35 \tau_p$$

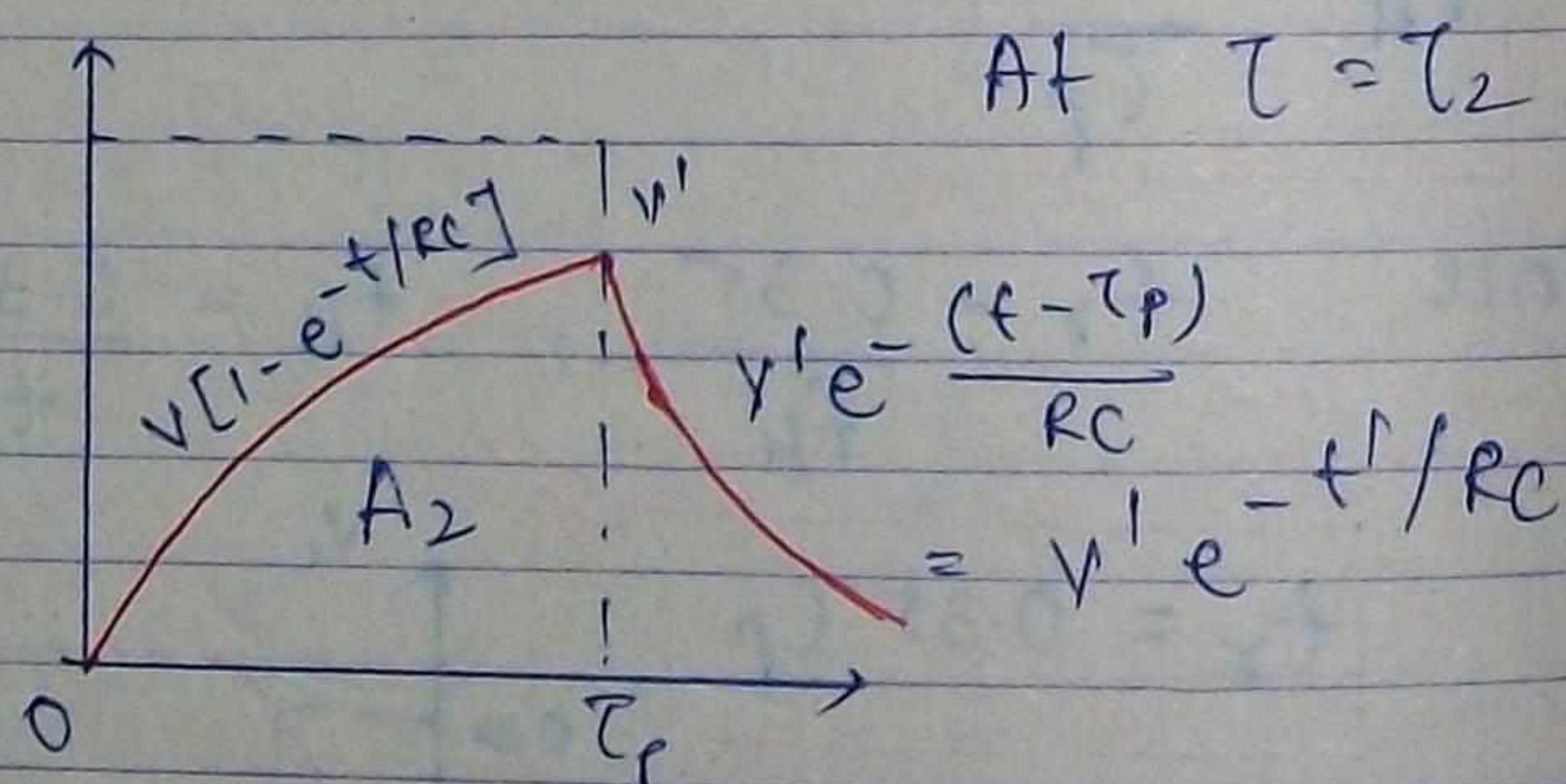
Que.



① $RC \ll \tau_p$



② $RC \gg \tau_p$



③ $RC \gg \gg \tau_p$

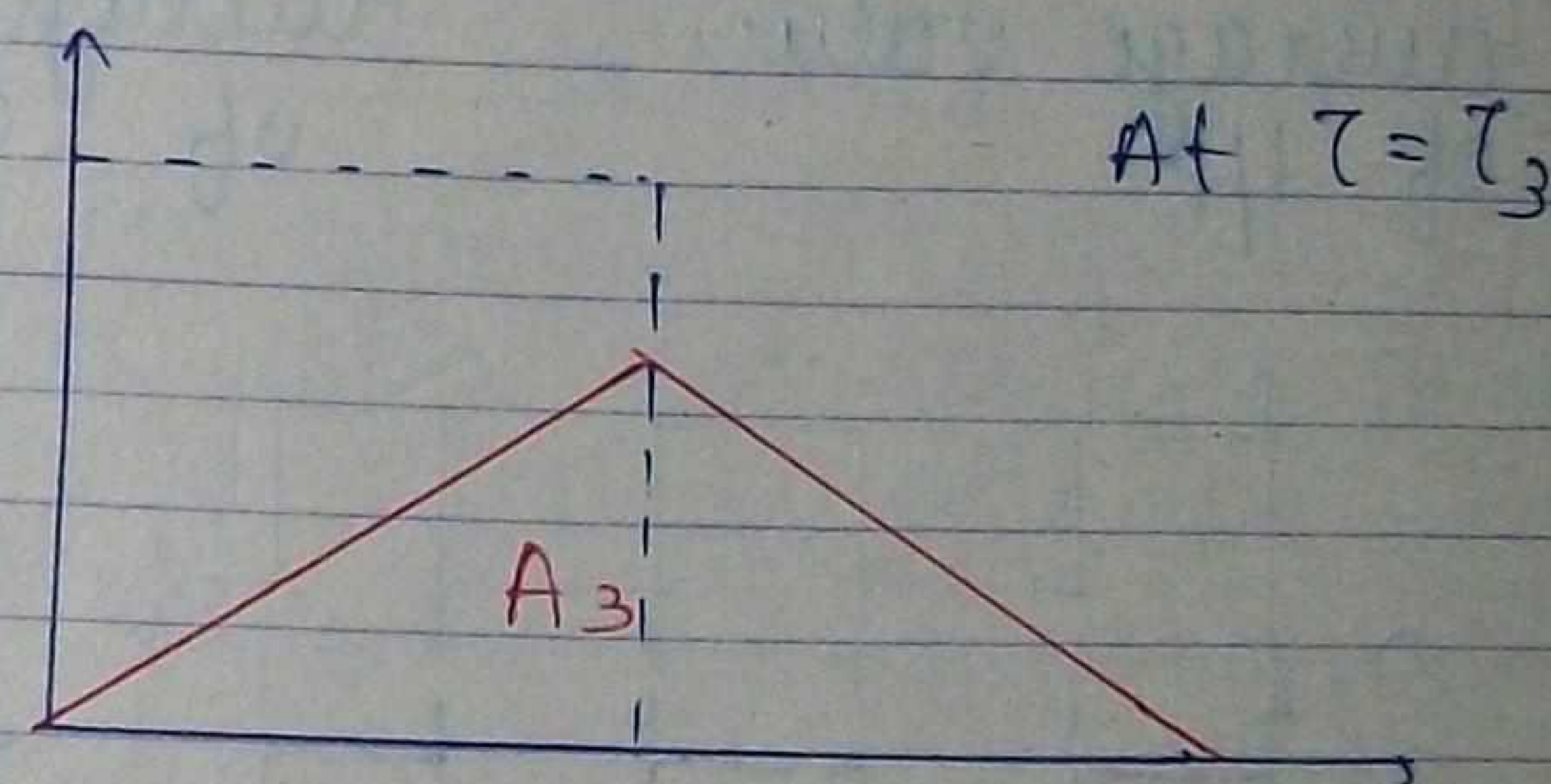
$$x = \frac{t}{RC} = \text{very very low.}$$

$$e^{-x} = [1 - x], \quad V_c = V[1 - e^{-t/RC}]$$

$$V_c = \frac{Vt}{RC} \quad (\text{charging})$$

$$\text{also, } V_c = V'e^{-t'/RC}$$

$$V_c = V'[1 - \frac{t'}{RC}] \quad (\text{discharging})$$

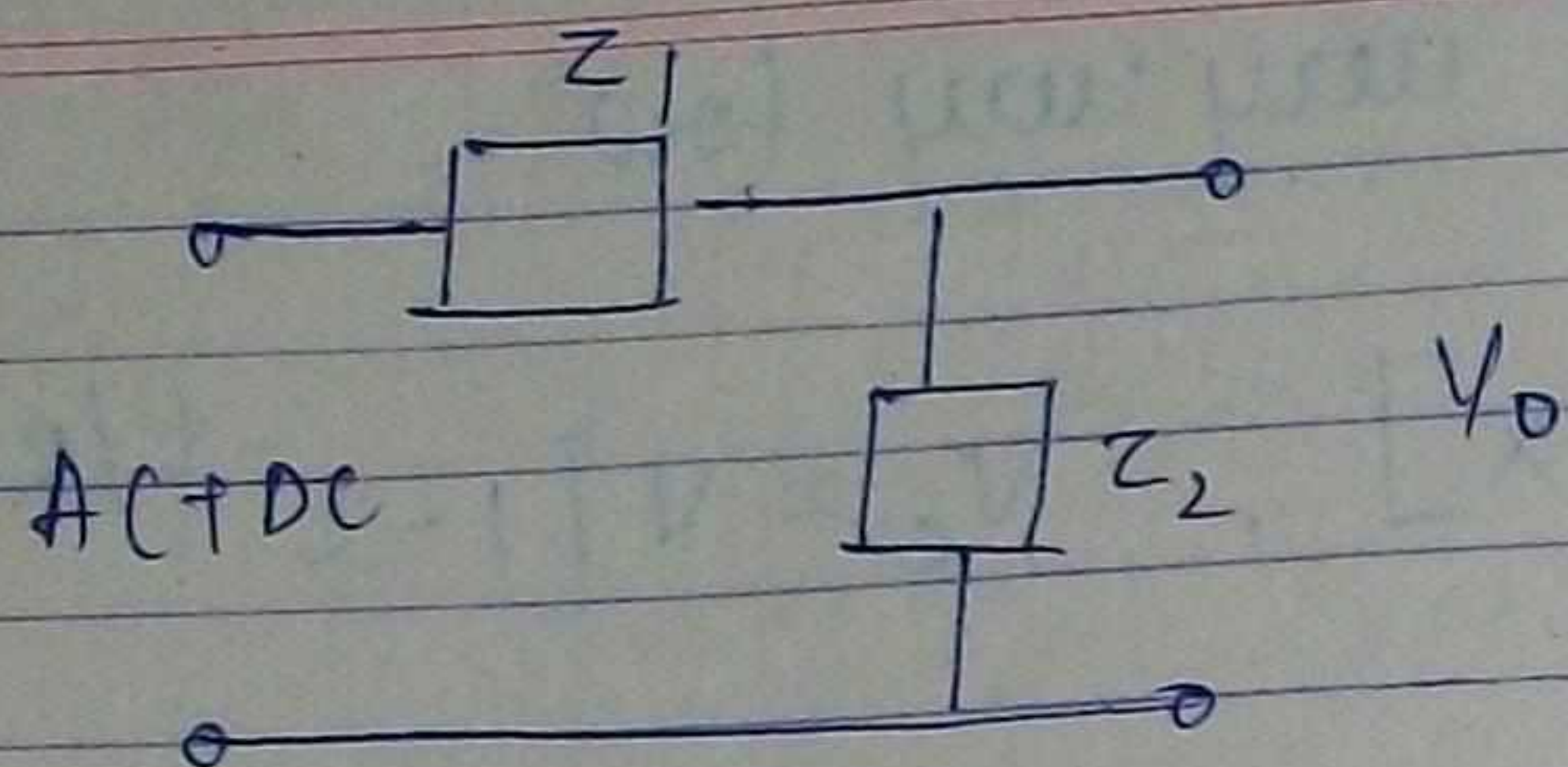


① ARRANGE τ_1, τ_2, τ_3 in ascending order:

$$\tau_1 < \tau_2 < \tau_3$$

② Relation b/w A_1, A_2, A_3 :

$$A = A_1 = A_2 = A_3$$



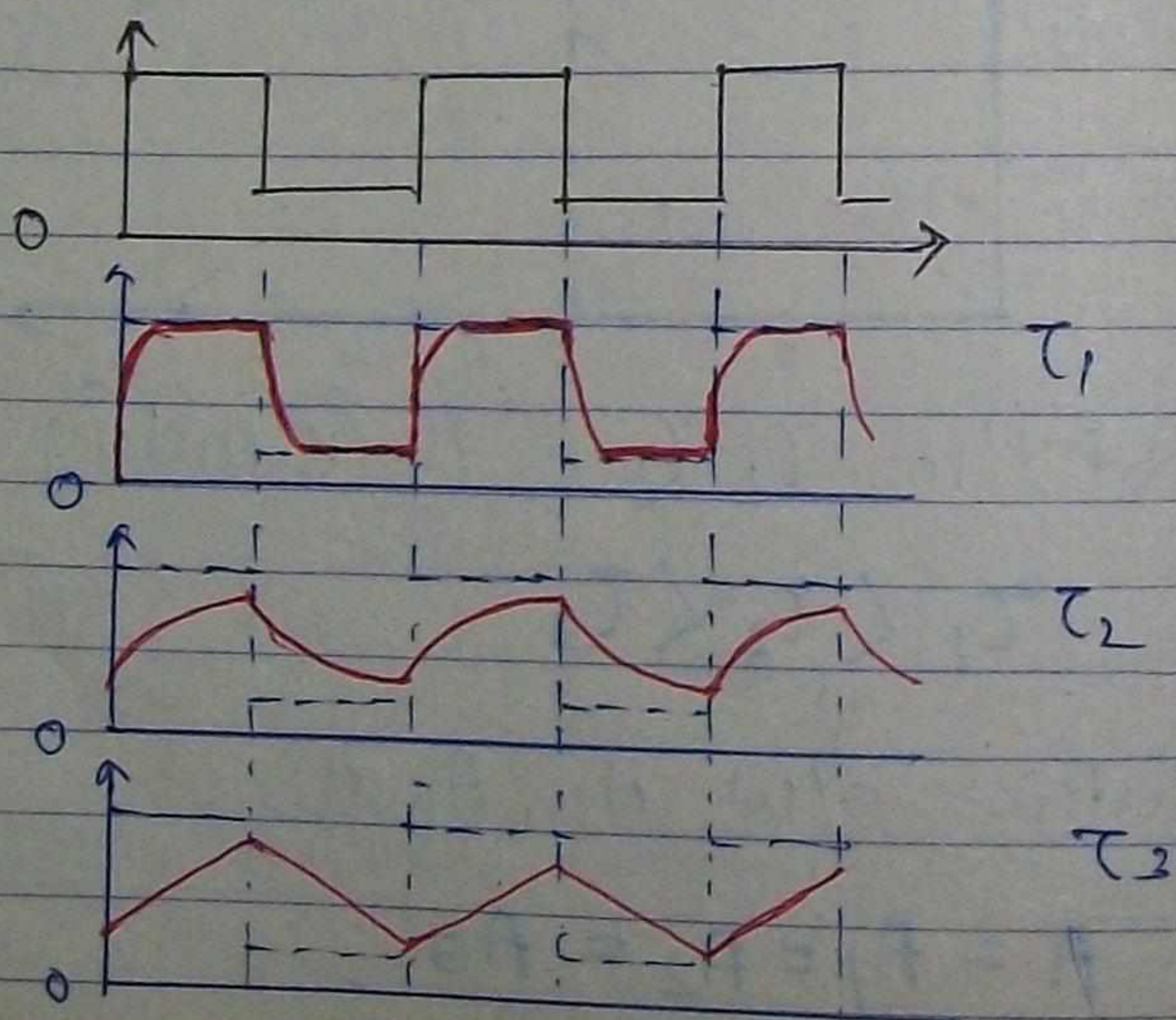
$$V_0 = \frac{Z_2}{Z_1 + Z_2} V_i$$

$$Z_C = \frac{1}{2\pi fC}$$

$$Z_C = \infty \text{ for dc } (f=0)$$

∴ average value = average value of I/P.

Ques.



① Arrange τ_1, τ_2, τ_3 in ascending order:
 $\tau_1 < \tau_2 < \tau_3$

② average value:

shown in figure

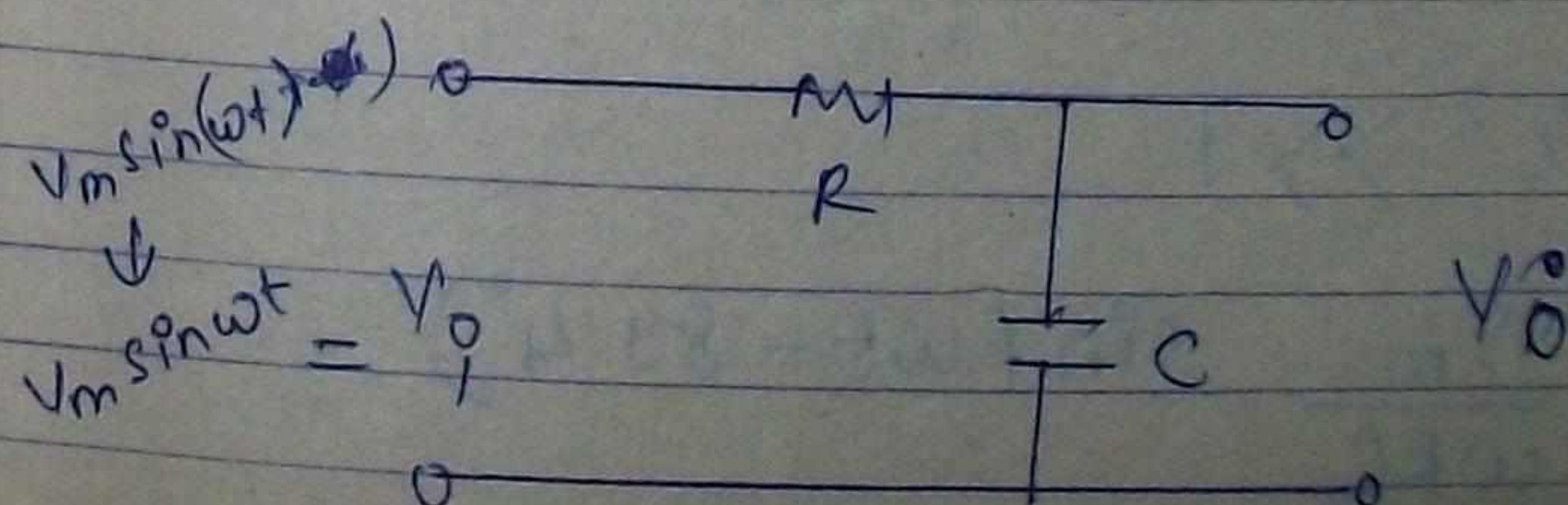
average value of I/P	=	average value of O/P.
----------------------	---	-----------------------

⇒ LOW PASS RC CIRCUIT AS AN INTEGRATOR -

- when time constant is very large in comparison with time period of the I/P signal then circuit is called an integrator.

$$RC \gg T_p$$

⇒ CRITERIA FOR GOOD INTEGRATOR -



$$|A| = \frac{1}{\sqrt{1 + \omega^2 R^2 C^2}}$$

$$\phi = -\tan^{-1} \omega RC$$

$$V_o = |A| V_m \sin(\omega t + \phi)$$

$$V_o = \frac{V_m}{\sqrt{1 + \omega^2 R^2 C^2}} \sin(\omega t + \phi)$$

	$ \phi $	$\omega RC = \frac{2\pi}{T} \times RC$
IDEAL INTEGRATOR	90°	∞ (practically not possible)

BEST INTGR.	89.4	$RC > 15T$
-------------	------	------------

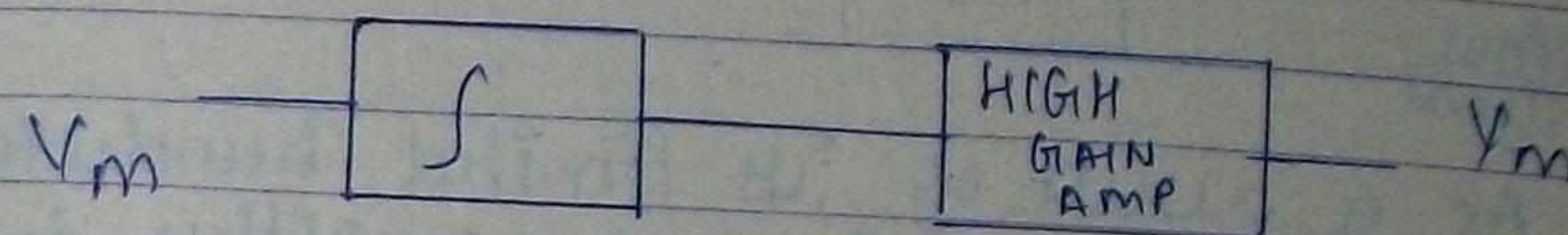
Let, $\phi = -89.4$ $RC > 15T$

$$\omega^2 R^2 C^2 \gg 1$$

$$V_o = \frac{V_m}{\omega RC} \sin(\omega t - 89.4^\circ)$$

Amplitude of output is very small

CONCLUSION -



- Integrator is also followed by a high gain amplifier.

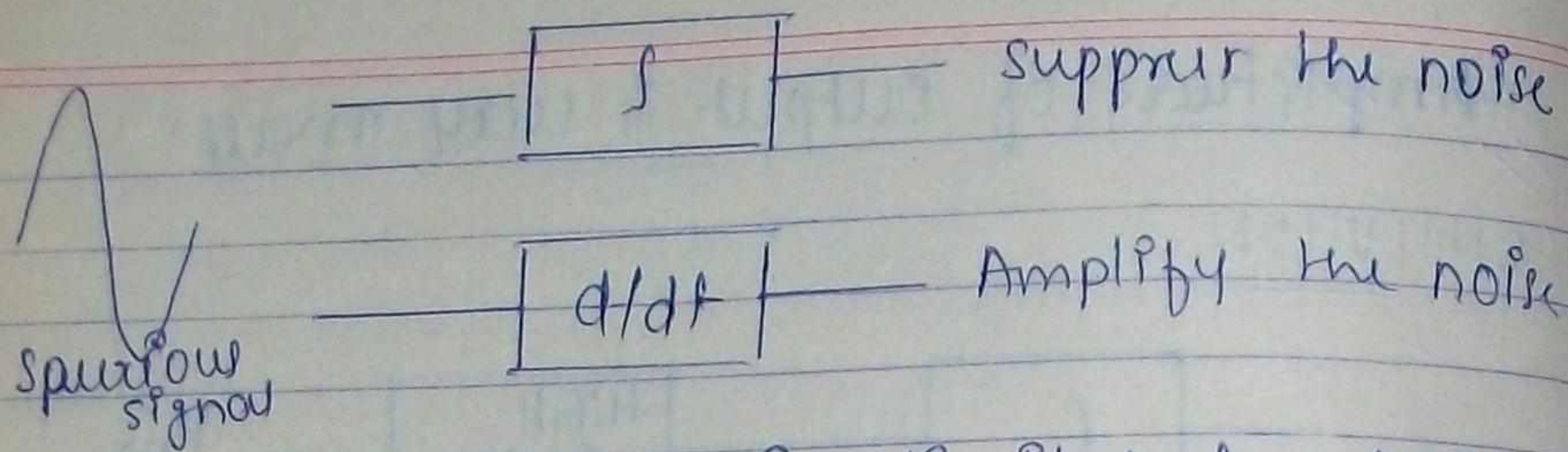
NOTE -

$$\frac{d}{dt} \rightarrow s \rightarrow j2\pi f$$

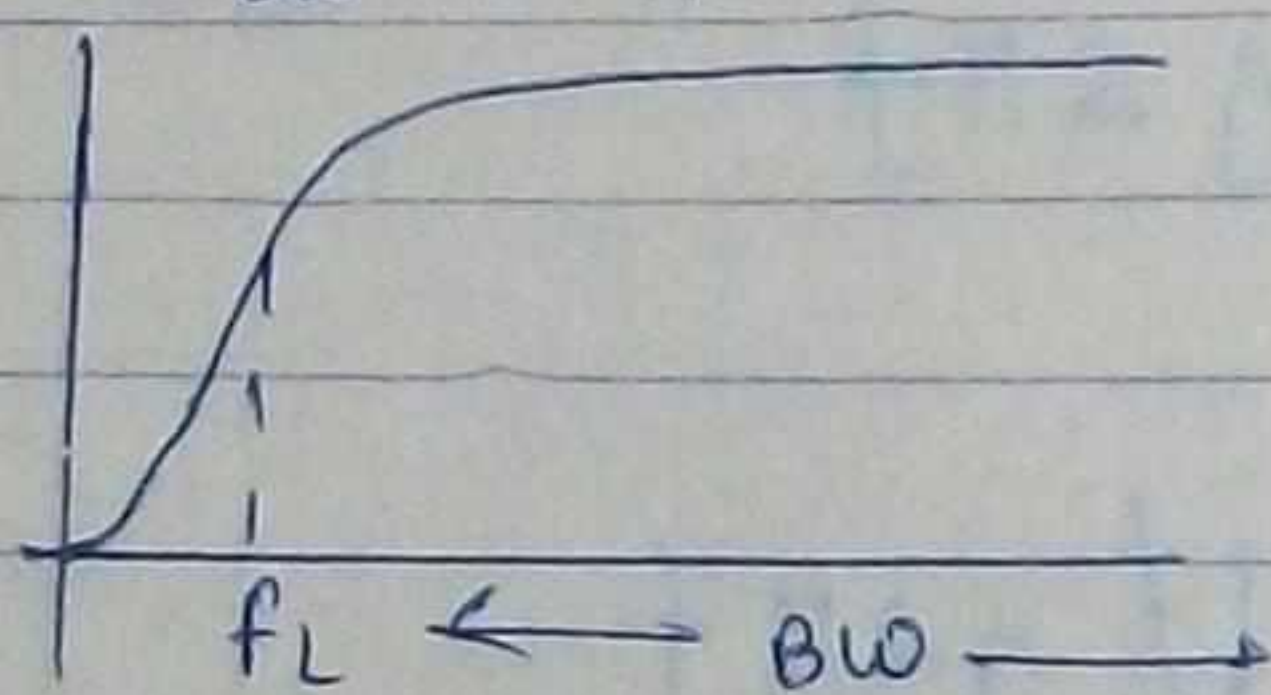
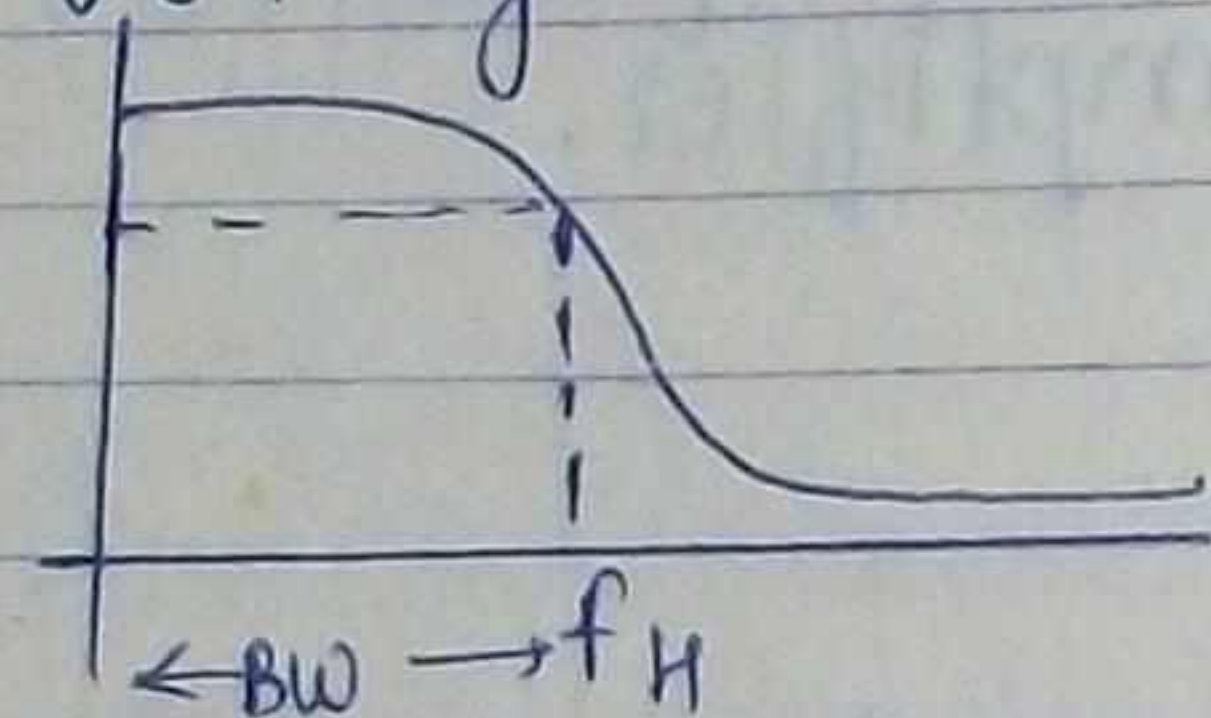
$$\int \rightarrow 1/s \rightarrow 1/j2\pi f$$

Integrator is ^{invariably} preferred over differentiator because:

- Since the gain of an integrator decreases with frequency whereas the gain of a differentiator increases with frequency, therefore it is easier to stabilize an integrator than a differentiator w.r.t. spurious oscillation (high frequency noise).

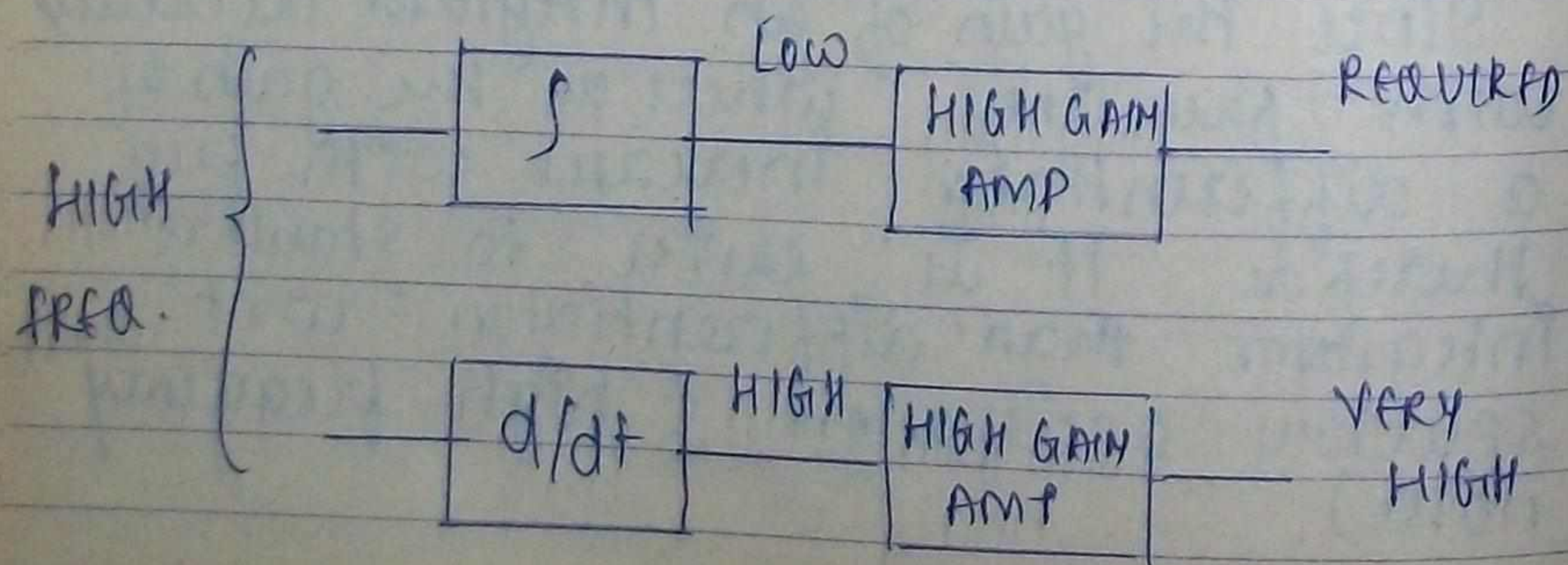


- ② As a result of its limited bandwidth an integrator is less sensitive to noise voltage than a differentiator



Noise $\propto$ BW

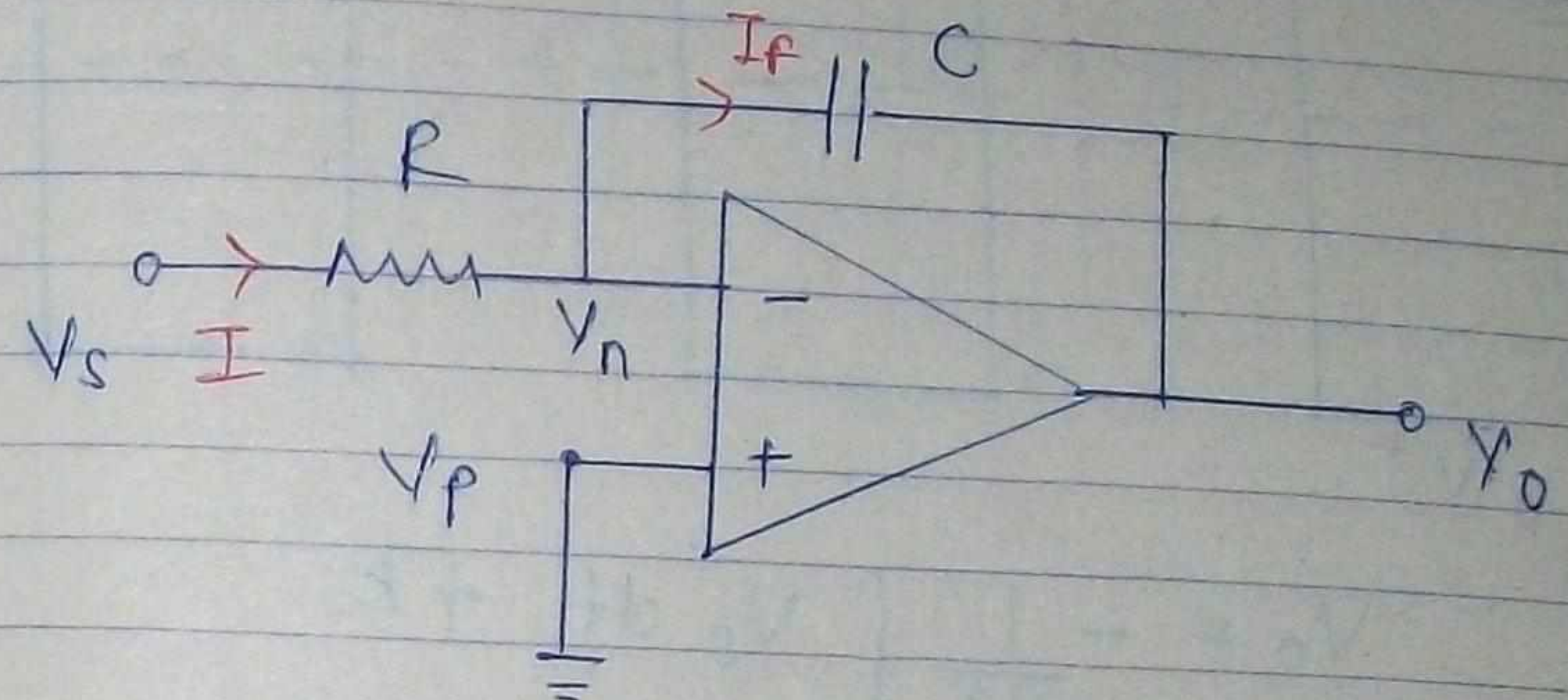
- ③ If input waveform changes very rapidly (high frequency signal) the amplifier of a differentiator may overload.



OP-AMP

CONT...

⇒ INTEGRATOR -



$$V_p = V_n = 0 \quad (\text{virtual ground})$$

$$I = I_f$$

$$\frac{V_s - V_n}{R} = C \frac{d[V_n - V_o]}{dt}$$

$$\frac{V_s}{R} = -C \frac{dV_o}{dt}$$

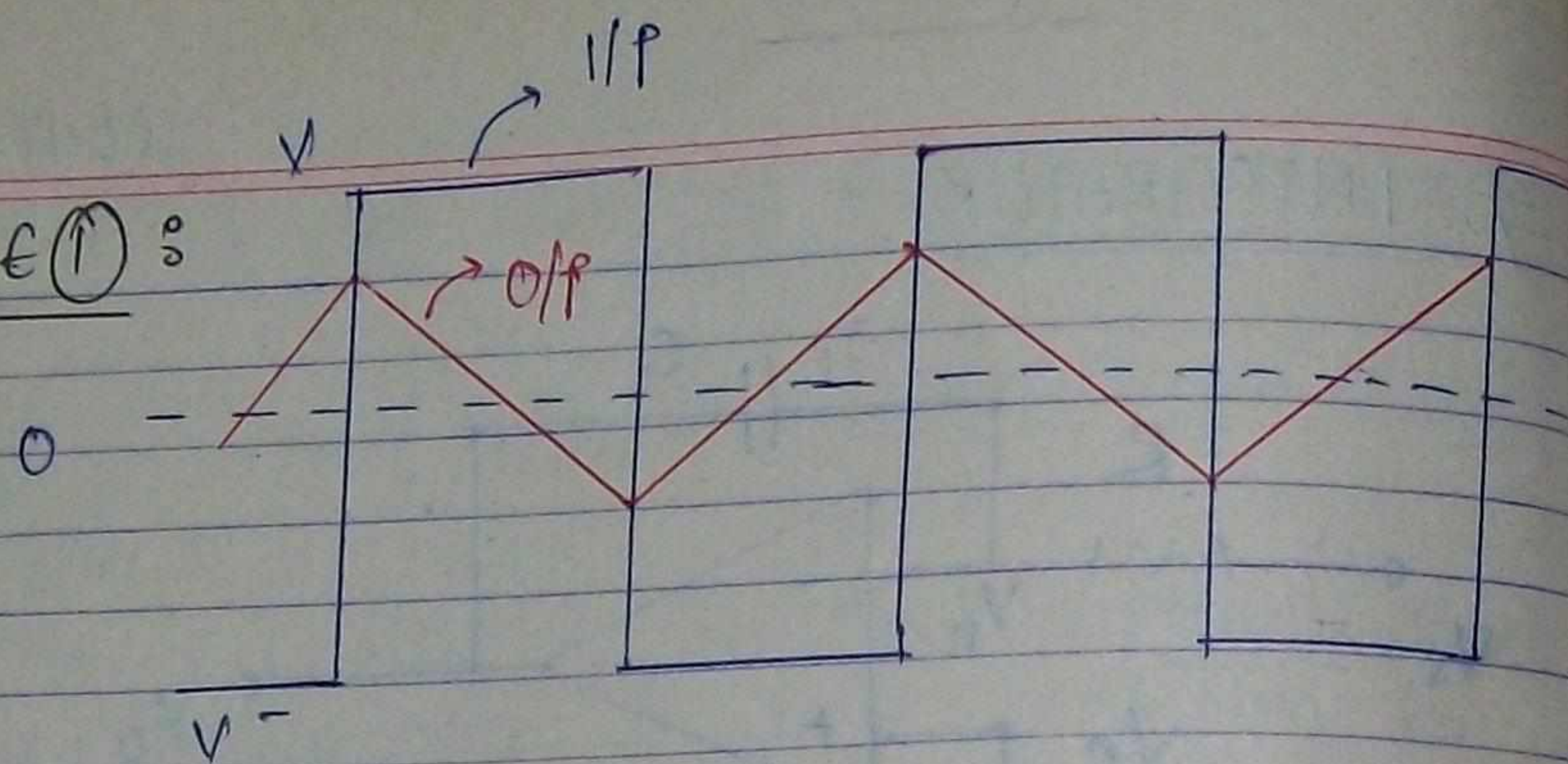
$$V_o = -\frac{1}{RC} \int V_s dt + K$$

CONSTANT

Since -ve sign comes hence called

Inverting Integrator

CASE (1) :



Since

$$V_o = -\frac{1}{RC} \int V_s dt + K$$

(or)

$$\frac{dV_o}{dt} = -\frac{V_s}{RC}$$

Rate of change of o/p voltage = slope of o/p.

(1) $V_s = V$

$$\frac{dV_o}{dt} = -\frac{V}{RC}$$

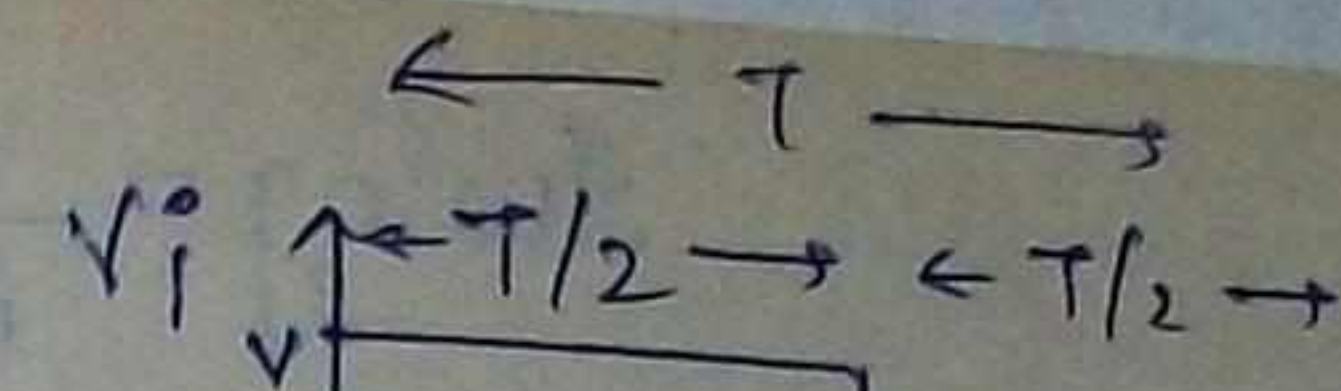
$V_o \downarrow$

(2) $V_s = -V$

$$\frac{dV_o}{dt} = \frac{V}{RC}$$

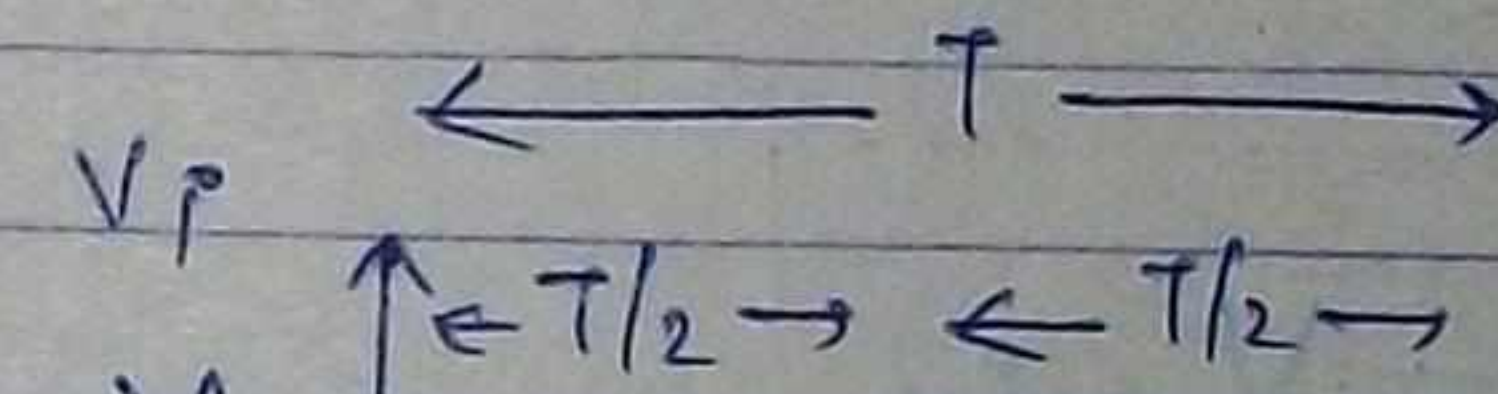
$V_o \uparrow$

(1)



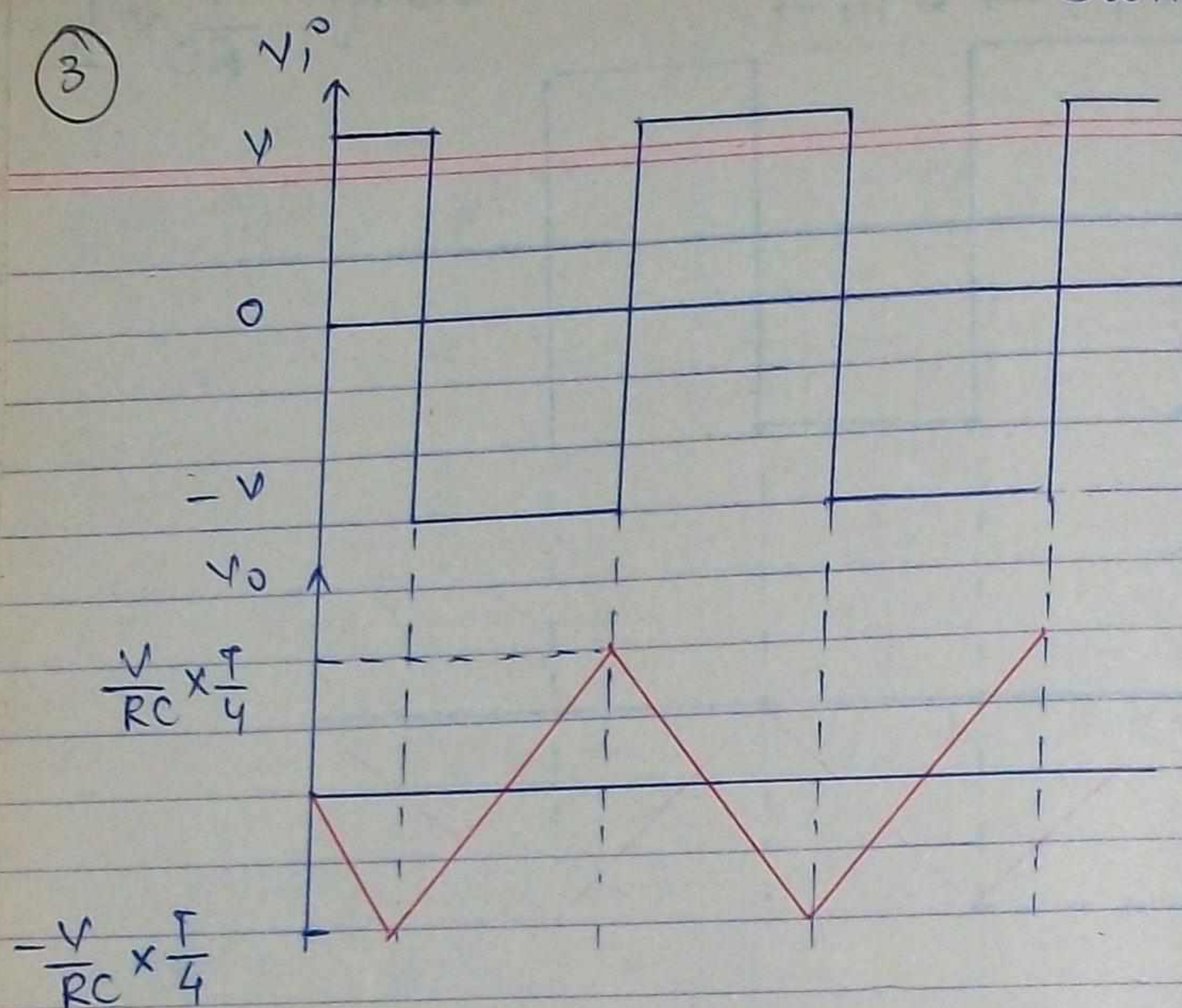
$$\text{Swing} = \frac{V}{RC} \times \frac{T}{2}$$

(2)



$$\text{Swing} = \frac{V}{RC} \times \frac{T}{2}$$

3



$$\text{Swing} = \frac{V}{RC} \times T/2$$

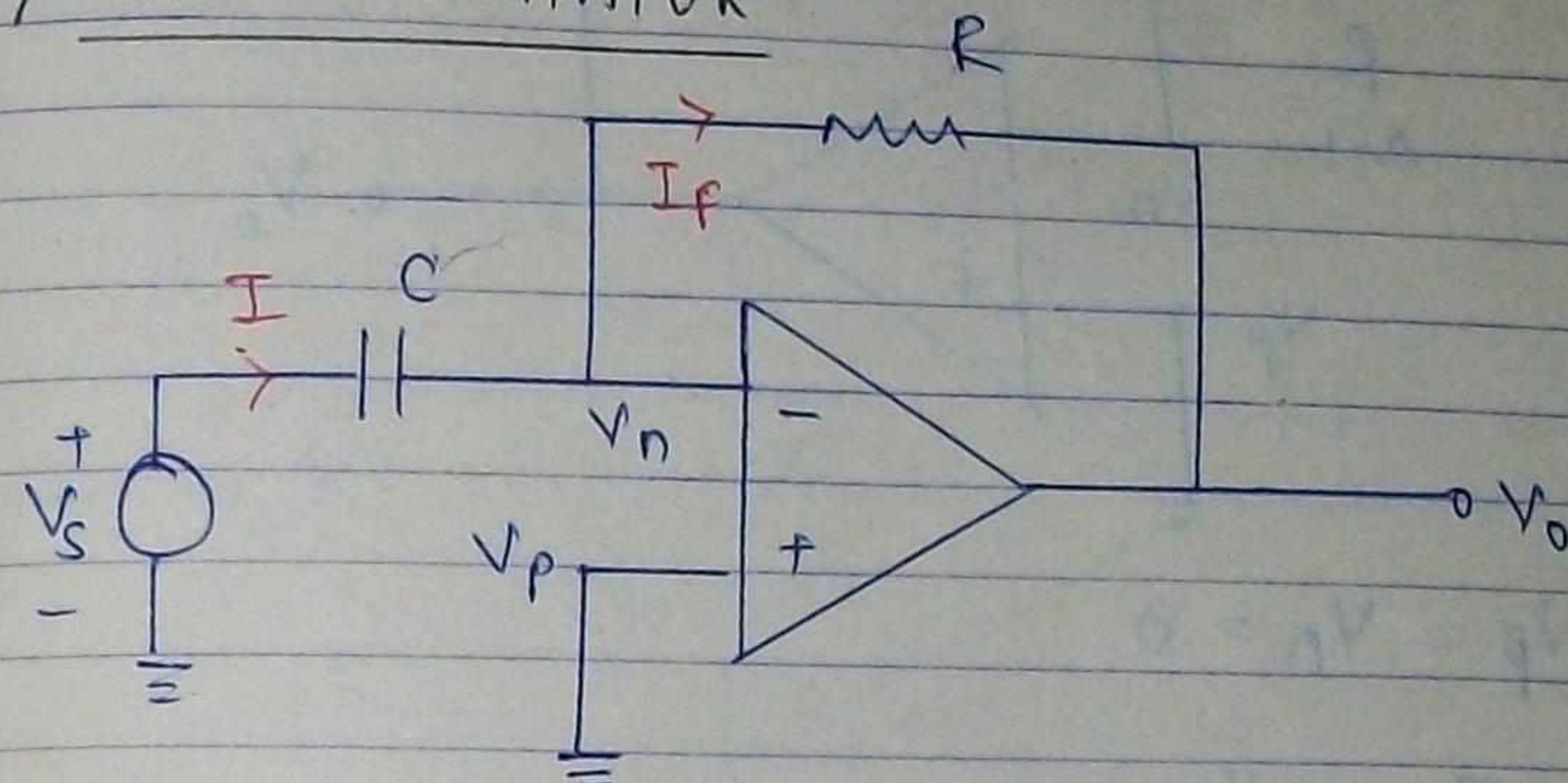
$$\text{SWING} = V_{\max} - V_{\min}$$

Que. Given integrator with ideal OPAMP $R = 1k\Omega$, $C = 1\mu f$. There is a square wave $f = 50\text{ Hz}$ and swing 20 V what is the swing of o/p triangular wave.

Ans.

$$\begin{aligned} \text{Swing} &= \frac{V}{RC} \times \frac{T}{4} - \left(-\frac{V}{RC} \times \frac{T}{4} \right) \\ &= \frac{V}{RC} \times \frac{T}{2} \\ &= \frac{10}{10^3 \times 10^{-6}} \times \frac{1}{2 \times 50} = 10^2 \end{aligned}$$

⇒ DIFFERENTIATOR -



$$V_p = V_n = 0 \quad (\text{Virtual ground})$$

$$I_f = I$$

$$\frac{V_n - V_o}{R} = C \frac{d}{dt} [V_s - V_n]$$

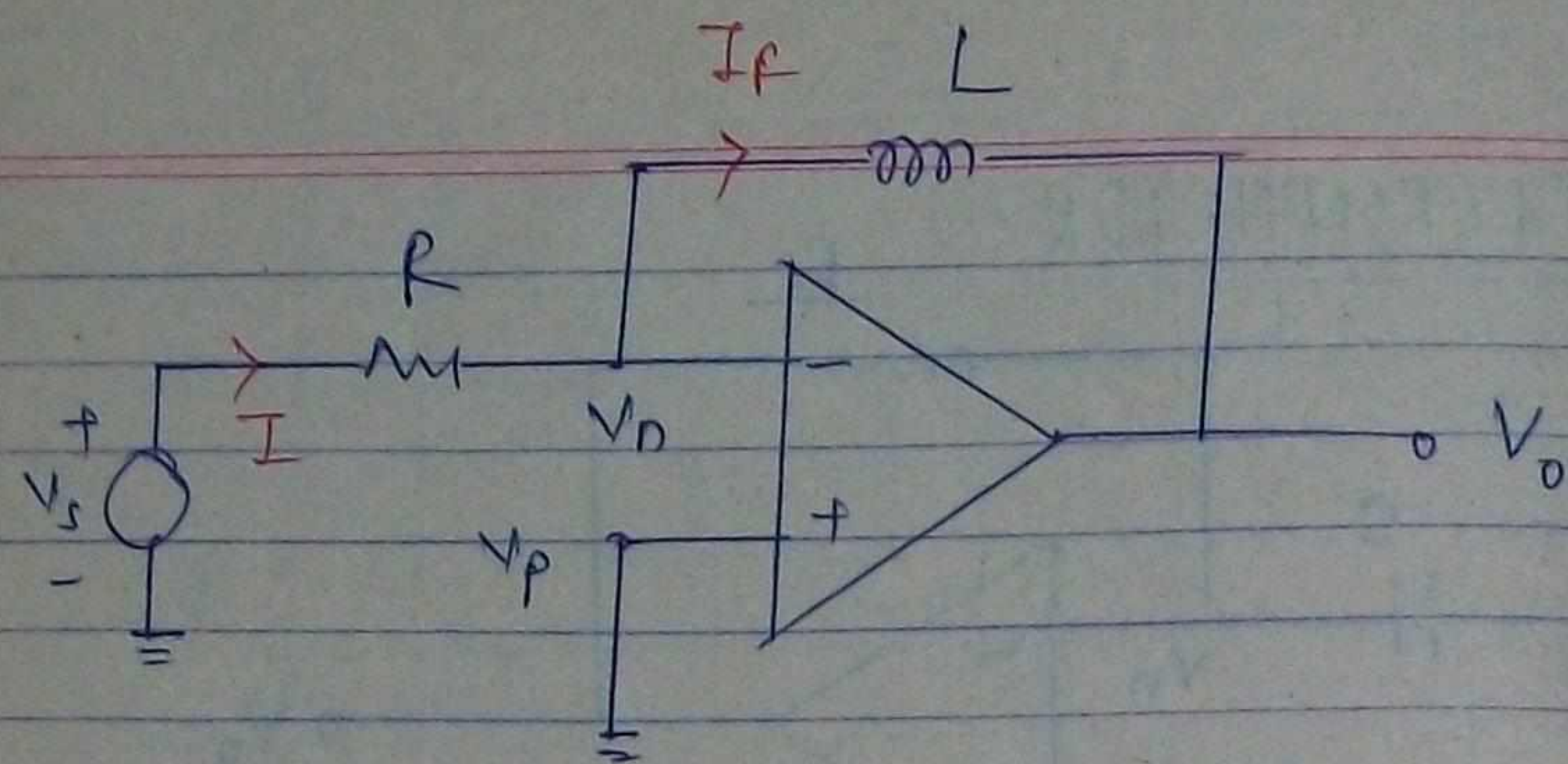
$$-\frac{V_o}{R} = C \frac{d}{dt} V_s$$

$$V_o = -RC \frac{dV_s}{dt}$$

INVERTING DIFFERENTIATOR

due to -ve sign

- Differentiator can also be design with the help of inductor.



$$V_p = V_n = 0$$

$$I_f = I$$

$$\frac{1}{L} \int (V_n - V_o) dt = \frac{V_s - V_o}{R}$$

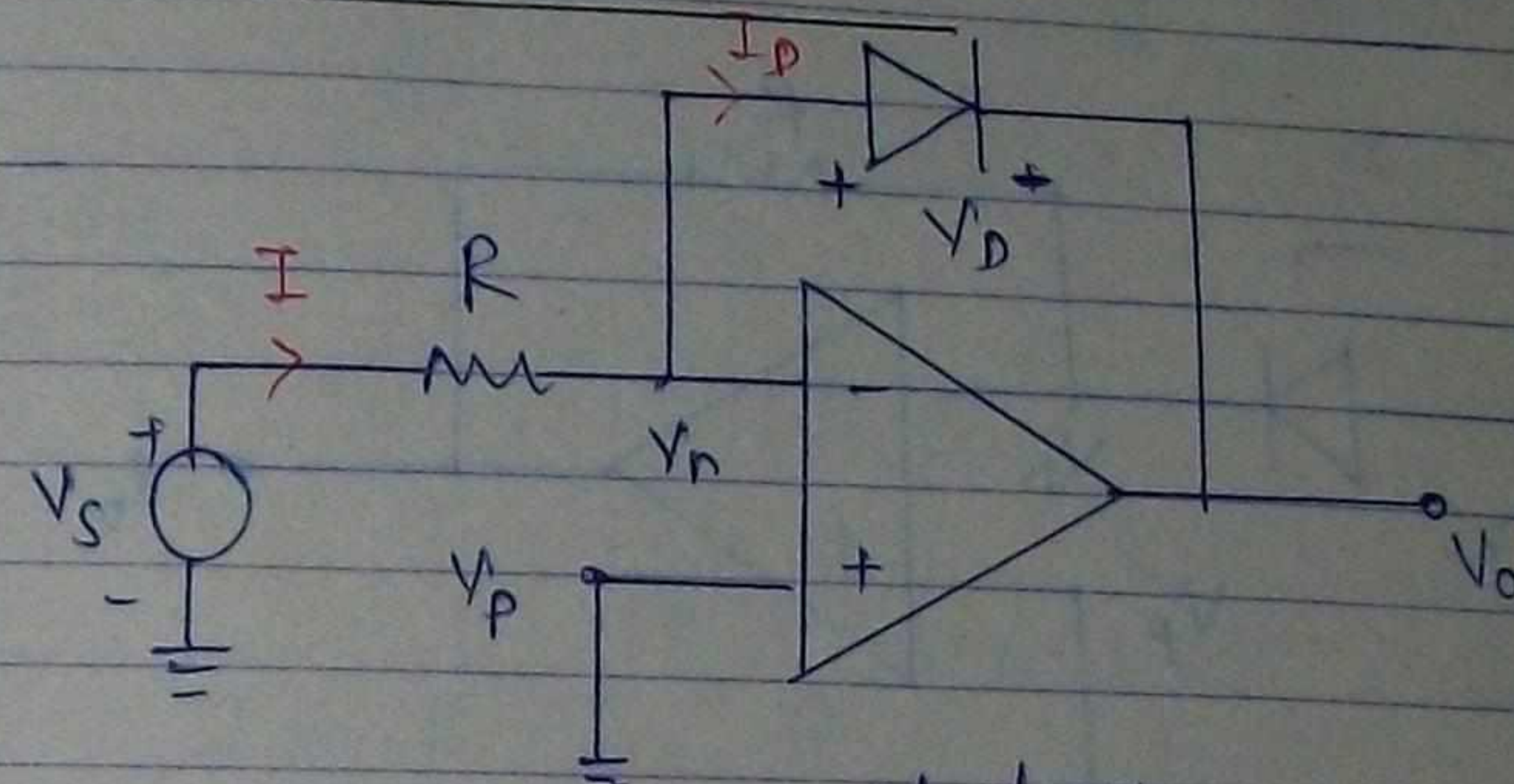
$$-\frac{1}{L} \int V_o dt = \frac{V_s}{R}$$

$$\boxed{V_o = -\frac{L}{R} \frac{dV_s}{dt}}$$

DISADVANTAGE -

The circuit using inductor is very bulky and expensive.

⇒ LOGARITHMIC AMPLIFIER -



$$\text{Since, } I_D = I_0 \left[e^{V_o / \eta V_T} - 1 \right]$$

$$I_D \approx I_0 e^{V_o / \eta V_T}$$

$$V_p = V_n = 0 \quad (\text{virtual ground})$$

$$I_D = I$$

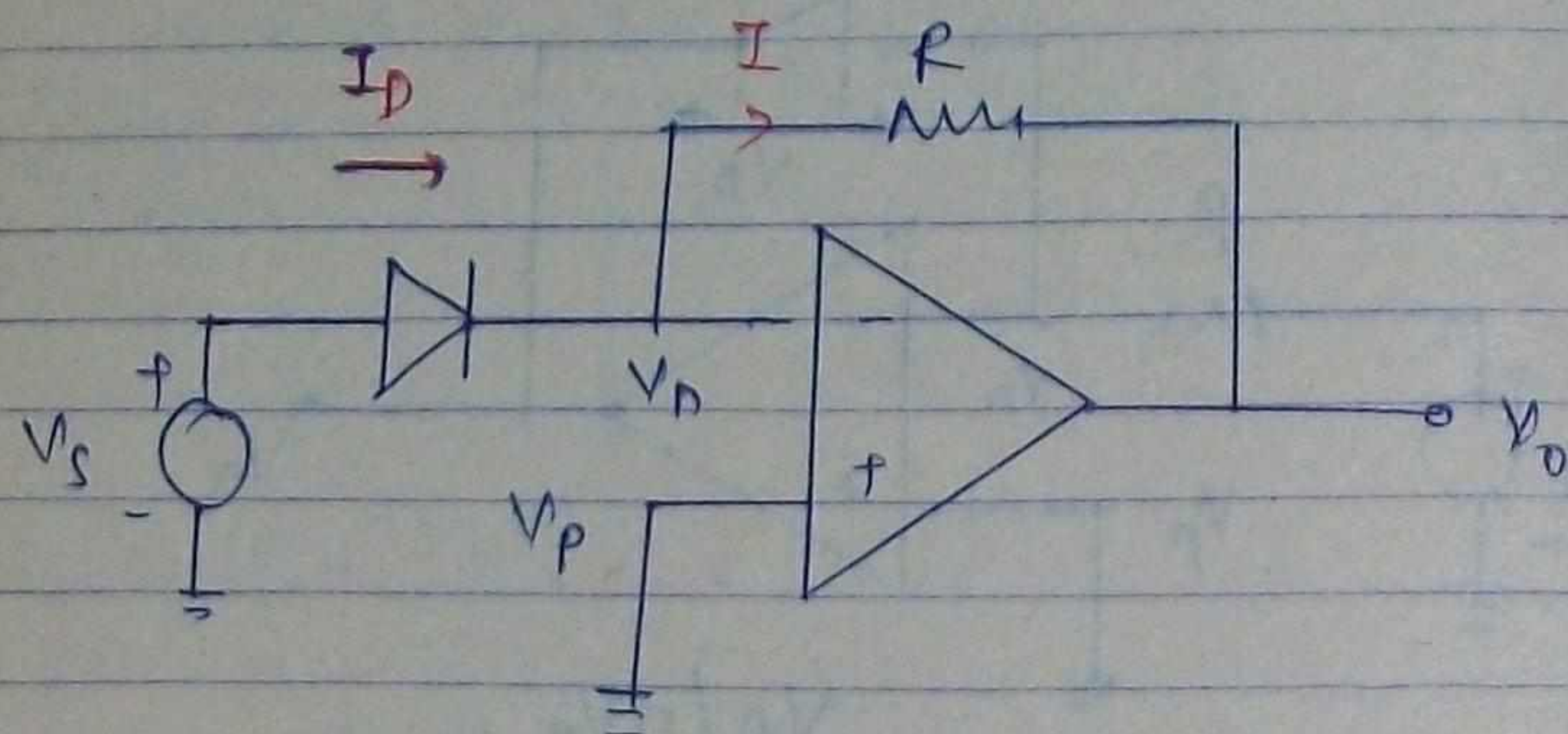
$$I_0 e^{-V_o / \eta V_T} = \frac{V_s}{R}$$

$$e^{-V_o / \eta V_T} = \frac{V_s}{I_0 R}$$

$$-\frac{V_o}{\eta V_T} = \ln \frac{V_s}{I_0 R}$$

$$\boxed{V_o = -\eta V_T \ln \left(\frac{V_s}{I_0 R} \right)}$$

⇒ ANTILOGRATHMIC AMPLIFIER -



$$V_p = V_n = 0$$

$$I_D = I$$

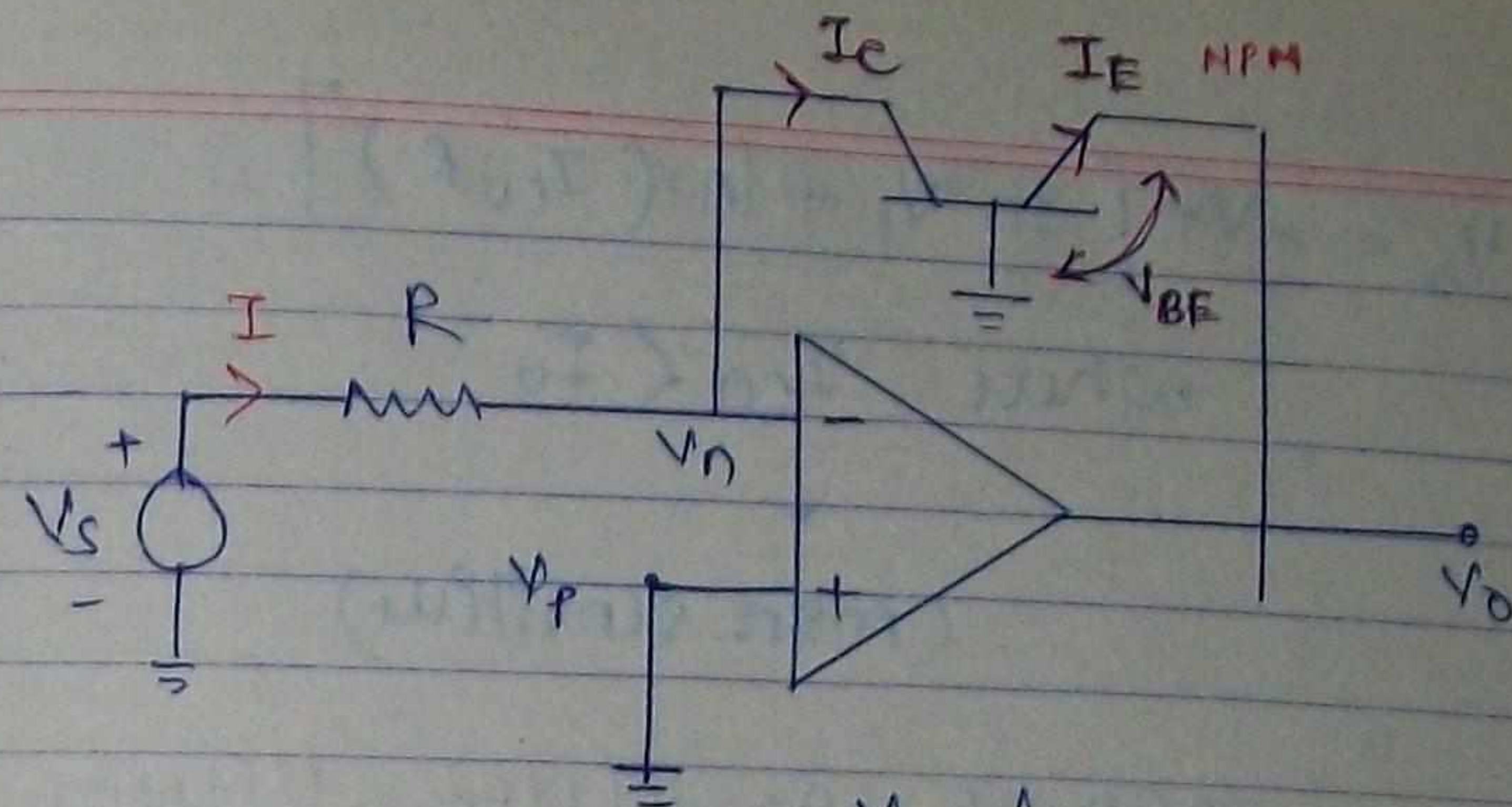
$$I_0 e^{V_s / n V_T} = \frac{0 - V_o}{R}$$

$$\textcircled{4} \quad V_o = -I_0 R \text{ antilog} \left(\frac{V_s}{n V_T} \right)$$

⇒ LOGRATHMIC AMPLIFIER BY USING BJT -

$$I_C \approx I_E$$

$$I_E = I_{CO} \left(e^{V_{BE} / V_T} - 1 \right)$$



$$I_E \approx I_{CO} e^{V_{BE} / V_T}$$

$$V_{BE} = -V_o$$

$$I = I_C \approx I_E$$

$$\frac{V_s}{R} = I_{CO} e^{V_{BE} / V_T}$$

$$\frac{V_s}{R} = I_{CO} e^{-V_o / V_T}$$

$$e^{-V_o / V_T} = \frac{V_s}{I_{CO} R}$$

$$-\frac{V_o}{V_T} = \ln \frac{V_s}{I_{CO} R}$$

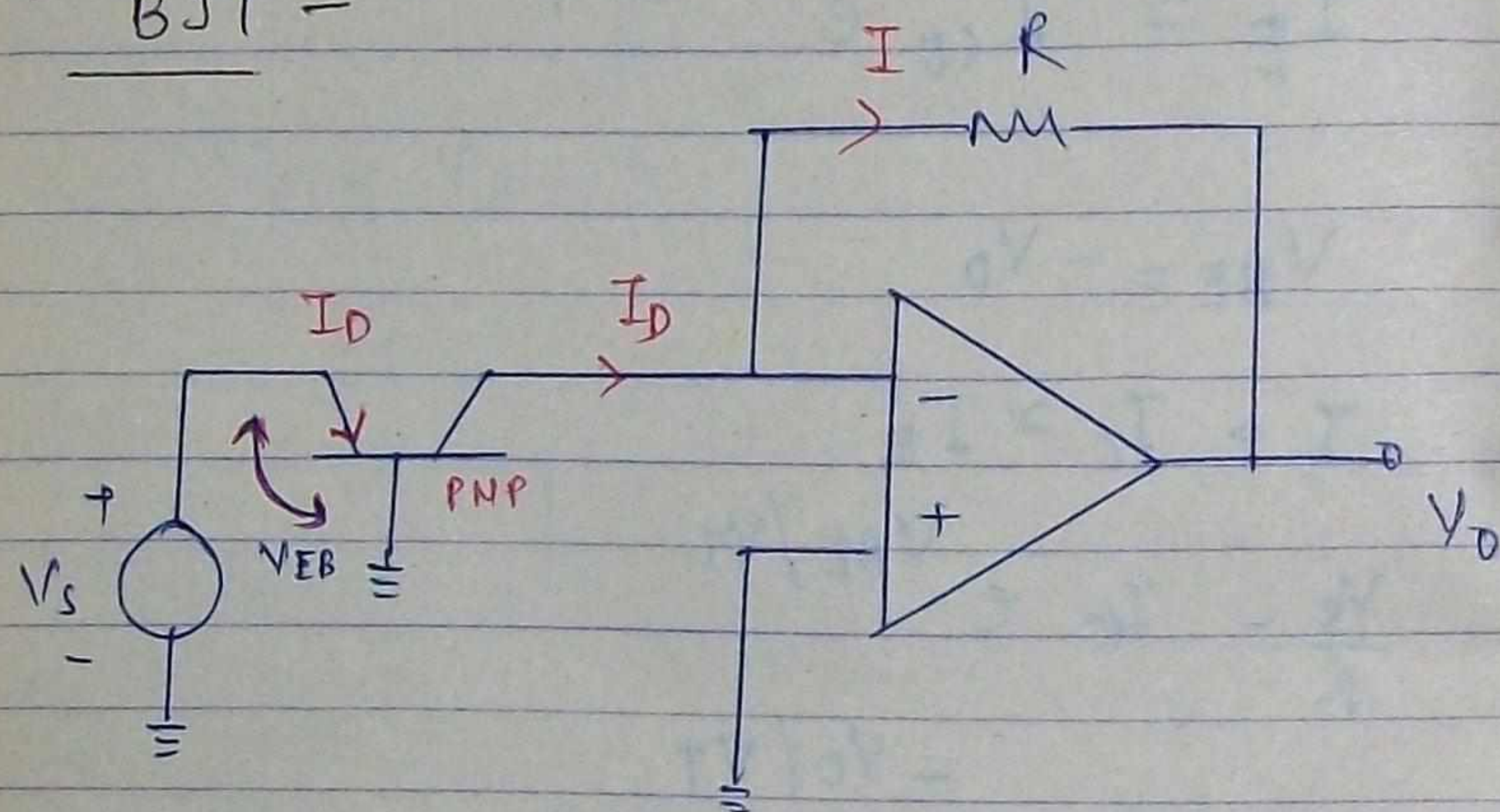
$$V_o = -V_T \ln \left(\frac{V_s}{I_{CO} R} \right)$$

$$V_o = -V_T \left[\ln V_s - \ln (I_{co} R) \right]$$

where $I_{co} < I_o$

(more sensitive)

⇒ ANTILOGRATHMIC AMPLIFIER USING BJT -



$$I = I_D$$

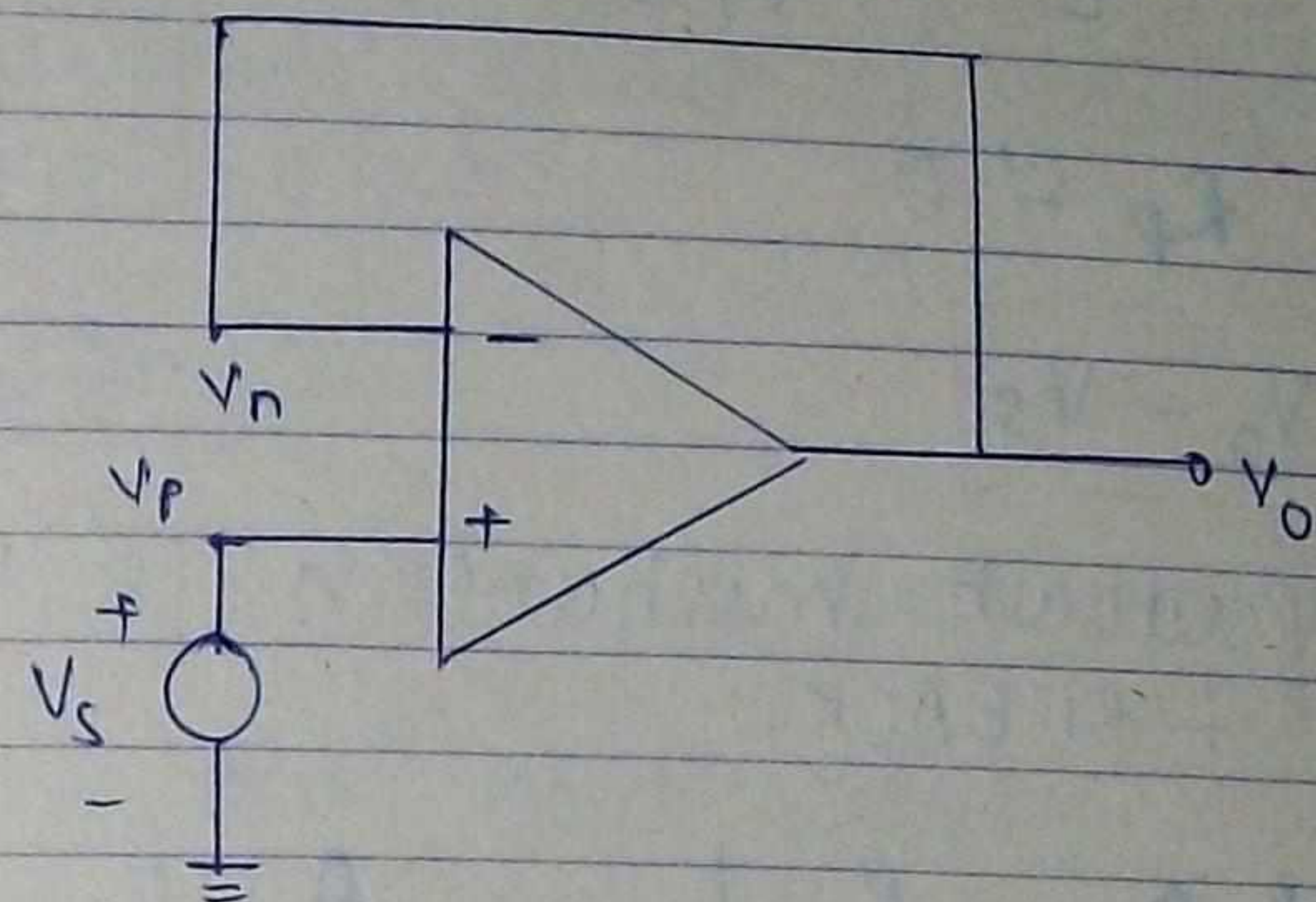
$$-\frac{V_o}{R} = I_{co} e^{V_s / \eta V_T}$$

$$V_o = -I_{co} R e^{V_s / \eta V_T}$$

$$V_o = -I_{co} R \text{ antilog} \left(\frac{V_s}{\eta V_T} \right)$$

⇒ VOLTAGE FOLLOWER -

①



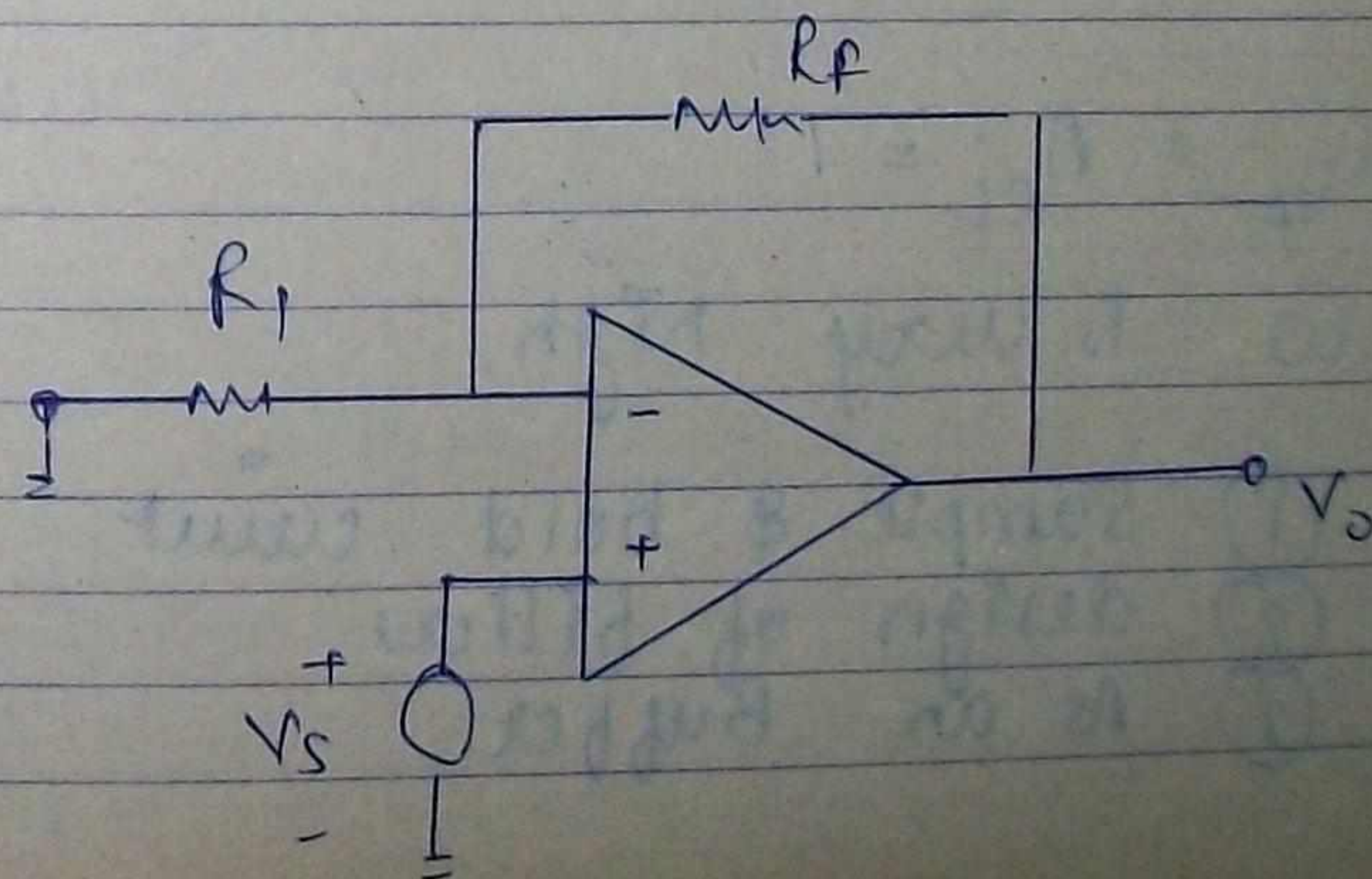
$$V_p = V_n = V_s$$

$$V_o = V_s, \quad A = \frac{V_o}{V_s} = 1$$

$$|A| = 1$$

$$\phi_{\text{shift}} = 0$$

②



$$V_o = \left[1 + \frac{R_f}{R_i} \right] V_s$$

$$R_f \approx 0$$

$$V_o = V_s$$

Here feedback mechanism is VOLTAGE SERIES FEEDBACK.

$$R_i \uparrow \quad R_o \downarrow \quad A_v \downarrow$$

$$R_i = 10^6 \Omega (1 \text{ M}\Omega), \quad R_o = 100 \Omega, \quad A_v = 10^6$$

$$R_{if} = 10^{10} \Omega = 10^4 \text{ M}\Omega$$

$$R_{of} \approx 10^{-4} \Omega \approx 0 \Omega$$

$$A_{vf} = A_{cl} = 1$$

B.W is very high.

- uses:
- ① Sample & hold circuit
 - ② design of filters
 - ③ As an Buffer.

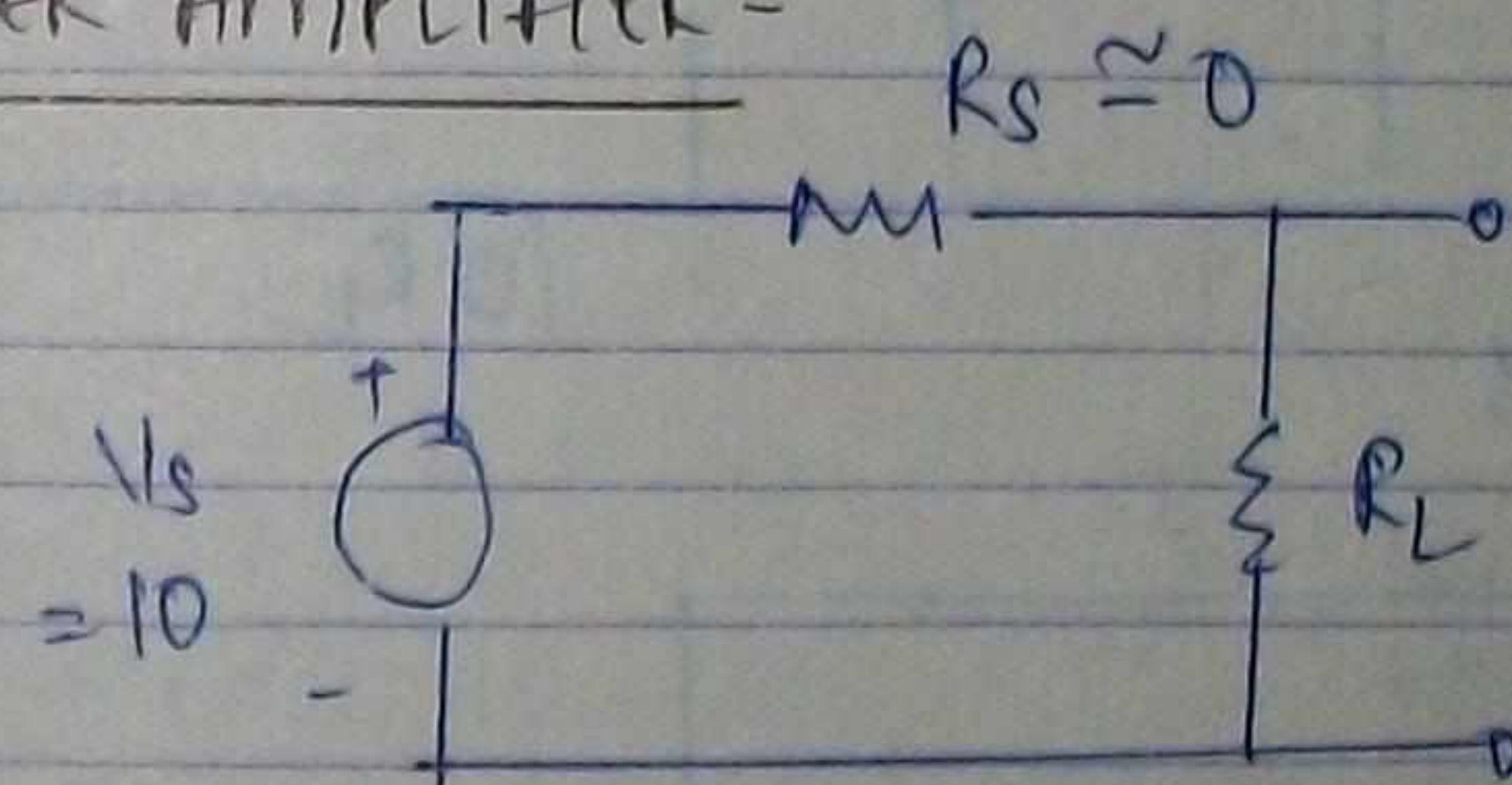
11

Ques. In a non inverting amplifier if the feedback resistance R_f is negligible small then circuit is called "Voltage follower."

- Applications:

- ① design of 1st order Butterworth filter
- ② Sample and hold circuit
- ③ used as an buffer amplifier or as an impedance matching device between high resistance and low resistance.

① BUFFER AMPLIFIER -



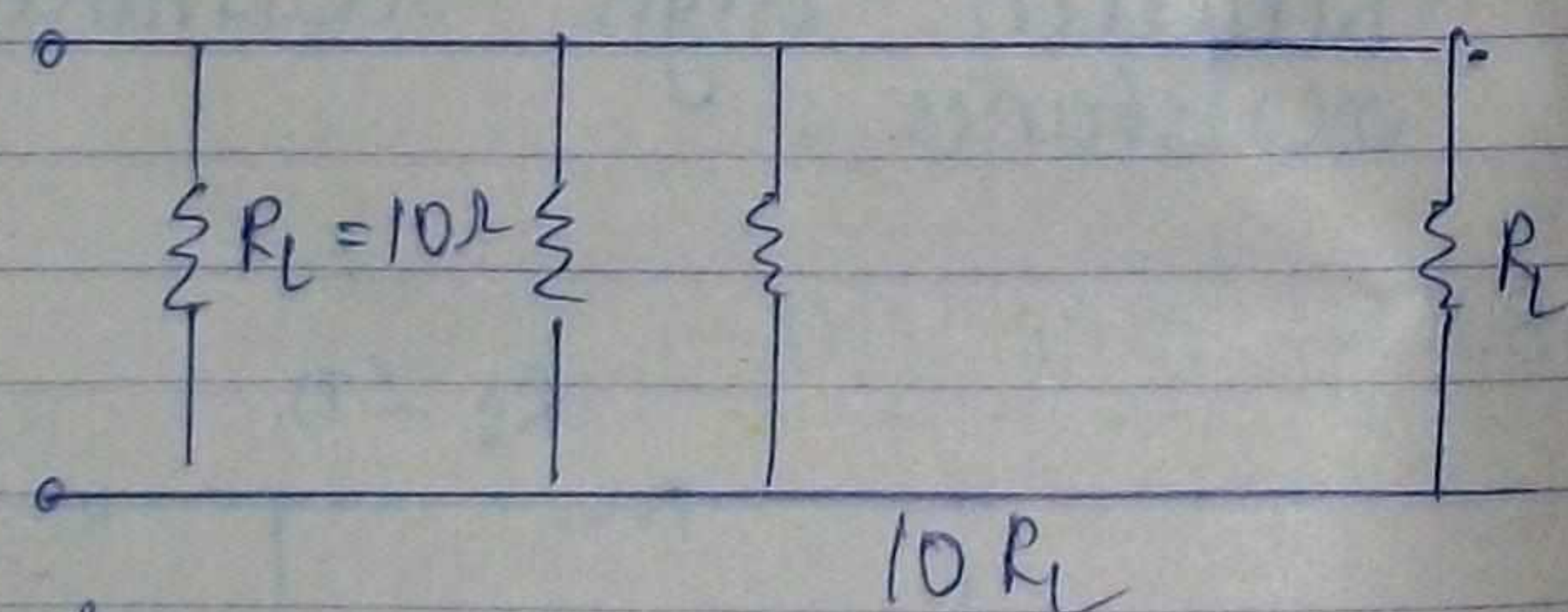
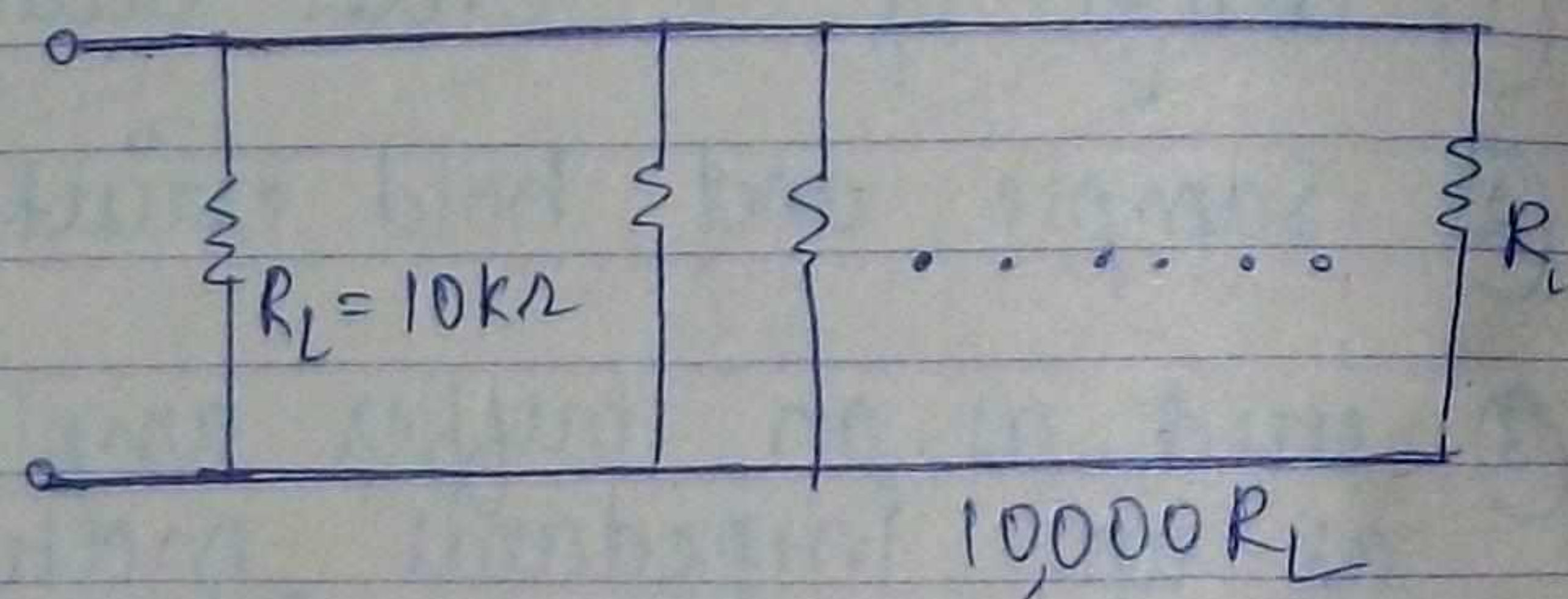
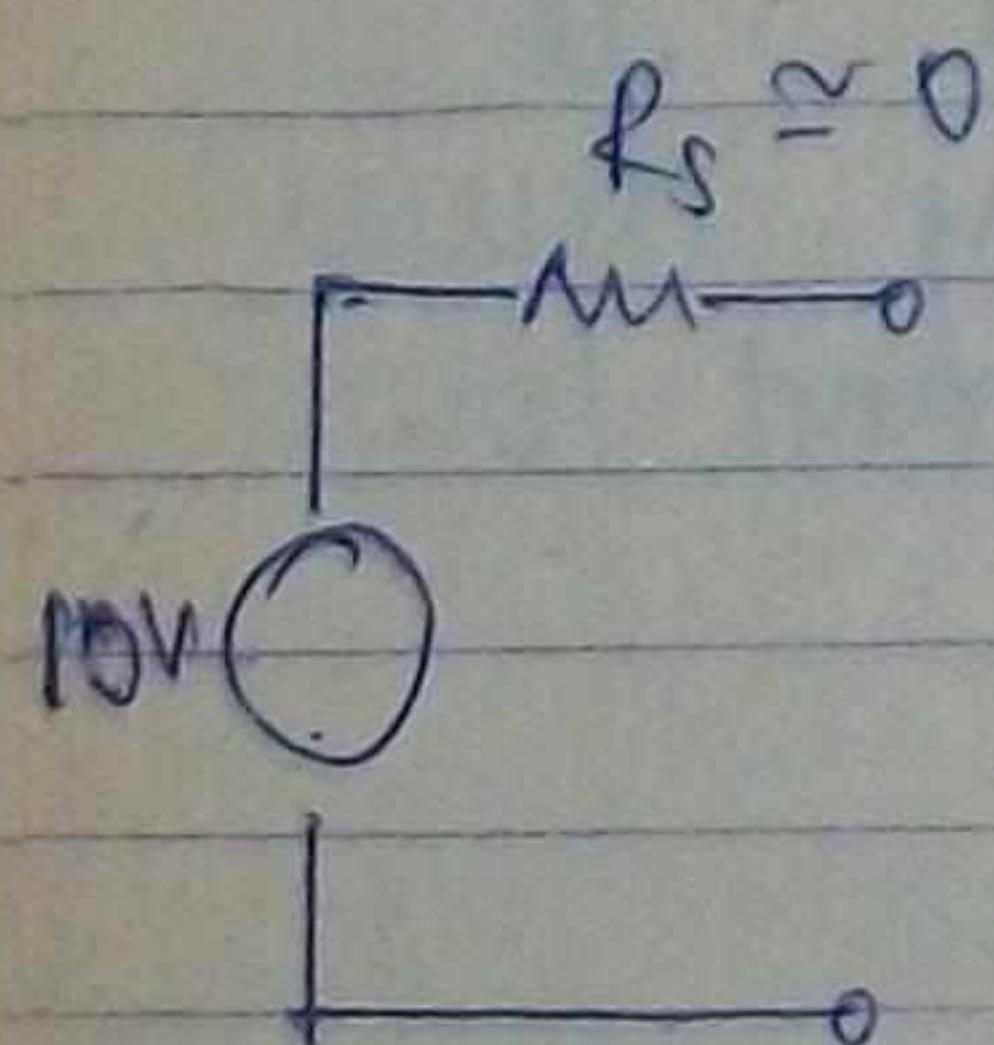
$$I_L = \frac{V_s}{R_s + R_L} \approx \frac{V_s}{R_L}$$

$$R_L = 10 \text{ k}\Omega \quad I_L = 10 / 10 \text{ k}\Omega = 1 \text{ mA}$$

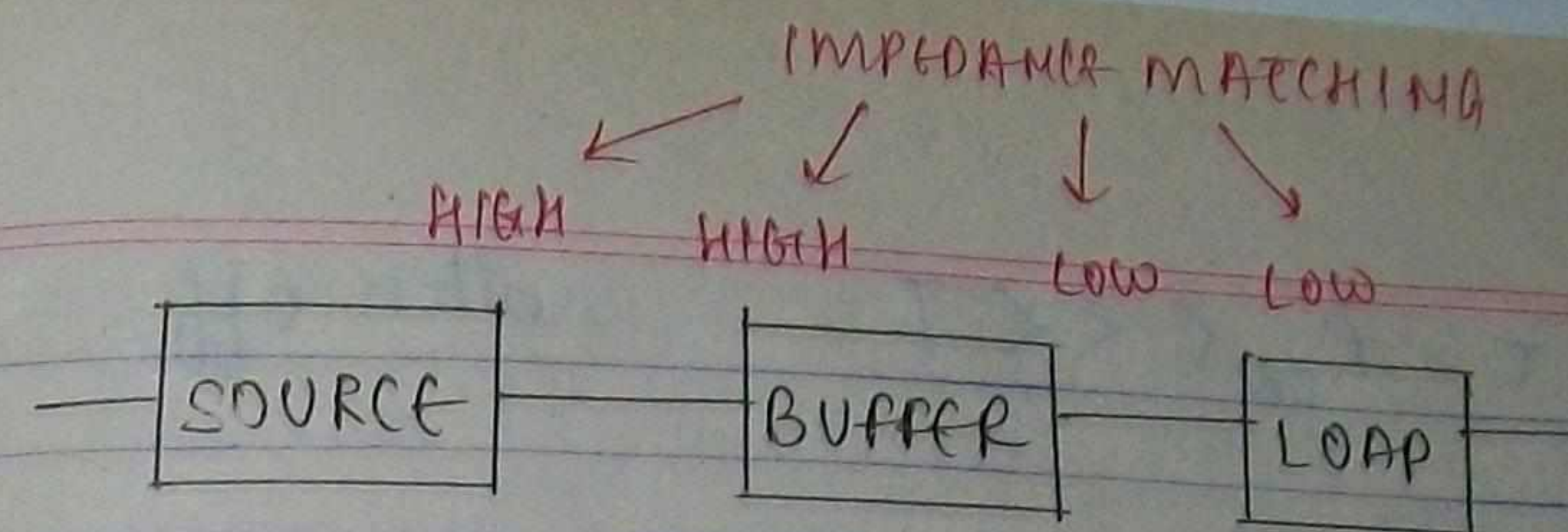
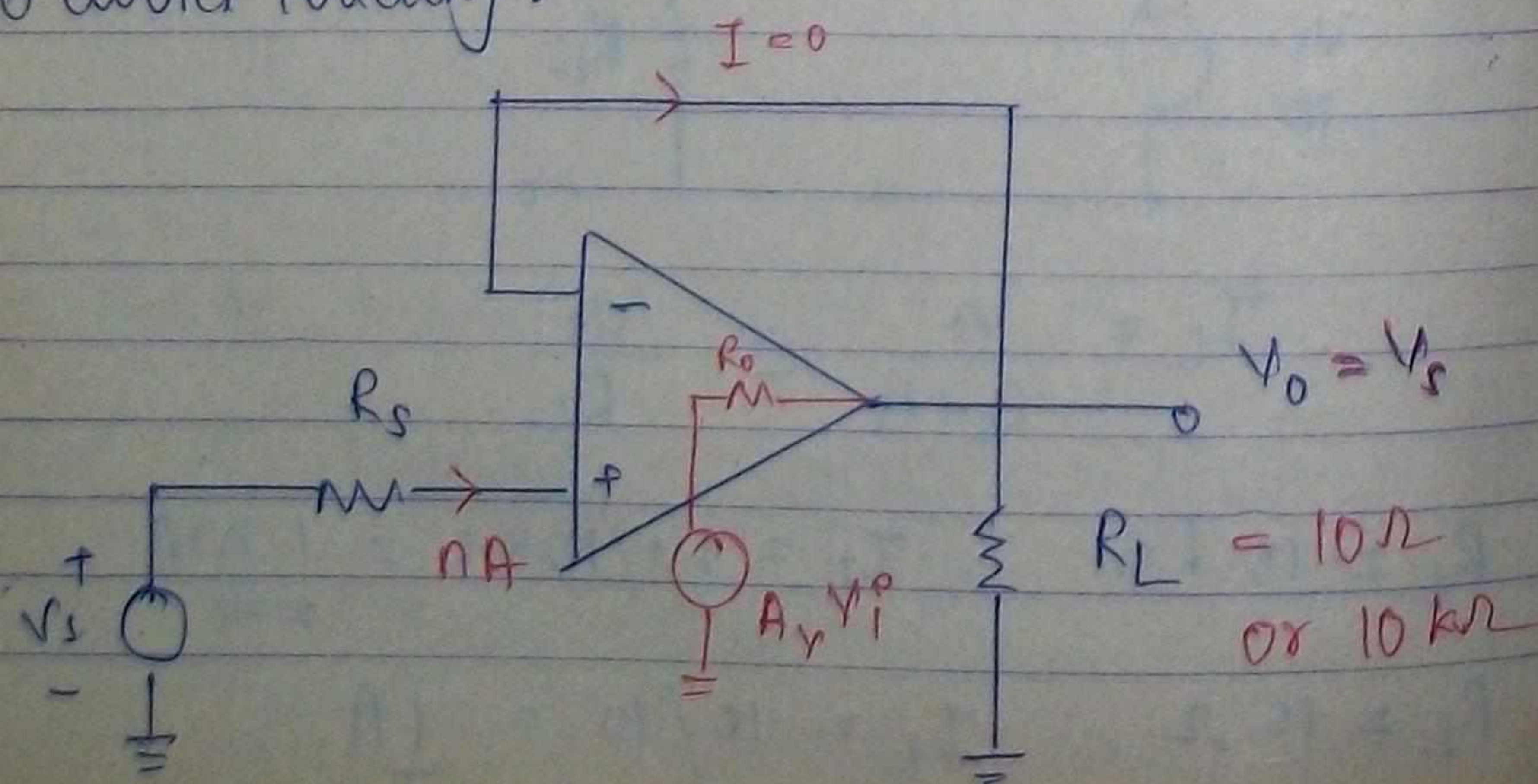
$$R_L = 10 \Omega \quad I_L = 10 / 10 = 1 \text{ A}$$

RESISTANCE	I_L	
HIGH	LOW	Low loading effect
LOW	HIGH	High loading effect

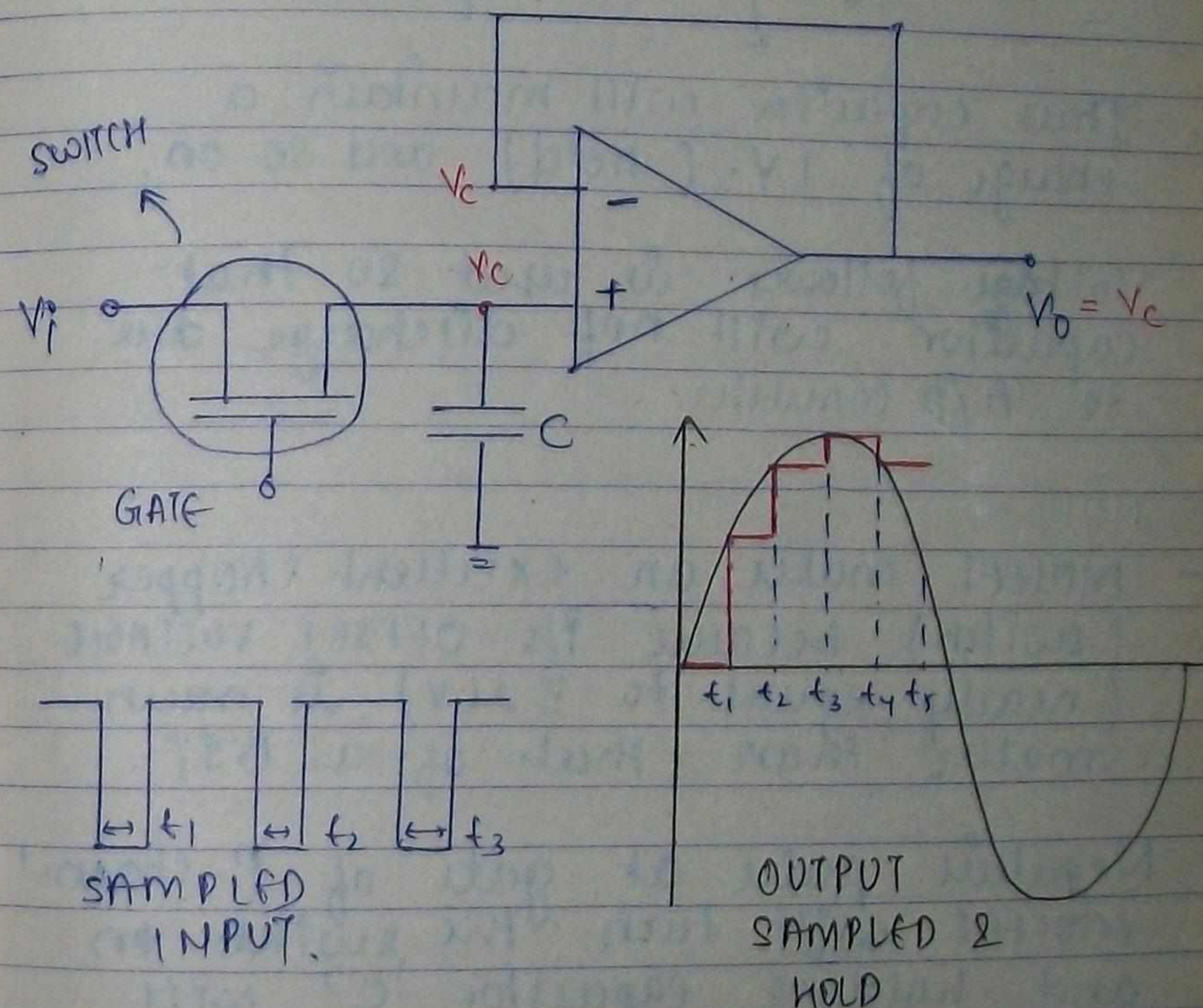
Let
 $P = 100W$
 $V = 10V$
 $I_{max} = 10A$



To avoid loading?

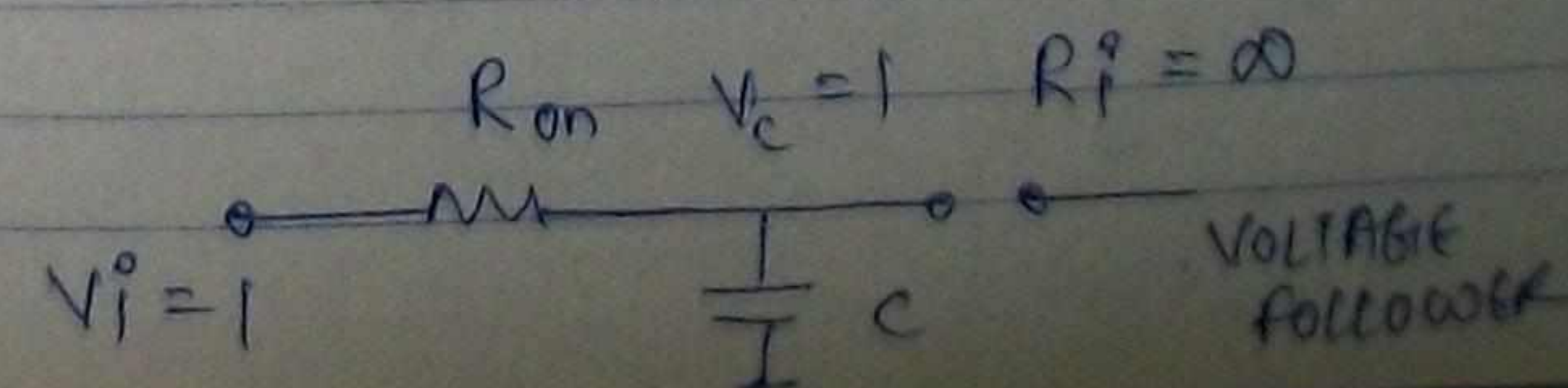


② SAMPLE AND HOLD CIRCUIT -

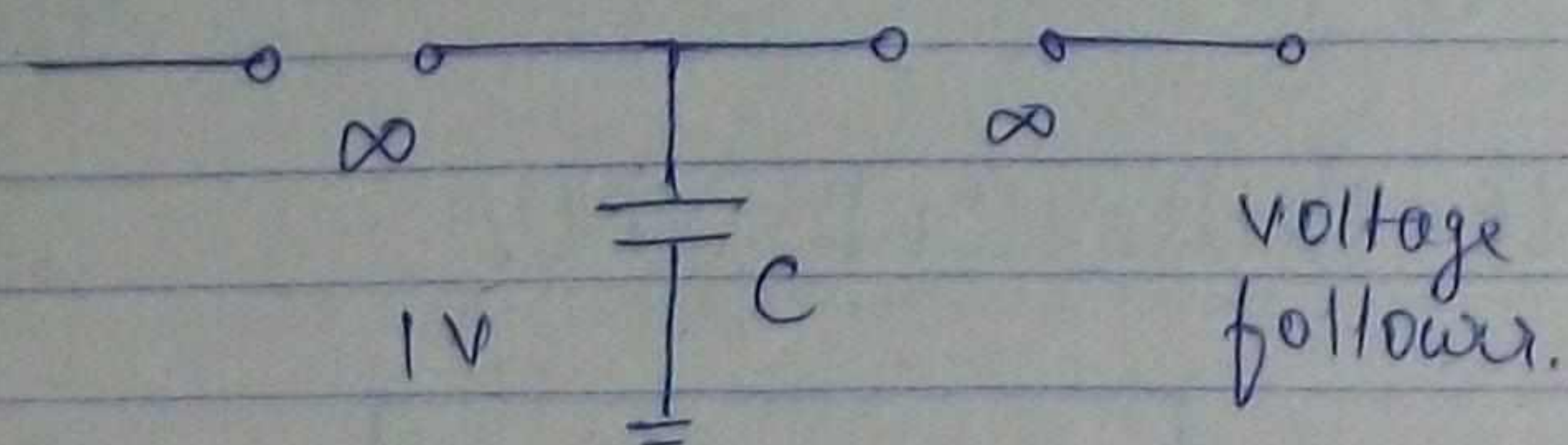


① At $t = t_1$ - we pulse appear at gate

MOSFET $\rightarrow$ ON



② $\tau = R_{ON} C \ll T$ switch off.

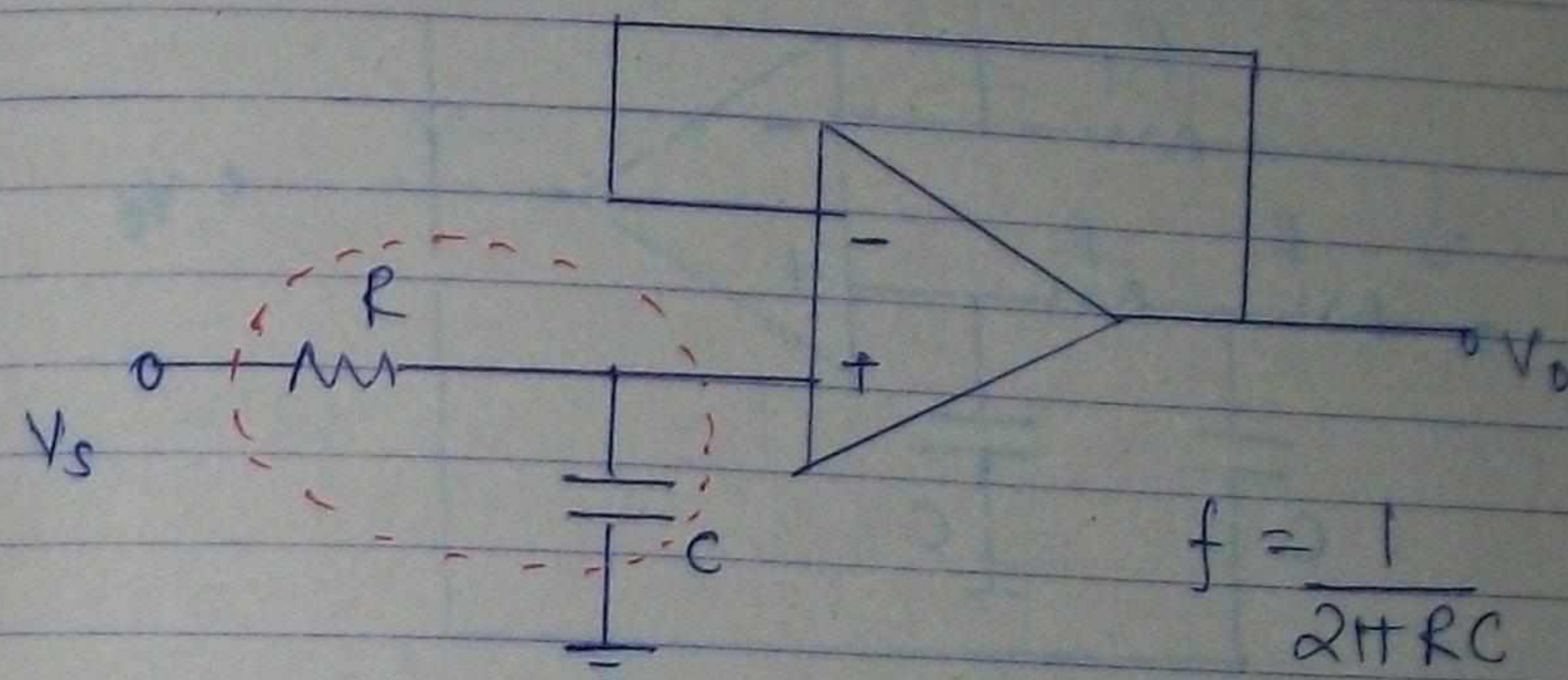


- Thus capacitor will maintain a charge of 1V. (hold) and so on.
- Voltage follower is used so that capacitor will not discharge due to A/D converter.

NOTE -

- MOSFET makes an excellent chopper (switch) because its offset voltage (nearly equal to 5 mV) is much smaller than that of a BJT.
- Negative pulse at gate of P-channel MOSFET will turn the switch on and holding capacitor 'C' will charge with a time constant $\tau = R_{ON} \cdot C$ to the instantaneous value of $1/P$ (V_P).

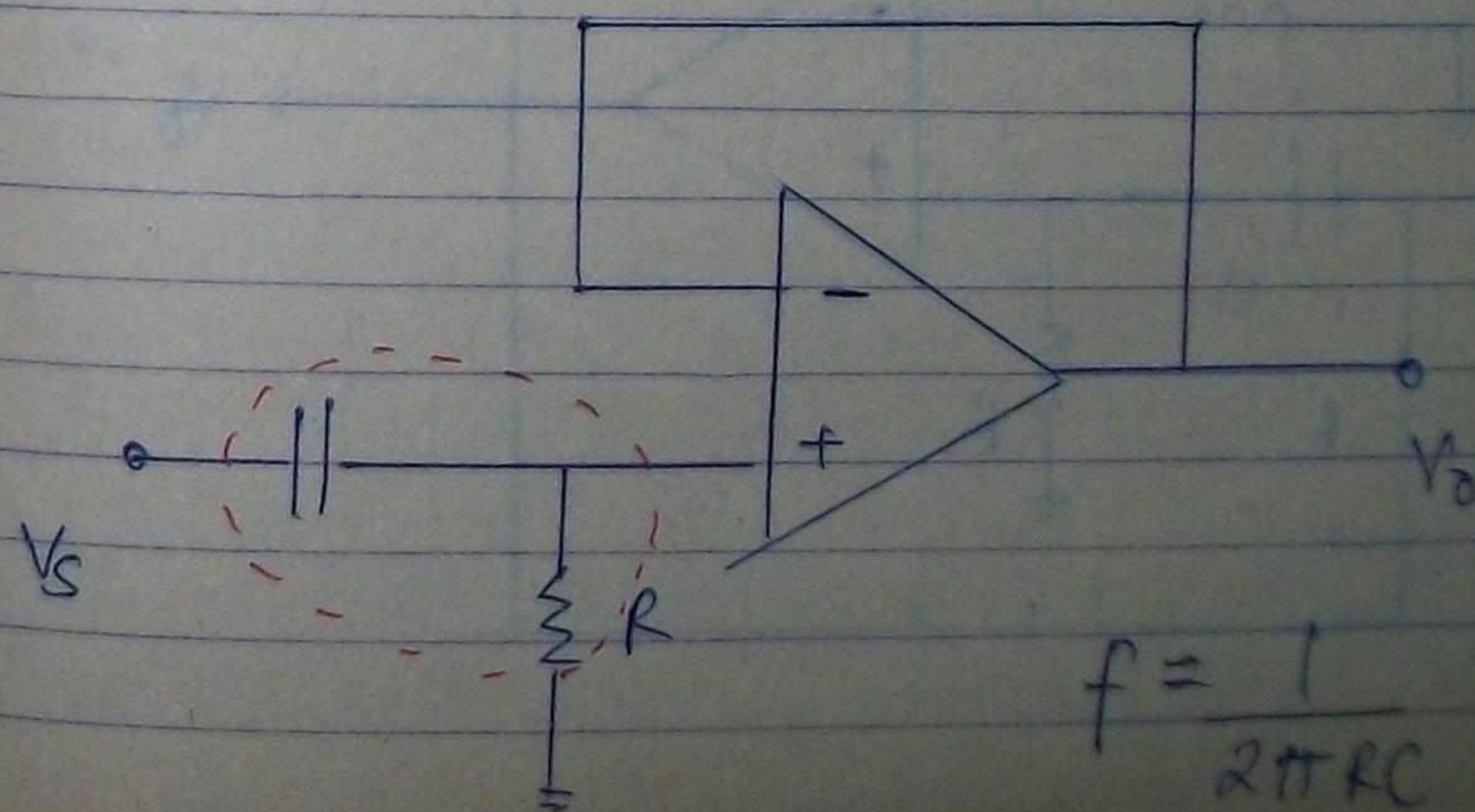
③ BUTTERWORTH FILTER (BWF) -
(1st ORDER LOW PASS BWF)



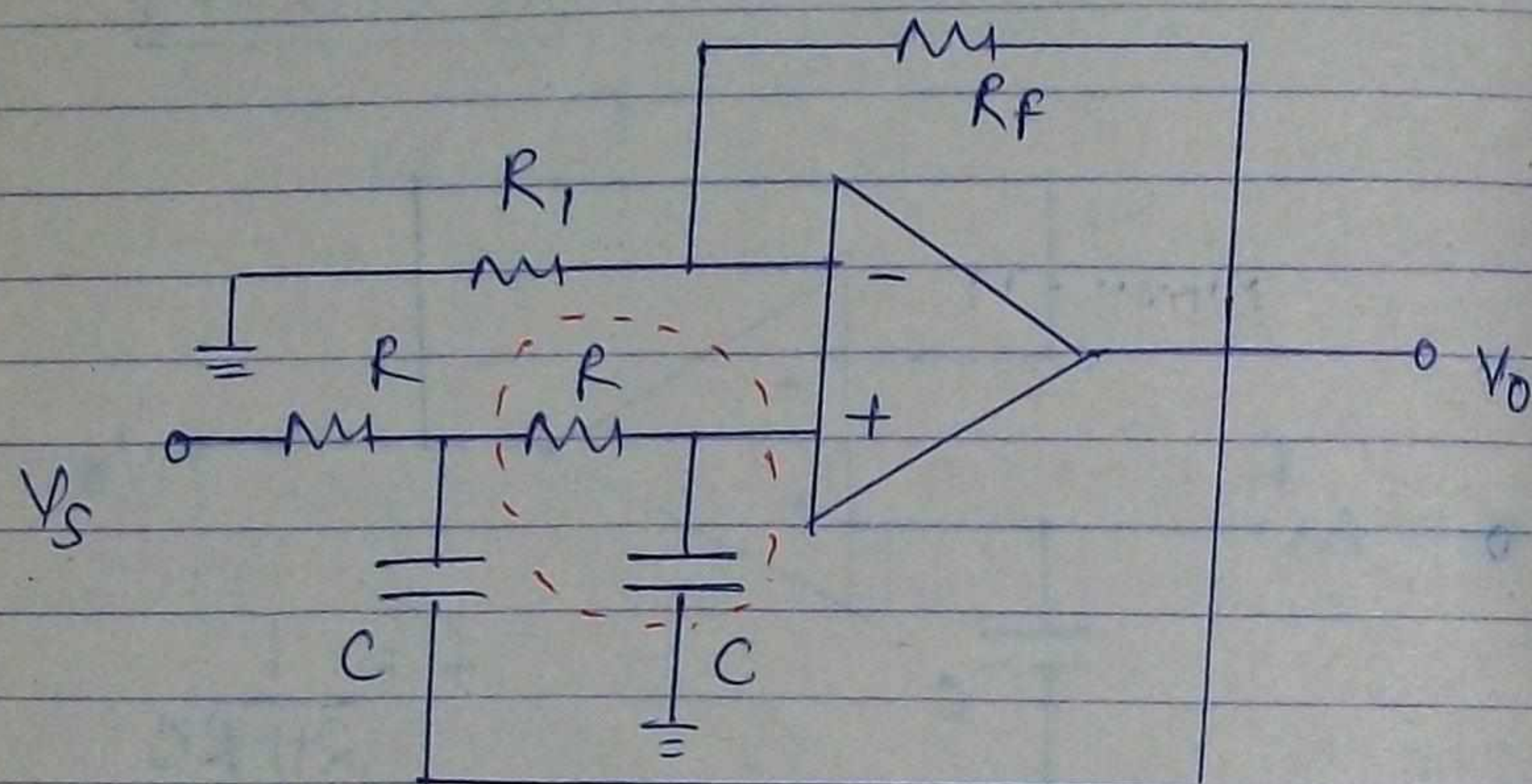
$$\frac{V_o}{V_i} = A = \frac{1 / j\omega C}{R + 1 / j\omega C}$$

$$A = \frac{1}{1 + j\omega RC}$$

④ 1st ORDER HIGH PASS BWF -

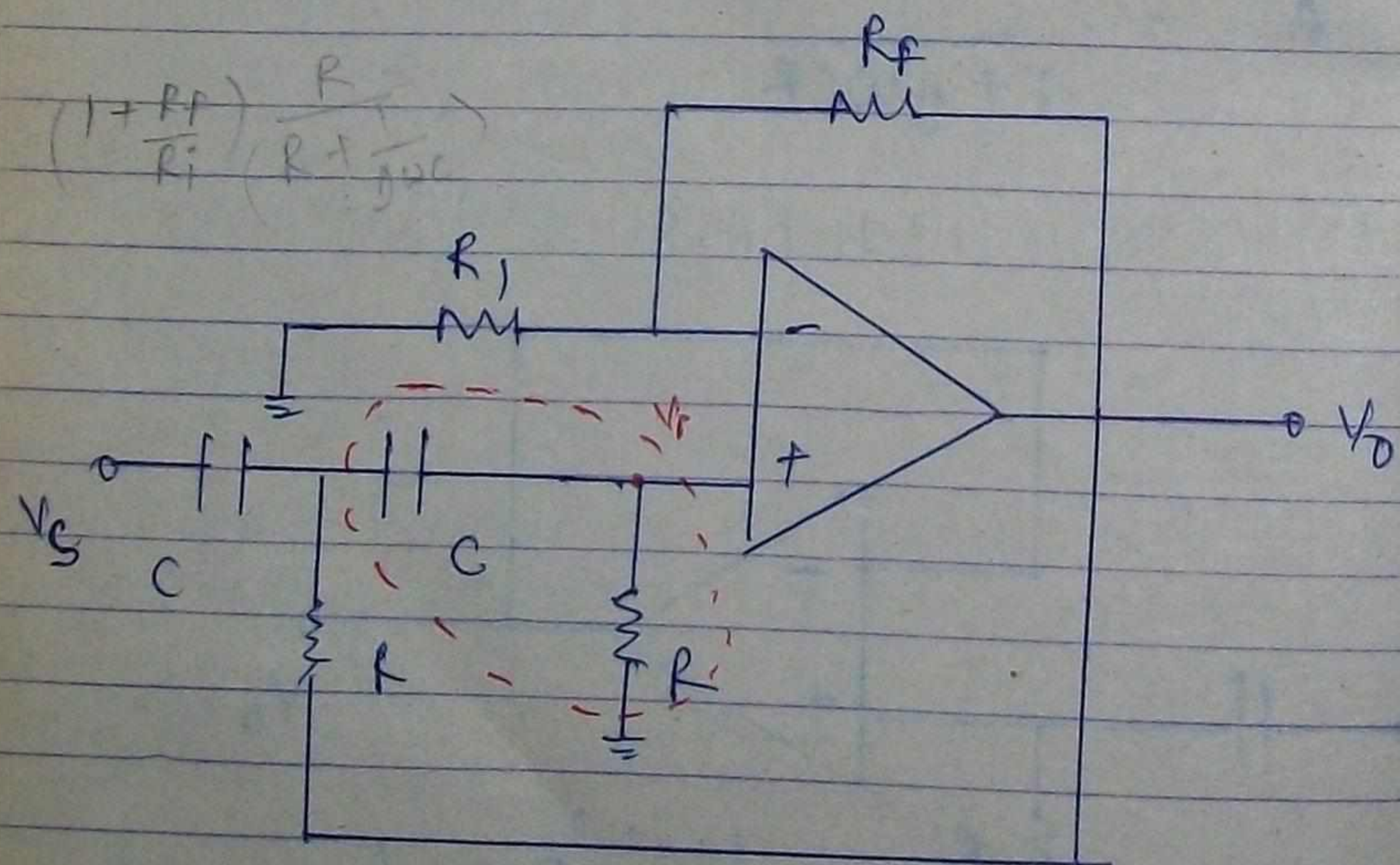


(C) 2nd ORDER LOW PASS BWF -

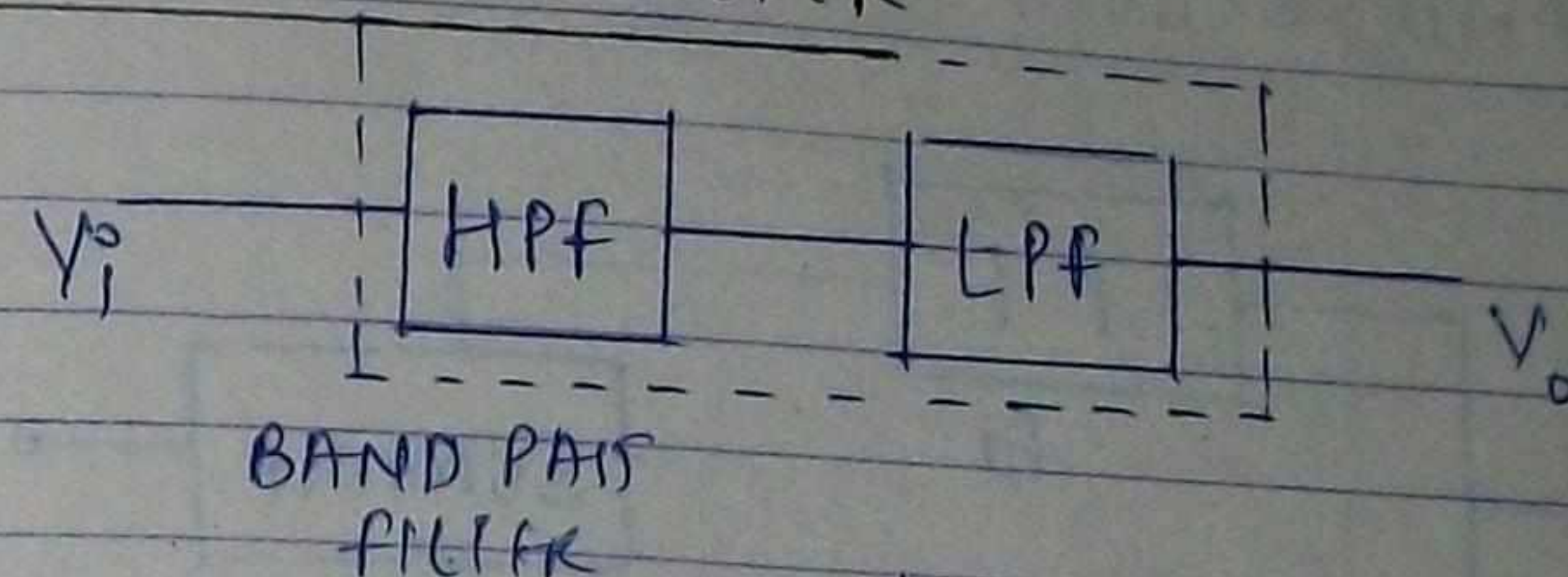


$$f = \frac{1}{2\pi RC}$$

(D) 2nd ORDER HIGH PASS BWF -



(E) BAND PASS FILTER -



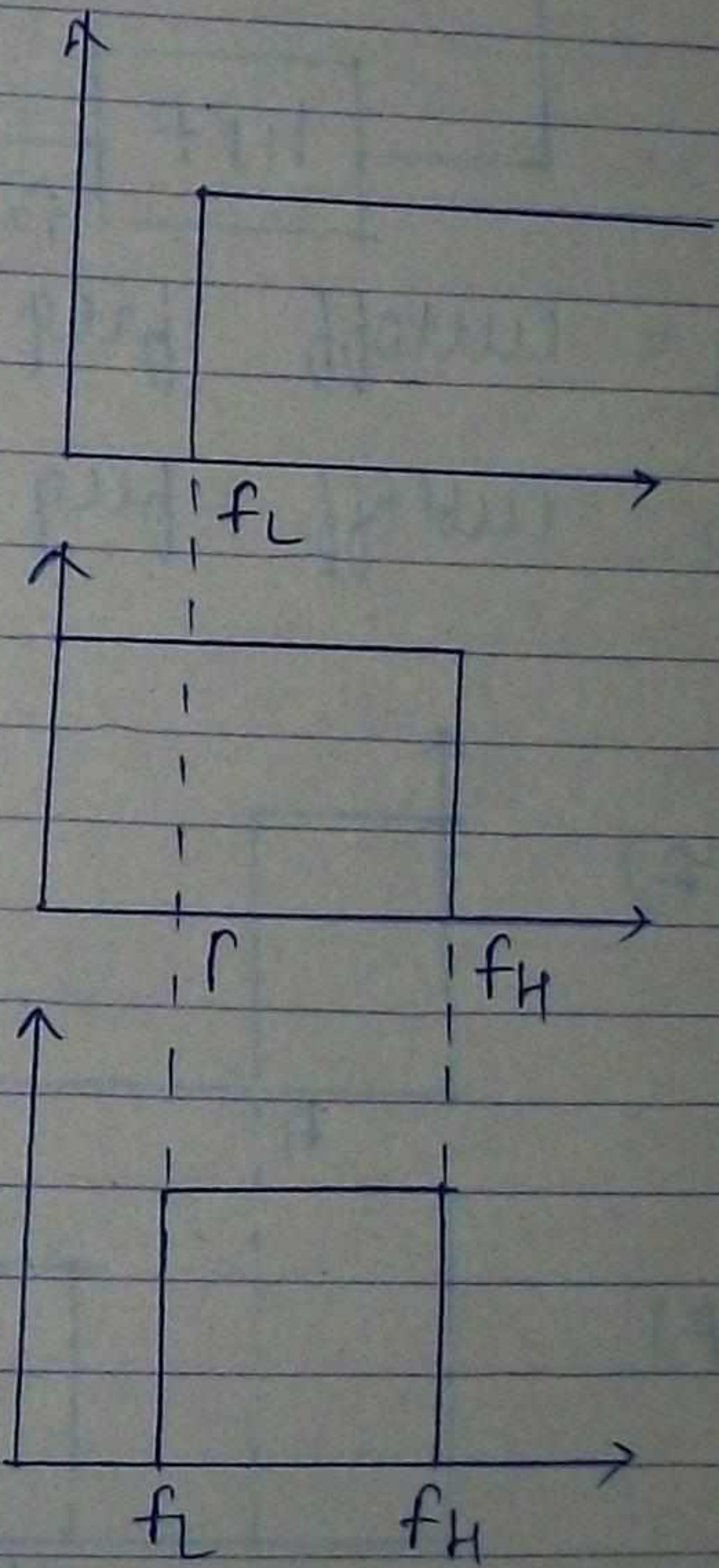
HIGH PASS FILTER

LOW PASS FILTER

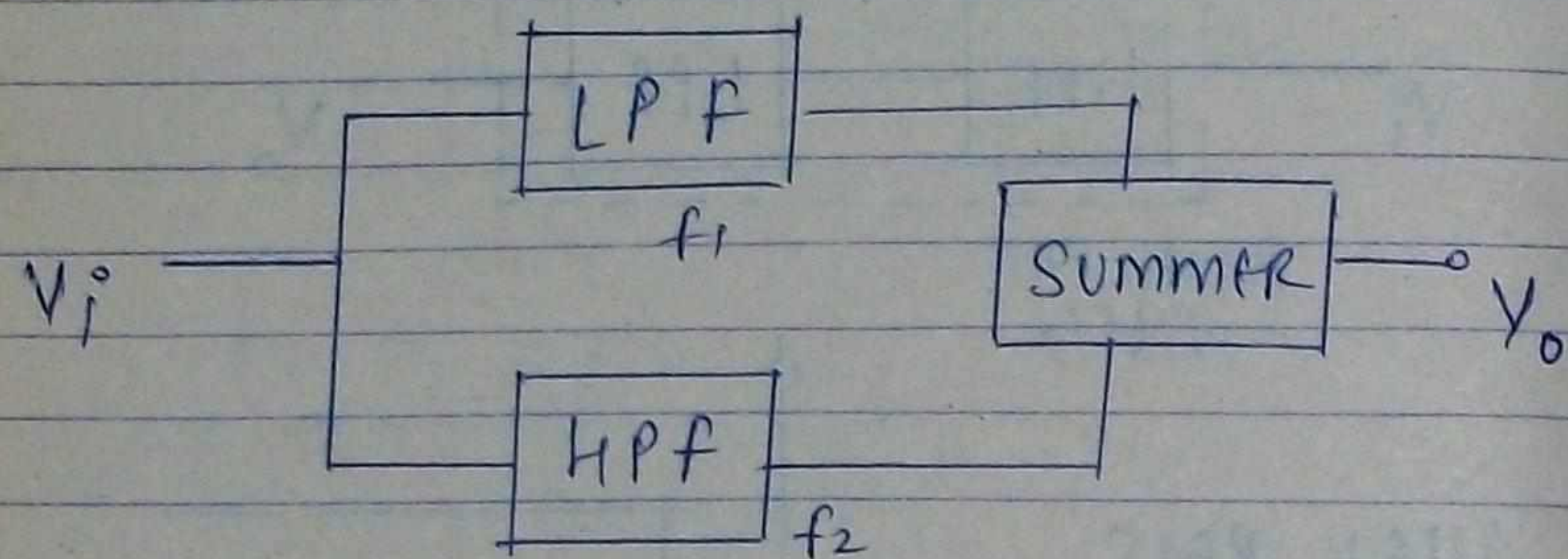
BAND PASS FILTER

$$f_H > f_L$$

f_H = 3 db cutoff freq of LPF
 f_L = 3db — — — — — HPF

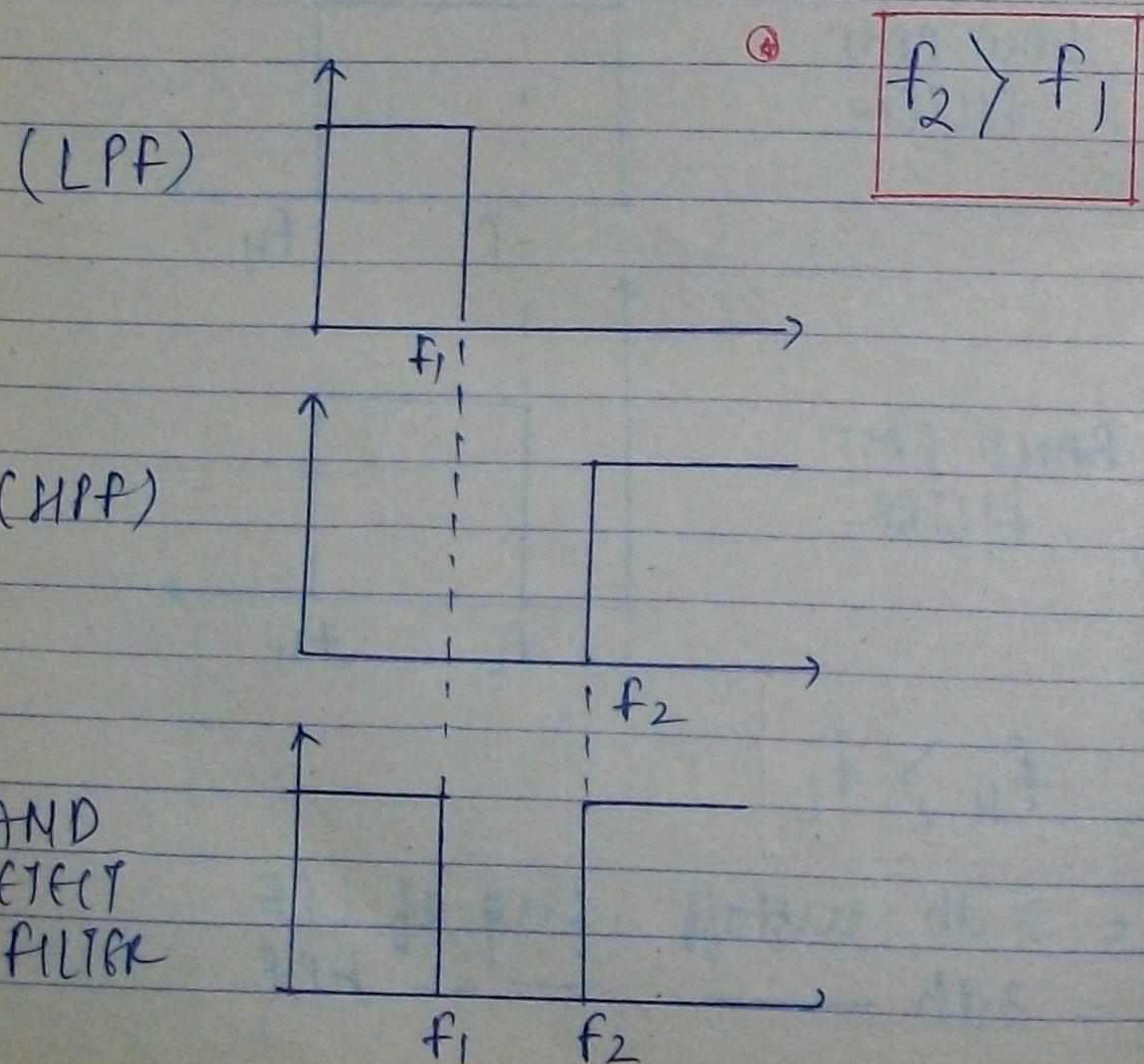


(F) BAND REJECT FILTER —

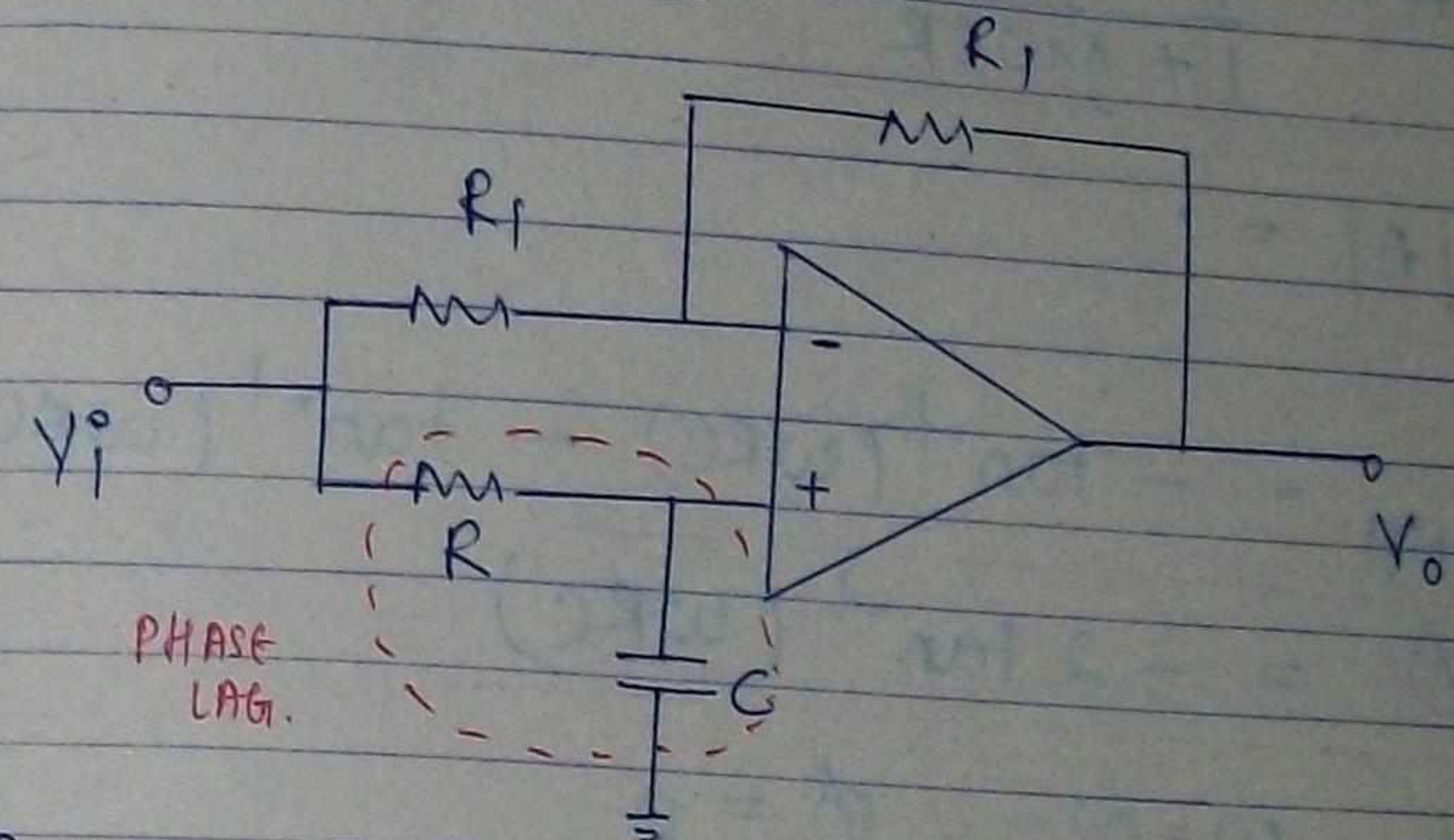


f_1 = cutoff freq of LPF

f_2 = cutoff freq of HPF.



(G) ALL PASS FILTER —



By superposition theorem:

$$V_o = -\frac{R_1}{R_1} V_i + \left(1 + \frac{R_1}{R_1}\right) V_p$$

$$V_p = \frac{1/sC}{R + 1/sC} V_i$$

$$V_p = \frac{V_i}{1 + sCR}$$

$$V_o = -V_i + \frac{2V_i}{1 + sCR}$$

$$A = \frac{V_o}{V_i} = \frac{1 - sCR}{1 + sCR}$$

$$A = \frac{1 - j\omega RC}{1 + j\omega RC}$$

$$|A| = 1$$

$$\phi_{\text{shift}} = -\tan^{-1}(\omega RC) - \tan^{-1}(\omega RC)$$

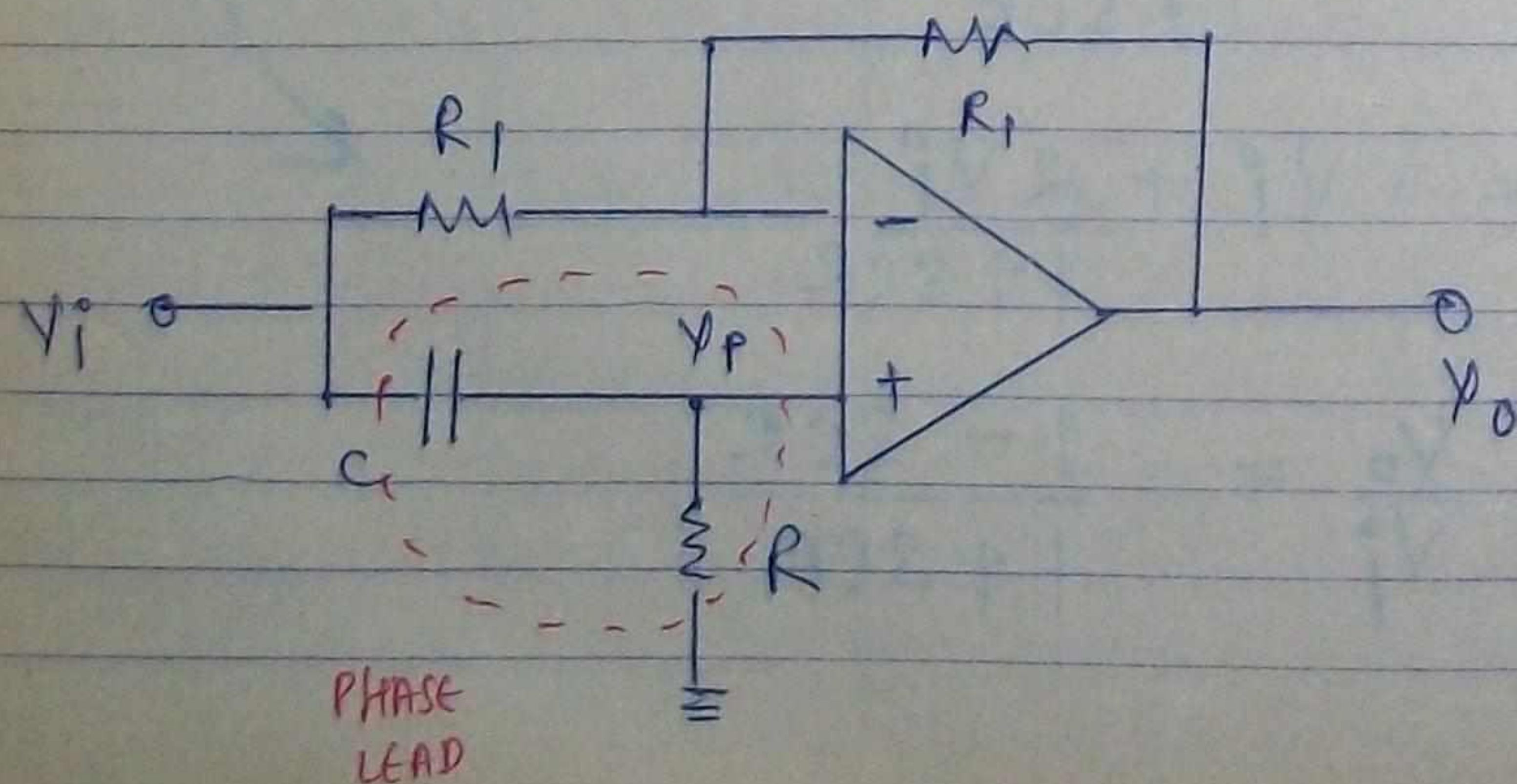
$$\phi = -2 \tan^{-1}(\omega RC)$$

$$\text{at } \omega = 0, \phi = 0$$

$$\omega = \infty, \phi = -\pi$$

Range of phase $\phi = 0 \text{ to } -\pi$
i.e. -ve phase.

for positive phase:-



$$V_p = \frac{R}{R+1/sC} \quad V_i^o = \frac{sCR}{1+sCR} V_i^o$$

$$V_o = -V_i^o + \frac{2sCR}{1+sCR} V_i^o$$

$$A = \frac{V_o}{V_i^o} = \frac{-1+sCR}{1+sCR} = -\frac{(1-sCR)}{(1+sCR)}$$

$$A = \frac{V_o}{V_i} = -\frac{(1-j\omega RC)}{(1+j\omega RC)}$$

$$|A| = 1$$

$$\phi = 180 - 2 \tan^{-1}(\omega RC)$$

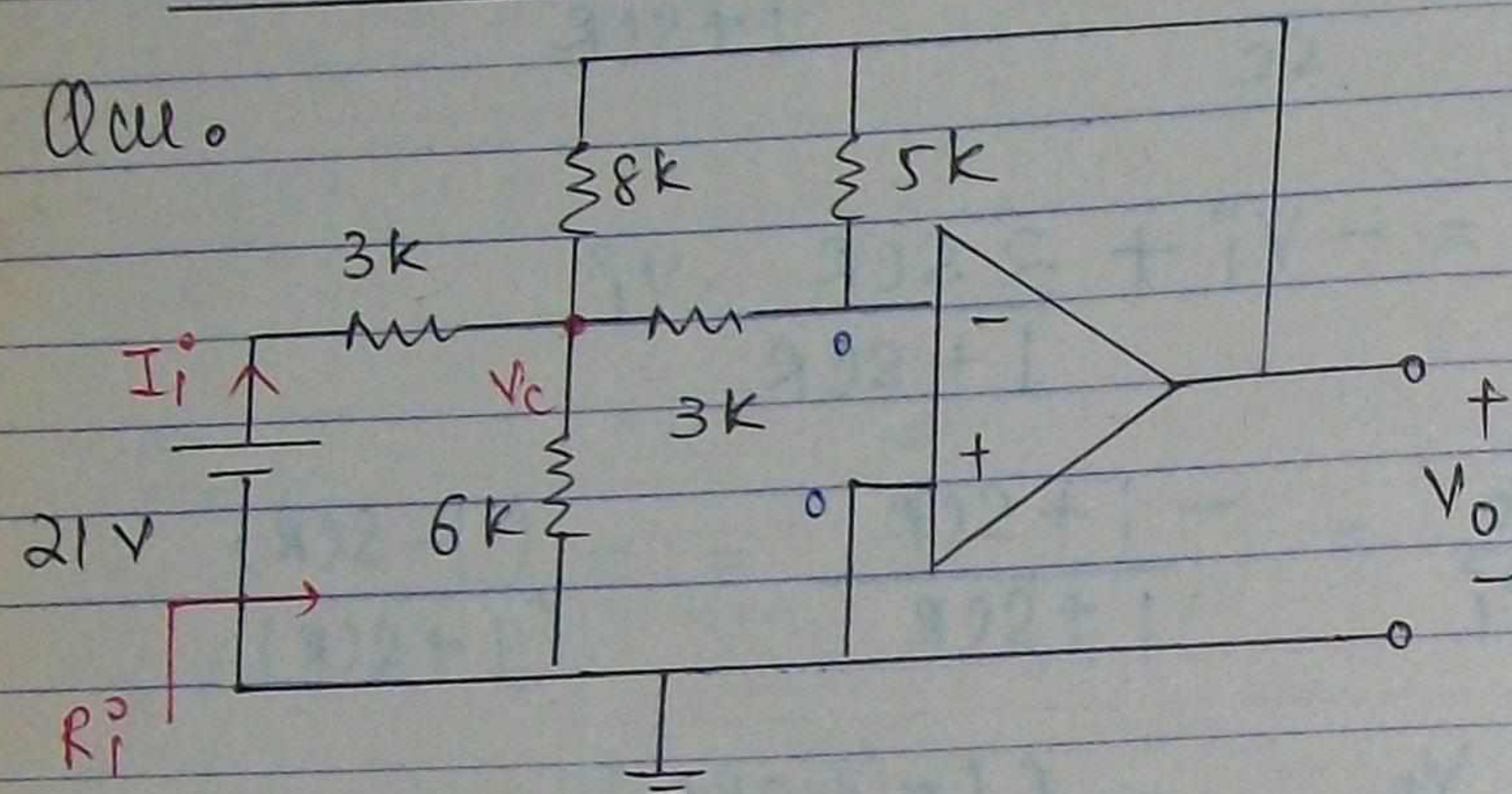
$$\omega = 0 \quad \phi = 180 = \pi$$

$$\omega = \infty \quad \phi = 0$$

$$\phi \text{ Range} = 0 \text{ to } \pi$$

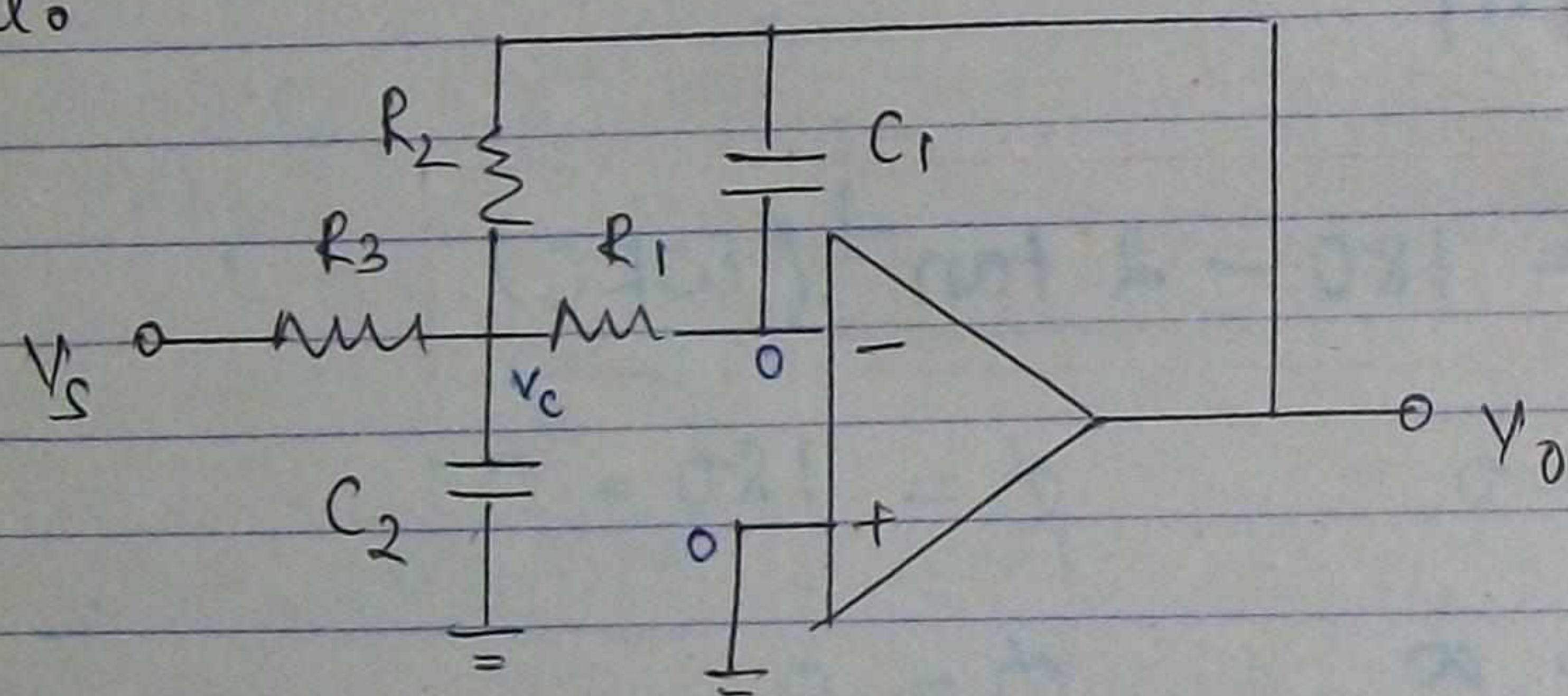
⇒ NUMERICALS -

Ques.



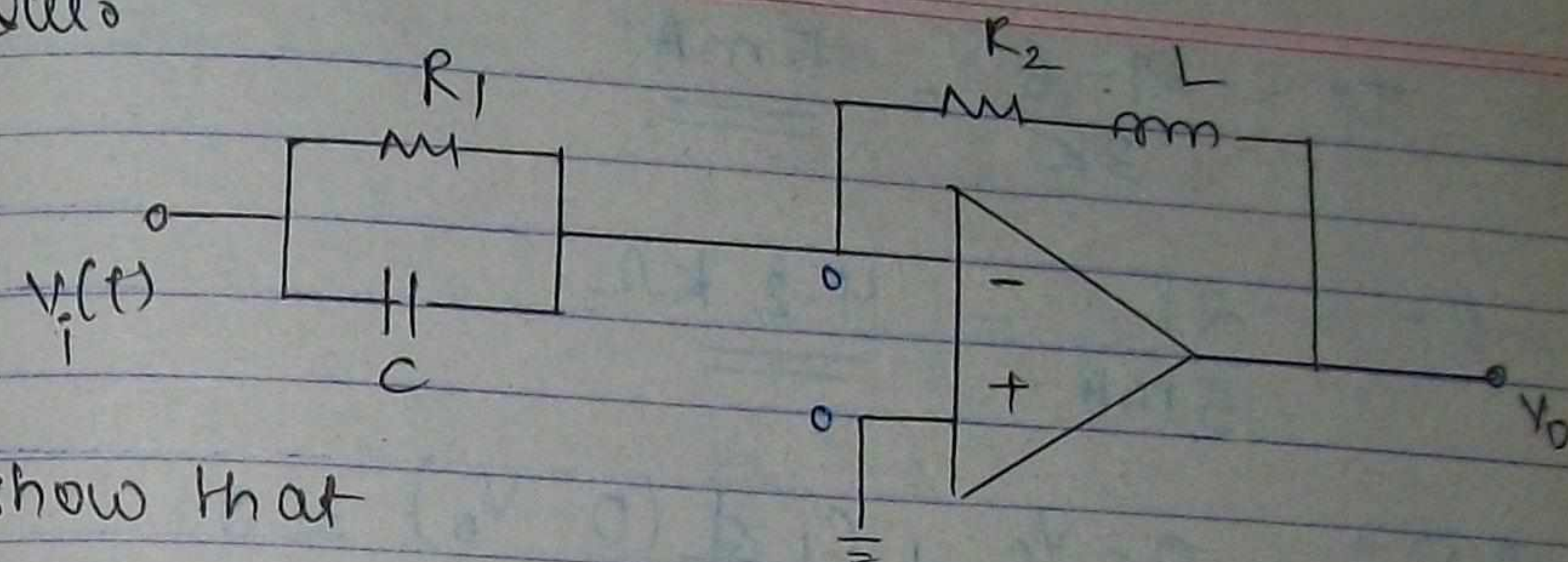
Calculate V_c , I_i , R_i^o

Ques.



for the circuit show and find $A_v = V_o/V_s$,
damping factor δ and cutoff
frequency. (ω_c).

Ques.



Show that

$$-V_o = \frac{R_2}{R_1} V + \left(R_2 C + \frac{L}{R_1} \right) \frac{dV}{dt} + LC \frac{d^2 V}{dt^2}$$

Solⁿ (1):

$$\frac{0 - V_c}{3k} + \frac{0 - V_o}{5k} = 0$$

$$-\frac{V_c}{3k} = \frac{V_o}{5k} \quad V_o = -V_c \frac{5k}{3k}$$

$$\text{also, } \frac{V_c - 21}{3k} + \frac{V_c}{6k} + \frac{V_c - 0}{3k} + \frac{V_c - V_o}{8k} = 0$$

$$V_c \left[\frac{1}{3k} + \frac{1}{6k} + \frac{1}{3k} + \frac{1}{8k} \right] = \frac{V_o}{8k} + \frac{21}{3k}$$

$$V_c \left[\frac{1}{3k} + \frac{1}{6k} + \frac{1}{3k} + \frac{1}{8k} + \frac{5k}{8 \cdot 3k^2} \right] = \frac{21}{3k}$$

$$V_c \left(\frac{7}{6} \right) = \frac{21}{3}$$

$$V_c = \frac{21}{3} \times \frac{6}{7} = 6 \quad V_c = \underline{\underline{6V}}$$

$$I_i = \frac{21-6}{3k} = \underline{\underline{5mA}}$$

$$R_i = \frac{21}{5mA} = \underline{\underline{4.2 k\Omega}}$$

Soln (2) : $\frac{0-V_c}{R_1} + C_1 \frac{d(0-V_o)}{dt} = 0$

$$\frac{V_c}{R_1} = -C_1 \frac{dV_o}{dt} \quad \text{--- (1)}$$

also,

$$\frac{V_c - V_s}{R_3} + \frac{V_c - V_o}{R_2} + \frac{V_c - 0}{R_1} + C_2 \frac{dV_c}{dt} = 0$$

$$\left(\frac{1}{R_3} + \frac{1}{R_2} + \frac{1}{R_1}\right) V_c - \left(\frac{V_s}{R_3} + \frac{V_o}{R_2}\right) + C_2 \frac{dV_c}{dt} = 0$$

$$C_2 \frac{d}{dt} \left[-R_1 C_1 \frac{dV_o}{dt} \right] + \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right) \left(-R_1 C_1 \frac{dV_o}{dt} \right) - \frac{V_s}{R_3} + \frac{V_o}{R_3} = 0$$

$$-R_1 C_1 C_2 \frac{d^2 V_o}{dt^2} + \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right) (-R_1 C_1) \frac{dV_o}{dt} + \frac{V_o}{R_3} = \frac{V_s}{R_3}$$

taking laplace on both side :

$$-R_1 C_1 C_2 S^2 V_o(s) + \left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right) (-R_1 C_1) S V_o(s) + \frac{1}{R_3} V_o(s) = \frac{V_s(s)}{R_3}$$

$$\frac{V_o}{V_s} = \frac{1/R_3}{S^2 (-R_1 C_1 C_2) - R_1 C_1 (R_1 R_2 + R_2 R_3 + R_3 R_1) S + \frac{1}{R_3}}$$

$$\frac{V_o}{V_s} = \frac{1}{S^2 (-R_1 R_3 C_1 C_2) - \frac{C_1}{R_2} (R_1 R_2 + R_2 R_3 + R_3 R_1) S + 1} = \frac{1}{S^2 + \frac{C_1}{R_2} (R_1 R_2 + R_2 R_3 + R_3 R_1) - \frac{1}{R_1 R_3 C_1 C_2}}$$

$$\frac{V_o}{V_s} = \frac{1}{S^2 + \cancel{\frac{C_1}{R_2} (R_1 R_2 + R_2 R_3 + R_3 R_1)} S - \frac{1}{R_1 R_3 C_1 C_2}}$$

$$\frac{V_o}{V_s} = \frac{1}{s^2 + \frac{(R_1 R_2 + R_2 R_3 + R_3 R_1)}{R_1 R_2 R_3 C_2} s - \frac{1}{R_1 R_3 C_1 C_2}}$$

$$\omega_n = \sqrt{\frac{-1}{R_1 R_3 C_1 C_2}}$$

$$2\zeta\omega_n = \frac{(R_1 R_2 + R_2 R_3 + R_3 R_1)}{R_1 R_2 R_3 C_2}$$

$$\zeta = \frac{(R_1 R_2 + R_2 R_3 + R_3 R_1)}{R_1 R_2 R_3 C_2}$$

$$2 \sqrt{\frac{-1}{R_1 R_3 C_1 C_2}}$$

Cutoff frequency ω_c

$$\frac{V_o}{V_s} = \frac{-R_2}{s^2 R_1 R_2 R_3 C_1 C_2 + s C_1 [R_1 R_2 + R_2 R_3 + R_3 R_1] + R_3}$$

$$= \frac{-R_2}{\frac{R_1 R_2 R_3 C_1 C_2}{s^2 + \frac{s C_1 [R_1 R_2 + R_2 R_3 + R_3 R_1]}{R_1 R_2 R_3 C_1 C_2} + \frac{R_3}{R_1 R_2 R_3 C_1 C_2}}}$$

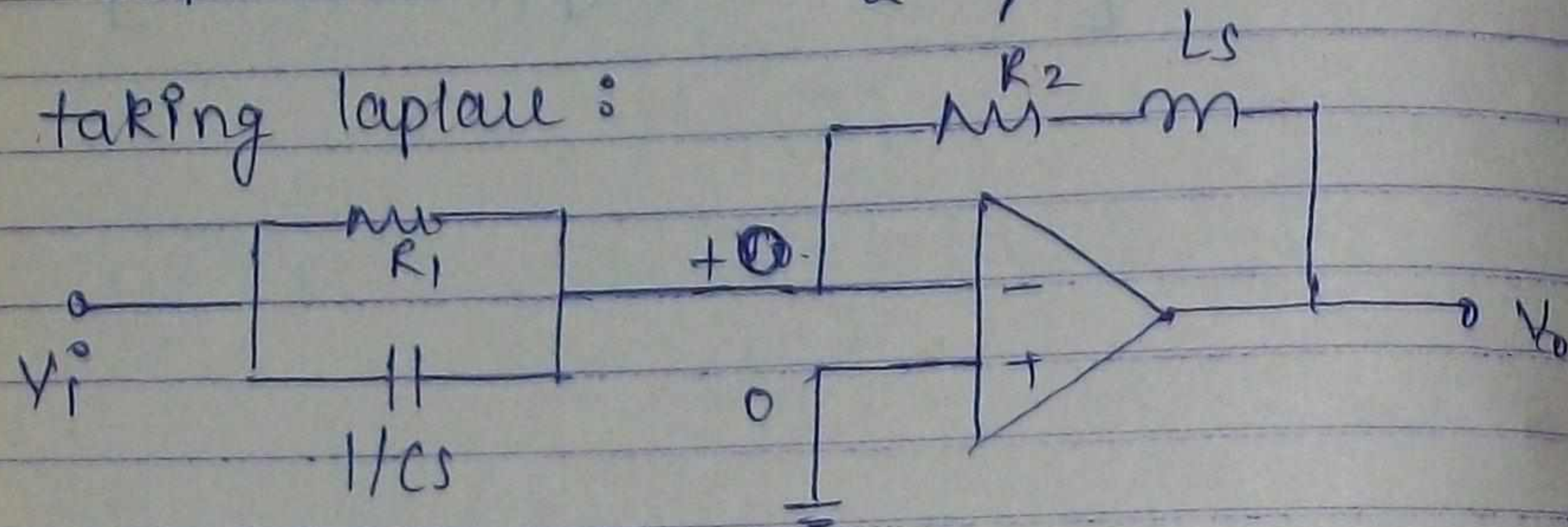
$$\zeta = \frac{1}{2} \sqrt{\frac{R_1 R_2 C_1}{C_2} \left[\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3} \right]}$$

$$\omega_c = \omega_n \left[(1 - 2\zeta^2) + \sqrt{(1 - 2\zeta^2)^2 + 1} \right]^{1/2}$$

Solⁿ (3):

$$\frac{0-V}{R_1} + \frac{C d(0-V)}{dt} + \frac{(0-V_0)}{R_2 + L \frac{d}{dt}} = 0$$

taking laplace:



$$Z_1 = \frac{R_1 \times 1/cs}{R_1 + \frac{1}{Cs}} = \frac{R_1}{R_1 Cs + 1}$$

$$Z_2 = R_2 + Ls$$

$$\frac{V_i}{1 + R_1 Cs} = \frac{0 - V_i}{R_1} = \frac{0 + V_0}{R_2 + Ls}$$

$$\frac{-V_0}{(R_2 + Ls)} = + \frac{V_i (1 + R_1 Cs)}{R_1}$$

$$\frac{-V_0}{V_i} = \frac{(R_2 + Ls)(1 + R_1 Cs)}{R_1}$$

$$\frac{-V_0}{V_i} = \frac{R_2}{R_1} + \left(R_2 C + \frac{L}{R_1} \right) s + L C s^2$$

$$-V_0 = \left[\frac{R_2}{R_1} + \left(R_2 C + \frac{L}{R_1} \right) s + L C s^2 \right] V_i$$

$$-V_0 = \frac{R_2}{R_1} V_i + \left(R_2 C + \frac{L}{R_1} \right) \frac{dV_i}{dt} + L C \frac{d^2 V_i}{dt^2}$$

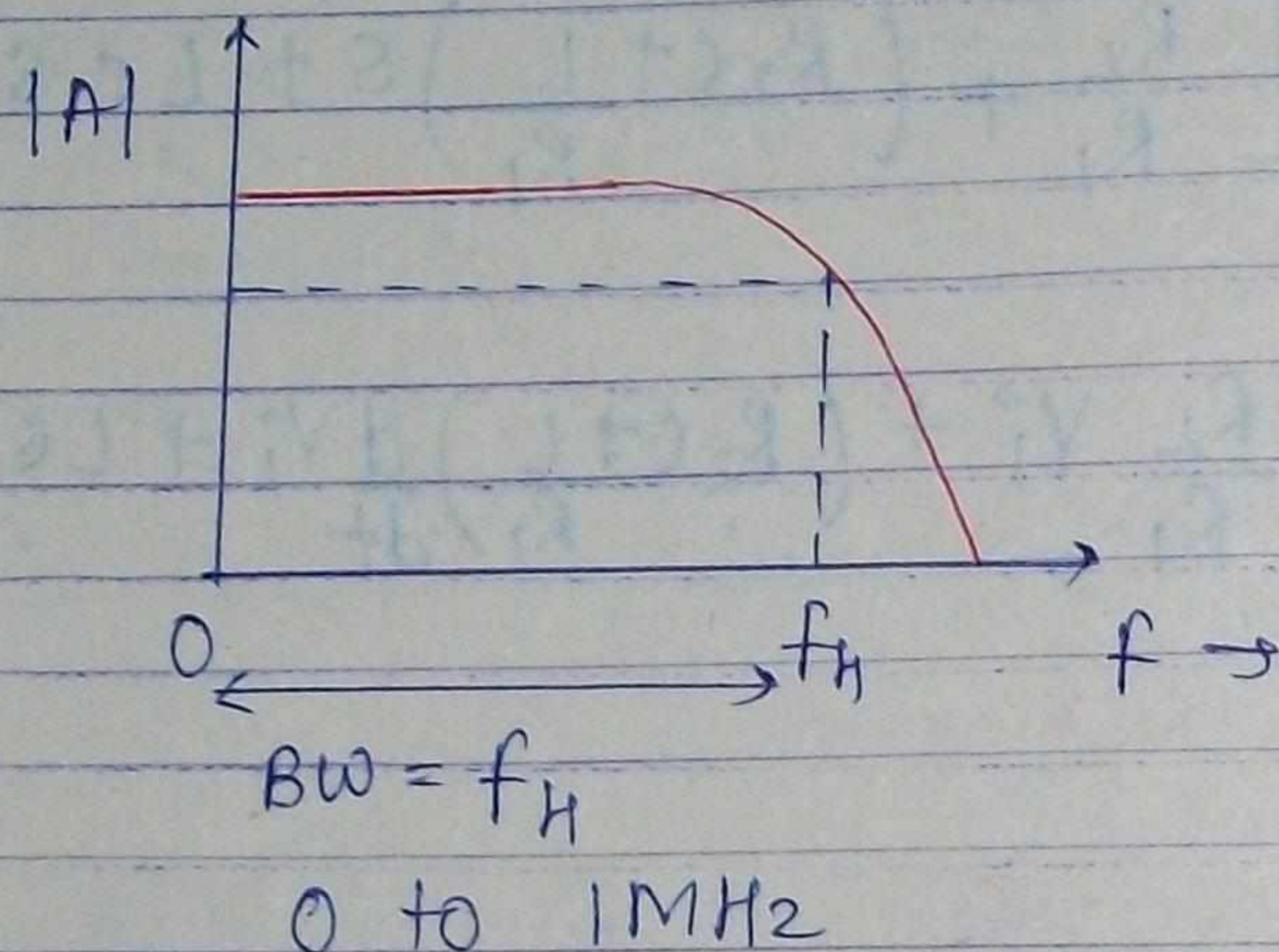
(*) SHORT TRICK —

$$A_v = - \frac{Z_2}{Z_1}$$

FREQUENCY RESPONSE OF PRACTICAL OPAMP

22/08/11

- frequency response of an OPAMP



Gain $\times$ bandwidth = product is constant

Gain $\uparrow$ BW $\downarrow$

(or)

Gain $\downarrow$ BW $\uparrow$

- OPAMP is basically a dc amplifier. It can amplify a DC signal and also an a.c. signal in a wide-band extending from 0 Hz to 1 MHz.

$\Rightarrow$ SLEW RATE (SR) -

- The slew rate is the time rate of change of closed loop amplifier output voltage under large signal condition. (dc)
- The slew rate will indicate how quick is the OPAMP in responding to small variation in I/P parameter in a time interval of 1 μ s.
- SR (slew rate) must be large. For ideal OPAMP $SR = \infty$ (IDEAL)

$$SR = \left. \frac{dV_o}{dt} \right|_{\max}$$

$$SR = \left. \frac{dV_o}{dV_i} \times \frac{dV_i}{dt} \right|_{\max}$$

$SR = A_{cl} \times \left. \frac{dV_i}{dt} \right|_{\max}$

Now,

let $V_i = V_m \sin \omega t$

$$\frac{dV_i}{dt} = V_m \omega \cos \omega t$$

$$= V_m \omega \cos \omega t$$

$$\left. \frac{dV_i}{dt} \right|_{\max} = V_m \omega_m$$

$$SR = A_{cl} \times V_m \omega_m$$

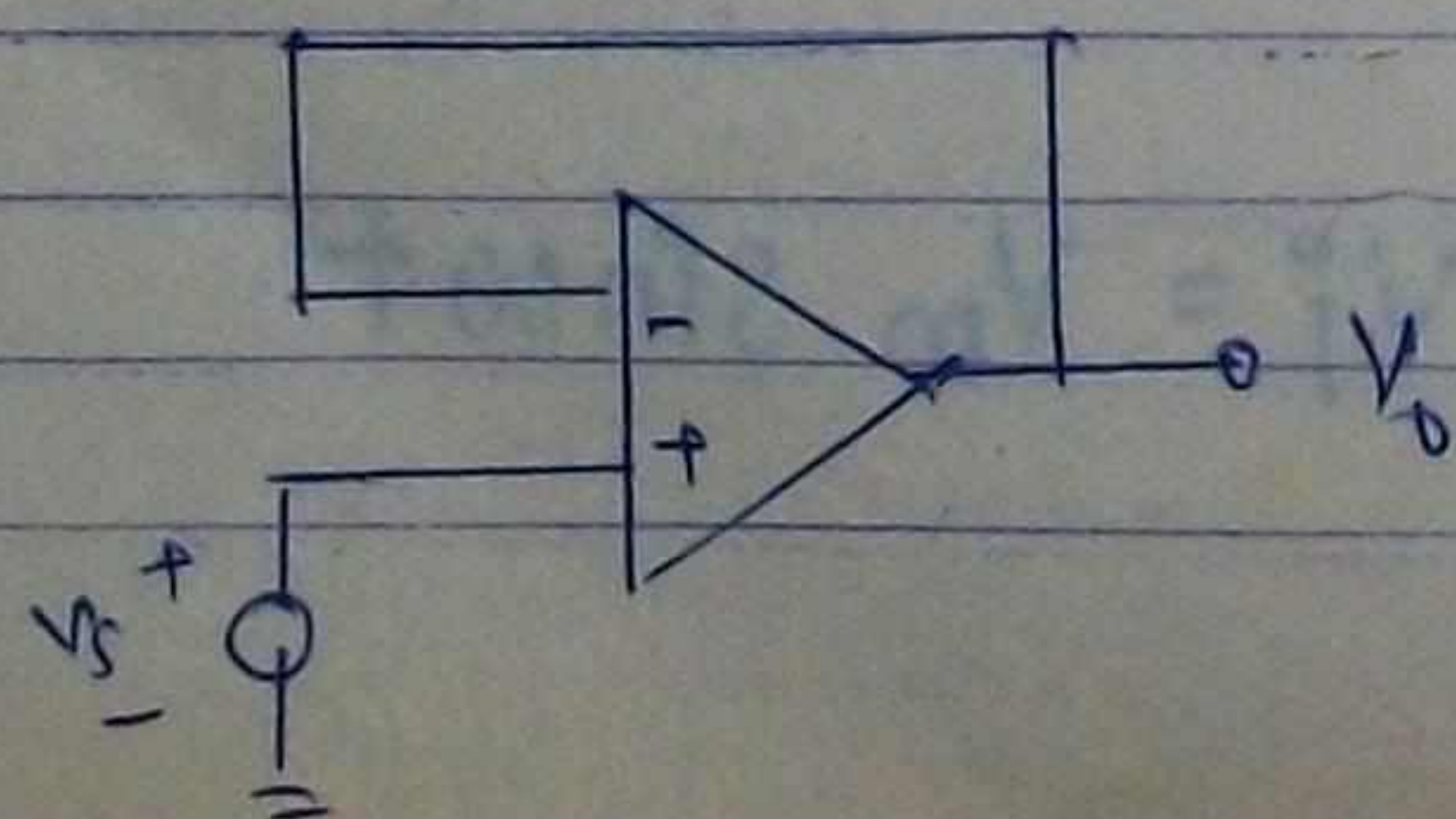
$$\omega_m = 2\pi f_m = \frac{SR}{|A_{cl}| V_m}$$

$$f_m = \frac{SR}{|A_{cl}| \cdot 2\pi V_m}$$

∴ $f \leq f_m$ There is 'NO DISTORTION'

$f > f_m$ There is 'DISTORTION'

for voltage follower:

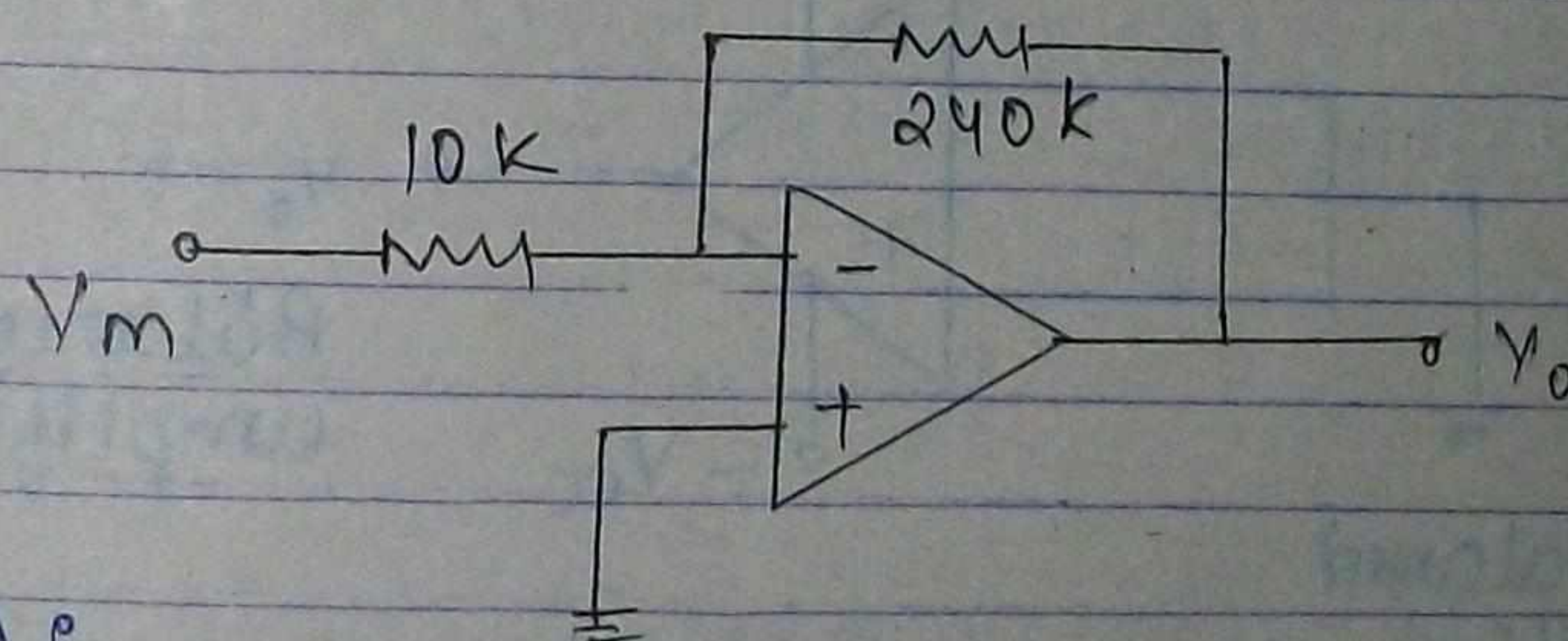


$$A_{cl} = 1$$

$$f_m = \frac{SR}{2\pi V_m}$$

Q. ① for an OPAMP having $SR = 2V/\mu s$ what is the max A_{cl} that can be used when i/p signal changes by $0.5V$ in $10\mu s$.

② for an OPAMP circuit determine max freq. of operation in radian/sec. that can be used by taking $SR = 0.5V/\mu s$ and $V_m = 0.02V$



Sol ① :

$$SR = A_{cl} \times \frac{dV_i}{dt}$$

$$2 = A_{cl} \times \frac{0.5}{10 \times 10^{-6}}$$

$$A_c = \frac{40 \times 10^{-6} \times 10^{-6}}{10^{-6}} = \underline{40}$$

$$② \quad |A_v| = -\frac{240}{10} = -24 = 24$$

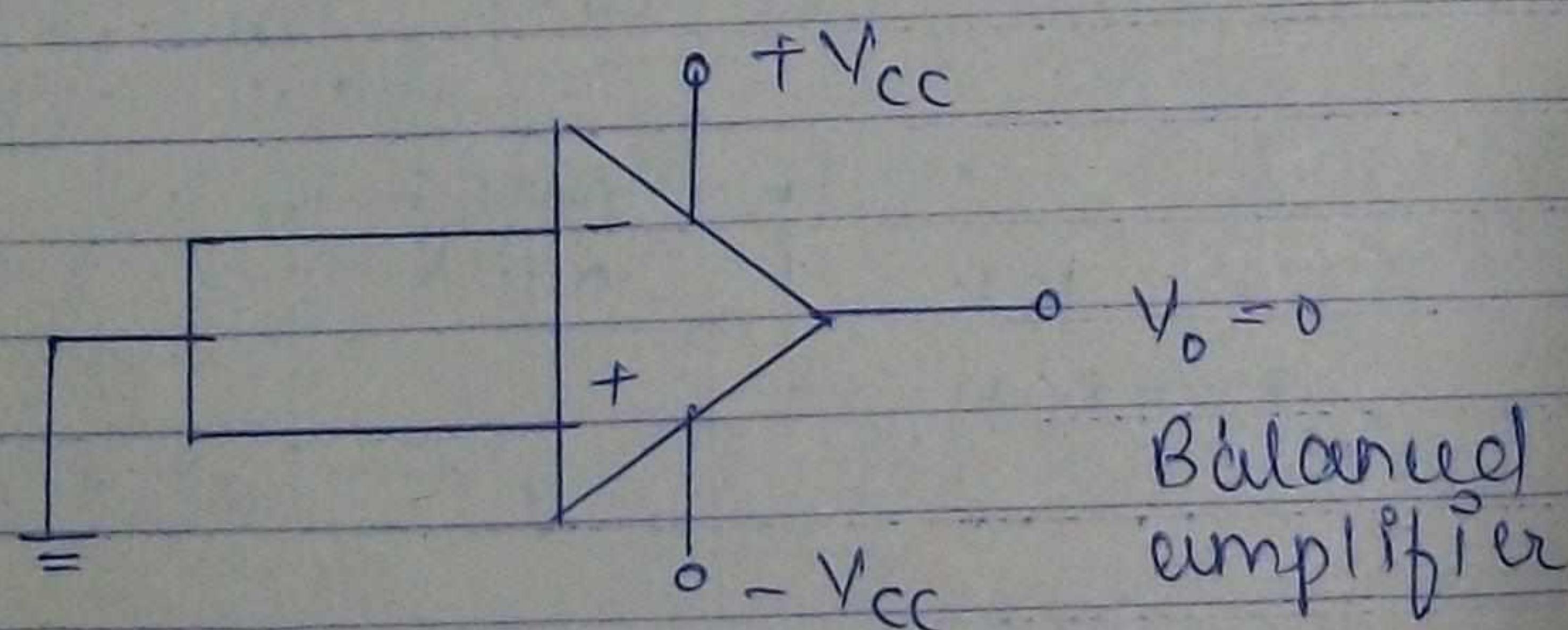
$$\omega_m = \frac{SR}{A_{cl} \times V_m}$$

$$\omega_m = \frac{0.5 \times 10^6}{+24 \times 0.02}$$

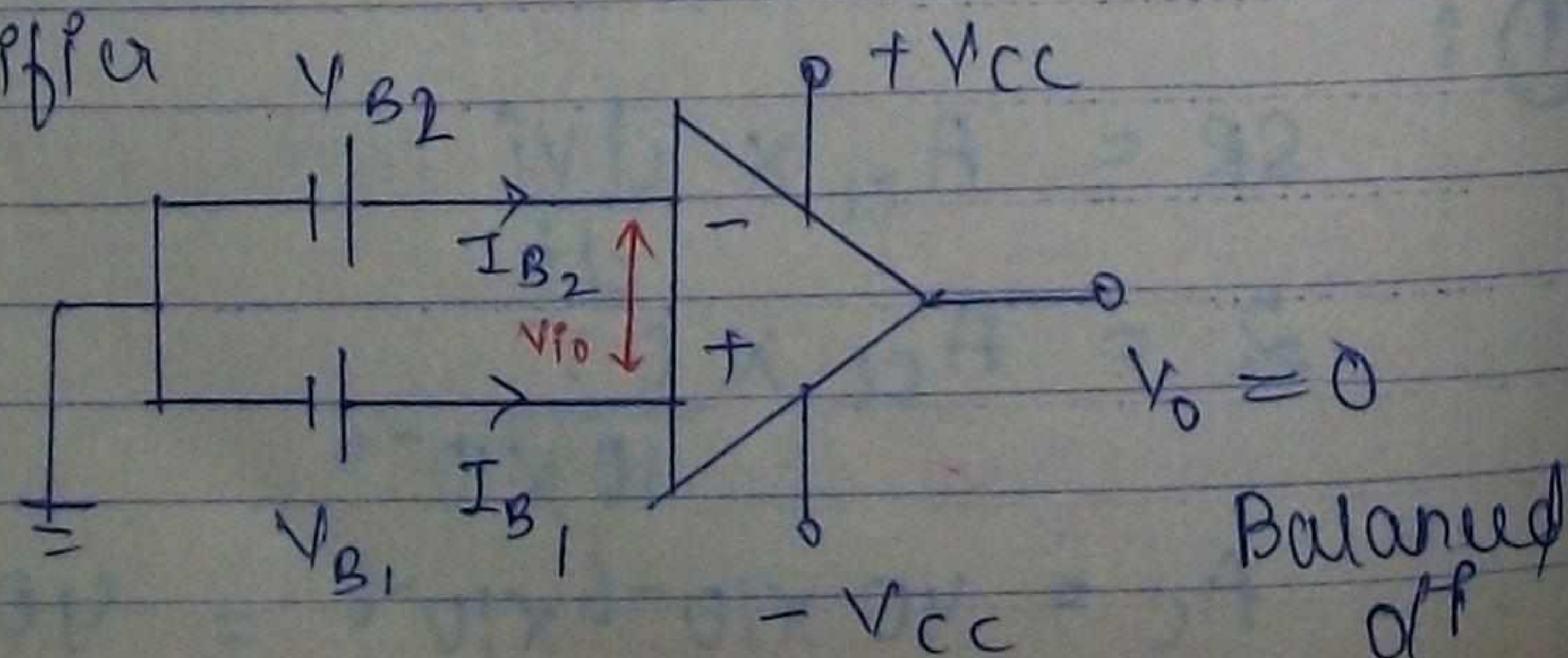
$$= 1.04 \times 10^6 \text{ rad/sec}$$

$$= 1.04 \times 10^6 \text{ rad/sec}$$

⇒ OFFSET VOLTAGE AND CURRENT -



for unbalanced amplifier



OUTPUT OFFSET VOLTAGE (V_{oo}) -

It is the difference between dc voltage present at output terminal and ground when two input terminals are grounded.

INPUT BIAS CURRENT -

This is one half of the sum of separate currents entering two i/p terminals, of a balanced amplifier.

$$I_B = \frac{I_{B1} + I_{B2}}{2}$$

INPUT OFFSET CURRENT (I_{io}) -

It is the difference between separate currents entering the i/p terminals of a balanced amplifier.

$$I_{io} = I_{B1} - I_{B2}$$

INPUT OFFSET CURRENT DRIFT -

It is the ratio of change of input offset current to change of temperature

$$\text{INPUT OFFSET CURRENT DRIFT} = \frac{\Delta I_{io}}{\Delta T}$$

INPUT OFFSET VOLTAGE (V_{io}) -

It is the voltage which must be applied between I/P terminals to balance amplifier.

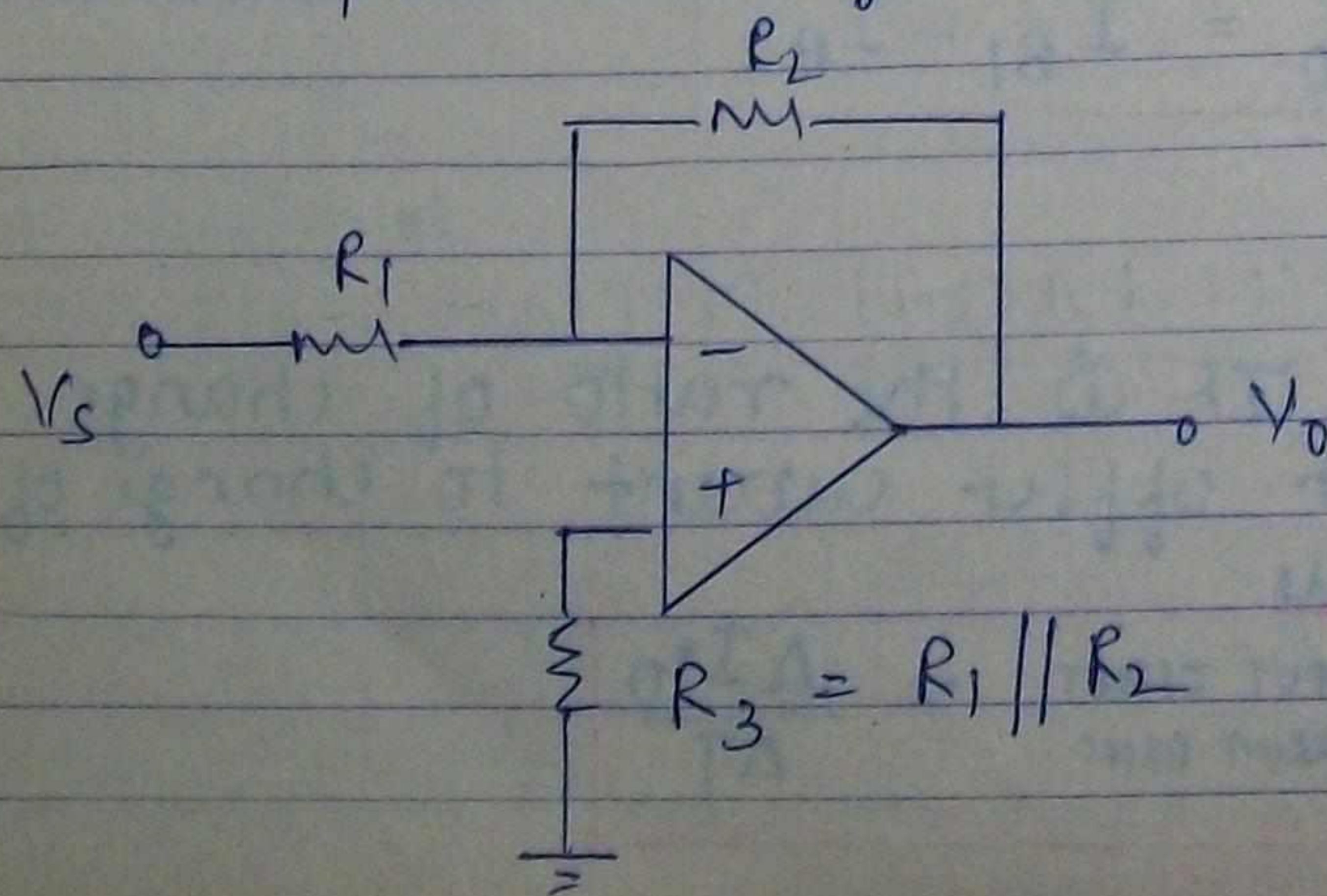
INPUT OFFSET VOLTAGE DRIFT -

Rate of change of V_{io} w.r.t change in T .

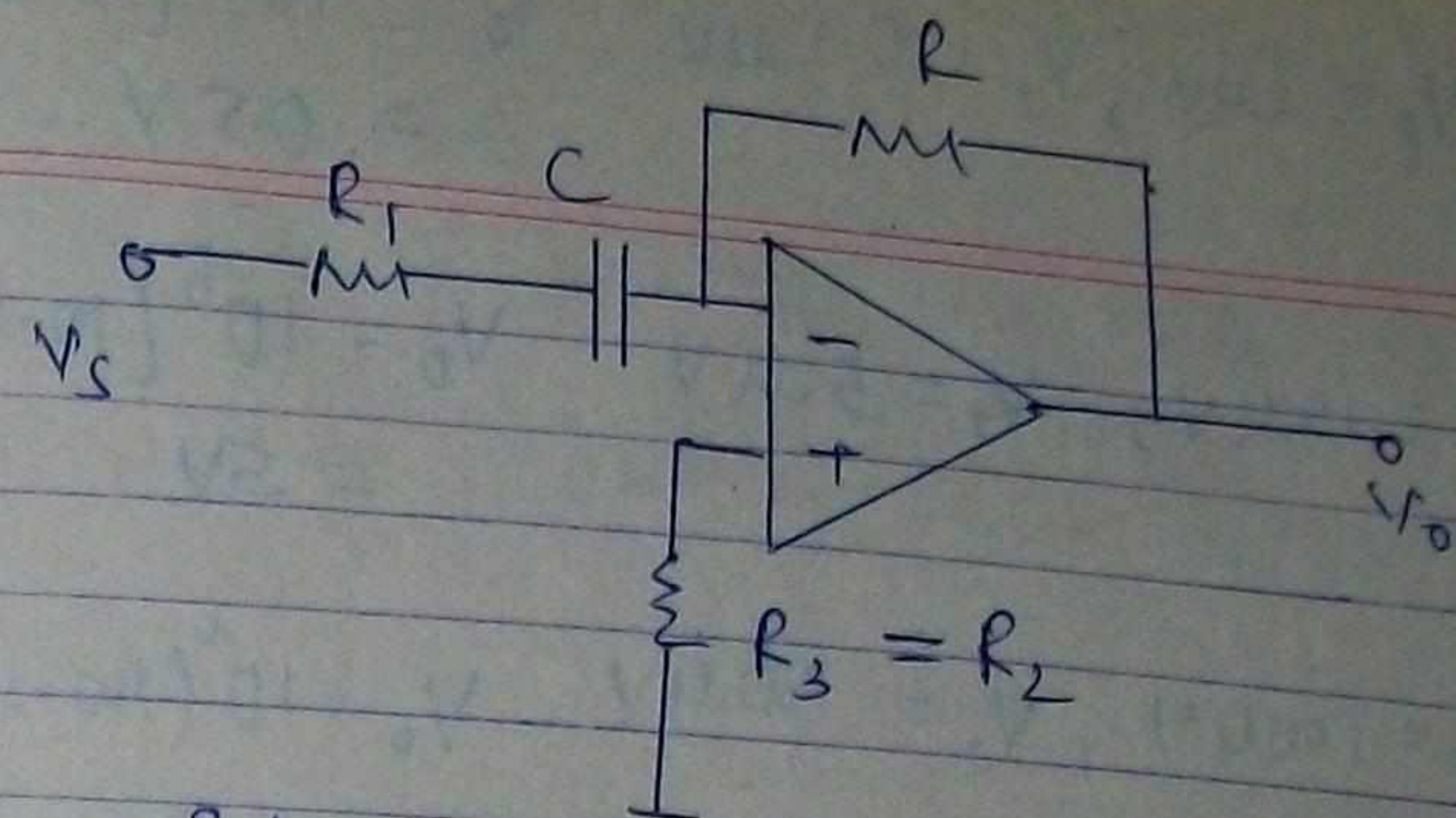
$$\text{INPUT OFFSET VOLTAGE DRIFT} = \frac{\Delta V_{io}}{\Delta T}$$

#NOTE -

- To minimize the effect of input bias current one should place in the positive lead (non-inverting terminal) A resistance equal to dc resistance seen by inverting terminal.



#



DC resistance seen by the inverting terminal.

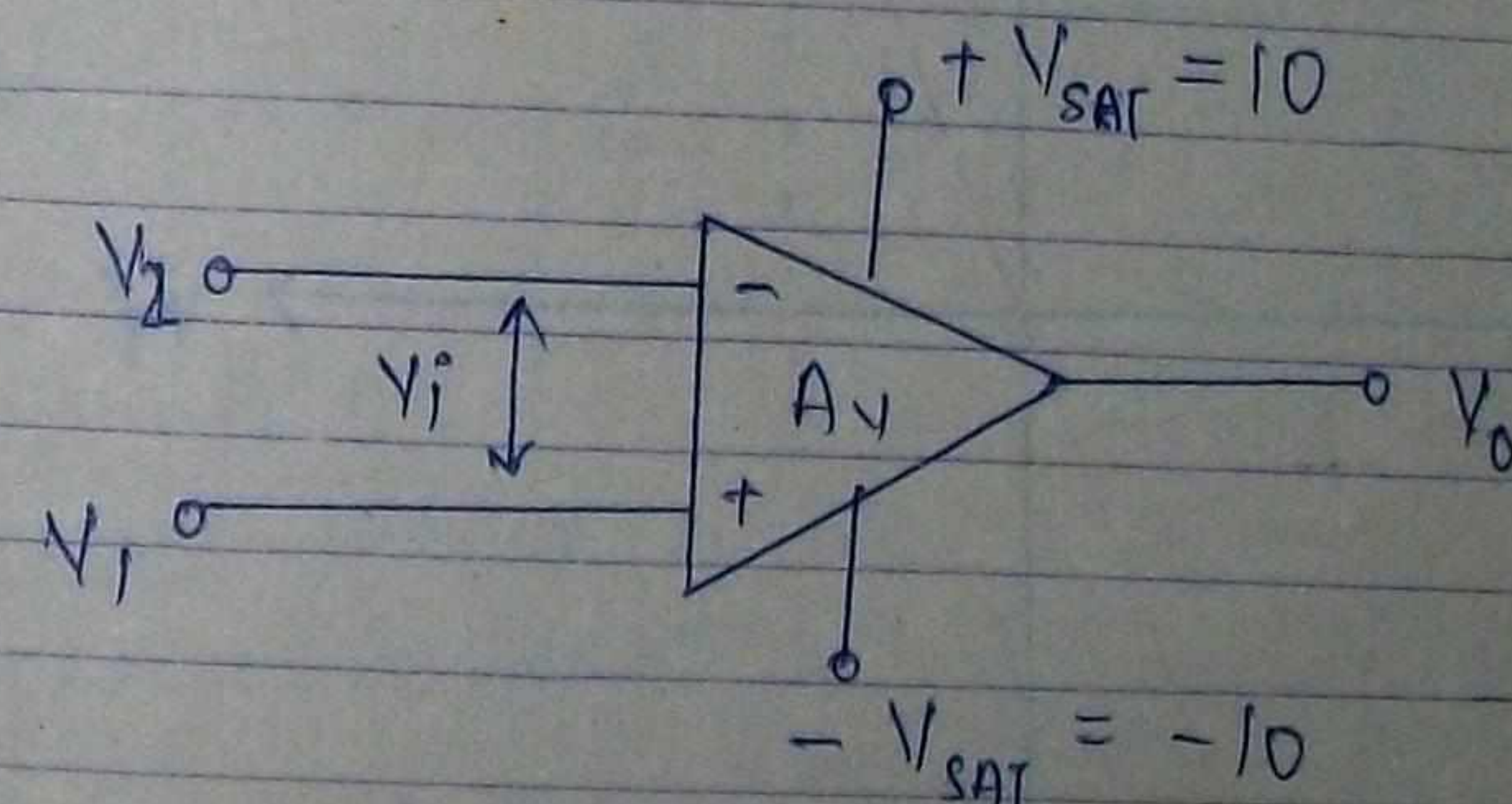
$$R_3 = R_2 || (R_1 + Z_c)$$

$$R_3 = R_2 || (R_1 + \infty)$$

$Z_c = \infty$
for dc

$$R_3 \approx R_2$$

⇒ TRANSFER CHARACTERISTICS OF PRACTICAL OPAMP -



$$A_v = 10^6$$

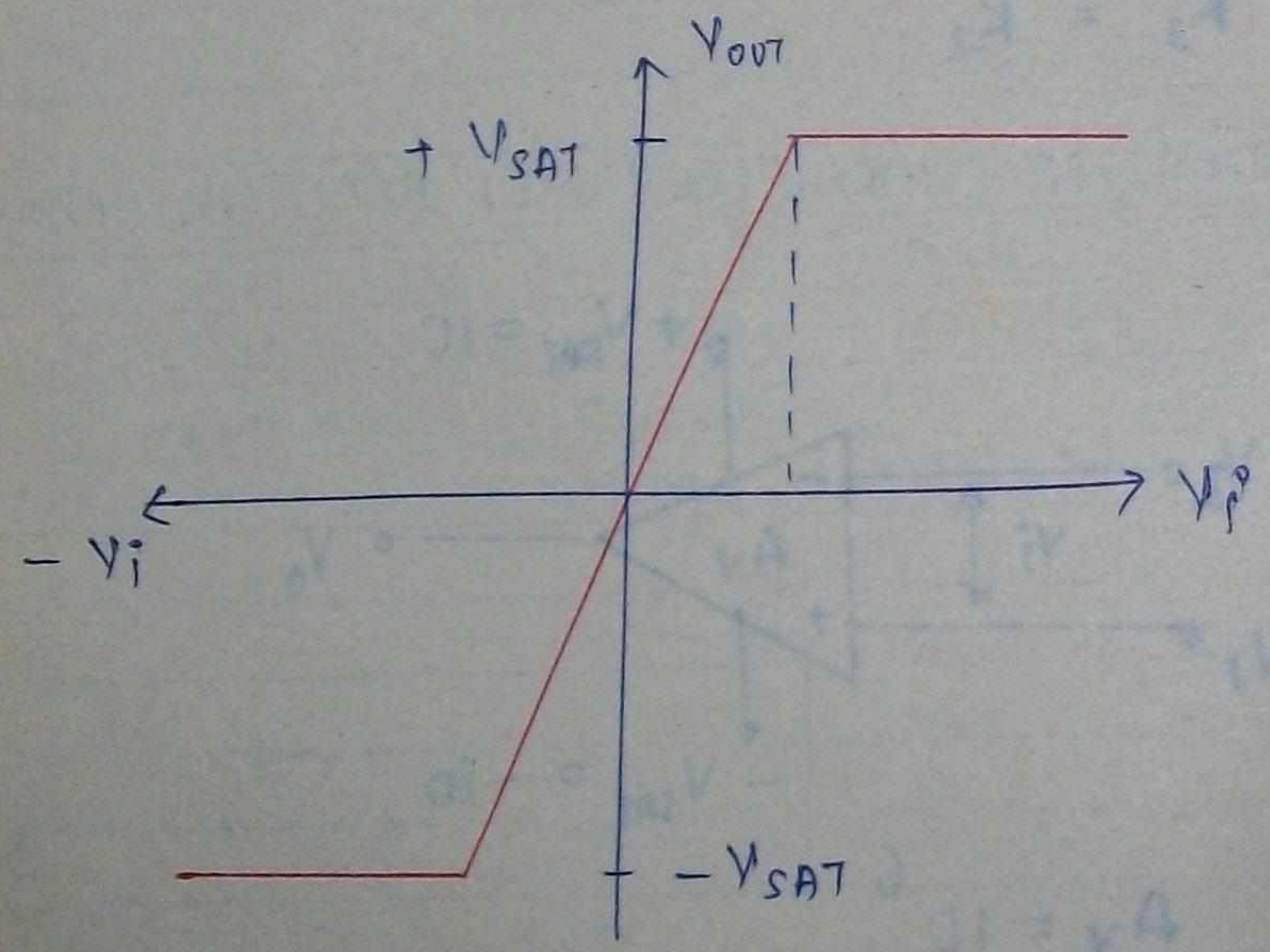
$$V_o = A_v V_i = A_v [V_1 - V_2]$$

for $V_1 = 1\mu V$, $V_2 = 0.5\mu V$ $V_o = 10^6 [1 - 0.5] \times 10^{-6} = 0.5V$

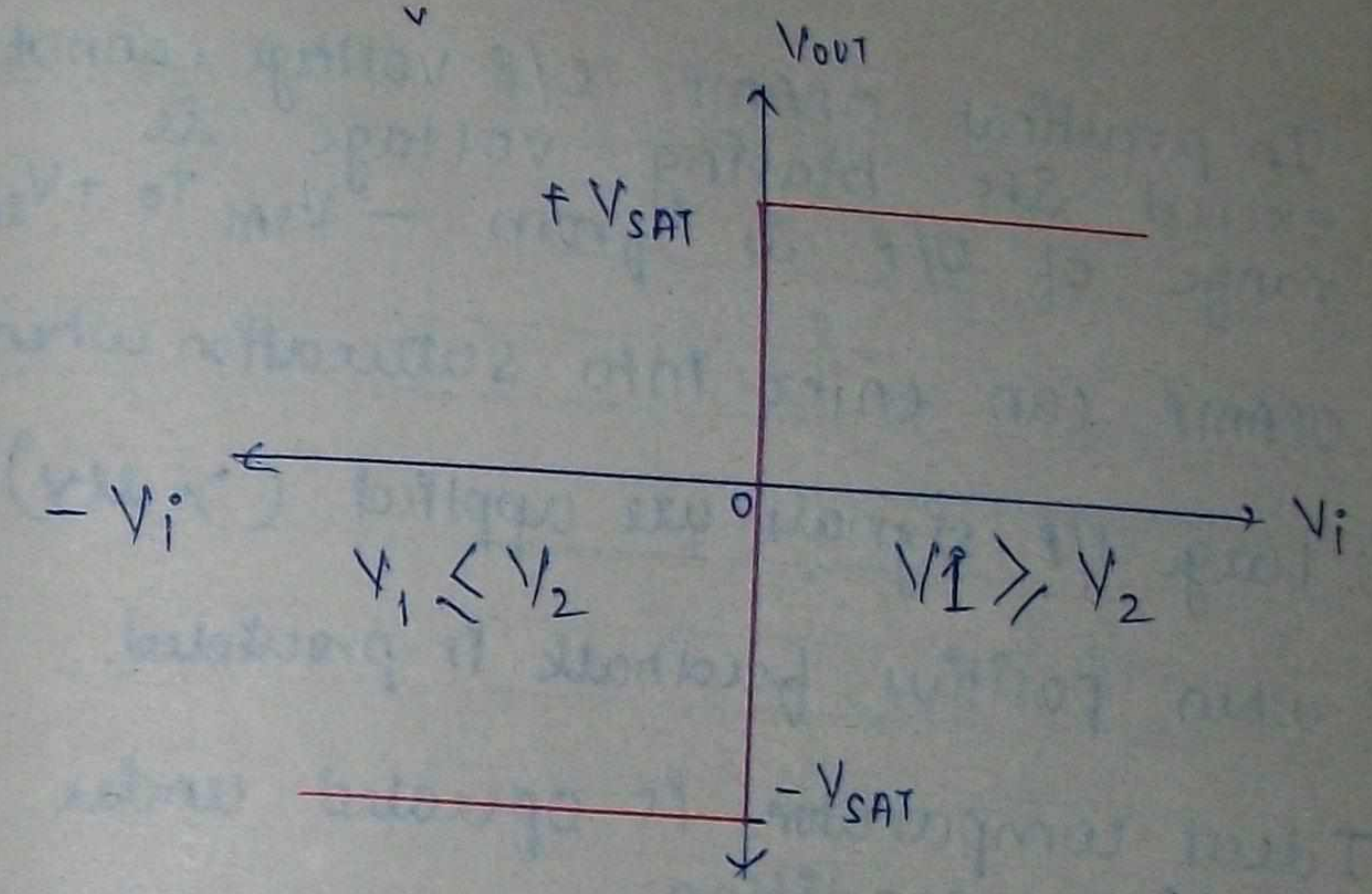
for $V_1 = 10\mu V$, $V_2 = 5\mu V$ $V_o = 10^6 [10 - 5] \times 10^{-6} = 5V$

for $V_1 = 100\mu V$, $V_2 = 50\mu V$ $V_o = 10^6 (100 - 50) \times 10^{-6} = 50V$
 $V_o = 10 (+V_{CC})$ not possible

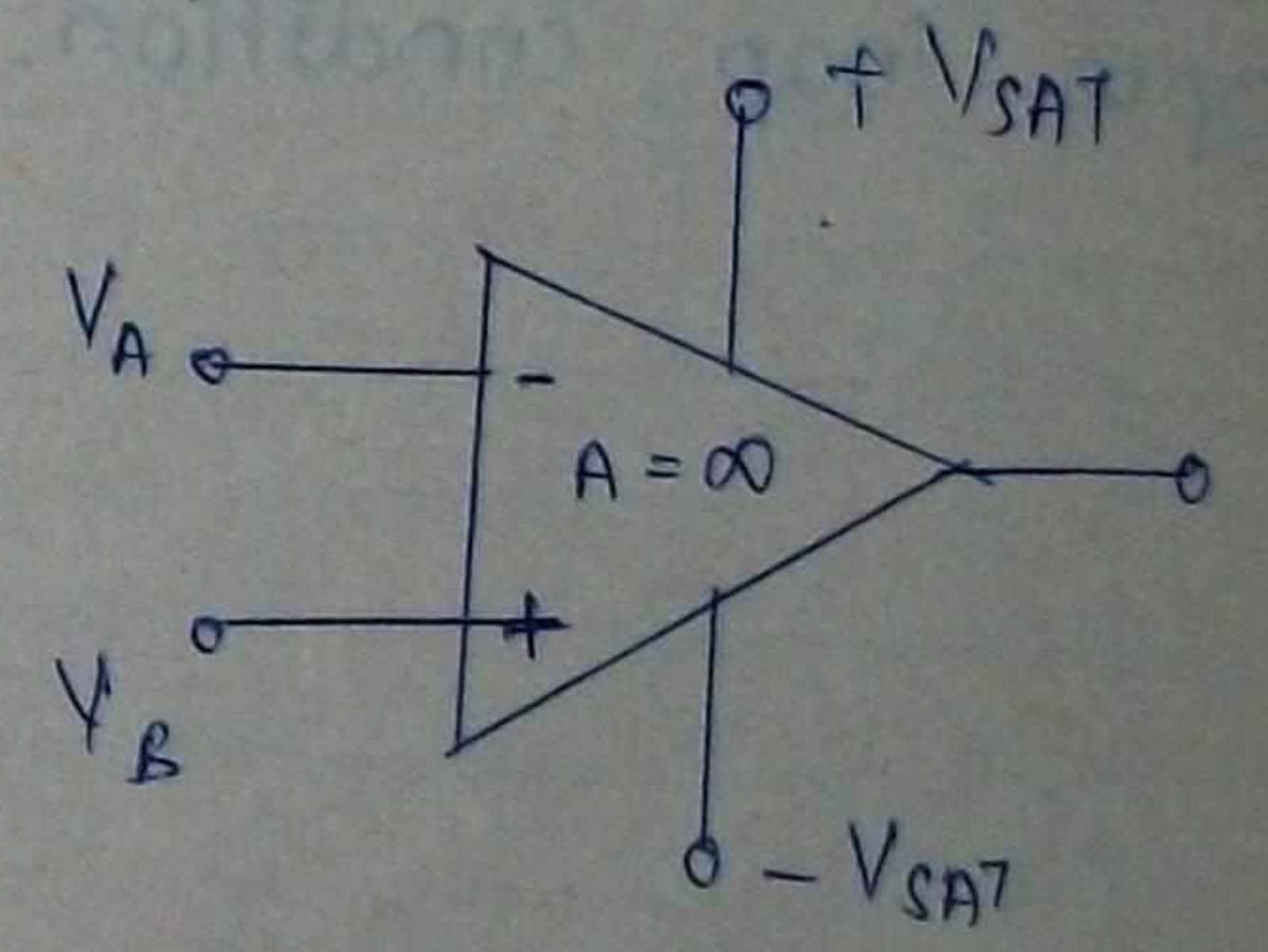
$\therefore$ Range of O/P = $[-V_{SAT}$ to $+V_{SAT}]$



$\Rightarrow$ TRANSFER CHARACTERISTIC OF IDEAL OPAMP -
 Gain, $A_v = \infty$

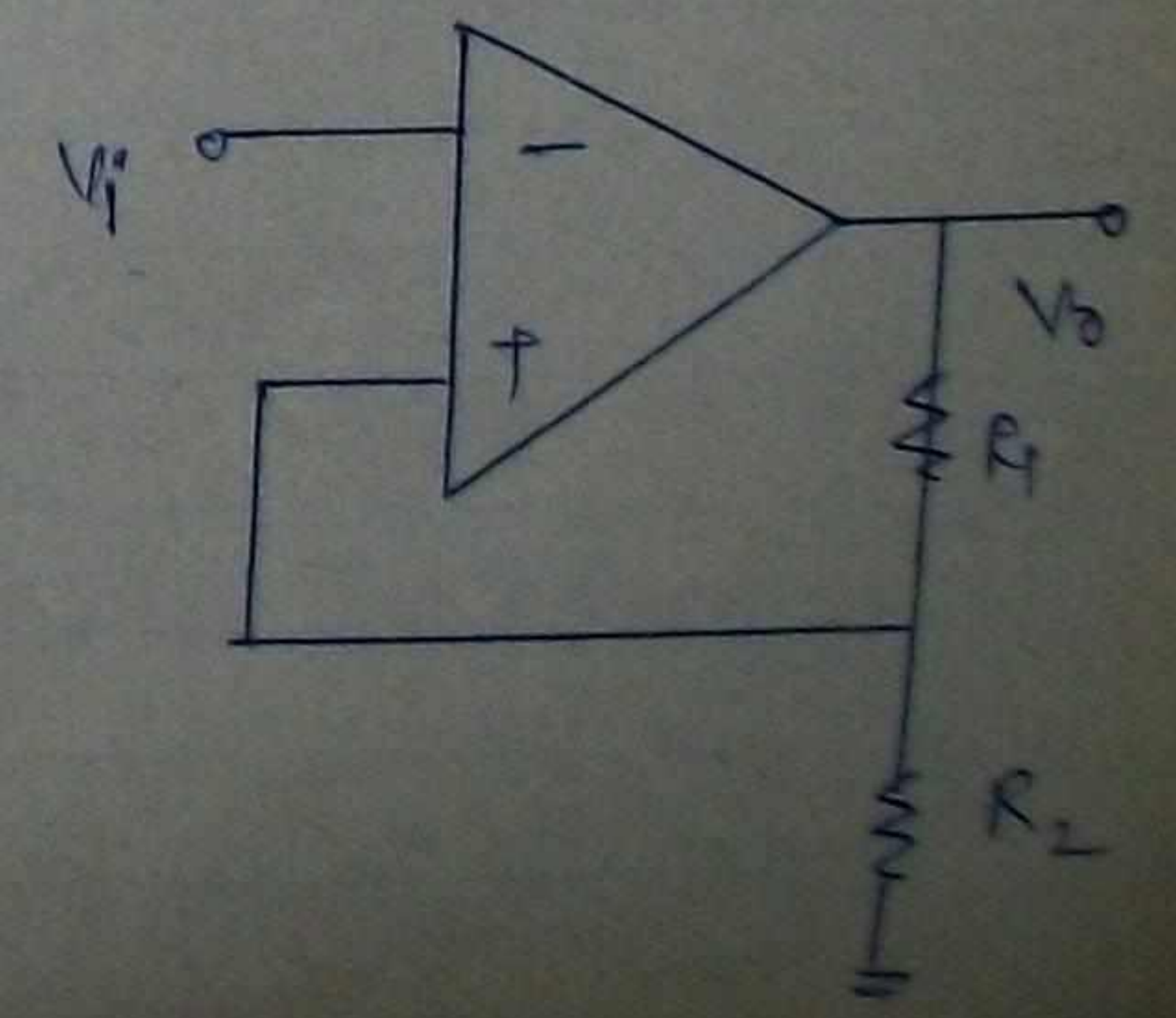
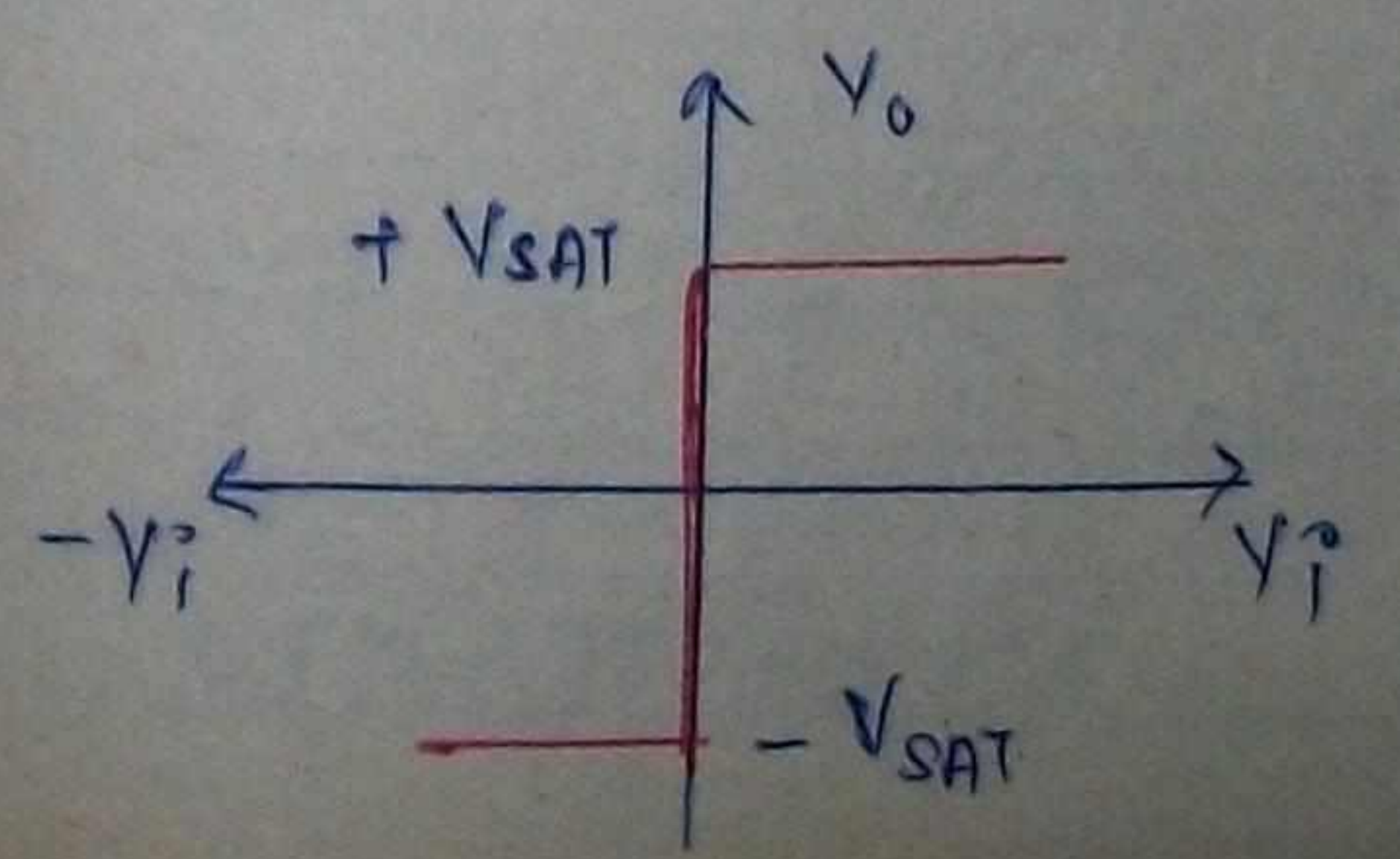


$V_B > V_A$	$+V_{SAT}$	1
$V_B < V_A$	$-V_{SAT}$	0



COMPARATOR -

PRACTICAL OPAMP WITH POSITIVE FEED BACK



- In practical OPAMP o/p voltage cannot exceed its biasing voltage i.e. range of o/p is from $-V_{SAT}$ to $+V_{SAT}$

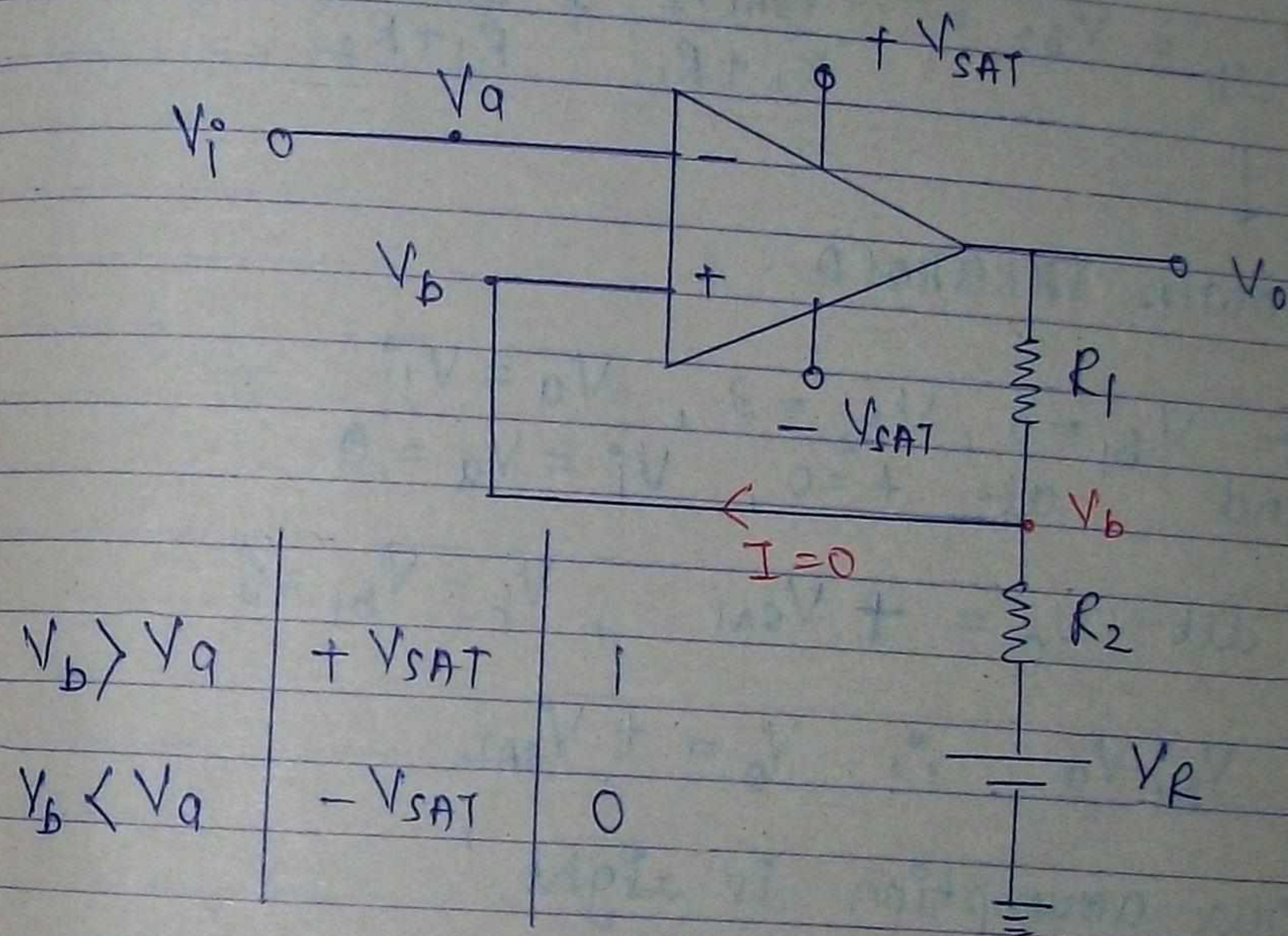
- OPAMP can enter into saturation when

① large i/p signals are applied ($> mV$)

② when positive feedback is provided.

- Ideal comparator is operated under open loop condition.

⇒ SCHMITT TRIGGER -



$V_b > V_a$	$+V_{SAT}$	1
$V_b < V_a$	$-V_{SAT}$	0

$$V_b = \frac{V_o R_2}{R_1 + R_2} + \frac{V_R R_1}{R_1 + R_2} \quad \left(\begin{array}{l} \text{by} \\ \text{superposition} \end{array} \right)$$

① $V_o = +V_{SAT}$

$$V_{UTH} = V_{b1} = \frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_R R_1}{R_1 + R_2}$$

↓
UPPER THRESHOLD

$$V_{UTH} > V_{LTH}$$

$$\textcircled{2} \quad V_o = -V_{SAT}$$

$$V_{LTH} = V_{b2} = \frac{-V_{SAT} R_2}{R_1 + R_2} + \frac{V_R R_1}{R_1 + R_2}$$



LOWER THRESHOLD.

let $V_{b1} = 6$, $V_{b2} = 3$, $V_a = V_i^\circ$
and at $t=0$, $V_i^\circ = V_a = 0$

$$\textcircled{1} \quad \text{let } V_o = \pm V_{SAT}, \quad V_b = V_{b1} = 6$$

$$V_b > V_a \quad \therefore V_o = +V_{SAT}$$

'Our assumption is right.'

$$\textcircled{2} \quad V_i^\circ \uparrow \text{ when } V_i^\circ = V_a \geq V_b$$

o/p switch from $+V_{SAT}$ to $-V_{SAT}$

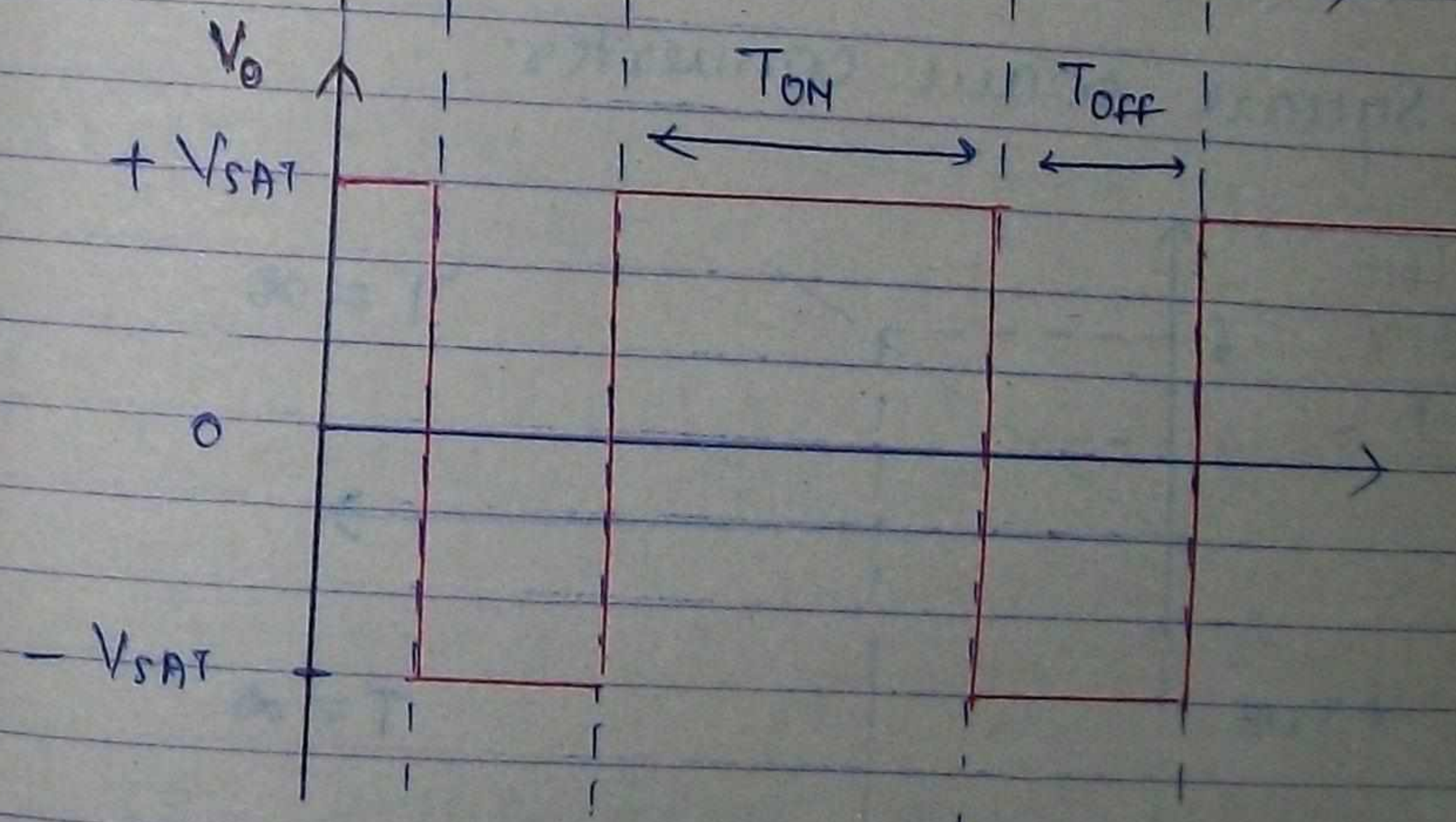
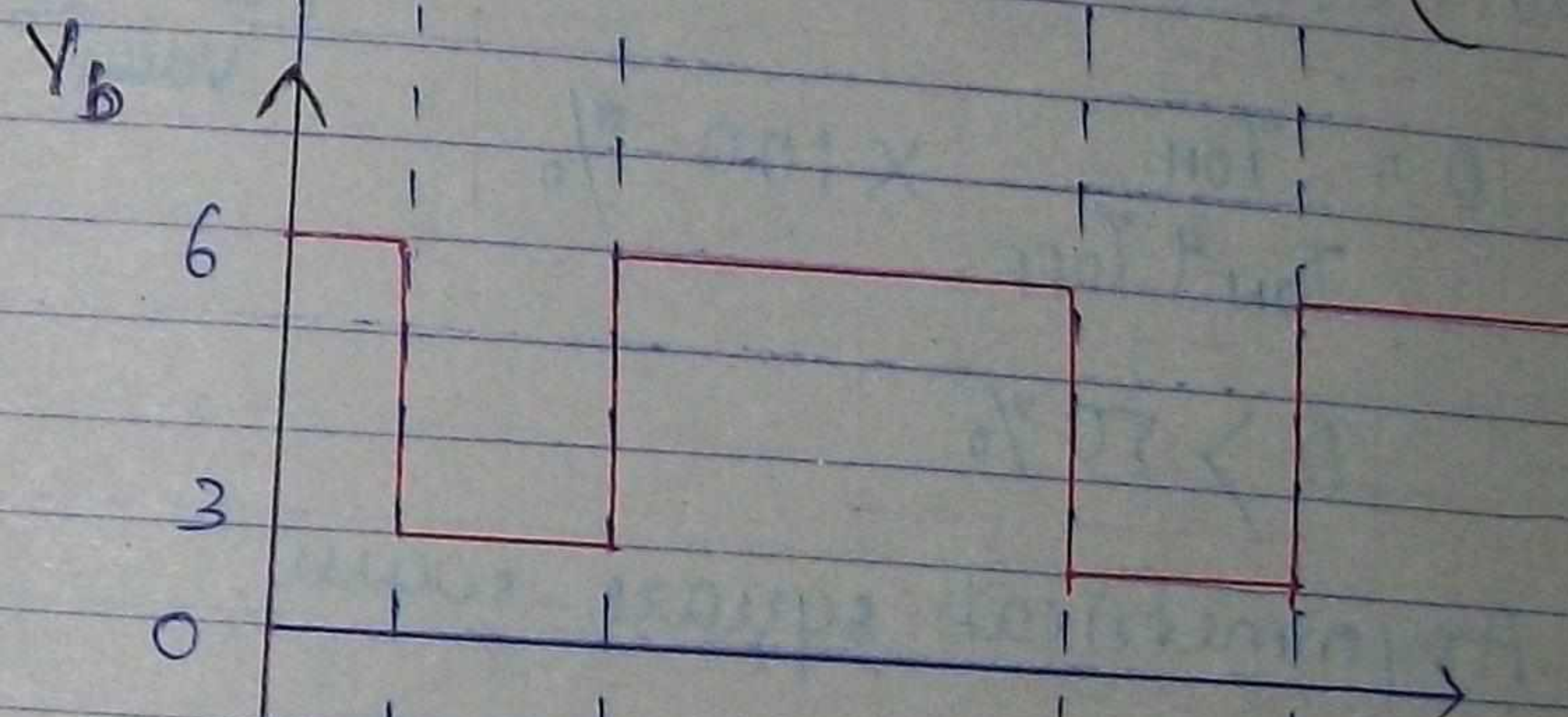
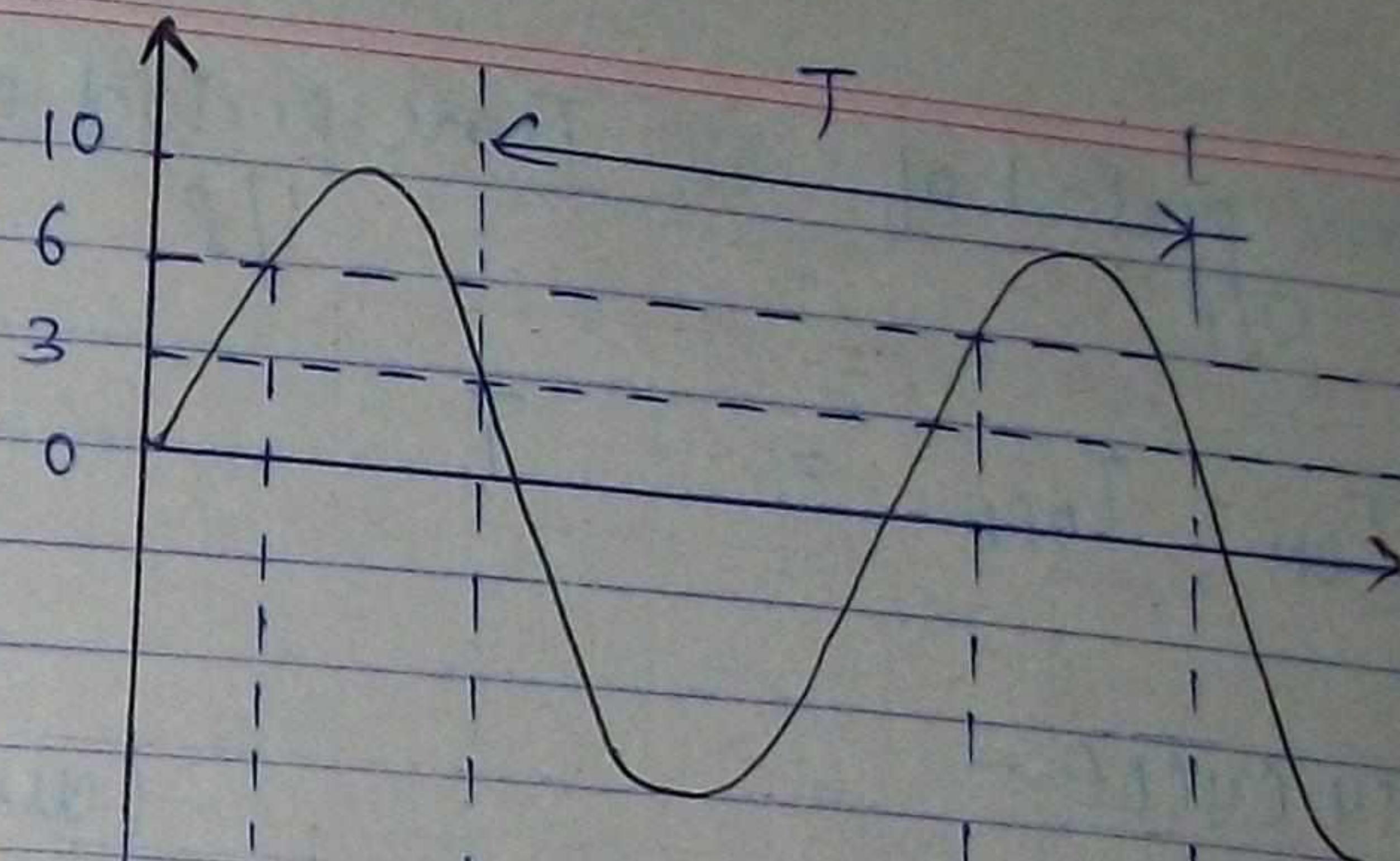
$$V_b \rightarrow 6 \text{ to } 3$$

$$\textcircled{3} \quad V_i^\circ \downarrow \text{ when } V_i^\circ = V_a \leq V_b$$

o/p switch from $-V_{SAT}$ to $+V_{SAT}$

$$V_b \rightarrow 3 \text{ to } 6.$$

$$V_a = V_i^\circ$$



$$\text{Time PERIOD OF I/P} = \text{Time PERIOD OF O/P}$$

- Time period of o/p = Time period of i/p

$$T_{ON} + T_{OFF} = T$$

- DUTY CYCLE -

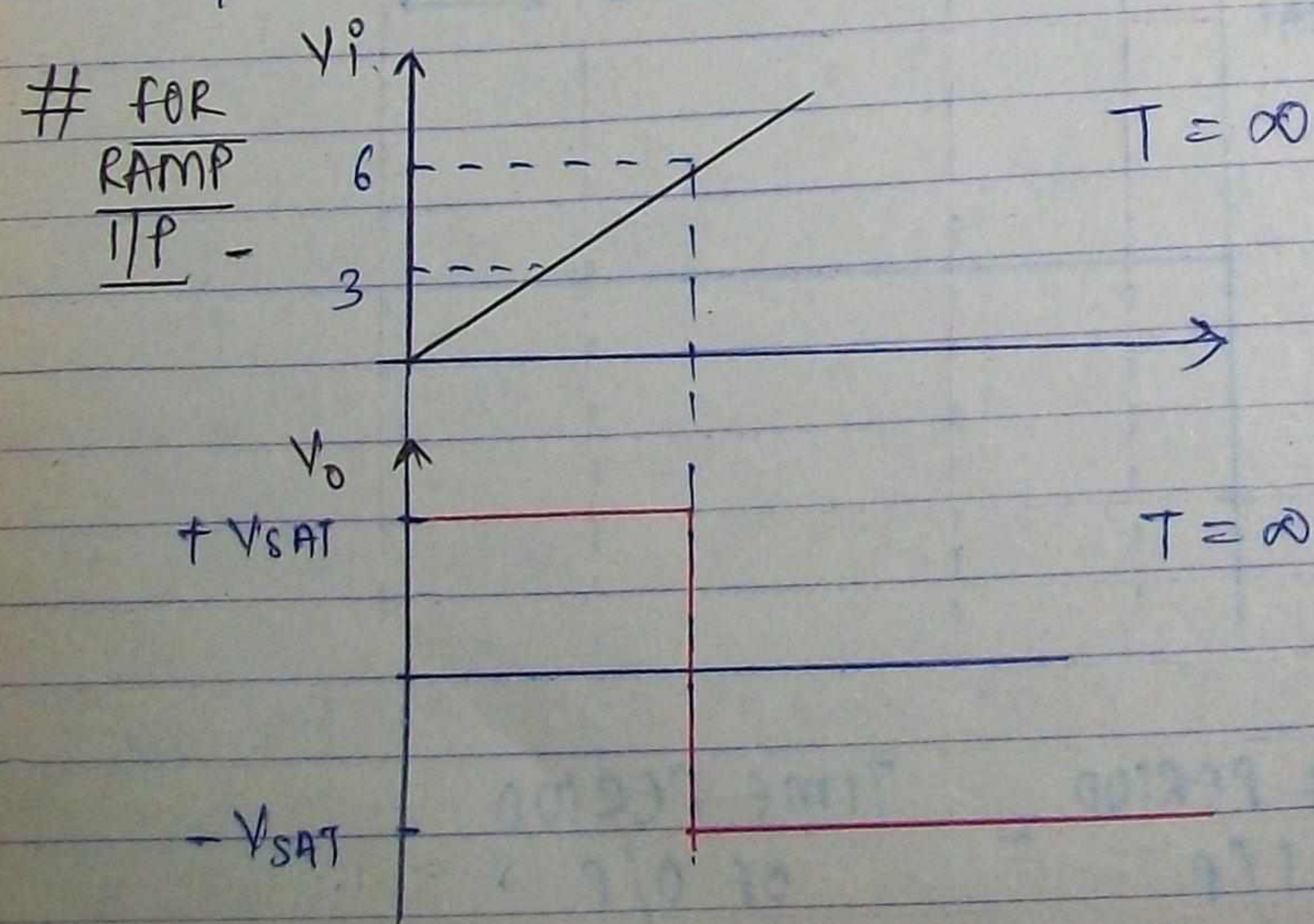
$$D = \frac{T_{ON}}{T_{ON} + T_{OFF}} \times 100\%$$

(average value = +ve)

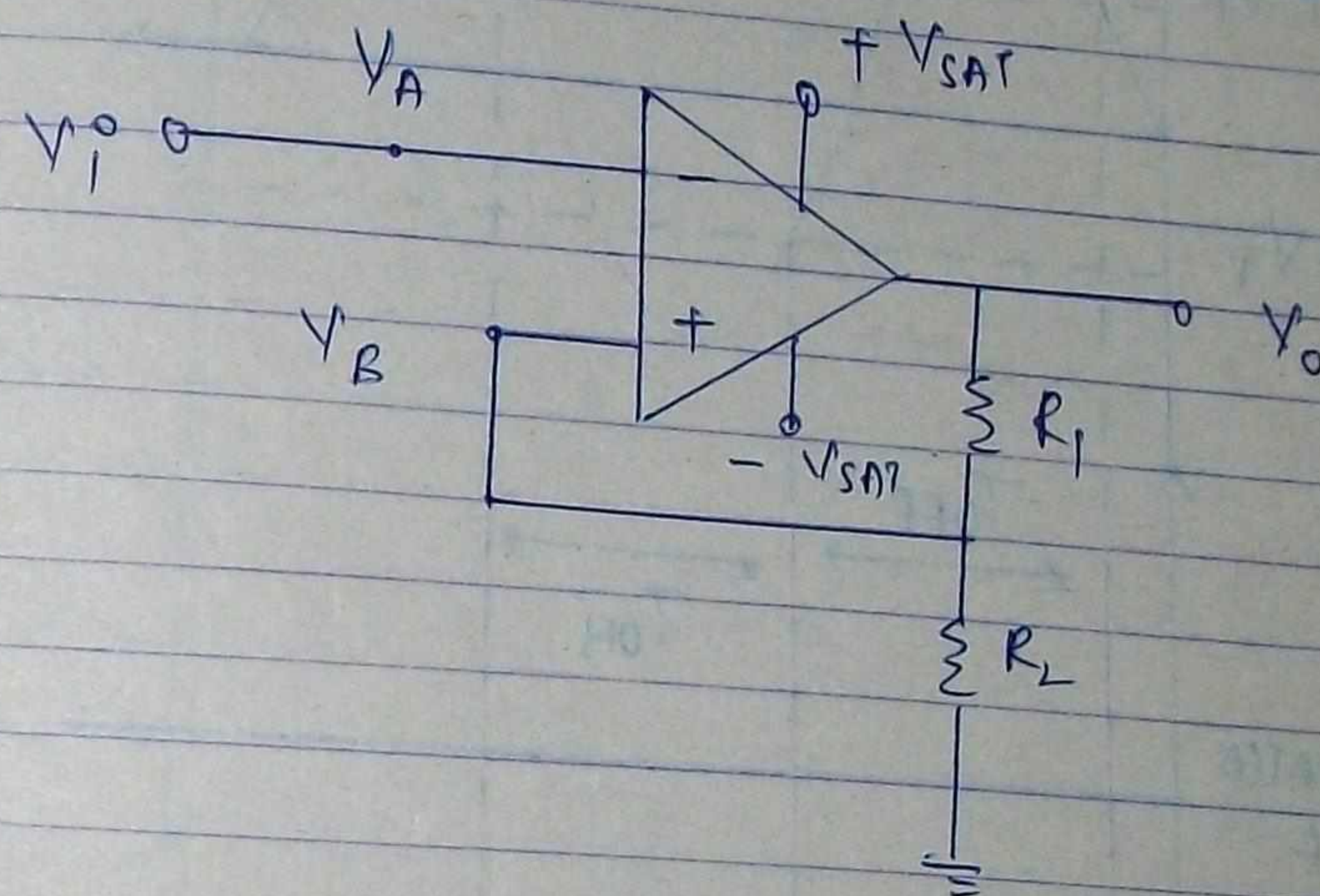
$$D > 50\%$$

Asymmetrical square wave

- Square wave converter.



FOR SYMMETRICAL SQUARE WAVE -



$$V_{b1} = \frac{V_{SAT} \cdot R_2}{R_1 + R_2} = +V_T$$

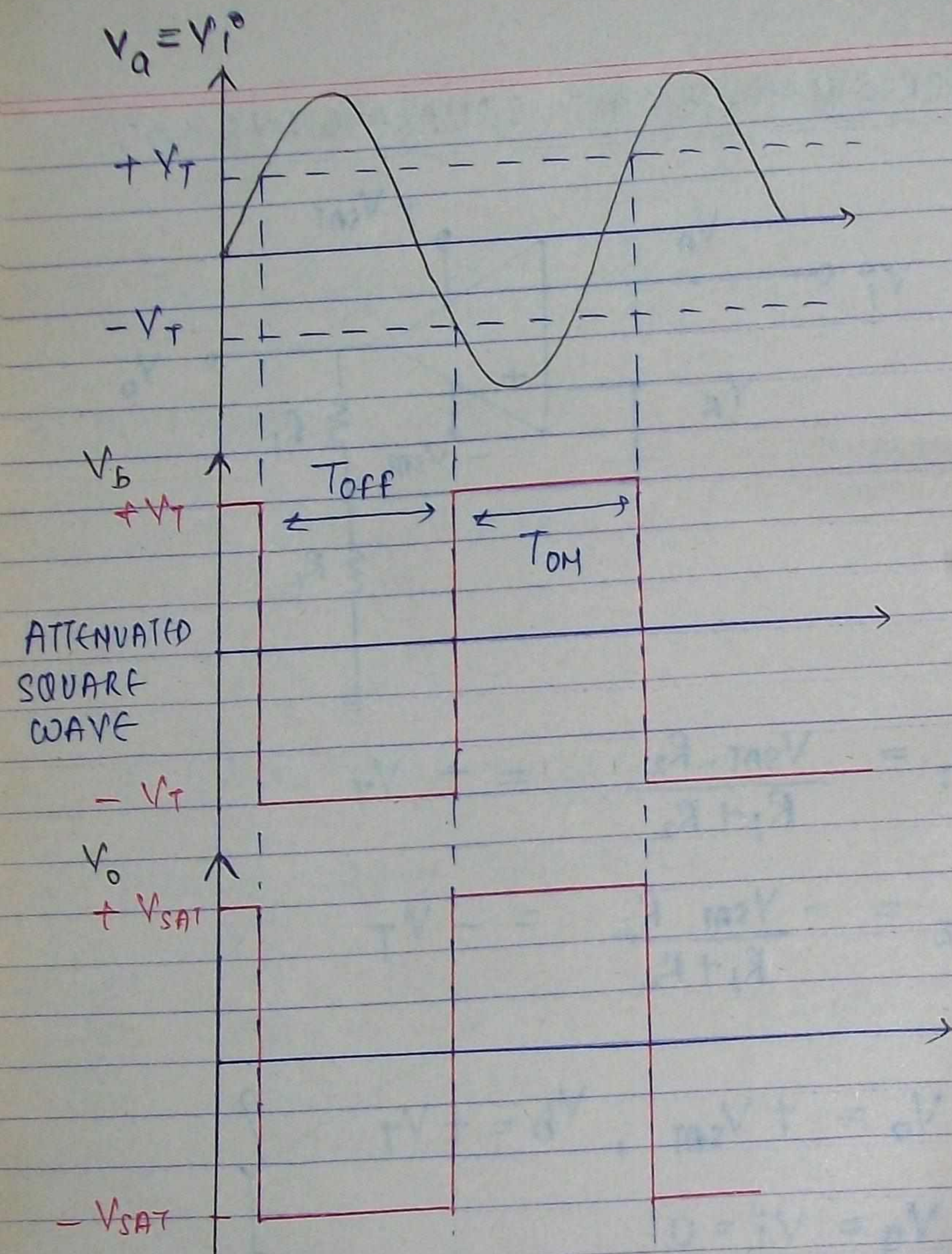
$$V_{b2} = -\frac{V_{SAT} \cdot R_2}{R_1 + R_2} = -V_T$$

$$\left. \begin{aligned} V_o &= +V_{SAT}, V_b = +V_T \\ V_a &= V_i = 0 \end{aligned} \right\}$$

* for symmetrical square wave

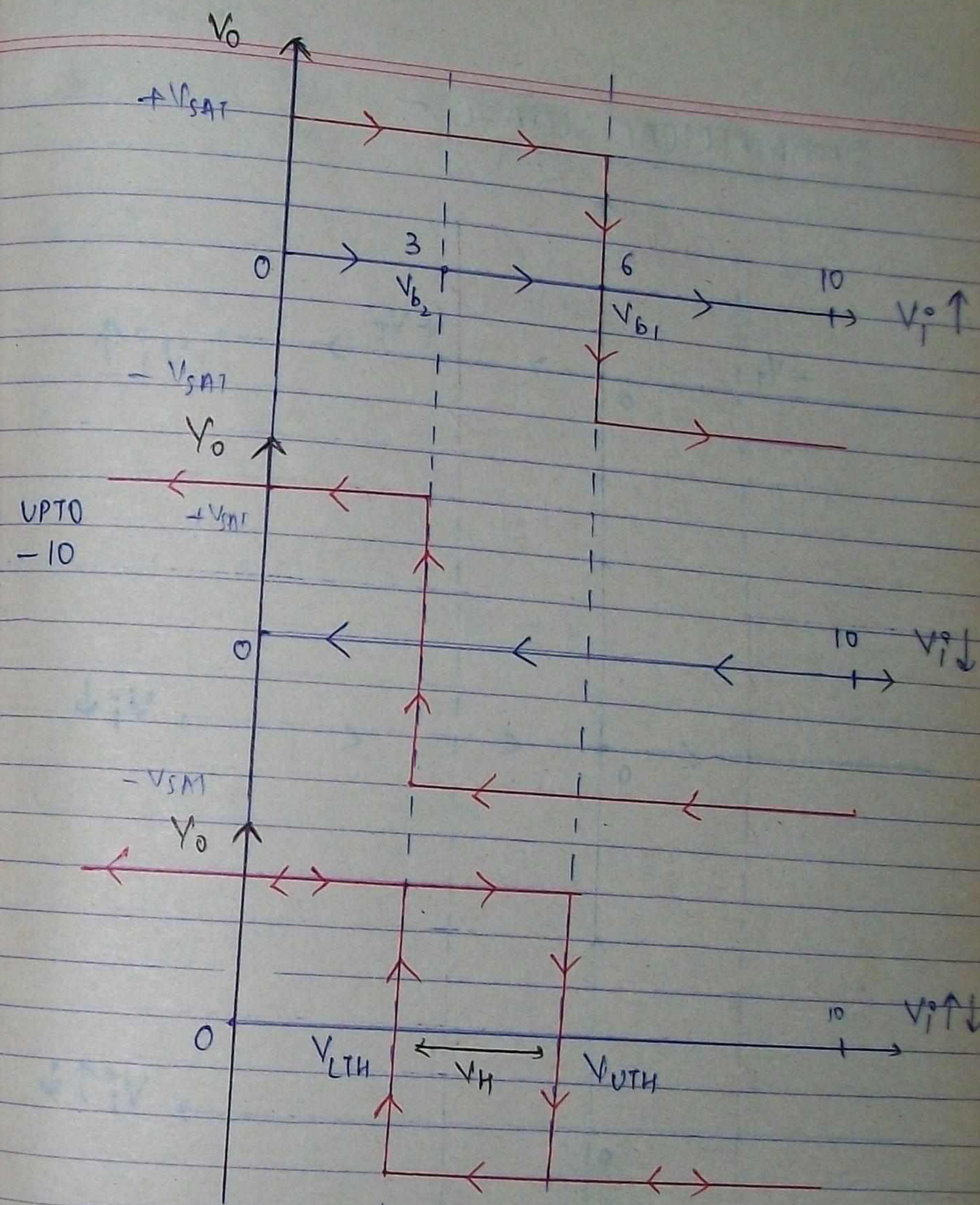
Duty cycle = 50%

$$\left. \begin{aligned} V_T &= \alpha V_{SAT} \\ \alpha &= \frac{R_2}{R_1 + R_2} < 1 \end{aligned} \right\} \text{ (ATTN)}$$



⇒ HYSTERISIS LOOP -

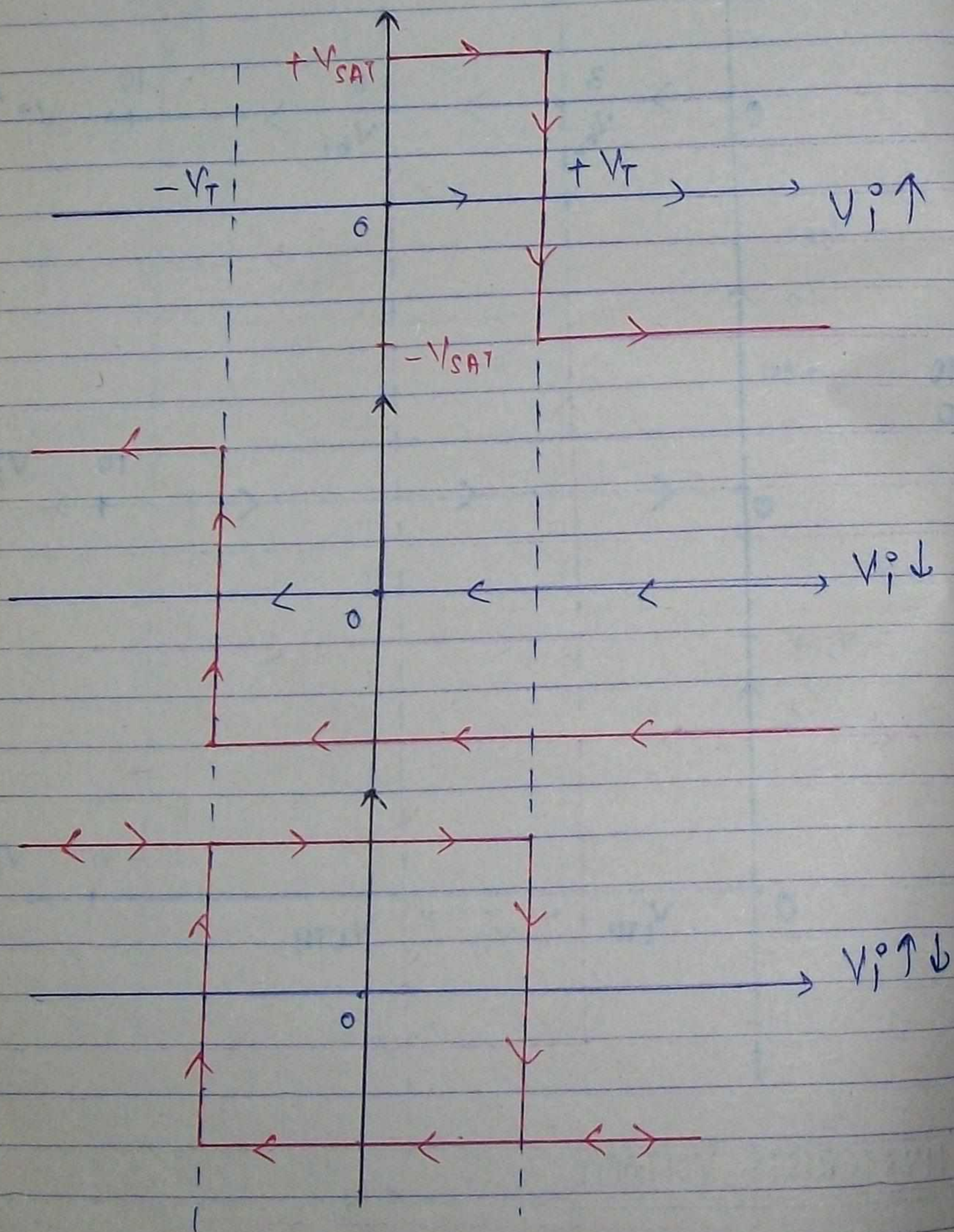
① ASYMMETRICAL WAVE -



HYSTERISIS VOLTAGE -

$$V_H = V_{UTH} - V_{LTH}$$

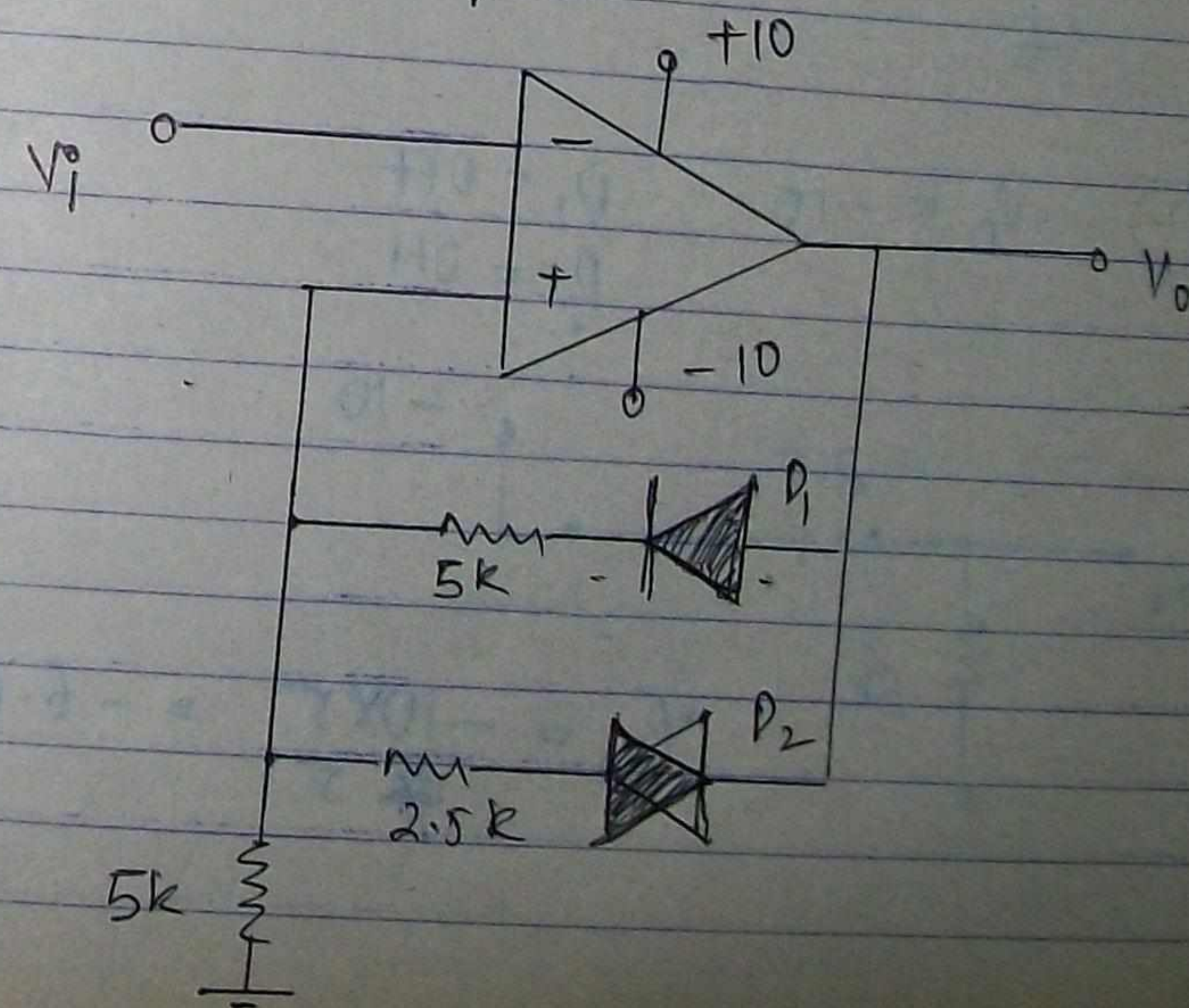
② SYMMETRICAL WAVE -

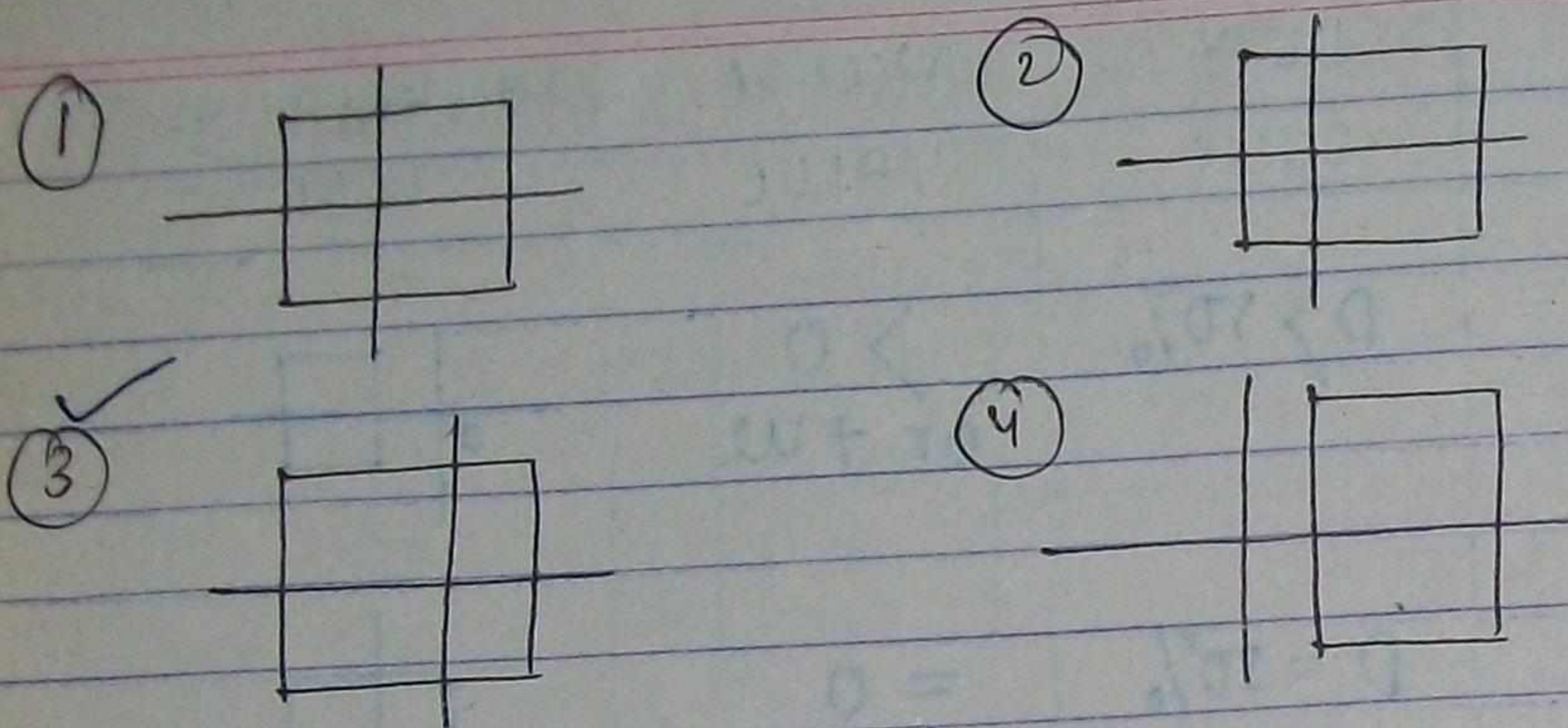


① CONCLUSION -

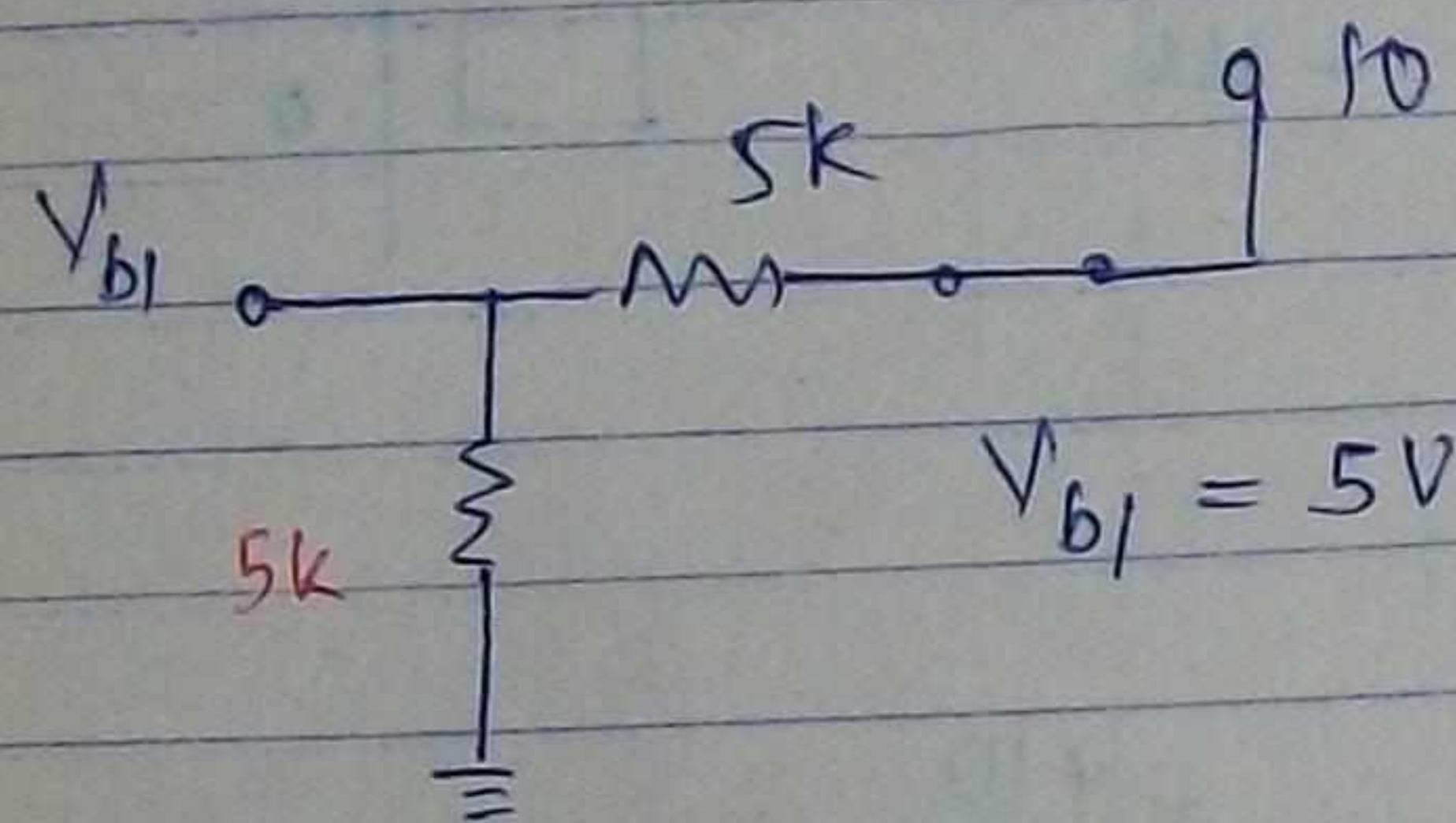
V_R	DUTY CYCLE	AVERAGE VALUE	HYSTERSIS LOOP
$V_R > 0$	$D > 50\%$	> 0 or $+ve$	
$V_R = 0$	$D = 50\%$	$= 0$	
$V_R < 0$	$D < 50\%$	< 0 or $-ve$	

Que. Hysteresis loop?

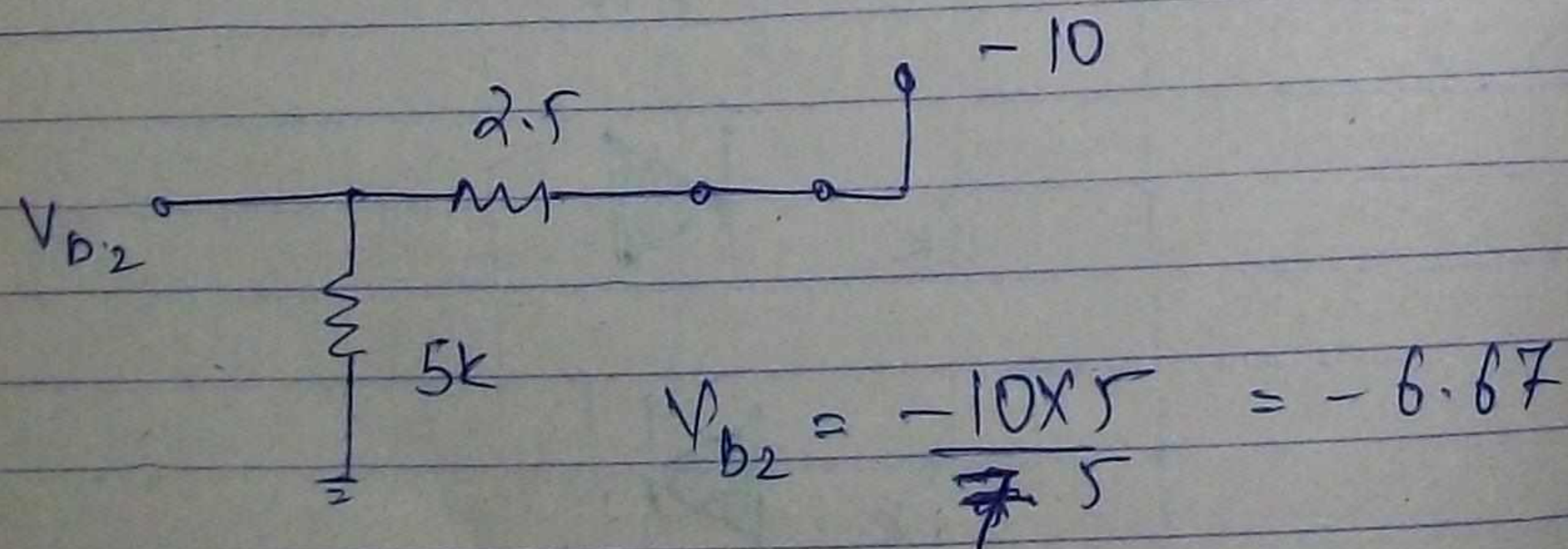




Solⁿ: ① $V_o = +10$ $D_1 - ON$
 $D_2 - OFF$



② $V_o = -10$ $D_1 - OFF$
 $D_2 = ON$



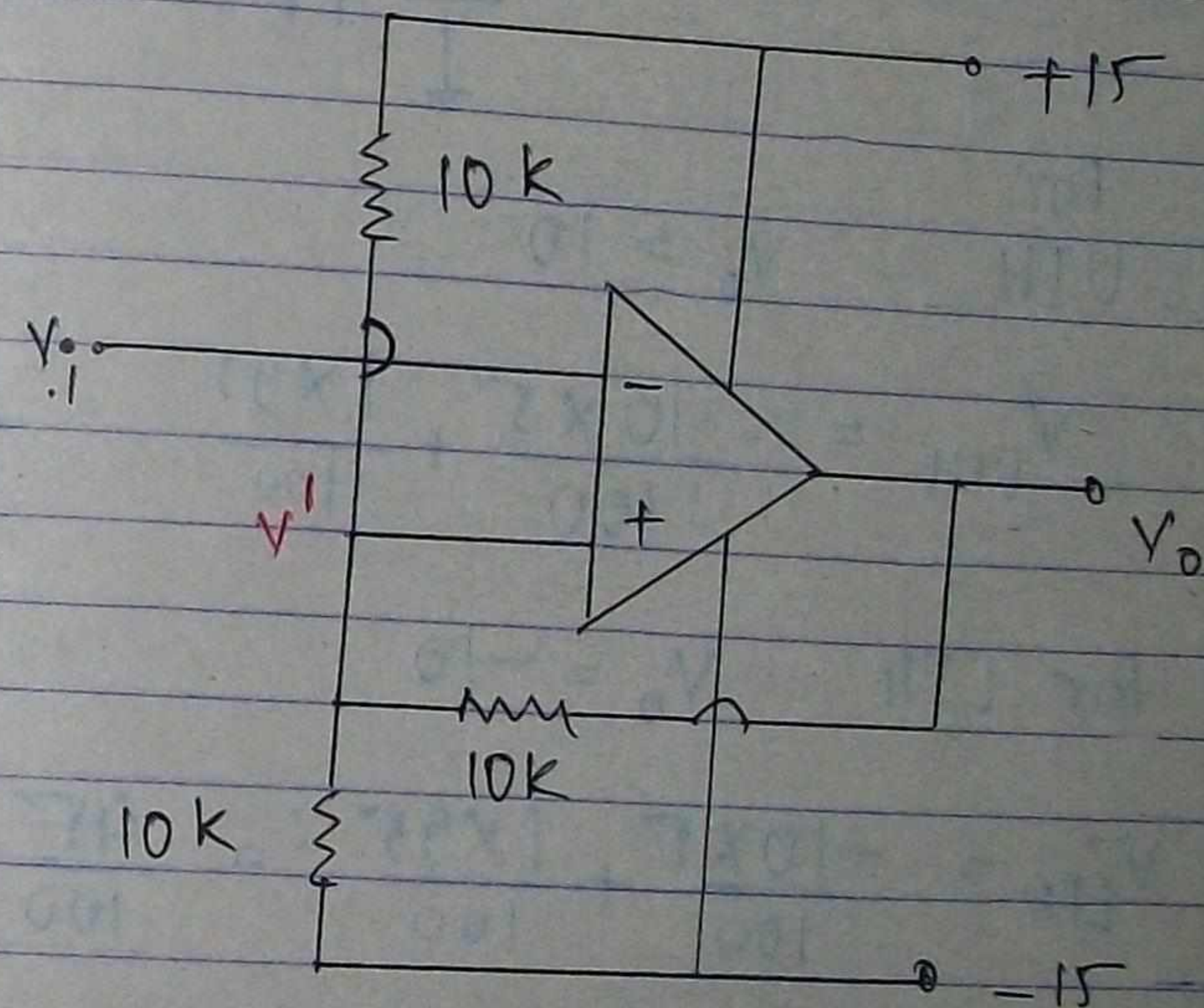
Ques. Consider Schmitt trigger ckt shown below a triangular wave which goes from -12 to +12V is applied to inverting I/P of OPAMP, assume that O/P of OPAMP swings from +15 to -15V the voltage at non-inverting I/P switches b/w

(A) -12 to +12

(B) -7.5 to +7.5

✓ (C) -5 to +5

(D) 0 to 5.



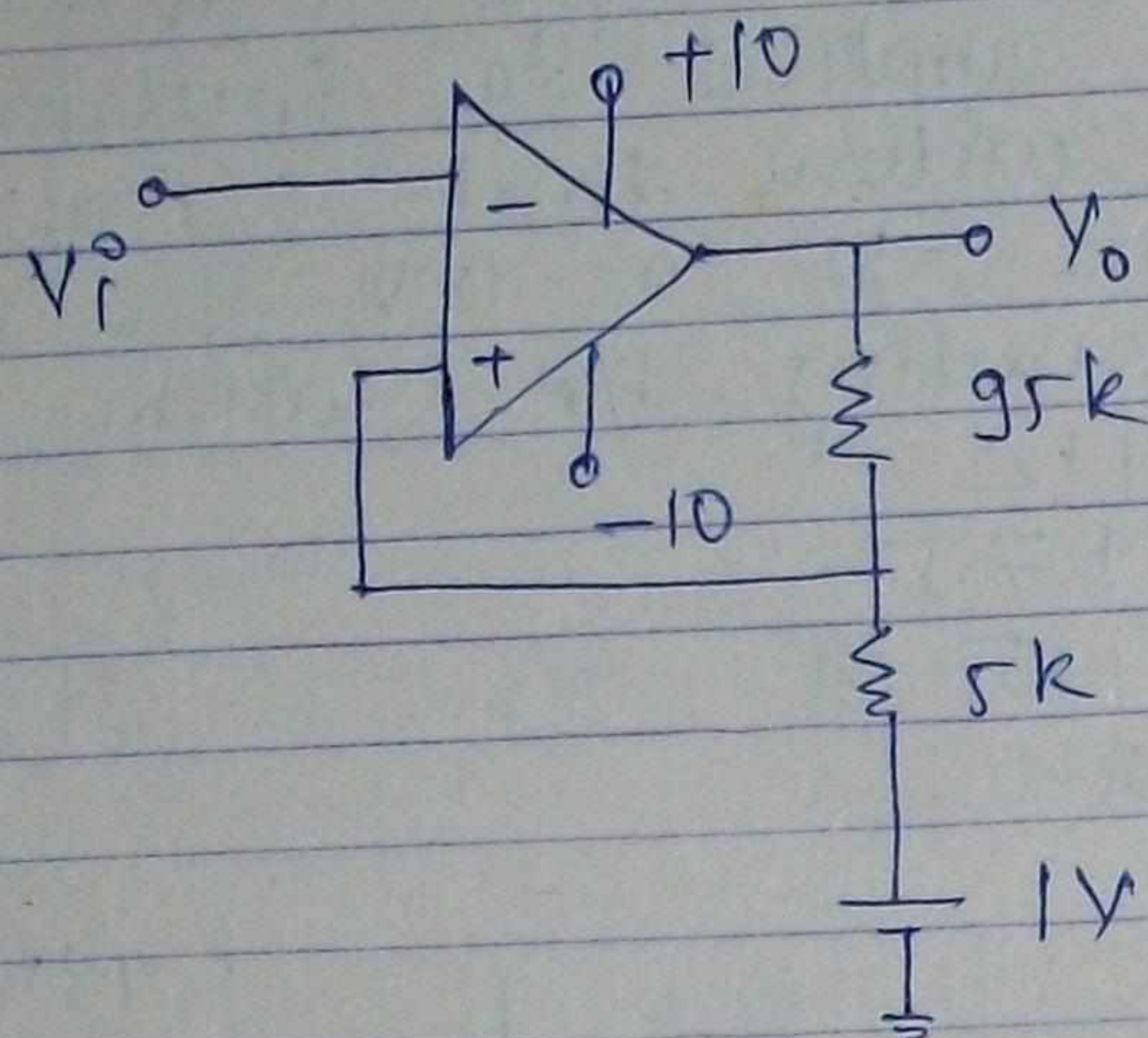
Solⁿ: $V' \left[\frac{1}{10} + \frac{1}{10} + \frac{1}{10} \right] = \frac{15}{10} - \frac{15}{10} + \frac{V_o}{10}$

$V' \times 3 = V_o$, $V' = V_o/3$

$V' = -5 \text{ to } +5$

Q. WORKBOOK P-60 Q3.

Solⁿ:



① for UTH $V_o = 10$

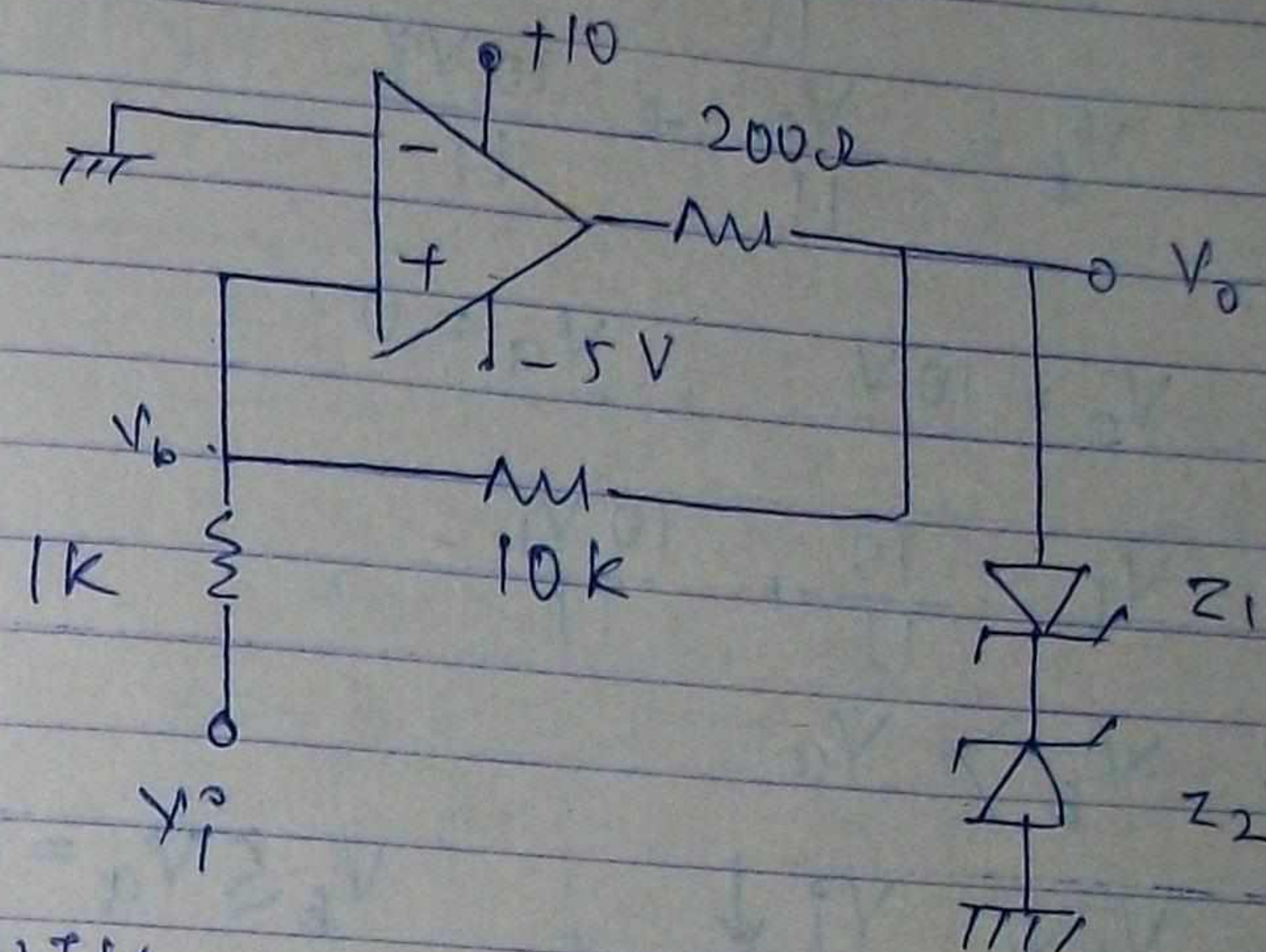
$$V_{UTH} = \frac{10 \times 5}{100} + \frac{1 \times 95}{100} = \frac{145}{100}$$

② for LTH $V_o = -10$

$$V_{LTH} = \frac{-10 \times 5}{100} + \frac{1 \times 95}{100} = \frac{45}{100}$$

$$V_H = V_{UTH} - V_{LTH} = \frac{145 - 45}{100} = 1$$

Q. WORKBOOK P-60 Q5.



① for UTH

$V_o = +10$, $V_a = 0$

$$V_{UTH} = \frac{10 \times 1}{11} + \frac{V_i \times 10k}{11} \quad V_b > V_a$$

② for VTH $V_o = -5$

$V_a = 0$

$$V_{VTH} = \frac{-5 \times 1}{11} + \frac{V_i \times 10k}{11} \quad V_b \leq V_a = 0$$

$$= 0 \quad \text{if } V_i = -1$$

Solⁿ:

$$V_b = -\frac{V_o \times 1}{11} + \frac{V_i \times 10}{11}$$

$$V_b = \frac{V_o}{11} + \frac{10 V_i}{11}$$

① $V_o = 10V$, $V_a = 0$

$$V_b = \frac{10}{11} + \frac{10 V_i}{11}$$

$$V_b > V_a$$

$$V_b \downarrow V_i \downarrow, V_b \leq V_a = 0$$

o/p switch from 10 to -5.

$$\frac{10}{11} + \frac{10 V_i}{11} = 0$$

$$V_i = -1$$

② $V_o = -5$, $V_a = 0$

$$V_b = \frac{-5}{11} + \frac{10 V_i}{11}$$

$$V_b < V_a$$

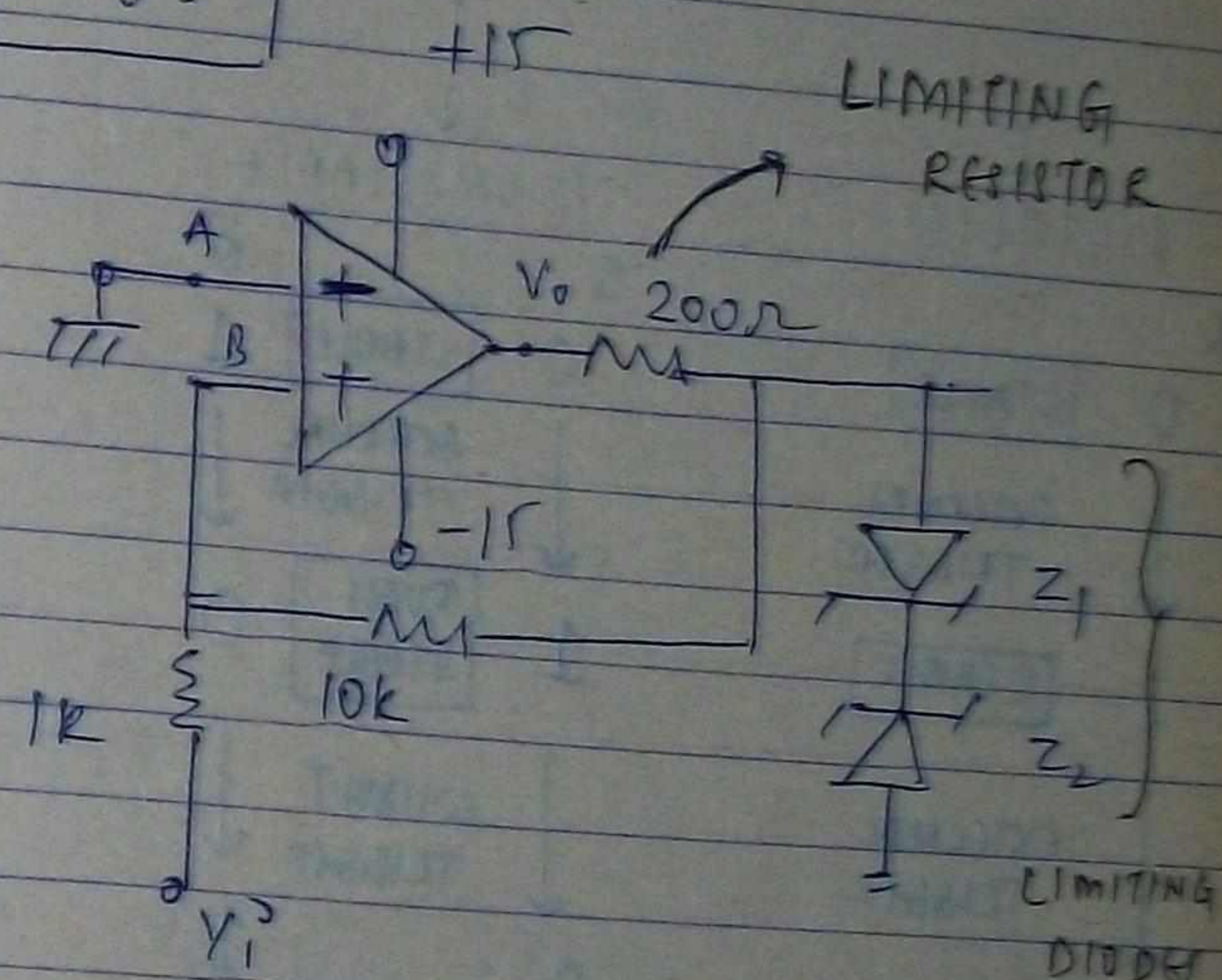
$$V_b \uparrow V_i \uparrow, \text{ when } V_b \geq V_a = 0$$

o/p switch from -5 to 10

$$-\frac{5}{11} + \frac{10 V_i}{11} = 0$$

$$V_i = 0.5$$

let

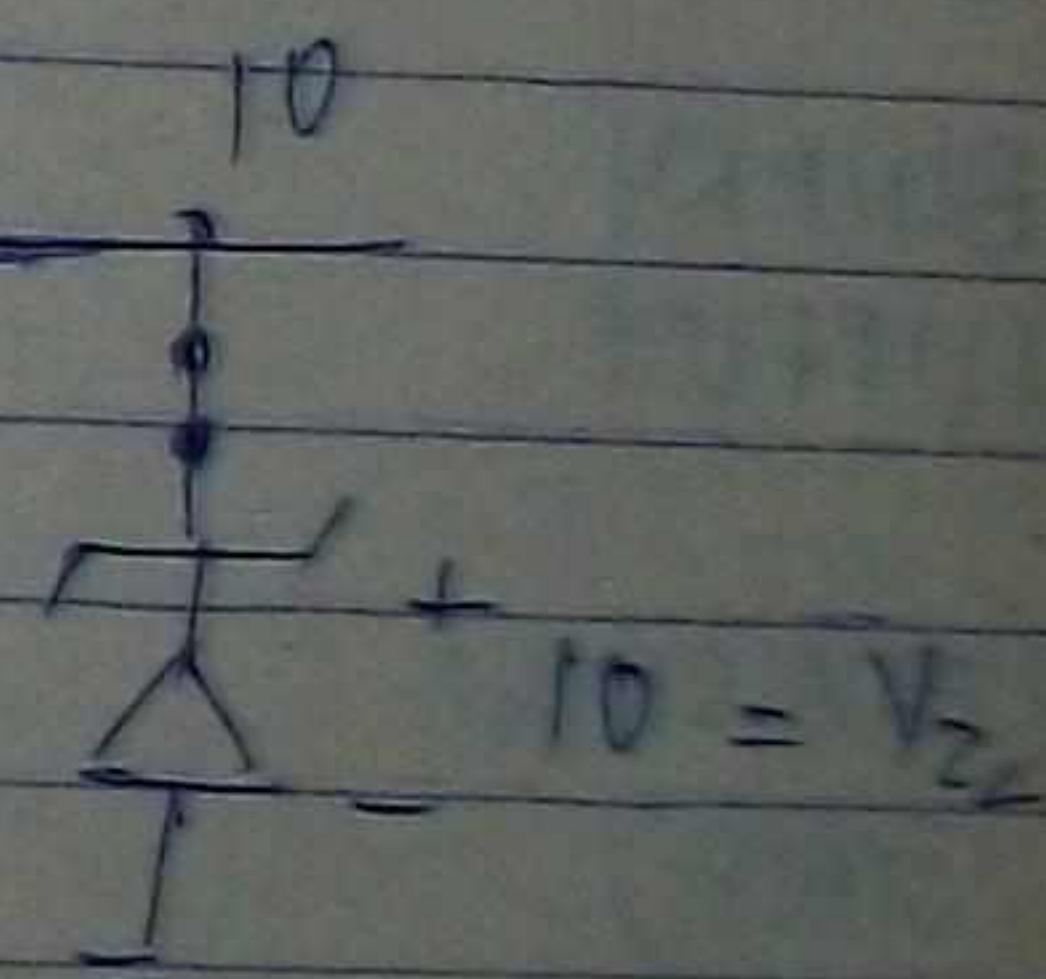


$$V_b > V_a$$

$$V_o = +15$$

$$+15$$

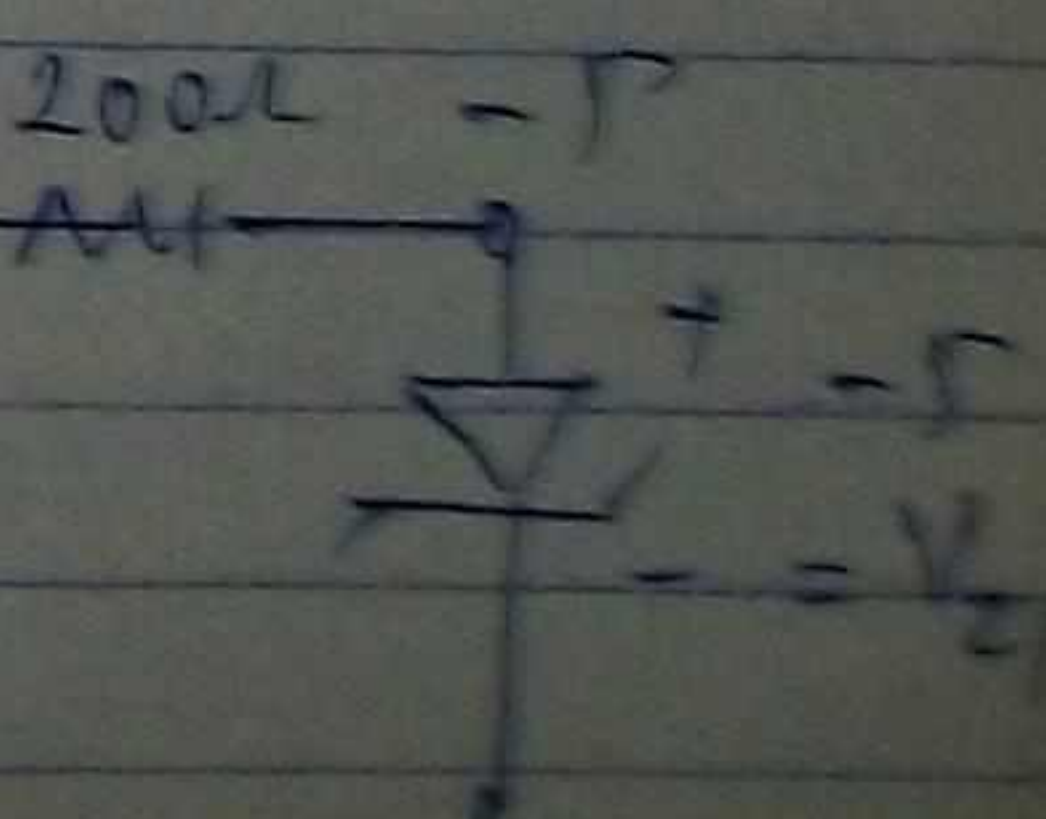
$$200$$



$$V_b < V_a$$

$$V_o = -15$$

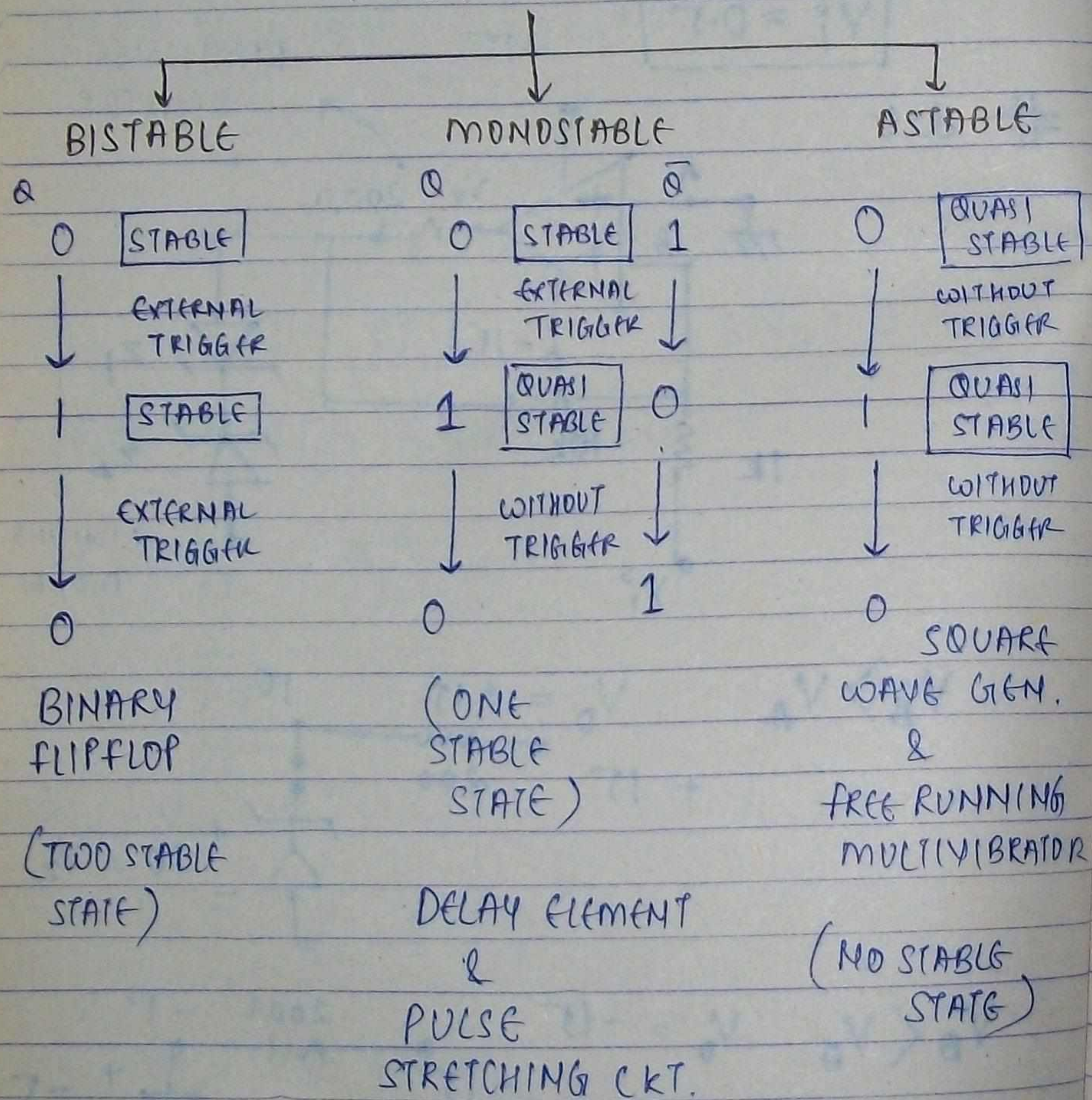
$$-15$$



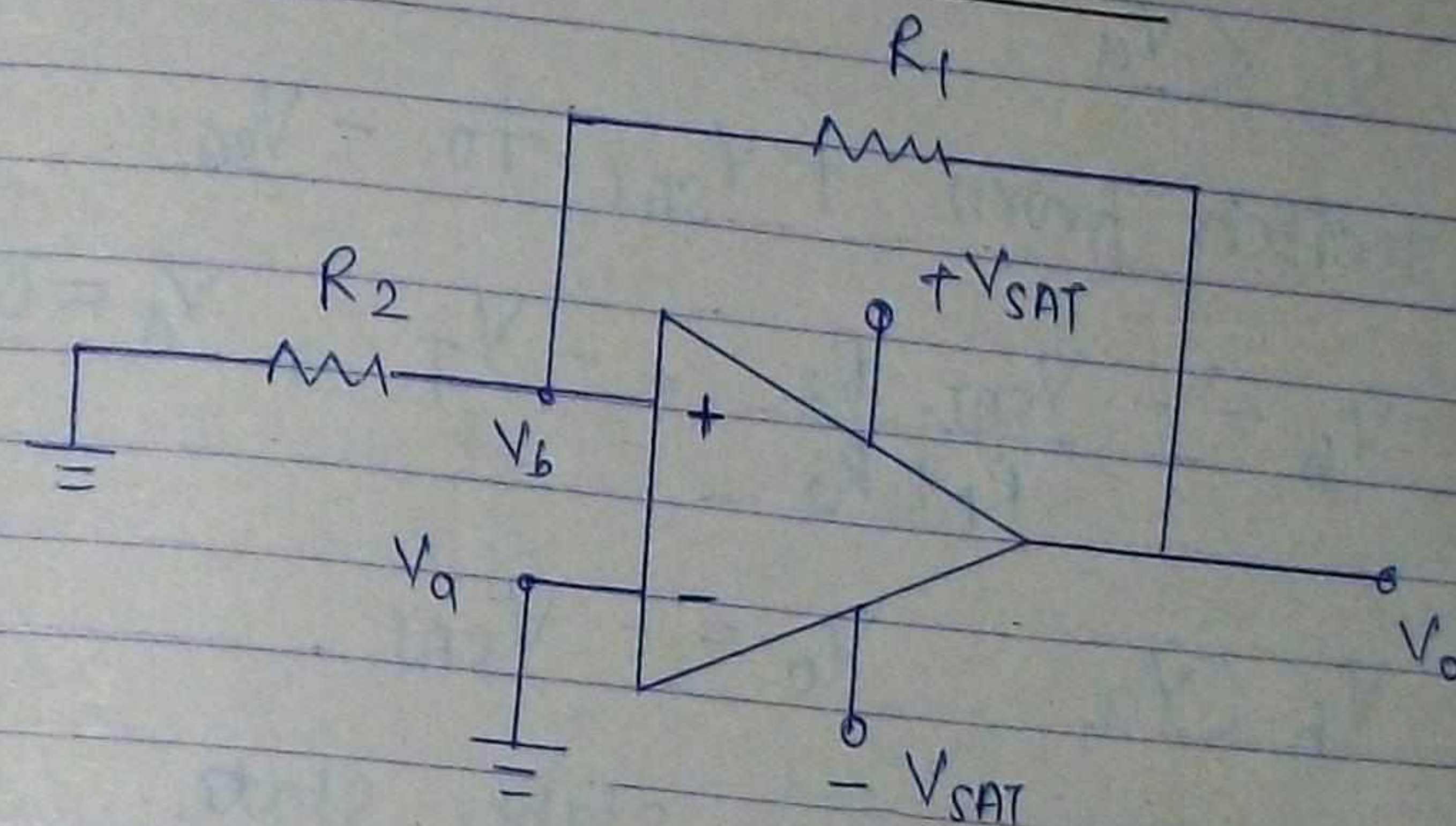
MULTI VIBRATORS

- Multivibrator is a device whose o/p vibrates between two level, low level and high level i.e 0 and 1.

MULTIVIBRATOR



⇒ MULTIVIBRATOR BY USING OPAMP -
① BISTABLE MULTIVIBRATOR -



$$\textcircled{1.} \quad V_b = \frac{R_2}{R_1 + R_2} V_0$$

$$V_o = 0, V_b = 0, V_g = 0$$

UNBALANCED EQUILIBRIUM

② Because of noise $V_b \uparrow$, $V_o = +V_{sat}$

$$V_b = + \frac{V_{SAT} R_2}{R_1 + R_2} = V_T =$$

$$V_a \doteq 0 \quad V_b > V_a \quad V_o \pm + V_{SAT}$$

ckt is in stable state

③ Negative trigger $\square$ at b or +ve trigger at a $\square$

$$V_b < V_a$$

V_o switch from $+V_{SAT}$ to $-V_{SAT}$

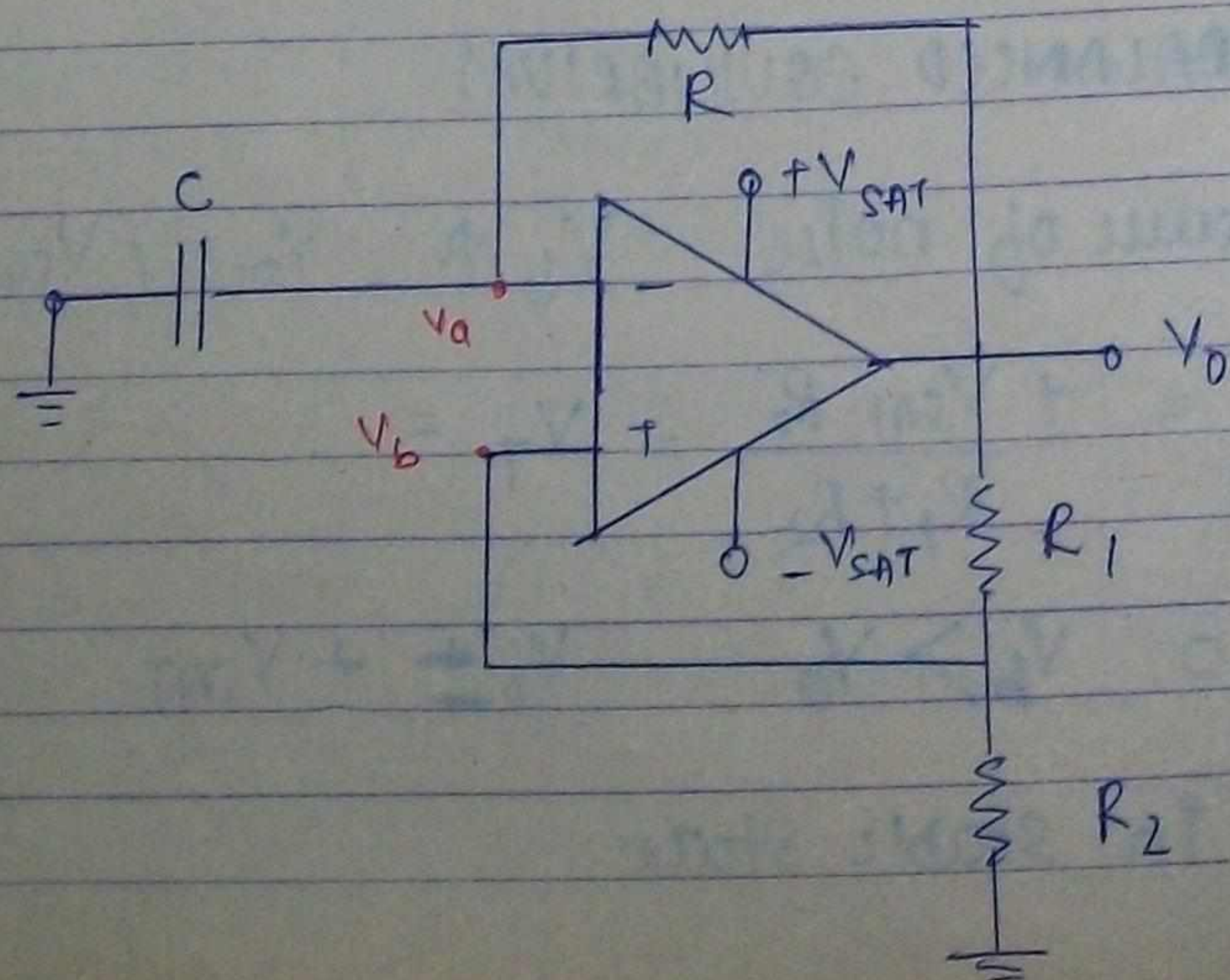
$$V_b = -\frac{V_{SAT} R_2}{R_1 + R_2} = -V_T, \quad V_a = 0$$

$$V_b < V_a \quad V_o = -V_{SAT}$$

circuit is in another stable state.

Acts as a volatile memory. (RAM).

② ASTABLE MULTIVIBRATOR - (SQUARE WAVE GENERATOR)



$$V_b = \frac{V_o R_2}{R_1 + R_2}$$

$$V_c(0^-) = V_c(0^+) = 0$$

$$\therefore V_a = 0$$

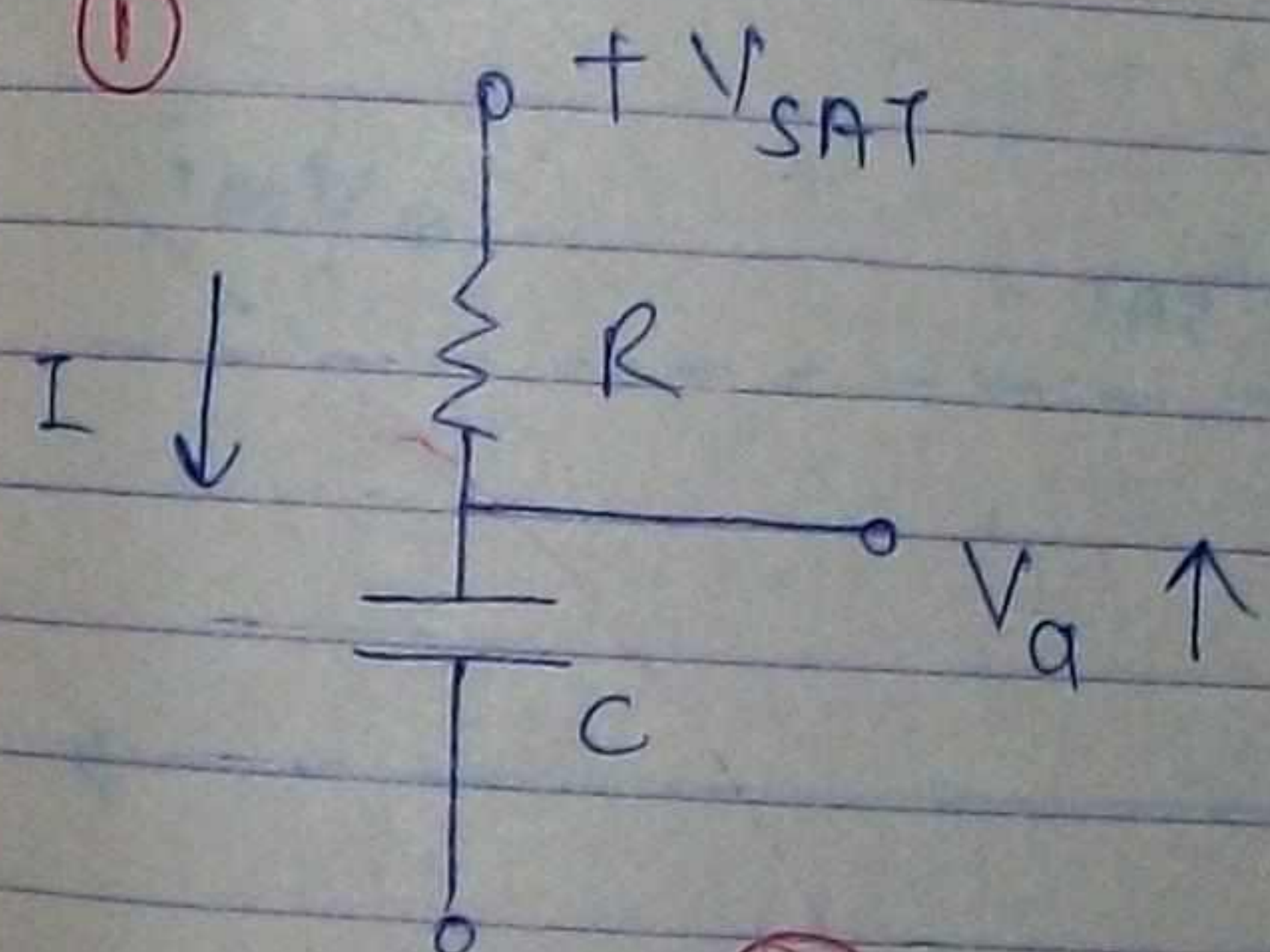
$$\text{Let } V_o = +V_{SAT}$$

$$V_b = \frac{V_{SAT} R_2}{R_1 + R_2} = +V_T$$

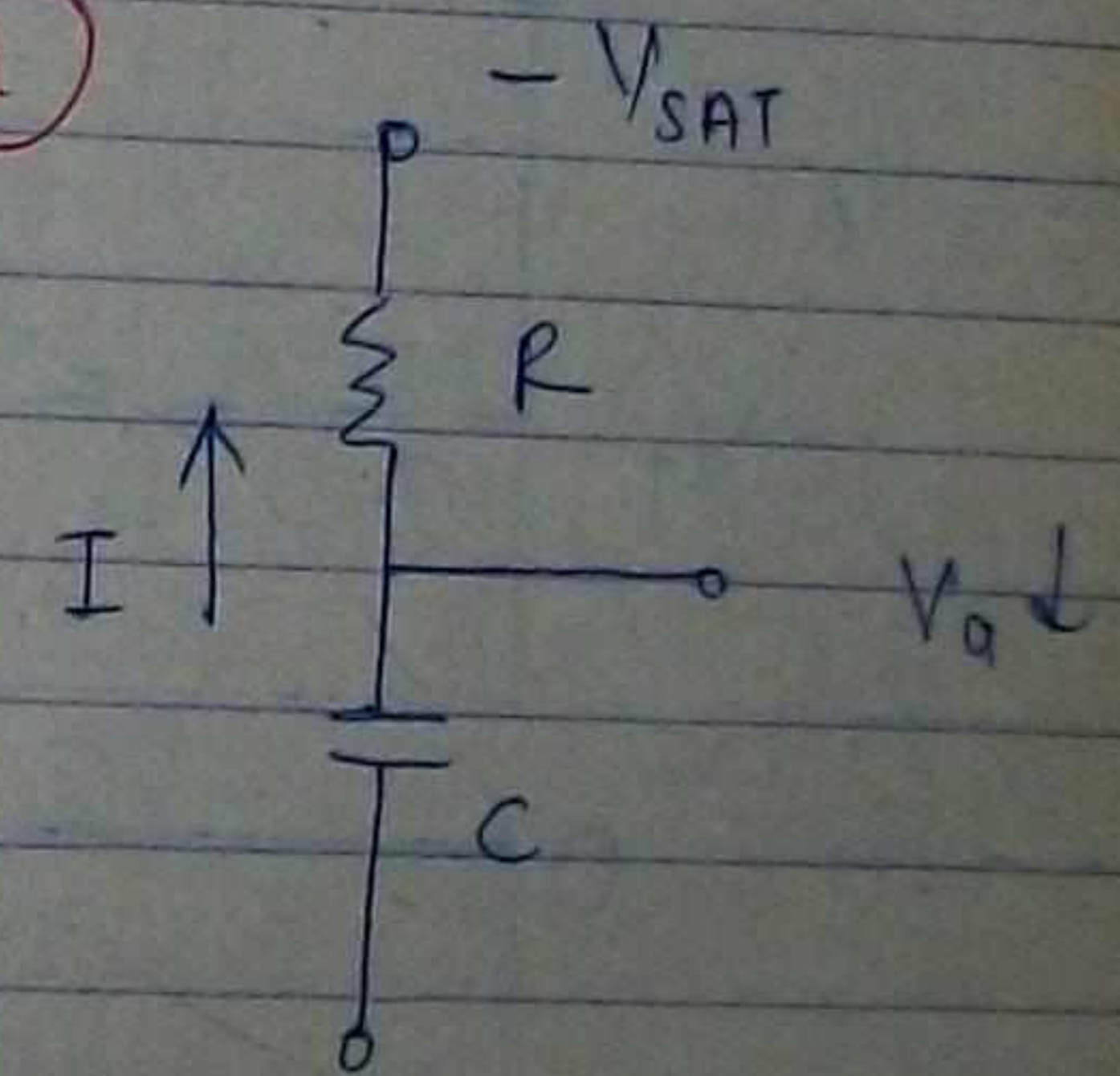
$$V_b > V_a \quad \therefore V_o = +V_{SAT}$$

$\therefore$ Our assumption is right.

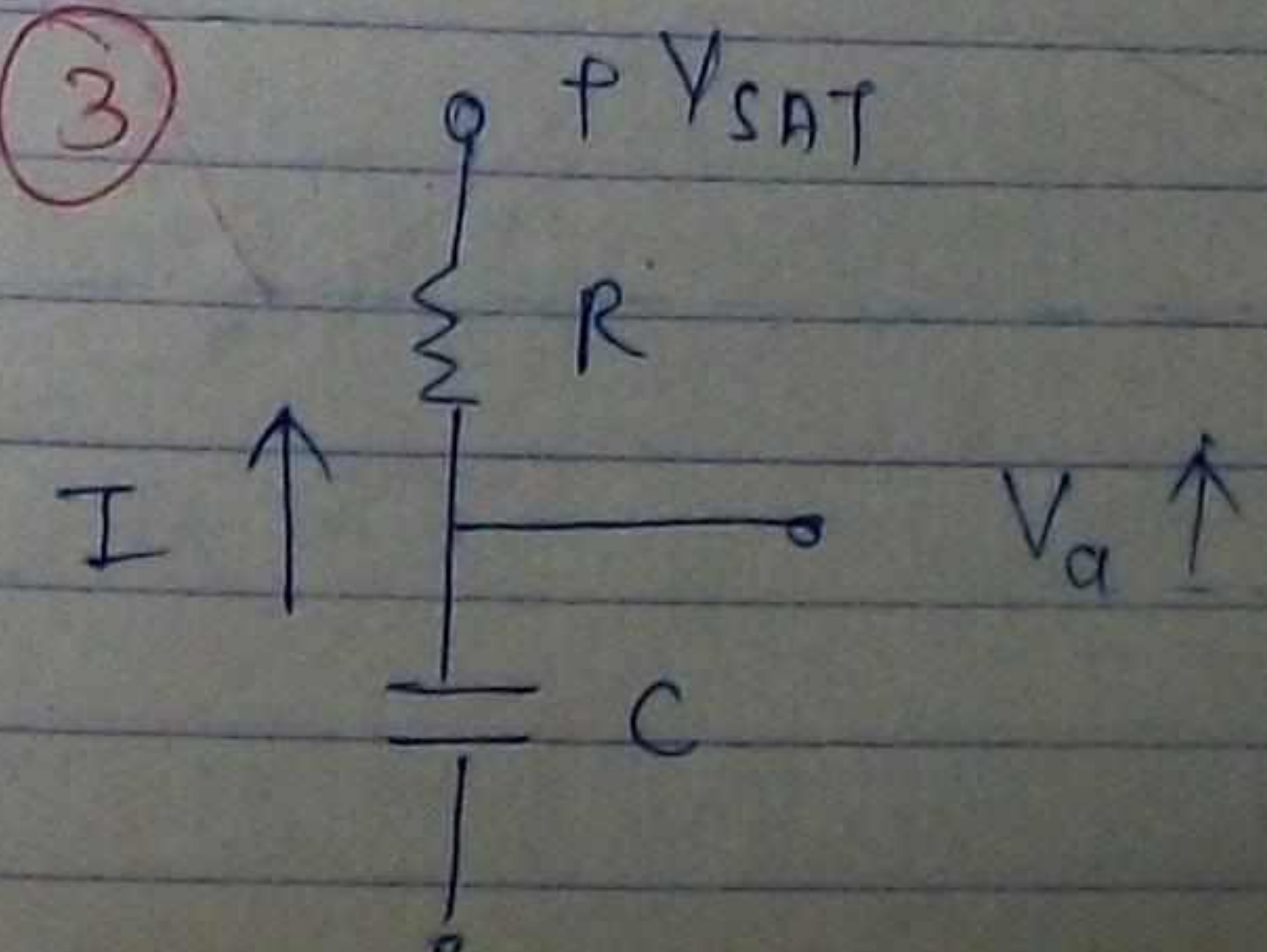
①

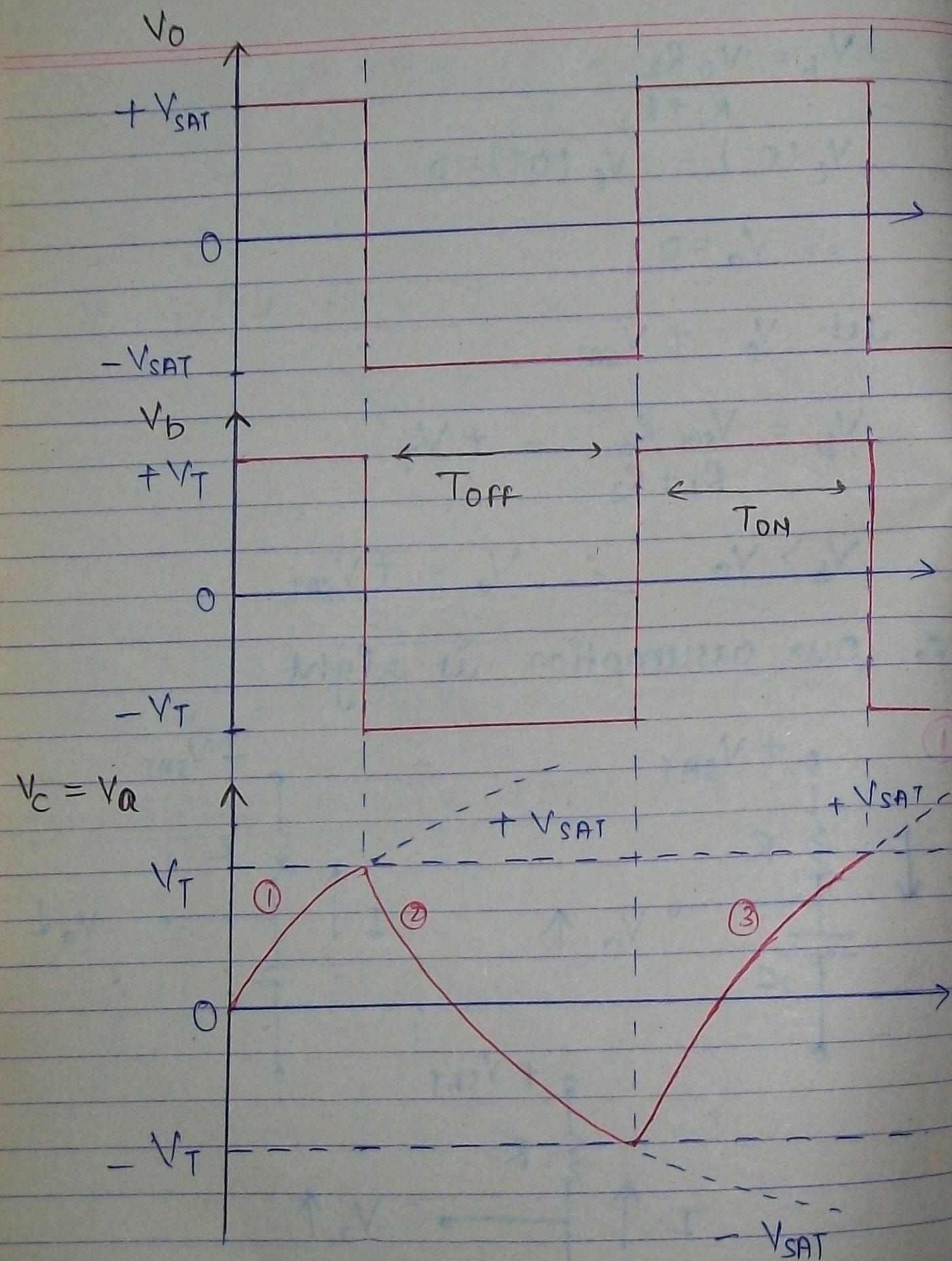


②



③





- At initial condition consider $V_o = +V_{SAT}$ and potential at point 'a' $V_a = V_c = 0$. Then $V_b = +V_T$. Now capacitor will charge towards $+V_{SAT}$ with $\tau = RC$. Capacitor charge upto $+V_T$ at this point $V_a = V_b = +V_T$ and operational amp comes out of saturation when cap. further charge above $+V_T$ then $V_a > V_b$ as a result of which O/P voltage V_o switch over to negative saturation level $-V_{SAT}$. Therefore $V_b = -V_T$.

- Now capacitor start discharging from $+V_T$ to $-V_T$ towards $-V_{SAT}$ with $\tau = RC$ thus when cap. discharge upto $-V_T$ then $V_a = V_b = -V_T$ and OPAMP comes out of saturation when cap. further discharge below $-V_T$ then $V_a < V_b$ as a result of which V_o switch over to $+V_{SAT}$ and again $V_b = +V_T$.

Thus cycle repeats..

$T_{ON} \propto RC$, $T_{OFF} \propto RC$

thus $T \propto RC$

$$T \propto V_T$$

$$V_T = \frac{R_2}{R_1 + R_2} V_{SAT}$$

$$T \propto \frac{R_2}{R_1 + R_2}$$

$$T \propto R_2, \quad T \propto \frac{1}{R_1}$$

DERIVATION OF T_{ON} -

C $\rightarrow$ charge from $-V_T$ to $+V_T$

$$\tau = RC$$

$$V_c = V_f - [V_f - V_i^0] e^{-t/\tau}$$

$$V_i^0 = -V_T \quad (t=0)$$

$$V_f = +V_{SAT} \quad (t=\infty)$$

$$V_c = V_{SAT} - [V_{SAT} + V_T] e^{-t/RC}$$

$$\text{at } t = T_{ON} \quad V_c = +V_T$$

$$V_T = V_{SAT} - [V_{SAT} + V_T] e^{-T_{ON}/RC}$$

$$T_{ON} = RC \ln \left(\frac{V_{SAT} + V_T}{V_{SAT} - V_T} \right)$$

DERIVATION OF T_{OFF} -

C $\rightarrow$ discharge from V_T to $-V_T$

$$\tau = RC \quad V_i^0 = +V_T \quad (t=0)$$

$$V_f = -V_{SAT} \quad (t=\infty)$$

$$V_c = -V_{SAT} + [-V_{SAT} - V_T] e^{-t/RC}$$

$$V_c = -V_{SAT} + [V_{SAT} + V_T] e^{-t/RC}$$

$$\text{at } t = T_{OFF} \quad V_c = -V_T$$

$$-V_T = -V_{SAT} + [V_{SAT} + V_T] e^{-T_{OFF}/RC}$$

$$T_{OFF} = RC \ln \left(\frac{V_{SAT} + V_T}{V_{SAT} - V_T} \right)$$

$$T_{ON} = T_{OFF}$$

$$\therefore D = 50\%$$

Symmetrical square wave.

Time Period -

$$T = T_{ON} + T_{OFF}$$

$$T = 2RC \ln \left(\frac{V_{SAT} + V_T}{V_{SAT} - V_T} \right)$$

$$V_T = \frac{R_2}{R_1 + R_2} V_{SAT}$$

$$T = 2RC \ln \left[1 + \frac{2R_2}{R_1} \right]$$

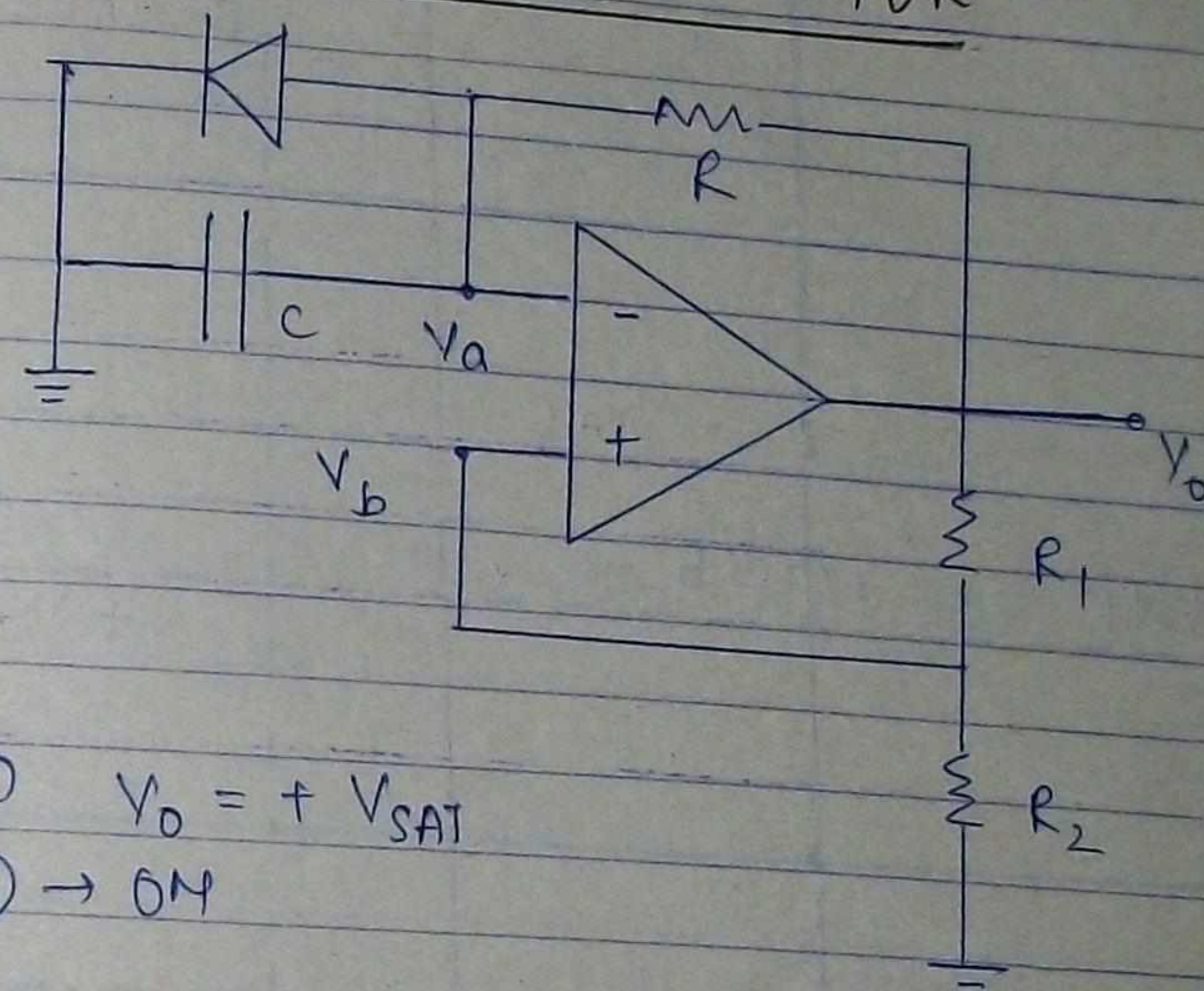
$$f = \frac{1}{T}$$

T = time period of square wave.

f = freq of square wave.

(*) To get asymmetrical waveform V_T is introduced in the circuit.

③ MONO STABLE MULTIVIBRATOR -

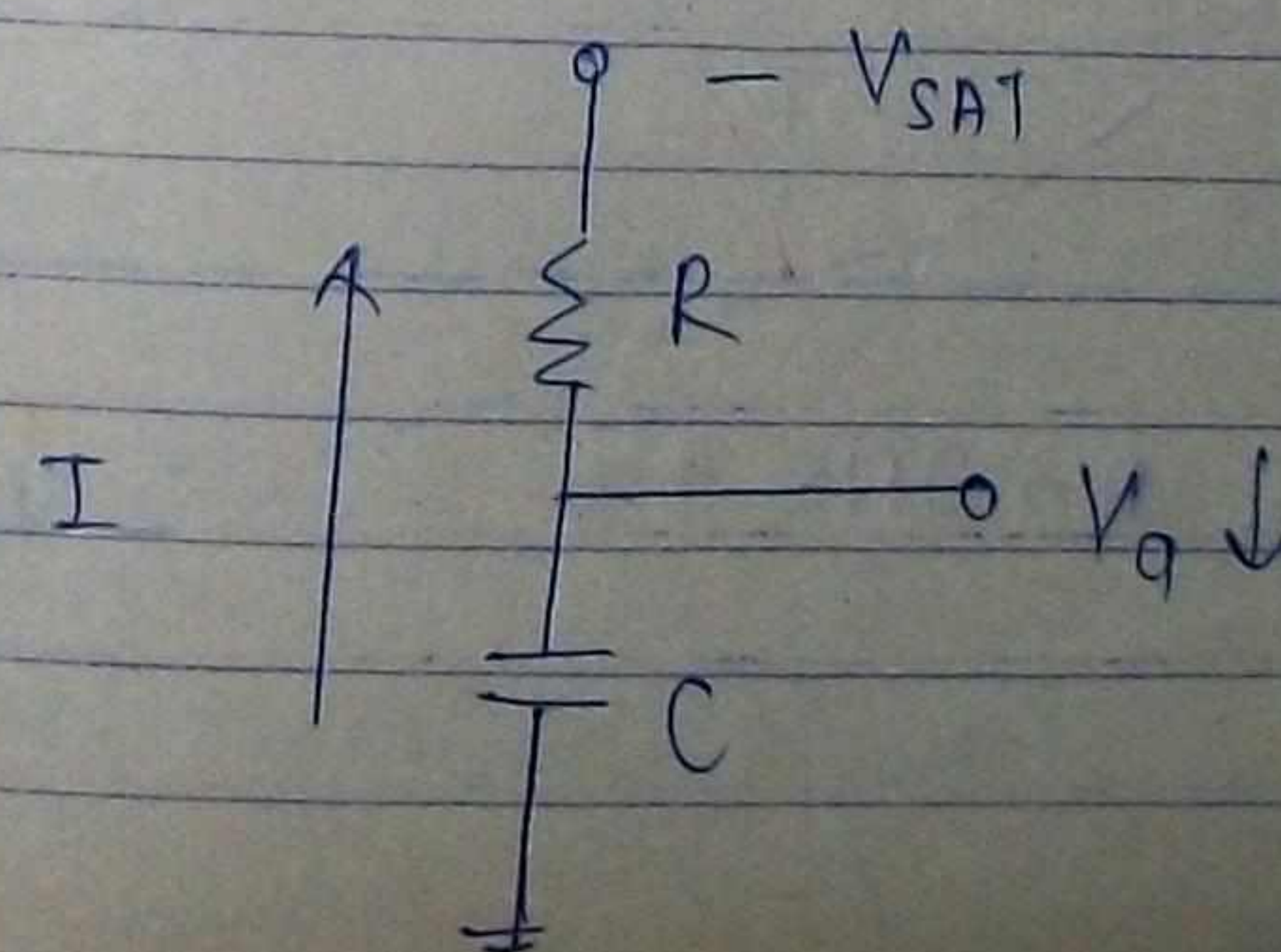


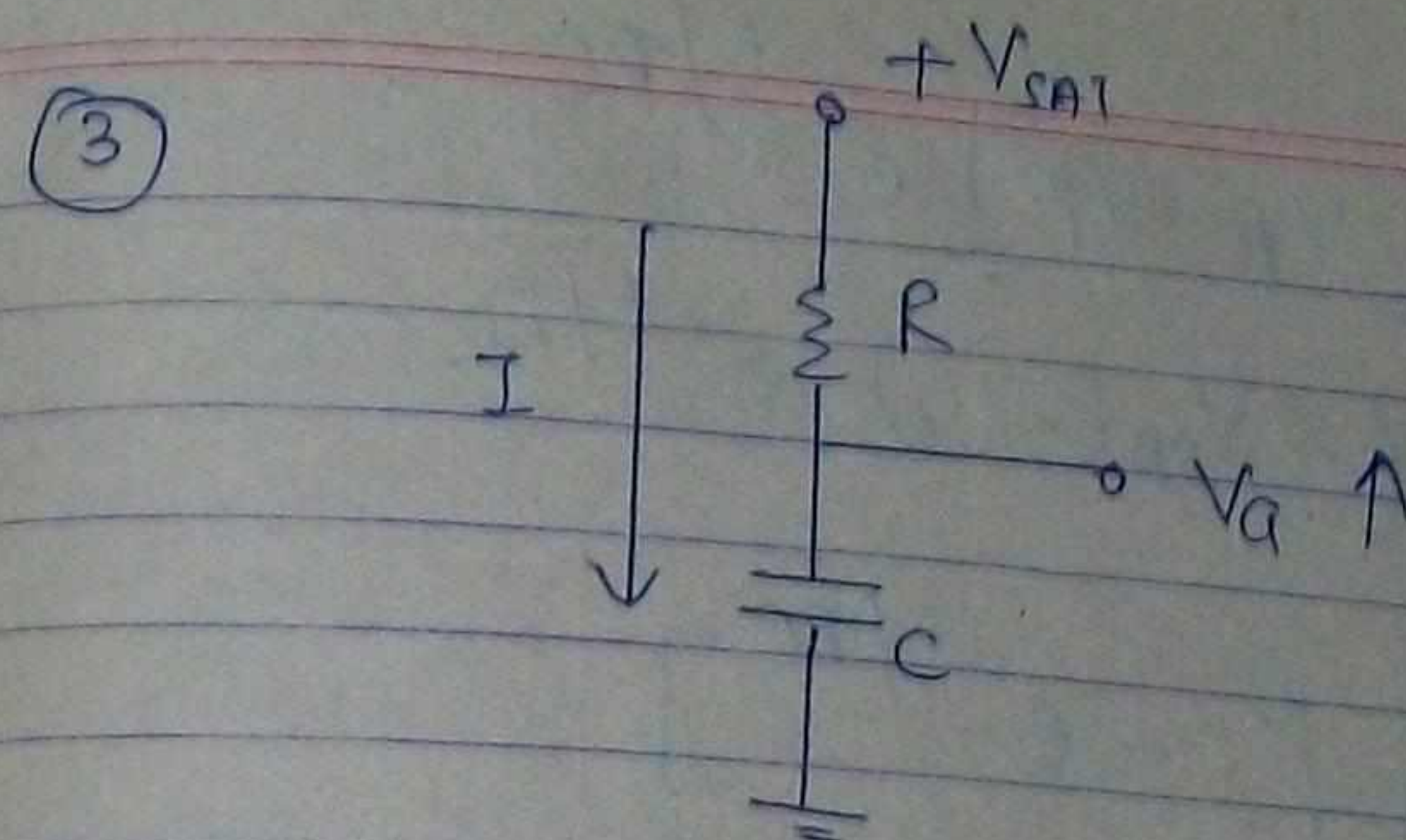
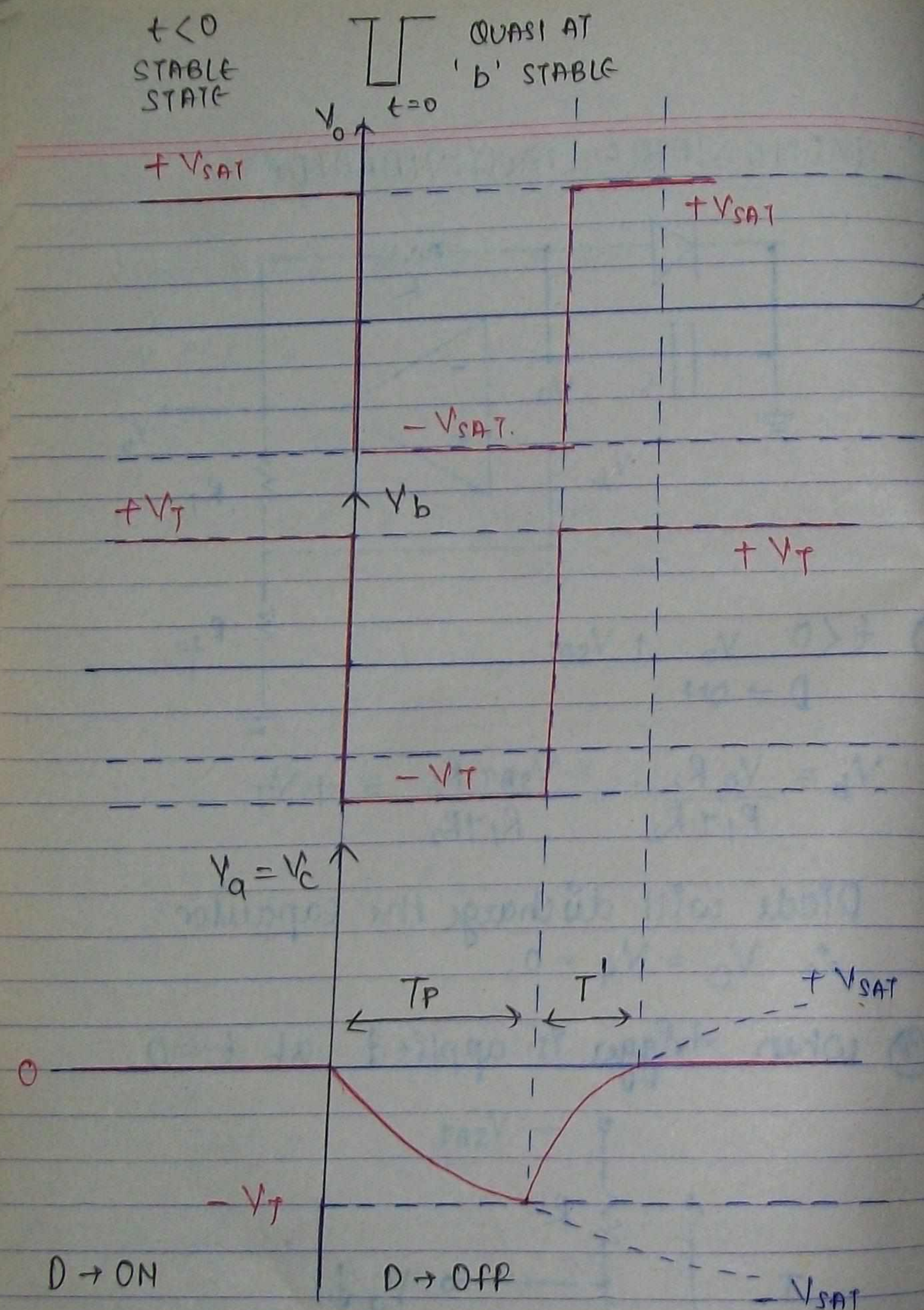
① $t < 0$ $V_0 = +V_{SAT}$
 $D \rightarrow ON$

$$V_b = \frac{V_0 R_2}{R_1 + R_2} = \frac{V_{SAT} R_2}{R_1 + R_2} = +V_T$$

Diode will discharge the capacitor
 $\therefore V_C = V_q = 0$.

② when trigger is applied at $t = 0$





④ minimum allowable interval b/w two triggers

$$= T_P + T'$$

$\downarrow$ $\downarrow$
 DESIRED UNWANTED

⑤ max freq. of trigger

$$= \frac{1}{T_P + T'}$$

DERIVATION OF T_P (PULSE WIDTH) -

$C \rightarrow$ discharge from 0 to $-V_T$

$$V_i^0 = 0 \quad (t=0)$$

$$V_f = -V_{SAT} \quad (t=\infty), \quad \tau = RC$$

$$V_c = V_f - [V_f - V_i] e^{-t/RC}$$

$$V_c = -V_{SAT} - (-V_{SAT} - 0) e^{-t/RC}$$

$$* \quad V_c = -V_{SAT} [1 - e^{-t/RC}]$$

Now,

$$\text{at } t = T_p \quad V_c = -V_T$$

$$-V_T = -V_{SAT} [1 - e^{-T_p/RC}]$$

$$* \quad T_p = RC \ln \left(\frac{V_{SAT}}{V_{SAT} - V_T} \right)$$

$$V_T = \frac{R_2}{R_1 + R_2} V_{SAT}$$

$$* \quad T_p = RC \ln \left[1 + \frac{R_2}{R_1} \right]$$

also,

$$* \quad T' = RC \ln \frac{V_{SAT} + V_T}{V_{SAT}}$$

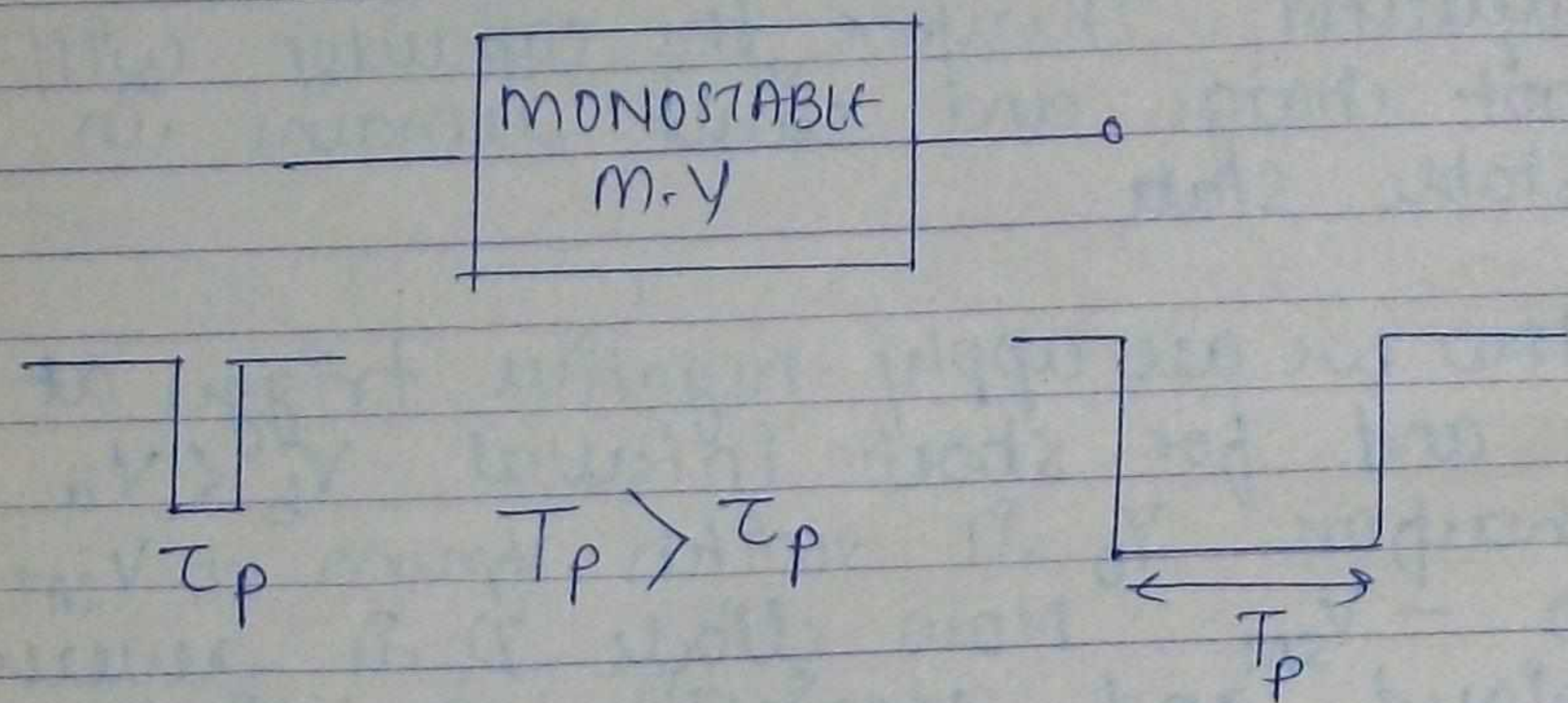
- Consider initial condition (for $t < 0$)
 $V_o = +V_{SAT}$ and $V_a = V_c = 0$ and no trigger input is applied then $V_b = +V_T$
 since $V_o = +V_{SAT}$ with diode D is forward biased and short circuit the capacitor. Therefore the capacitor will not charge and circuit remains in stable state.

- Now we are apply negative trigger at b and for short interval $V_b < V_a$
 therefore V_o is switch from $+V_{SAT}$ to $-V_{SAT}$. Now diode D is reverse biased and capacitor will discharge below zero toward $-V_{SAT}$ with $T = RC$. when capacitor discharge upto $-V_T$ then $V_a = V_b = -V_T$ and OPAMP comes out of saturation. When capacitor further discharge below $-V_T$ then $V_a < V_b$ as a result of which V_o switch over to $+V_{SAT}$ and again $V_b = +V_T$.

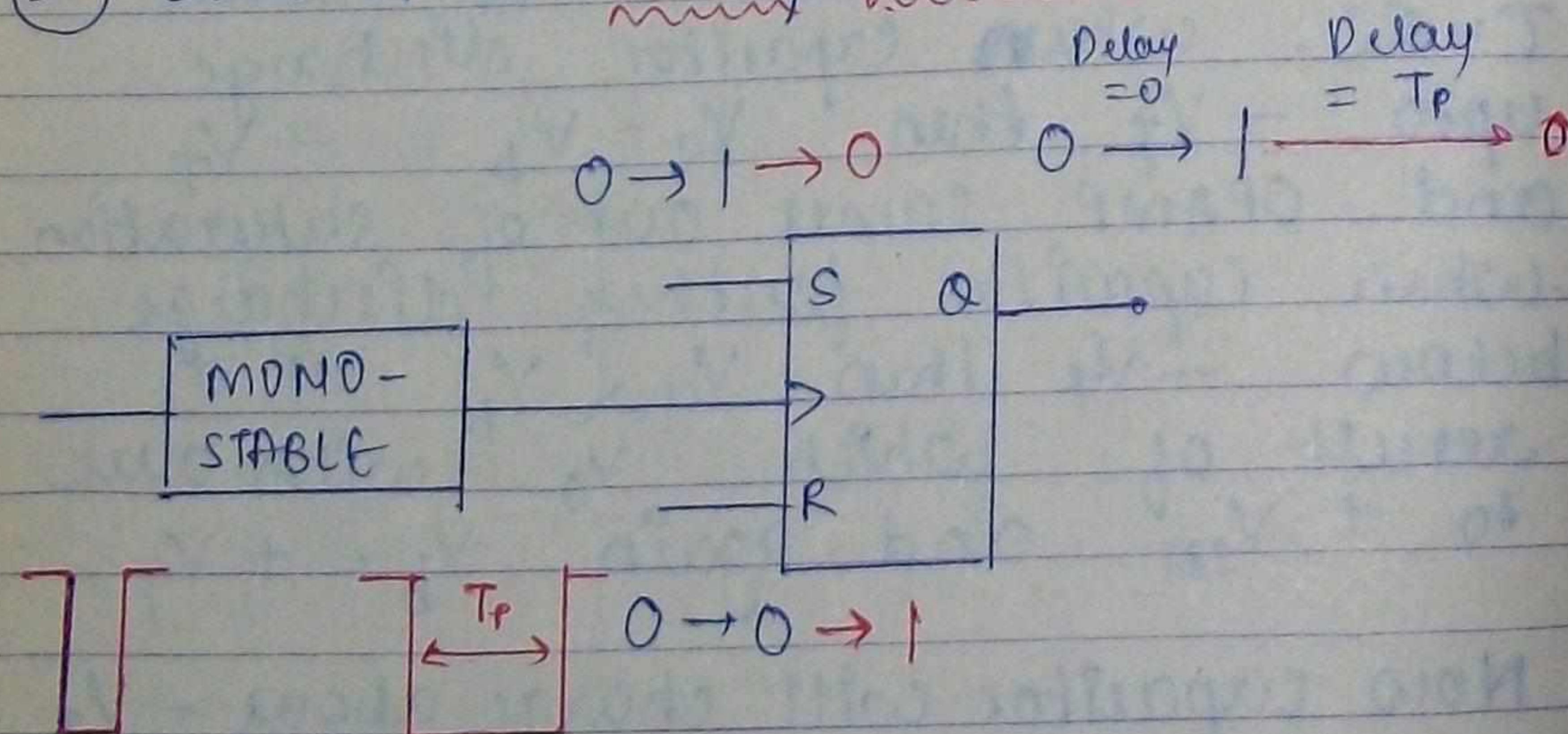
- Now capacitor will charge above $-V_T$ towards $+V_{SAT}$ but the capacitor can charge only upto '0' because when capacitor charge above zero, the diode 'D' become forward bias and short circuit the capacitor.

APPLICATION -

① used as pulse stretcher circuit.



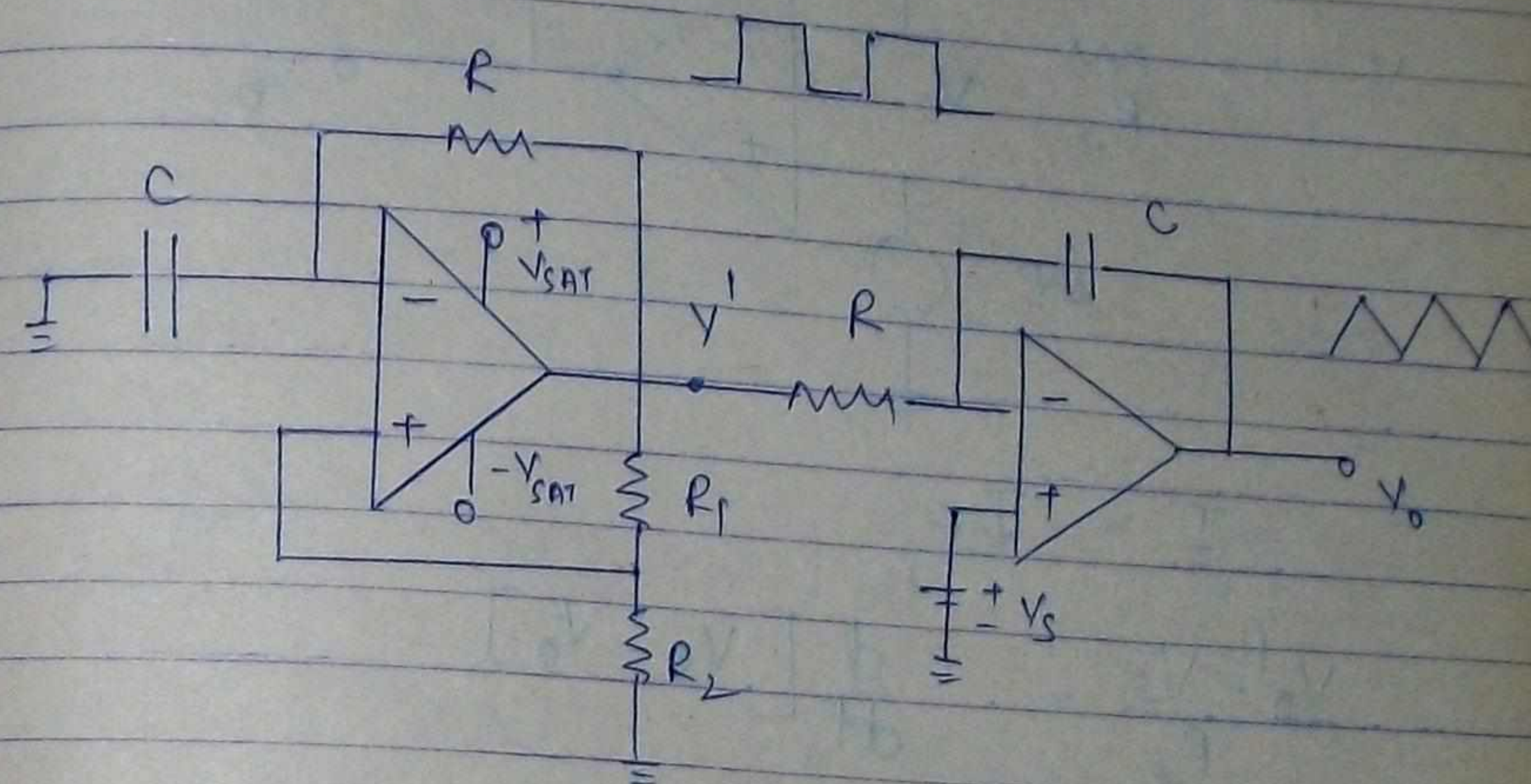
② used as a delay element.



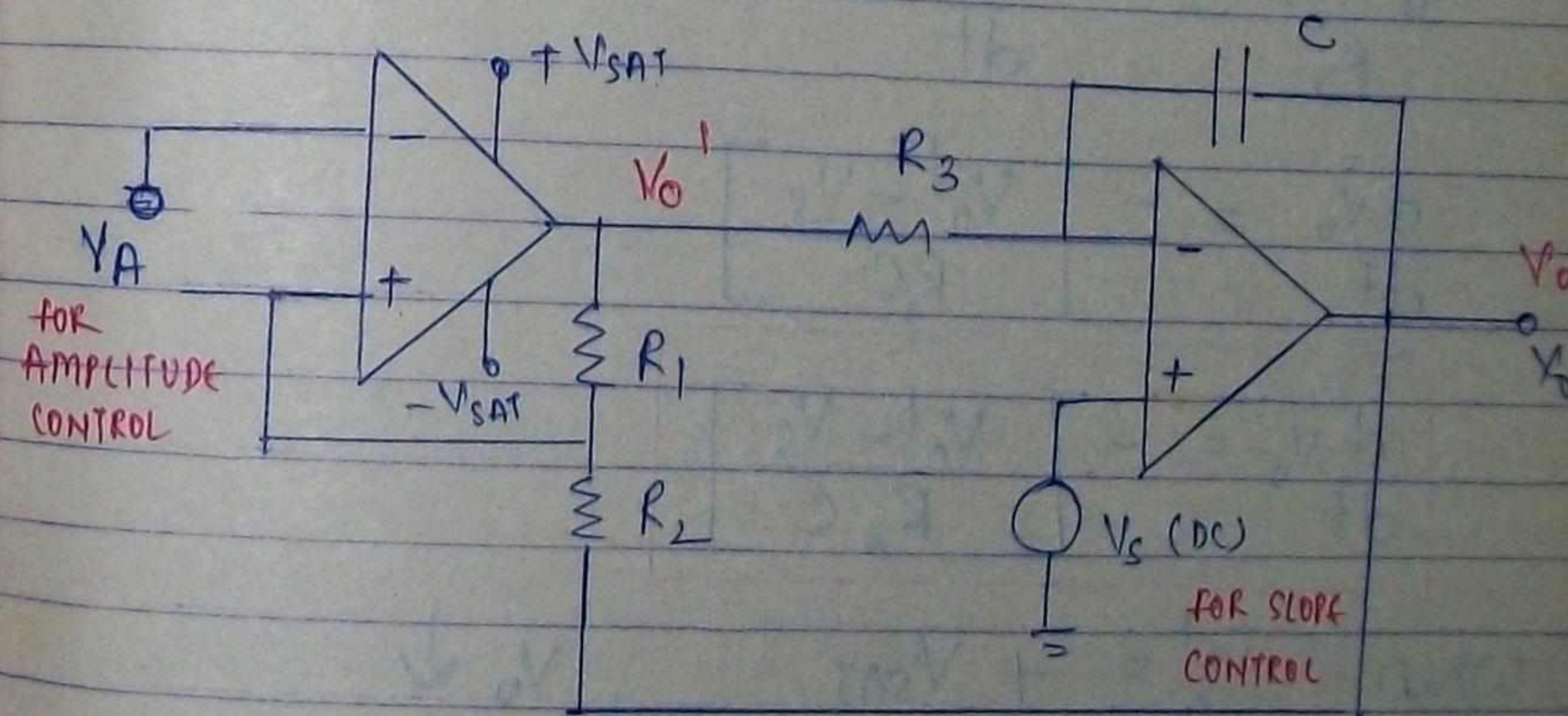
delay b/w execution of instructions.

23/08/11

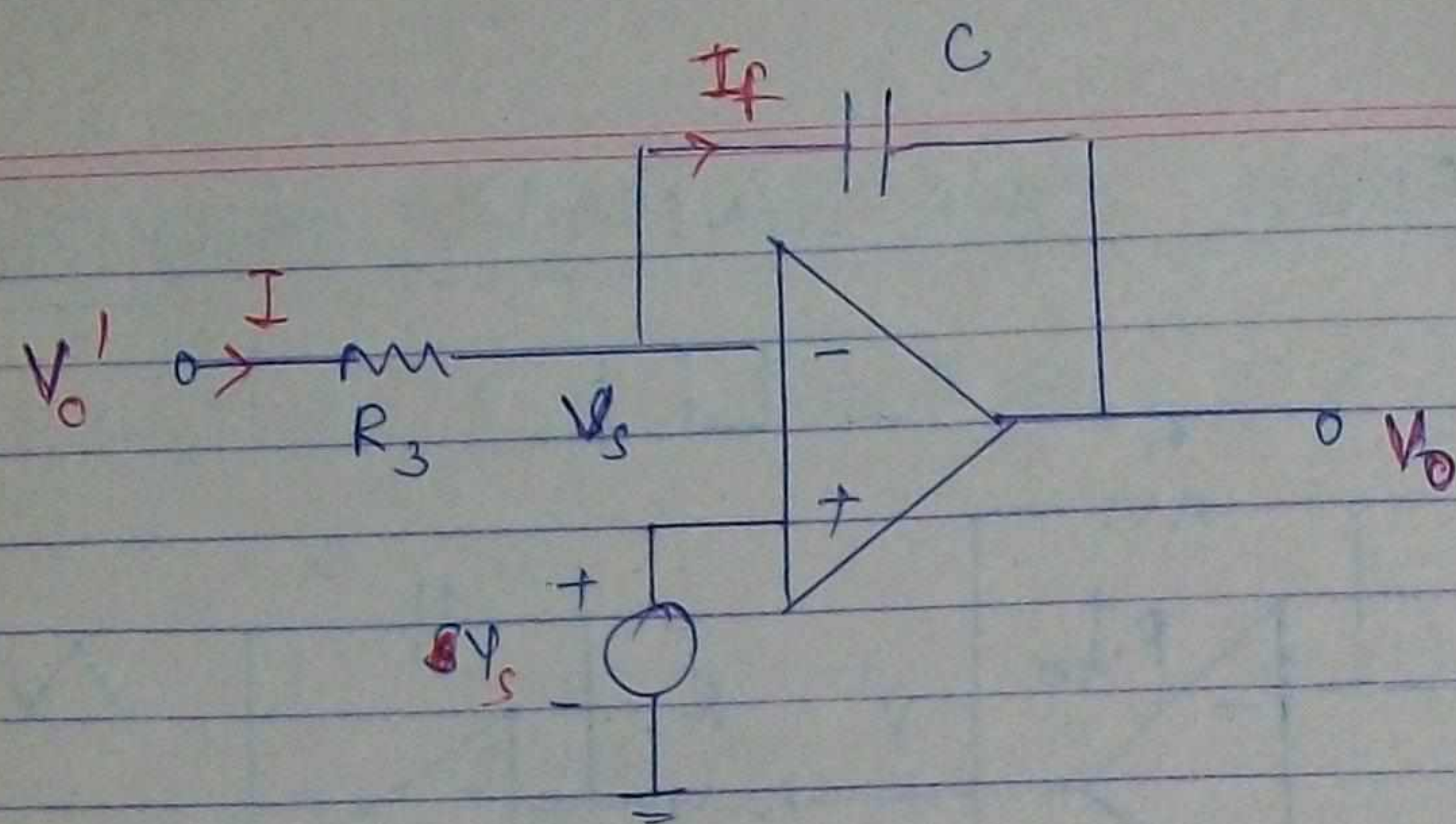
⇒ TRIANGLE WAVE GENERATOR -



④ Above circuit is generally not used due to presence of large no of components.



① SLOPE CONTROL METHOD -



$$I = I_f$$

$$\frac{V_o' - V_s}{R_3} = C \frac{d}{dt} [V_s - V_o]$$

$V_s = \text{constant (dc voltage)}$

$$\frac{V_o' - V_s}{R_3} = -C \frac{dV_o}{dt}$$

$$\frac{dV_o}{dt} = - \left[\frac{V_o' - V_s}{R_3 C} \right]$$

$$\frac{dV_o}{dt} = - \left[\frac{V_o' - V_s}{R_3 C} \right]$$

Ⓐ when $V_o' = +V_{SAT}$, $V_o \downarrow$

$$\frac{dV_o}{dt} = - \left[\frac{V_{SAT} - V_s}{R_3 C} \right]$$

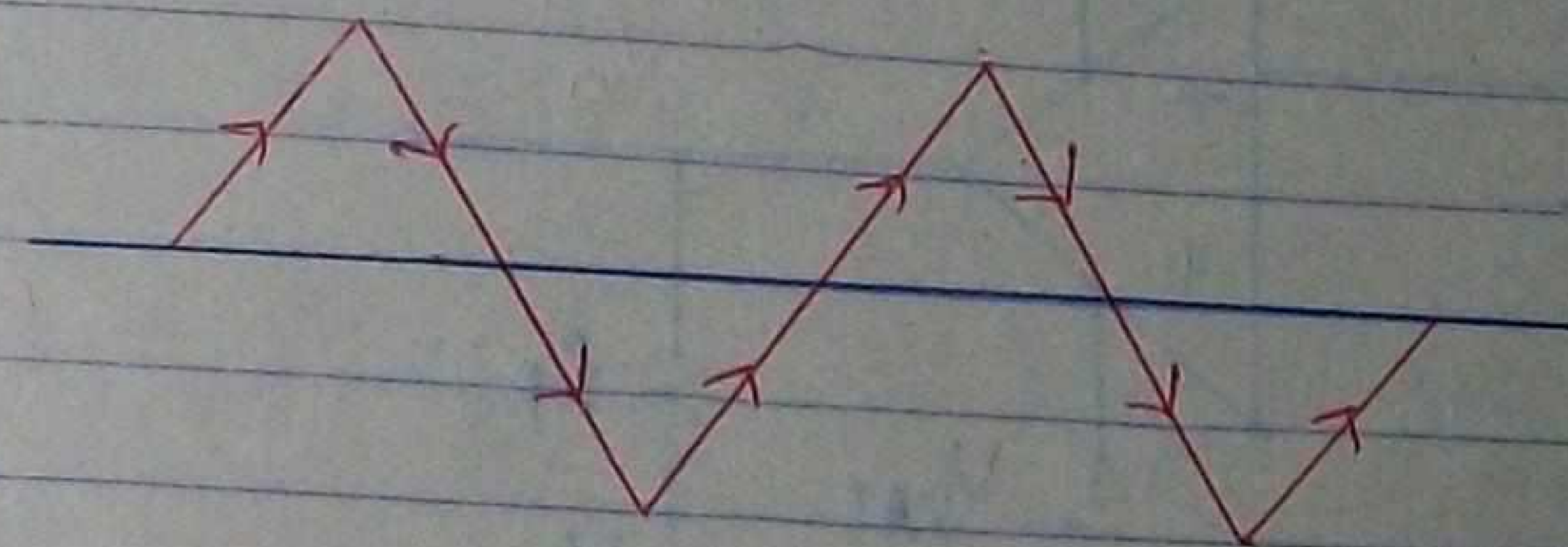
Ⓑ when $V_o' = -V_{SAT}$, $V_o \uparrow$

$$\frac{dV_o}{dt} = \left[\frac{V_{SAT} + V_s}{R_3 C} \right]$$

Ⓘ let $V_s = 0$:

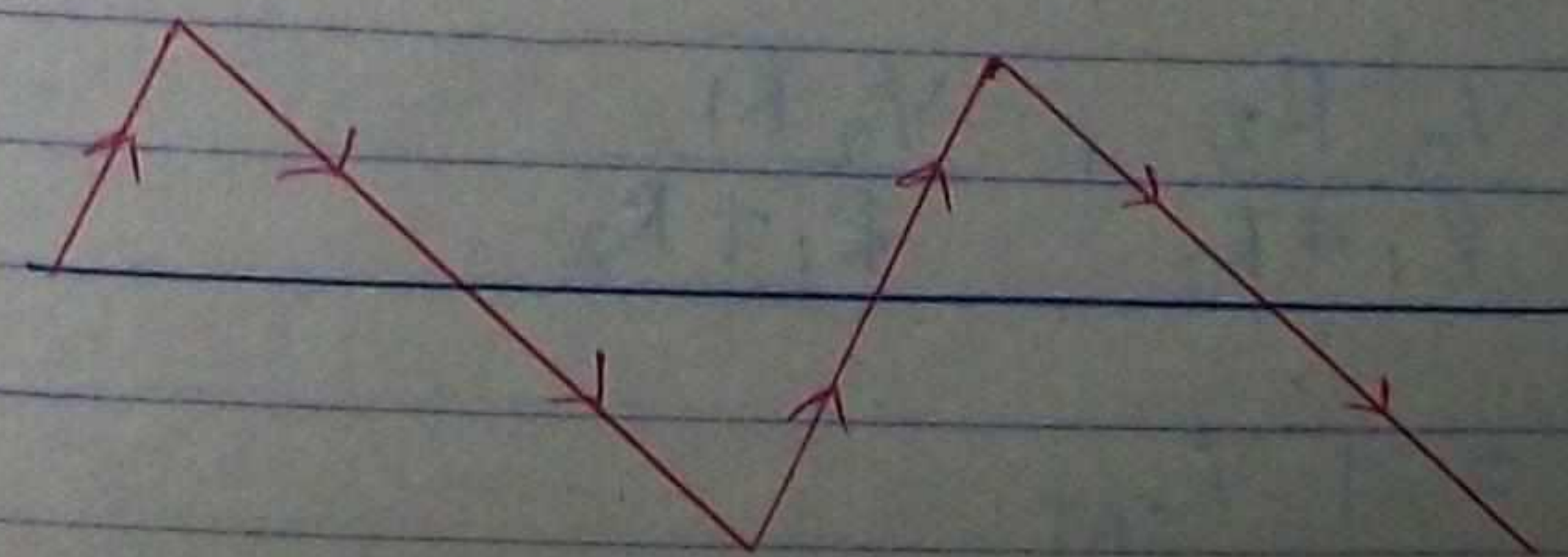
$$\frac{dV_o}{dt} = - \frac{V_{SAT}}{R_3 C}, \quad \frac{dV_o}{dt} = \frac{V_{SAT}}{R_3 C}$$

↑ slope = ↓ slope



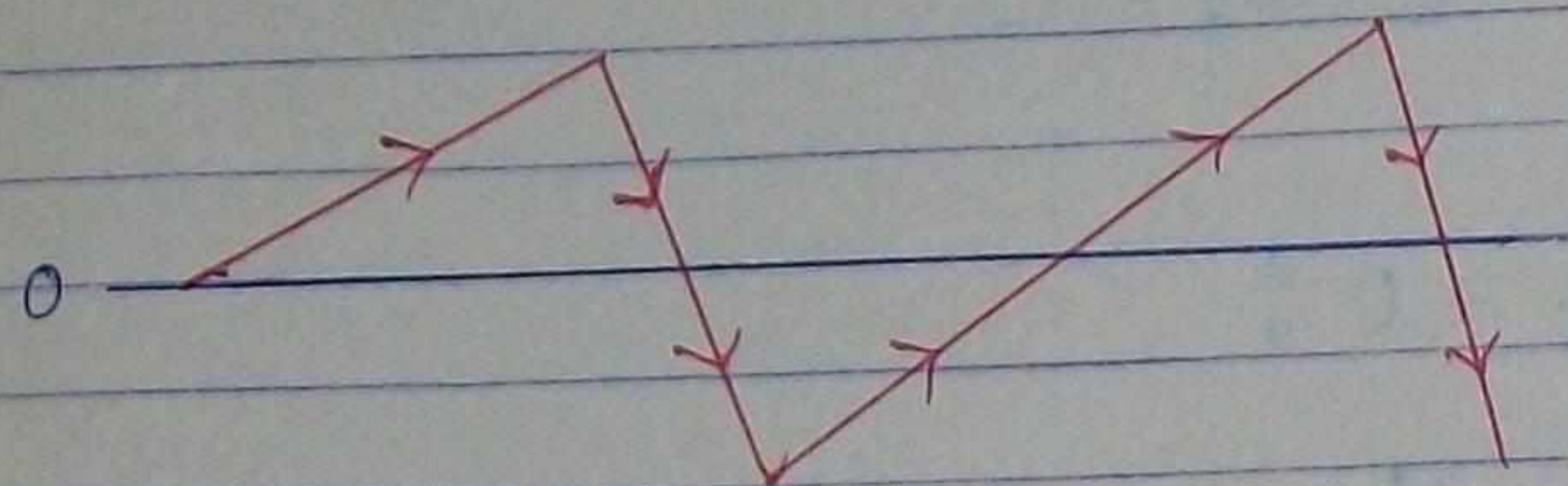
Ⓙ when $V_s > 0$:

↑ slope > ↓ slope

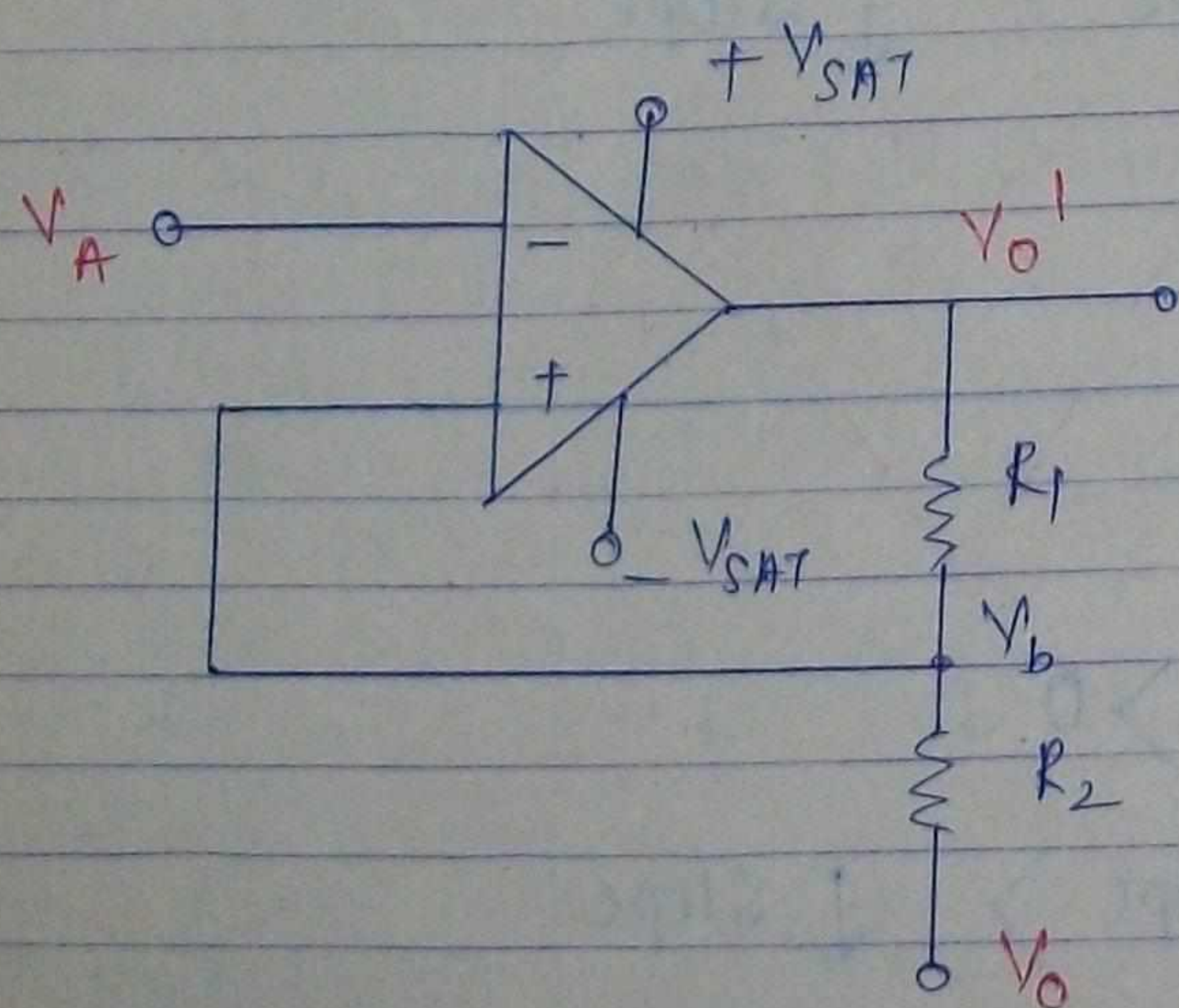


③ when $V_s < 0$:

↑ slope < ↓ slope



② AMPLITUDE CONTROL METHOD -



$$V_b = \frac{V_O' R_2}{R_1 + R_2} + \frac{V_O R_1}{R_1 + R_2}$$

(A) $V_O' = +V_{SAT}$

$$V_b = \frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_O R_1}{R_1 + R_2} = V_A$$

$$V_{OL} = \frac{R_1 + R_2}{R_1} \left[V_A - \frac{V_{SAT} R_2}{R_1 + R_2} \right]$$

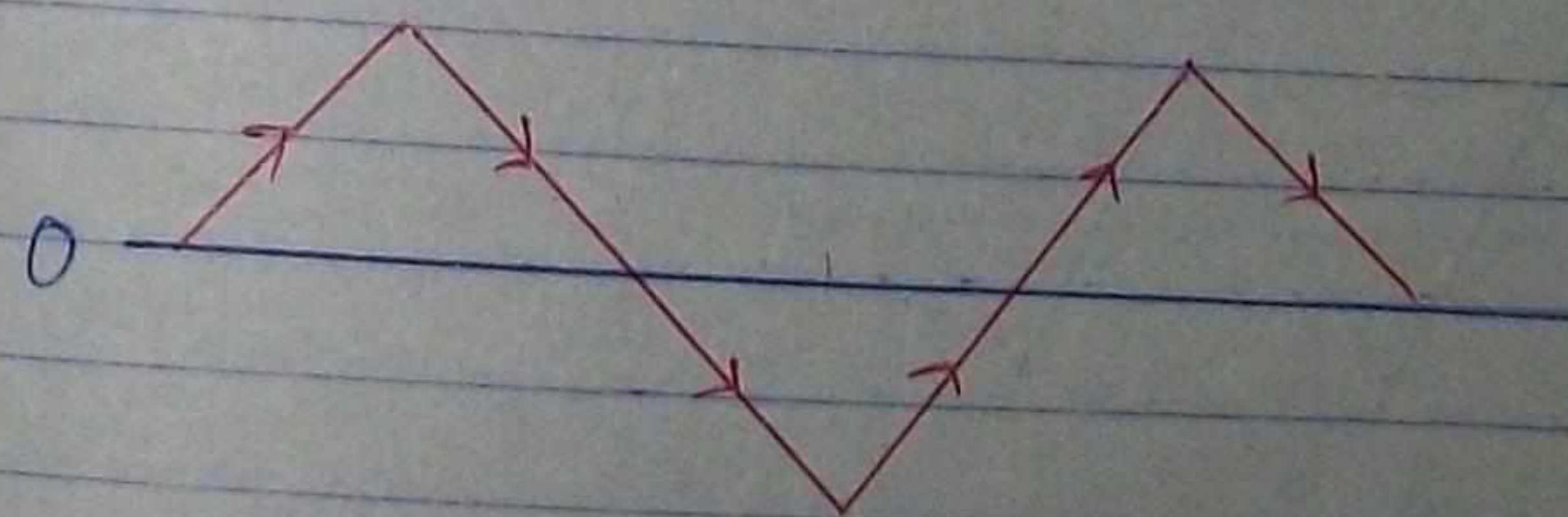
(B) $V_O' = -V_{SAT}$

$$V_b = -\frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_O R_1}{R_1 + R_2} = V_A$$

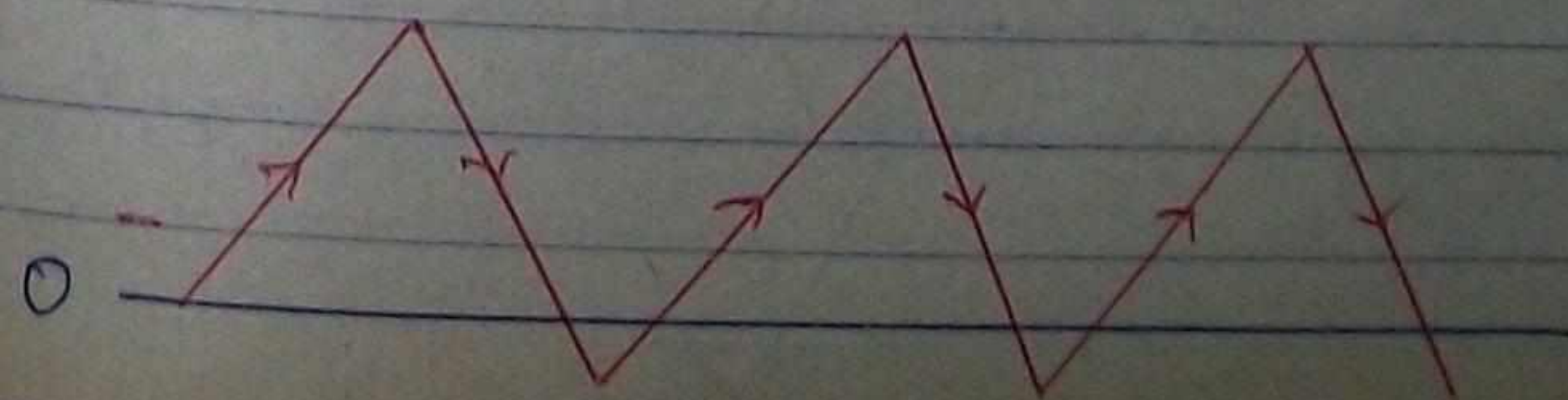
$$V_{OU} = \frac{R_1 + R_2}{R_1} \left[V_A + \frac{V_{SAT} R_2}{R_1 + R_2} \right]$$

(1) when $V_A = 0$

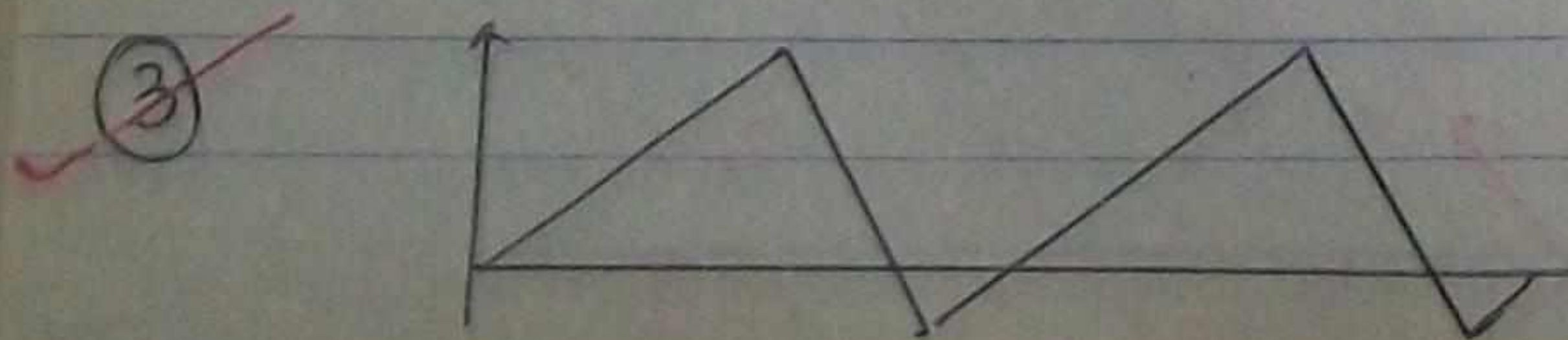
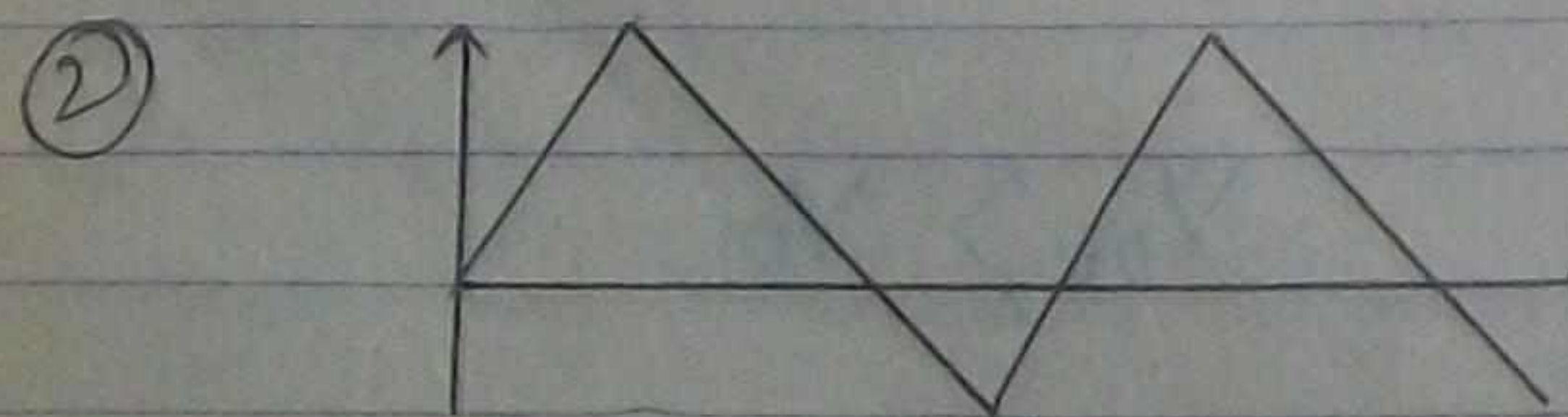
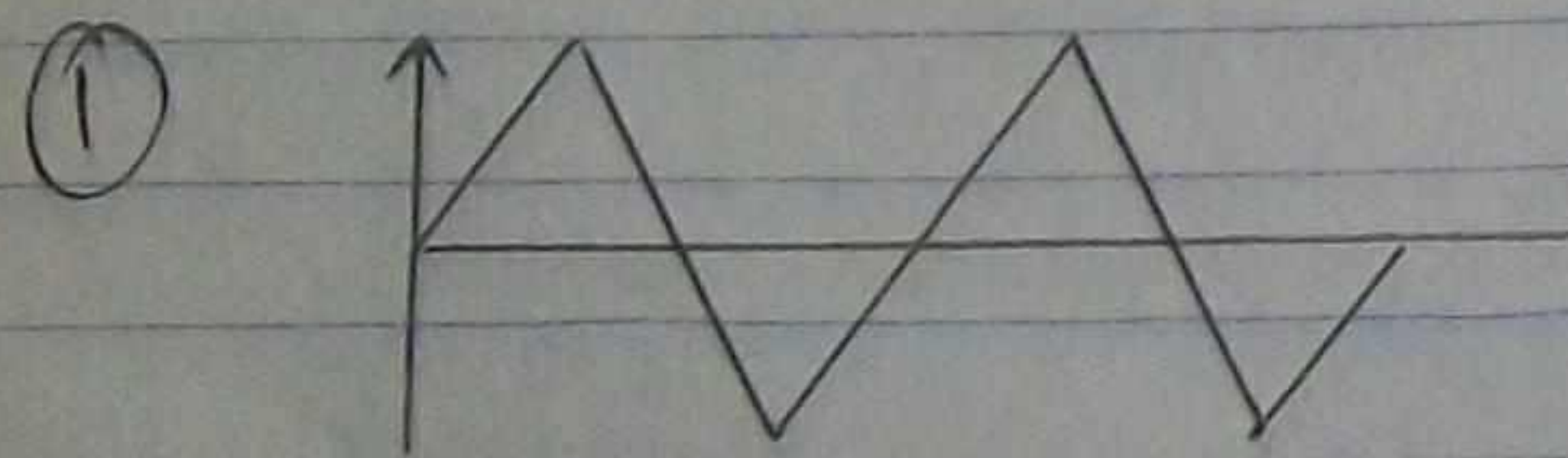
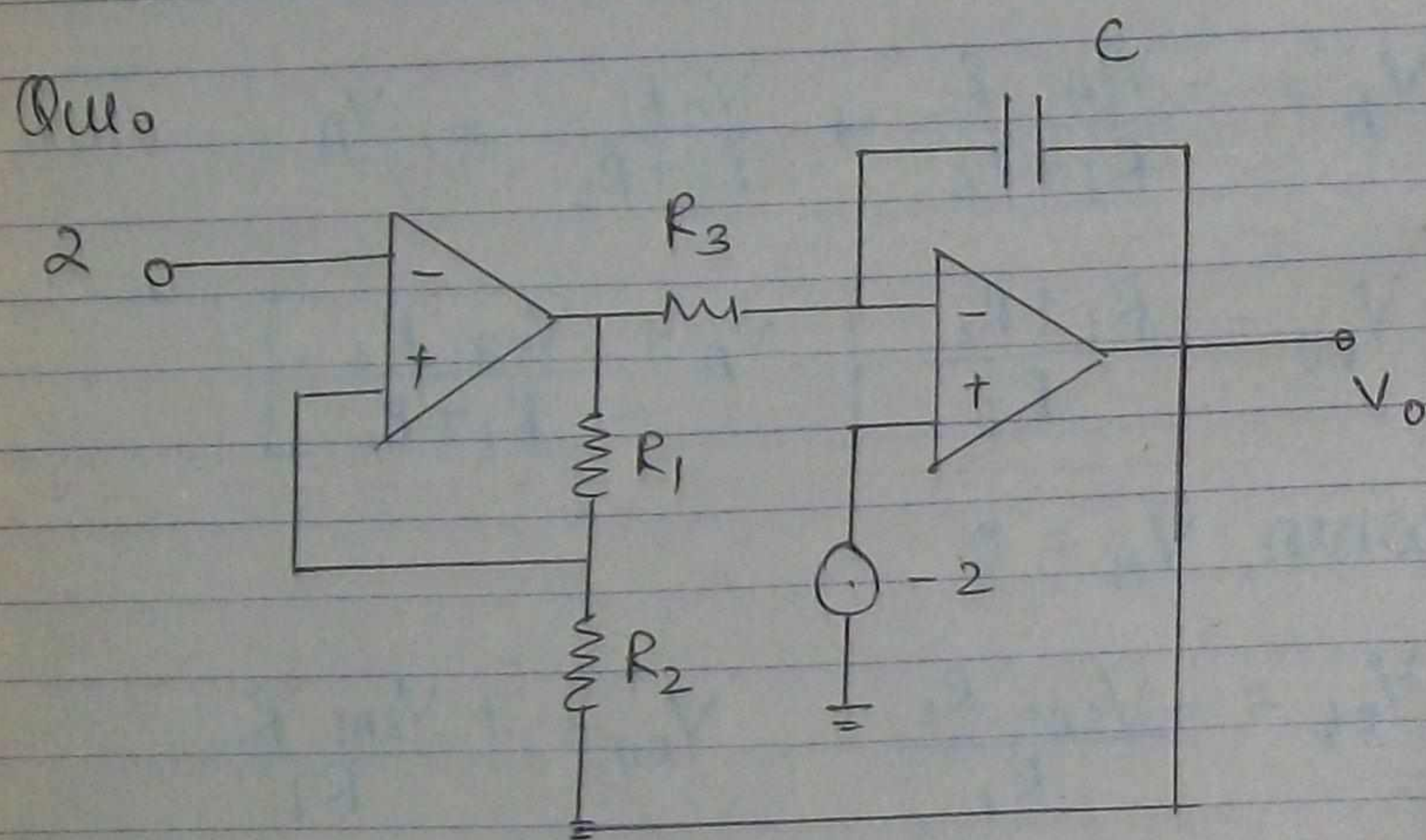
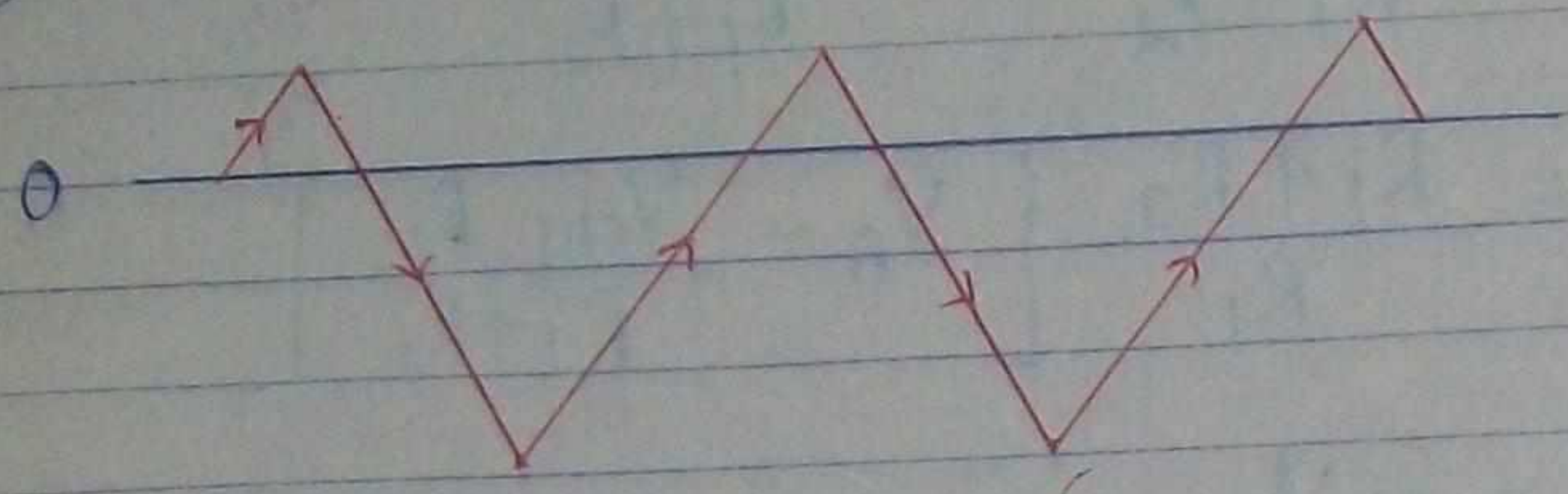
$$V_{OL} = -\frac{V_{SAT} R_2}{R_1}, \quad V_{OU} = +\frac{V_{SAT} R_2}{R_1}$$



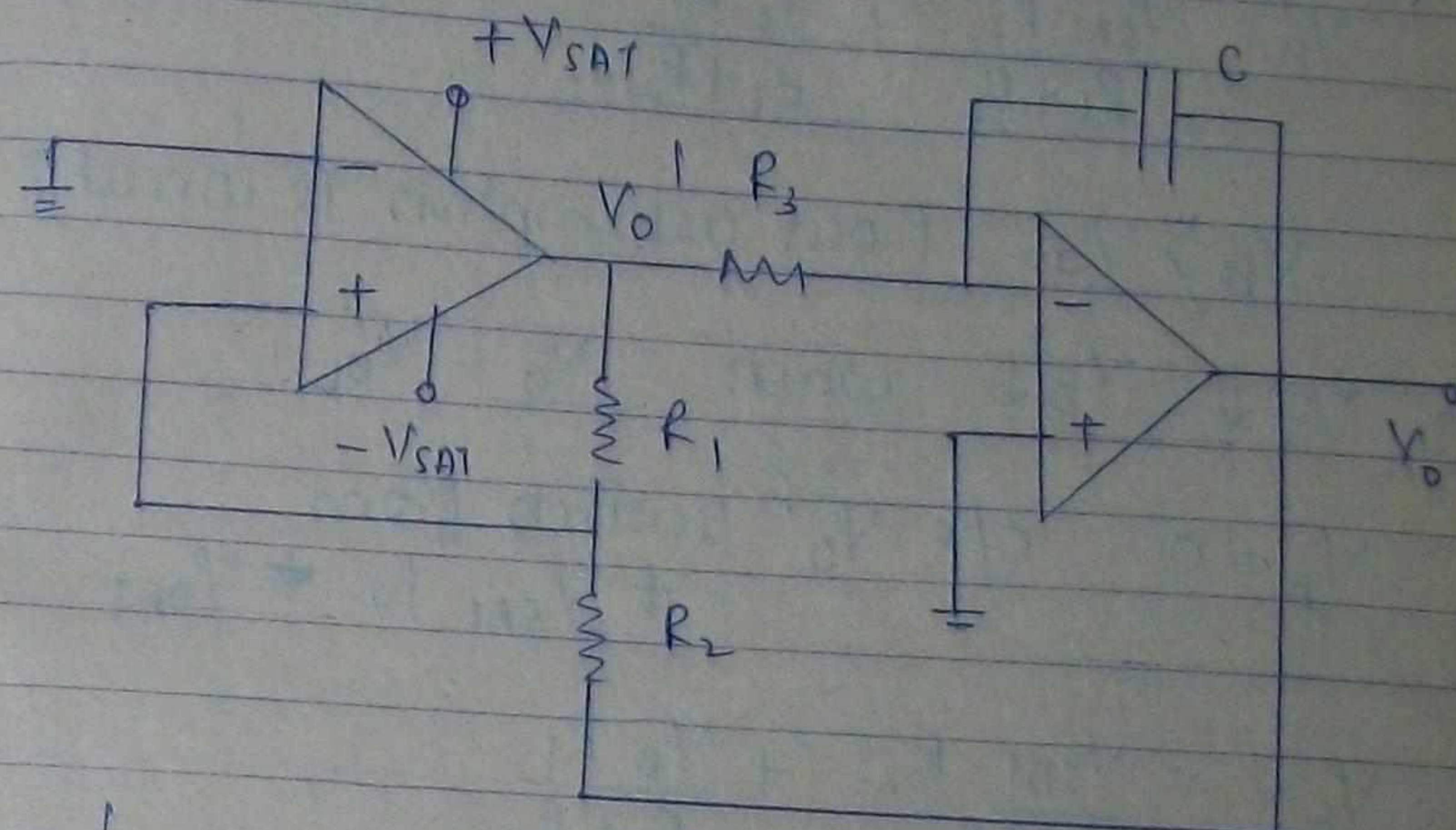
(2) when $V_A > 0$ $V_{OU} > V_{OL}$



⑤ when $V_A < 0$, $V_{OV} < V_{OL}$



⇒ SYMMETRICAL TRIANGULAR WAVE —
($V_S = 0$ & $V_A = 0$)



① $V_O' = +V_{SAT}$, $V_O \downarrow$

$$\frac{dV_O}{dt} = -\frac{V_{SAT}}{R_3 C}$$

② $V_O' = -V_{SAT}$, $V_O \uparrow$

$$\frac{dV_O}{dt} = \frac{V_{SAT}}{R_3 C}$$

③ $V_{OV} = +\frac{R_2}{R_1} V_{SAT}$

$$V_{OL} = -\frac{R_2}{R_1} V_{SAT}$$

① when $V_o = +V_{SAT}$ (let)

$$V_b = \frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2}, \quad V_a = 0$$

$V_b > V_a$ (our assumption is correct)

② $V_o \downarrow$, $V_b \downarrow$ when $V_o = V_{OL}$

$V_b = 0$ o/p V_o switch from $+V_{SAT}$ to $-V_{SAT}$

$$V_b = -\frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2}$$

$$V_b < 0, \quad V_a = 0$$

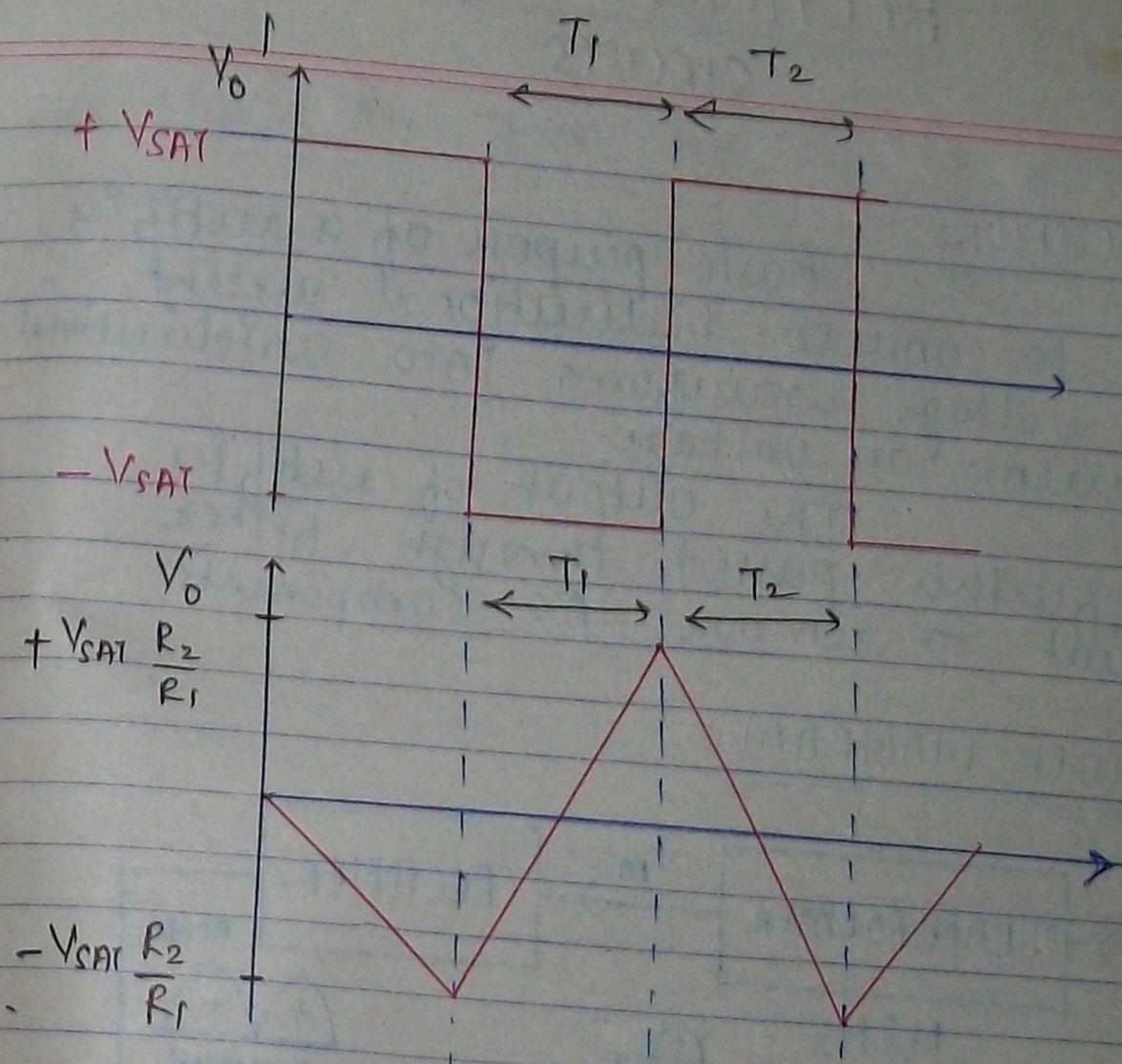
③ $V_o \uparrow$, $V_b \uparrow$ when $V_o = V_{OH}$

$V_b = 0$ o/p V_o switch from $-V_{SAT}$ to $+V_{SAT}$

$$V_b = \frac{V_{SAT} R_2}{R_1 + R_2} + \frac{V_o R_1}{R_1 + R_2}$$

$$V_b > 0, \quad V_a = 0$$

and so on.



$$\text{TIME} = \frac{\text{CHANGE}}{\text{RATE OF CHANGE}} = \frac{\text{FINAL} - \text{INITIAL}}{\text{SLOPE}}$$

$$T_1 = \frac{V_{SAT} R_2 / R_1 - (-V_{SAT} R_2 / R_1)}{V_{SAT} / R_3 C} = \frac{2 R_2 R_3 C}{R_1}$$

$$T_2 = \frac{-V_{SAT} R_2 / R_1 - V_{SAT} R_2 / R_1}{-V_{SAT} / R_3 C} = \frac{2 R_2 R_3 C}{R_1}$$

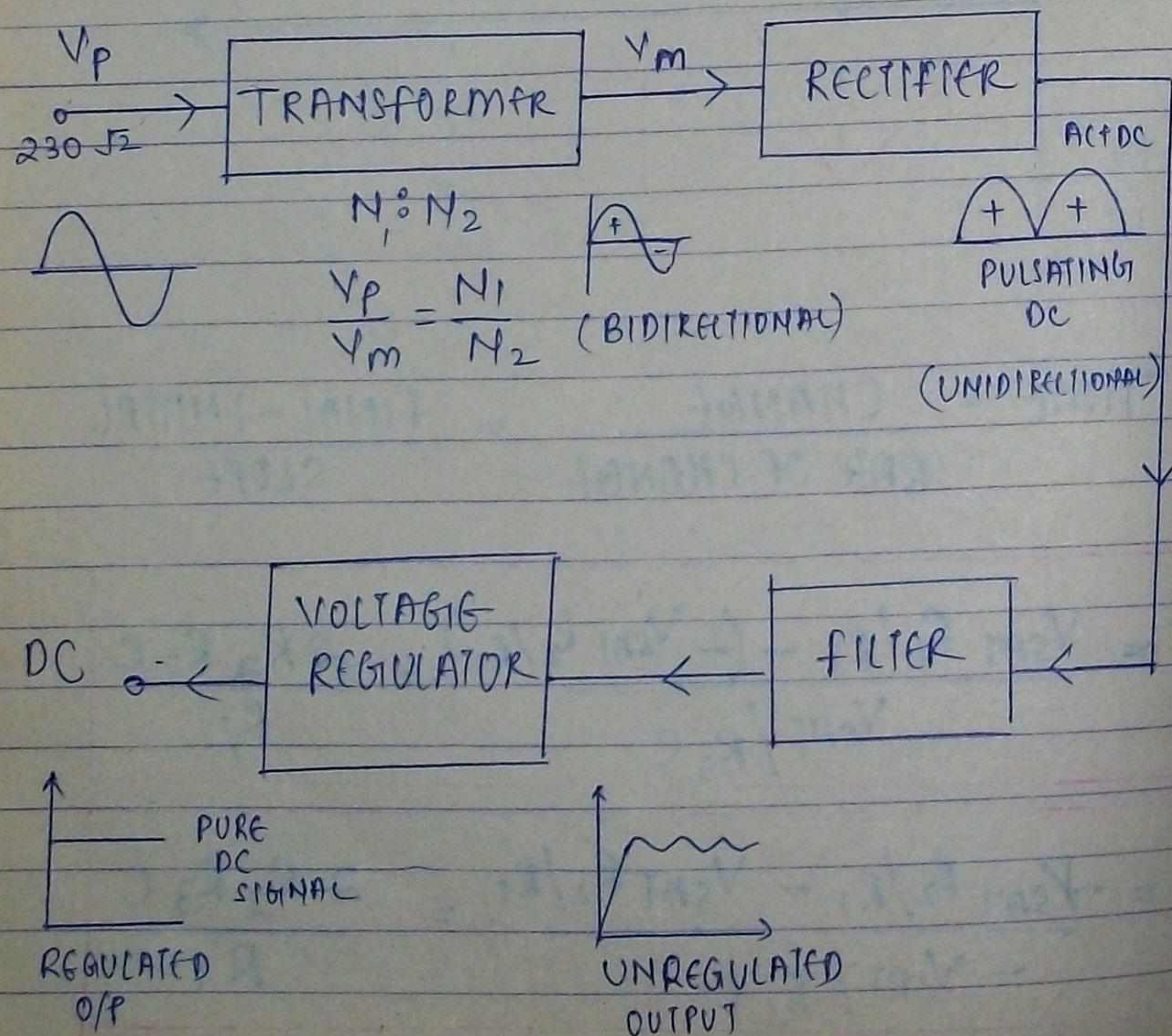
RECTIFIERS & DIODE CIRCUITS

⇒ RECTIFIERS -

Basic purpose of a rectifier is to convert bidirectional current or voltage waveform into unidirectional current or voltage.

The output of rectifier is further passed through filter circuit to remove AC components.

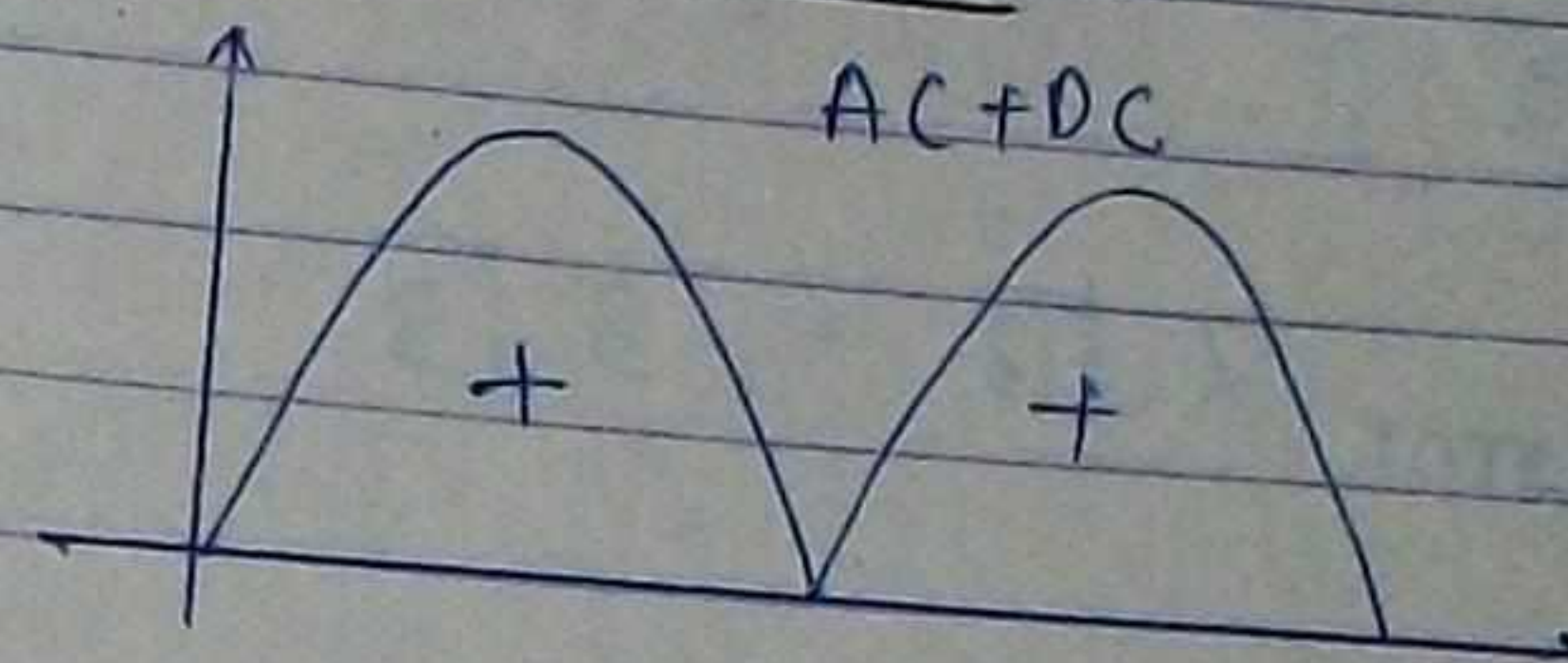
⇒ BLOCK DIAGRAM -



$$V_{rms} = \frac{V_p}{\sqrt{2}}$$

⇒ IMPORTANT TERMS -

① RIPPLE VOLTAGE -



$$V = V_{AC} + V_{DC}$$

$$V_{rms} = \text{rms value of } (AC + DC)$$

$$V_{DC} = \text{DC value}$$

$$V_{rms} = \sqrt{V_{acrms}^2 + V_{DC}^2}$$

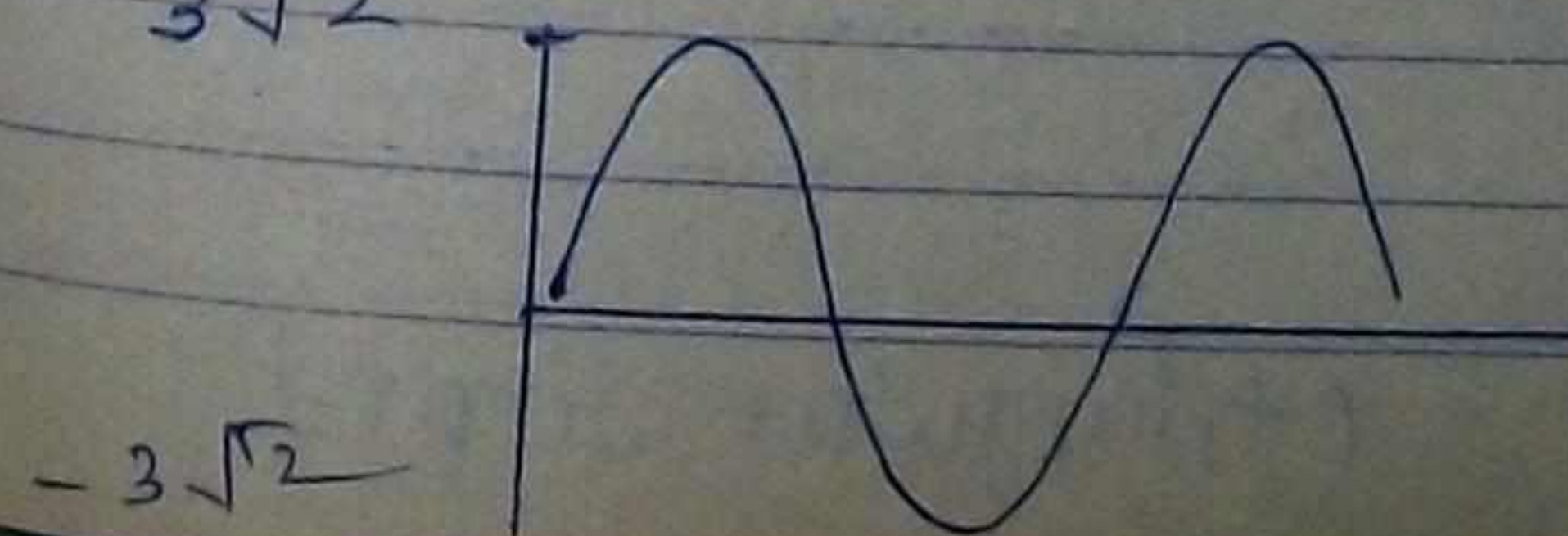
$$\therefore V_{acrms} = \sqrt{V_{rms}^2 - V_{DC}^2}$$

$$\text{Let } V_{acrms} = 3$$

① Sinusoidal representation of AC component:

$$V_p = 3\sqrt{2}$$

$$3\sqrt{2}$$

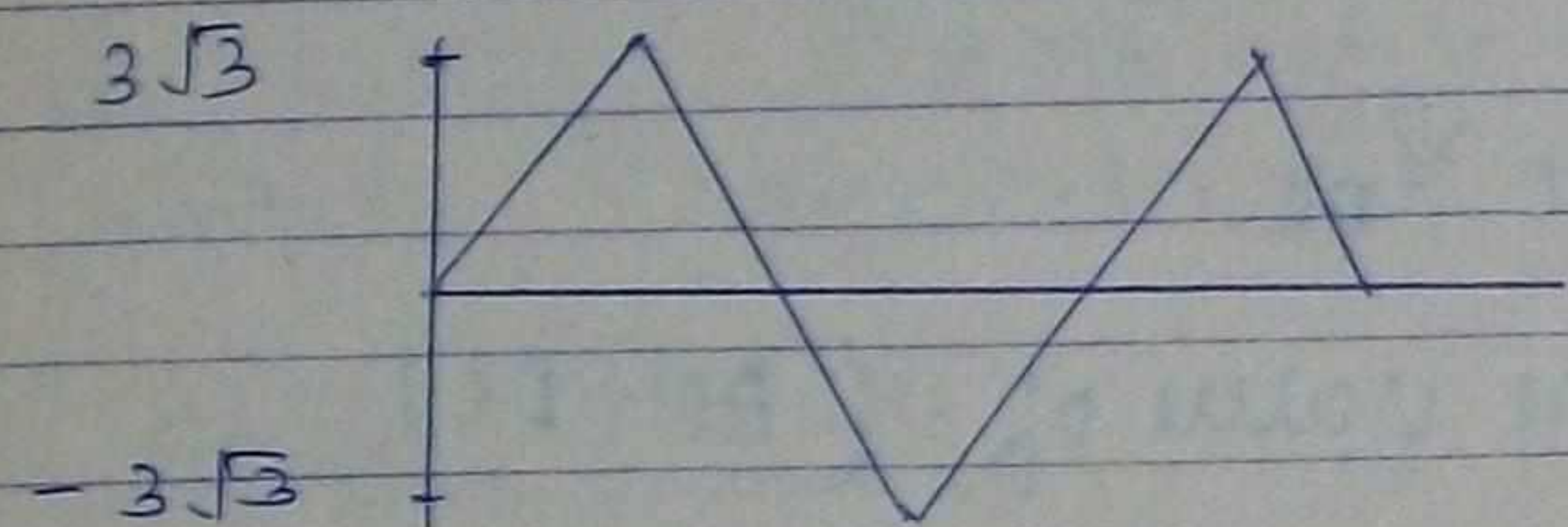


$$V_{acrms} = \frac{V_p}{\sqrt{2}}$$

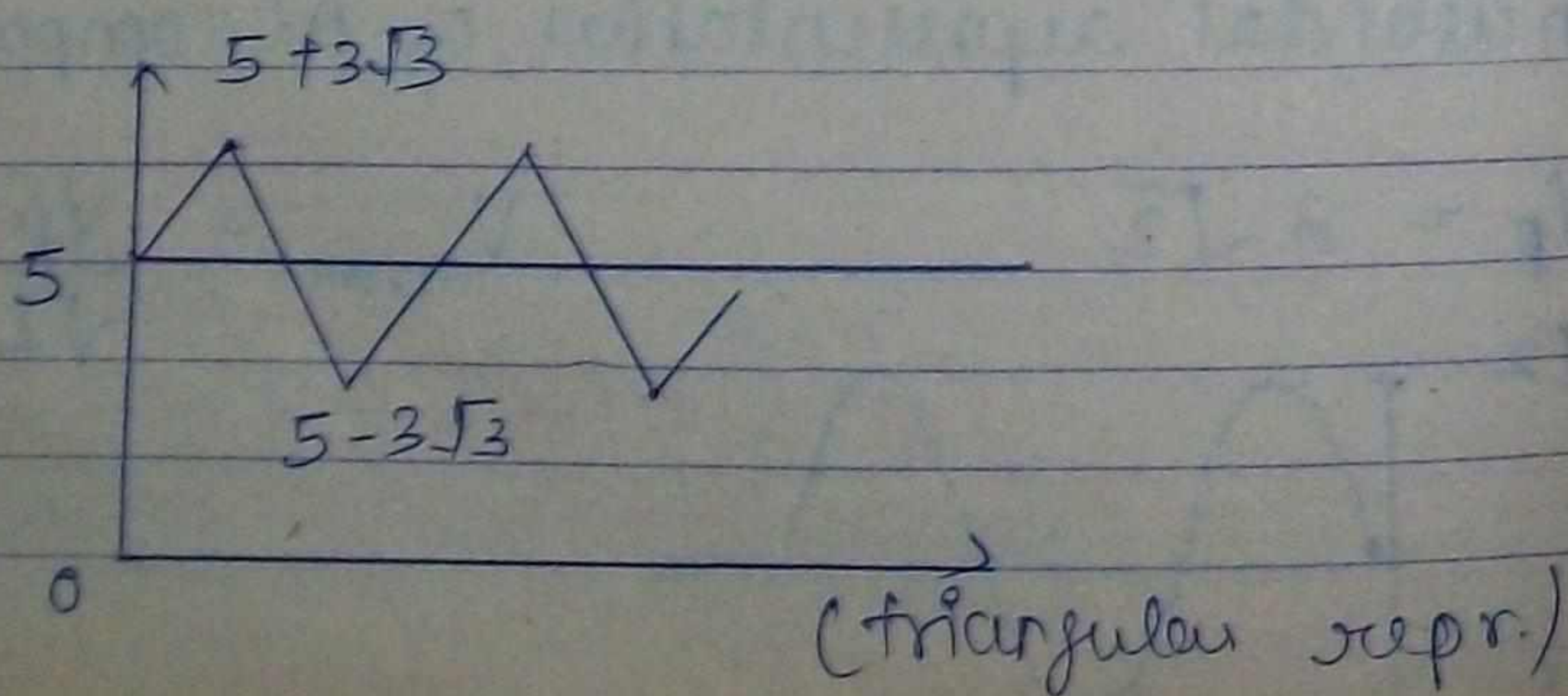
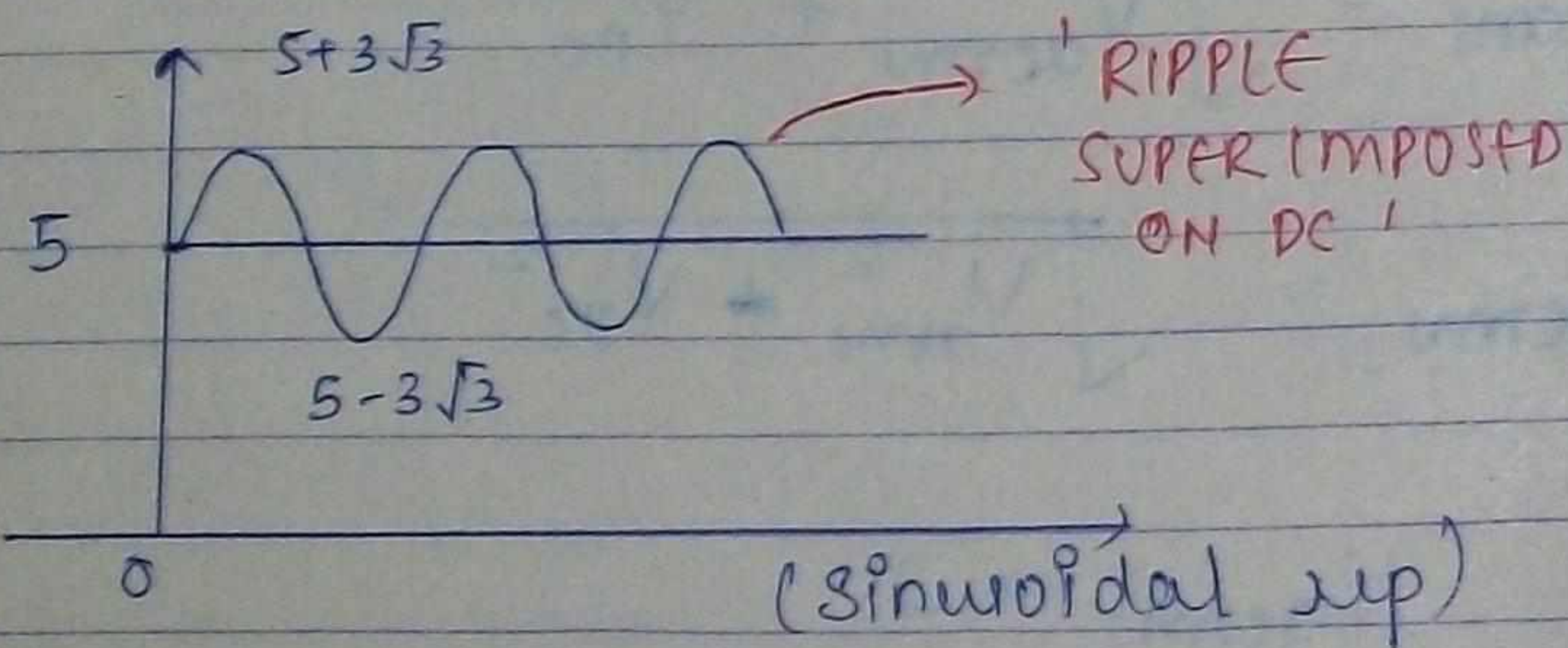
② Triangular representation of AC component

$$V_{acrms} = \frac{V_p}{\sqrt{3}}$$

$$V_p = V_{acrms} \times \sqrt{3} = 3\sqrt{3}$$



let $DC = 5$ $acrms = 3\sqrt{3}$



DEFINATION -

Ripple voltage (V_r) is defined as deviation of load voltage (output voltage) from its DC or average value.

The waveform after rectification is unidirectional but not pure DC, it has an ac component called 'RIPPLE SUPERIMPOSED ON DC'.

② RIPPLE FACTOR -

It is ratio of rms value of ac component to DC value, i.e.

$$r = \frac{\text{rms value of ac component}}{\text{DC value}}$$

$$r = \frac{V_{acrms}}{V_{DC}}$$

$$r = \frac{\sqrt{V_{rms}^2 - V_{DC}^2}}{V_{DC}}$$

$$r = \sqrt{\left(\frac{V_{rms}}{V_{DC}}\right)^2 - 1}$$

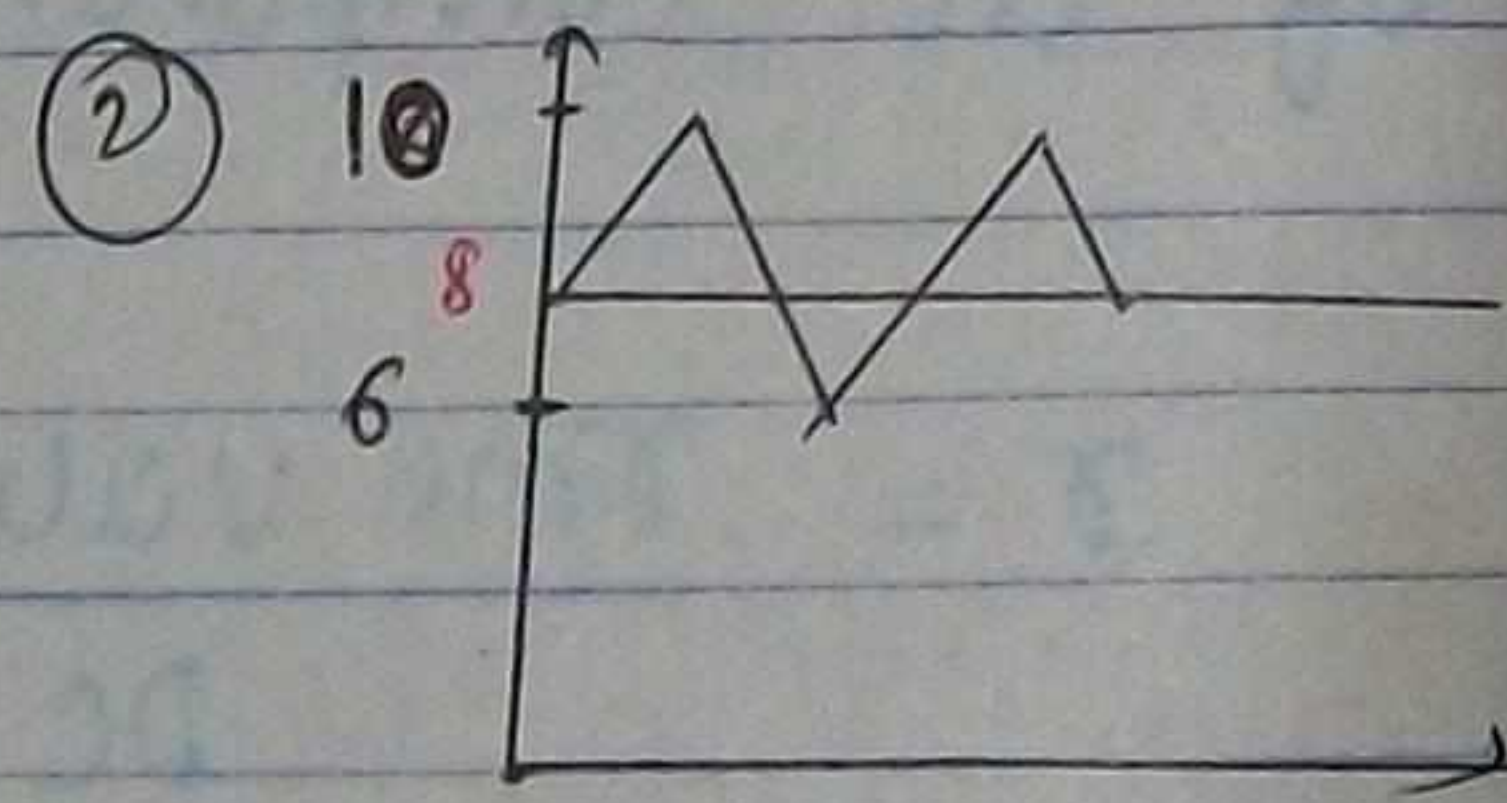
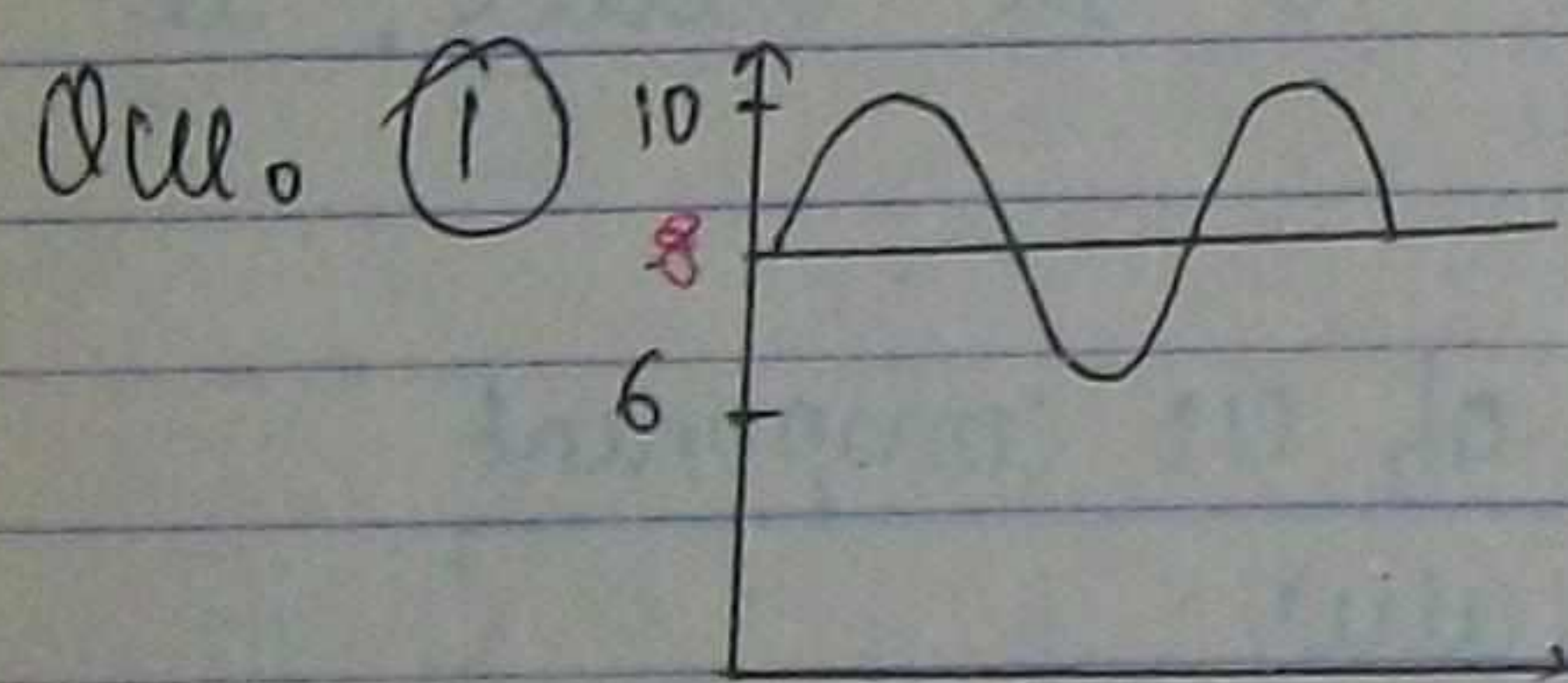
Form factor (F) -

$$F = \frac{V_{rms}}{V_{dc}}$$

$$r = \sqrt{F^2 - 1}$$

Ideally $r = 0$

$\therefore F = 1$



Calculate r ?

Solⁿ ① $V_{acrms} = \frac{V_p}{\sqrt{2}}$

$$V_{dc} = 8$$

$$V_{rms} = 8 + 2 \sin \omega t$$

$$r = \frac{V_{acrms}}{V_{dc}} = \frac{2/\sqrt{2}}{8} = \frac{1}{4\sqrt{2}}$$

② $V_{acrms} = \frac{V_p}{\sqrt{3}}$

$$V_{rms} = 8 + 2 \sin \omega t$$

$$V_{dc} = 8 \quad V_{acrms} = ? \quad V_p = 2$$

$$r = \frac{2/\sqrt{3}}{8} = \frac{1}{4\sqrt{3}}$$

Ques Draw equivalent representation of ① in triangular form.

Solⁿ: $DC = 8$

from ① $V_{rms} = 8 + 2 \sin \omega t$

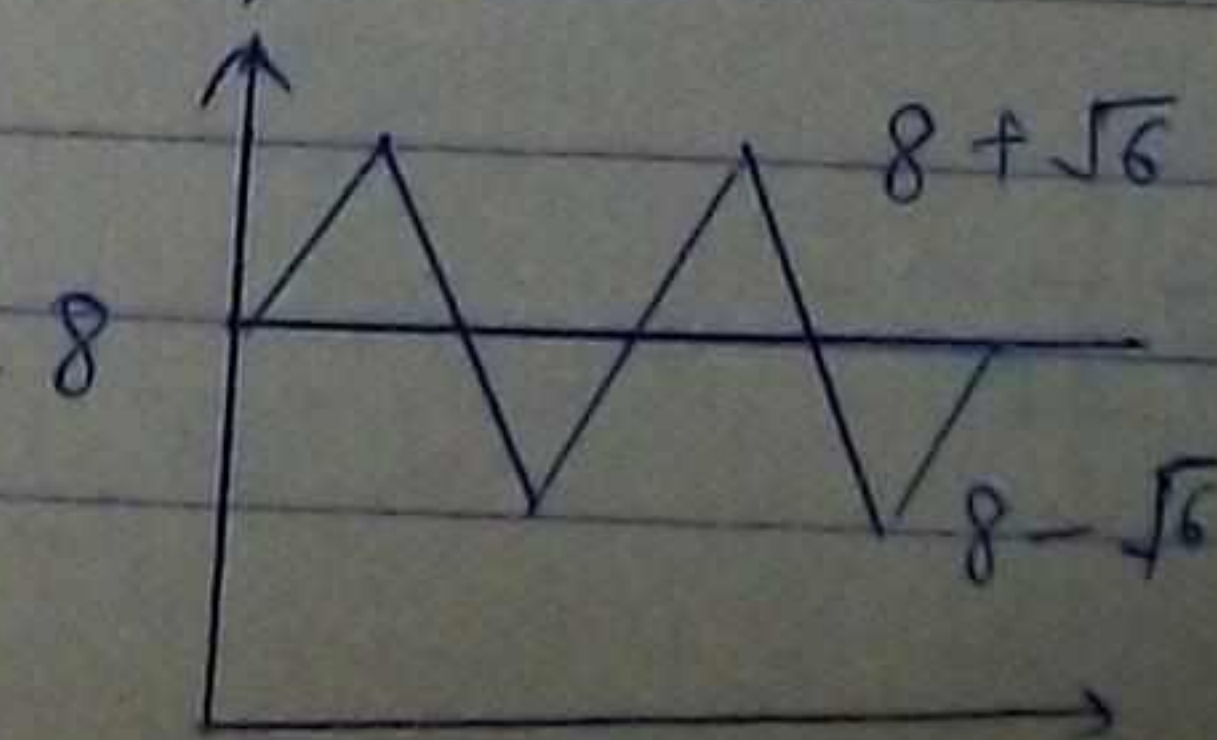
$$V_{dc} = 8 \quad V_{p-p} = 4, \quad V_p = 2$$

$$V_{acrms} = \frac{2}{\sqrt{2}} = \sqrt{2}$$

Now triangle,

$$V_{acrms} = \frac{V_p}{\sqrt{3}}$$

$$V_p = \sqrt{3} \cdot \sqrt{2} = \sqrt{6}, \quad V_{dc} = 8$$



③ CREST FACTOR (C) -

$$C = \frac{\text{PEAK VALUE}}{\text{RMS VALUE}}$$

(should be low)

④ RECTIFIER EFFICIENCY -

$$\eta = \frac{\text{O/P}}{\text{I/P}} \times 100\%$$

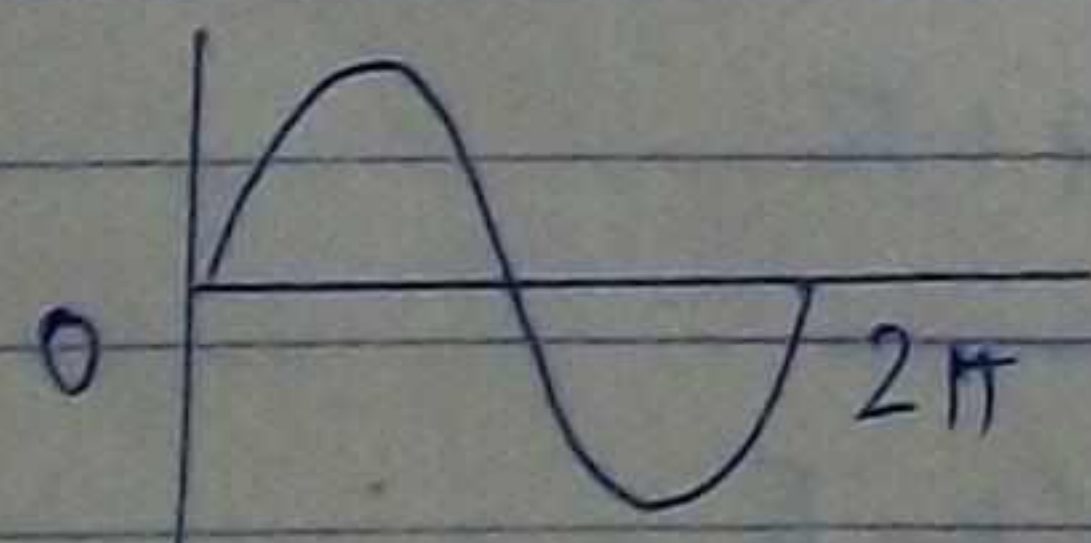
$$\eta = \frac{\text{O/P DC POWER}}{\text{I/P AC POWER}} \times 100\%$$

⑤ AVG. VALUE 'OR' DC VALUE -

$$I_{DC} = \frac{1}{2\pi} \int_0^{2\pi} I(t) d(\omega t)$$

⑥ RMS VALUE -

$$I_{rms} = \left[\frac{1}{2\pi} \int_0^{2\pi} I^2(t) d(\omega t) \right]^{1/2}$$



⑦ PEAK INVERSE VOLTAGE (PIV) -

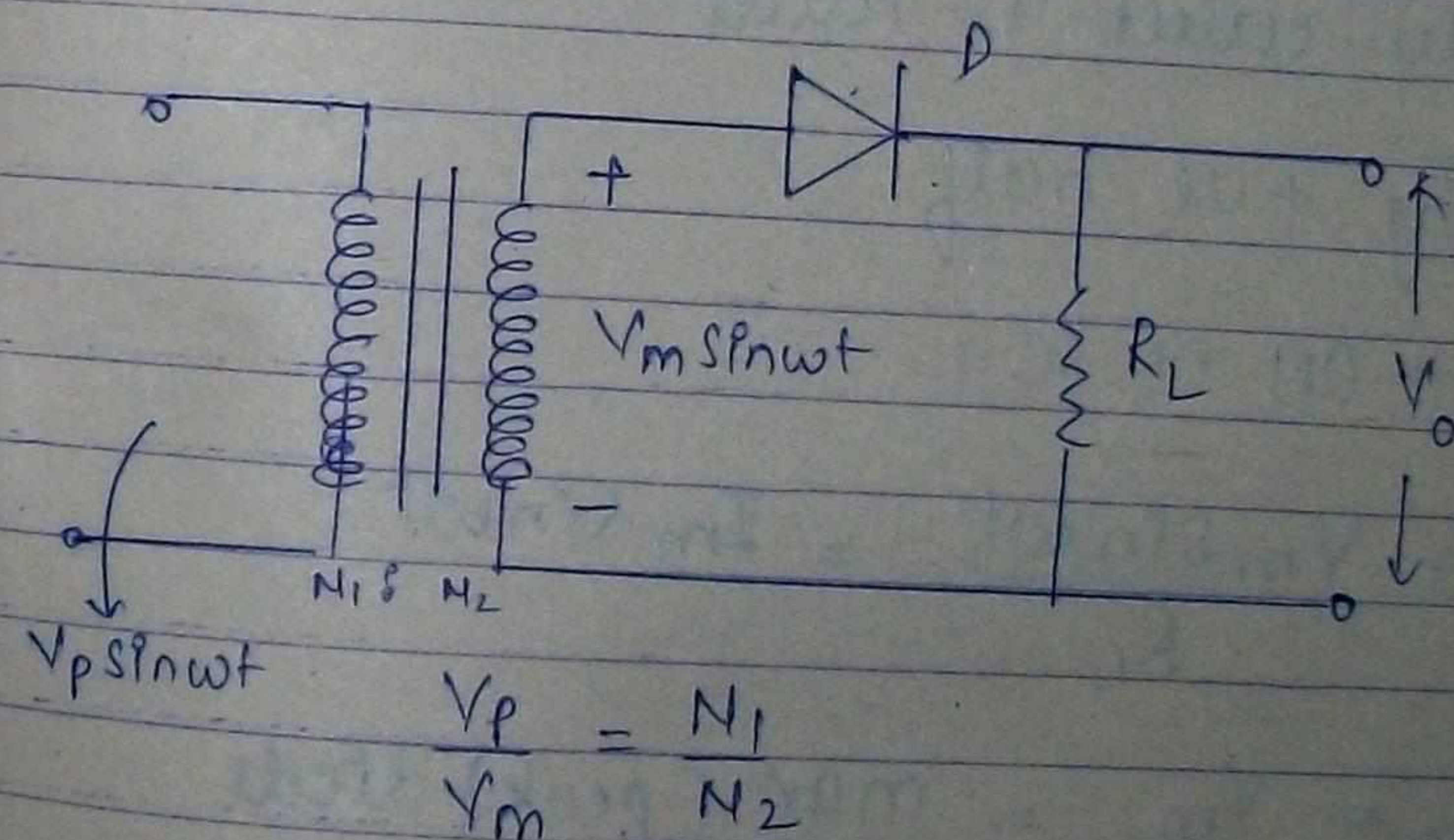
The maximum voltage across the diode when diode is reverse biased.

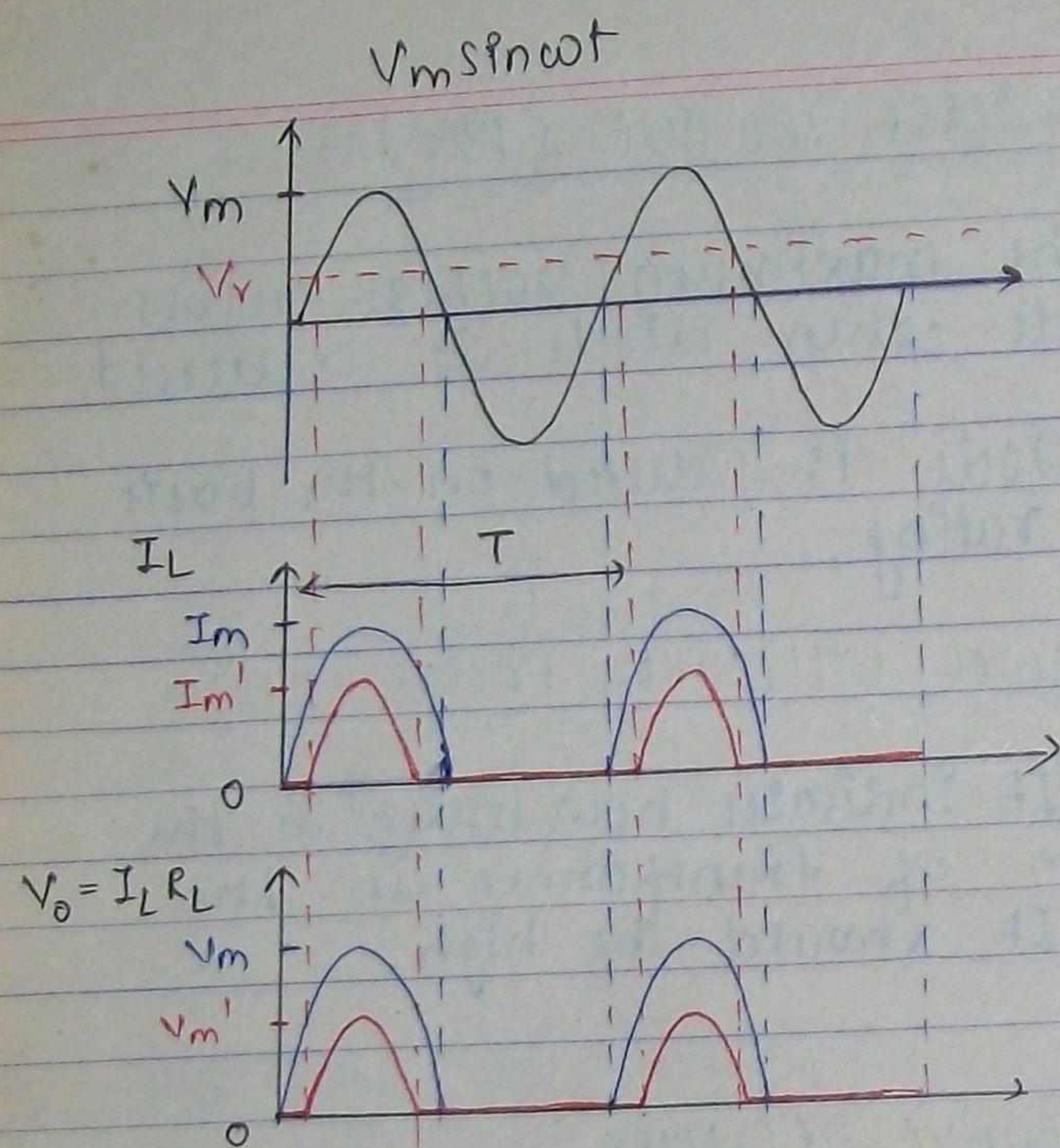
Diode is selected on the basis of PIV rating.

⑧ TRANSFORMER UTILISATION FACTOR (UTF) -

It indicates how much is the utilisation of transformer in the circuit. It should be high.

⇒ HALF WAVE RECTIFIER -





Initially circuit is ideal

(1) During +ve half

$D \rightarrow ON \rightarrow SC$

$$I_L = \frac{V_m \sin \omega t}{R_L} = I_m \sin \omega t$$

$$I_m = \frac{V_m}{R_L} = \text{max (peak) diode current.}$$

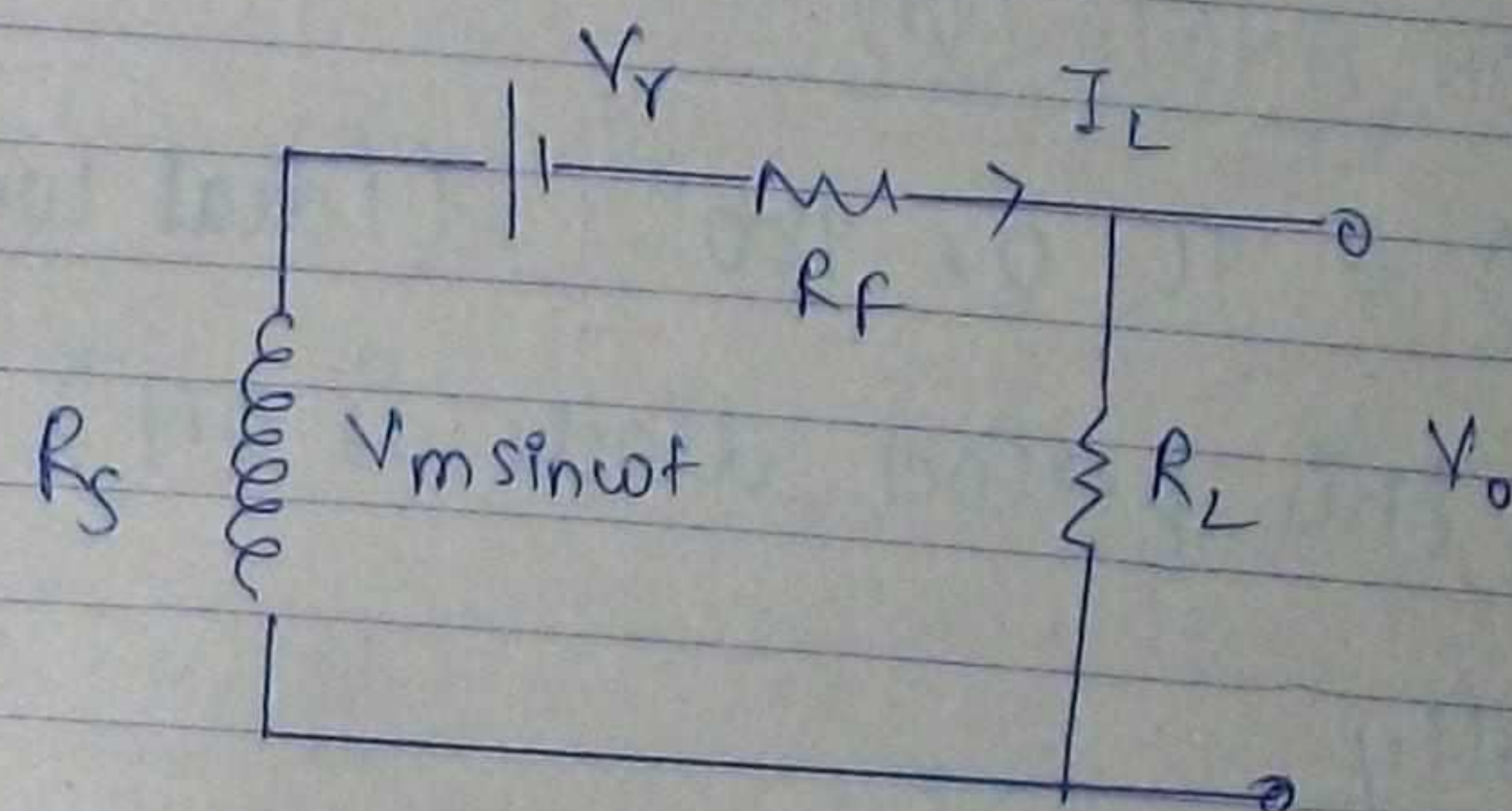
O/P freq = Ripple freq.

(2) During -ve half

$$I_L = 0$$

$$\therefore V_o = I_L R_L = V_m \sin \omega t$$

under practical condition -



R_s = resistance in secondary coil.

$$V_i < V_Y \quad D \rightarrow OFF$$

$$V_i > V_Y \quad D \rightarrow ON$$

$$I_L = \frac{V_m \sin \omega t - V_Y}{R_s + R_f + R_L}$$

$$I_m' = \frac{V_m - V_Y}{R_s + R_f + R_L} < I_m$$

$$V_o = I_L R_L$$

Output frequency = Ripple freq.

$$\text{Supply freq (F)} = f_r$$

$$\therefore \text{time period of o/p} = \text{time period of i/p.}$$

CONDUCTION ANGLE (ϕ) -

$$\phi = \pi \text{ or } 180^\circ \quad (\text{ideal cond}^n)$$

(During this period diode is ON)

practically

$$\phi = \pi - 2\theta$$

where,

$$\theta = \sin^{-1} \left(\frac{V_r}{V_m} \right)$$

AVERAGE VALUE -

$$I_{dc} = \frac{1}{2\pi} \int_0^\pi I_m \sin \omega t \, d(\omega t)$$

$$I_{dc} = \frac{I_m}{\pi}$$

Similarly,

$$V_{dc} = \frac{V_m}{\pi}$$

RMS VALUE -

$$I_{rms} = \left[\frac{1}{2\pi} \int_0^\pi I_m^2 \sin^2 \omega t \, d(\omega t) \right]^{1/2}$$

$$I_{rms} = \frac{I_m}{2}$$

Similarly,

$$V_{rms} = \frac{V_m}{2}$$

FORM FACTOR -

$$F = \frac{R_{ms}}{Dc} = \frac{V_m/2}{V_m/\pi} = \frac{\pi}{2}$$

$$F = 1.58 \text{ or } \pi/2$$

RIPPLE FACTOR -

$$\gamma = \sqrt{F^2 - 1}$$

$$\gamma = \sqrt{1.58^2 - 1}$$

$$\gamma = 1.21$$

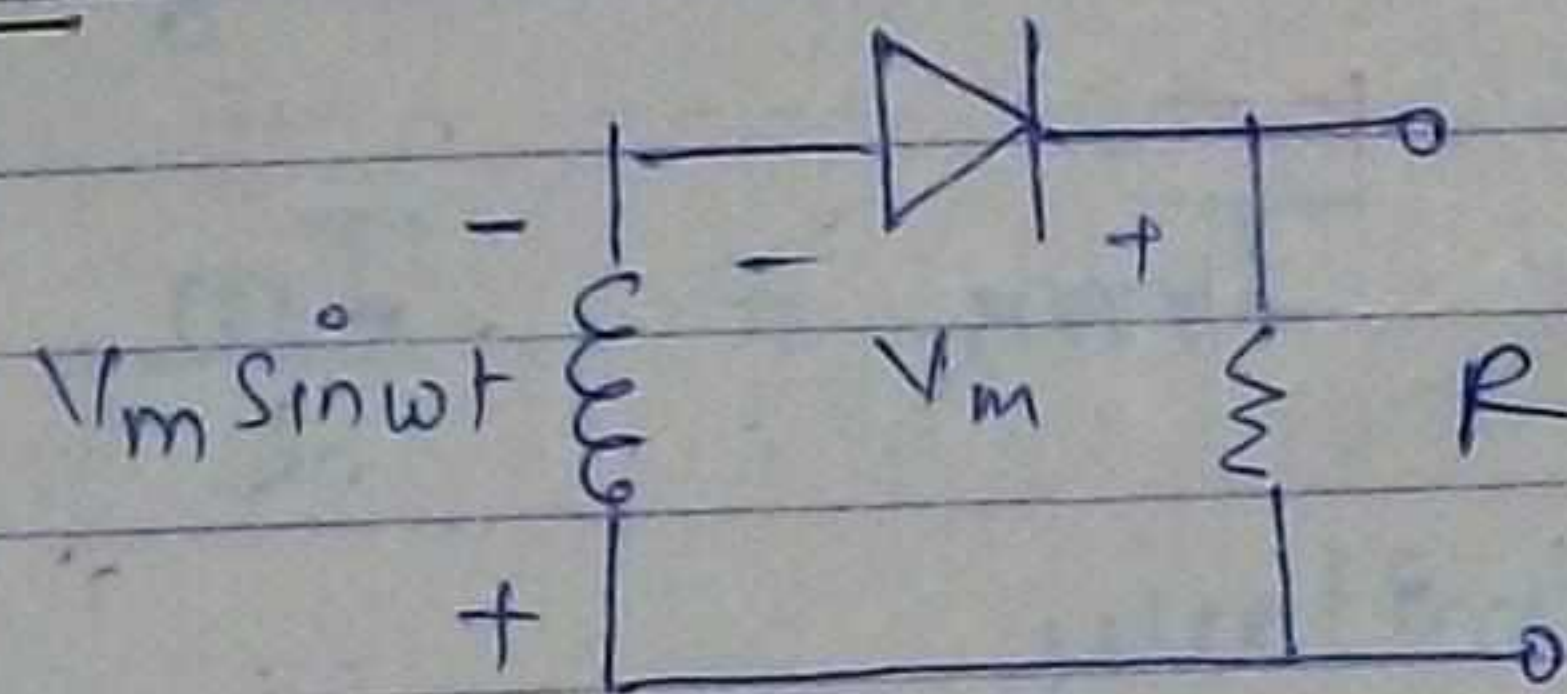
CREST FACTOR -

$$C = \frac{\text{Peak}}{\text{RMS}} = \frac{V_m}{V_m/2}$$

$$C = 2$$

PEAK INVERSE VOLTAGE -

$$\text{PIV} = V_m$$



RECTIFIER EFFICIENCY (η) -

$$\eta = \frac{\text{O/P DC POWER}}{\text{I/P AC POWER}} \times 100\%$$

$$\eta = \frac{I_{DC}^2 R_L}{I_{RMS}^2 [R_s + R_f + R_L]} \times 100\%$$

$$\eta = \frac{\left(\frac{I_m}{\pi}\right)^2 R_L}{\left(\frac{I_m}{2}\right)^2 (R_s + R_f + R_L)}$$

$$\eta = \frac{0.406}{\frac{R_s + R_f}{R_L} + 1} \times 100\%$$

for $R_L \gg R_s + R_f$

$$\eta_{\text{max}} = 40.6\%$$

⊛ If $\eta = 40\%$ then remaining 60% power is in form of supple.

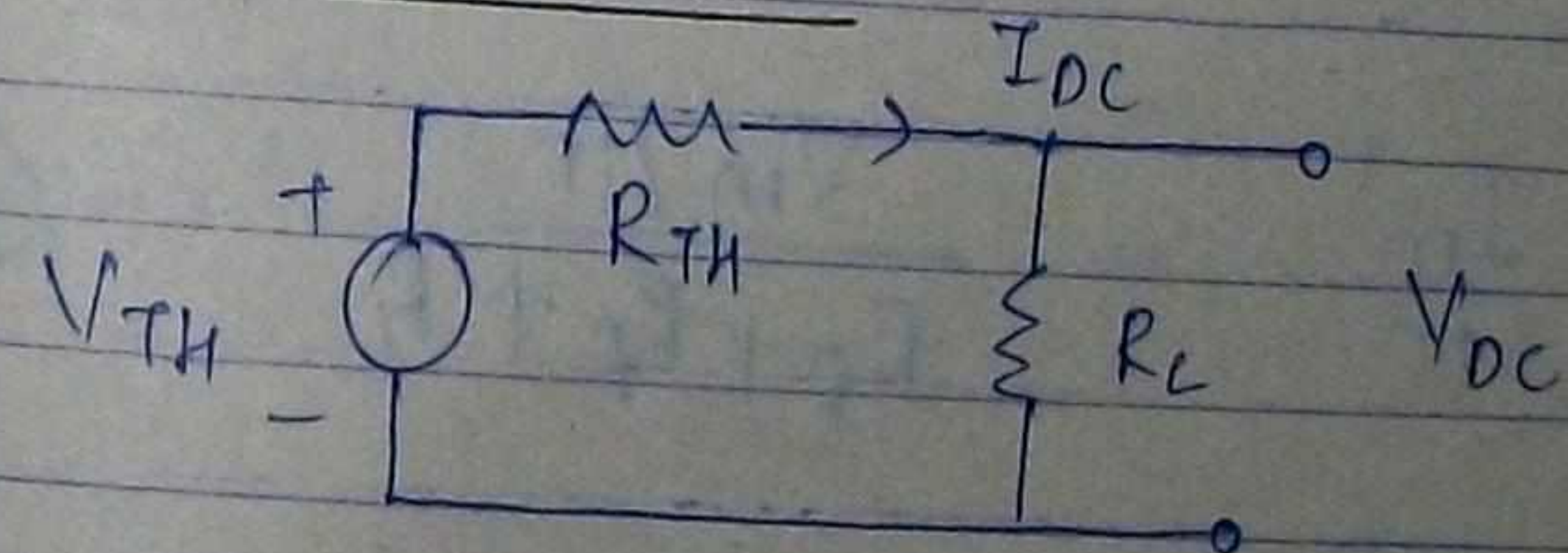
TRANSFORMER UTILIZATION FACTOR -

$$\text{TUF} = 0.286$$

(drawback)

means transformer is under utilization.

⇒ THEVENIN'S EQUIVALENT OF HALF WAVE RECTIFIER - (PRACTICALLY CONDⁿ)



$$I_{DC} = \frac{V_{TH}}{R_{TH} + R_L} \quad - \textcircled{1}$$

also,

$$I_L = \frac{V_m \sin \omega t - V_r}{R_s + R_f + R_L}$$

$$I_{dc} = \frac{1}{2\pi} \int_0^{\pi-\theta} I_L d(\omega t) \quad - (2)$$

Comparing (1) and (2) V_{TH} and R_{TH} can be determined.

let $V_r = 0$, $\theta = \sin^{-1}\left(\frac{V_r}{V_m}\right) = 0$

$$I_L = \frac{V_m \sin \omega t}{R_s + R_f + R_L} = I_m' \sin \omega t$$

$$I_m' = \frac{V_m}{R_s + R_f + R_L}$$

$$I_{dc} = \frac{1}{2\pi} \int_0^{\pi} I_m' \sin \omega t d(\omega t)$$

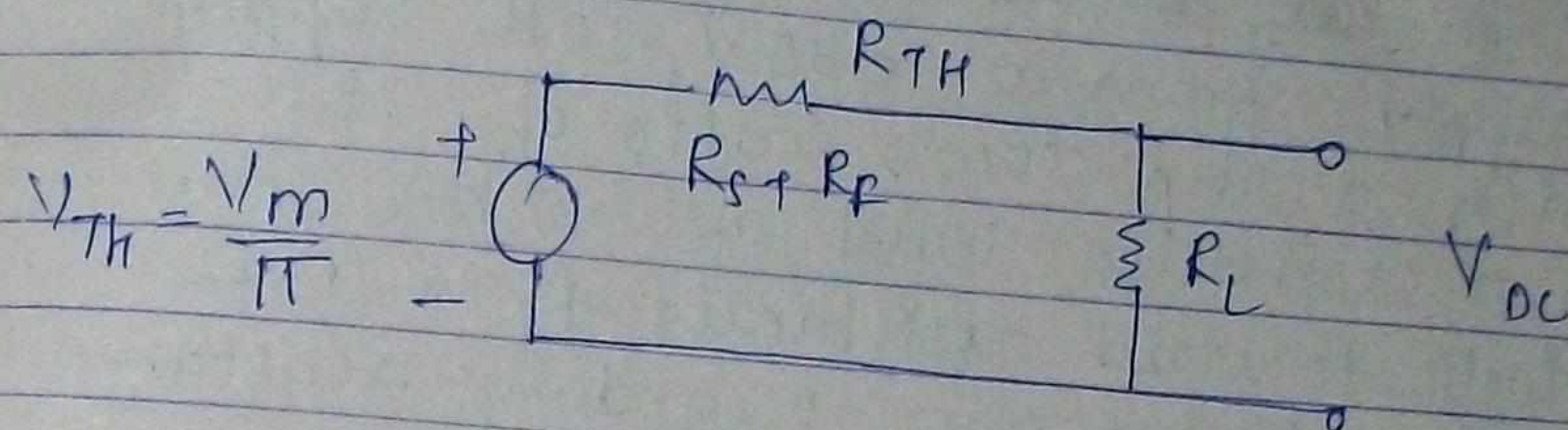
$$I_{dc} = \frac{I_m'}{\pi} = \frac{V_m/\pi}{R_s + R_f + R_L} \quad - (3)$$

from (1) and (3) :

$$V_{TH} = \frac{V_m}{\pi}$$

$$R_{TH} = R_s + R_f$$

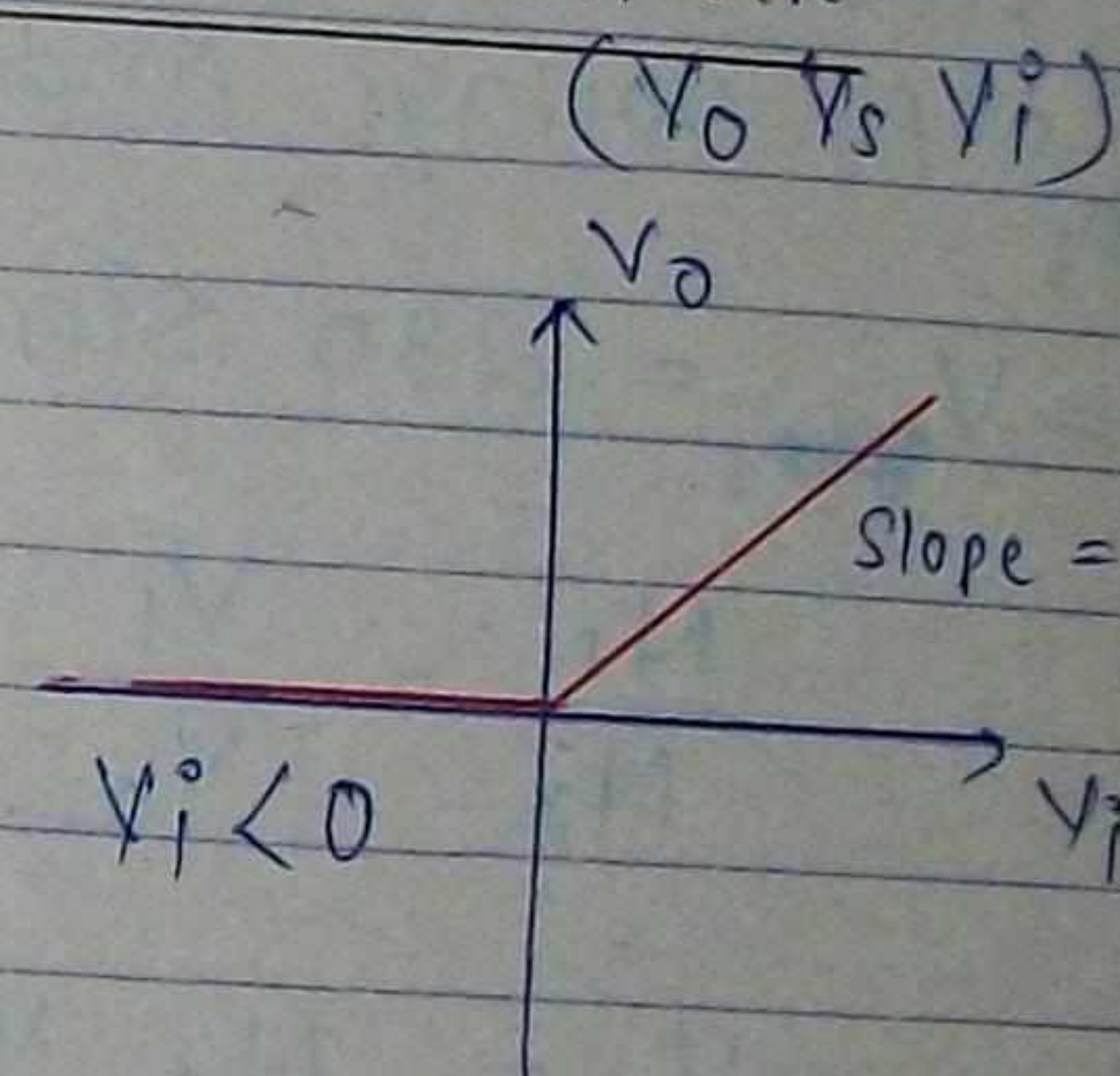
also from circuit :



⇒ TRANSFER CURVE FOR HW RECTIFIER -

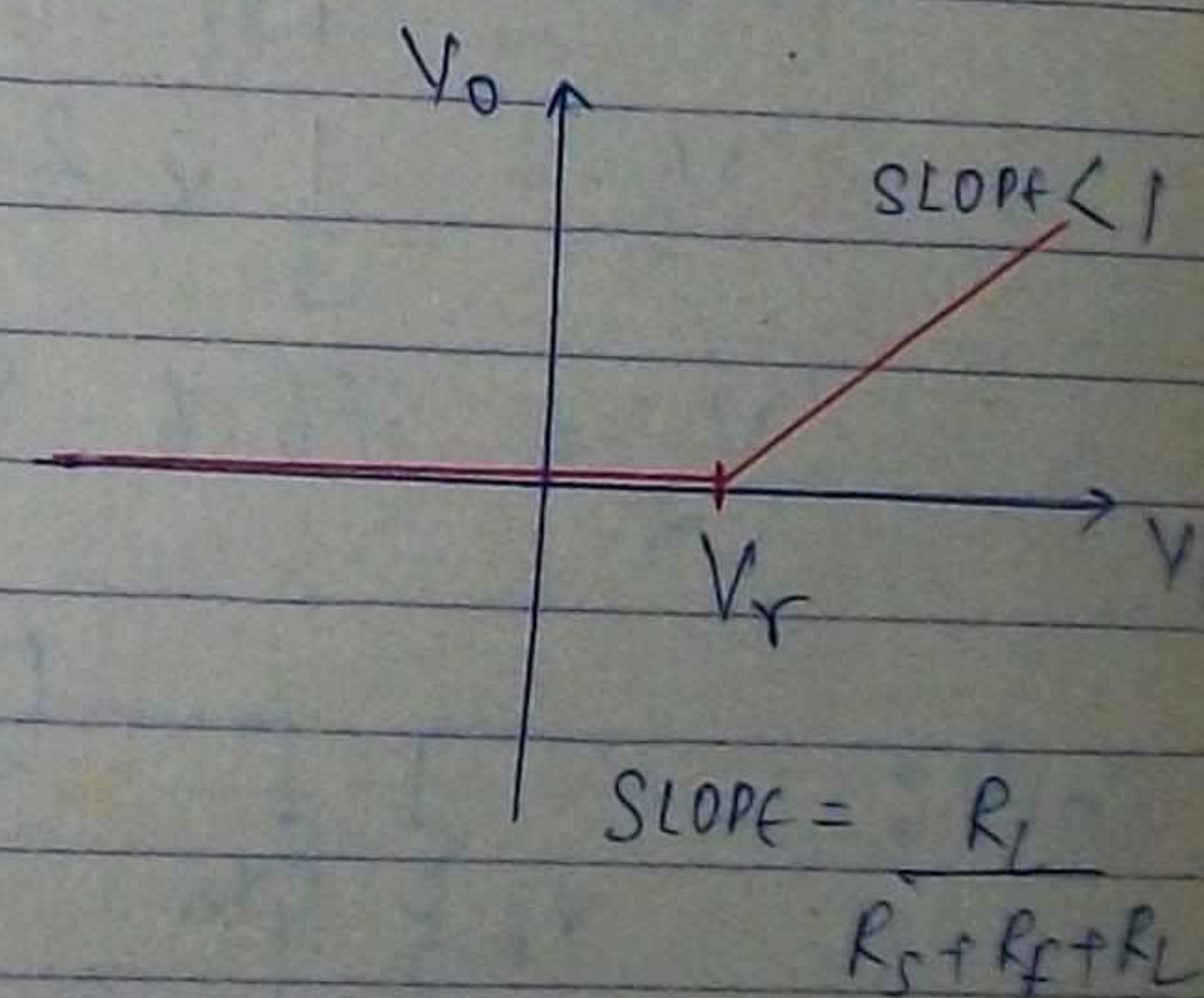
(1) IDEAL CIRCUIT -

V_i	D	V_o
$V_i < 0$	OFF	0
$V_i \geq 0$	ON	V_i



(2) PRACTICAL CIRCUIT -

V_i	D	V_o
$V_i < V_r$	OFF	0
$V_i \geq V_r$	ON	$V_i - V_r$



$$V' = I_L R_L = \frac{V_i - V_r}{R_s + R_f + R_L} \times R_L$$

$$V' = \frac{R_L}{R_s + R_f + R_L} V_i - \frac{V_r R_L}{R_s + R_f + R_L}, \quad y = mx + c$$

Ques. A Half wave rectifier circuit is supplied from a 230V, 50Hz supply with a step down ratio of 3:1 to a resistive load of 10k Ω . The diode forward resistance $R_f = 75\Omega$ while transformer secondary resistance R_s is 10 Ω . Calculate max., avg., rms value of current, DC value of o/p voltage and $\eta\%$.

Soln:

$$V_{rms} = V'_{rms} = 230 \sin(50 \times 2\pi t)$$

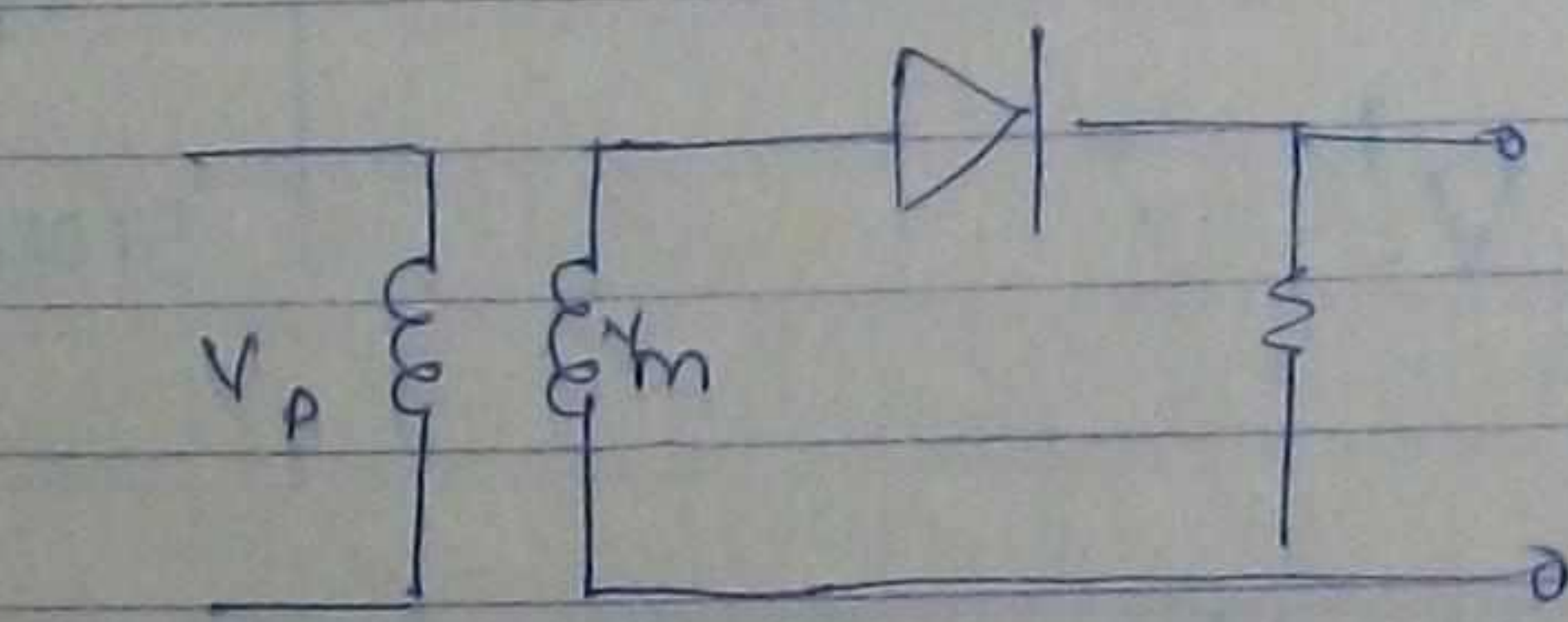
$$\frac{N_1}{N_2} = \frac{V_1}{V_2}$$

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

$$V_2 = \frac{1}{3} \times 230 \sin(2\pi 50t)$$

$$V_2 = 76.6 \sin(2\pi 50t)$$

Soln:



230V
50Hz

$$V_{rms} = 230$$

$$V_p = 230\sqrt{2}$$

$$V_m = \frac{230\sqrt{2}}{3} = 108.42$$

$$I_m = \frac{230\sqrt{2}/3}{R_s + R_f + R_L} = \underline{\underline{10.75 \text{ mA}}}$$

$$I_{rms} = \frac{I_m}{2} = \underline{\underline{5.37 \text{ mA}}}$$

$$I_{dc} = \frac{I_m}{\pi} = \underline{\underline{3.422 \text{ mA}}}$$

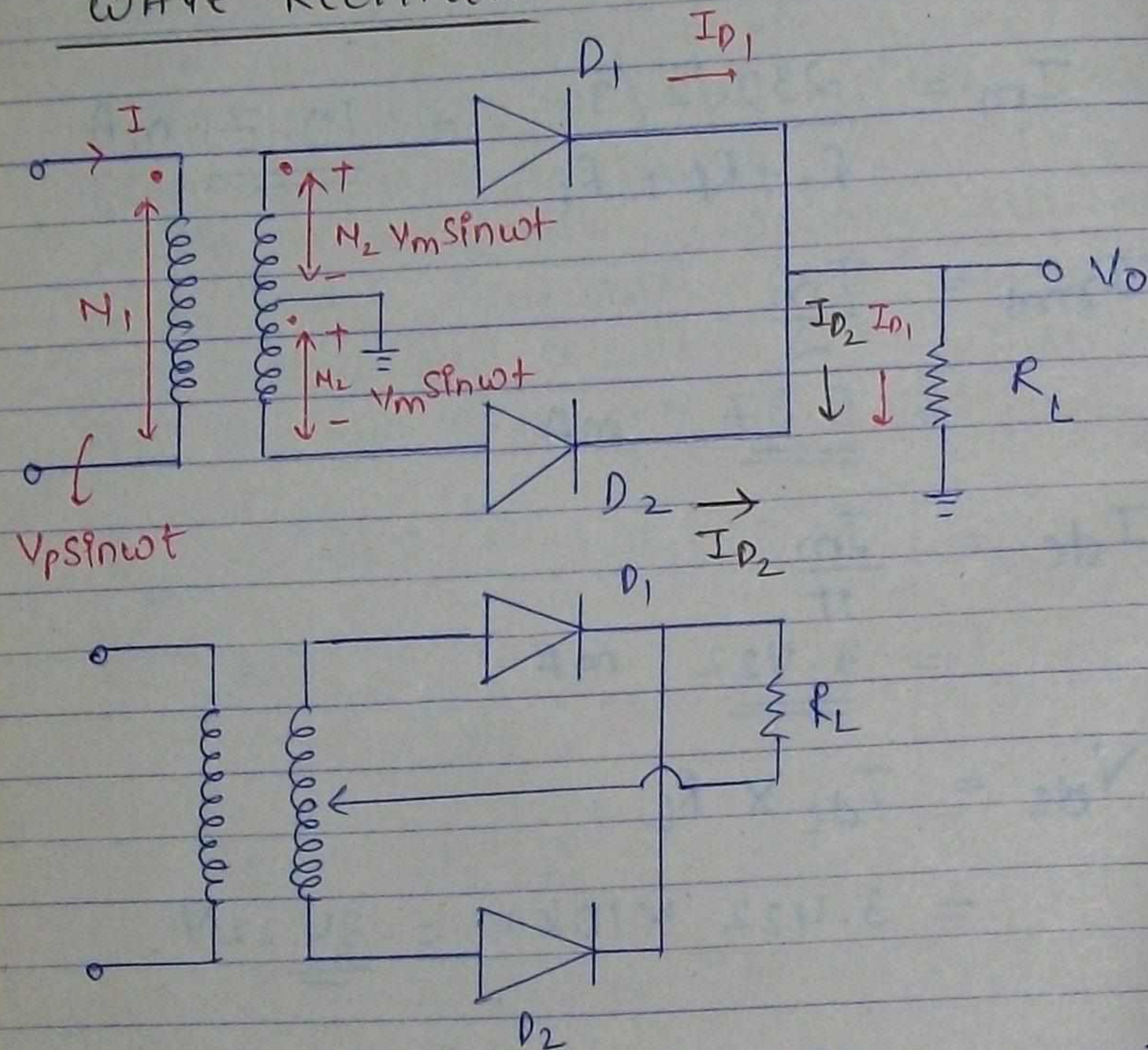
$$V_{dc} = I_{dc} \times R_L = 3.422 \times 10k\Omega = \underline{\underline{34.22 \text{ V}}}$$

$$\eta\% = \frac{I_{dc}^2 R_L}{I_{rms}^2 (R_s + R_L + R_f)} \times 100\%$$

$$\eta\% = \underline{\underline{40.19\%}}$$

24/08/11

② CENTRE-TAPPED TRANSFORMER TYPE FULL WAVE RECTIFIER -



$$\frac{V_p}{V_m} = \frac{N_1}{2N_2} \quad \left[\frac{N_1}{2N_2} = \frac{V_p}{2V_m} \right]$$

under ideal condition -

① During +ve half :
D₁ ON D₂ OFF

$$I_L = \frac{V_m \sin \omega t}{R_L} = I_m \sin \omega t$$

$$I_m = \frac{V_m}{R_L} = \text{max diode current}$$

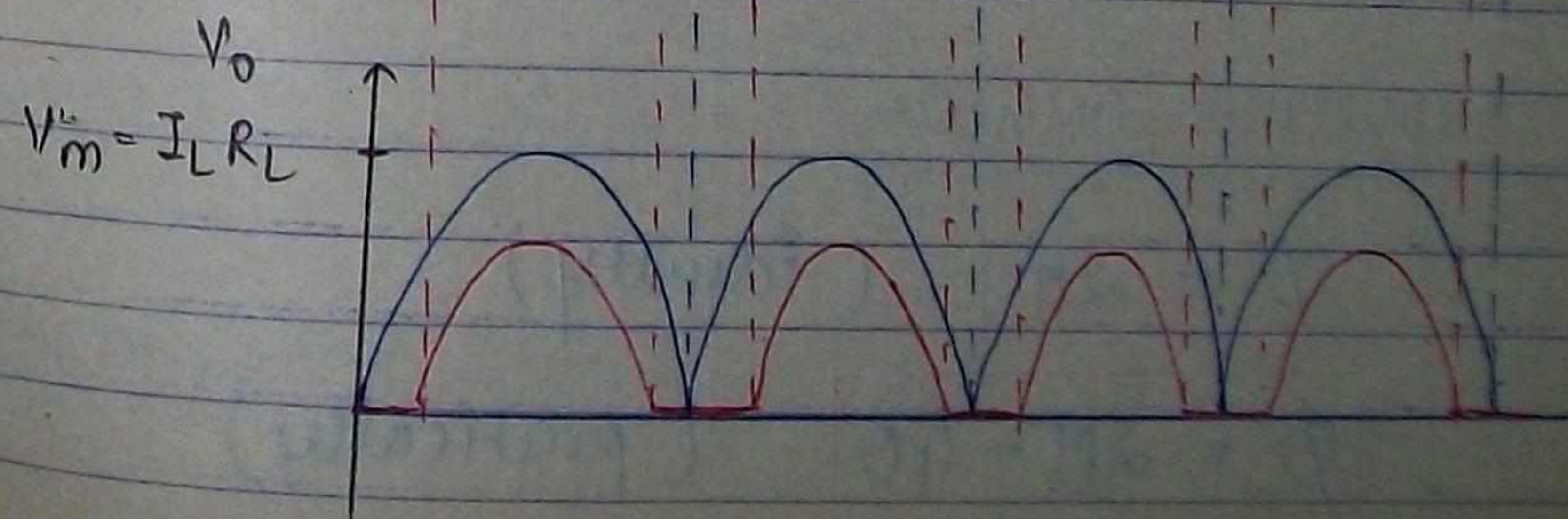
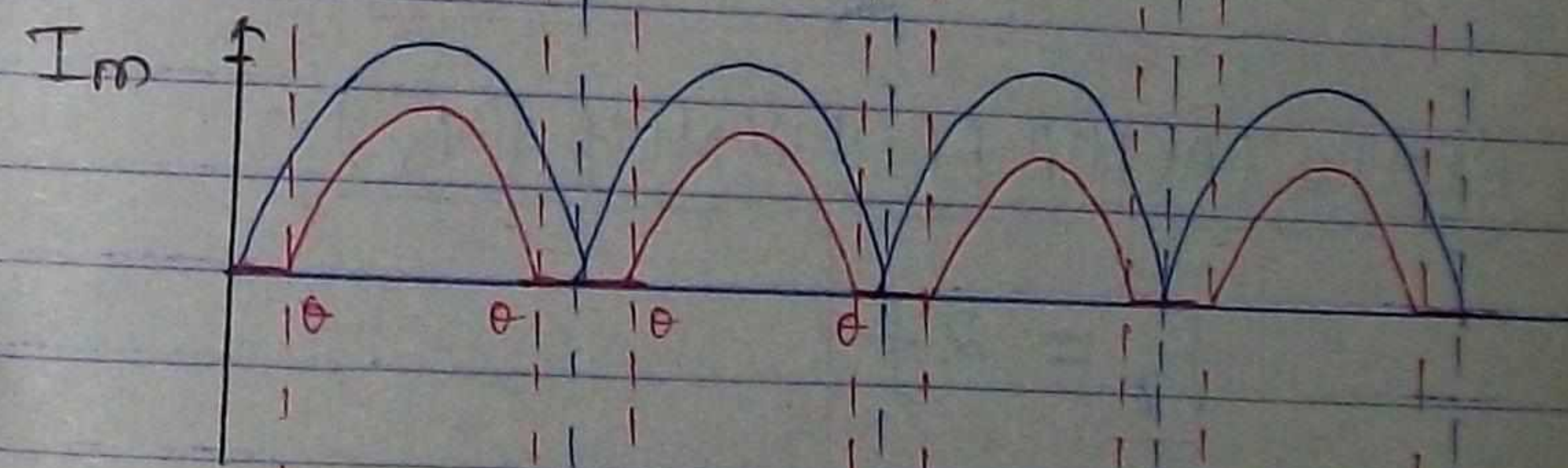
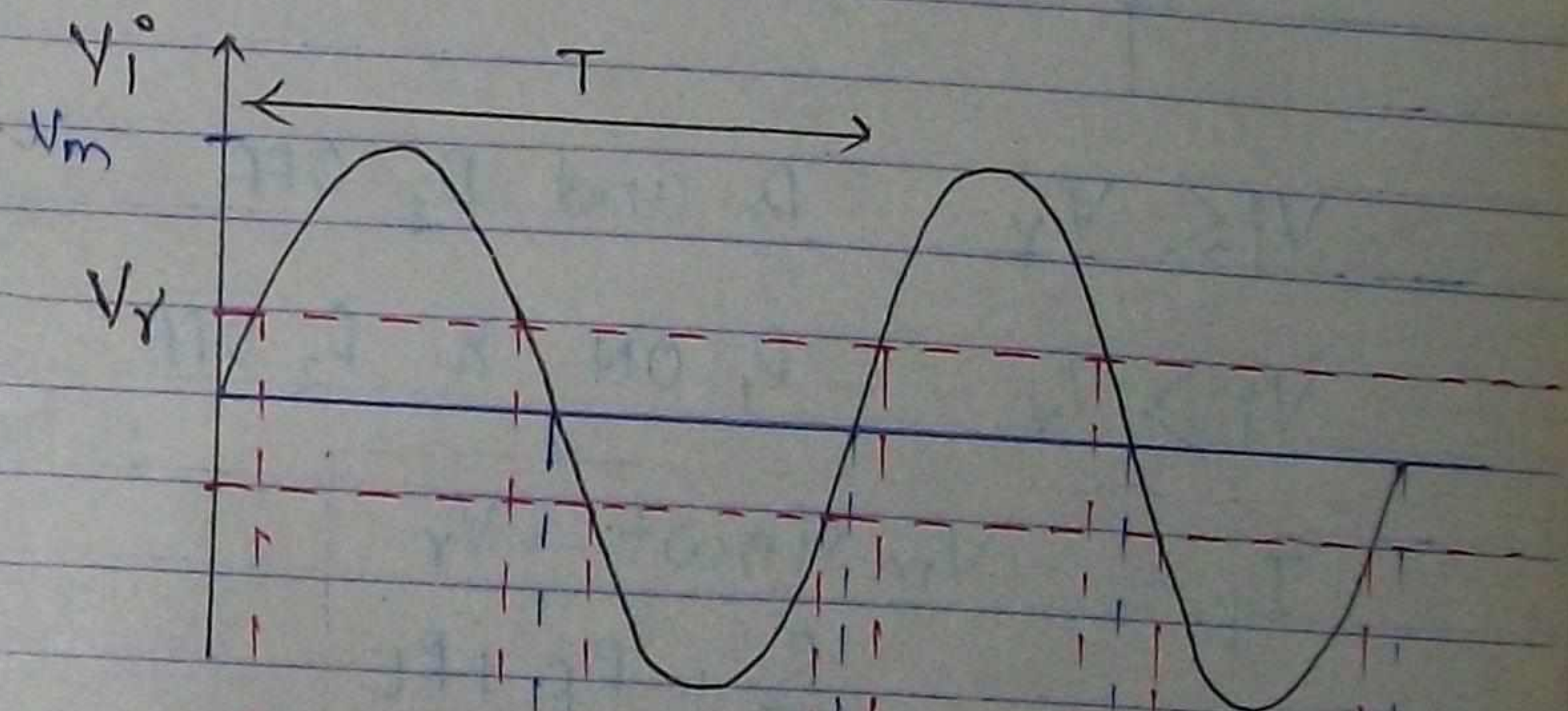
$$V_o = I_L R_L$$

② During -ve half :

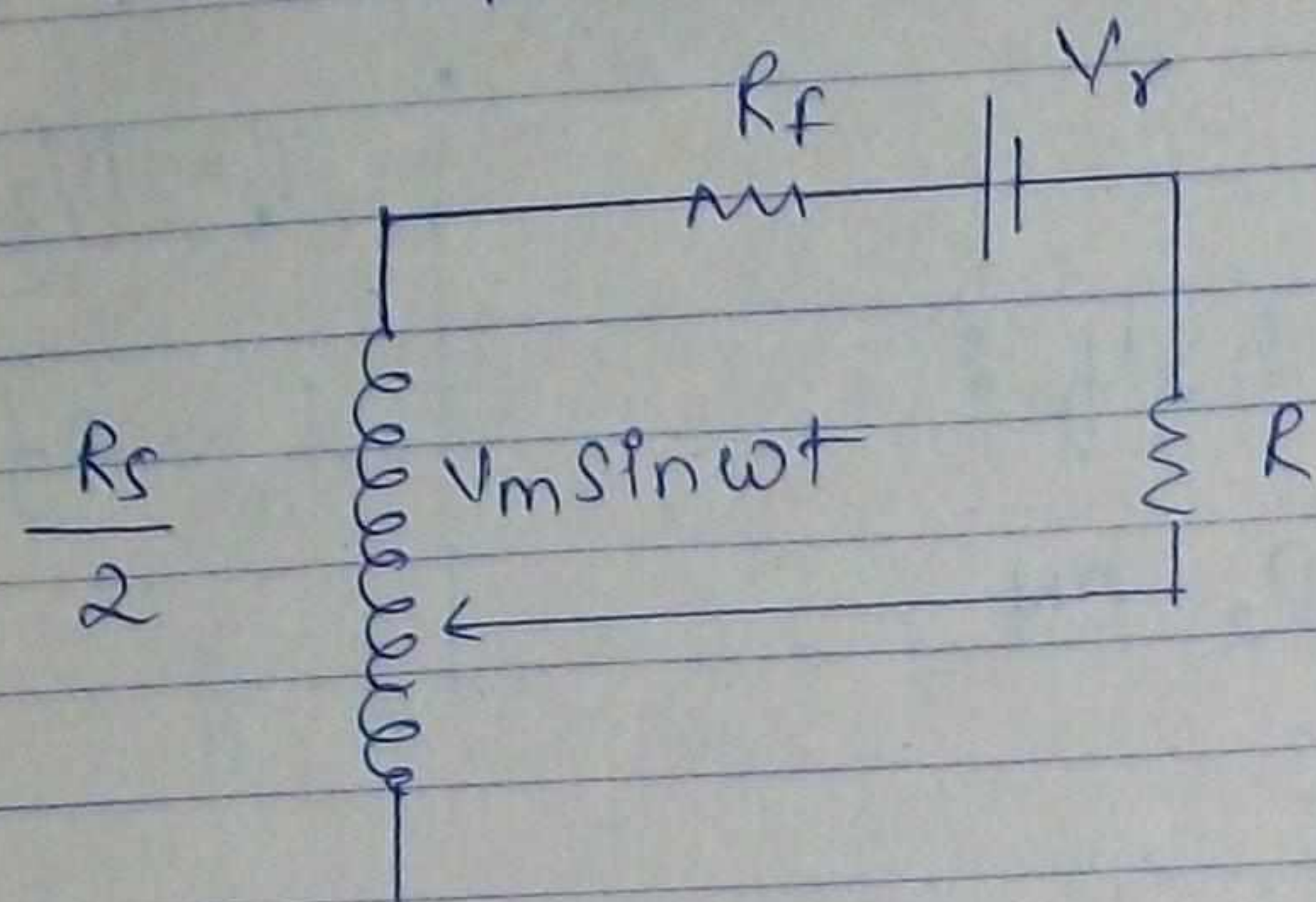
D₁ OFF D₂ ON

$$T_o = T/2$$

$$f_o = 2f_i$$



under practical condition:



$V_i \leq V_r$ D_1 and D_2 OFF

$V_i \geq V_r$ D_1 ON & D_2 OFF

$$I_L = \frac{V_m \sin \omega t - V_r}{\frac{R_s}{2} + R_f + R_L}$$

RIPPLE FACTOR FREQUENCY (f_r) -

$$f_r = 2f$$

CONDUCTION ANGLE -

$$\phi = 2\pi \text{ (ideally)}$$

$$\phi = 2\pi - 4\theta \text{ (practically)}$$

where, $\theta = \tan^{-1}$

$$\theta = \sin^{-1} \left(\frac{V_r}{V_m} \right)$$

AVERAGE VALUE -

$$I_{DC} = \frac{1}{\pi} \int_0^{\pi} I_m \sin \omega t \, d(\omega t)$$

$$I_{DC} = \frac{2 I_m}{\pi}$$

Similarly,

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

RMS VALUE -

$$I_{rms} = \left[\frac{1}{\pi} \int_0^{\pi} I_m^2 \sin^2 \omega t \, d(\omega t) \right]^{1/2}$$

$$I_{rms} = \frac{I_m}{\sqrt{2}}$$

Similarly,

$$V_{rms} = \frac{V_m}{\sqrt{2}}$$

FORM FACTOR (F) -

$$F = \frac{V_{rms}}{V_{dc}}$$

$$f = \frac{V_m / \sqrt{2}}{V_m / \pi} = 1.11$$

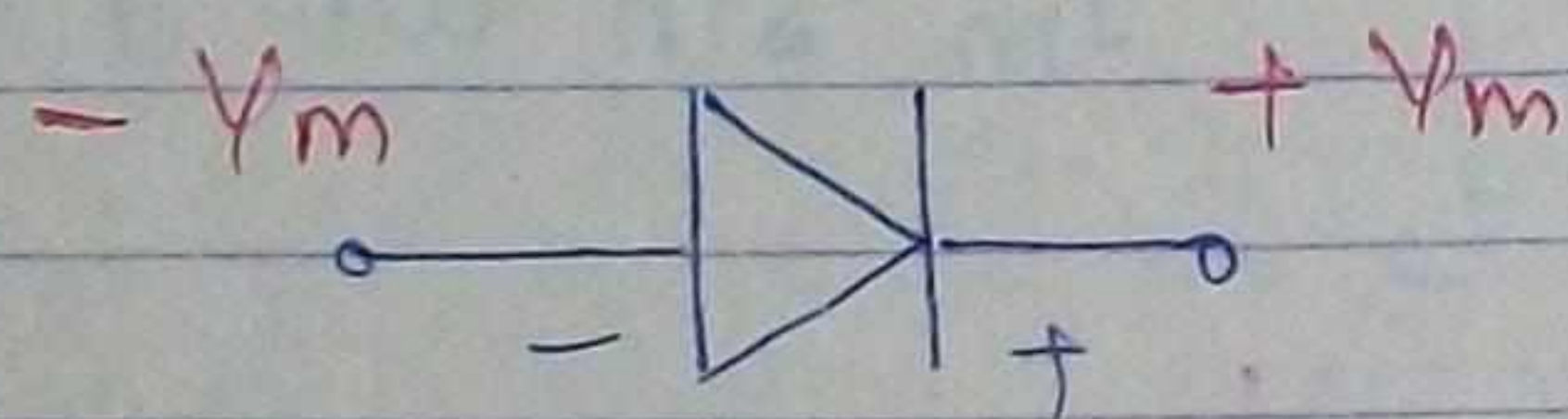
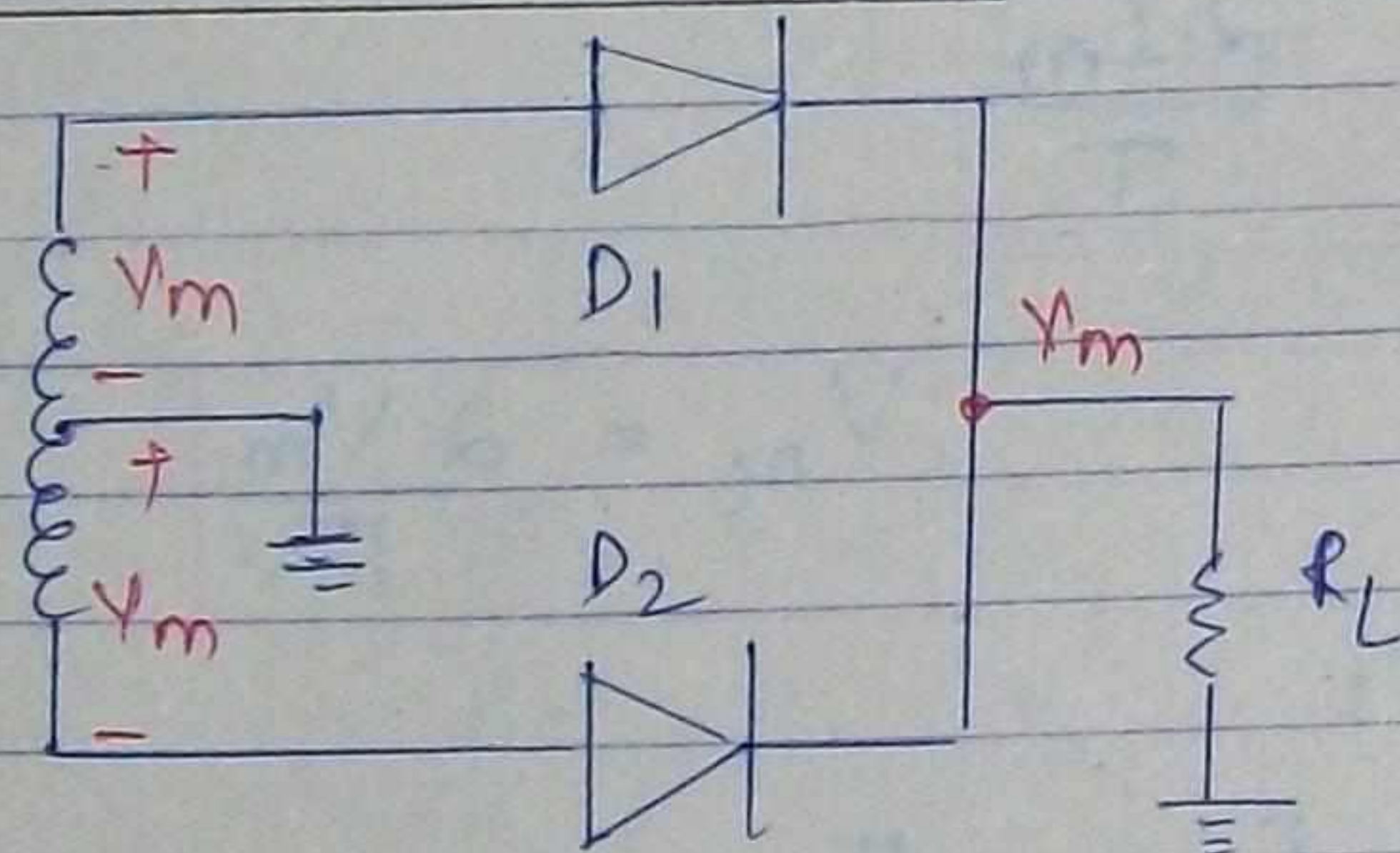
$$F = 1.11$$

RIPPLE FACTOR (γ) -

$$\gamma = \sqrt{F^2 - 1} = \sqrt{(1.11)^2 - 1}$$

$$\gamma = 0.48$$

PEAK INVERSE VOLTAGE (PIV) -

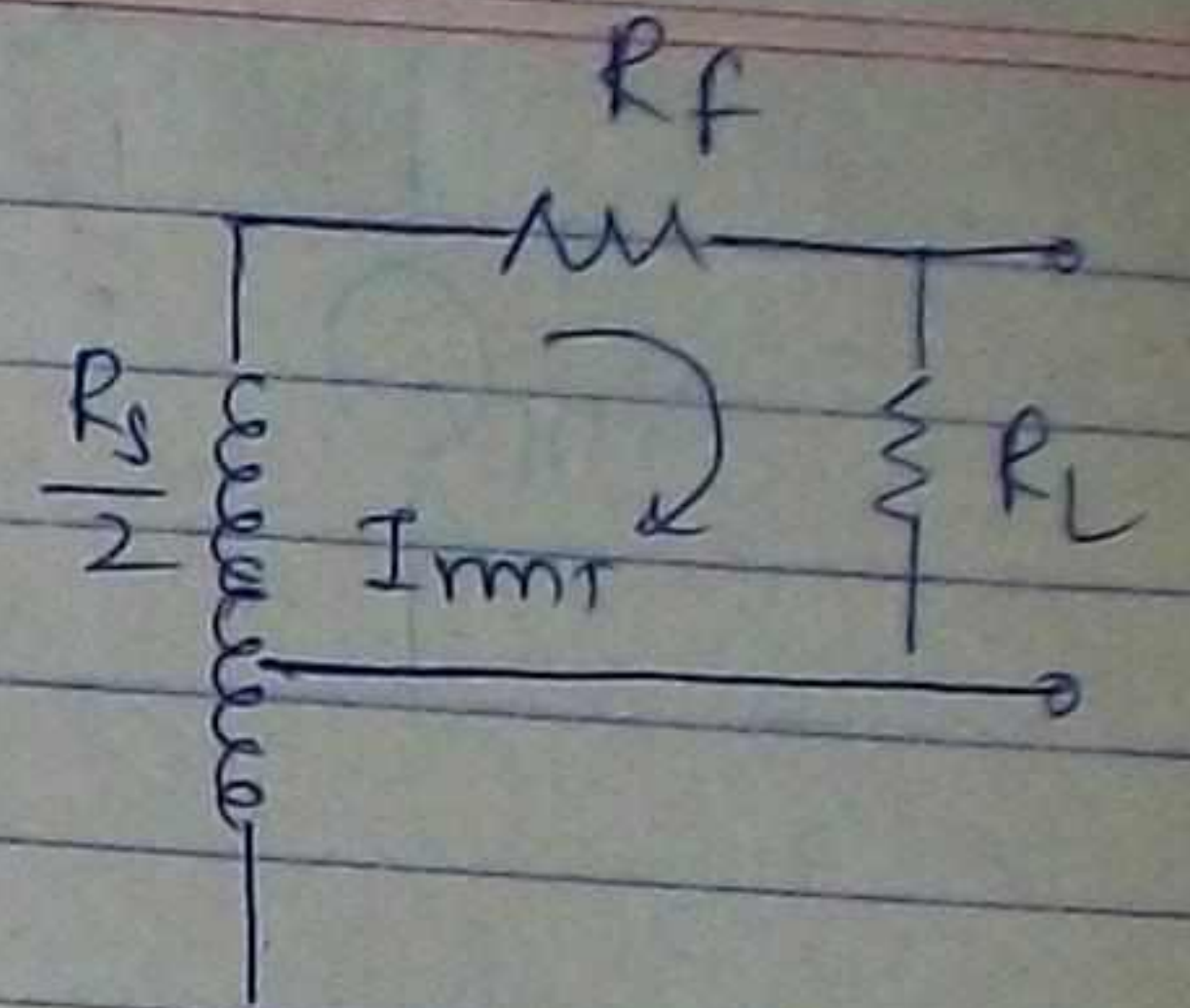


$$PIV = 2V_m$$

RECTIFIER EFFICIENCY (η) -

$$\eta = \frac{\text{D/c output power}}{\text{Ac i/p power}} \times 100\%$$

$$\eta = \frac{I_{dc}^2 \times R_L}{I_{rms}^2 \left[\frac{R_s}{2} + R_f + R_L \right]}$$



$$\eta = \frac{\left(\frac{2I_m}{\pi} \right)^2 \times R_L}{\left(\frac{I_m}{\sqrt{2}} \right)^2 \left(\frac{R_s}{2} + R_f + R_L \right)}$$

$$\eta = \frac{8}{\pi^2} \times \frac{1}{\frac{R_s/2 + R_f}{R_L} + 1} \times 100\%$$

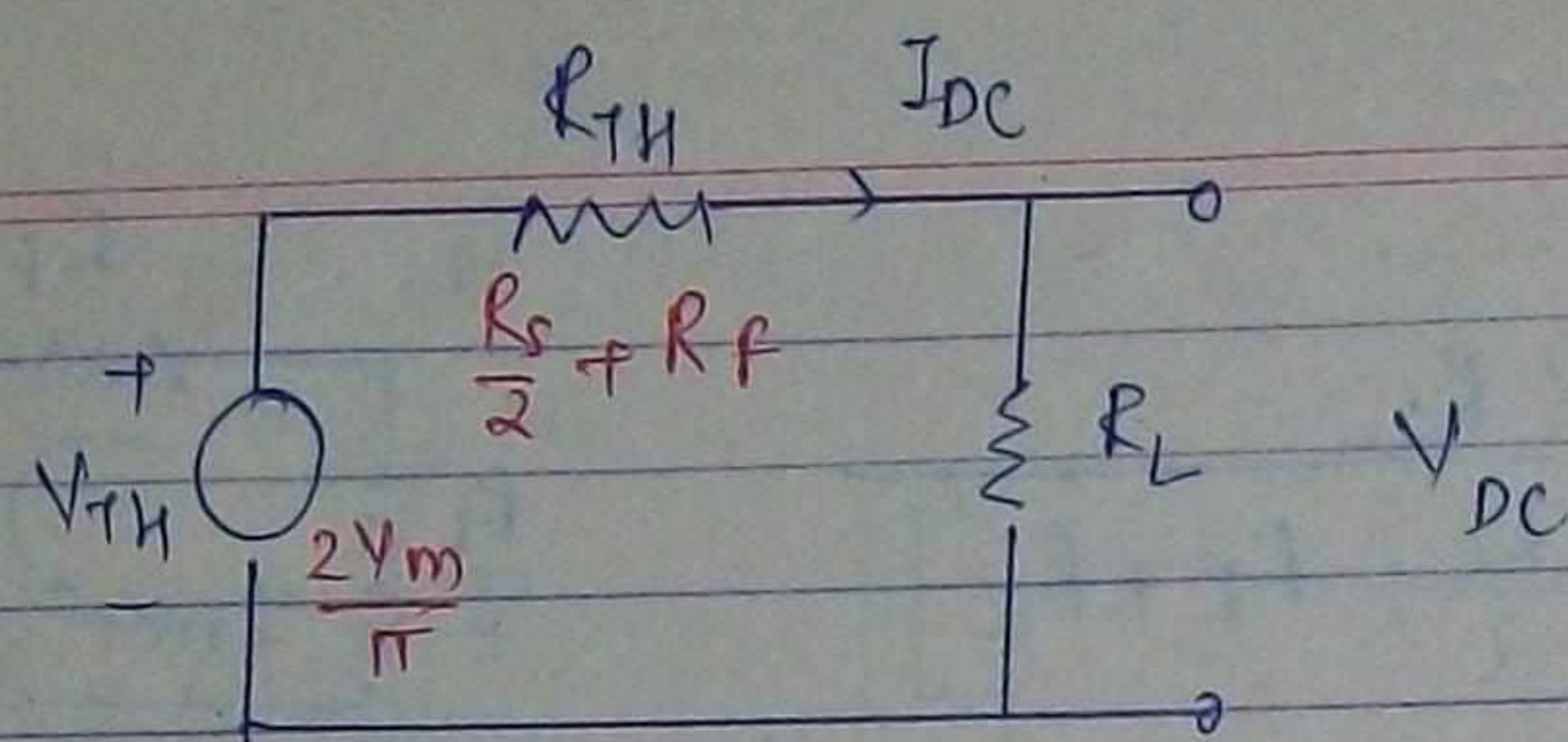
$$\eta_{max} = 81.2\%$$

when, $\frac{R_s}{2} + R_f \ll R_L$

TRANSFORMER UTILIZATION FACTOR -

$$TUF = 0.693$$

$\Rightarrow$ THEVENIN'S EQUIVALENT OF CENTRE TAP FULL WAVE RECTIFIER -



$$I_{DC} = \frac{V_{TH} \times 1}{R_{TH} + R_L}$$

$$I_{DC} = \frac{V_{TH}}{R_{TH} + R_L} \quad \text{--- (1)}$$

$$I_L = \frac{V_m \sin \omega t - V_r}{\frac{R_s}{2} + R_f + R_L}, \quad \theta = \sin^{-1} \left(\frac{V_r}{V_m} \right)$$

$$I_{DC} = \frac{1}{\pi} \int_0^{\pi-\theta} I_L d(\omega t)$$

let $V_r = 0$ $\theta = 0$

$$I_L = \frac{V_m \sin \omega t}{\frac{R_s}{2} + R_f + R_L} = I_m' \sin \omega t$$

$$I_m' = \frac{V_m}{\frac{R_s}{2} + R_f + R_L}$$

$$I_{DC} = \frac{1}{\pi} \int_0^{\pi} I_m' \sin \omega t \cdot d(\omega t)$$

$$I_{DC} = \frac{2 I_m'}{\pi}$$

$$I_{DC} = \frac{2 V_m / \pi}{\frac{R_s}{2} + R_f + R_L} \quad \text{--- (2)}$$

Compare (1) and (2):

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

$$R_{TH} = \frac{R_s}{2} + R_f$$

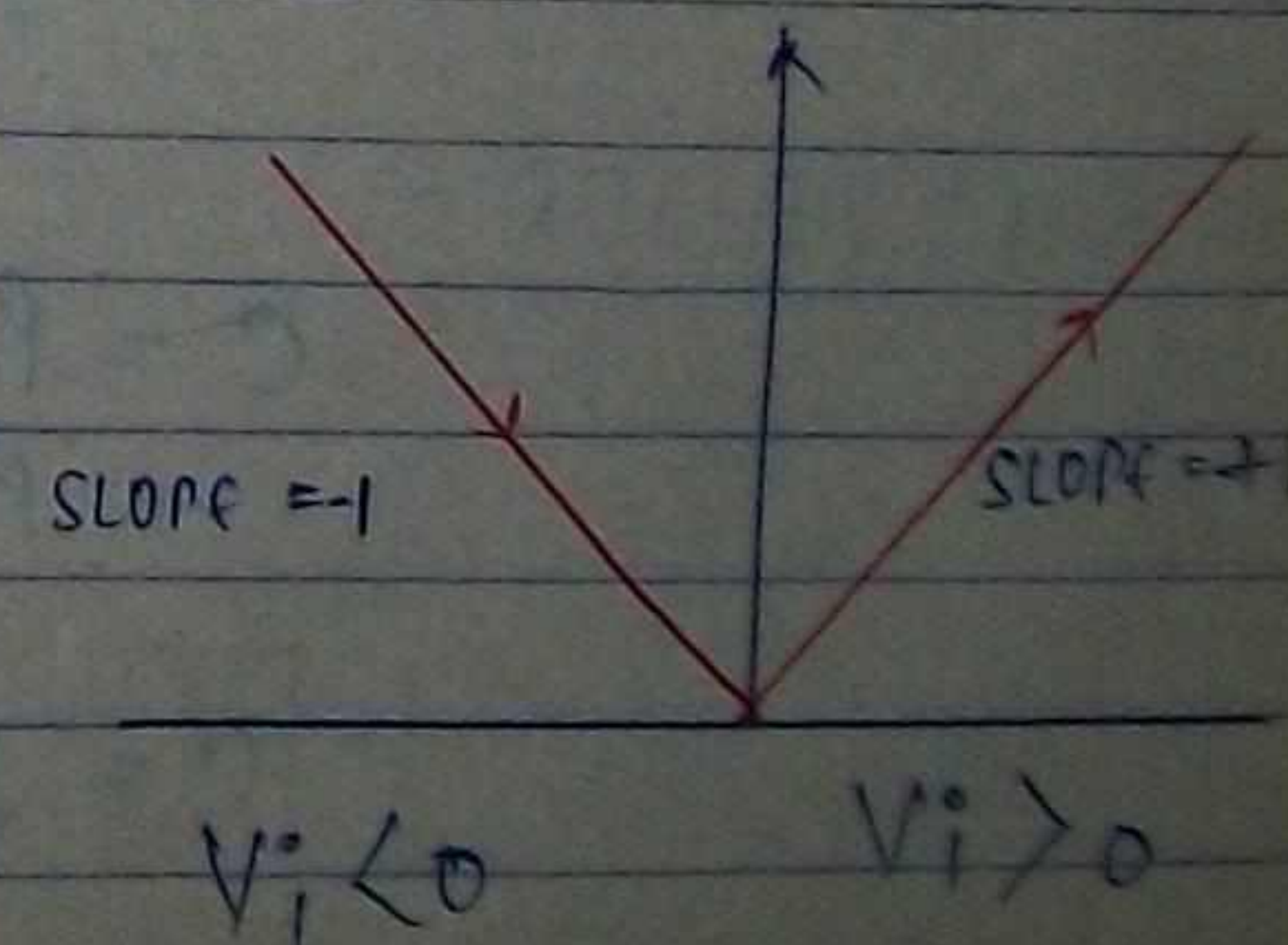
also from circuit:

$$\text{when } \frac{R_s}{2} + R_f = 0 \quad V_{DC} = \frac{2 V_m}{\pi}$$

⇒ TRANSFER CURVE OF FULL WAVE RECTIFIER -

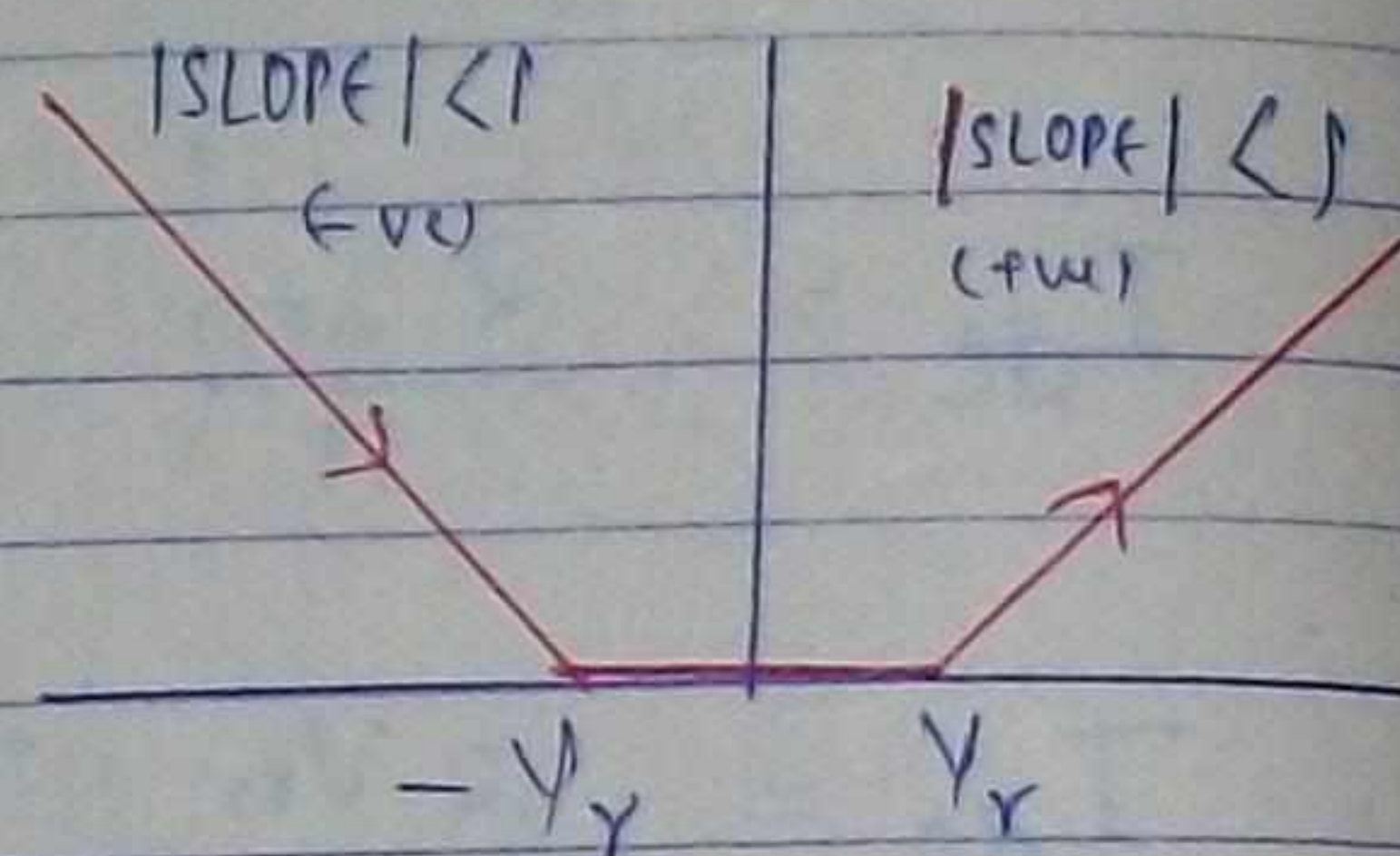
(1) FOR IDEAL CONDITION -

V_i	D_1	D_2	$V_o = V_i $
$V_i < 0$	OFF	ON	$-V_i$
$V_i > 0$	ON	OFF	V_i



② FOR PRACTICAL CONDITION -

V_i	D_1	D_2	V_o
$V_i < -V_r$	OFF	ON	V'
$-V_r \leq V_i \leq V_r$	OFF	OFF	0
$V_i \geq V_r$	ON	OFF	V''



$$V_o = V'' = I_L R_L = \frac{V_i - V_r}{\frac{R_s}{2} + R_f + R_L} \times R_L$$

$$V'' = \frac{R_L V_i}{\frac{R_s}{2} + R_f + R_L} - \frac{V_r R_L}{\frac{R_s}{2} + R_f + R_L}$$

$$y = mx + c$$

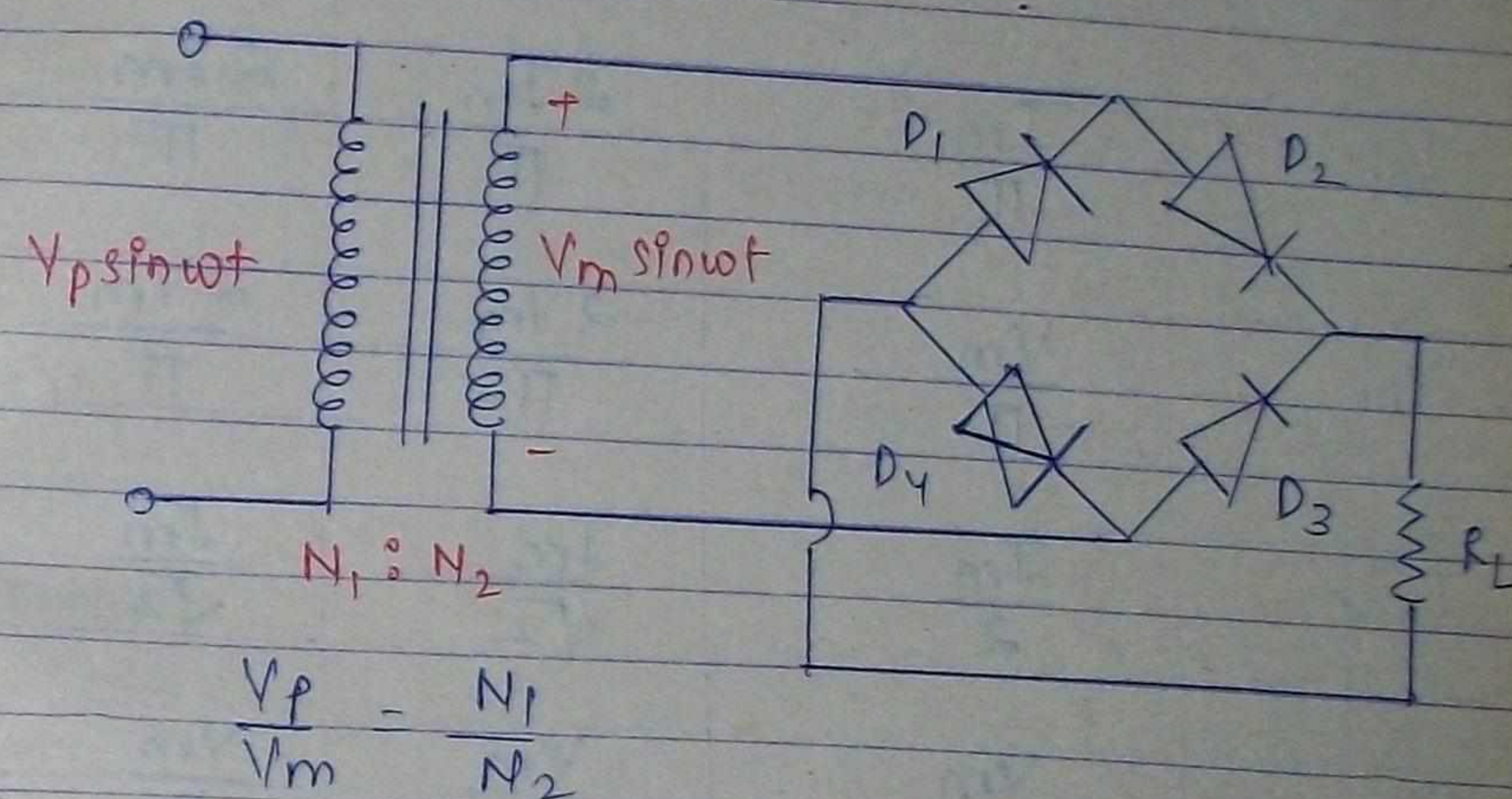
$$\text{SLOPE} = \frac{R_L}{\frac{R_s}{2} + R_f + R_L}$$

CREST FACTOR -

$$C = \frac{\text{Peak}}{\text{RMS}} = \frac{V_m}{V_m/\sqrt{2}}$$

$$C = \sqrt{2}$$

⇒ BRIDGE RECTIFIER -

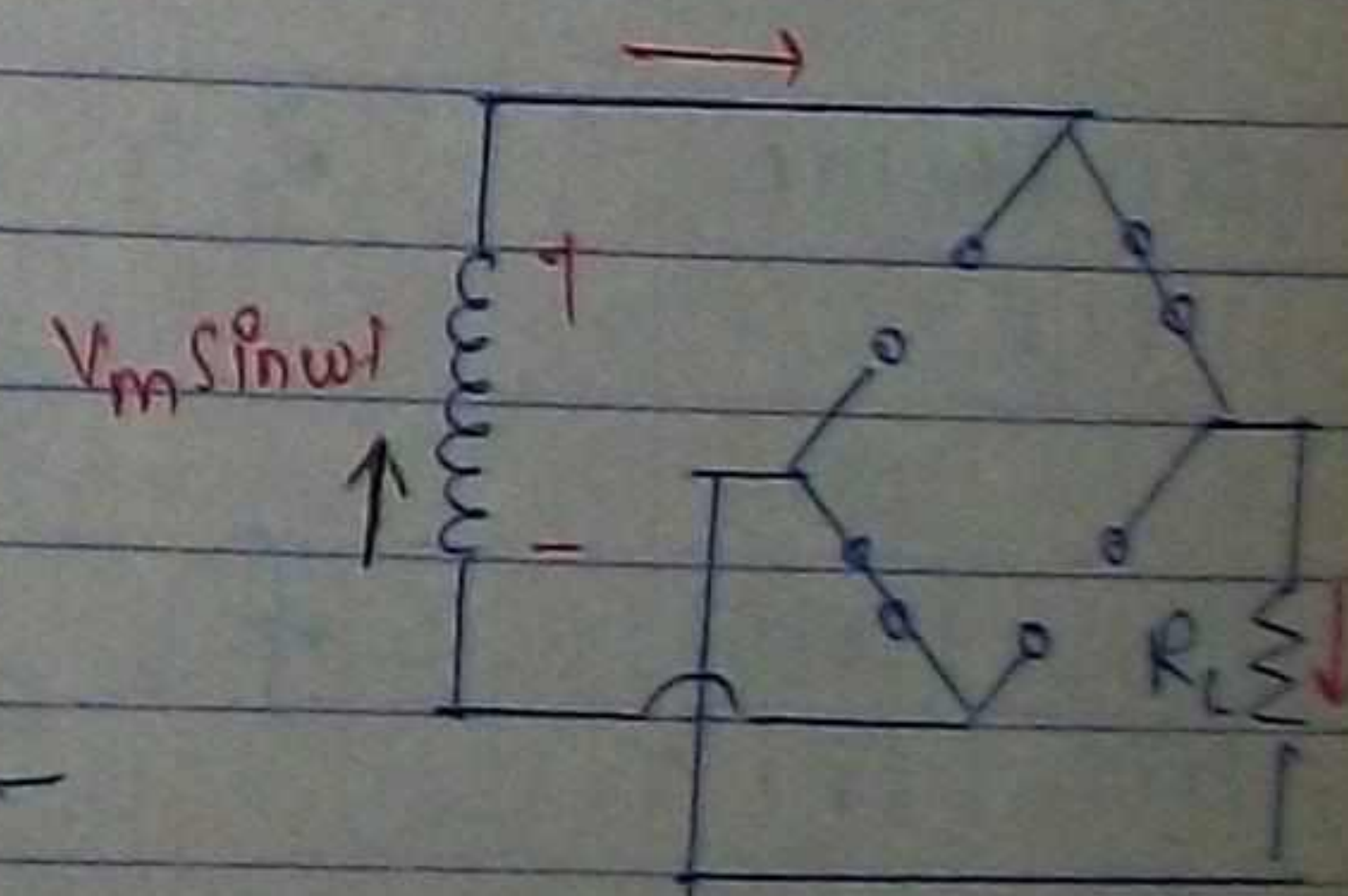


under ideal conditions :

① During positive half-

D_2 and D_4 are ON

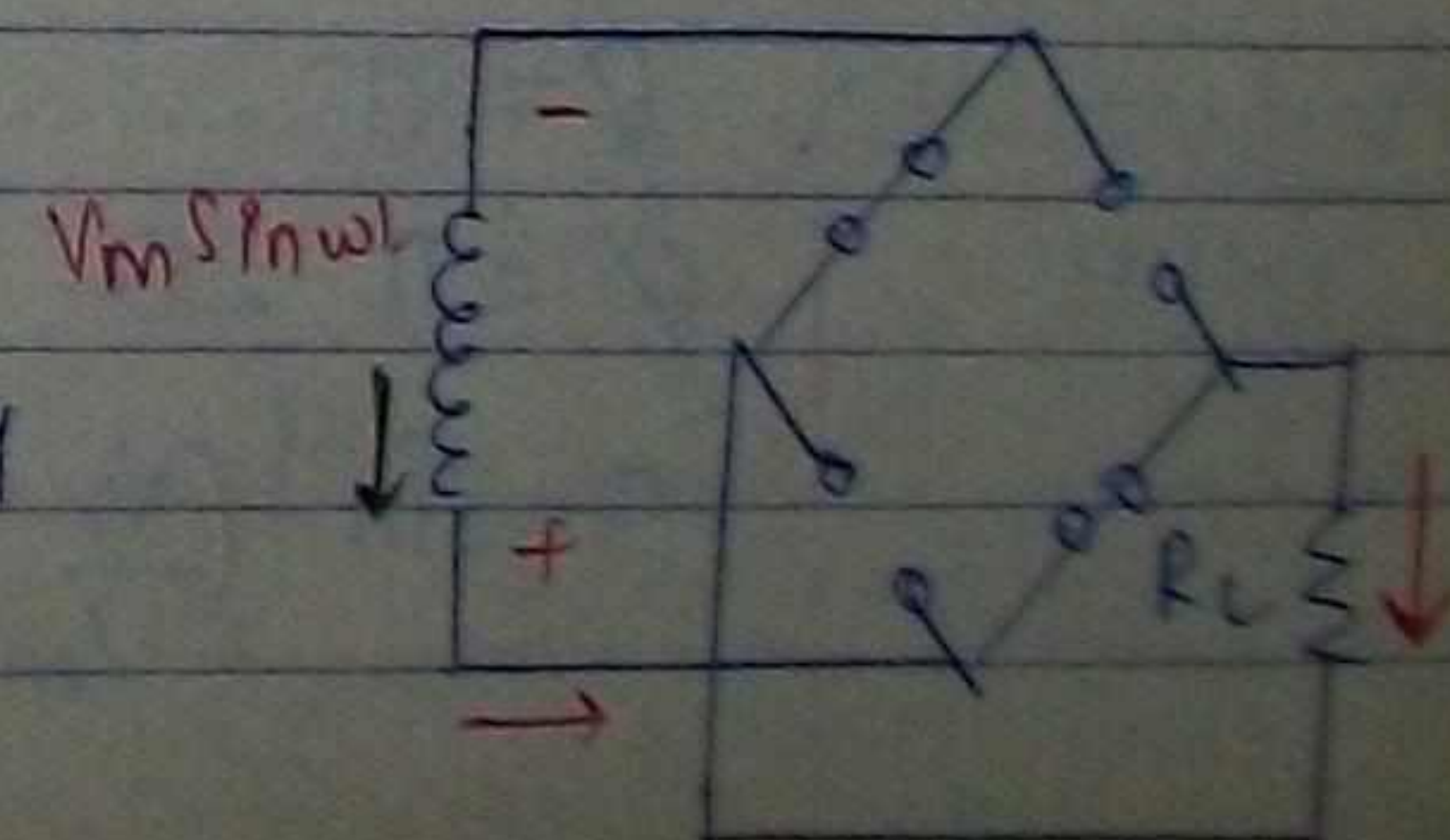
and D_1 & D_3 are OFF.



② During negative half -

D_2 and D_4 are OFF

and D_1 and D_3 are ON



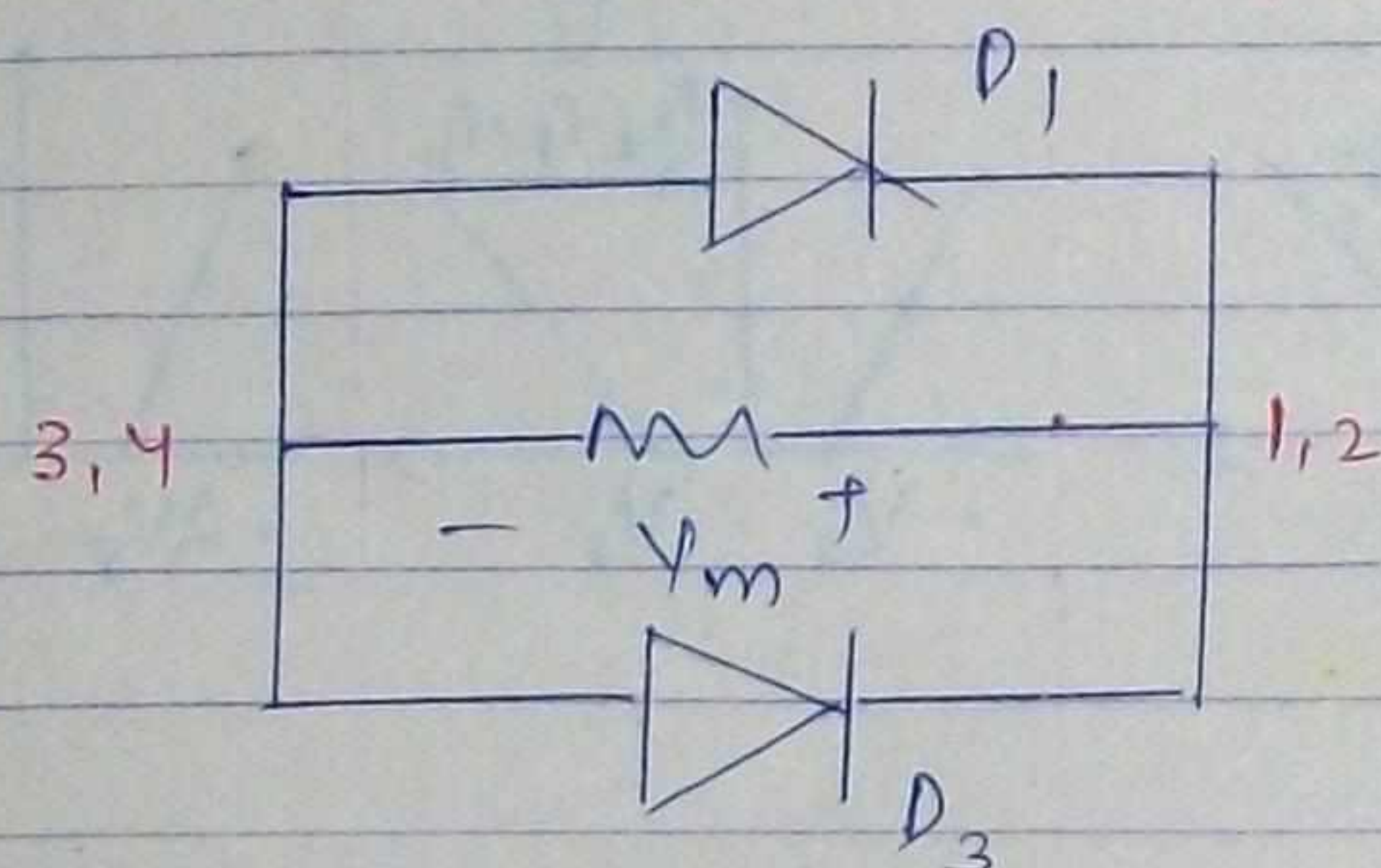
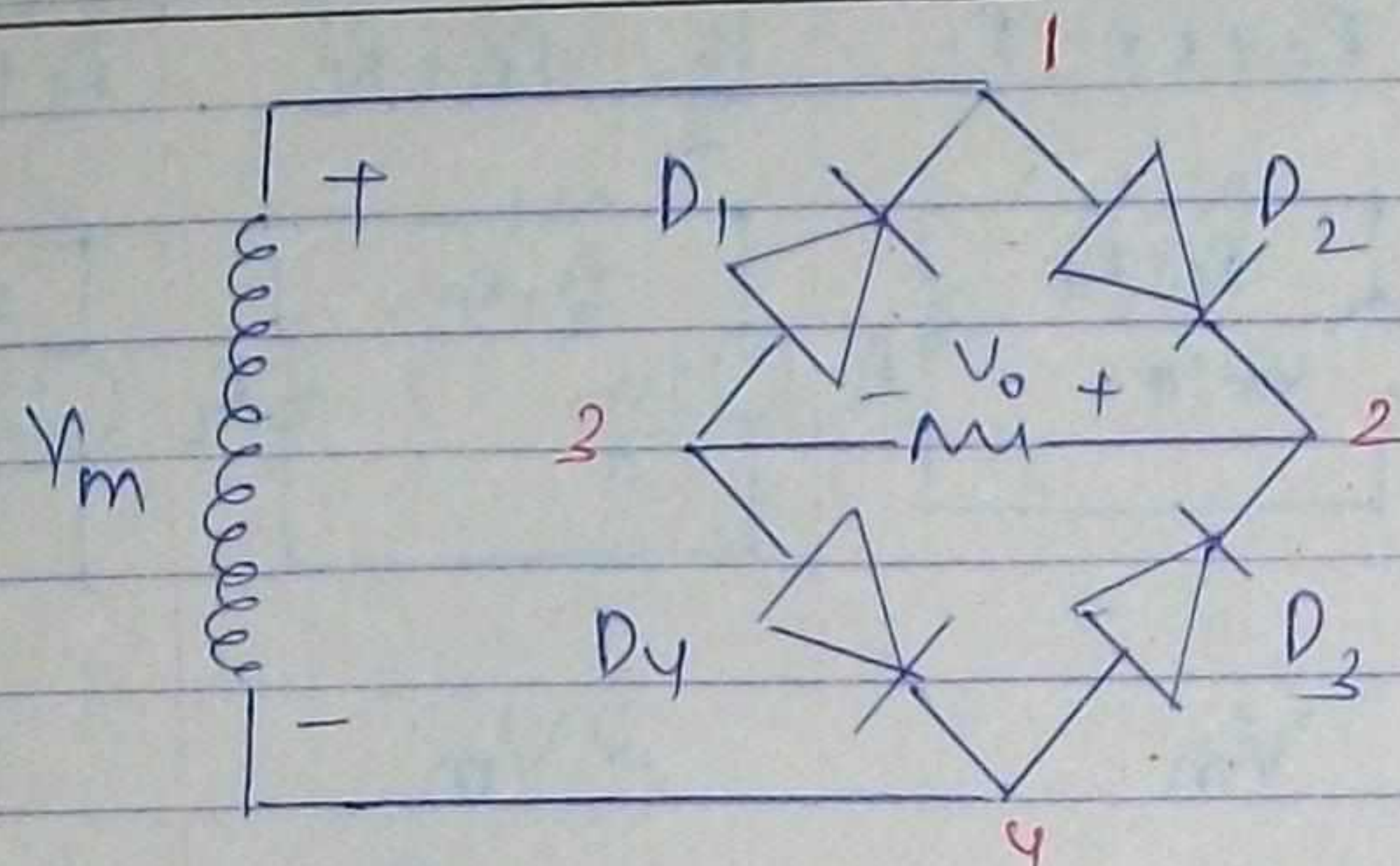
* Avg value of current across transformer is zero.

SPECIFICATION	HALF WAVE RECTIFIER	CENTRE TAPPED	BRIDGE RECTIFIER
I_{DC}	$\frac{I_m}{\pi}$	$\frac{2I_m}{\pi}$	$\frac{2I_m}{\pi}$
V_{DC}	$\frac{V_m}{\pi}$	$\frac{2V_m}{\pi}$	$\frac{2V_m}{\pi}$
I_{rms}	$\frac{I_m}{2}$	$\frac{I_m}{\sqrt{2}}$	$\frac{I_m}{\sqrt{2}}$
V_{rms}	$\frac{V_m}{2}$	$\frac{V_m}{\sqrt{2}}$	$\frac{V_m}{\sqrt{2}}$
FORM FACTOR (F)	$F = 1.58$	1.11	1.11
CREST FACTOR (C)	2	$\sqrt{2}$	$\sqrt{2}$
RIPPLE FACTOR FREQ (fr)	f	2f	2f
CONDUCTION ANGLE	$\phi = \pi$ $\phi = \pi - 2\theta$ $\theta = \sin^{-1}\left(\frac{V_r}{V_m}\right)$	$\phi = 2\pi$ $\phi = 2\pi - 4\theta$ $\theta = \sin^{-1}\left(\frac{V_r}{V_m}\right)$	$\phi = 2\pi$ $\phi = 2\pi - 4\theta$ $\theta = \sin^{-1}\left(\frac{2V_r}{V_m}\right)$

f = supply freq.

I_L	$\frac{V_i - V_r}{R_s + R_f + R_L}$	$\frac{V_i - V_r}{\frac{R_s}{2} + R_f + R_L}$	$\frac{V_i - 2V_r}{R_s + 2R_f + R_L}$
THEVENIN'S EQUIVALENT			
PIV	V_m	$2V_m$	V_m
TRANSFER CURVE			
RIPPLE FACTOR	1.21	0.48	0.48
EFFICIENCY η_{max}	40.6%	$\frac{81.2}{\frac{R_s/2 + R_f}{R_L} + 1}$ $\eta_{max} = 81.2\%$	$\frac{81.2}{\frac{R_s + 2R_f}{R_L} + 1}$ $\eta_{max} = 81.2\%$
TUF	0.286	0.693	0.812

PIV OF BRIDGE RECTIFIER -

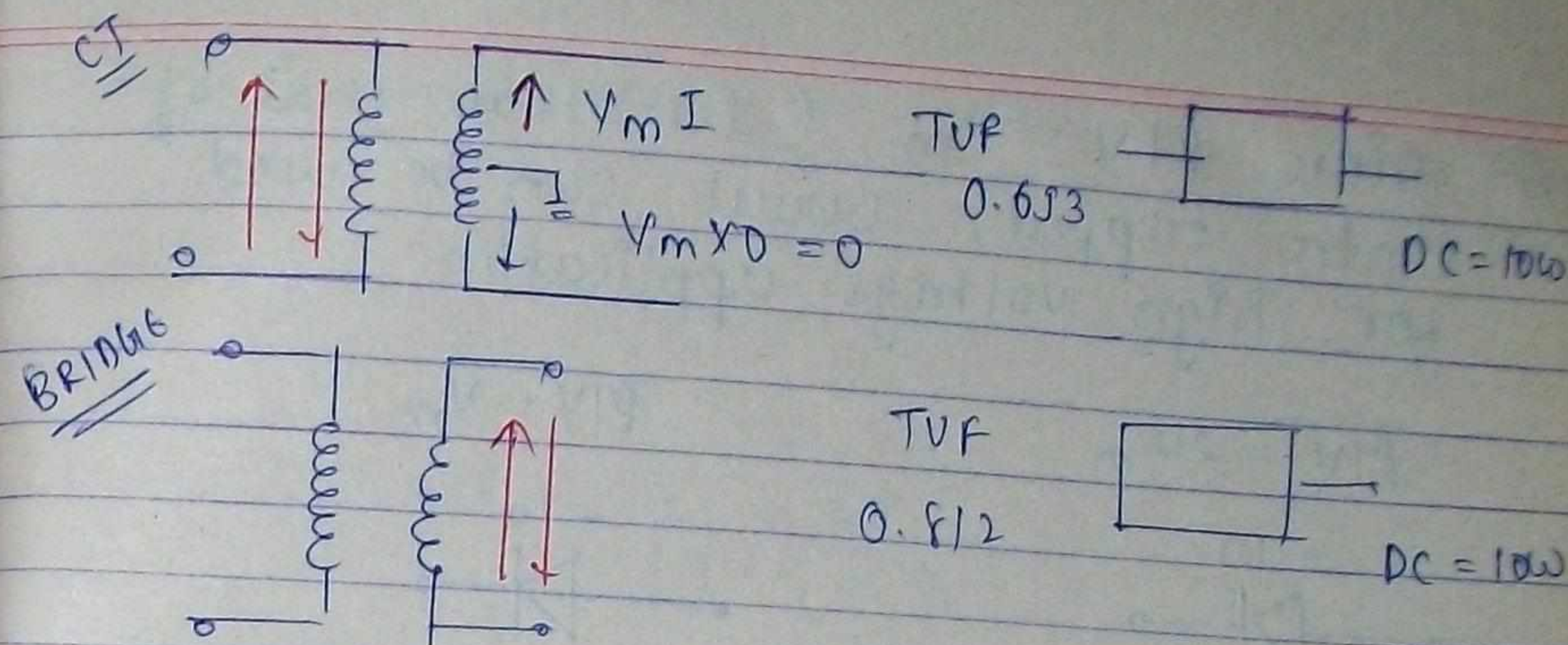
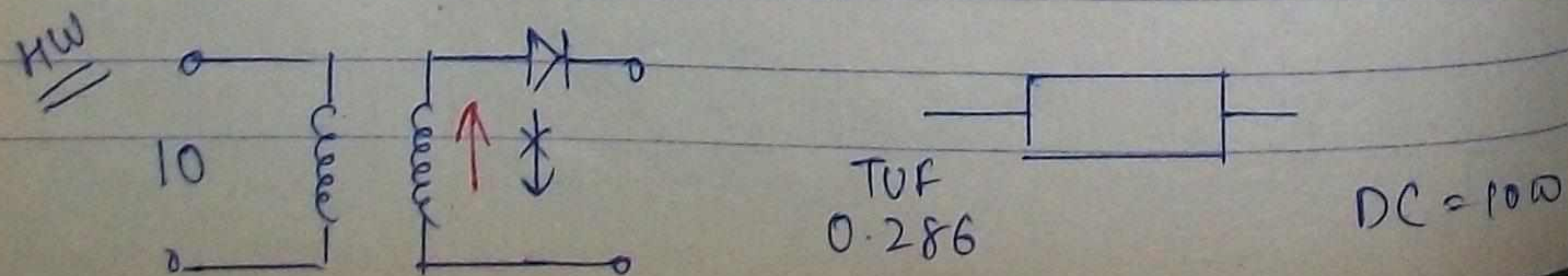


During F.B
D₂ & D₄ ARE
SHORT

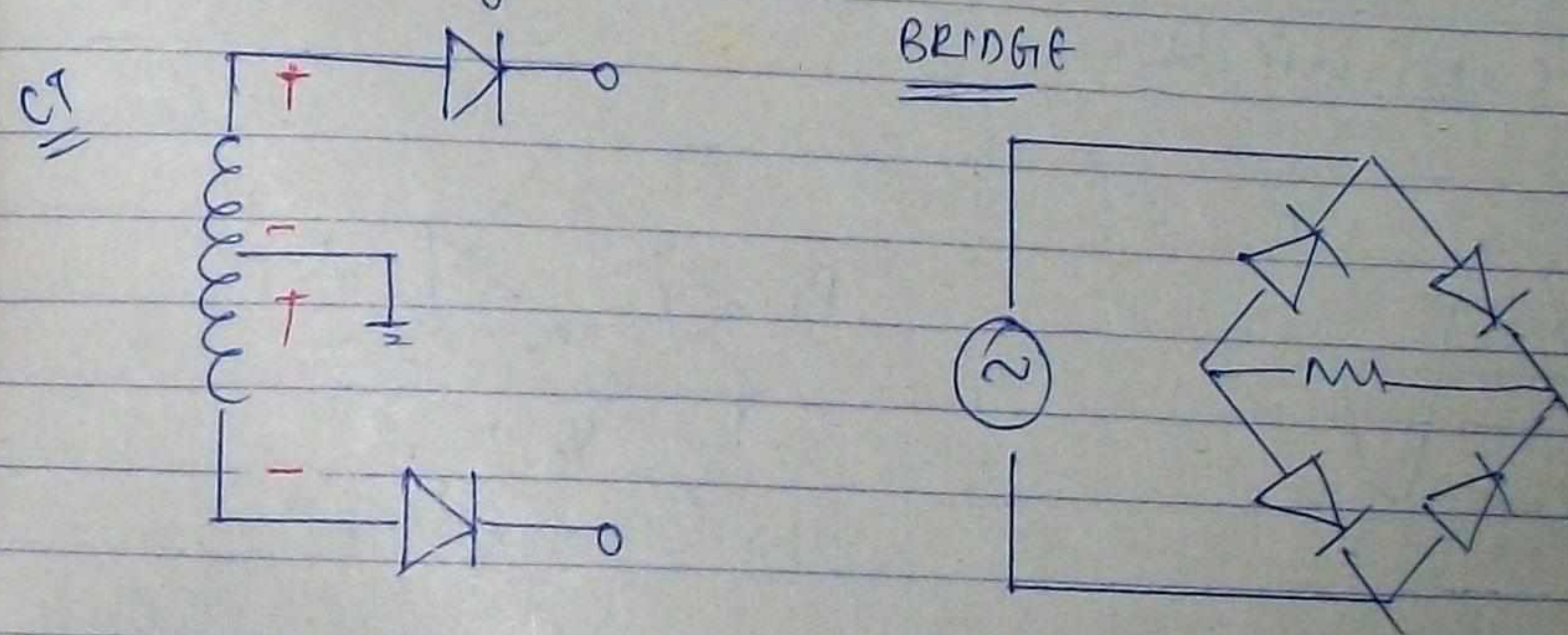
∴ $PIV = V_m$

⇒ ADVANTAGES OF BRIDGE RECTIFIER -

- ① The current in both the primary & secondary of power transformer is present for entire cycle and hence for given power O/P, power transformer of a small size & low cost may be used.



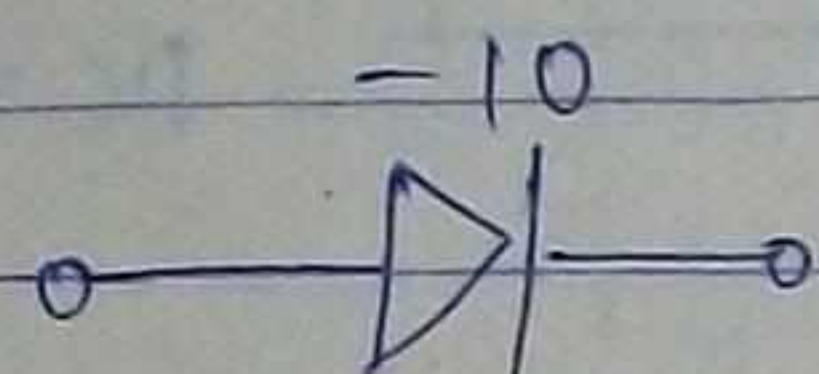
- ② No centre tap is required in transformer secondary hence, wherever possible ac voltage can directly be supplied to bridge.



- ③ The current in secondary of transformer is in opposite direction in two half cycle hence net dc component is zero, which reduces the loss and reduces the danger of saturation of transformer.

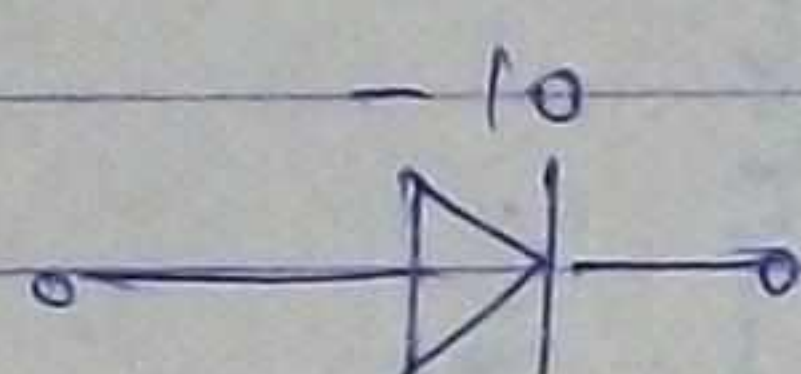
④ Since $PIV = V_m$ (less than that of centre tapped) circuit can be used for high voltage application.

$$PIV = 2V_m$$



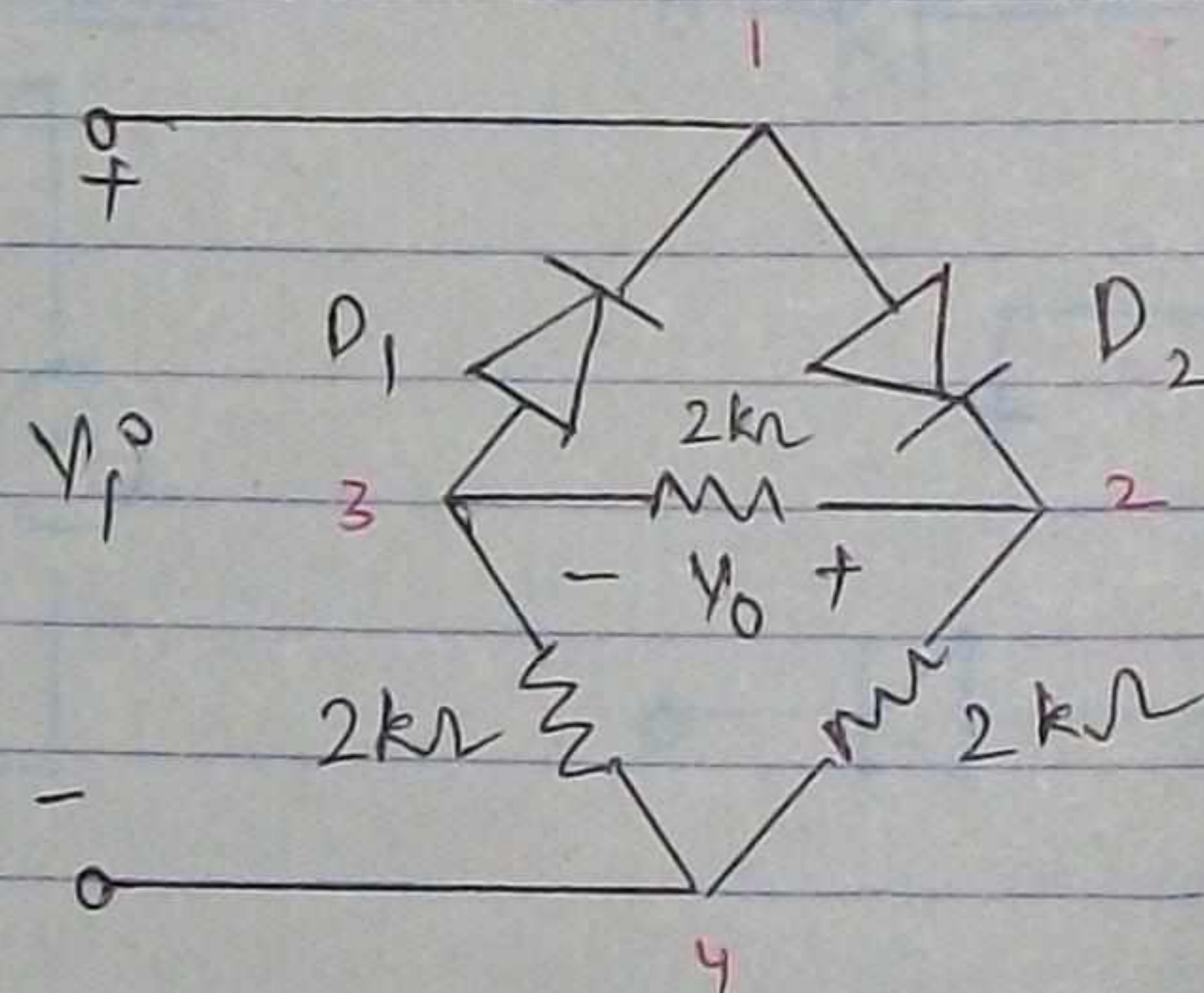
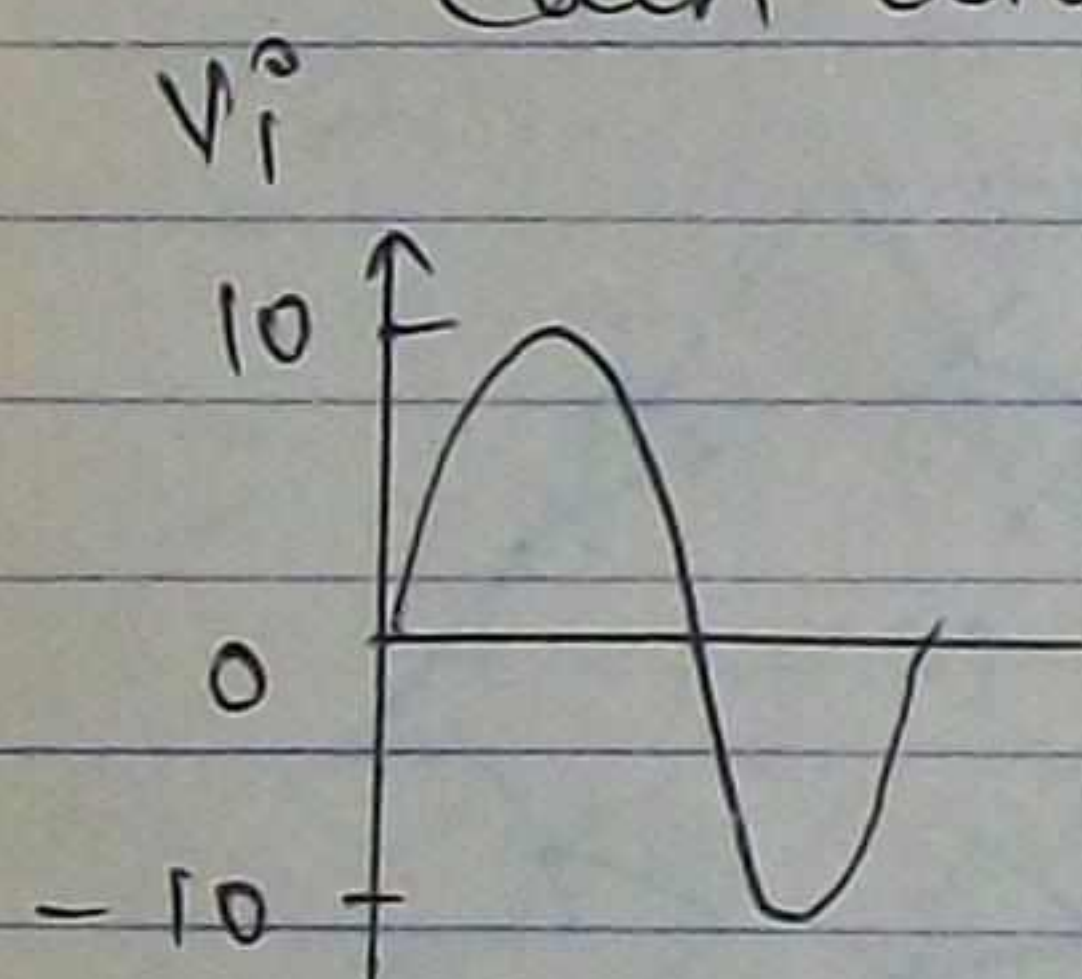
$$2V_m \leq 10$$

$$PIV = V_m$$



$$V_m \leq 10$$

Ques. Determine the o/p waveform of the figure and calculate the o/p DC level and required PIV of each diode.

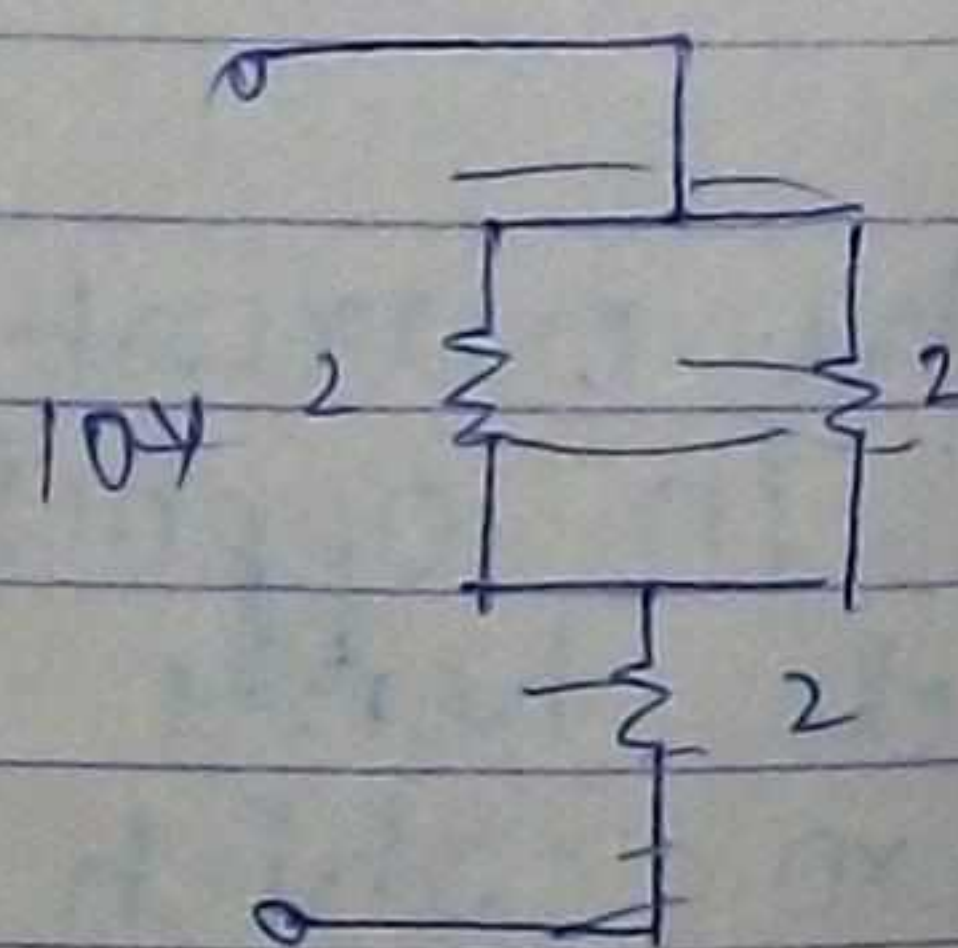


Solⁿ: During +ve half:

$$10 = 3k\Omega I$$

$$I = \frac{10}{3}$$

$$V_o = 2 \times \frac{10}{3 \times 2} = \frac{10}{3}$$



During +ve half

$$10 = \frac{8}{6} I$$

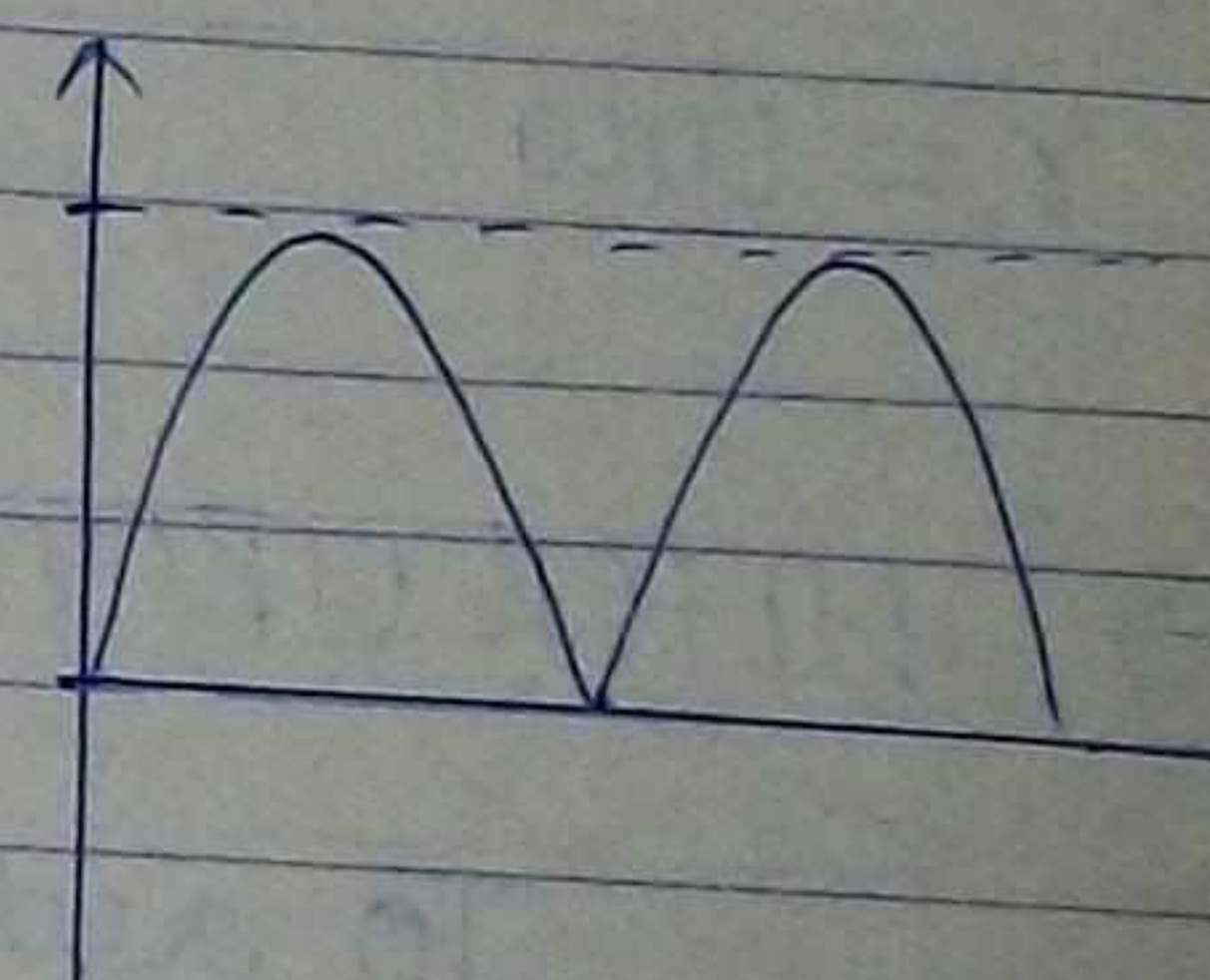
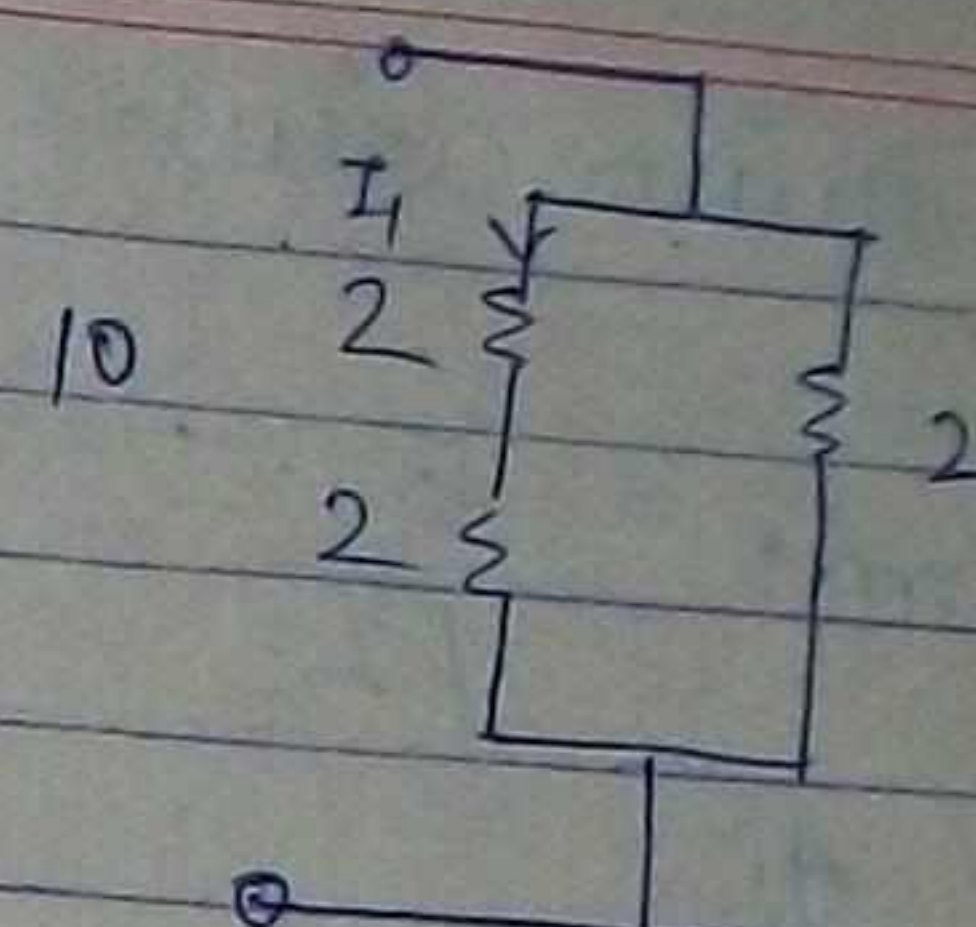
$$I = \frac{60}{8}$$

$$V_o = 2 \times \frac{2}{2+4} \times \frac{60}{8} \times \frac{10}{2}$$

$$V_o = \frac{10}{2} = 5$$

During -ve half $5 = \frac{10}{2}$

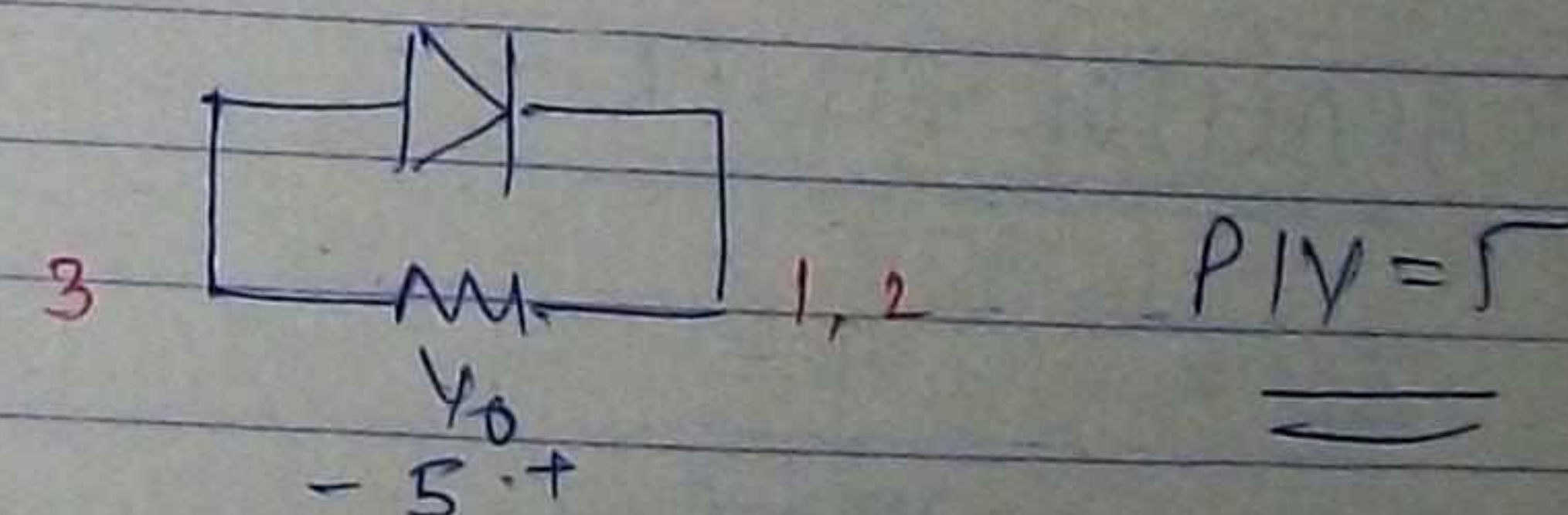
$$V_o = -(-\frac{10}{2}) = 5$$



$$\text{Average} = V_{DC} = \frac{2V_m}{\pi} = \frac{2 \times 5}{\pi} = \frac{10}{\pi}$$

PIV -

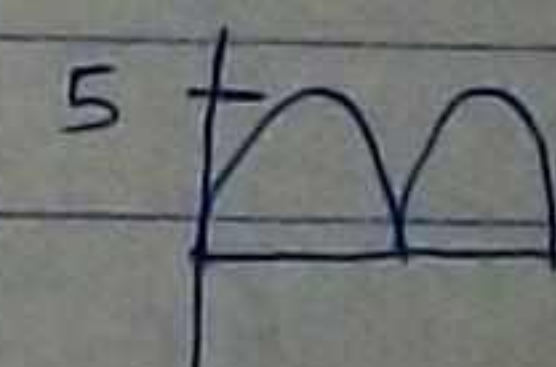
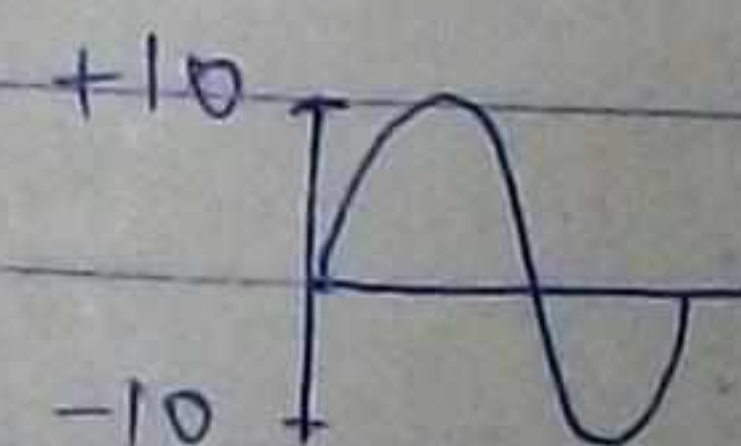
Used for higher rating



I/P

O/P for V_{DC} PIV

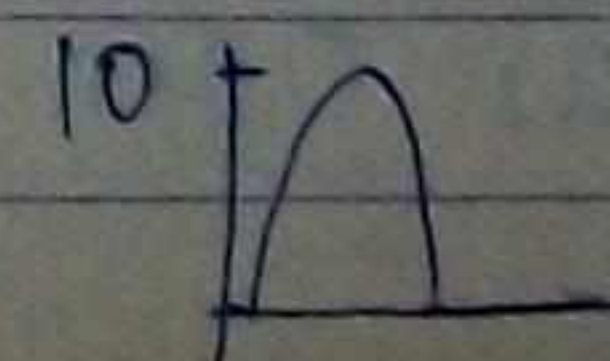
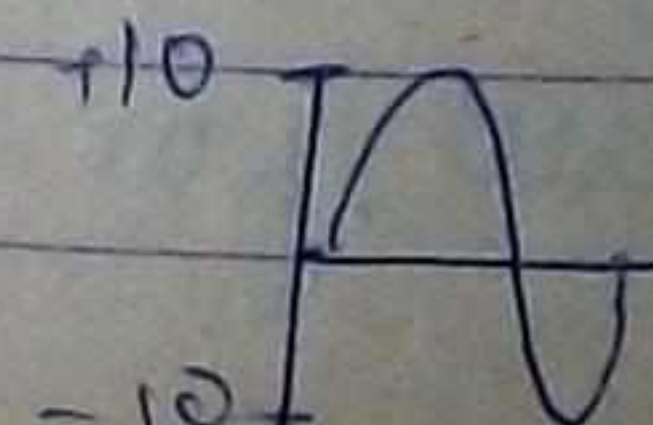
Given circuit



$$2 \times \frac{10}{\pi}$$

$$5 = \frac{V_m}{2}$$

HW Rectifier



$$+ \frac{10}{\pi}$$

$$V_m$$

(for given circuit

$$V_{rms} = \frac{5}{\sqrt{2}}$$

↓ ac component at o/p

∴ $r = \text{Low}$

for half wave

$$V_{rms} = \frac{10}{2} = 5$$

rms value ↑
ac component ↑
(disadvantage)

∴ $r = \text{HIGH}$

⇒ FILTER CIRCUITS -

To reduce ac component in the o/p filter circuits are used, we are using L and C in filter circuit.

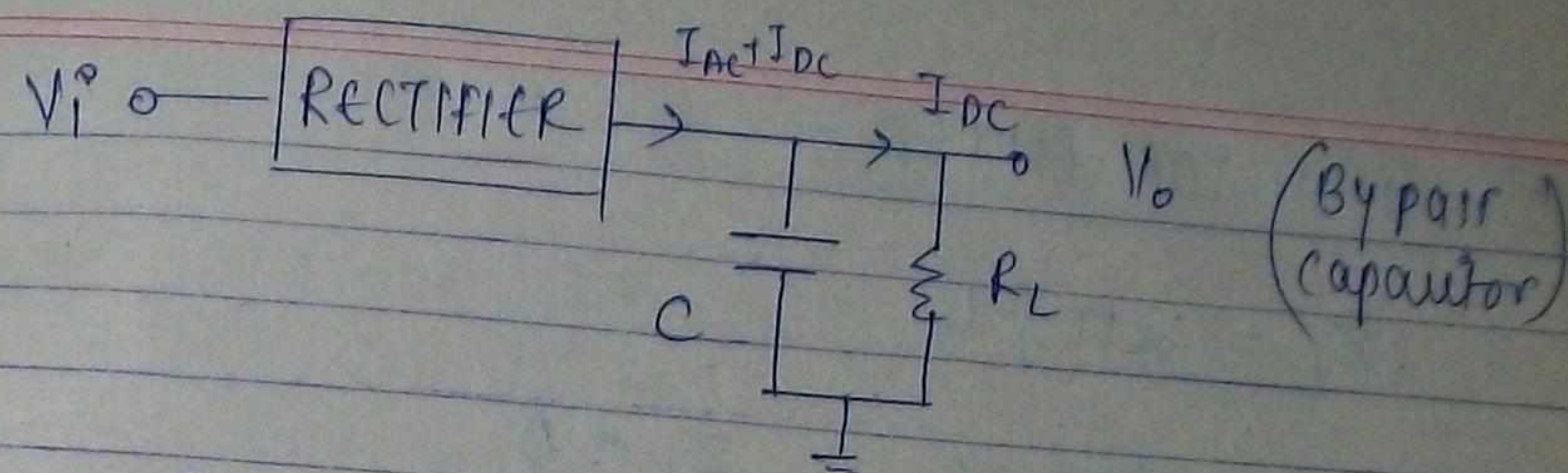
① CAPACITOR FILTER -

$$Z_c = \frac{1}{2\pi f C}$$

for dc $f = 0$ $Z_c = \infty$

for ac $f \neq 0$ $Z_c = Z$

C → acts as open circuit for dc.



* C → V.V. HIGH (1000 μF approx)

Z_c → V.V. LOW for AC
and act as short circuit.

for HW

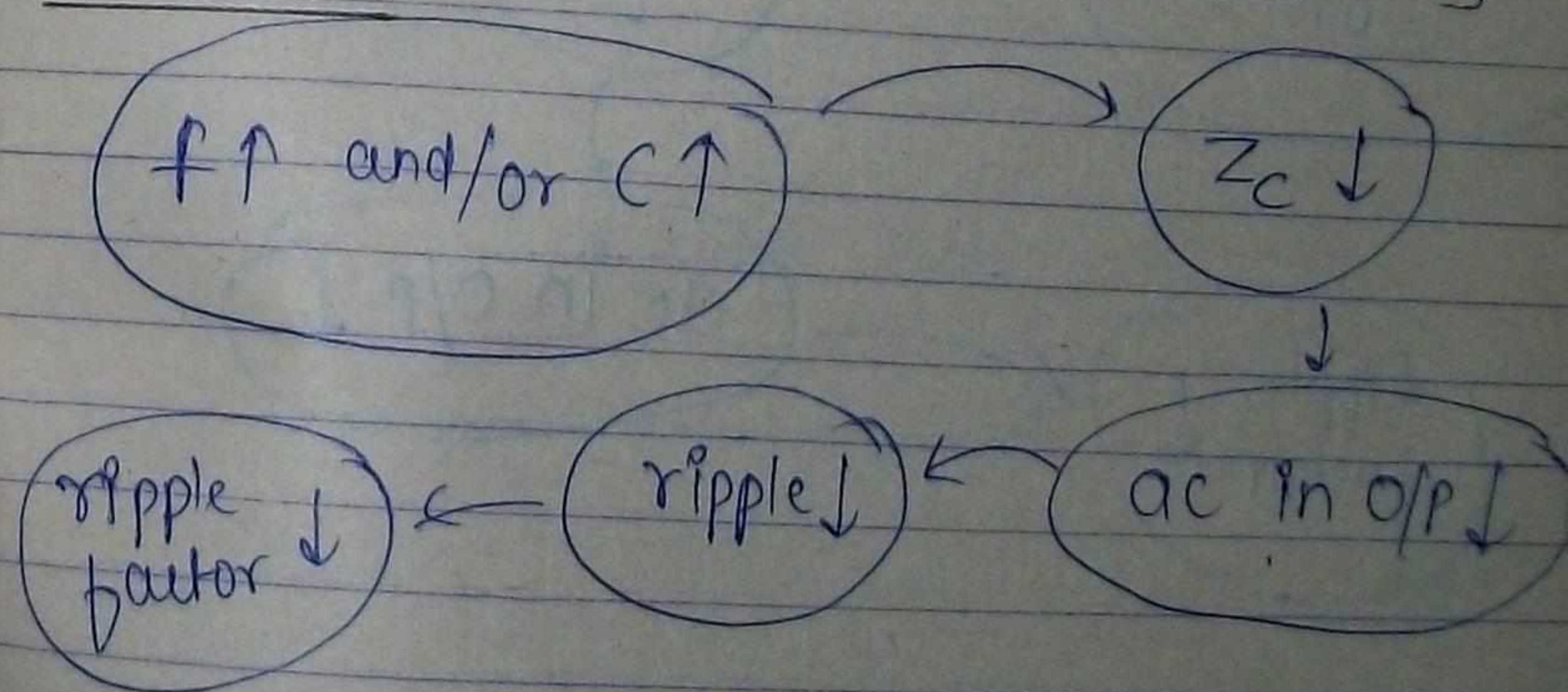
$$Z_c = \frac{1}{2\pi f \times C} = Z$$

for f.w

$$Z_c = \frac{1}{2\pi \times 2f \times C} = Z/2$$

} Thus ripple freq should be high

CONCLUSION -



$$\therefore \boxed{r \propto \frac{1}{fC}}$$

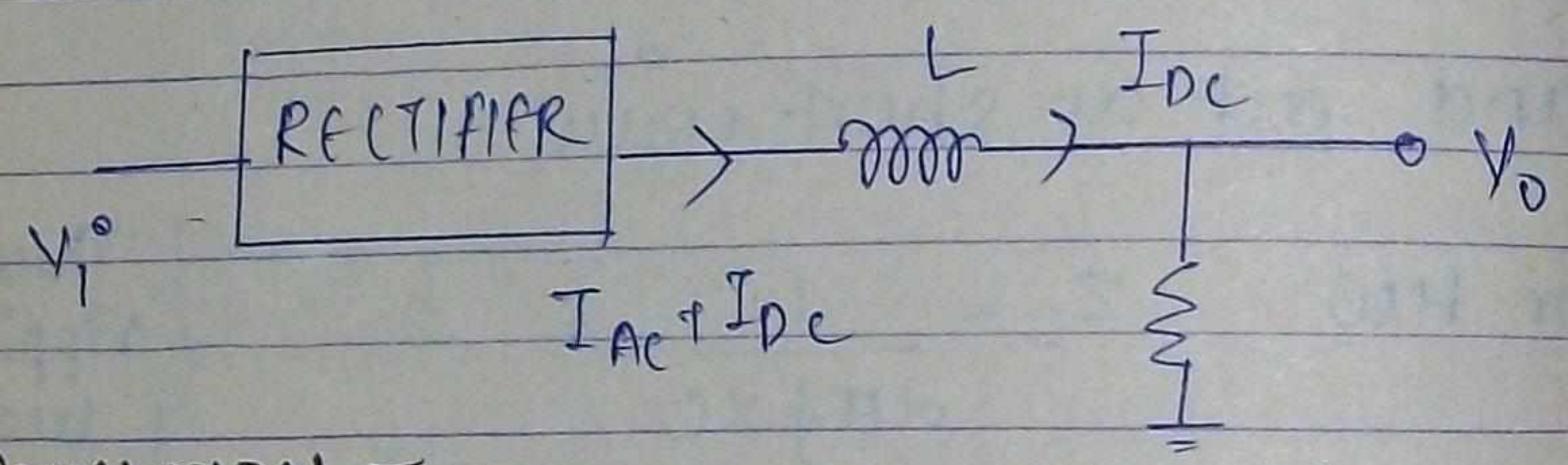
- 1. CAPACITOR FILTER
- 2. INDUCTOR FILTER
- 3. L SECTION FILTER
- 4. π FILTER

② INDUCTOR FILTER -

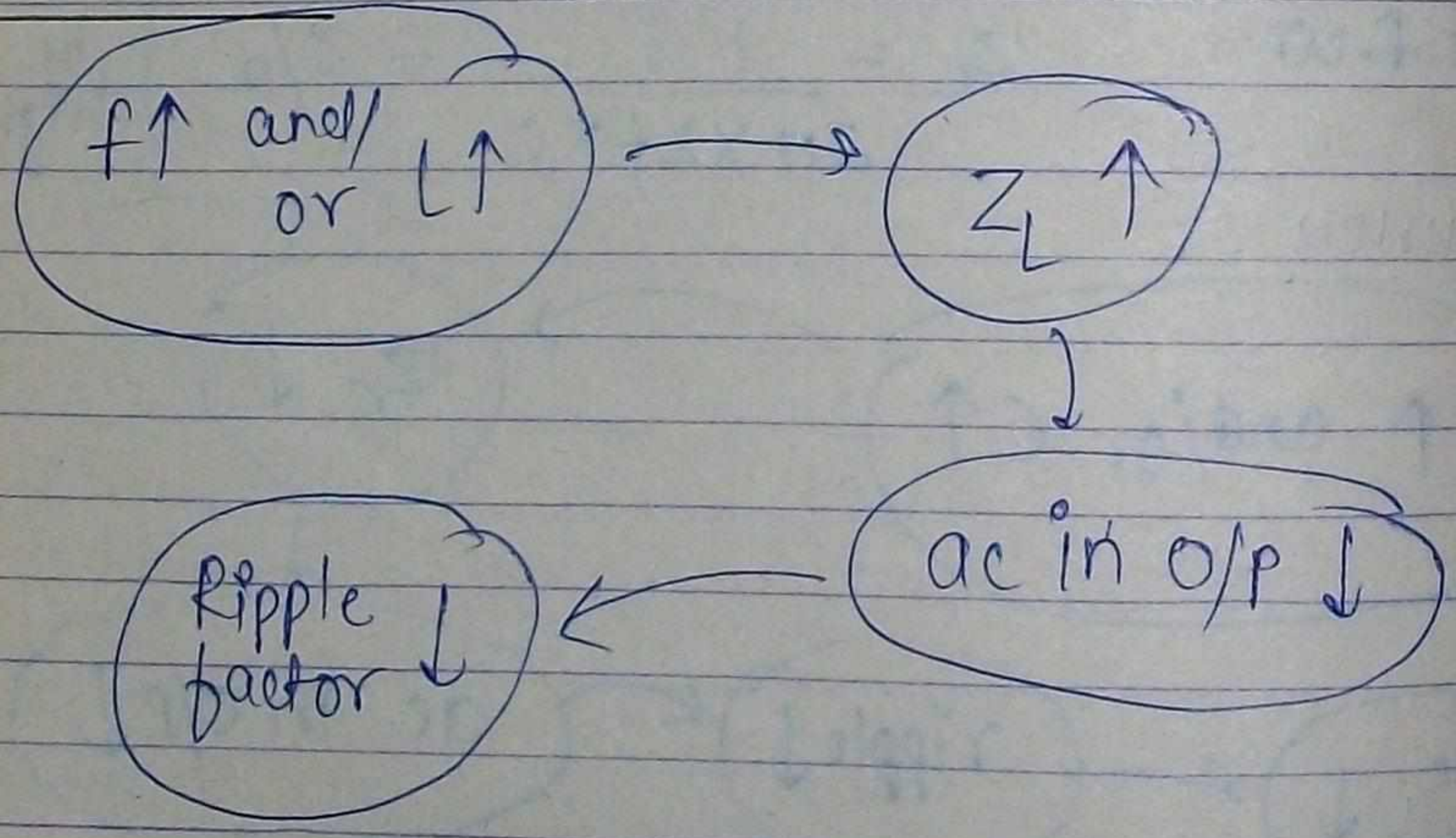
$$Z_L = 2\pi fL$$

$f=0$ for dc, $Z_L = 0$

L acts as short circuit for dc.



CONCLUSION -

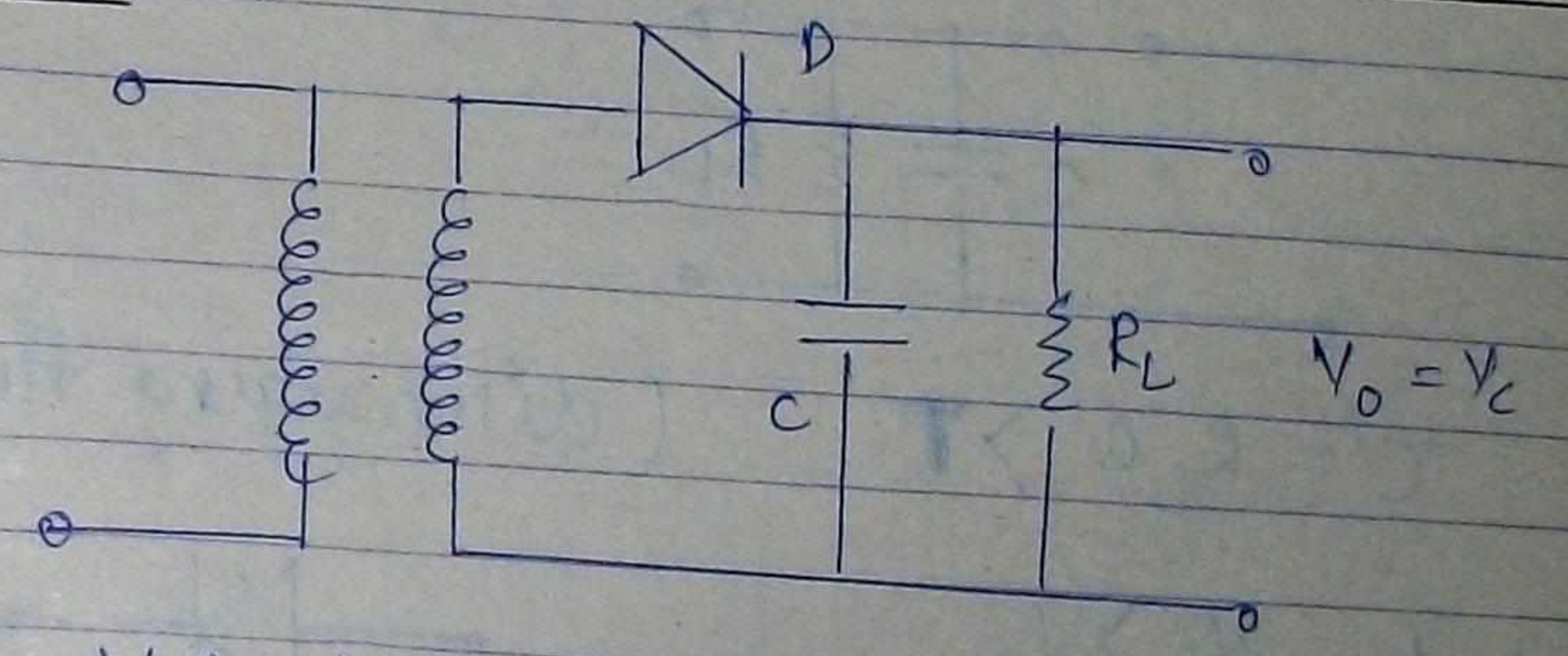


$$\gamma \propto \frac{1}{fL}$$

⇒ FILTERS -

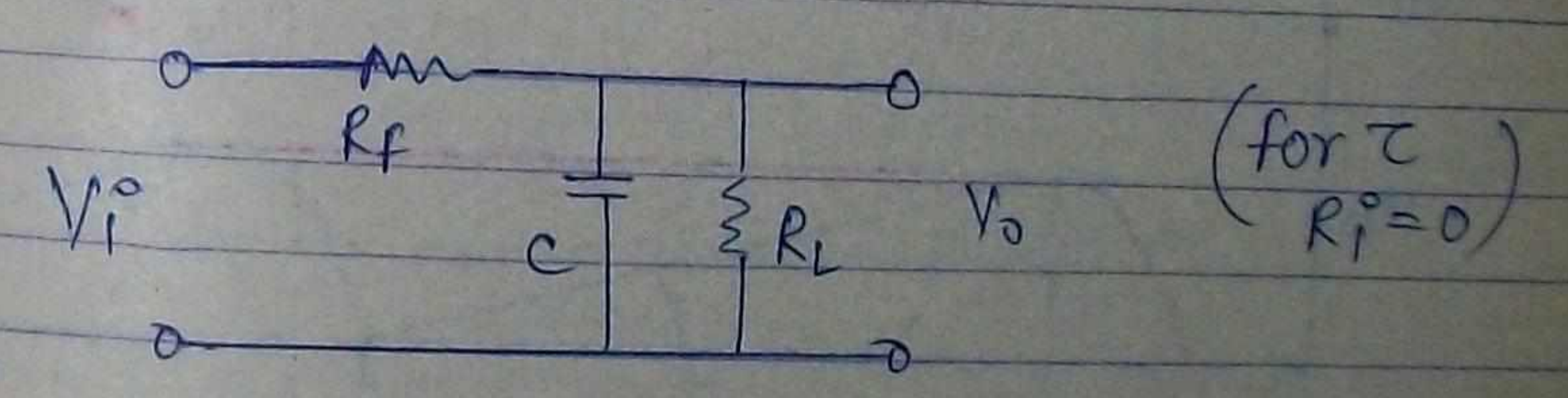
- Capacitor filter
- Inductor filter
- L section filter
- π filter.

⇒ HALF WAVE RECTIFIER WITH CAPACITOR FILTER -



$$V_c(0^-) = V_c(0^+) = 0$$

During 1st +ve half cycle $D \rightarrow ON$



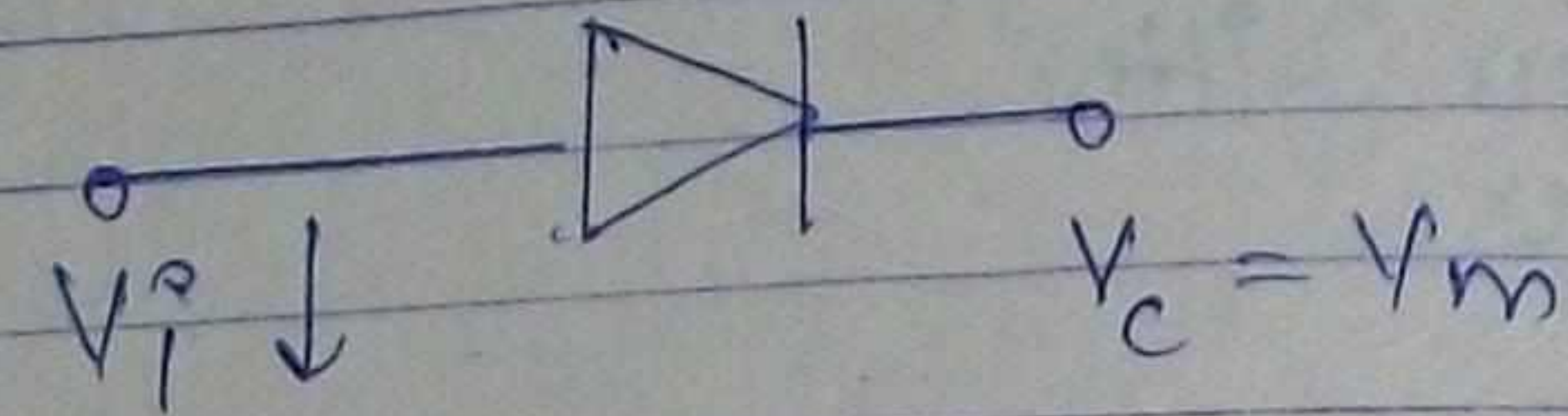
(Charging time) $\tau = (R_f || R_L)C \approx R_f C < T_2$

$R_f \ll R_L$

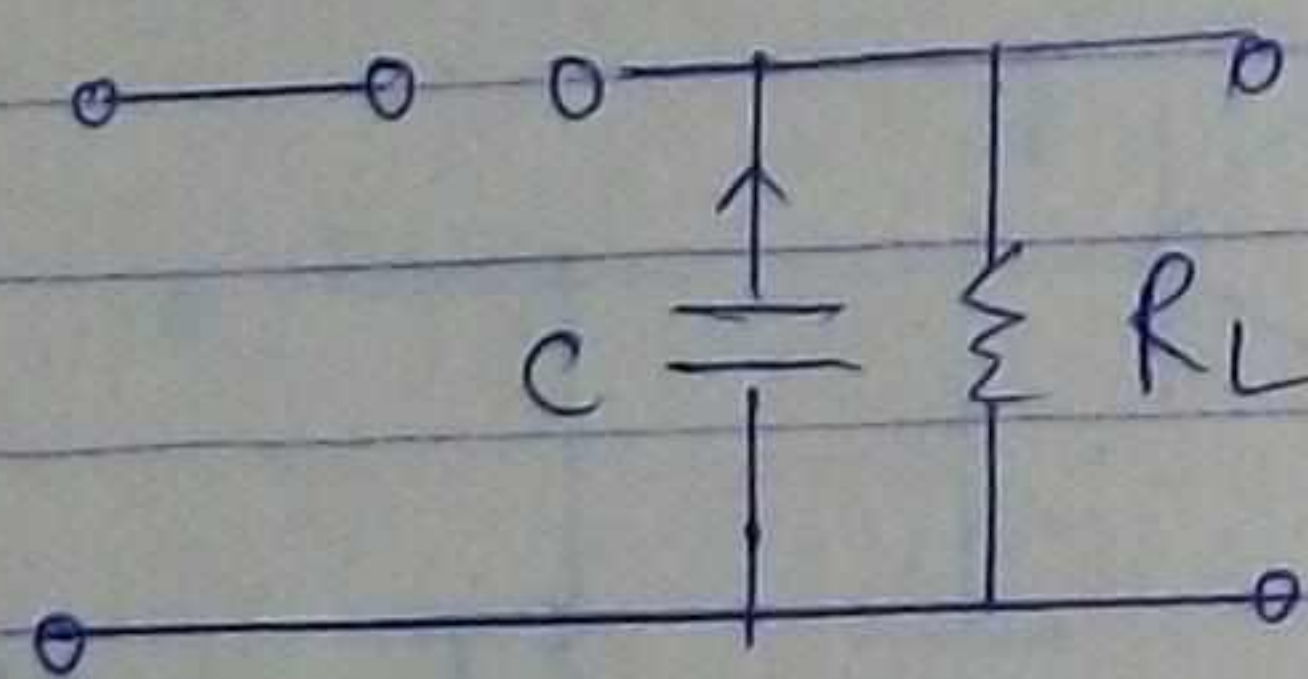
TIME PERIOD OF I/P SIGNAL

$\therefore C \rightarrow$ charges towards V_m

at $t = T/4^-$ $V_i = V_m$, $V_c = V_o = V_m$



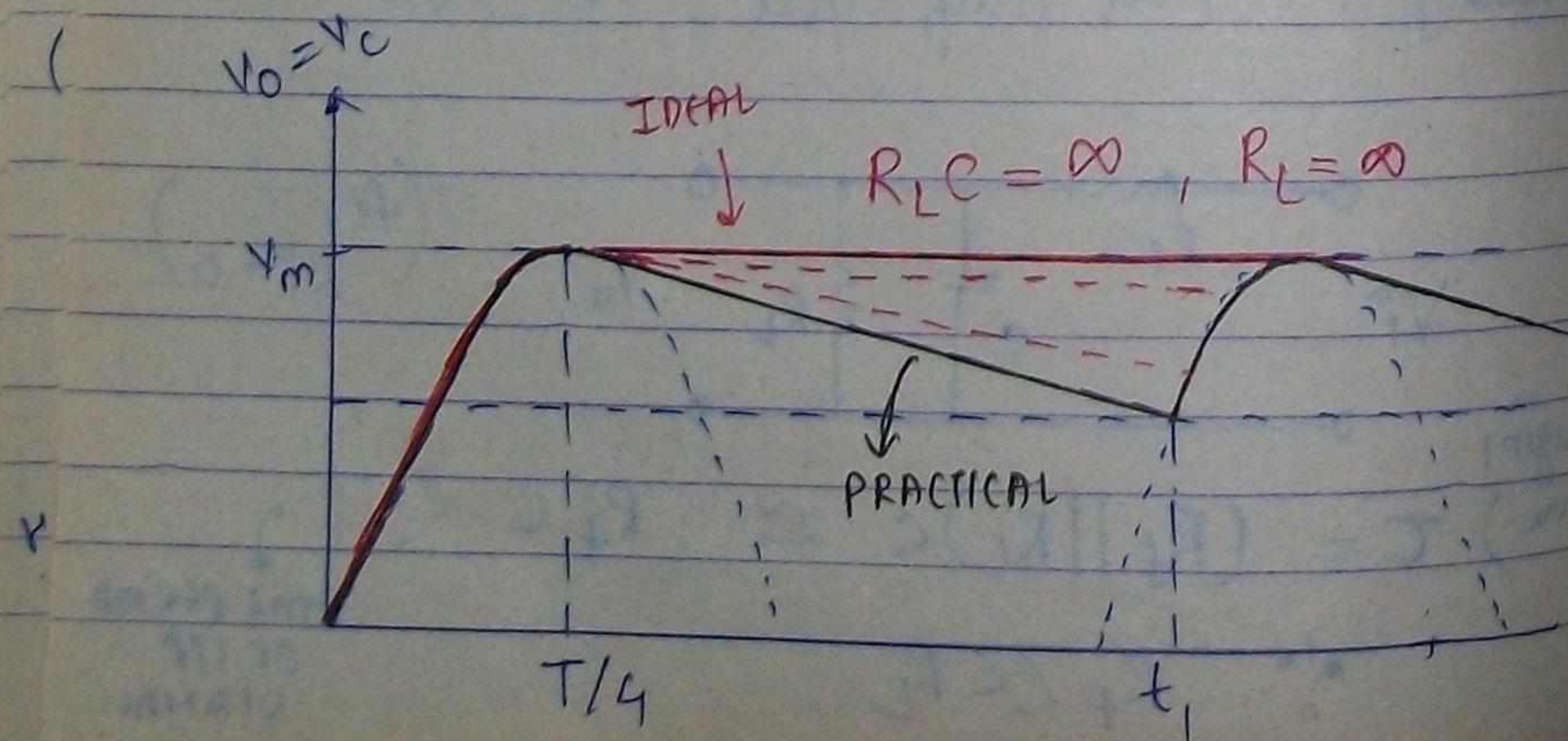
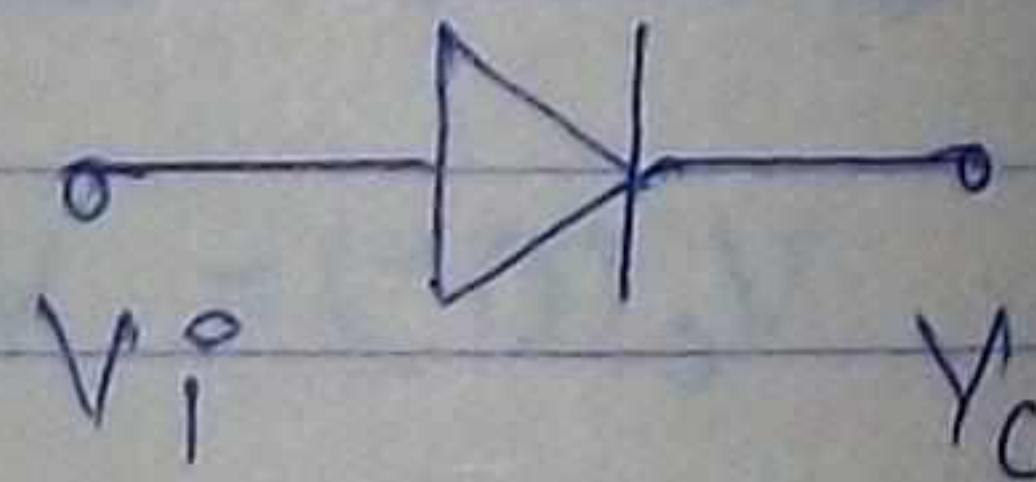
$V_i < V_m$ $\therefore D = \text{OFF}$



$\tau' = R_L C > T$ (discharging time)

at t_1 $V_i > V_c$

$\therefore D \rightarrow \text{ON}$ $\tau = R_f C$

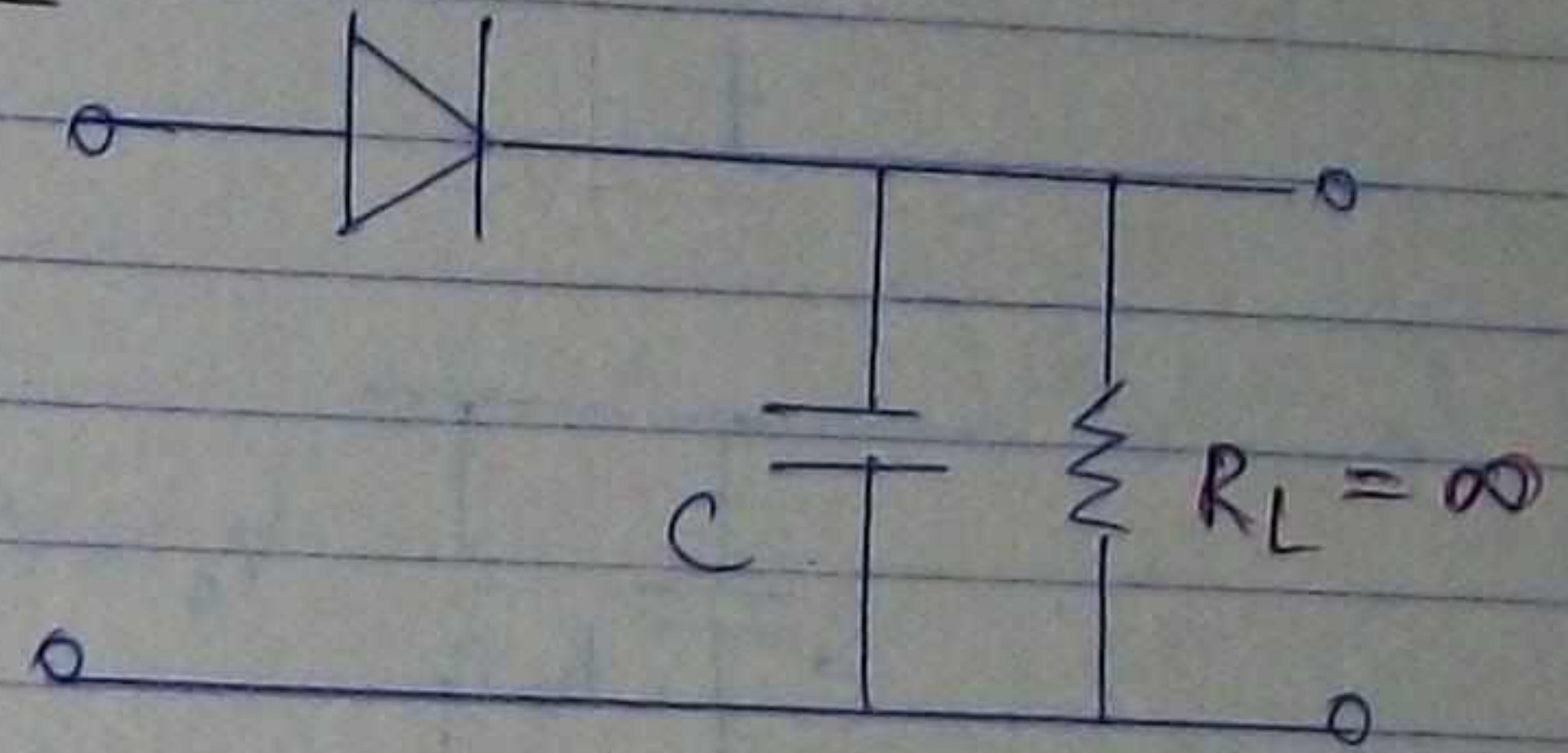


$R_L = \infty$

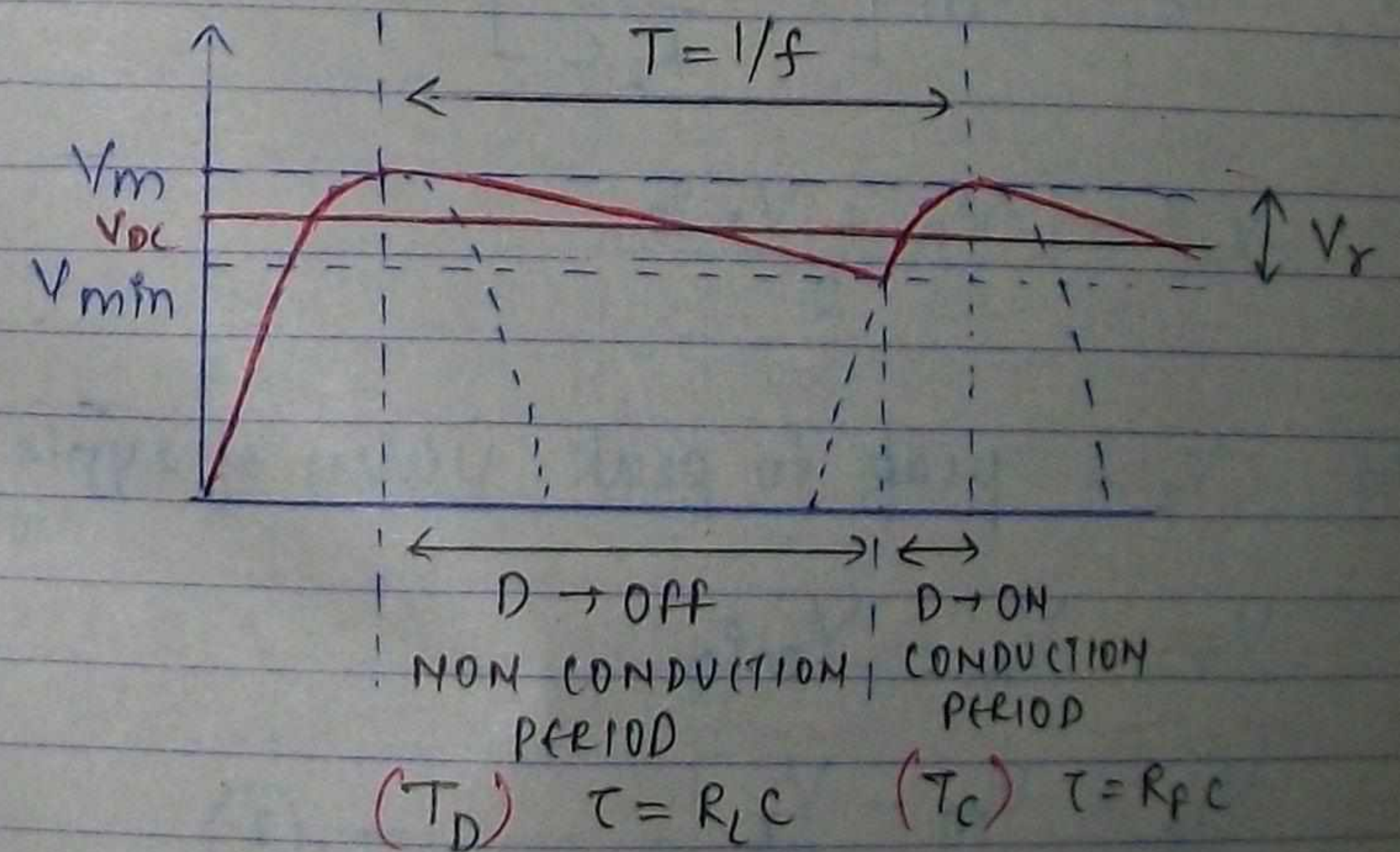
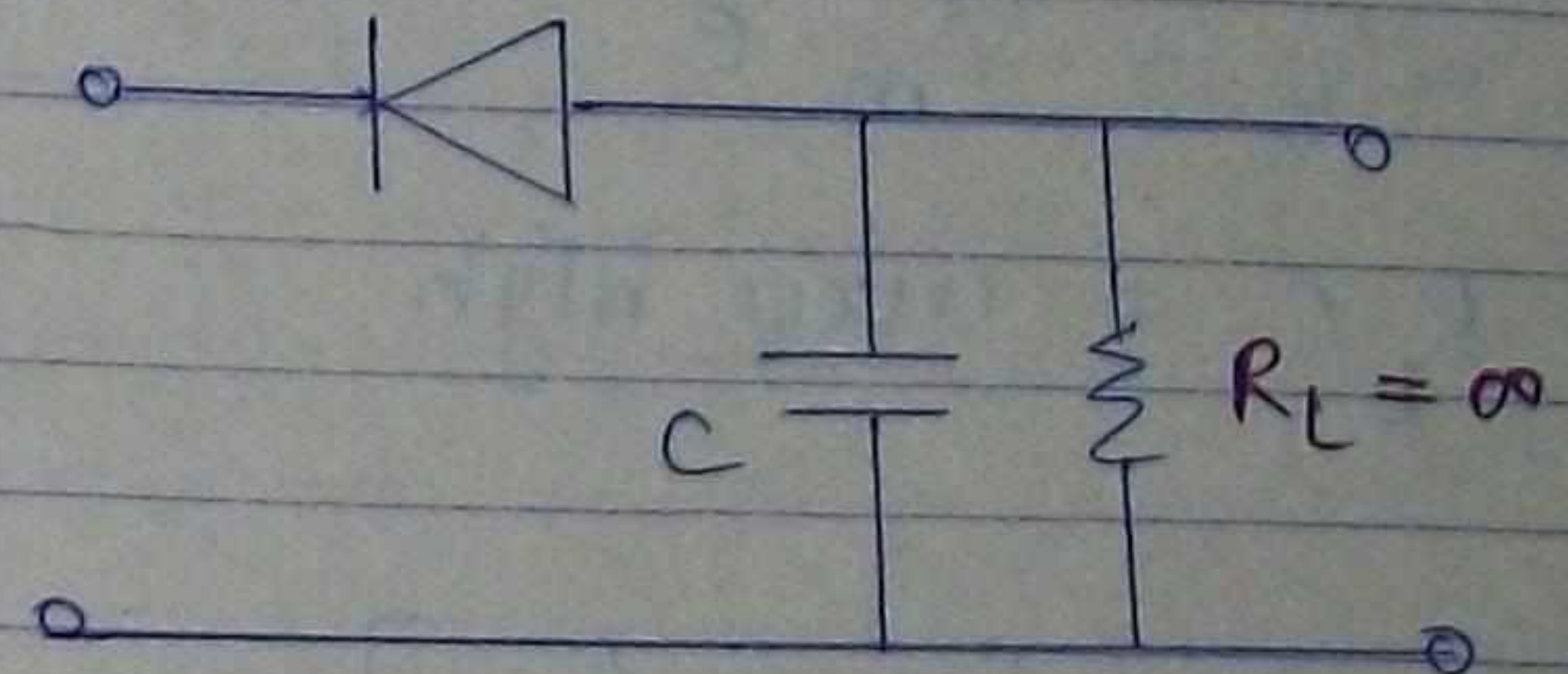
CREST FACTOR = 1

PEAK DETECTOR -

POSITIVE
PEAK
DETECTOR



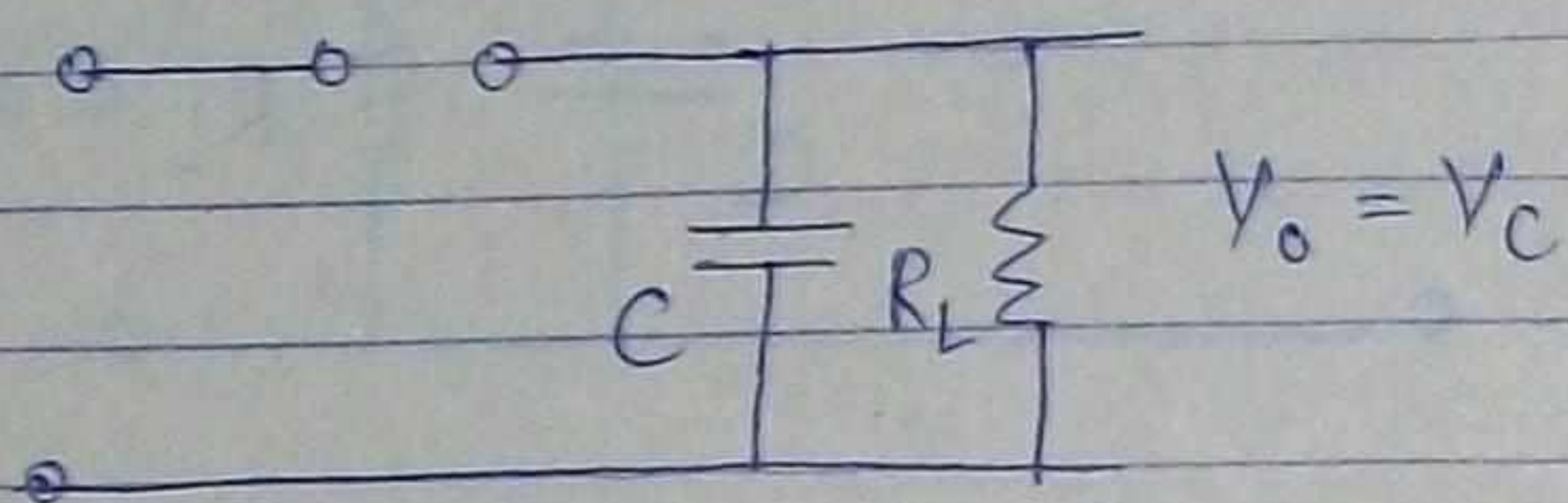
NEGATIVE
PEAK
DETECTOR



$$T_D + T_C = T$$

for best filter $T_D \gg T_C$

Ans $T_D \cong T = \frac{1}{f}$ — (1)



$$V_c = V_o = V_m e^{-t/R_L C}$$

∴ $R_L C = \text{very high}$

∴ $V_c = V_m \left[1 - \frac{t}{R_L C} \right]$

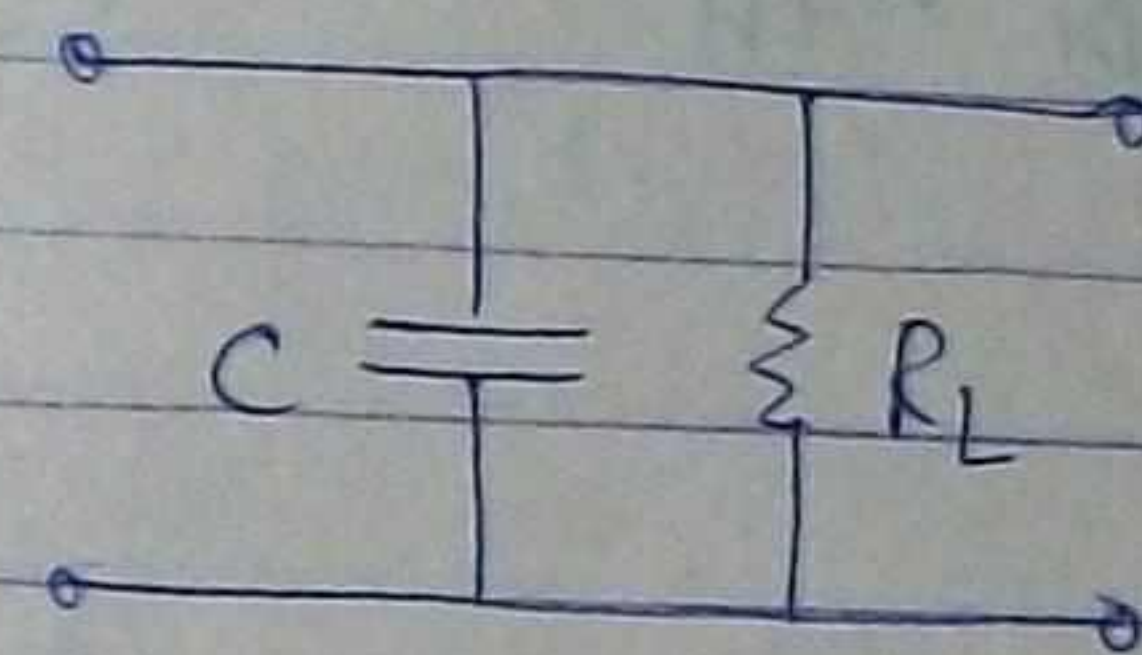
$$V_{dc} = \frac{V_m + V_{min}}{2}$$

Let $V_r = \text{peak to peak value of ripple voltage}$

$$V_r = V_m - V_{min}$$

∴ $V_{dc} = V_m - \frac{V_r}{2}$ — (2)

$V_r \rightarrow \text{Change in } V_c \text{ during } T_D$



$$I_{dc} = \frac{V_{oc}}{R_L} \quad (\text{average current})$$

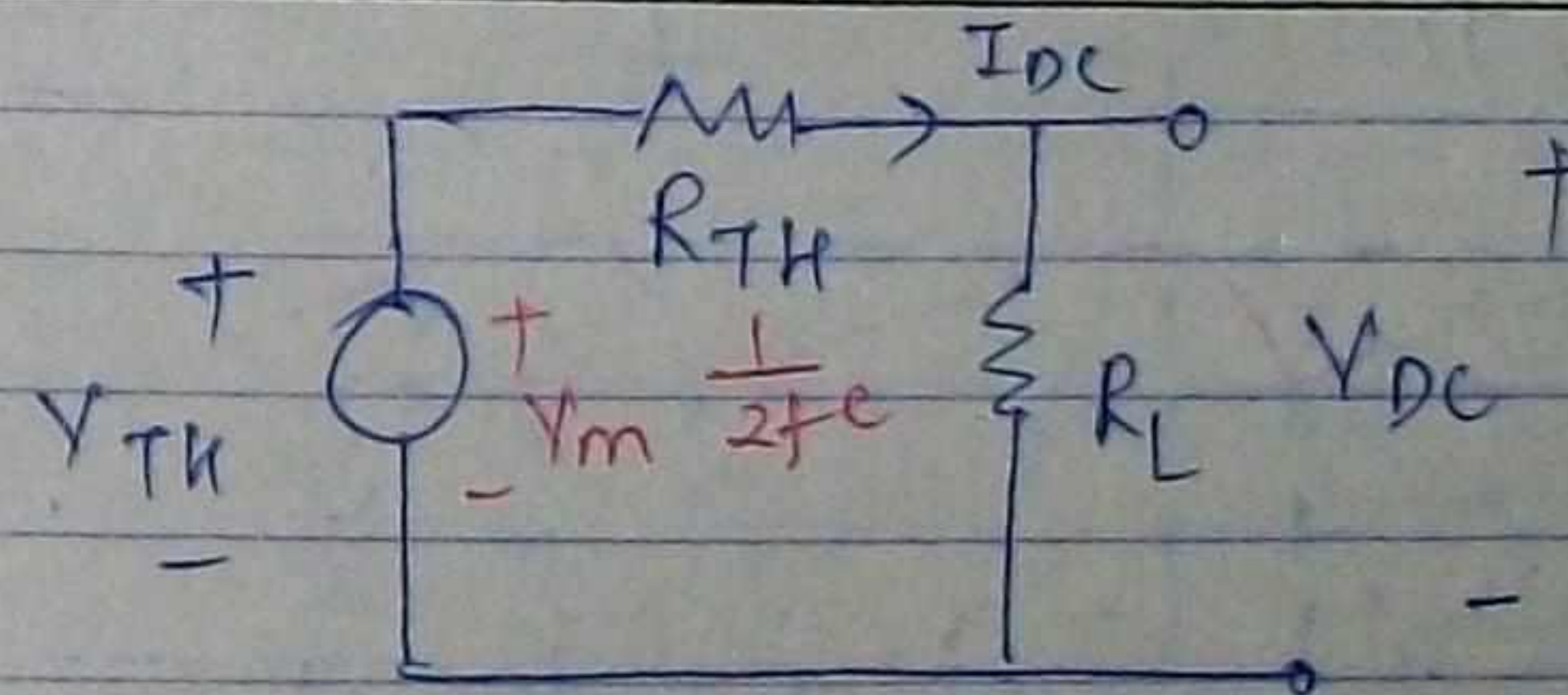
$$Q(\text{discharge}) = I_{dc} \times T_D$$

$$V_r = \frac{Q(\text{discharge})}{C} = \frac{I_{dc} T_D}{C}$$

Ans $V_r = \frac{I_{dc} T}{C} = \frac{I_{dc}}{Cf}$

Ans $V_{dc} = V_m - \frac{I_{dc}}{2Cf}$

THEVENIN'S EQUIVALENT OF HALF WAVE RECTIFIER WITH CAPACITOR FILTER —



$$V_{DC} = \frac{V_{TH} \cdot R_L}{R_{TH} + R_L}$$

25/08/11

$$V_{DC} = V_{TH} - I_{DC} R_{TH} \quad - (2)$$

$$V_{TH} = V_m$$

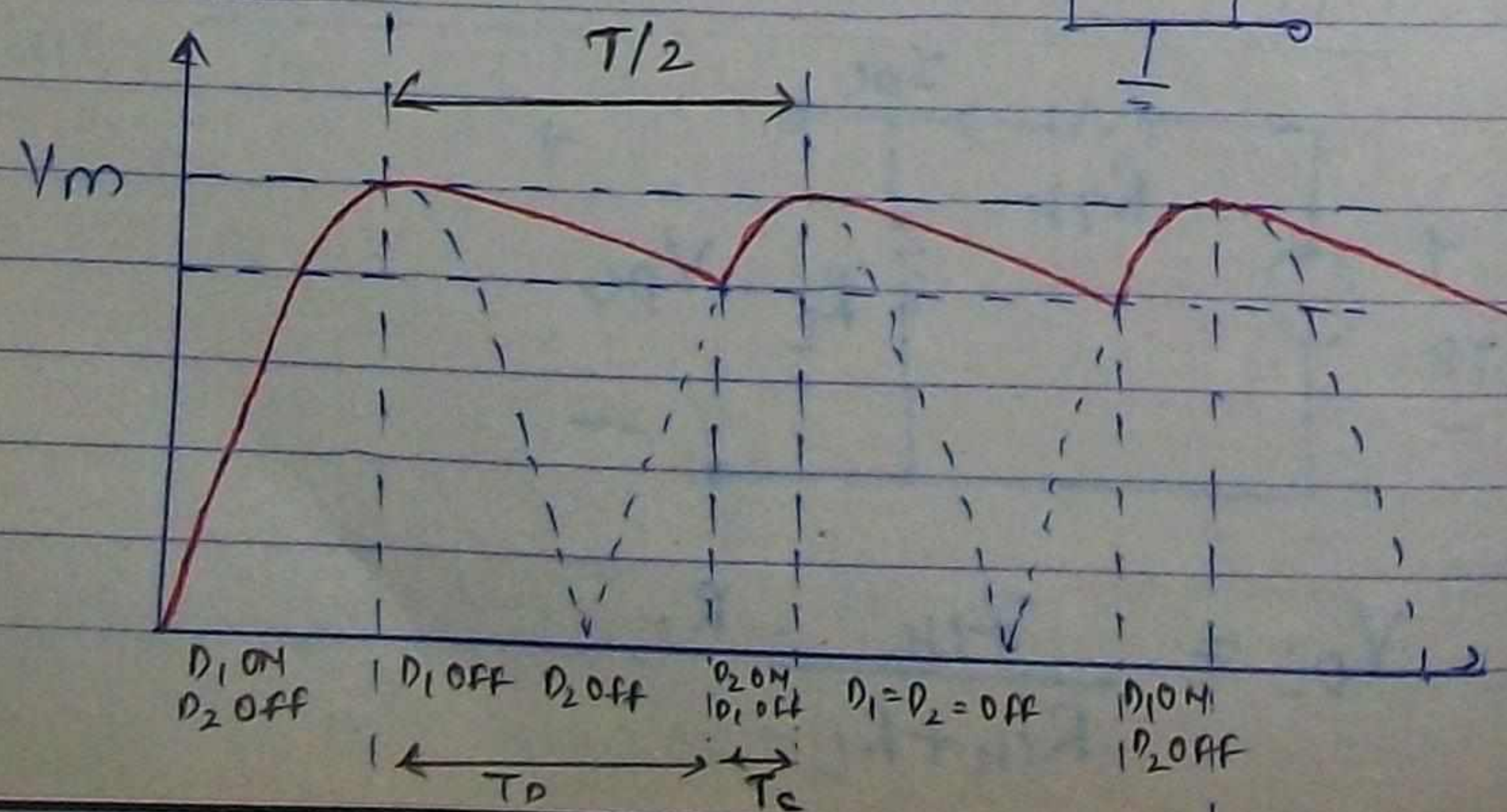
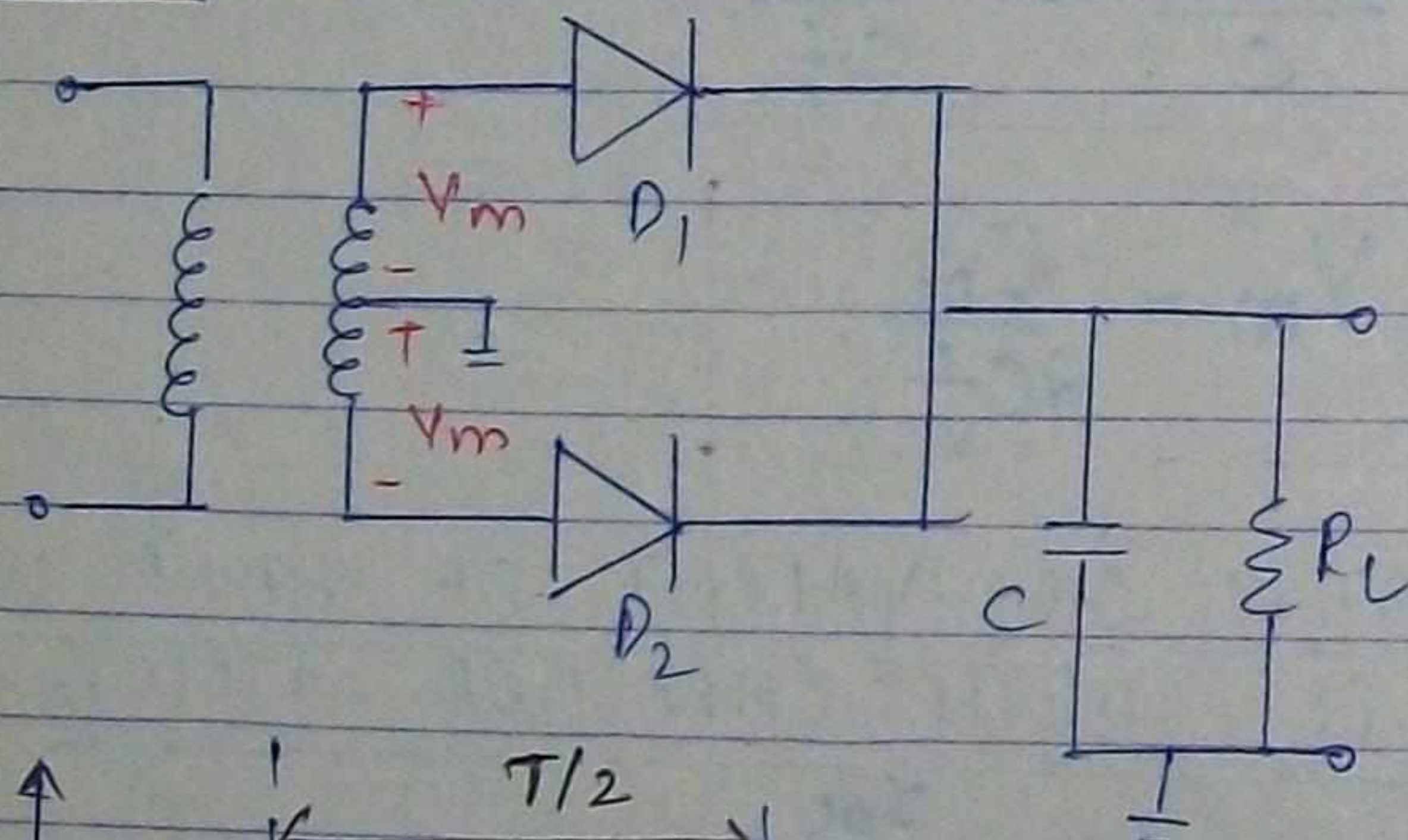
$$\text{also } V_{DC} = V_m - \frac{I_{DC}}{2fc} \quad - (1)$$

from (1) and (2) :

$$V_{TH} = V_m$$

$$R_{TH} = \frac{1}{2fc}$$

⇒ FULL WAVE RECTIFIER WITH CAPACITOR FILTER -



T_D = non conduction period
 T_C = conduction period

$$T_D + T_C = T/2$$

Since for best filter $T_D \gg T_C$

$$T_D \approx T/2$$

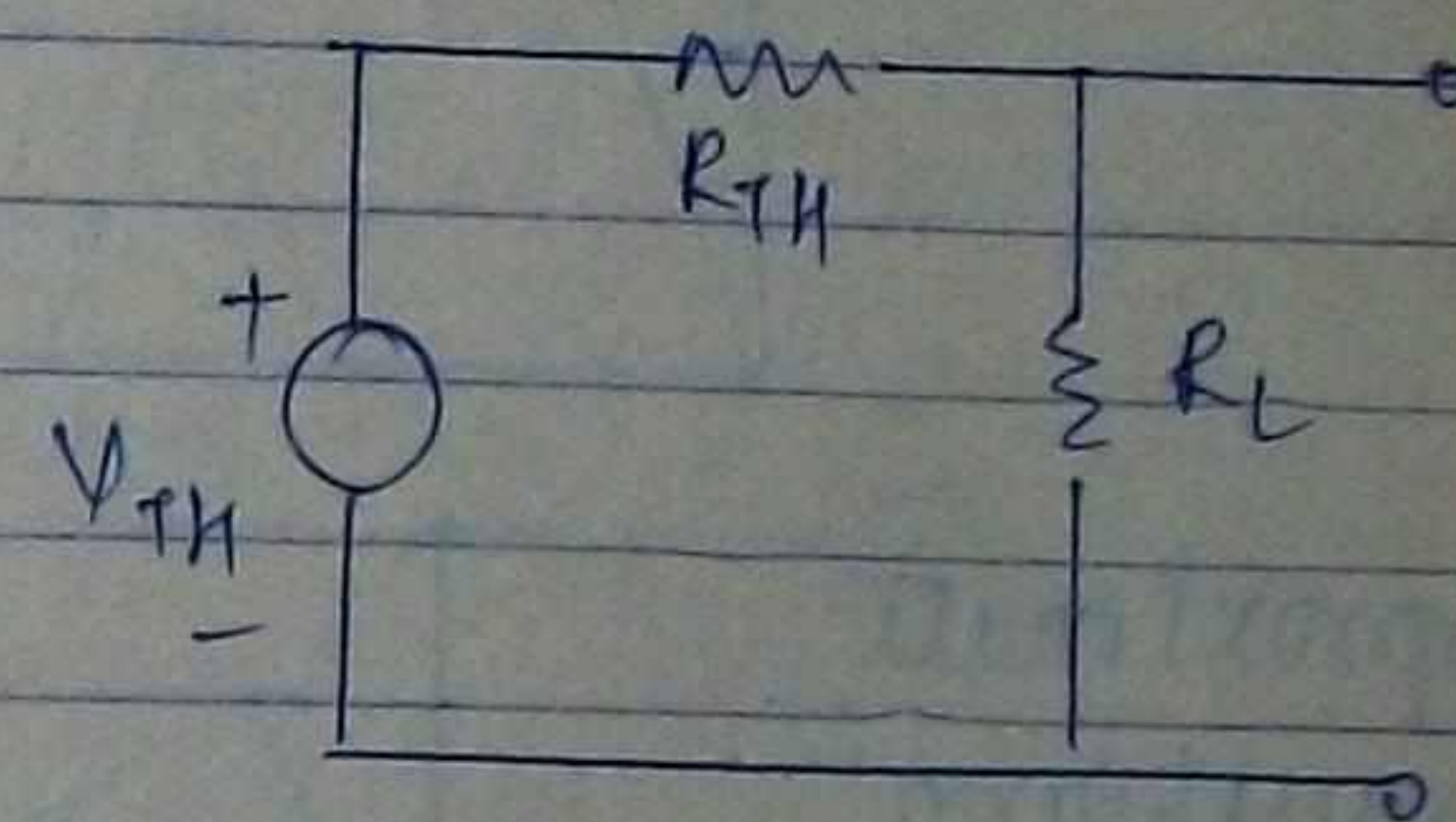
or

$$T_D \approx 1/2f$$

THEVENIN'S EQUIVALENT -

Here

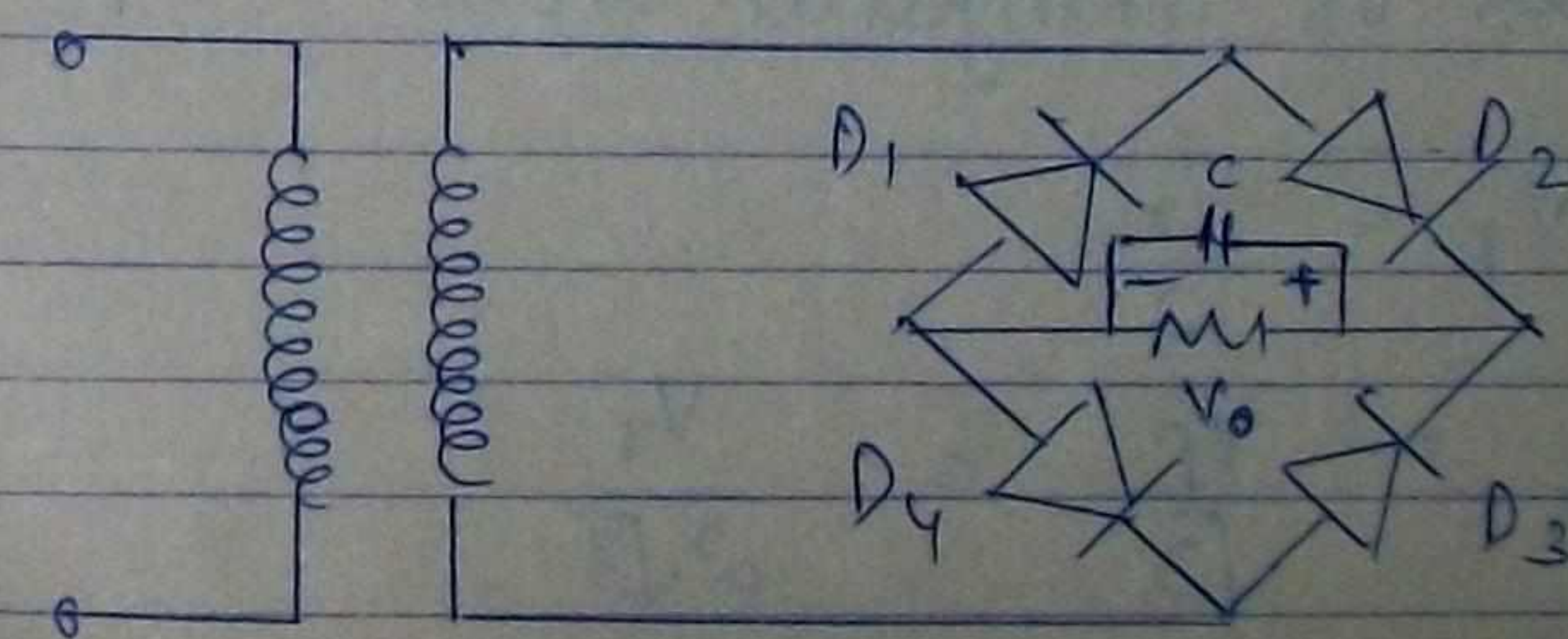
$$V_r = \frac{I_{DC}}{2fc}$$



$$V_{DC} = V_m - \frac{V_r}{2} = V_m - \frac{I_{DC}}{4fc}$$

$$V_{TH} = V_m, \quad R_{TH} = \frac{1}{4fc}$$

⇒ BRIDGE RECTIFIER WITH CAPACITOR FILTER -

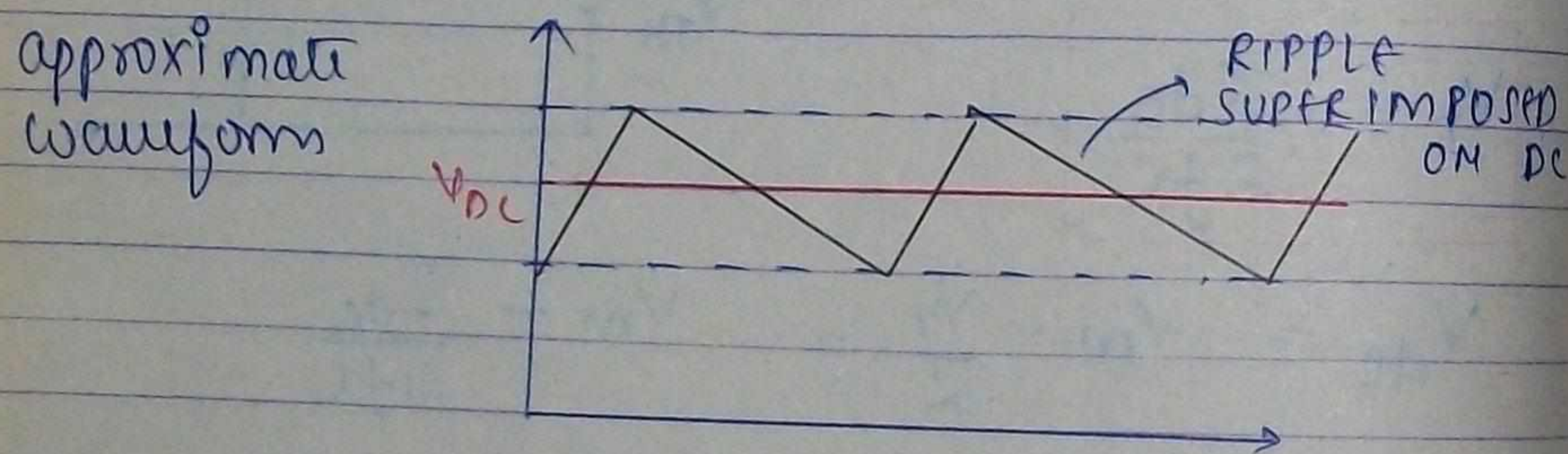
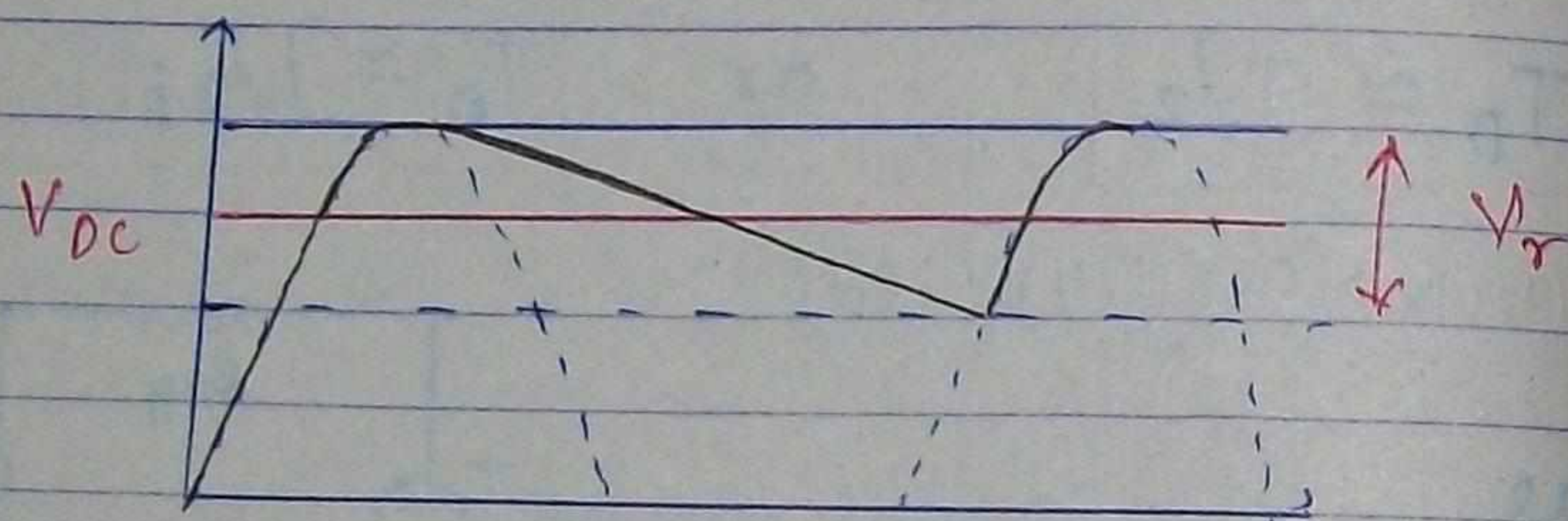


$$\tau = 2R_f C$$

Capacitor charges upto V_m
Same waveform and formula.

⇒ RIPPLE FACTOR -

① HALF WAVE RECTIFIER -



$$r = \frac{V_{acrms}}{V_{dc}}$$

p-p value of triangular wave = V_r

$$V_p = \frac{V_r}{2}$$

$$V_{acrms} = \frac{V_p}{\sqrt{3}} = \frac{V_r}{2\sqrt{3}}$$

$$r = \frac{V_r / 2\sqrt{3}}{V_{dc}} = \frac{I_{dc} / fc}{\frac{2\sqrt{3}}{I_{dc} \times R_L}}$$

Since $V_r = \frac{I_{dc}}{fc}$ and $V_{dc} = I_{dc} \times R_L$

$$r = \frac{1}{2\sqrt{3} f C R_L}$$

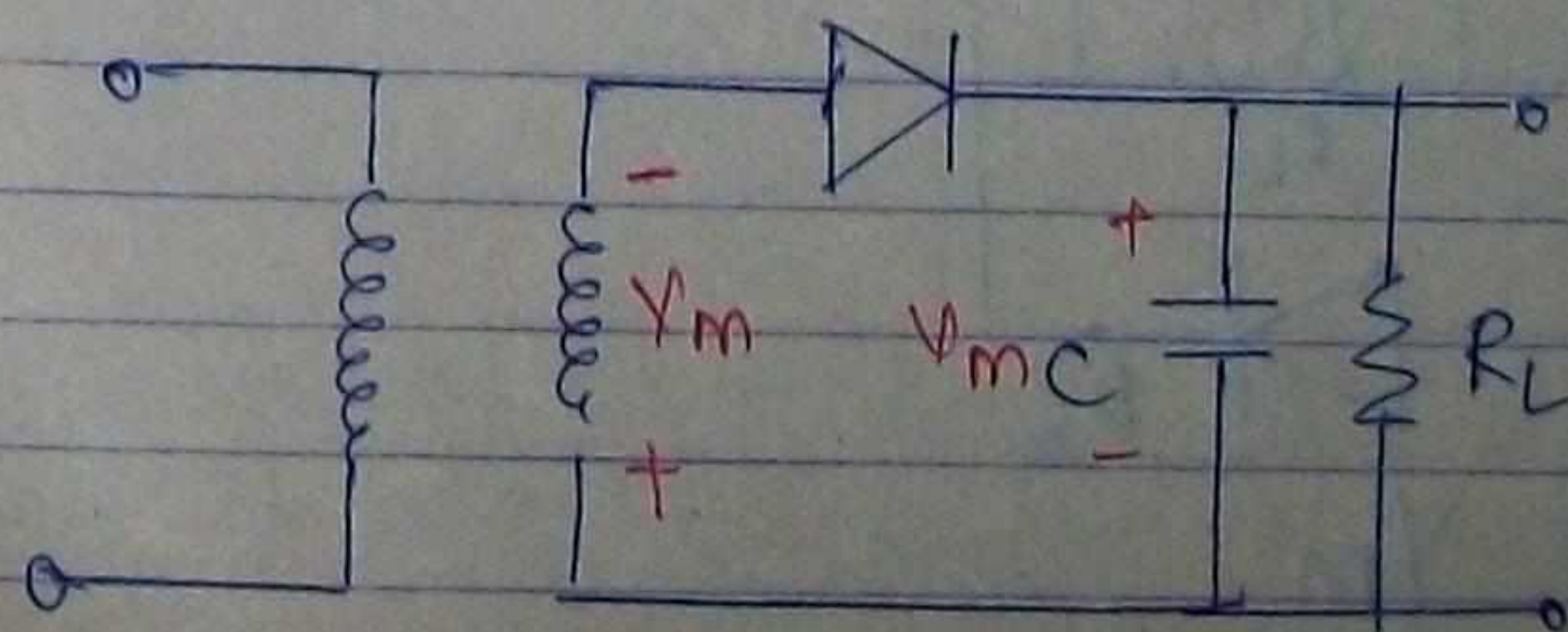
② FOR FULL WAVE RECTIFIER -

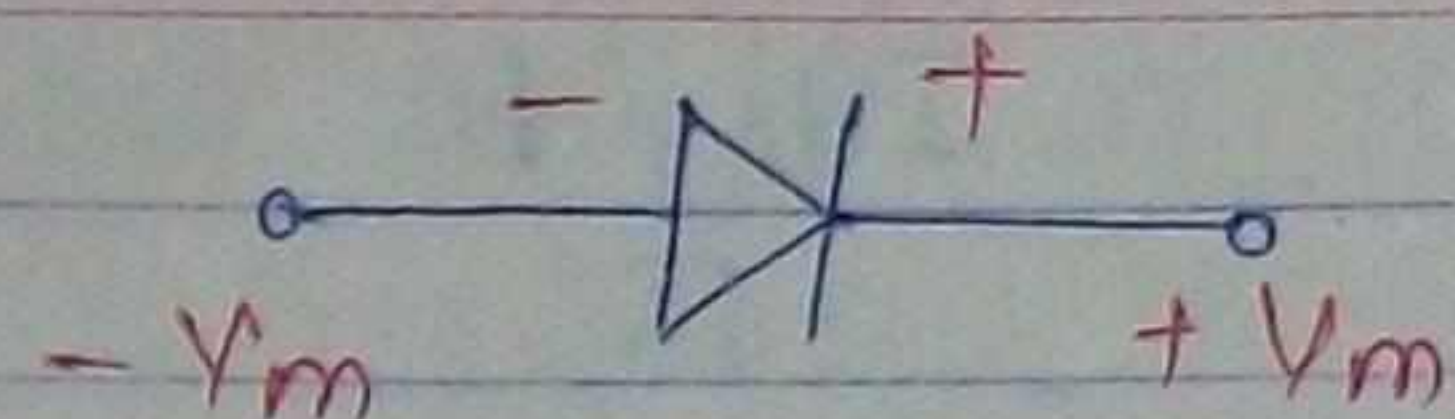
$$V_r = \frac{I_{dc}}{2fc} \quad \text{and} \quad V_{dc} = I_{dc} \times R_L$$

$$r = \frac{1}{4\sqrt{3} f C R_L}$$

⇒ PEAK INVERSE VOLTAGE WITH CAPACITOR FILTER -

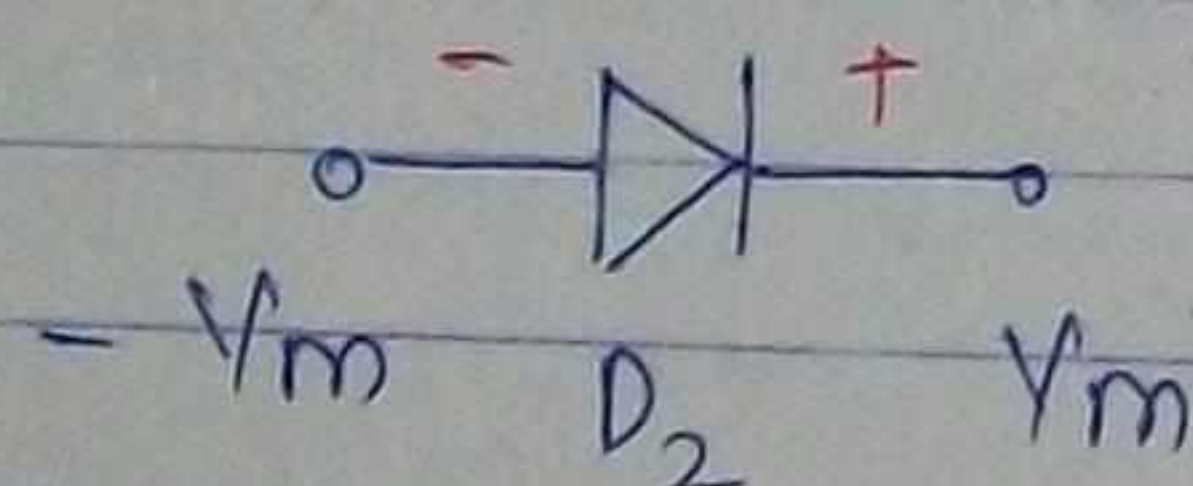
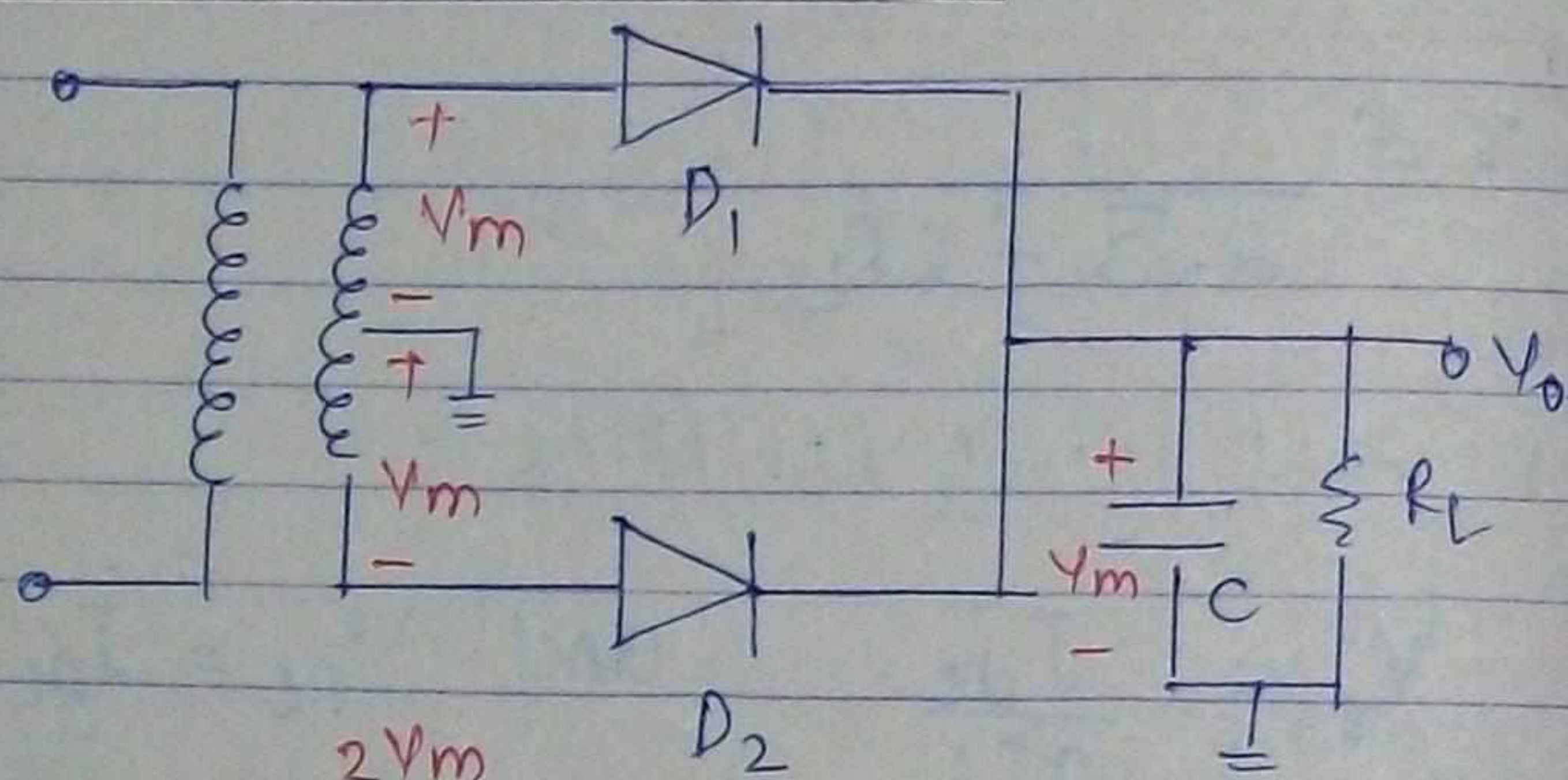
① FOR HALF WAVE RECTIFIER -





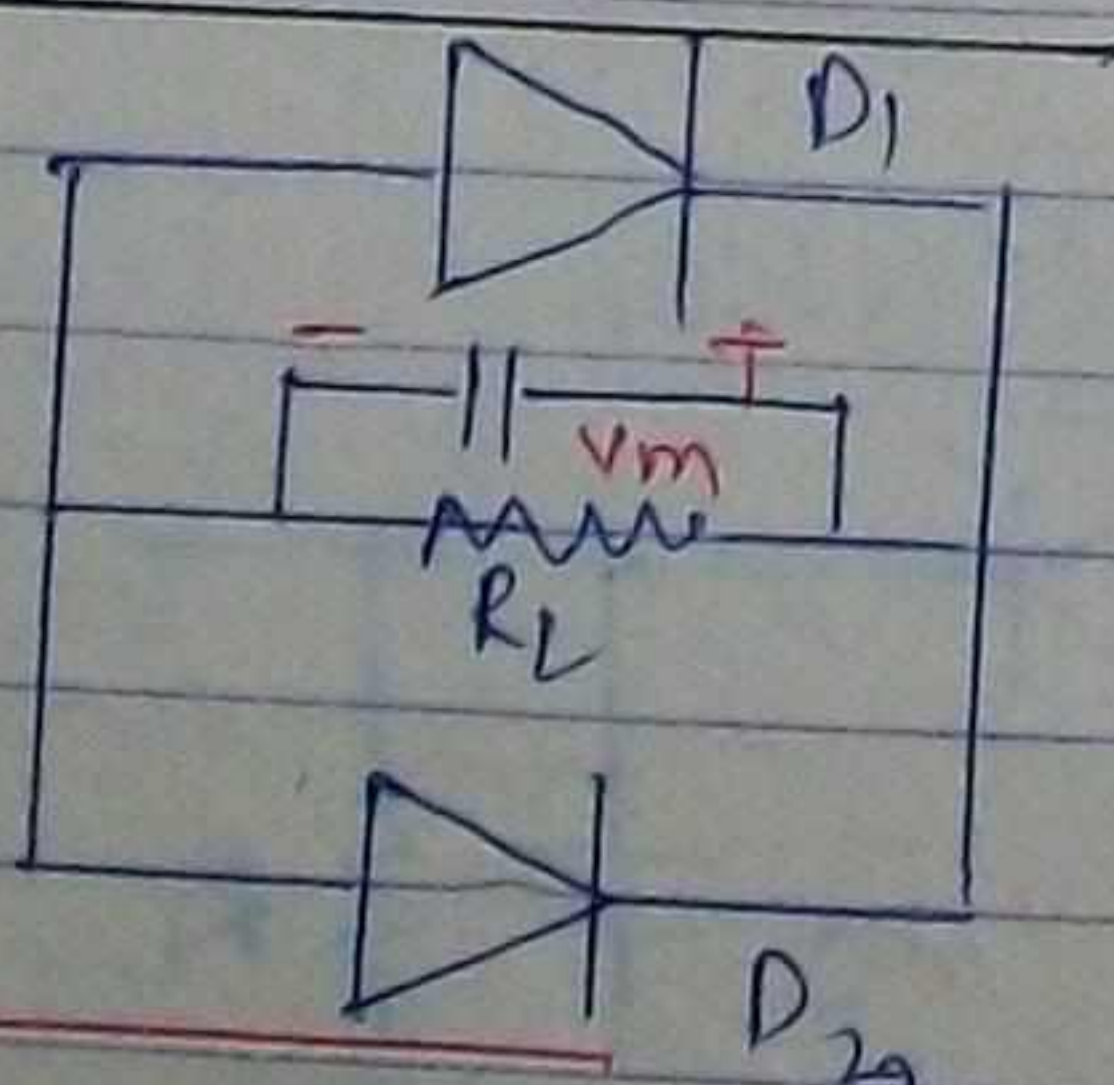
$$PIV = 2V_m$$

② FOR CENTRE TAP RECTIFIER -



$$PIV = 2V_m$$

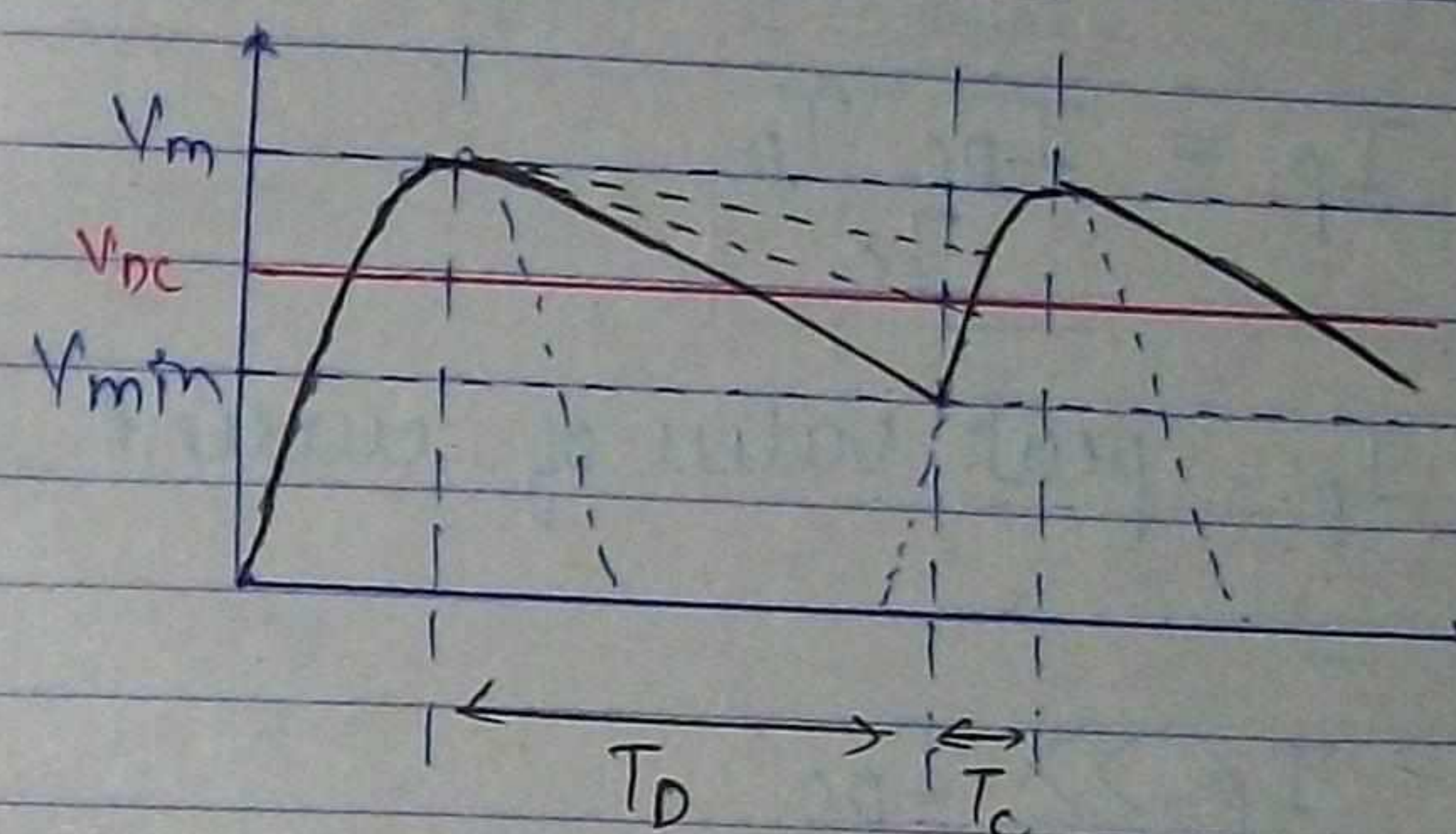
③ FOR BRIDGE RECTIFIER -



$$PIV = V_m$$

⇒ PEAK DIODE CURRENT OR SURGE CURRENT -

① FOR HALF WAVE RECTIFIER -



During T_D , $C \rightarrow$ discharge

$$Q(\text{discharge}) = I_{dc} \times T_D \quad \text{--- (1)}$$

Change in voltage $V_r = V_m - V_{min}$

During T_c , $C \rightarrow$ charging

$$Q(\text{charging}) = I_p \times T_c \quad \text{--- (2)}$$

Then from the law of Conservation of charges:

$$Q(\text{charge}) = Q(\text{discharge})$$

$$I_p \times T_c = I_{DC} \times T_D$$

$$I_p = \frac{I_{DC} T_D}{T_c}$$

where I_p = peak value of current.

thus:

$$I_p \gg I_{DC}$$

$$C \uparrow, T_D \uparrow, T_c \downarrow, r \downarrow, I_p \uparrow$$

Ex -

$$\text{let } f = 50 \text{ kHz} \quad T = 1/f = 20 \text{ msec}$$

$$T_D + T_c = 20 \text{ ms}$$

$$T_D \gg T_c$$

$$19.98 + 0.02$$

$$\text{let } I_{DC} = \frac{V_{OC}}{R_L} \approx \frac{V_m}{R_L} = \frac{10}{10k} = 1 \text{ mA}$$

$$I_p = 1 \text{ mA} \times \frac{19.98}{0.02} \approx 1 \text{ Amp}$$

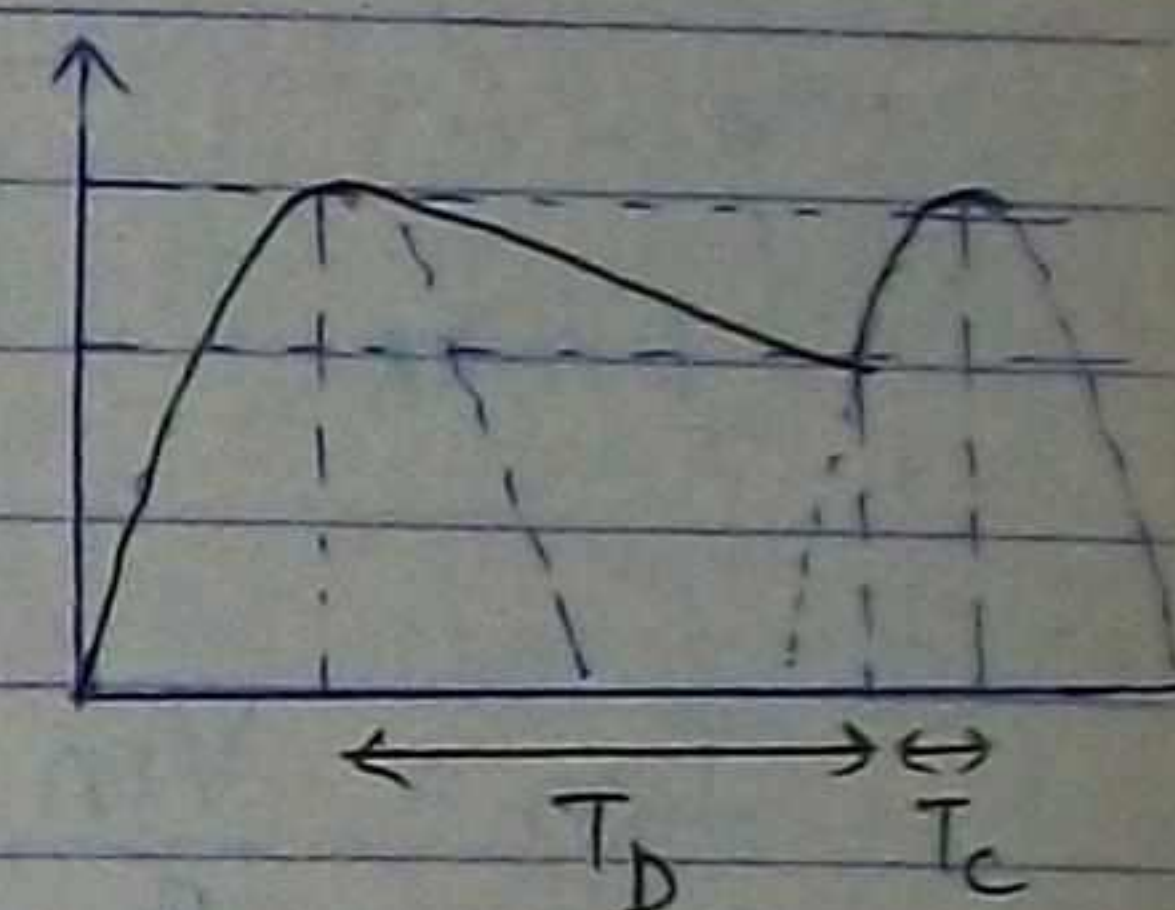
$$I_p = \frac{I_{DC} T_D}{T_c} \approx \frac{I_m T_D}{T_c} \approx \frac{I_m T}{T_c}$$

⇒ MAXIMUM DIODE CURRENT -

① HALF WAVE RECTIFIER -

$$V_m \sin \omega t$$

$$\omega = \frac{d\theta}{dt}$$



CONDUCTION ANGLE -

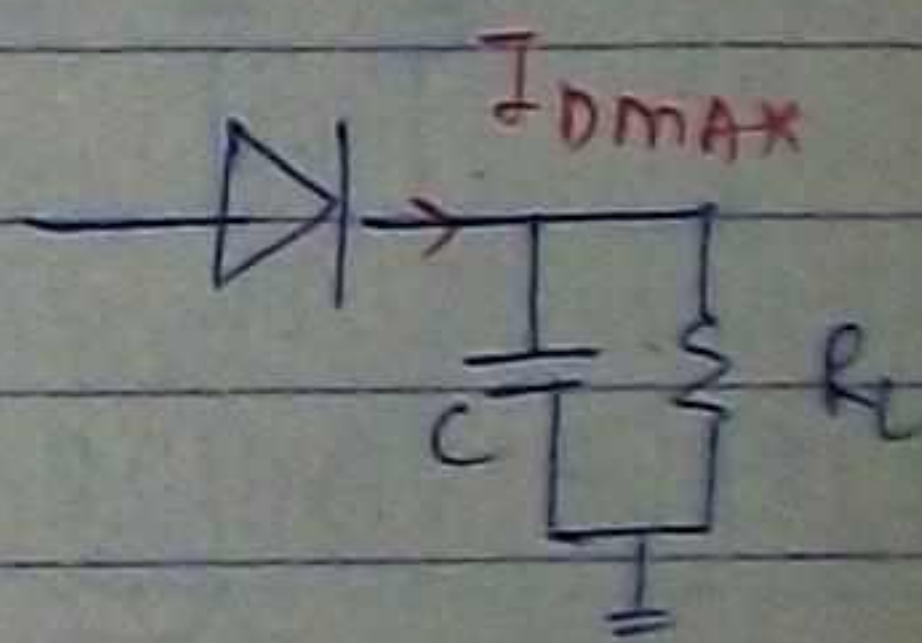
$$\phi = \omega T_c$$

$$\phi = \sqrt{\frac{2V_r}{V_m}}$$

$$V_r = \frac{I_{DC}}{fC} \approx \frac{I_m}{fC}$$

I_{Dmax} -

$$I_{Dmax} = I_m \left[1 + 2\pi \sqrt{\frac{2V_m}{V_r}} \right]$$



② FULL WAVE RECTIFIER -

CONDUCTION ANGLE -

$$\phi = \omega T_c = \sqrt{\frac{2V_r}{V_m}}$$

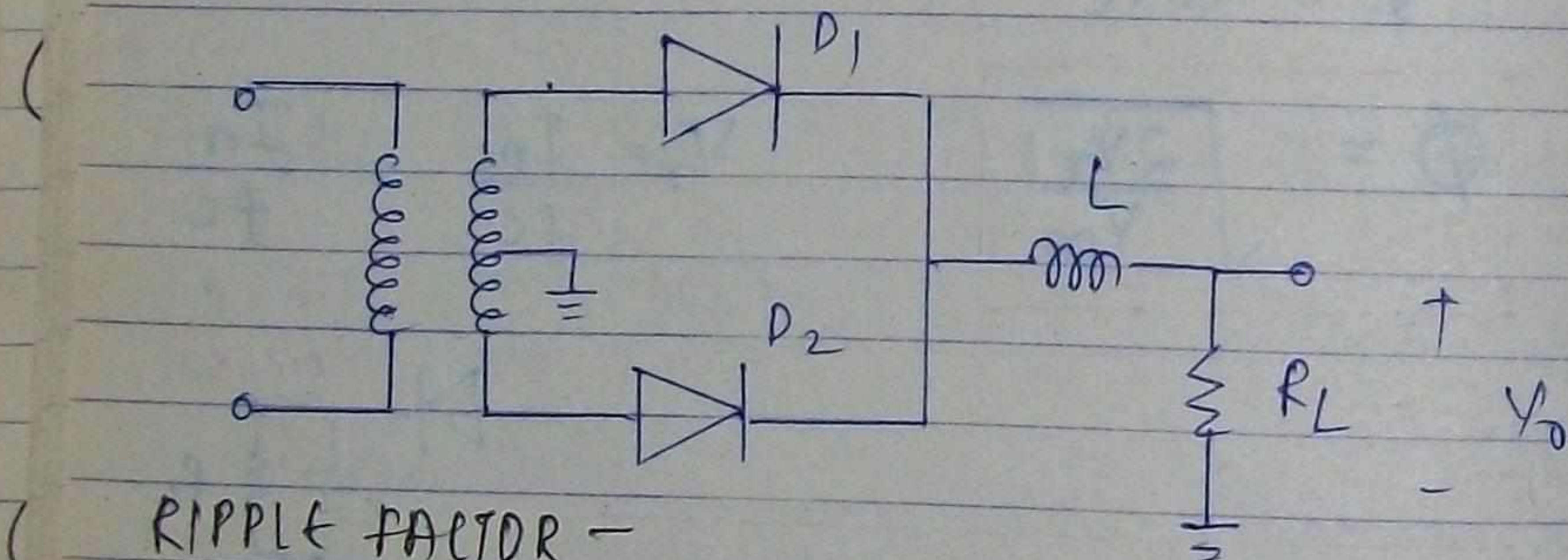
$$V_r = \frac{I_{dc}}{2fc} \approx \frac{I_m}{2fc}$$

$$I_{Dmax}$$

$$I_{Dmax} = I_m \left[1 + 2\pi \sqrt{\frac{V_m}{2V_r}} \right]$$

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

⇒ INDUCTOR FILTER [CHOKER FILTER] -



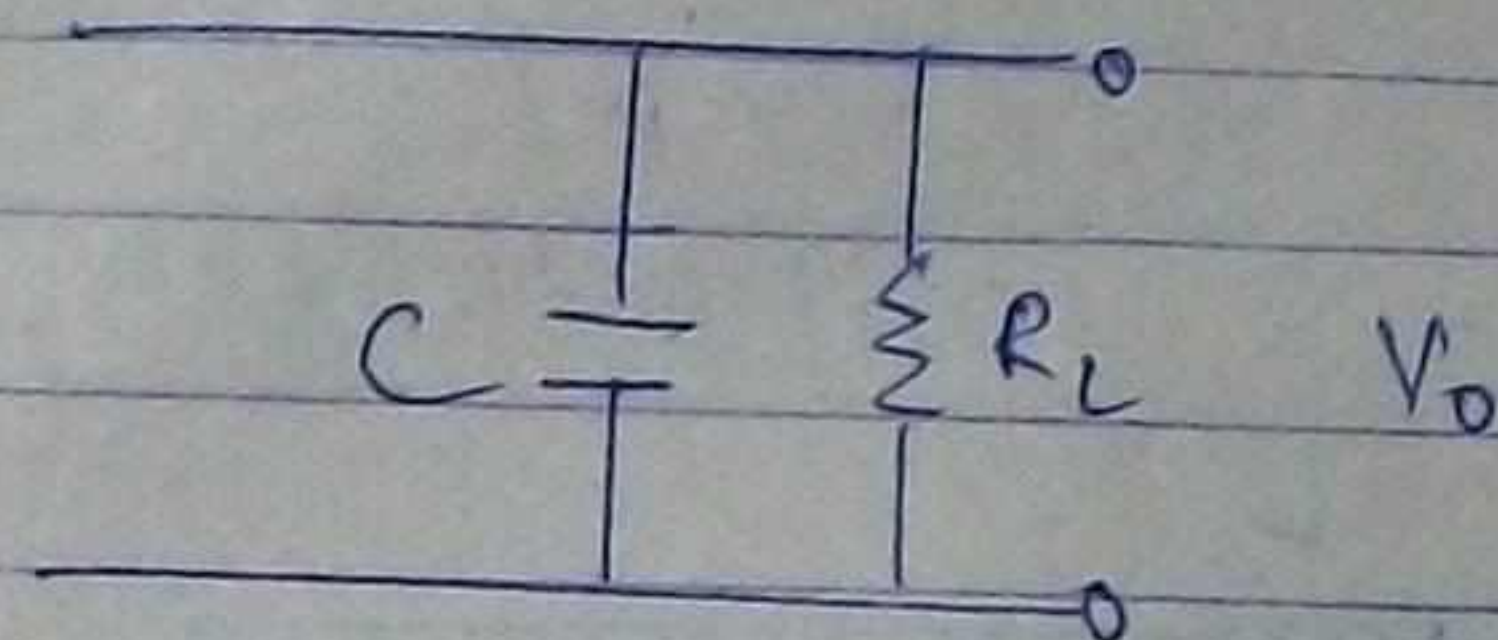
RIPPLE FACTOR -

$$\gamma = \frac{2}{3\sqrt{2}} \times \frac{1}{\sqrt{1 + \left(\frac{X_L}{R_L}\right)^2}}$$

$$X_L = \omega L \text{ for HWR}$$

$$X_L = 2\omega L \text{ for FWR}$$

C-FILTER



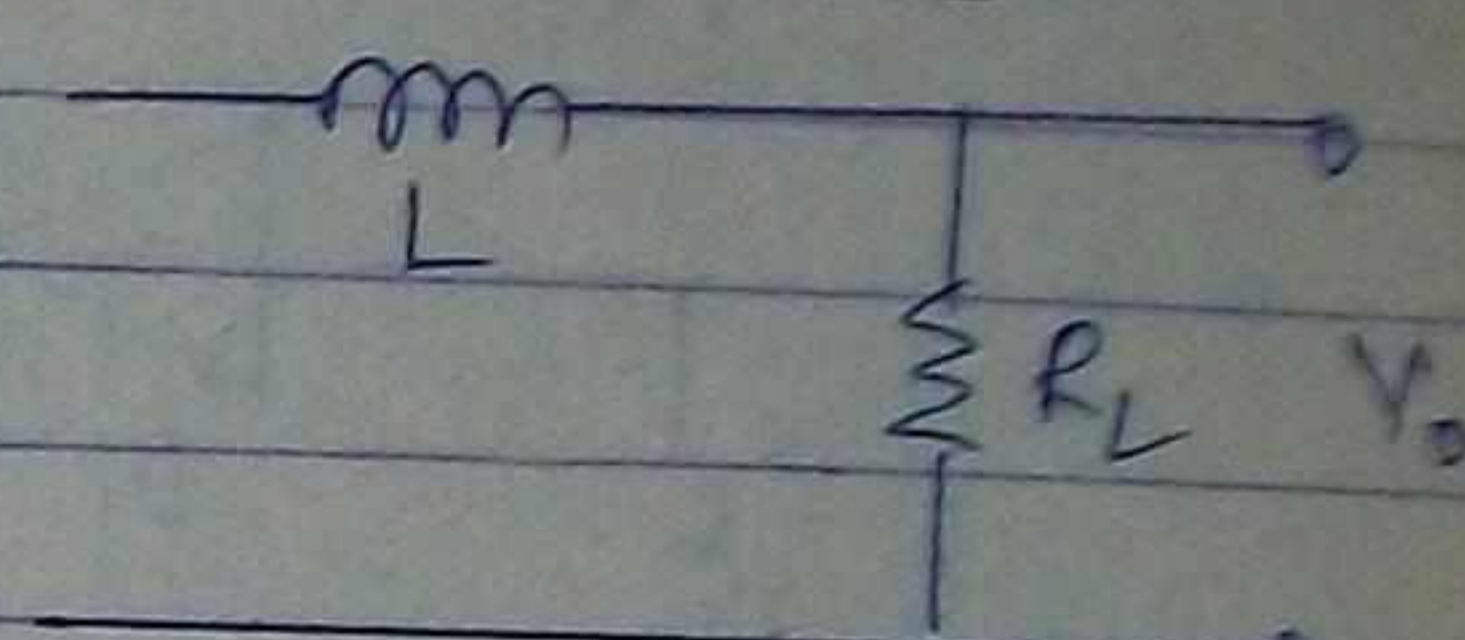
$$\tau = R_L C = \text{V.V high}$$

∴ Variation in V_0 = V.V low

$R_L \rightarrow \text{HIGH}$ $C \rightarrow \text{HIGH}$

$\tau \rightarrow \text{HIGH}$ $\gamma \rightarrow \text{LOW}$

L-FILTER



$$\tau = L/R_L = \text{V.V. HIGH}$$

Variation in I_0 = V.V. Low

$L \rightarrow \text{HIGH}$ $R_L \rightarrow \text{LOW}$

$\tau \rightarrow \text{HIGH}$ $\gamma \rightarrow \text{LOW}$

CONCLUSION -

for low load resistance L filter is preferable

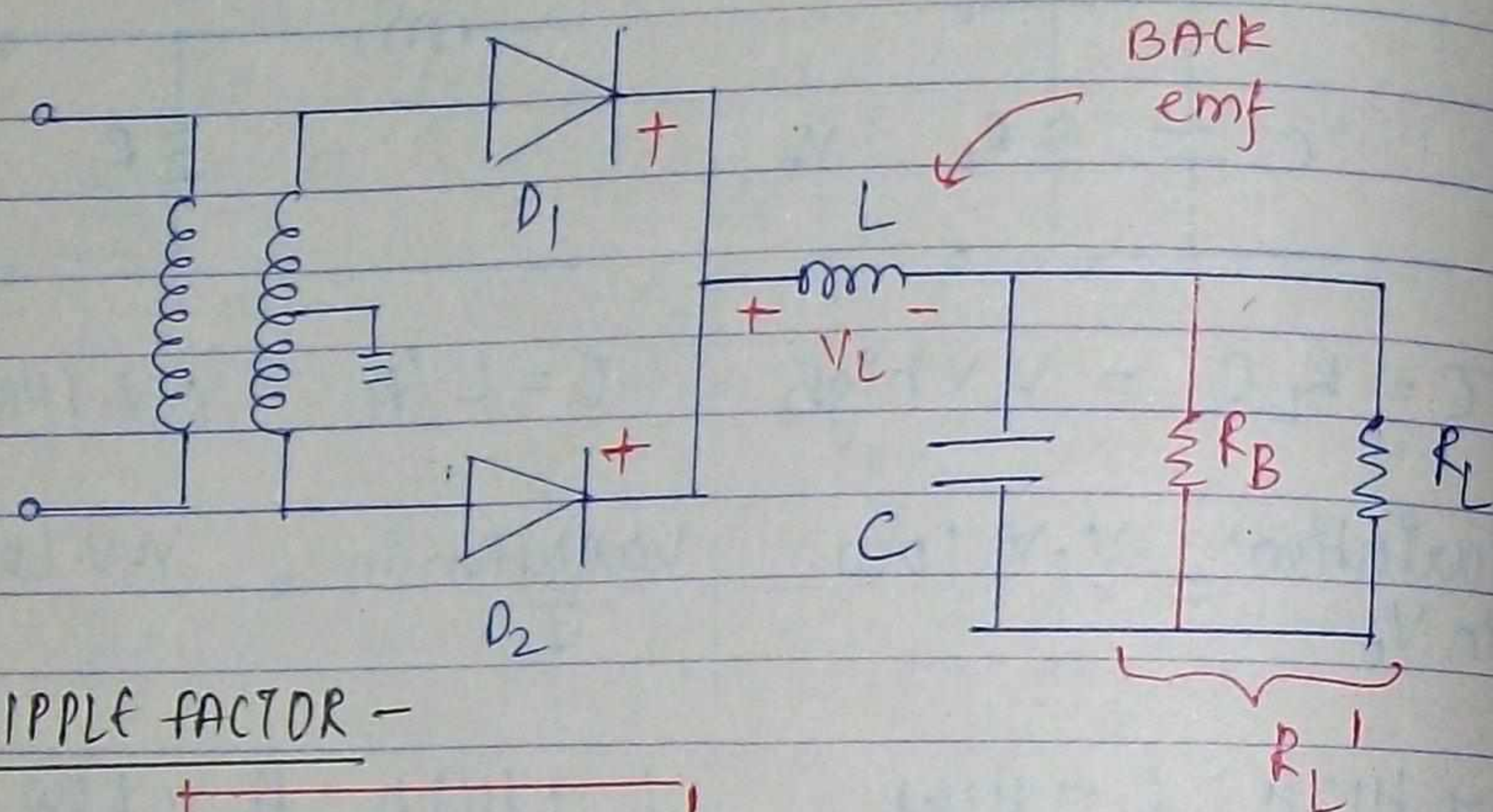
for high load resistance C filter is preferable

also from ①

$$\gamma \propto \frac{R_L}{\omega L} = \frac{R_L}{2\pi f L} = \frac{1}{2\pi f L/R_L} = \frac{1}{2\pi f \tau}$$

$$\gamma \propto \frac{1}{\tau} \text{ and } \gamma \propto \frac{1}{f}$$

⇒ L-SECTION FILTER [LC FILTER] -



RIPPLE FACTOR -

$$r = \frac{\sqrt{2}}{3} \frac{X_C}{X_L}$$

($X_C = \frac{1}{\omega C}$, $X_L = \omega L$ for HWR

$X_C = \frac{1}{2\omega C}$, $X_L = 2\omega L$ for FWR

(① $r = \frac{\sqrt{2}}{3} \times \frac{1}{\omega^2 LC}$

r is independent of R_L

Since, $r \propto \frac{1}{\tau_L \tau_C} = \frac{1}{LC}$

$\tau_L = \frac{L}{R_L}$ $\tau_C = R_L C$ $\tau_L \tau_C = LC$

②

$$r \propto \frac{1}{f^2}$$

Now when R_B is placed in circuit

$R_B \rightarrow$ Bleeder Resistance. $R_B \gg R_L$

① Inductor does not allow sudden change in current

$$V_L = L \frac{dI}{dt}$$

$$\frac{dI}{dt} = \text{very high}$$

$$V_L = \text{very high}$$

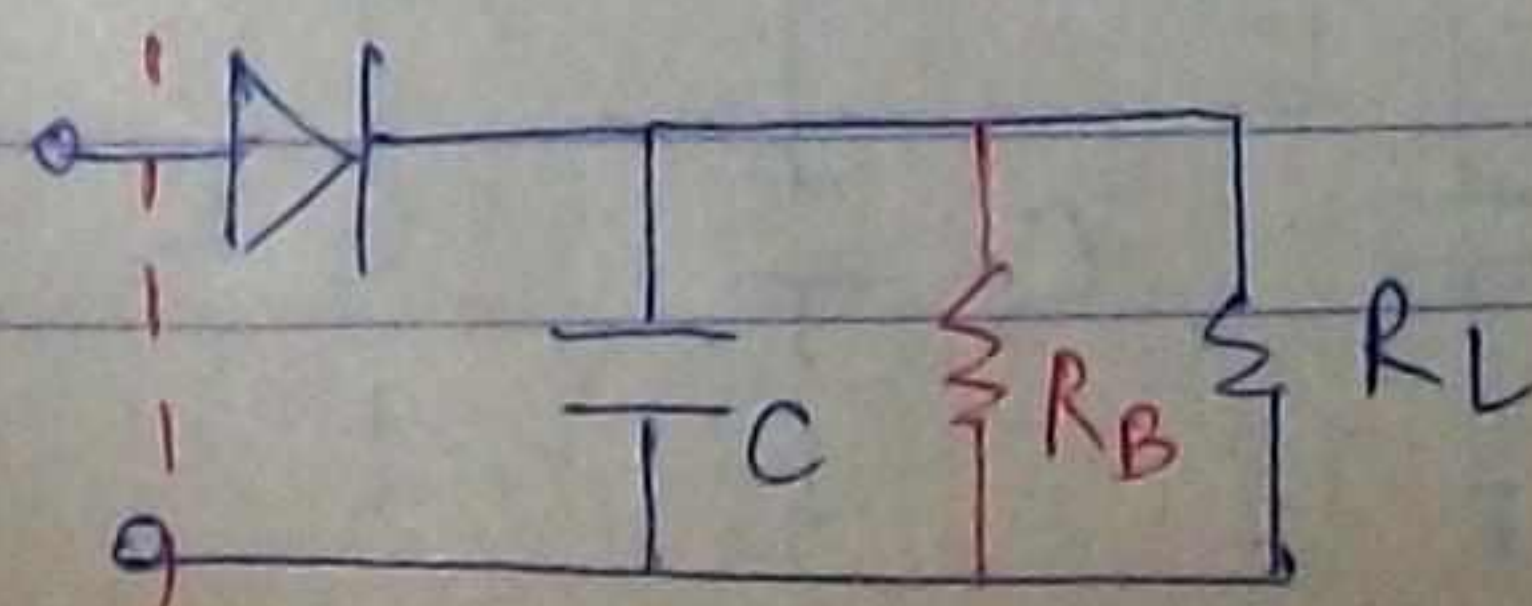
$$\tau = \frac{L}{R_L} \approx 0$$

$$R_L = \infty$$

without R_B , $\tau \approx 0$

with R_B , $\tau = L/R_B$

②



affect ripple

← Intermittent

← Continuous

* Here R_L does not affect ripple factor

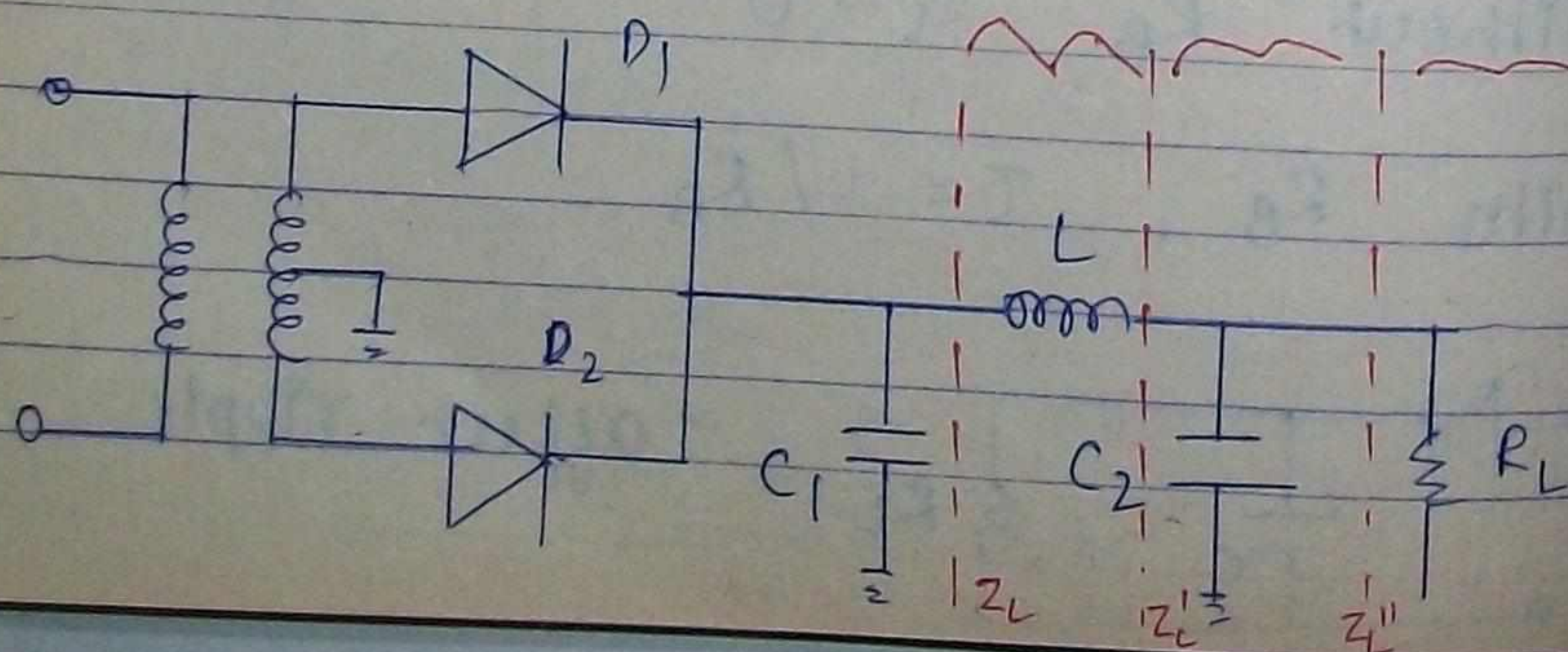
- The basic requirement of this filter is that the current through the choke is continuous and not interrupted. An interrupted current may develop a large back emf which may be in excess of PIV rating of diodes and/or max. rating of capacitor.

To eliminate back emf the current through the choke must be maintained continuous. This is assured by connecting R_B across output terminals.

- Another reason for R_B is to bleed off voltage stored in filter capacitor when the supply is turned off.

for high voltage supplies R_B is essential.

⇒ π-FILTER - (CLC FILTER)



RIPPLE FACTOR -

$$\gamma = \frac{\sqrt{2} X_C X_L}{R_L X_L} = \frac{\sqrt{2} (X_C)^2}{R_L X_L}$$

If $C_1 = C_2 = C$

$$X_C = \frac{1}{2\omega C} \quad X_L = 2\omega L \quad \text{for FWR}$$

$$X_C = \frac{1}{\omega C} \quad X_L = \omega L \quad \text{for HWR}$$

$$\gamma \propto \frac{1}{\omega^3 C_1 C_2 L R_L}$$

$$\gamma \propto \frac{1}{f^3}$$

ripple factor → very very low (least).

Since,

$$\gamma \propto \frac{1}{T_1 T_2 T_3}$$

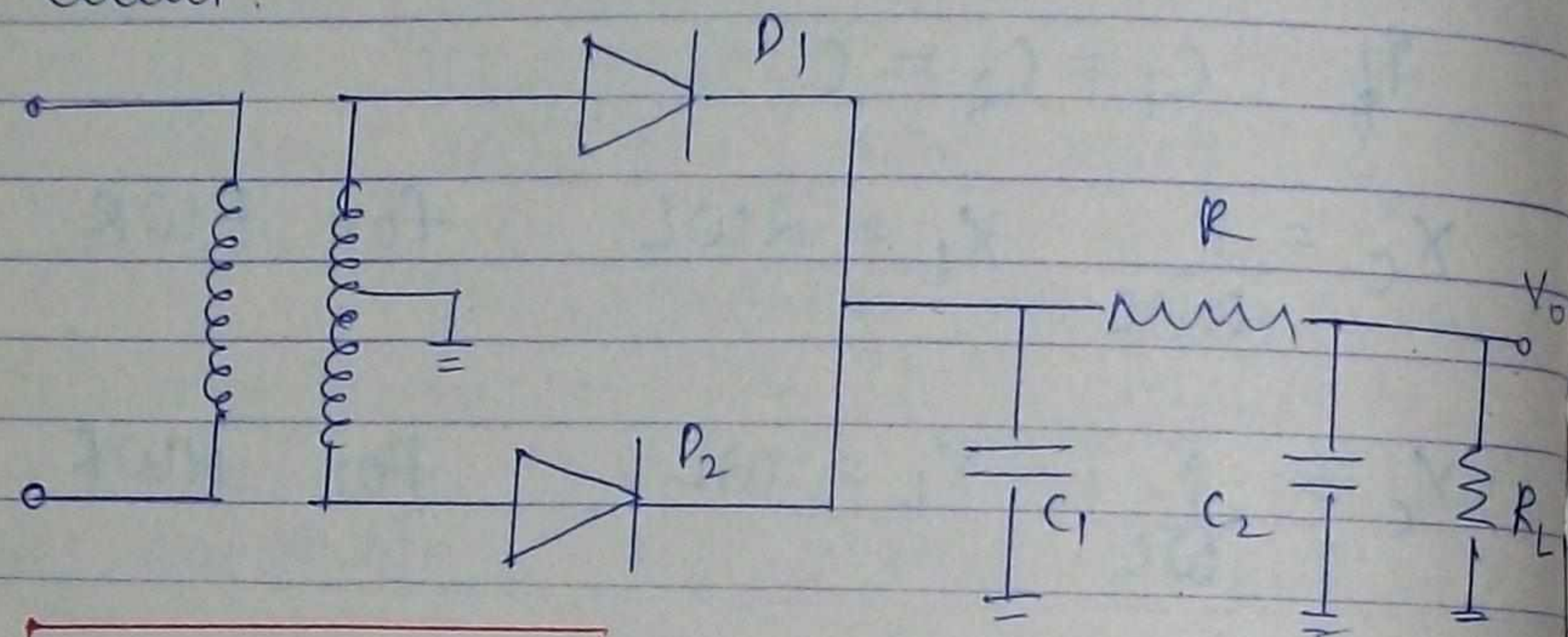
$$T_1 = R_L C_1 \quad T_2 = L/R_L \quad T_3 = R_L C_2$$

$$R_L C_1 C_2 L$$

$$\gamma \propto \frac{1}{R_L C_1 C_2 L}$$

⇒ Π FILTER (OR) CRC FILTER -

Since size of L is very large hence L is replaced by R for compact circuit.



$$\gamma = \frac{\sqrt{2} X_{L1} X_{C2}}{R_L R}$$

If $C_1 = C_2 = C$

$X_C = \frac{1}{\omega C}$ for HWR

$X_C = \frac{1}{2\omega C}$ for FWR

$$\gamma \propto \frac{1}{f^2}$$

γ is very low

Q. CONVENTIONAL Q.1.

Solⁿ: $V_{DC} = 30 \text{ V}$ $R_L = 500 \Omega$ $\gamma \leq 0.01$
 $f = 50 \text{ Hz}$

$C = ?$ $I_p = I_{Dmax} = ?$

$$\gamma = \frac{1}{2\sqrt{3} f C R_L} \leq 0.01$$

$$C \geq \frac{1}{2\sqrt{3} f \times R_L \times 0.01}$$

$$C \geq 1154 \mu\text{F}$$

$$I_{Dmax} = I_m \left[1 + 2\pi \sqrt{\frac{2V_m}{V_r}} \right]$$

$$I_m = \frac{V_m}{R_L} \approx \frac{V_{DC}}{R_L} \quad (\gamma = \text{very low})$$

$$V_m \approx V_{DC} \quad V_r = \frac{I_{DC}}{fC}$$

$$I_{DC} = \frac{V_{DC}}{R_L}$$

also, $V_{DC} = V_m - V_r/2$

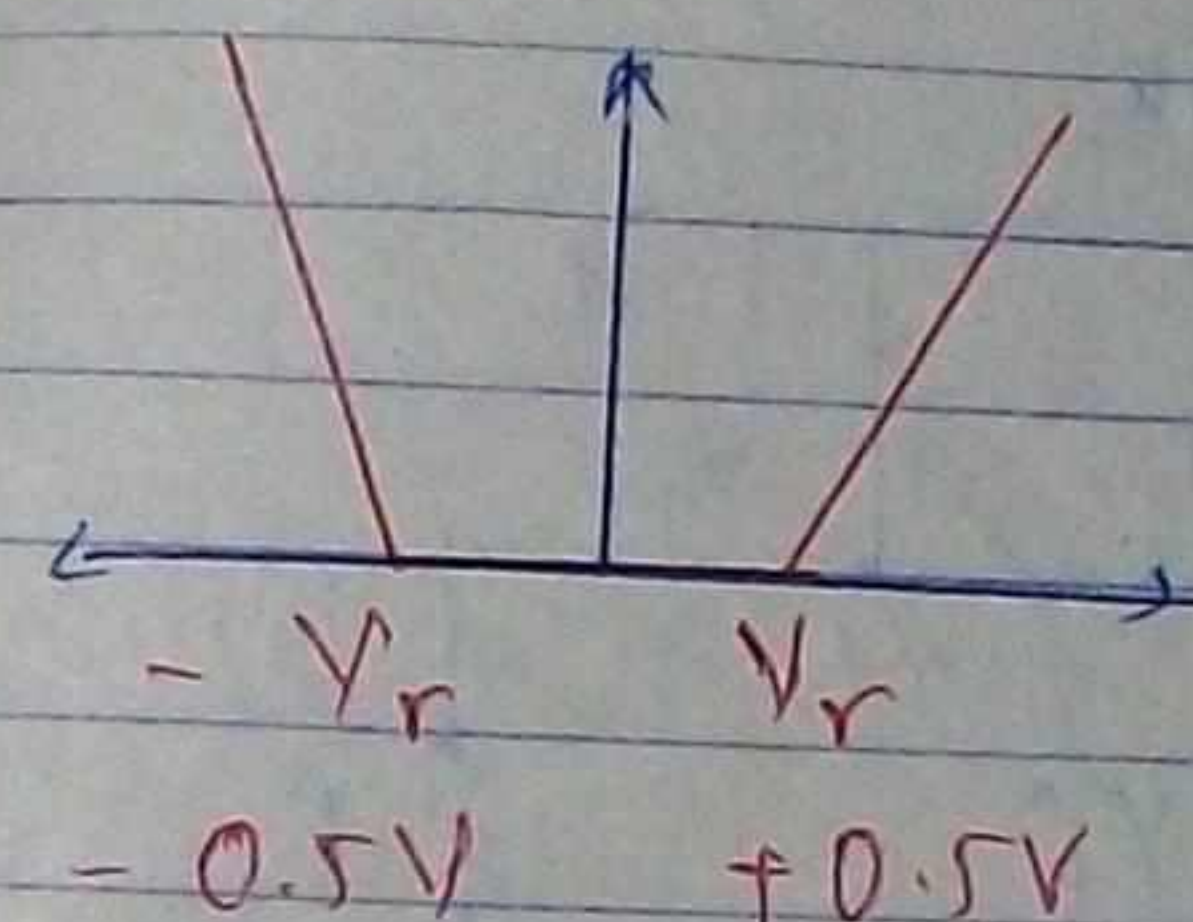
Q. Conventional Q2.

Solⁿ:

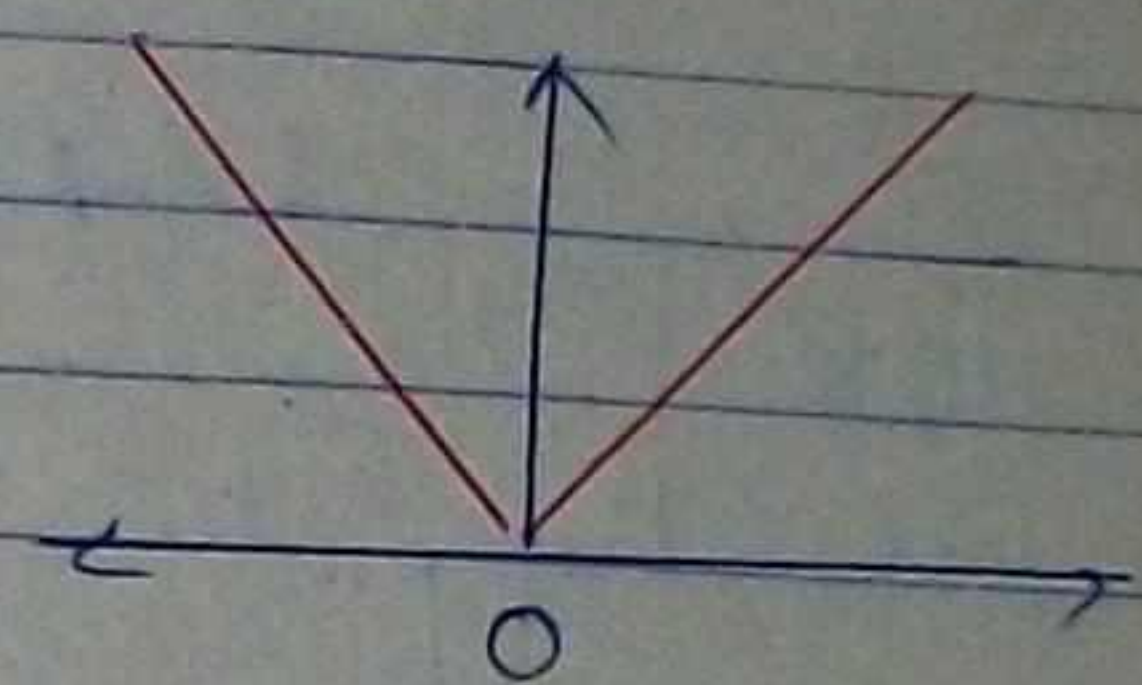
sine transformer is not given

$$V_m = 240\sqrt{2} \quad R_L = 10k\Omega$$

⇒ PRECISION RECTIFIER -

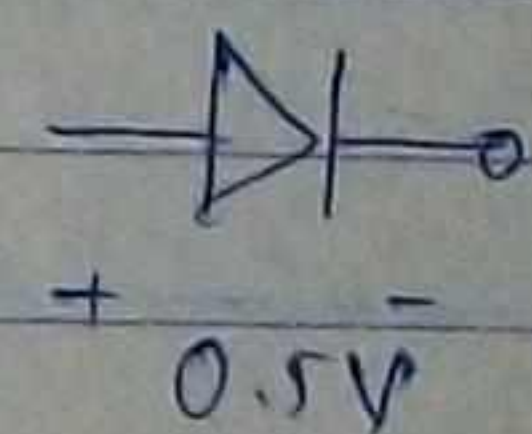
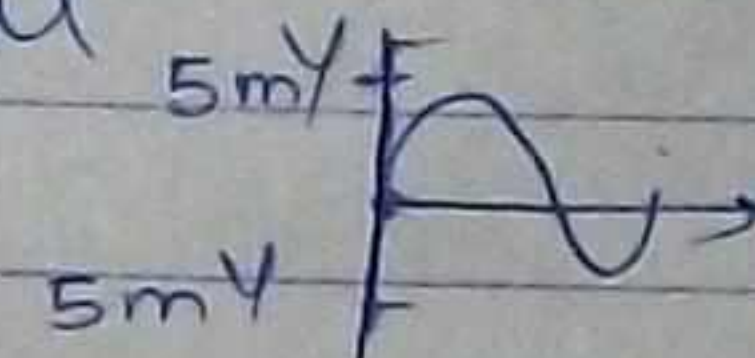


practical rectifier
transfer curve

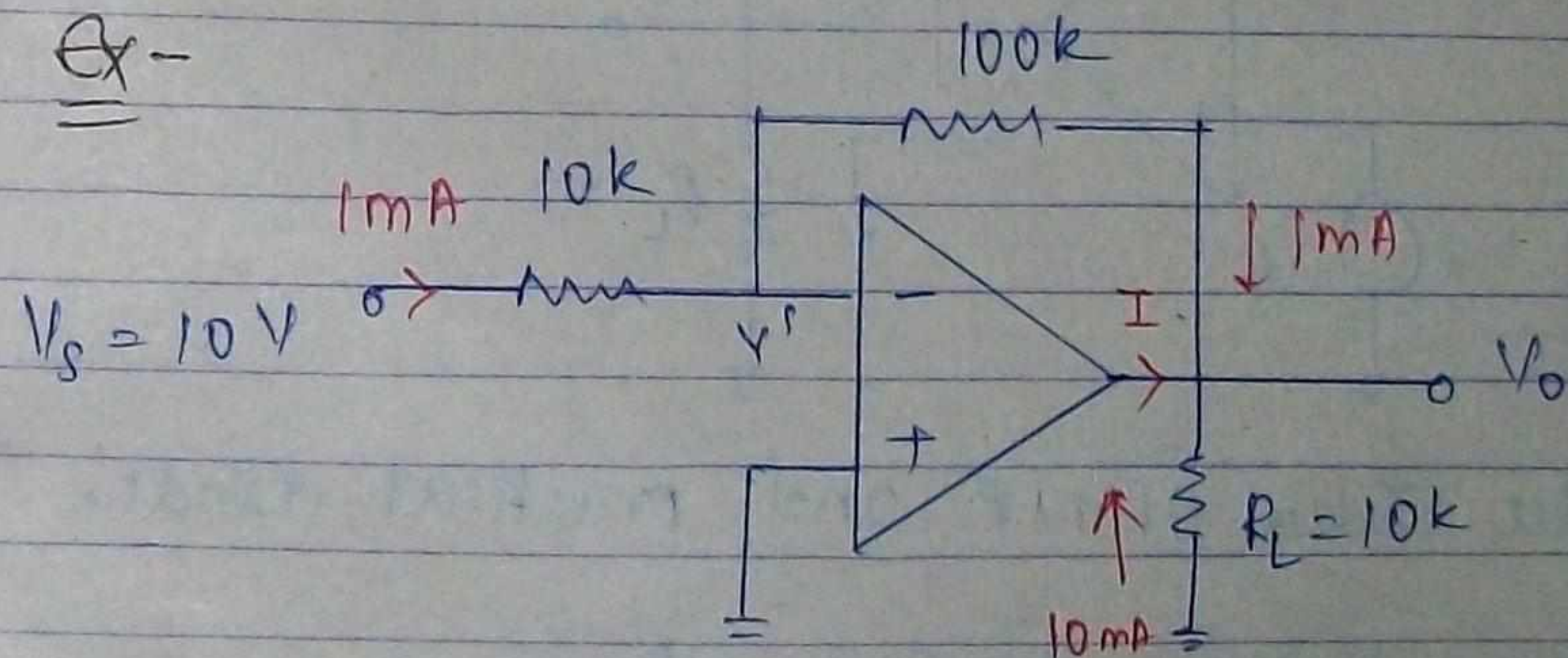


precision
rectifier

does not rectify small
signal



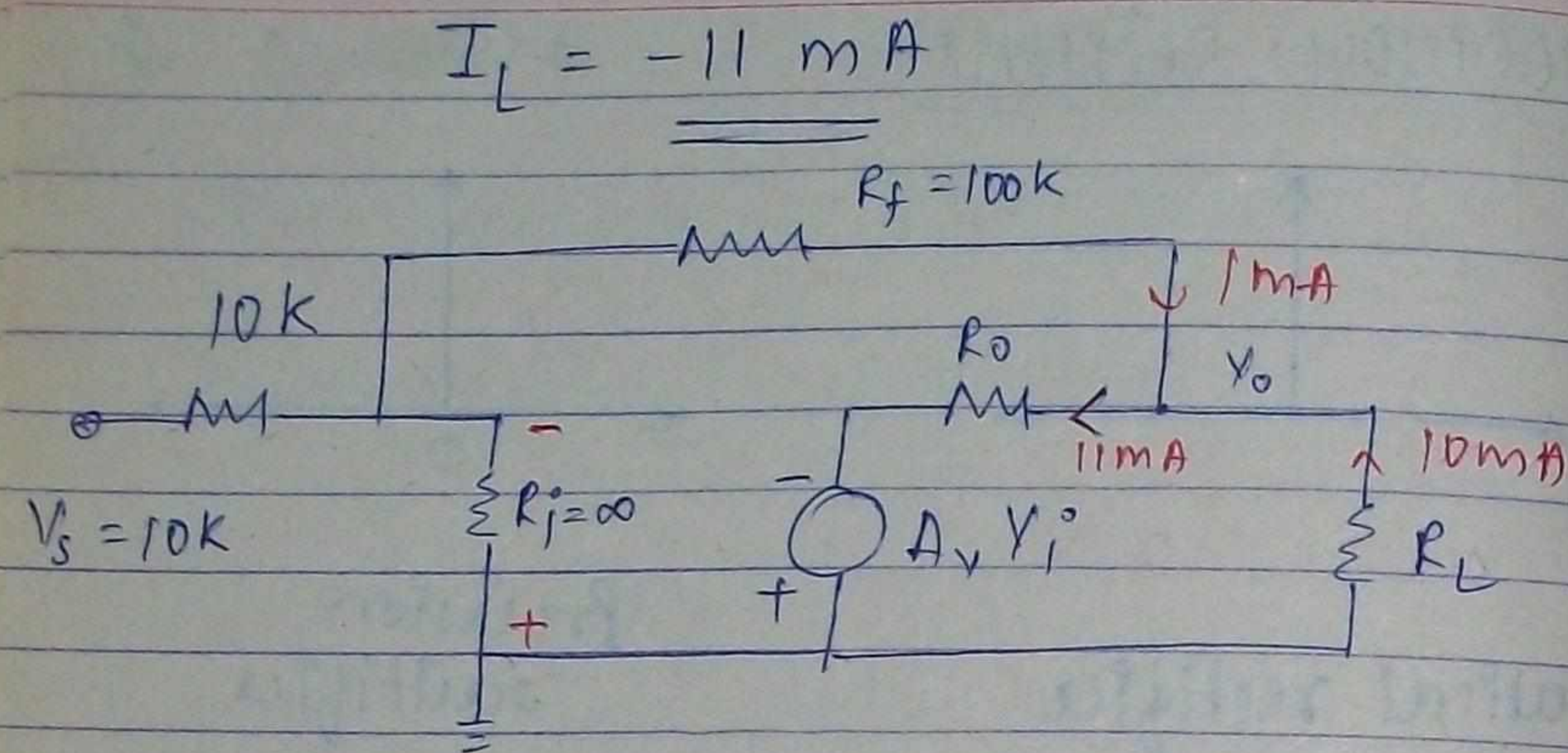
Ex -



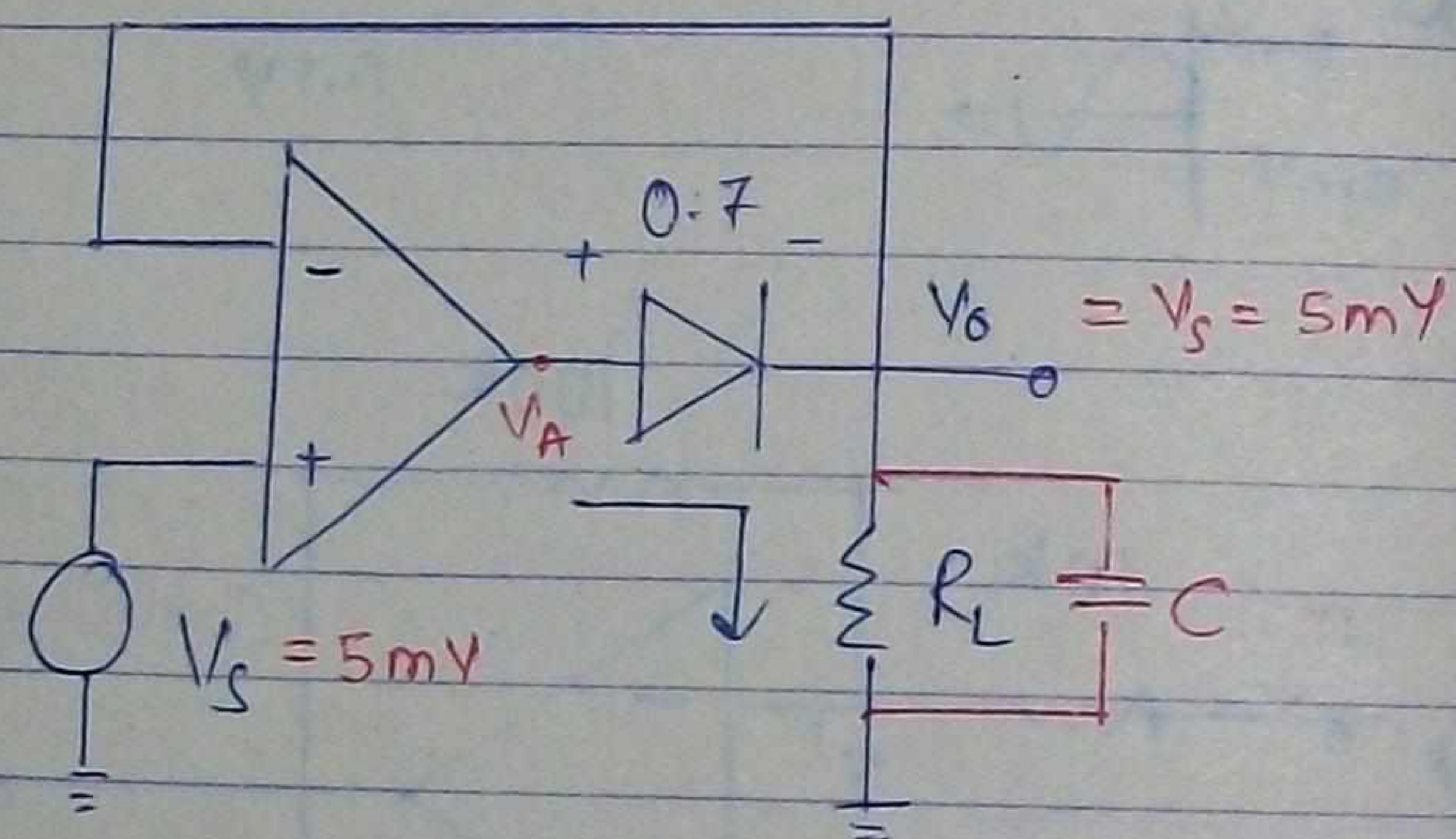
$$\frac{0-10}{10} + \frac{0-V_o}{100} = 0 \quad V_o = -100$$

$$R_L = 10k$$

$$I_L = \frac{V_o}{R_L} = -10mA$$



⇒ PRECISION HALF WAVE RECTIFIER -



Consider Ideal OPAMP and practical diode:

① when $V_s > 0$ $D \rightarrow ON$

$V_A \uparrow$ toward $+V_{SAT}$

But diode will saturate ^{not allow} O/P to become $+V_{SAT}$ Due to -ve f-B (closed loop)

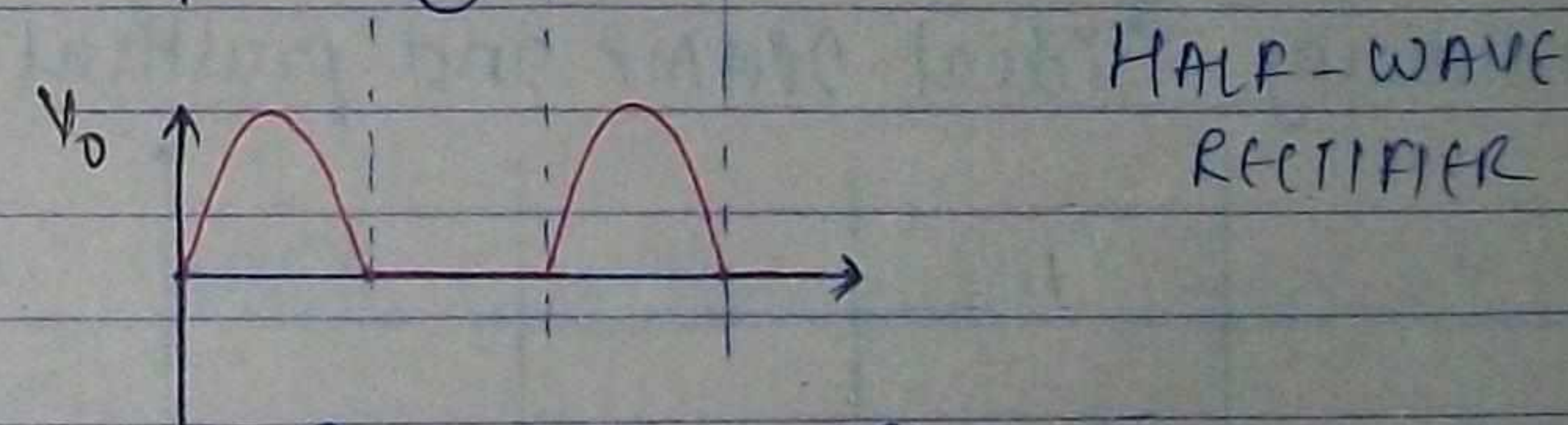
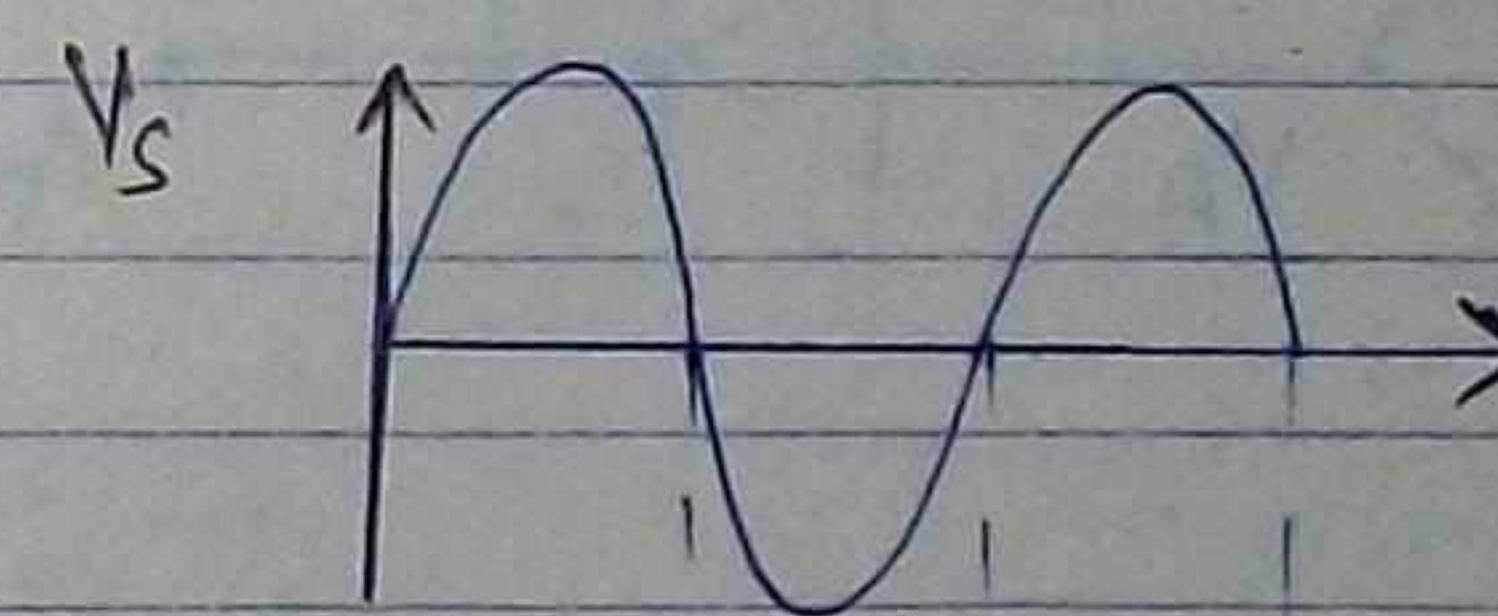
② when $V_s < 0$ $D \rightarrow OFF$

$V_A \downarrow$ toward $-V_{SAT}$

V_s	D	V_A	V_o
$V_s > 0$	ON	$0.7 + V_s$	V_s
$V_s < 0$	OFF	$-V_{SAT}$	0

NOTE -

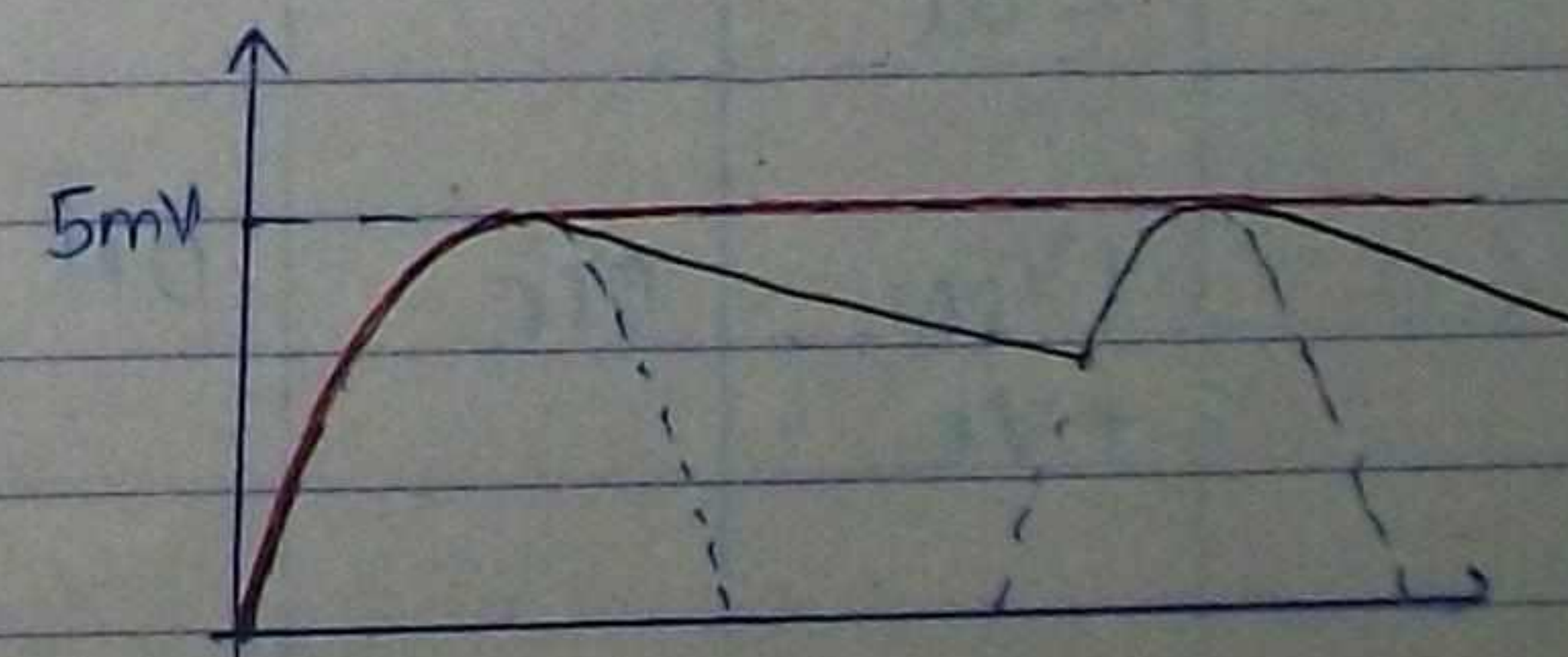
Diode will avoid the saturation of OPAMP.



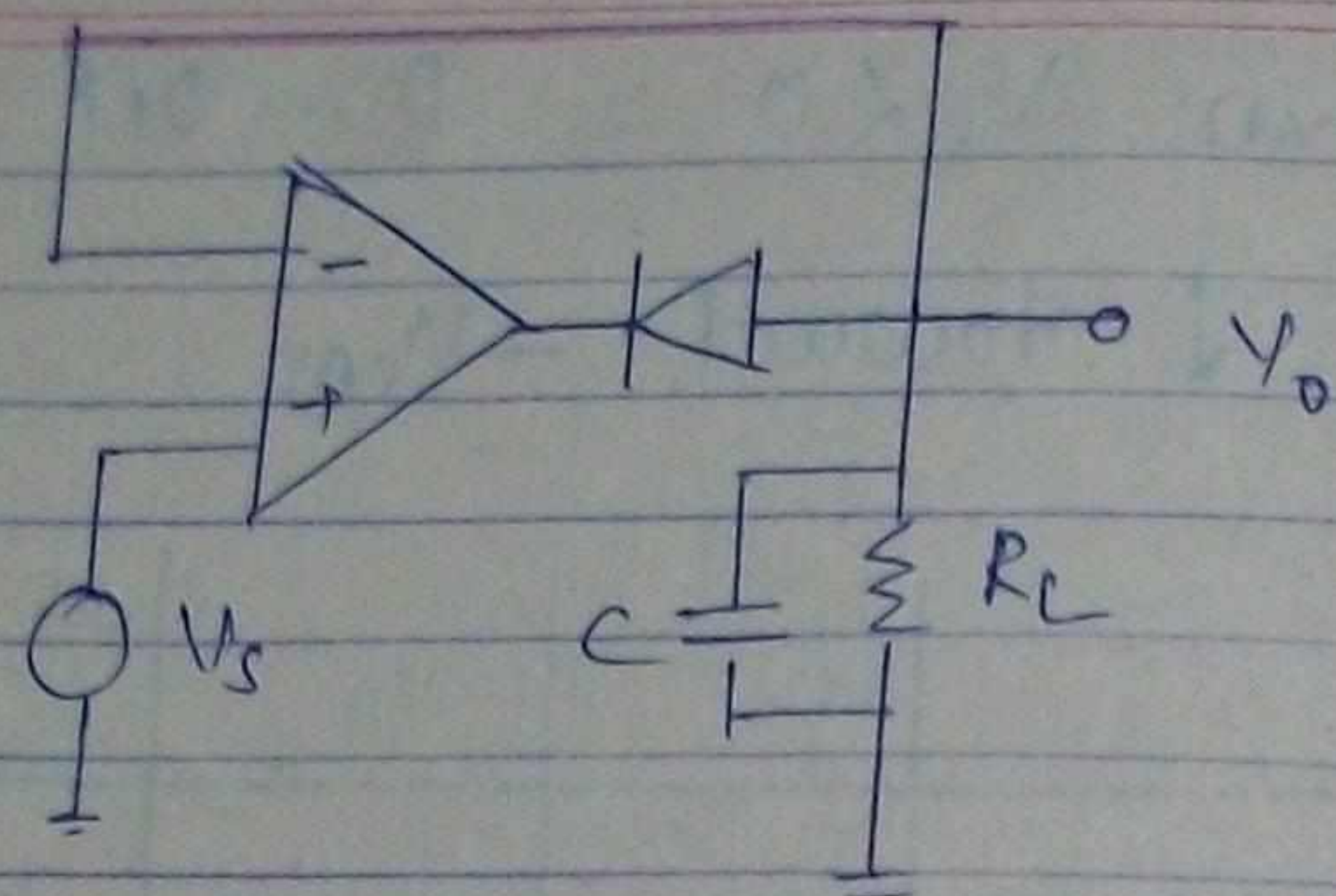
when capacitor is applied.

$R_i = \infty$

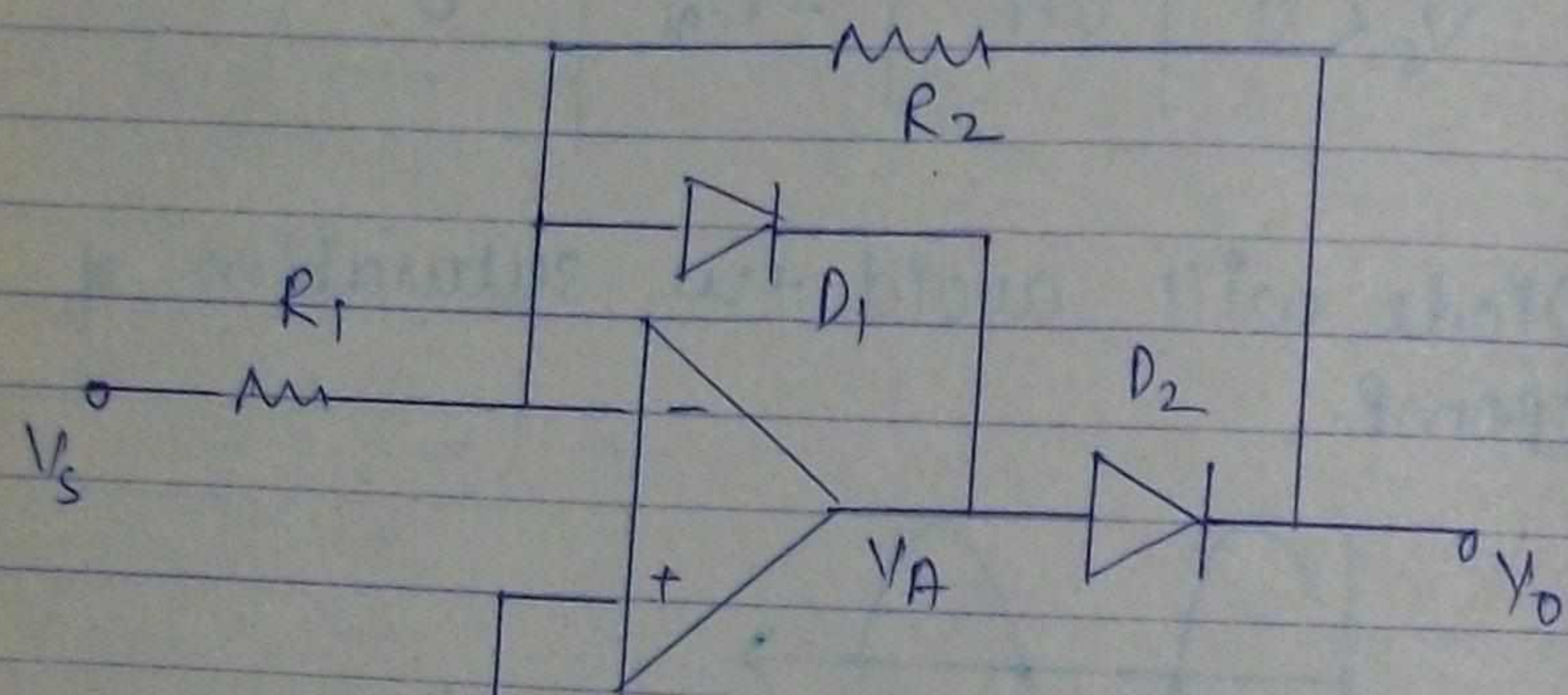
positive
Peak
Detector



NEGATIVE
PEAK
DETECTOR



⇒ INVERTING HALF WAVE RECTIFIER -

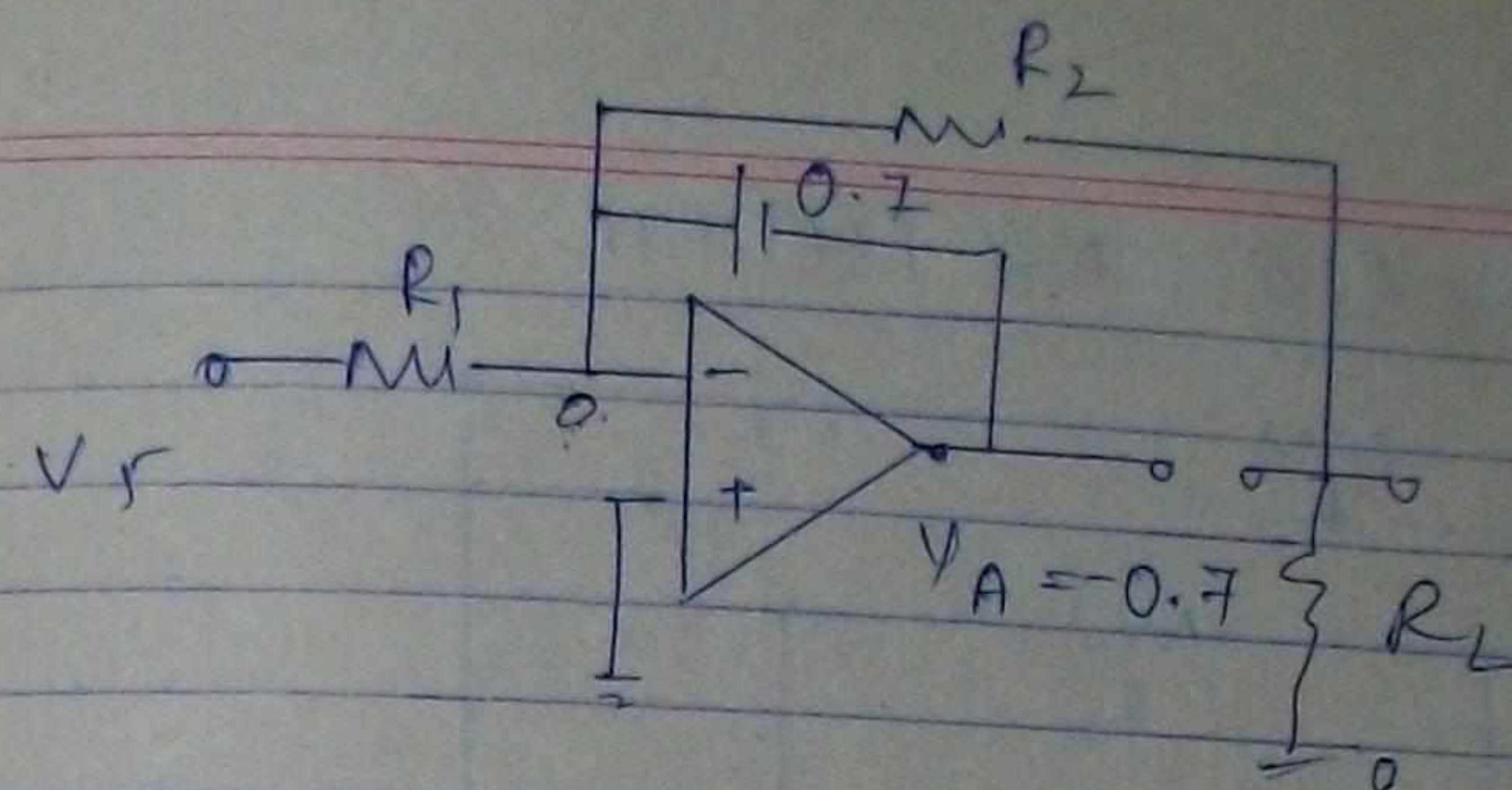


Considering ideal OPAMP and practical diode.

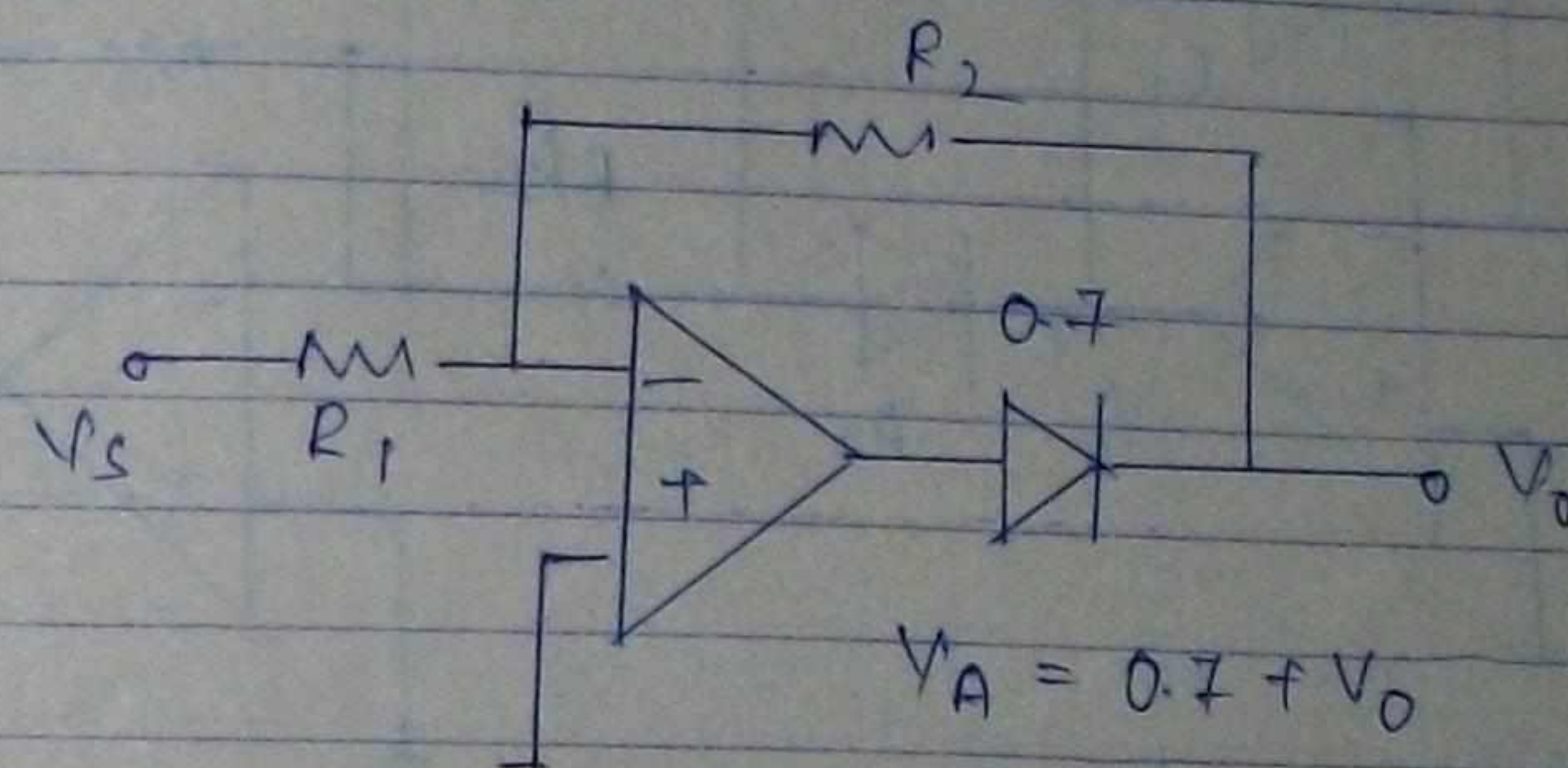
V_s	V_A	D_1	D_2	V_o
$V_s > 0$	$-V_{SAT}$ $-u$	ON	OFF	0
$V_s < 0$	$+V_{SAT}$ $+V_e$	OFF	ON	$-V_s$

(SUPER DIODE)

①



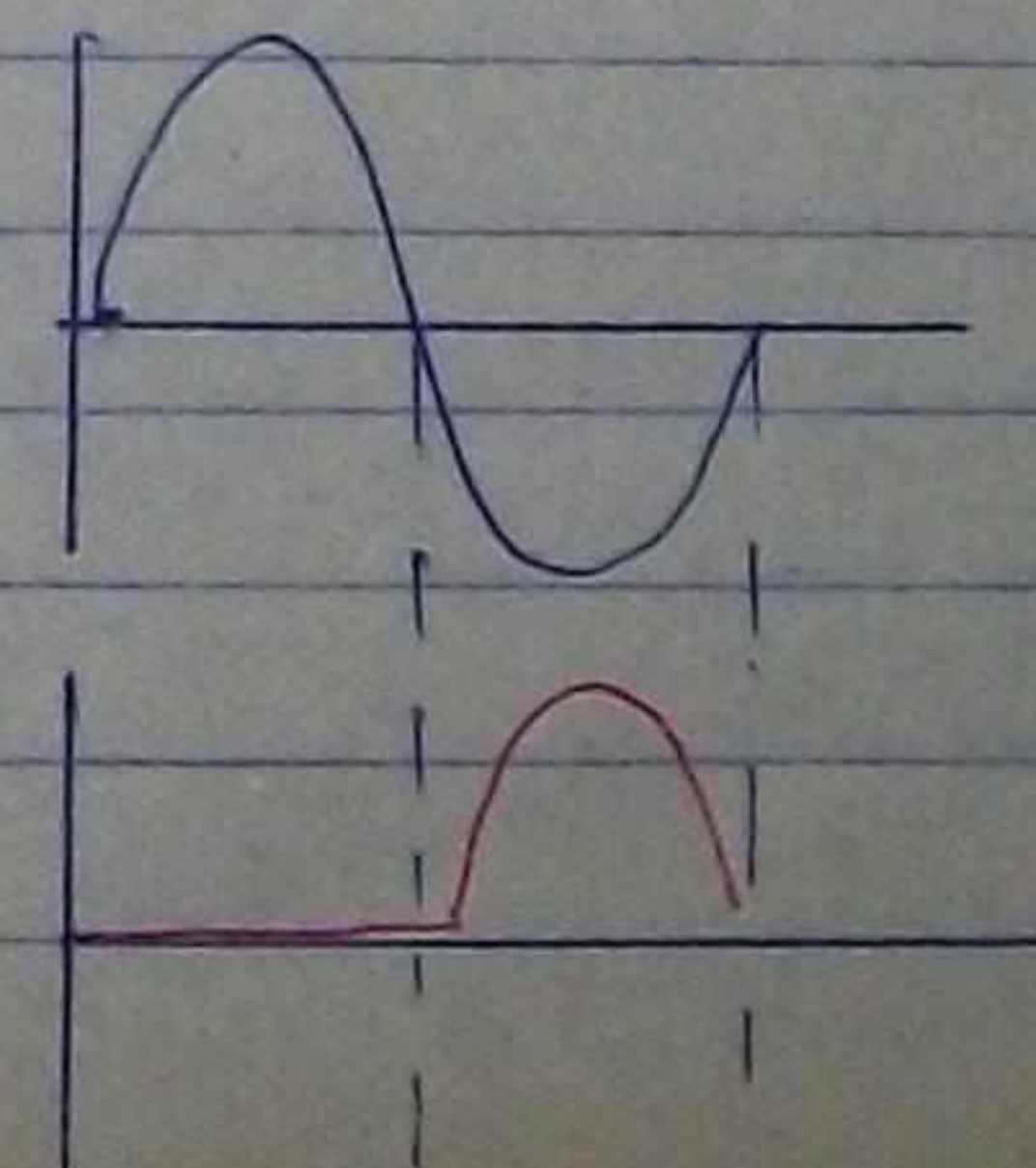
②



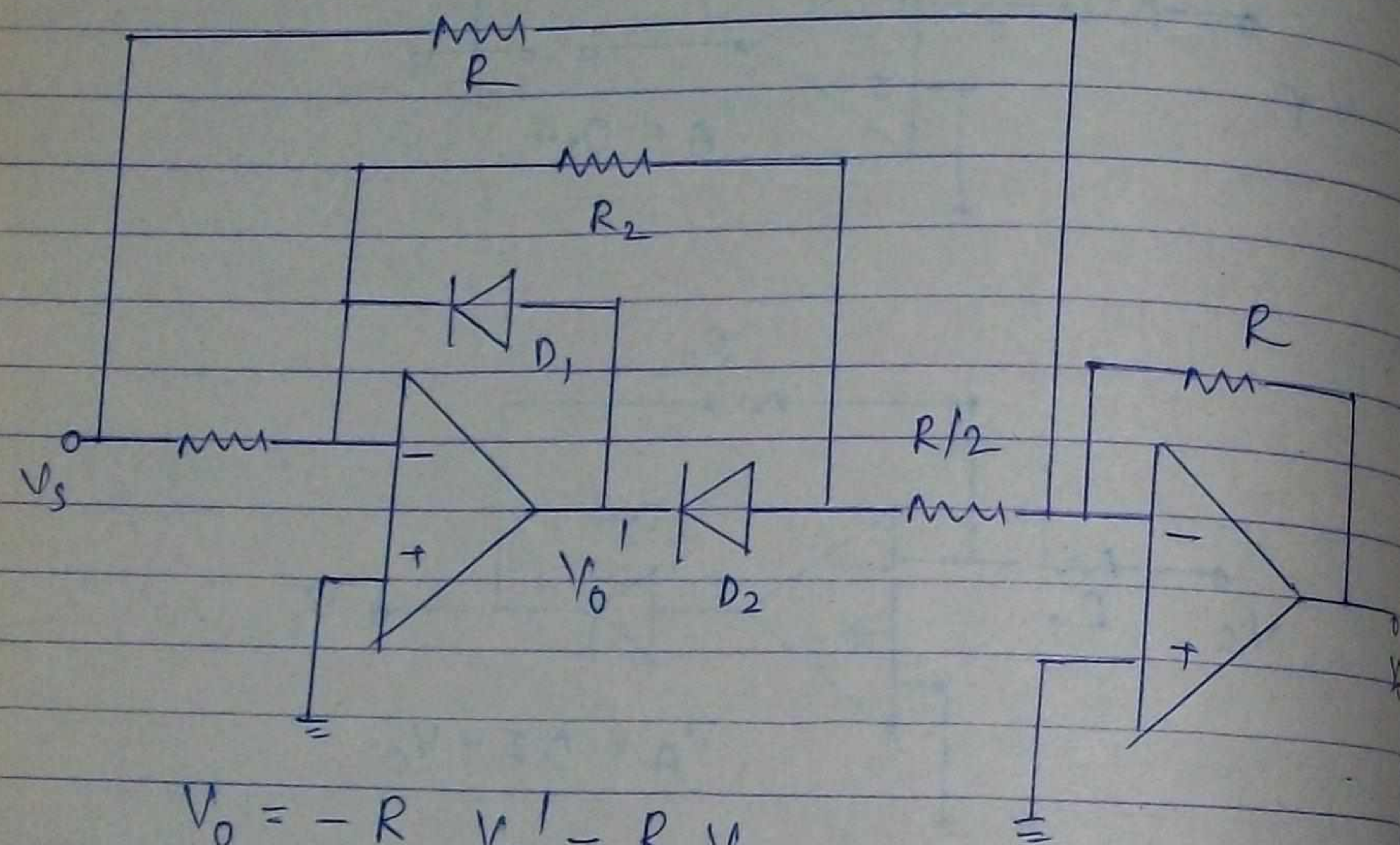
$$V_o = -\frac{R_2}{R_1} V_s \quad V_o = -V_s \quad \text{if } (R_2 = R_1)$$

- D_1 will avoid $-u$ saturation when $1/p(V_s)$ is $+u$.

- D_2 will avoid $+u$ saturation when $1/p(V_s)$ is $-u$.



⇒ PRECISION FULL WAVE RECTIFIER ←



$$V_o = -\frac{R}{R/2} V_o' - \frac{R}{R} V_s$$

$$V_o = -2V_o' - V_s$$

for +ve half:

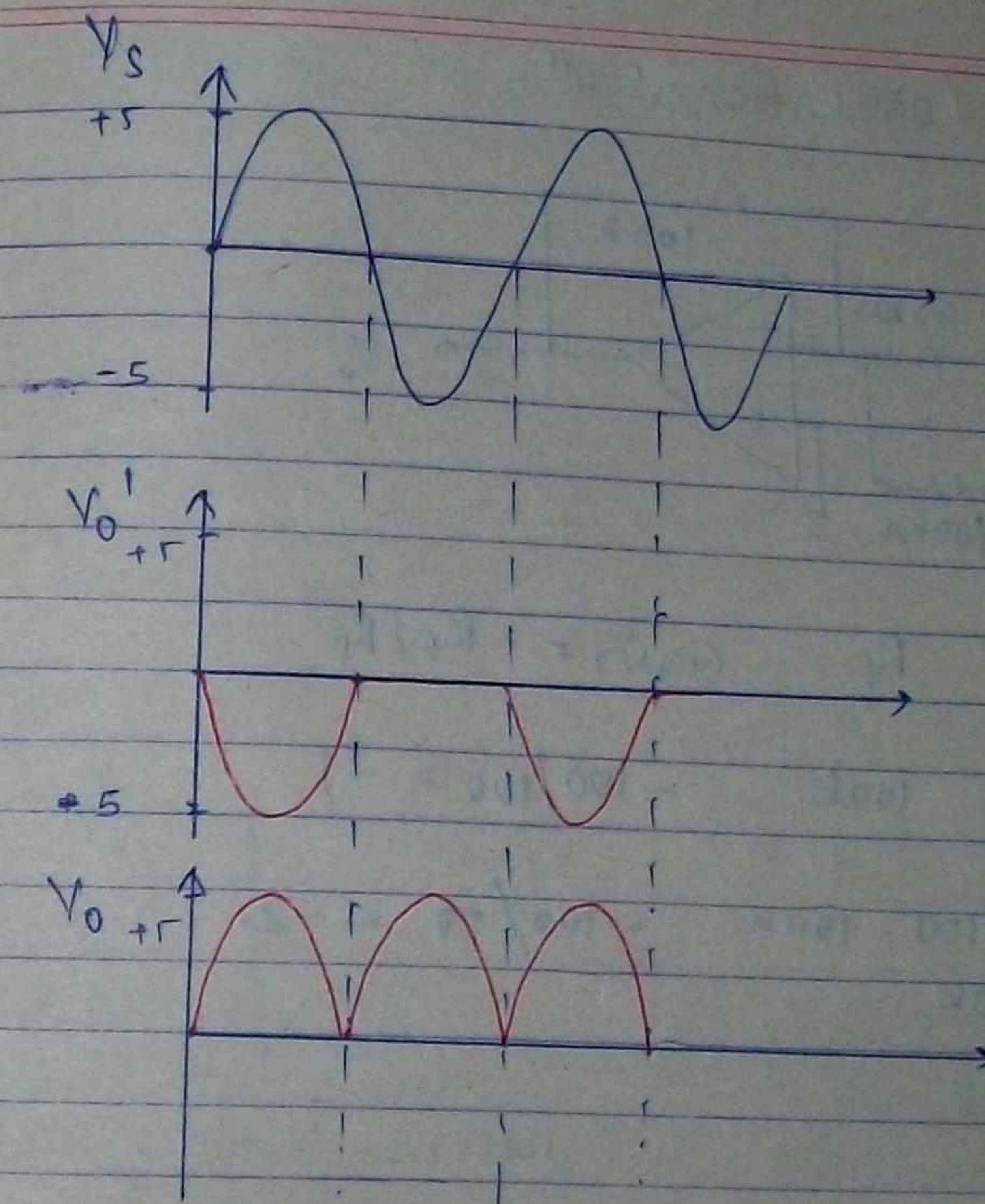
$$V_o = -2 \times (-5) - 5$$

$$V_o = 10 - 5 = 5$$

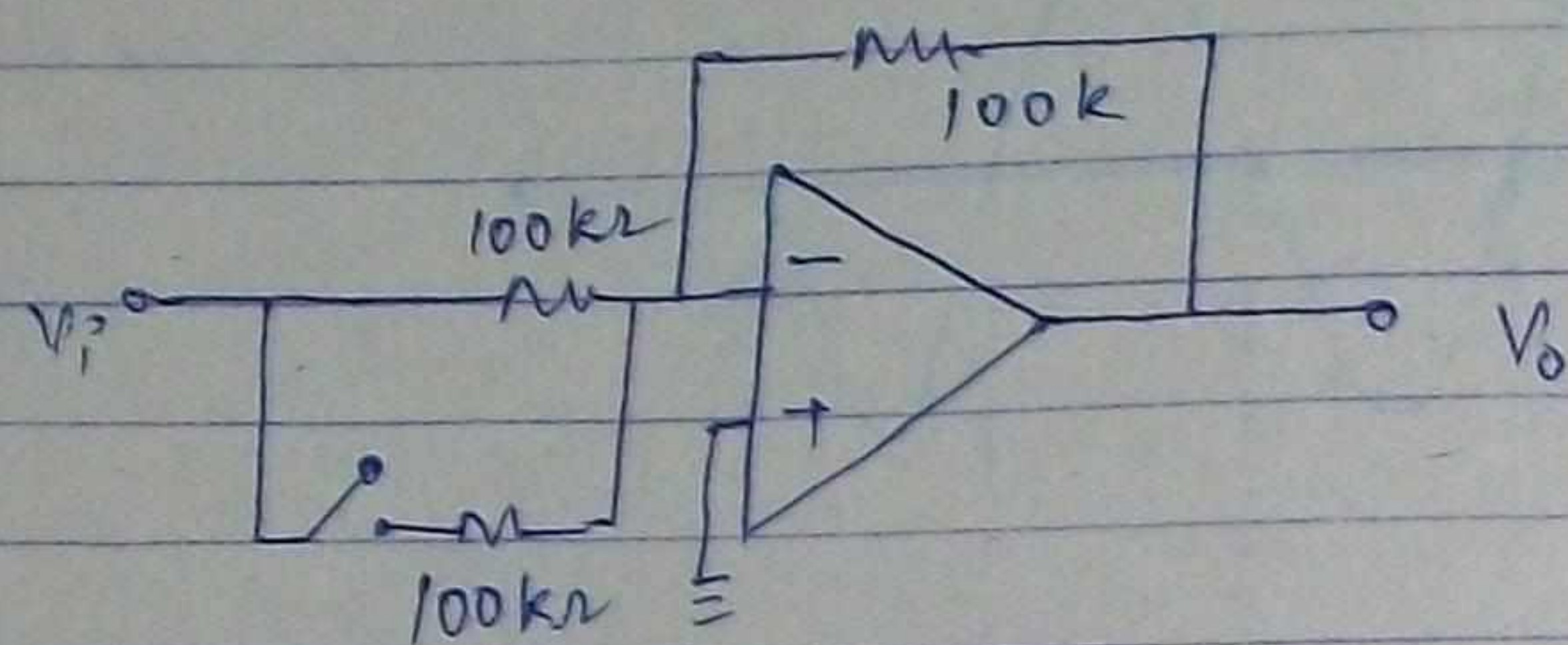
for -ve half:

$$V_o = -2 \times (0) - (-5)$$

$$V_o = 5$$



Que. P-55 Conventional Q₂.

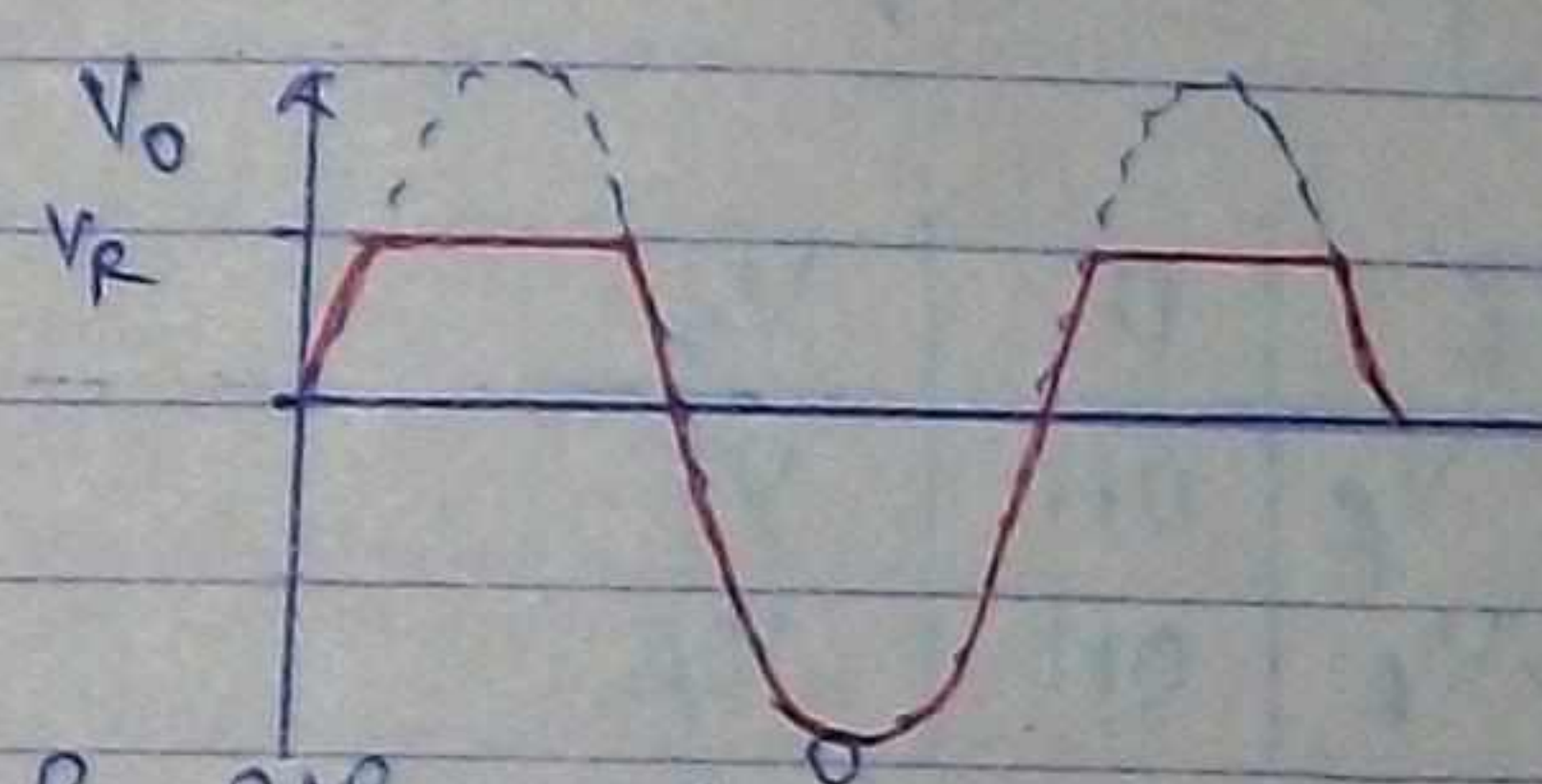
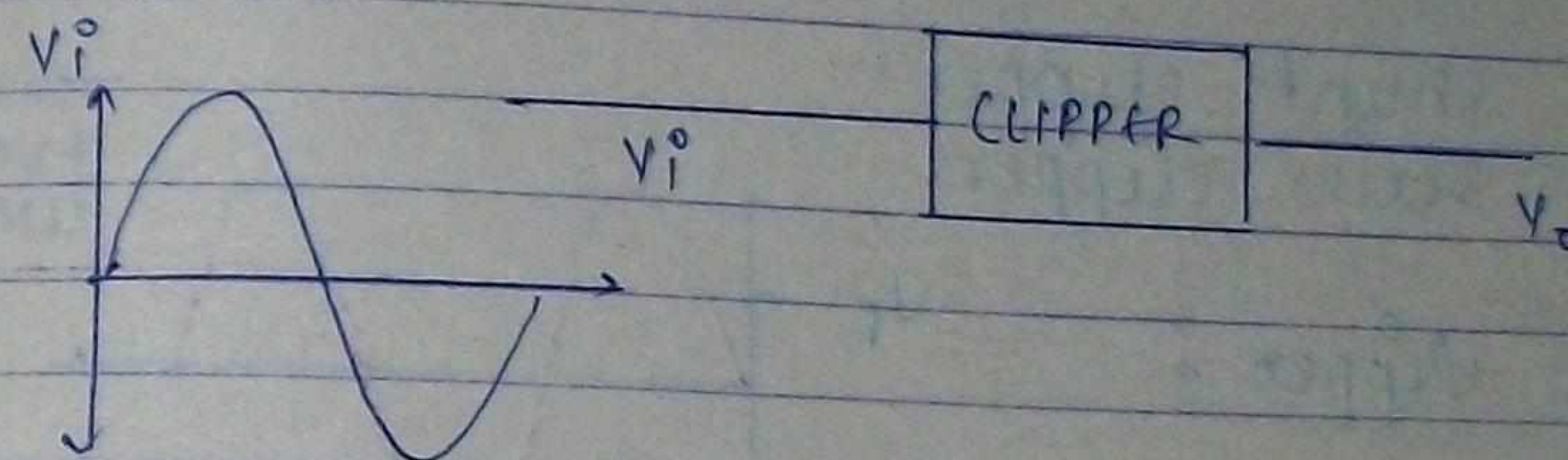


SWITCH	R_{eq}	R_f	Gain $= -R_f/R_i$
OFF	100k	100k	$-100/100 = -1$
ON	$100 // 100 = 50k$	100k	$-100/50 = -2$

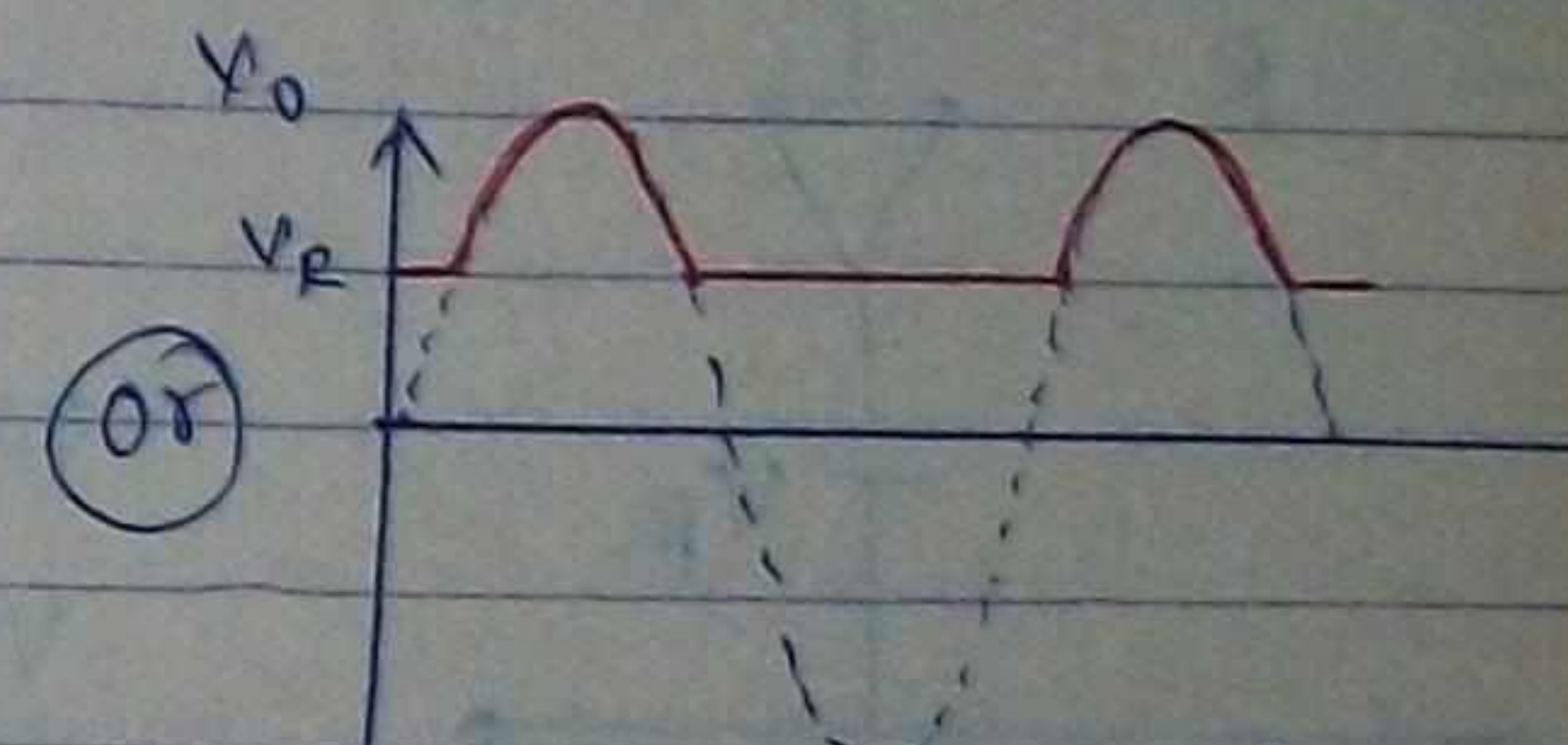
CLIPPER AND CLAMPERS

26/08/11

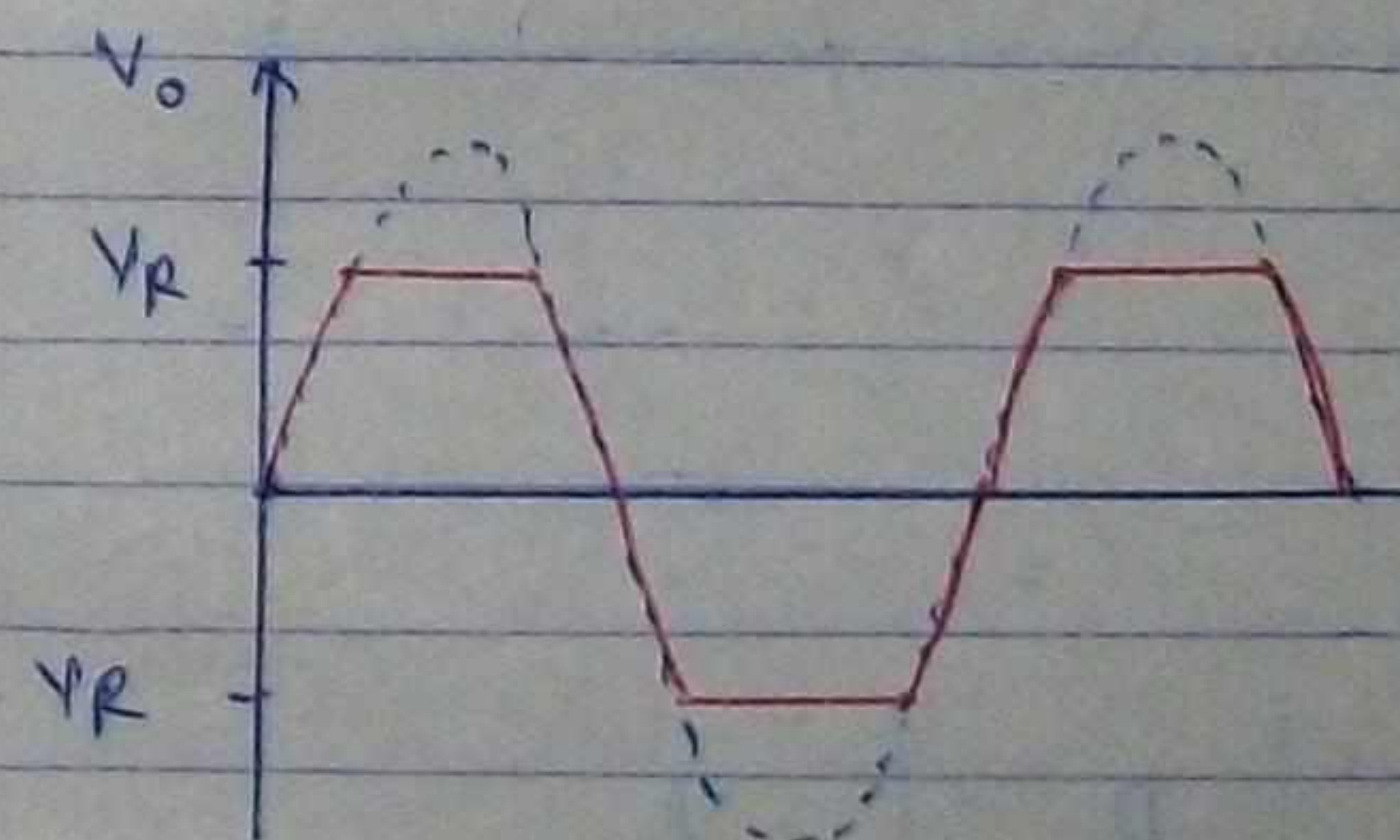
⇒ CLIPPING CIRCUIT OR LIMITING CIRCUIT -



Positive clipper
Since clipping is
above some reference
level



Negative clipper
Since clipping is
below some reference
level.



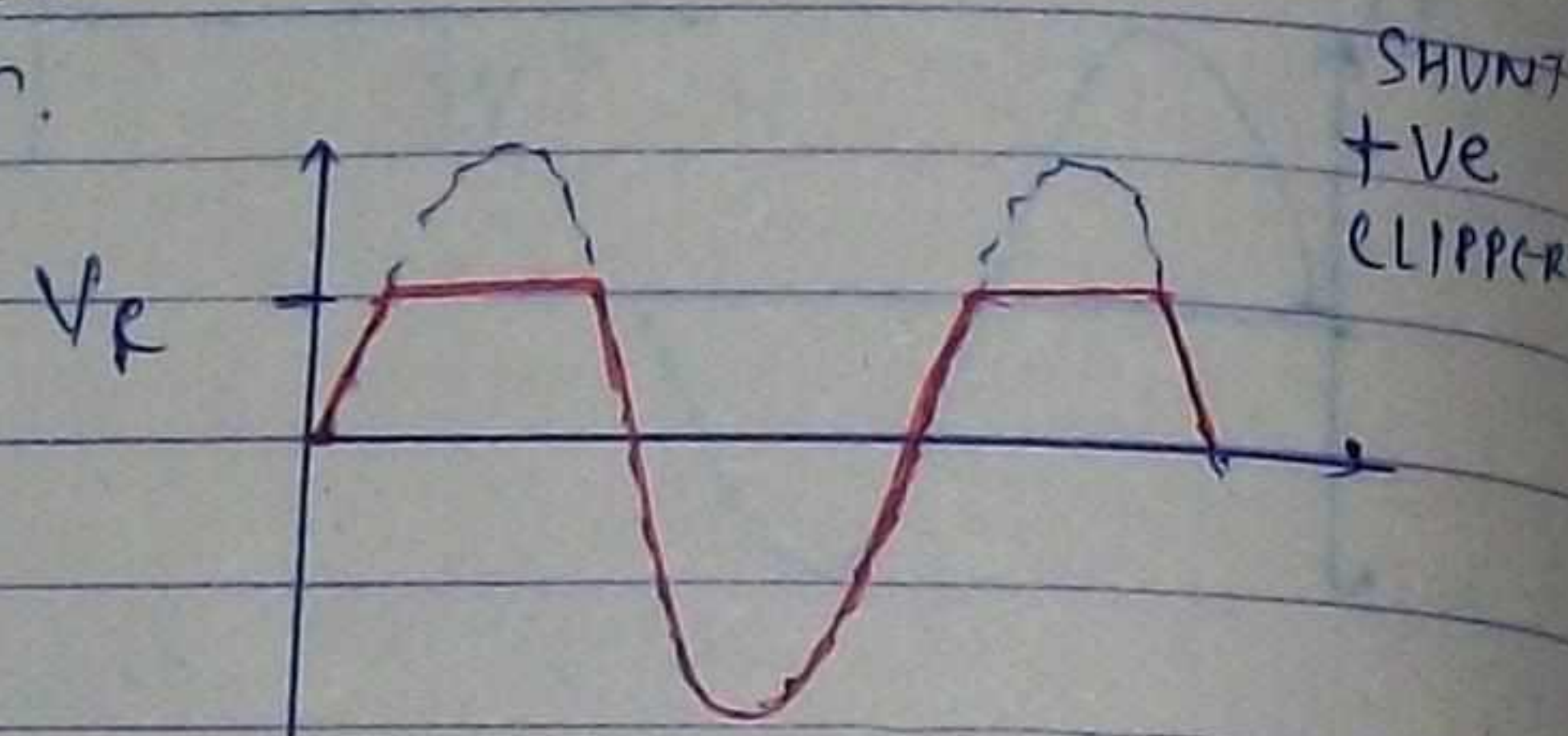
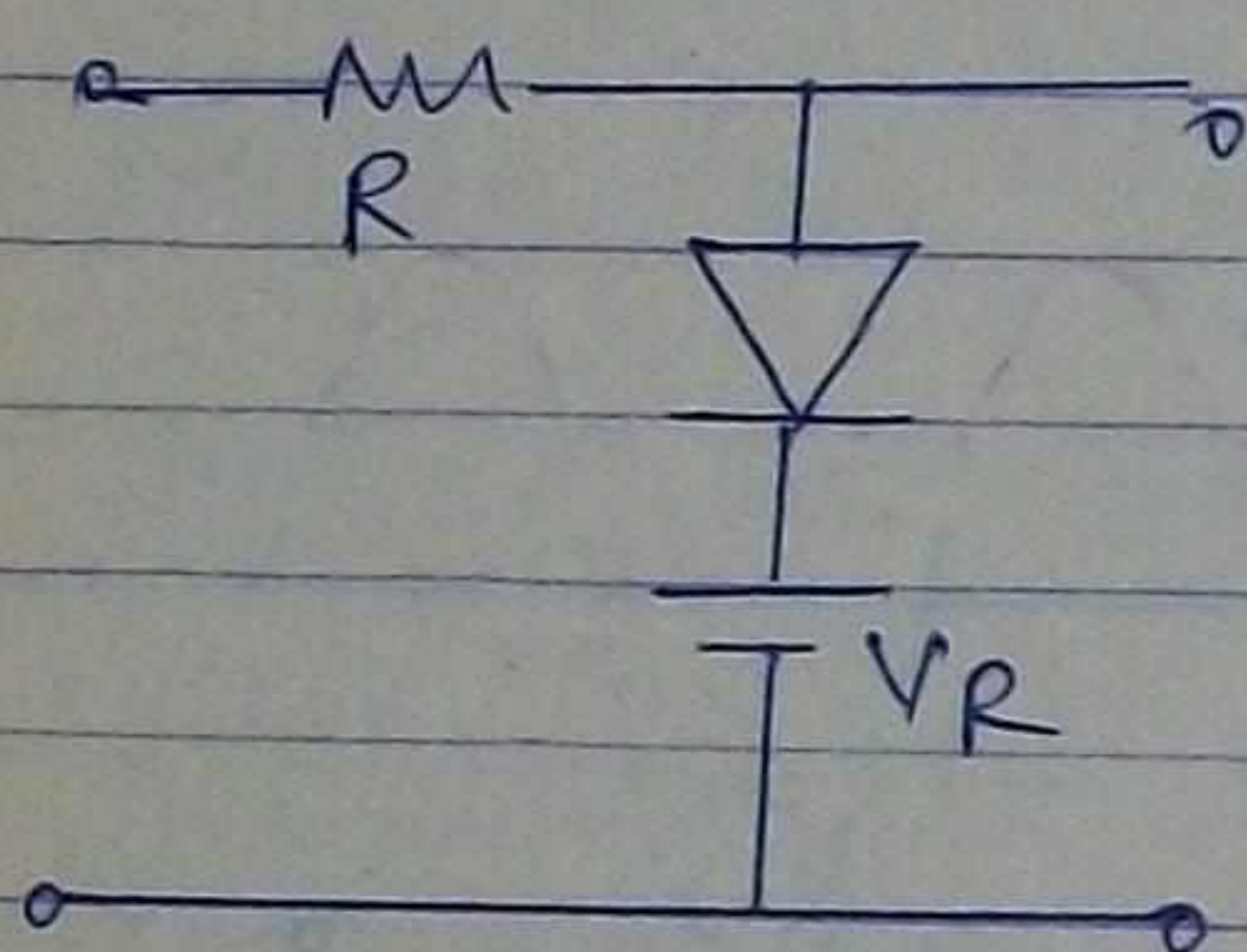
Two Independent
level clipper

- Clipper circuits are used to select that part of waveform which lies above or below some reference level. They are also referred to as amplitude selector, voltage or current limiter or slider.

- There are of two types (according to the position of diode w.r.t load):

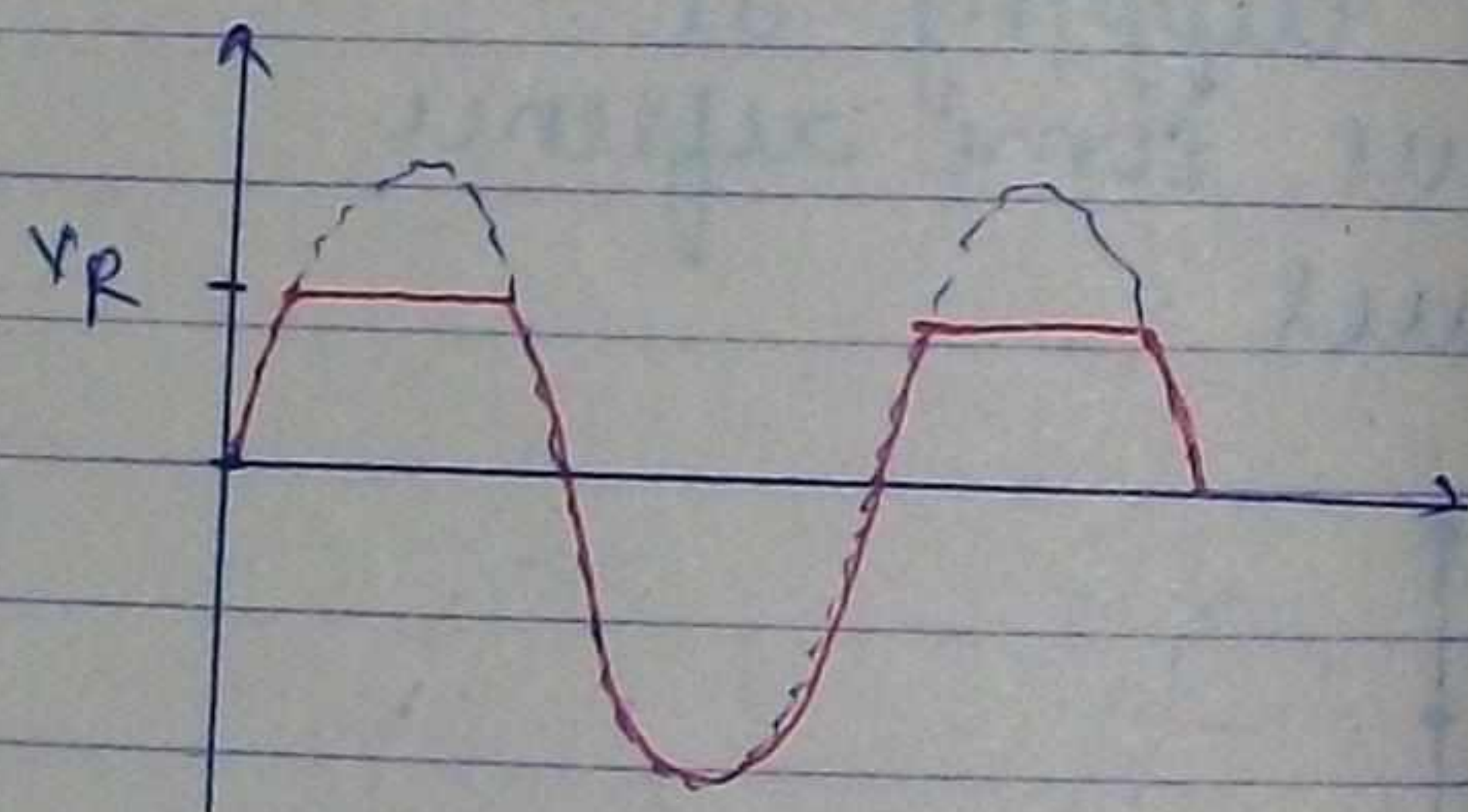
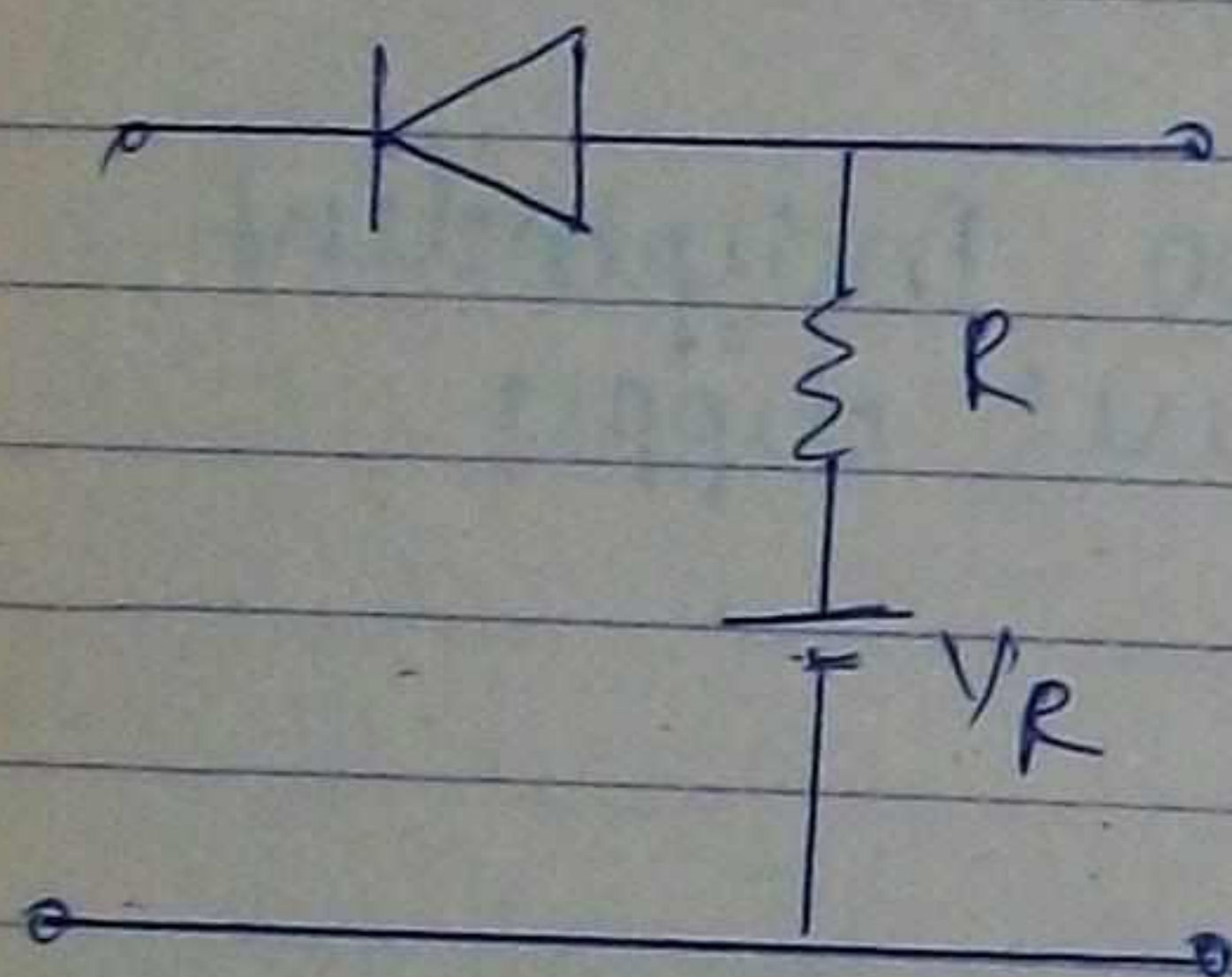
1. Shunt clipper.
2. Series clipper.

Shunt clipper:



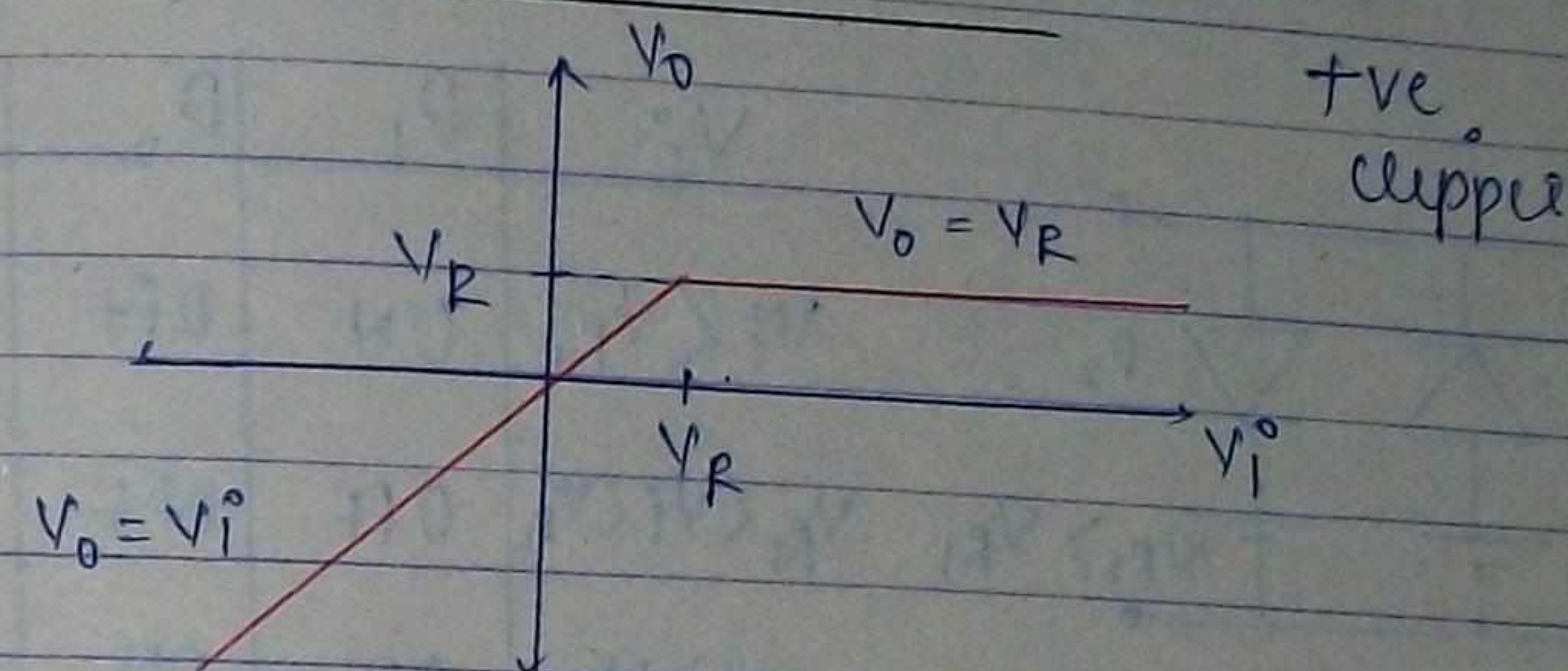
V_i	D	V_o
$V_i < V_R$	OFF	V_i
$V_i > V_R$	ON	V_R

Series clipper:

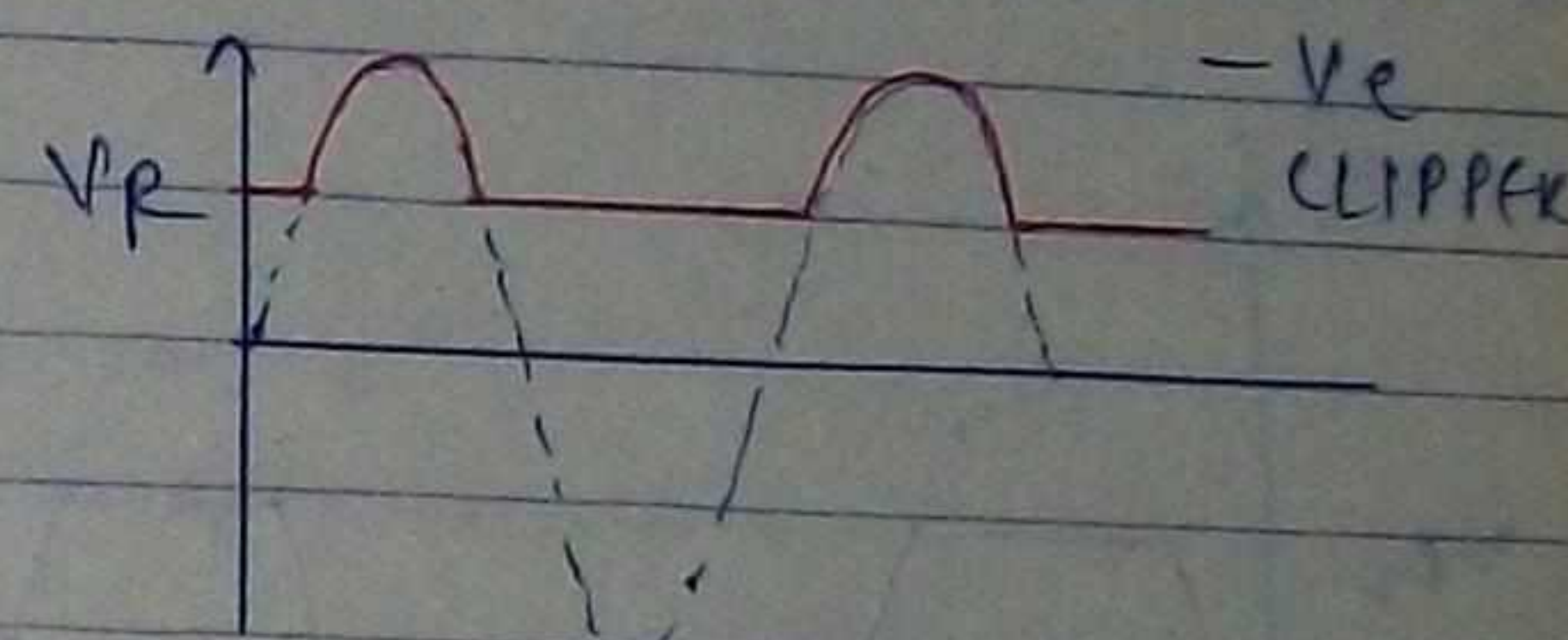
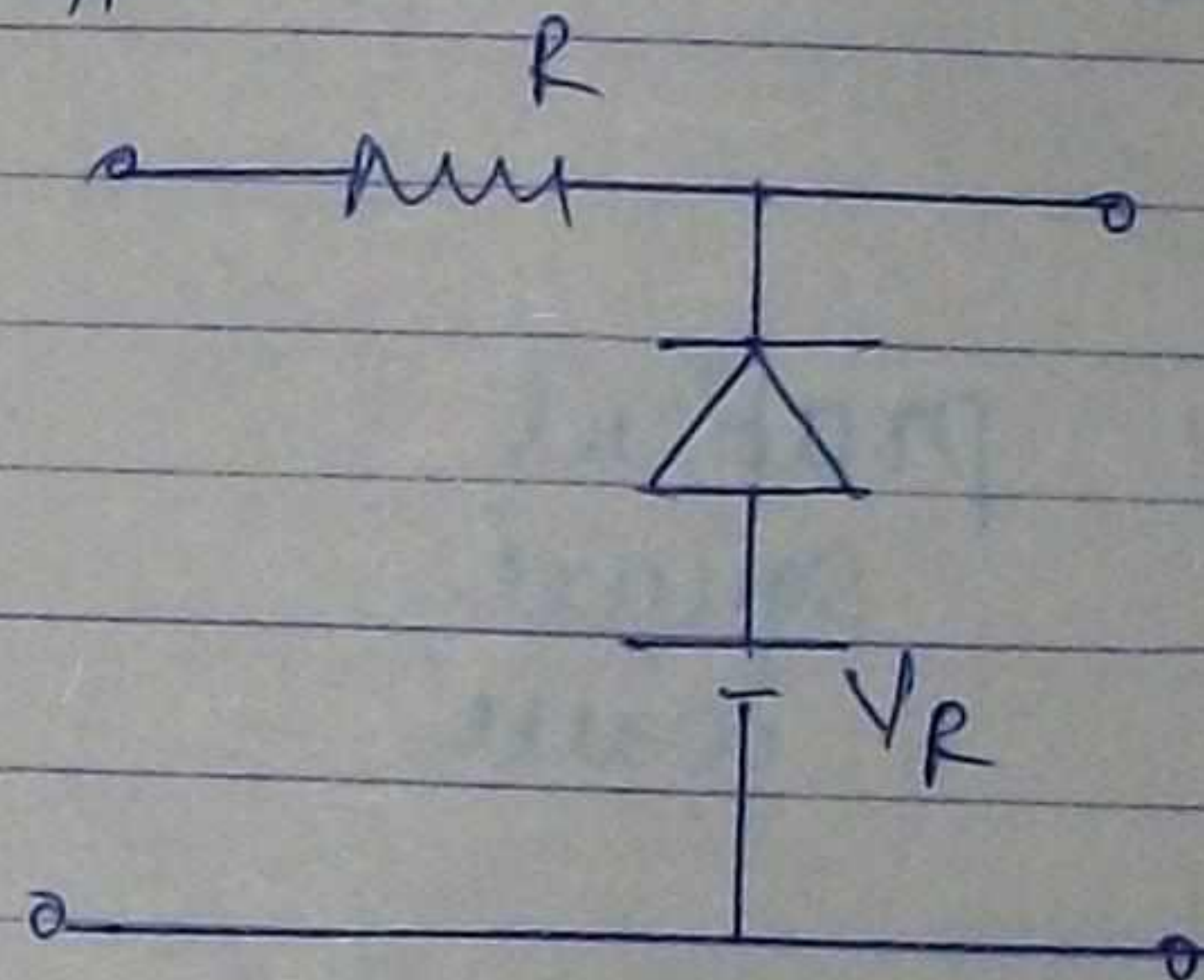


V_i	D	V_o
$V_i < V_R$	ON	V_i
$V_i > V_R$	OFF	V_R

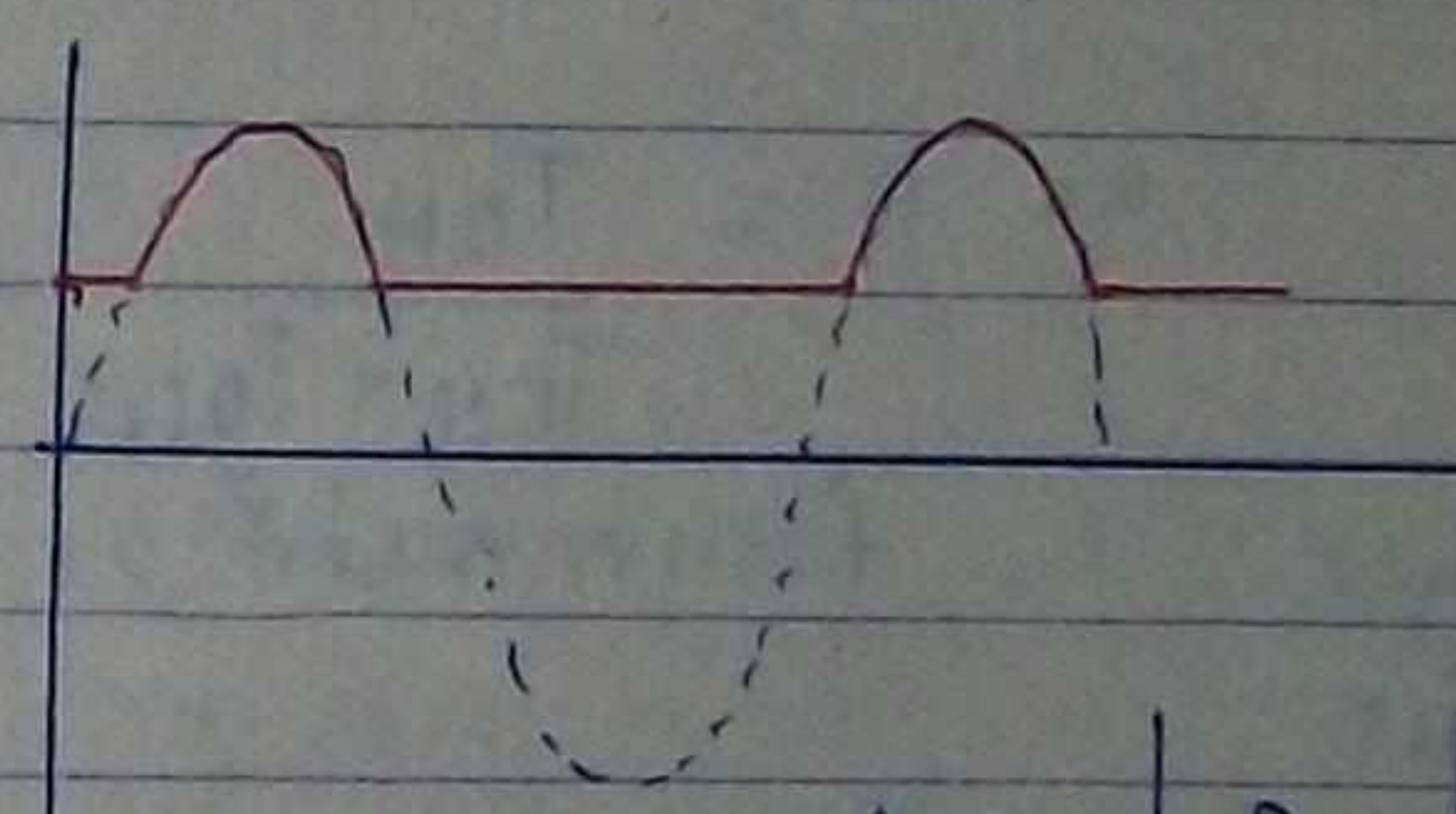
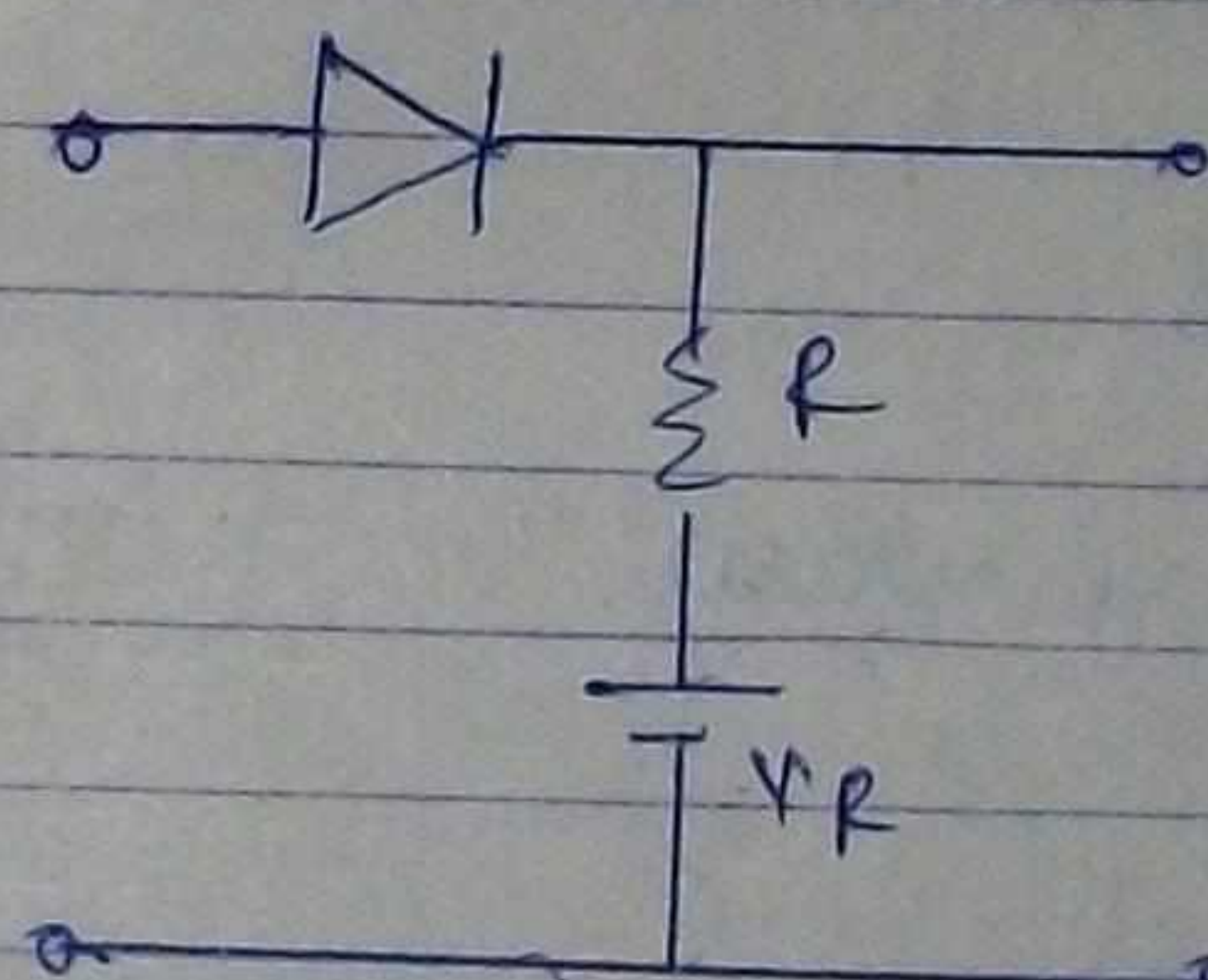
TRANSFER CURVE (V_o vs V_i) -



#

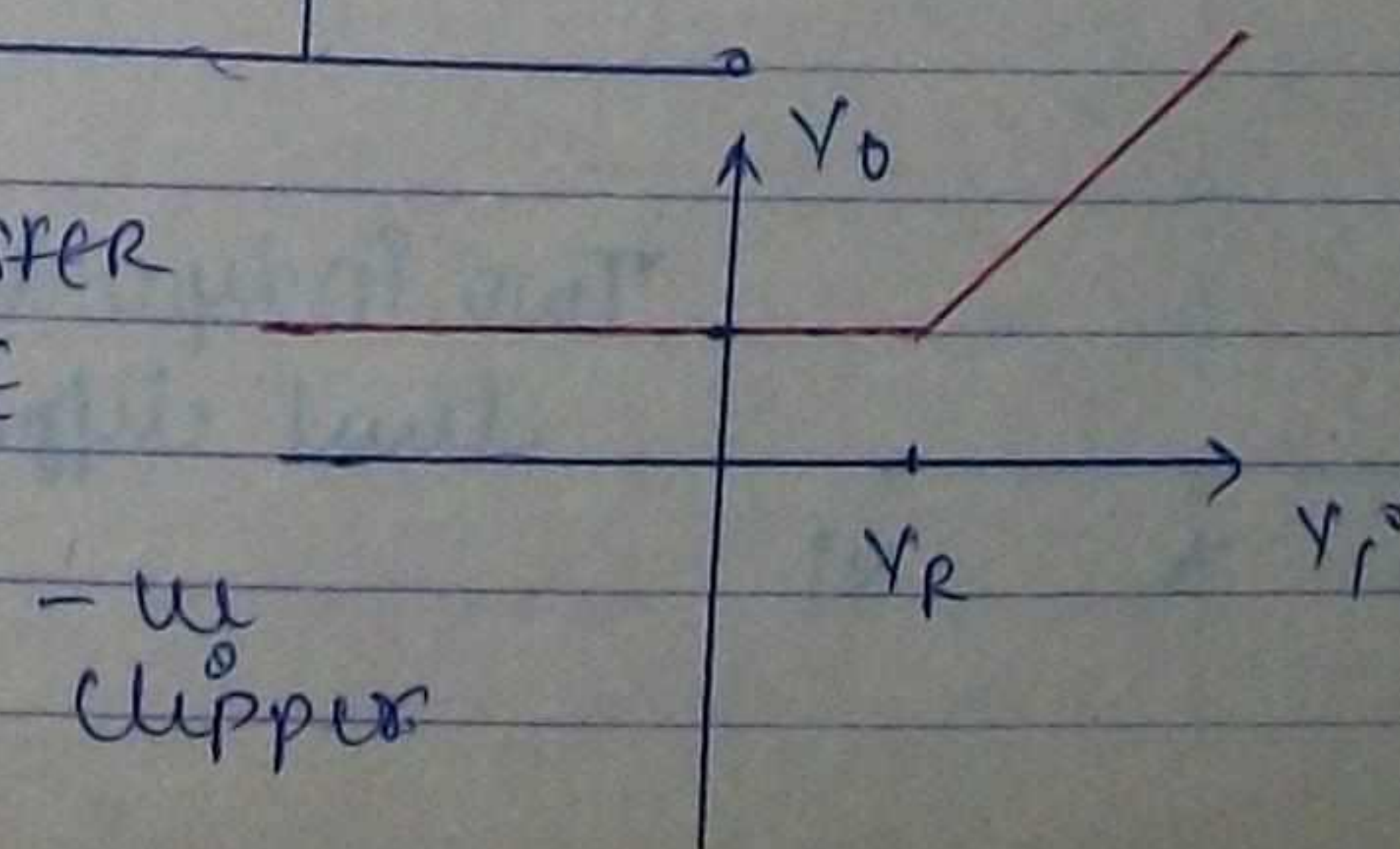


V_i	D	V_o
$V_i < V_R$	ON	V_R
$V_i > V_R$	OFF	V_i

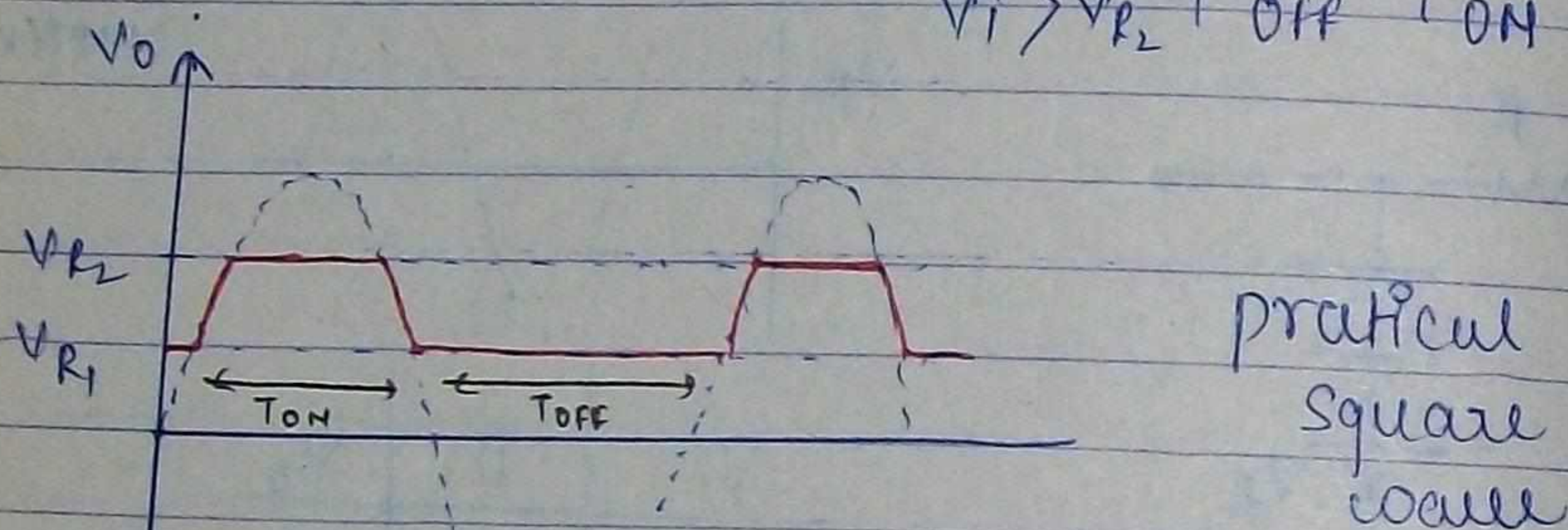
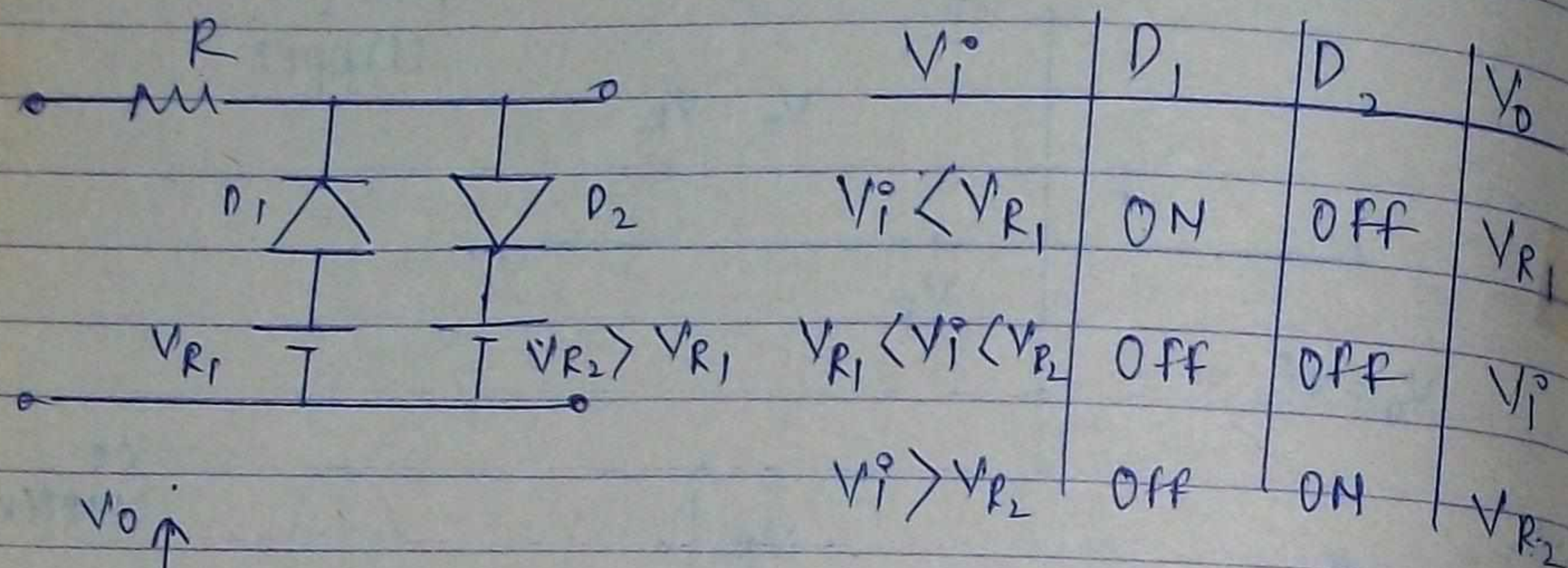


V_i	D	V_o
$V_i < V_R$	ON	V_R
$V_i > V_R$	OFF	V_i

TRANSFER CURVE



⇒ TWO INDEPENDENT LEVEL CLIPPER -

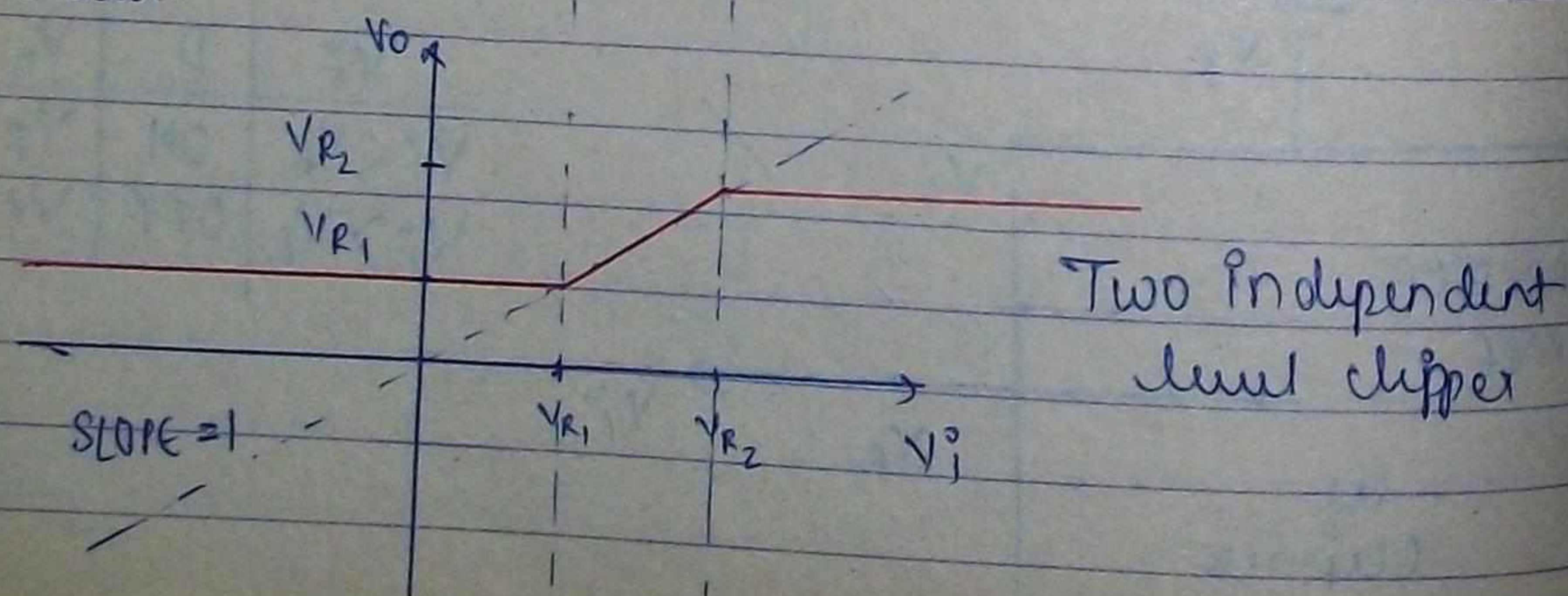


$$T_{OFF} > T_{ON} \quad D < 50\%$$

$$\therefore D = \frac{T_{ON}}{T_{ON} + T_{OFF}} \times 100\%$$

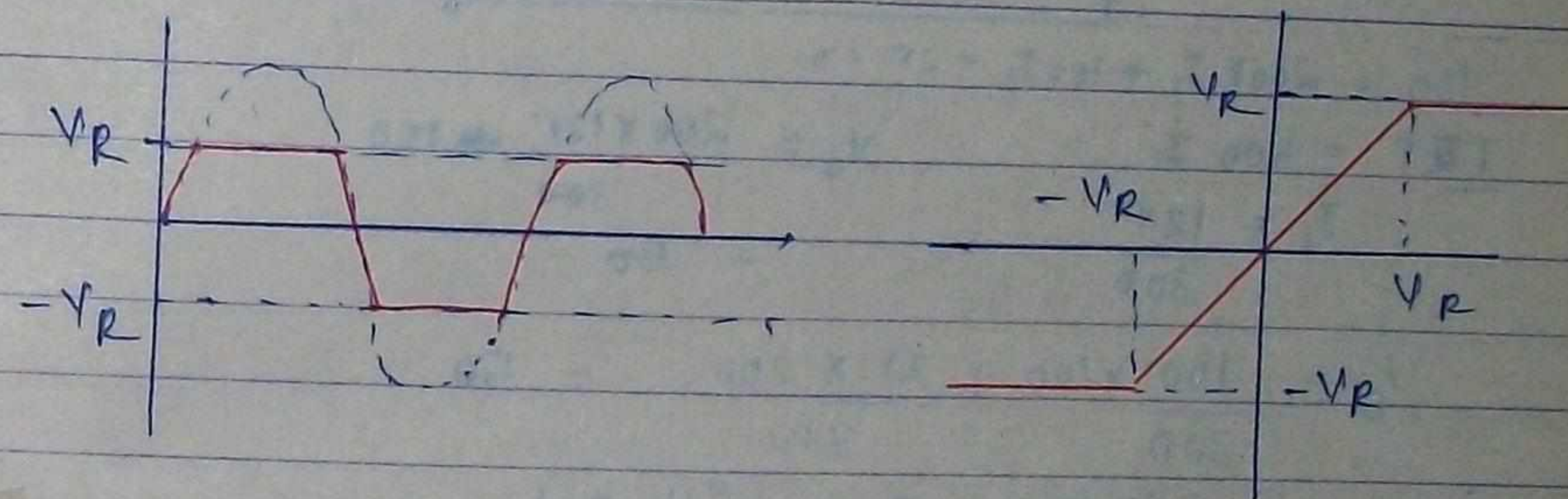
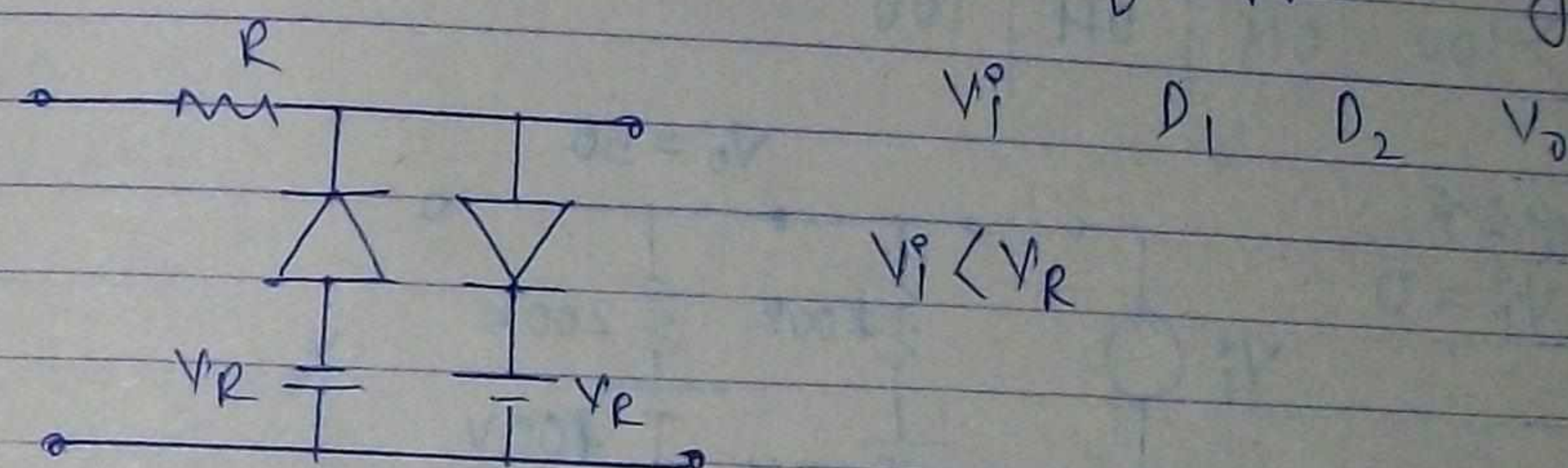
Transfer curve

Asymmetrical square wave.

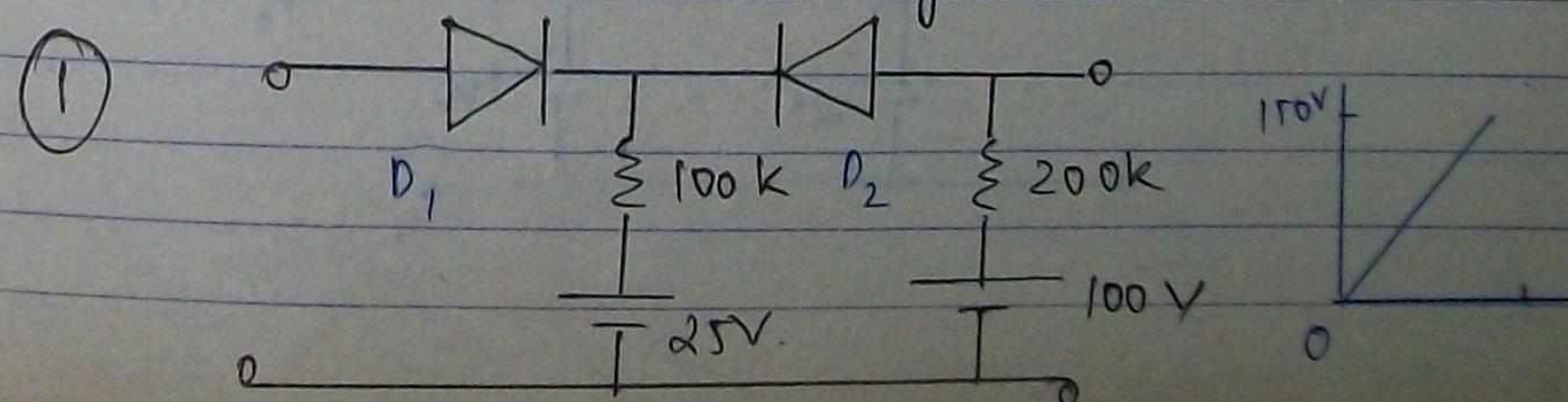


The circuit is used as means of converting a sinusoidal waveform into a square wave (Asymmetrical square wave)

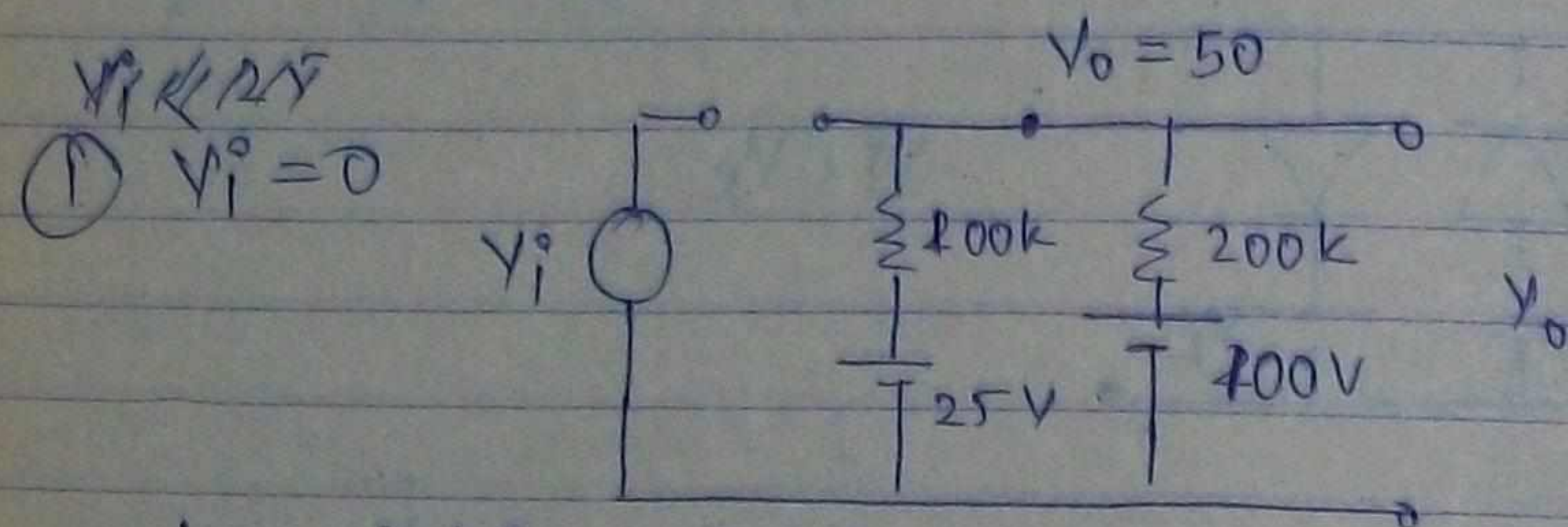
To generate a symmetrical square wave V_{R1} and V_{R2} are adjusted to be numerically equal but of opposite sign.



Ques. The I/P voltage V_i to a level clipper shown in figure varies linearly from 0 to 150 V. Sketch the O/P voltage V_o to the same time scale as the I/P voltage assume ideal diodes.



V_i	D_1	D_2	V_o
$V_i < 25$	OFF	ON	50V
$50 < V_i < 100$	ON	ON	V_i
$V_i > 100$	ON	OFF	100



$$100 = 200k I_1 + 100 I_1 + 25 = 0$$

$$125 = 300 I_1$$

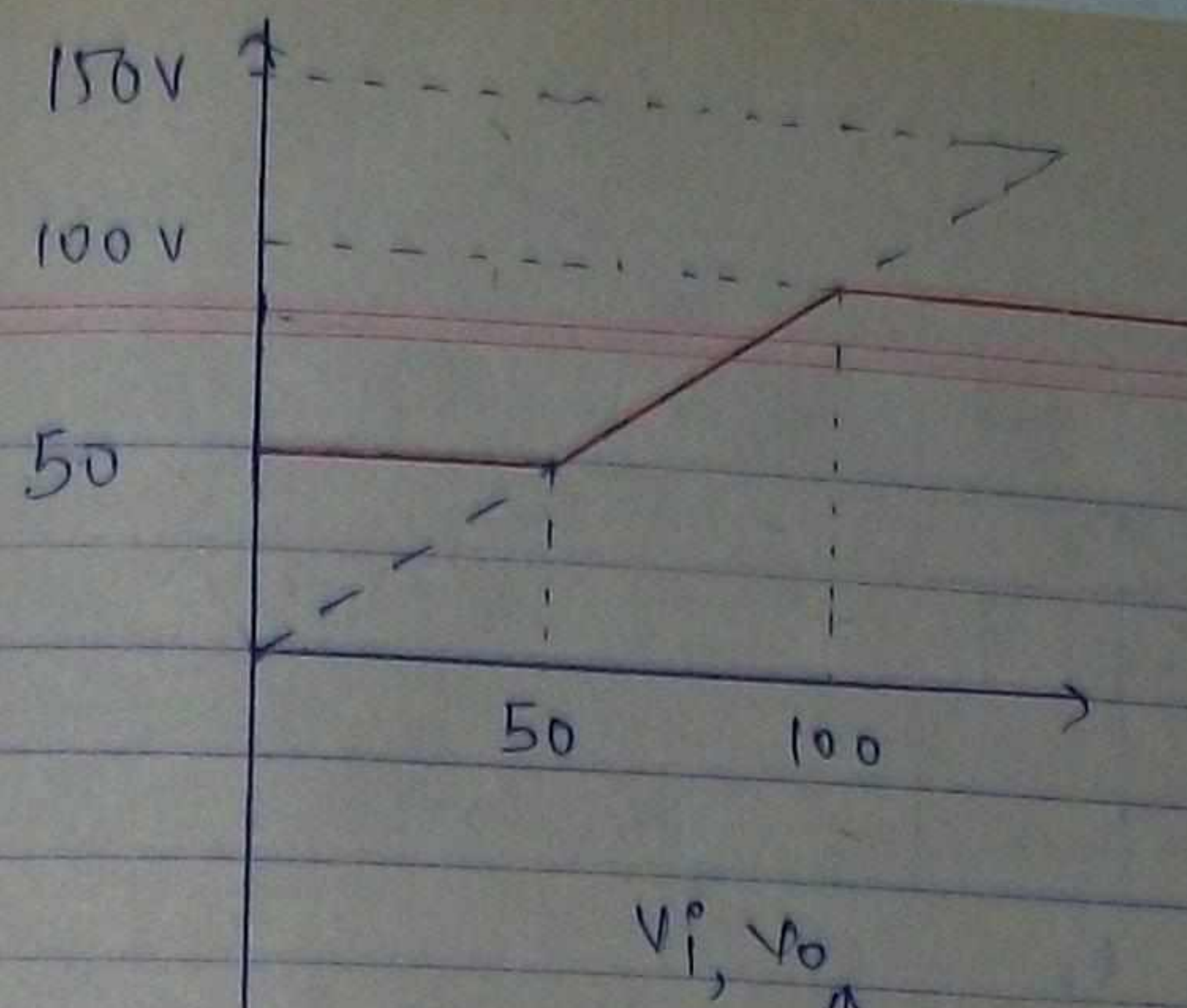
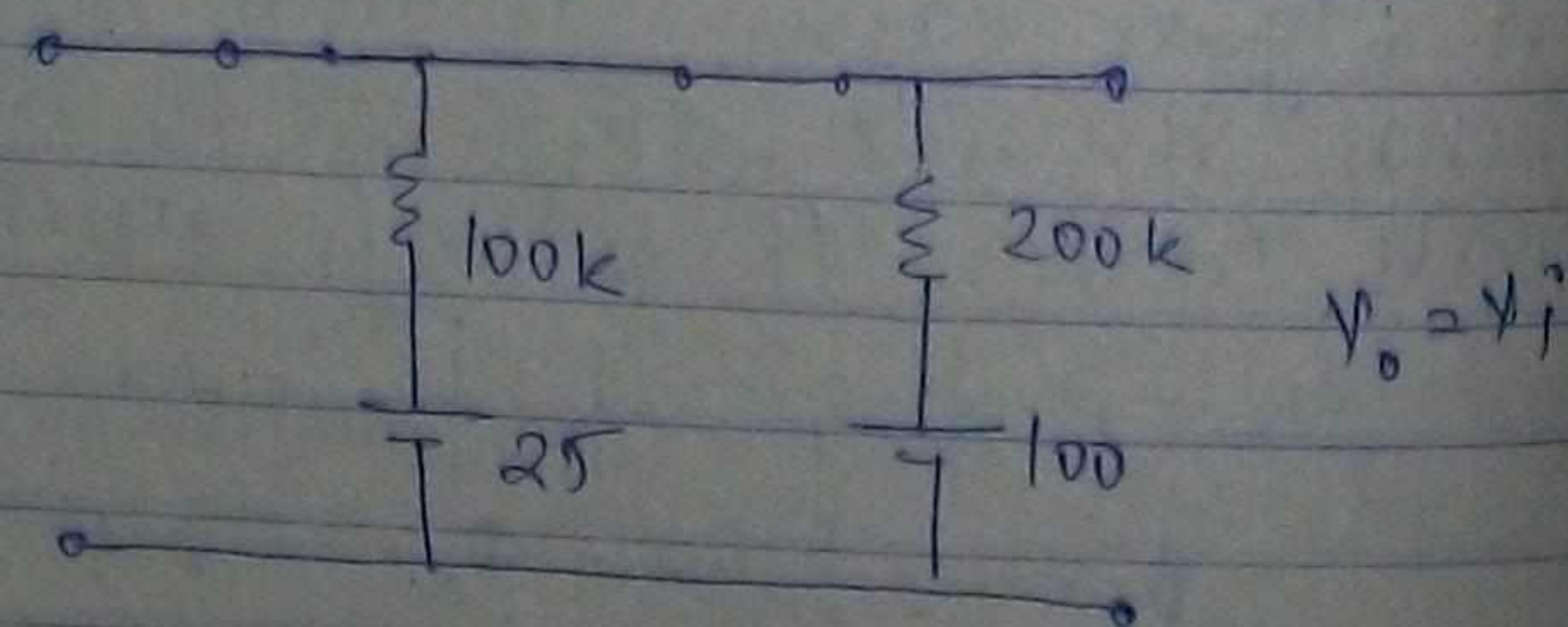
$$I_1 = \frac{125}{300}$$

$$V_o = \frac{200 \times 125}{300} + 100 = 50$$

$$V_o = \frac{100 \times 100 + 25 \times 200}{300} = 50$$

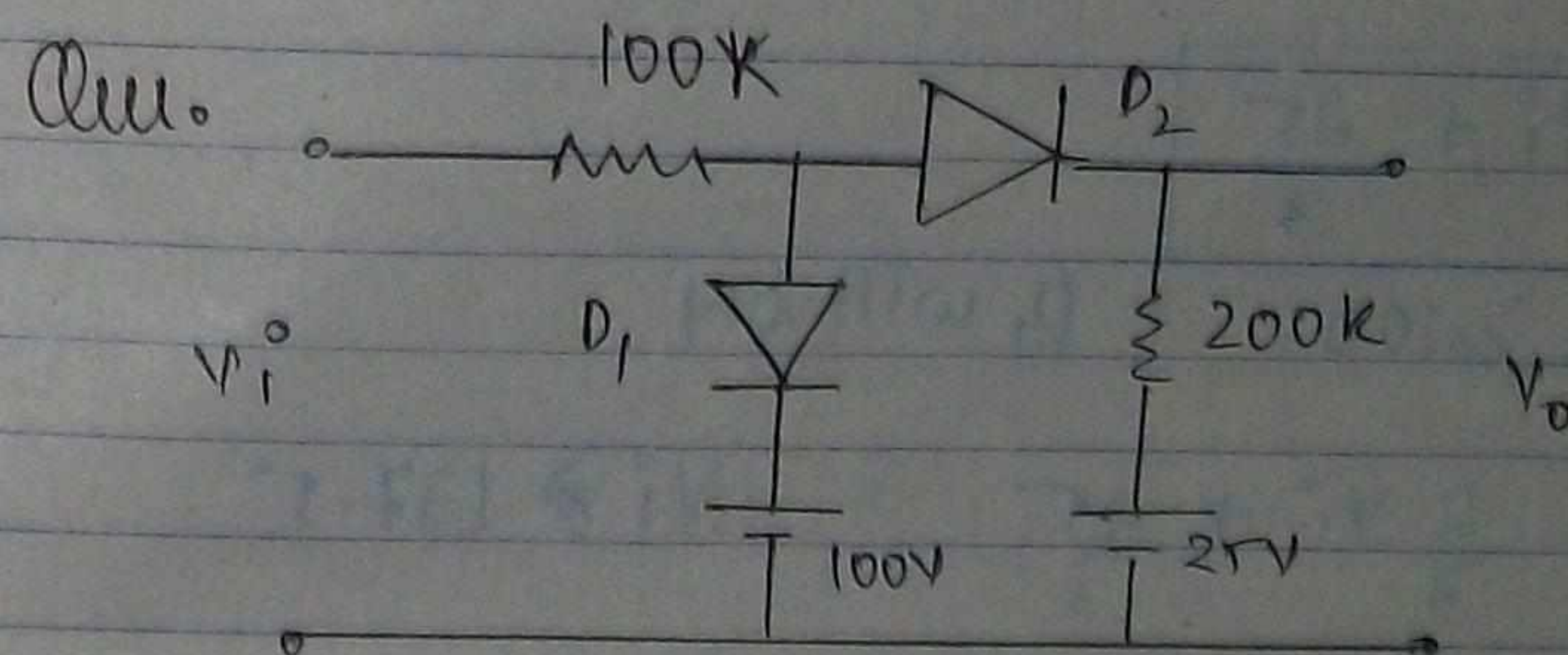
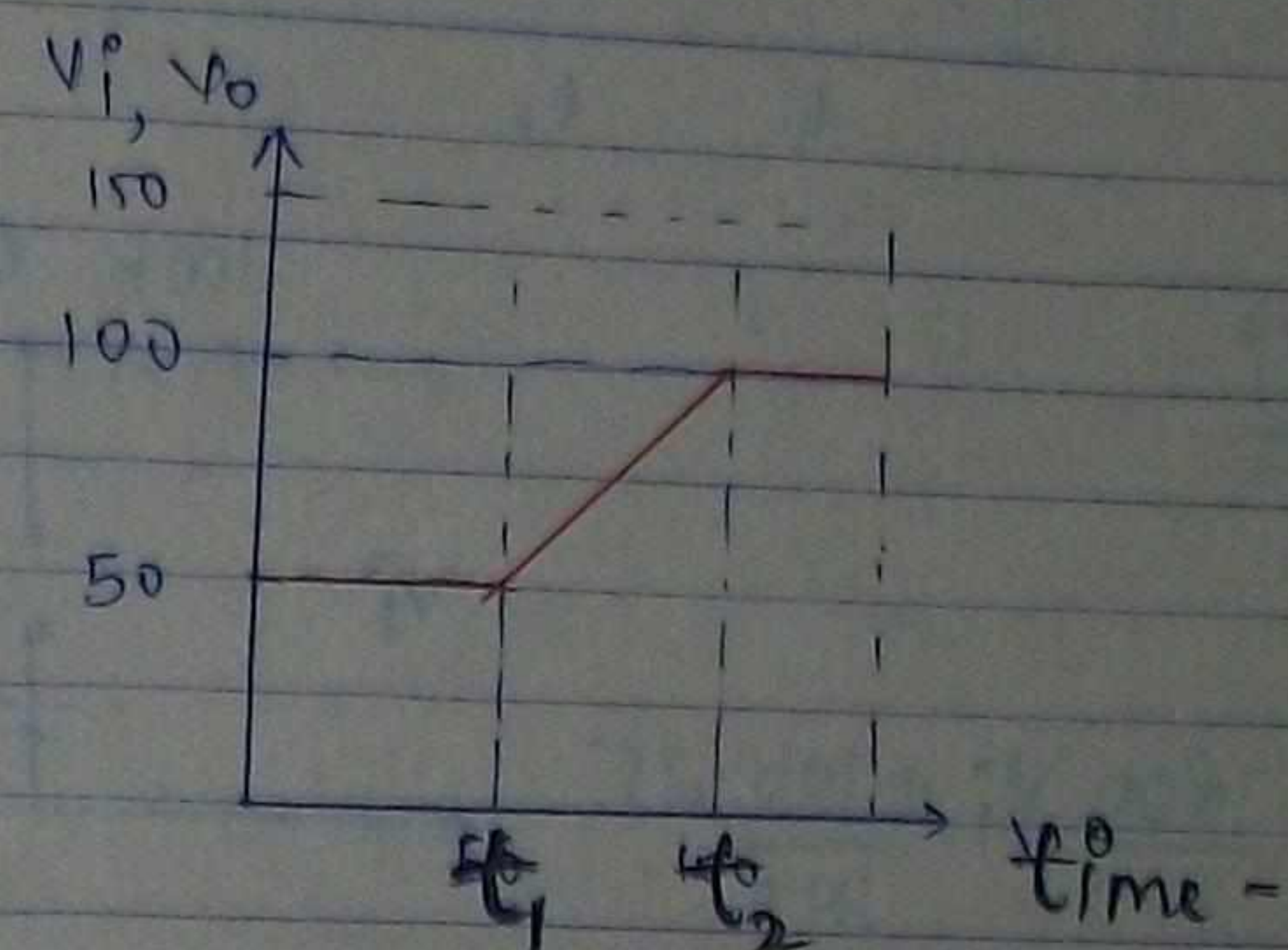
when $V_i > 50$ D_1 will ON.

② $V_i \geq 50$

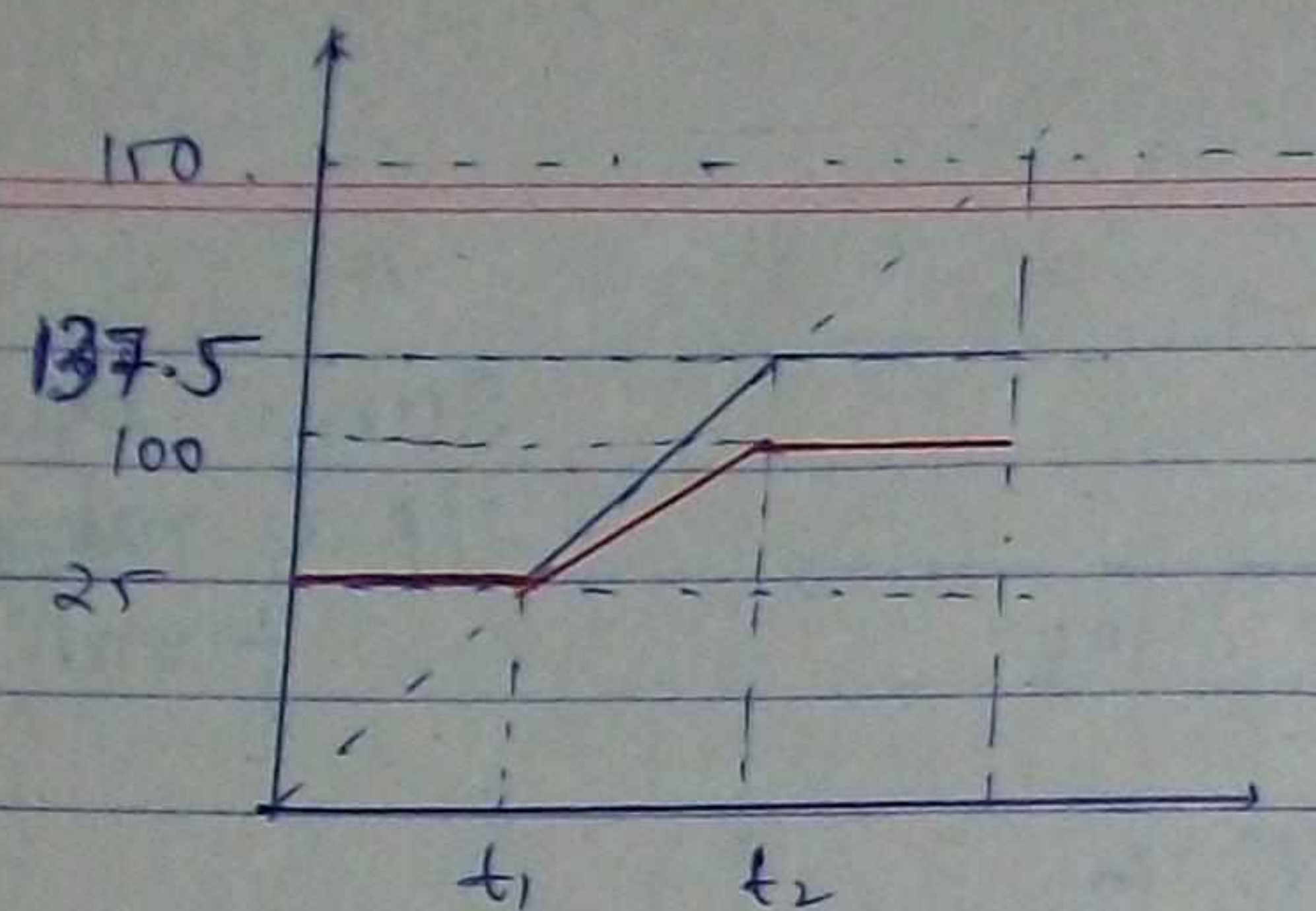


Same because
i/p is ramp
function.

TRANSFER
CURVE



V_i	D_1	D_2	V_o
$0 < V_i < 25$	OFF	OFF	25
$25 < V_i < 137.5$	OFF	ON	$\frac{2V_i}{3} + \frac{25}{3}$
$V_i > 137.5$	ON	ON	100



$25 < V_i$

$$V_o = \frac{200 V_i + 100 \cdot 25}{300}$$

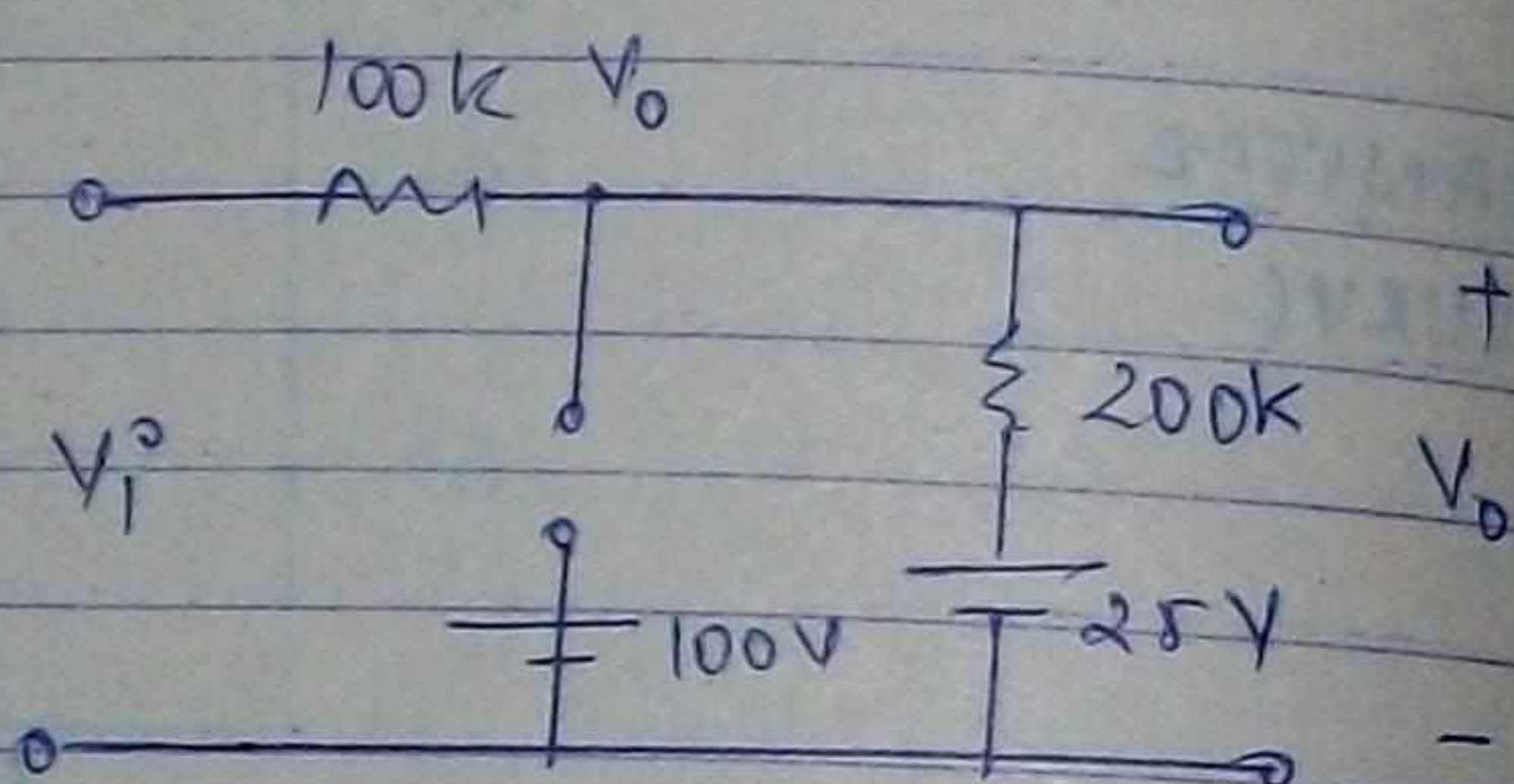
$$V_o = \frac{2}{3} V_i + \frac{25}{3}$$

when $V_o \geq 100$ D_1 will ON

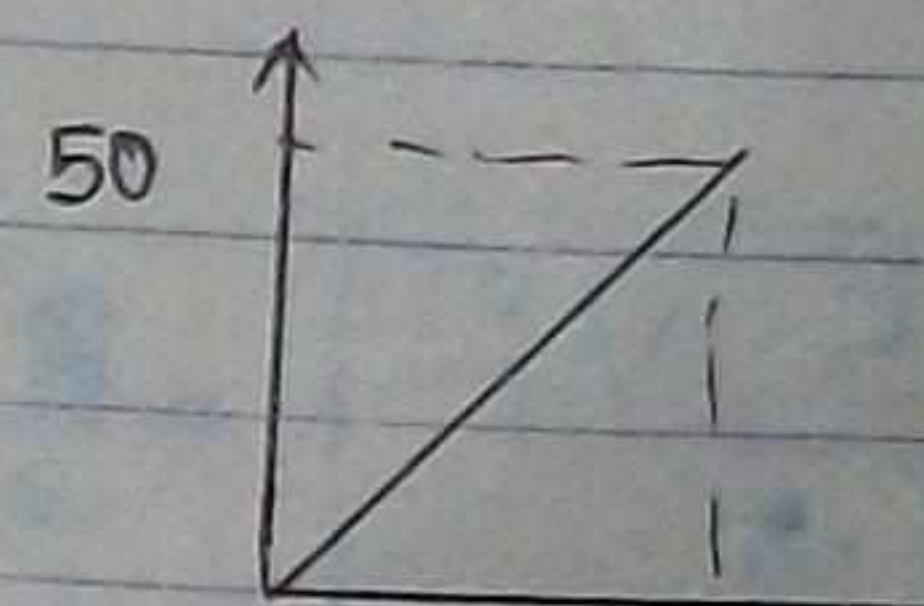
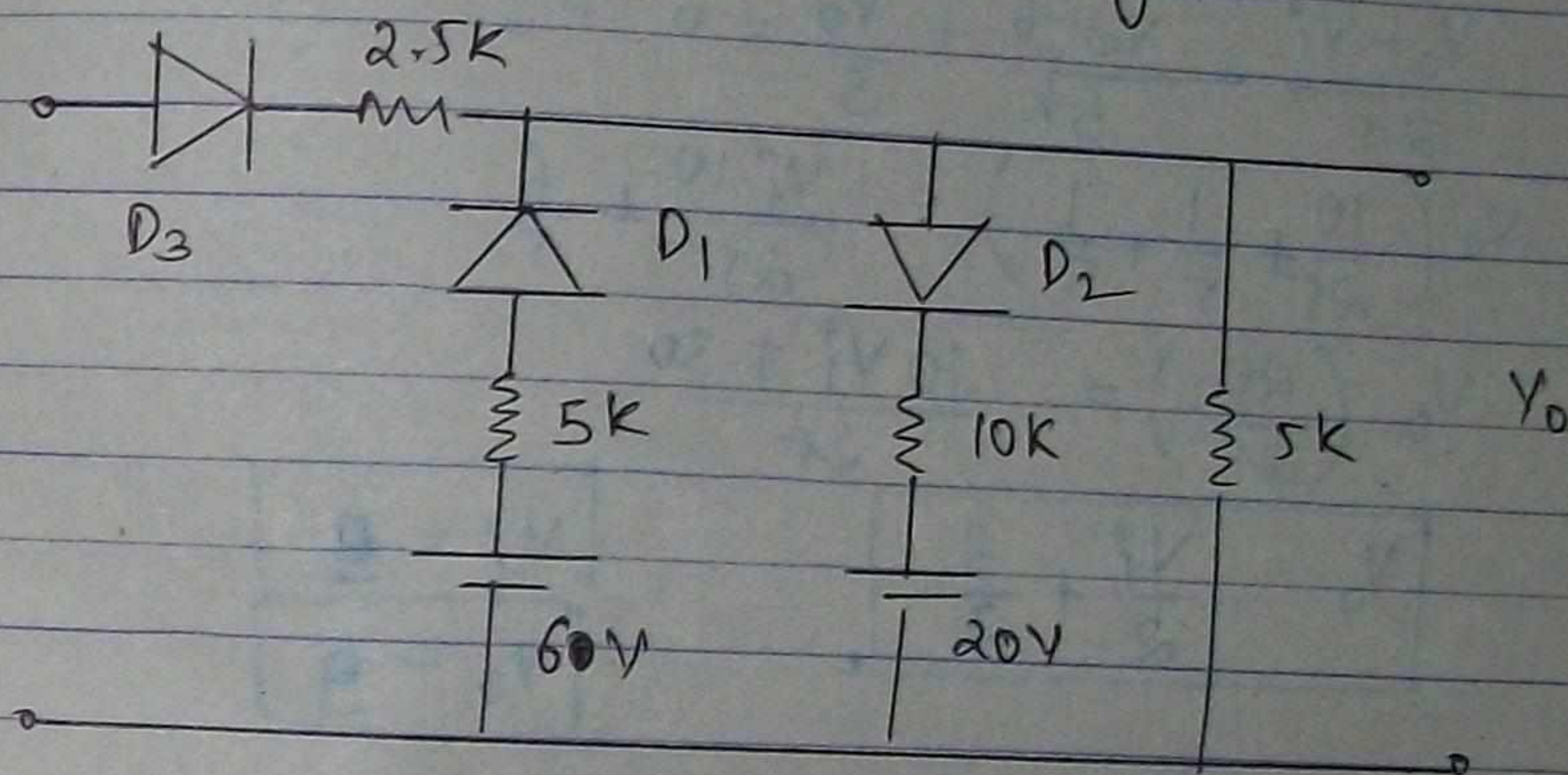
$$100 = \frac{2}{3} V_i + \frac{25}{3}, \quad V_i \geq 137.5$$

$V_i = 137.5$

$V_o = 100$

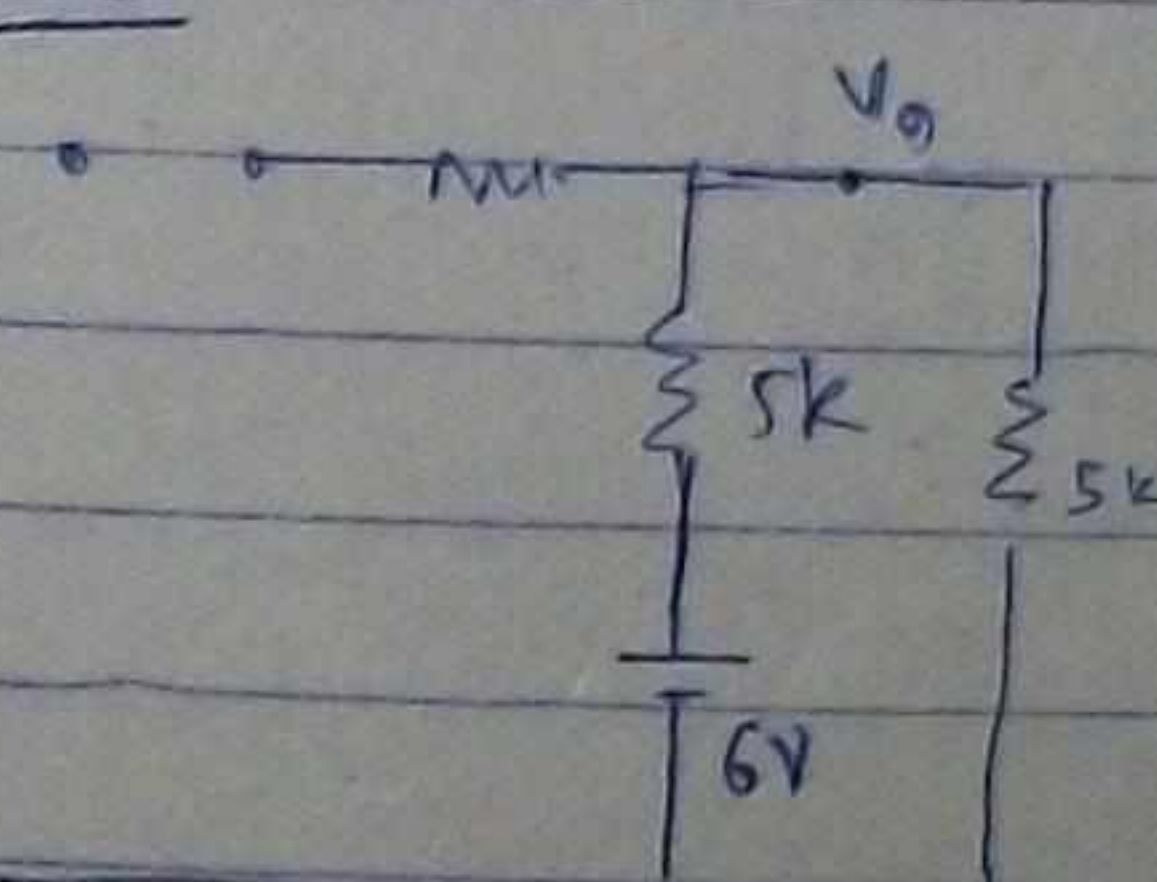


Ques. Assume that diodes are ideal make a plot of V_o against V_i for the range of V_i from 0 to 50V. Indicate all slope and voltage levels. Indicate for each region which diodes are conducting.



V_i	D_1	D_2	D_3	V_o
$0 < V_i < 3$	ON	OFF	ON	3
$3 < V_i < 9$	ON	OFF	ON	$\frac{V_i}{2} + \frac{3}{2}$
$9 < V_i < 30$	OFF	OFF	ON	$\frac{2}{3} V_i$
$30 < V_i$	OFF	ON	ON	$\frac{4}{7} V_i + \frac{20}{7}$

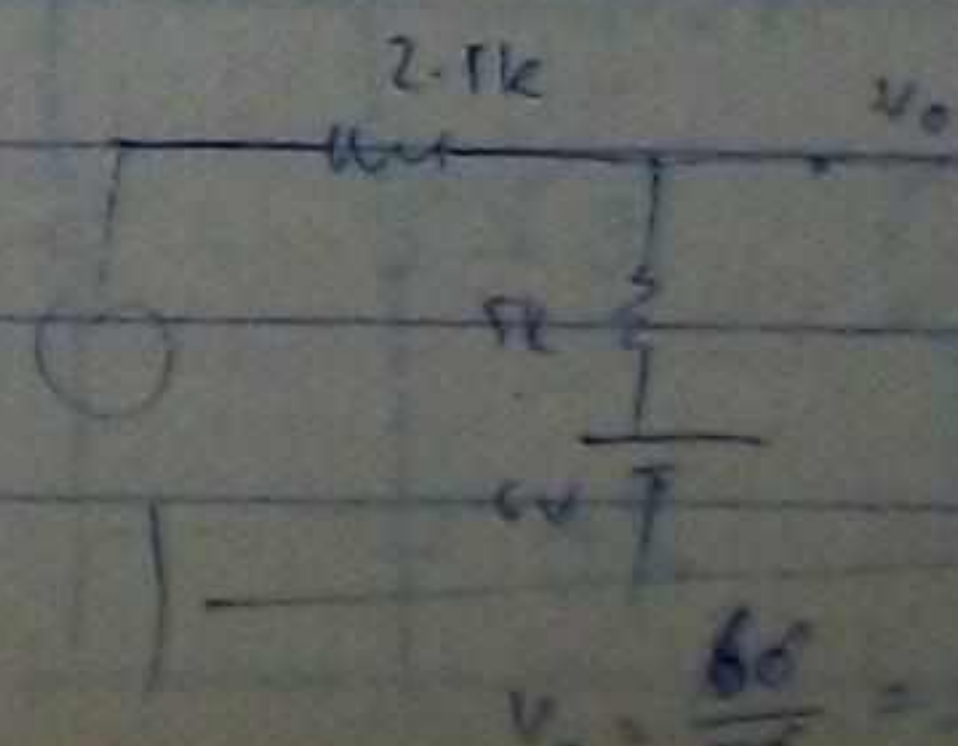
$V_i = 0$



$6 = 10I$
 $I = \frac{6}{10}$

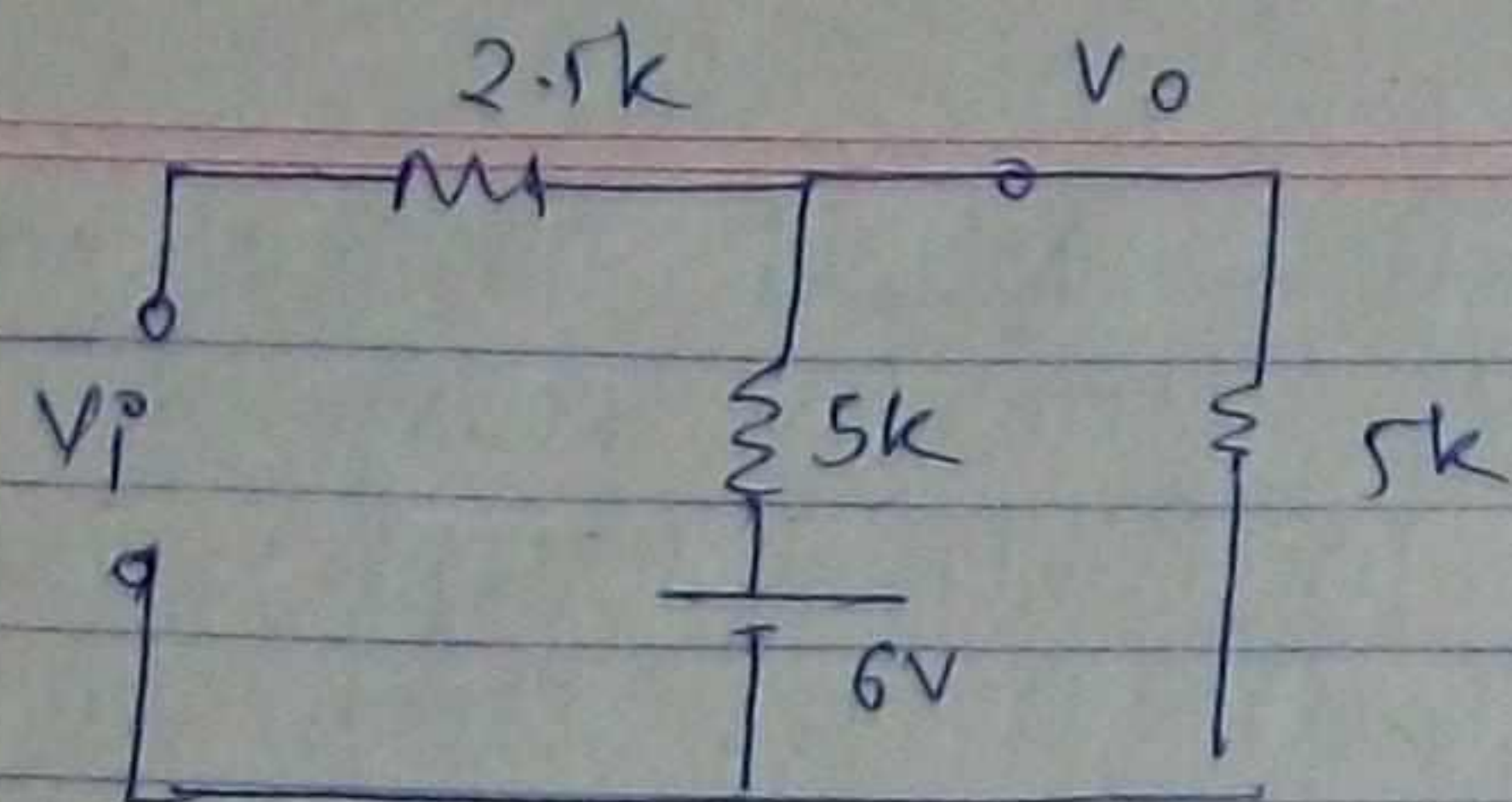
$\frac{V-6}{5} + \frac{V}{5} = 0$
 $2V = 6$
 $V = 3$

$V_i > 3$



$\frac{V_o - 6}{5} + \frac{V_o}{2.5} = 0$
 $V_o (\frac{1}{5} + \frac{2}{5}) = \frac{6}{5}$
 $V_o (\frac{3}{5}) = \frac{6}{5}$
 $V_o = \frac{6}{3} = 2$

$$V_i > 3$$



$$\frac{V_o - V_i}{2.5} + \frac{V_o - 6}{5} + \frac{V_o}{5} = 0$$

$$V_o \left(\frac{10}{25} + \frac{1}{5} + \frac{1}{5} \right) = \frac{V_i \cdot 10}{25} + \frac{6}{5}$$

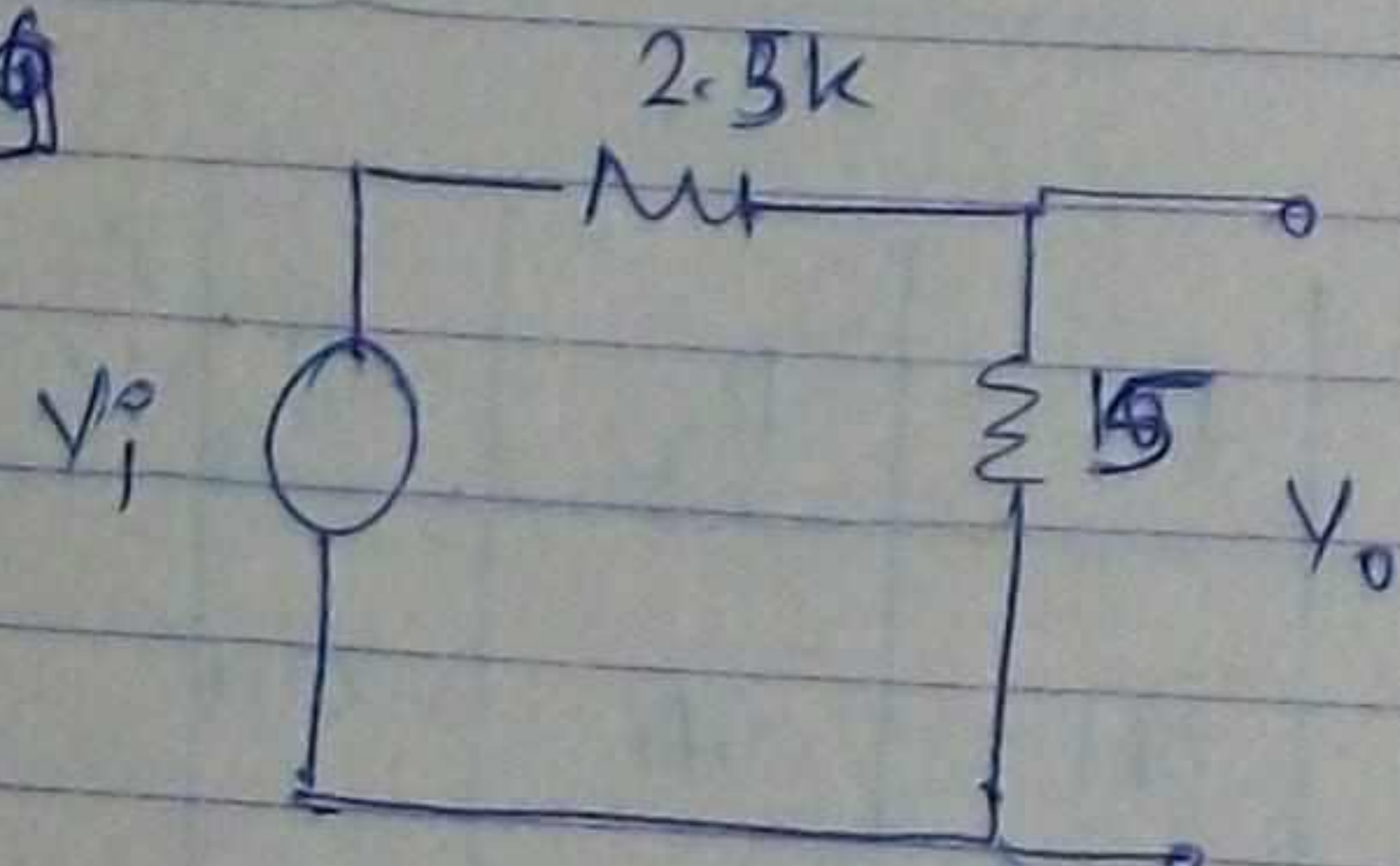
$$V_o \left(\frac{20}{25} \right) = \frac{10 V_i + 30}{25}$$

$$V_o = \frac{V_i}{2} + \frac{3}{2}$$

$$\boxed{V_i = 9}$$

$$\boxed{V_o = 9}$$

$$V_i > 9$$

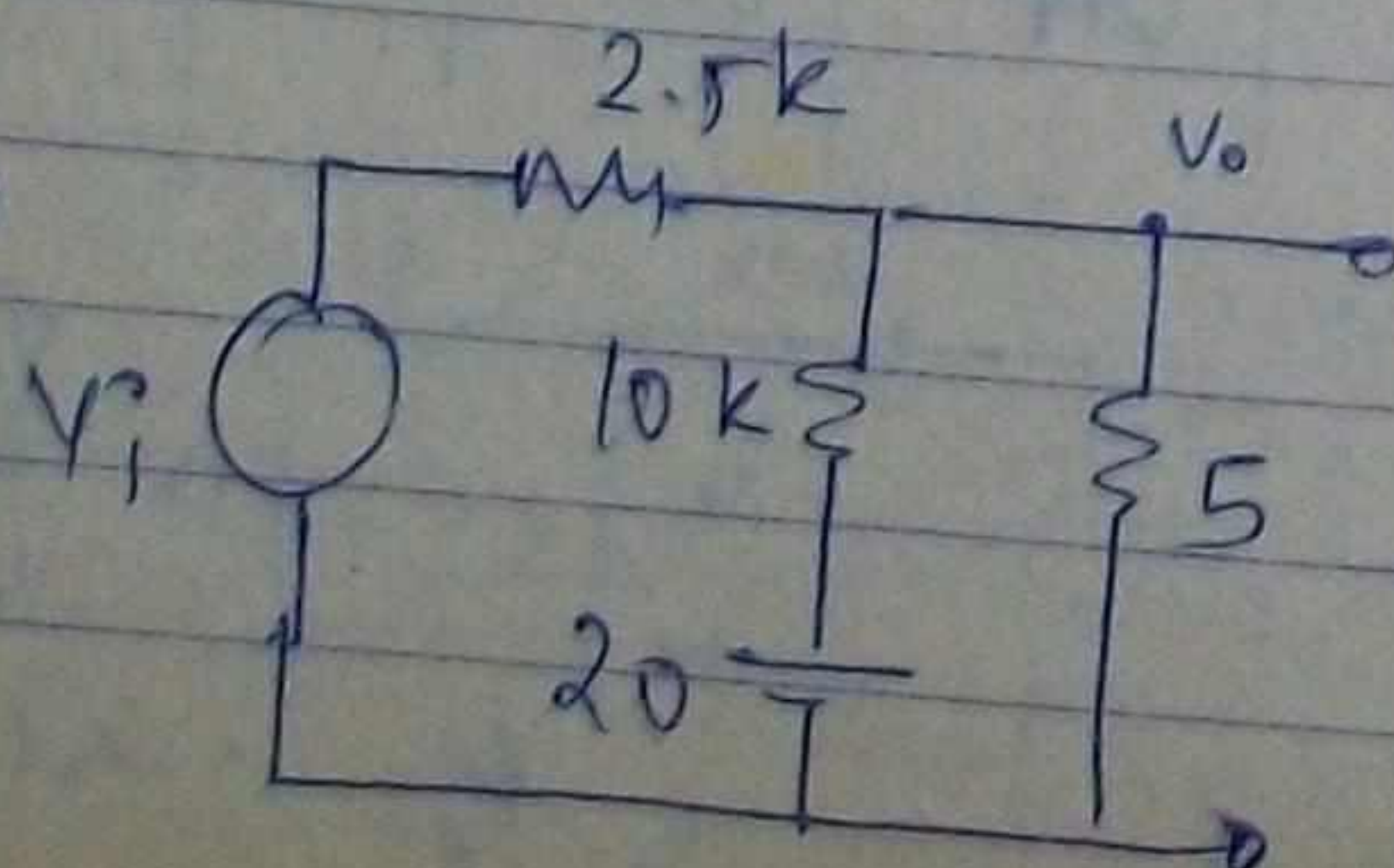


$$\boxed{V_o = \frac{5}{7.5} V_i} = \frac{2}{3} V_i$$

when $V_o \geq 20$ D_2 will ON

$$\frac{2}{3} V_i \geq 20 \quad V_i \geq 30$$

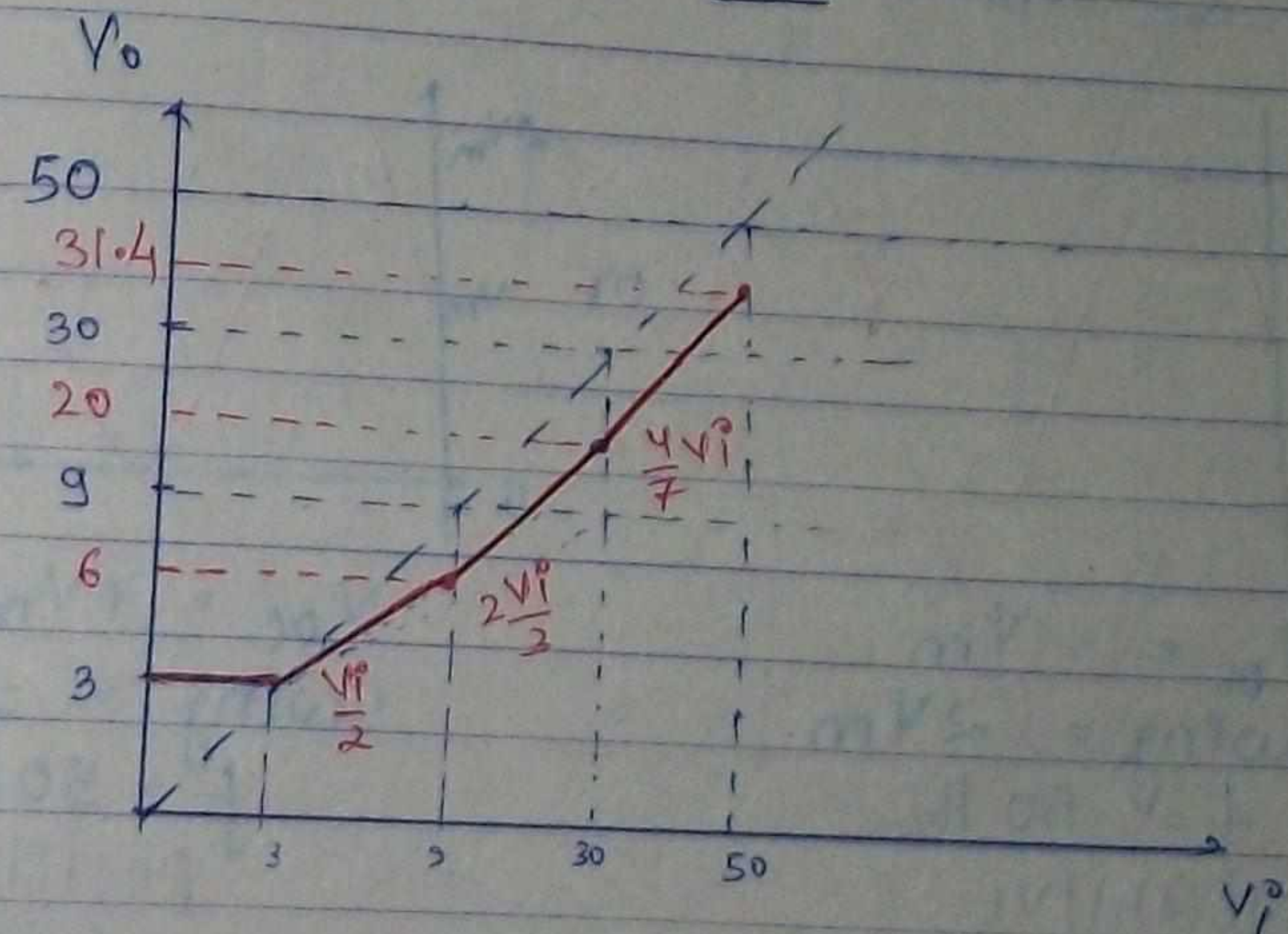
$$V_i > 30$$



$$V_o \left[\frac{1}{2.5} + \frac{1}{10} + \frac{1}{5} \right] = \frac{20}{10} + \frac{V_i}{25}$$

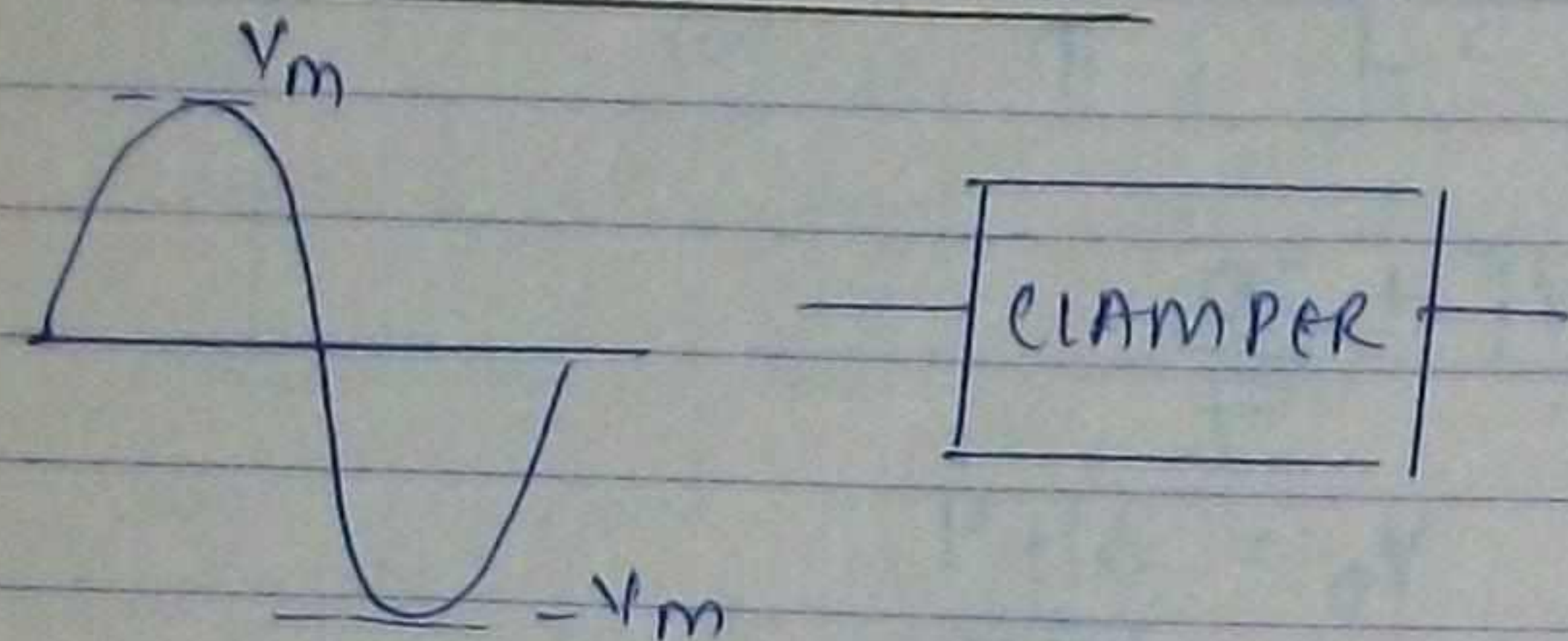
$$V_o = \frac{4}{7} V_i + \frac{20}{7}$$

for $V_i = 50$ $V_o = \underline{\underline{31.4}}$

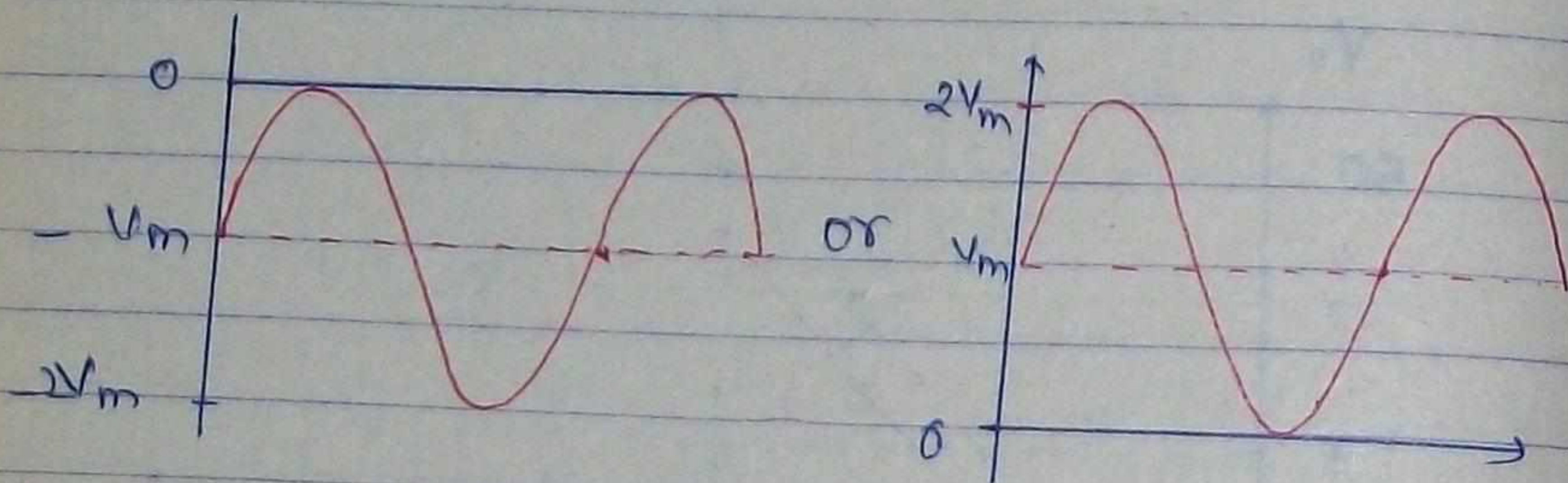


DC restorer
DC translator
DC inserter

⇒ CLAMPER CIRCUITS -



$$\begin{aligned} DC &= 0 \\ \text{Swing} &= V_{\max} - V_{\min} \\ &= 2V_m \\ f &= 50 \text{ Hz} \end{aligned}$$



$$\begin{aligned} V_{DC} &= -V_m \\ \text{Swing} &= 2V_m \\ f &= 50 \text{ Hz} \end{aligned}$$

NEGATIVE CLAMPER

$$\begin{aligned} V_{DC} &= +V_m \\ \text{Swing} &= 2V_m \\ f &= 50 \text{ Hz} \end{aligned}$$

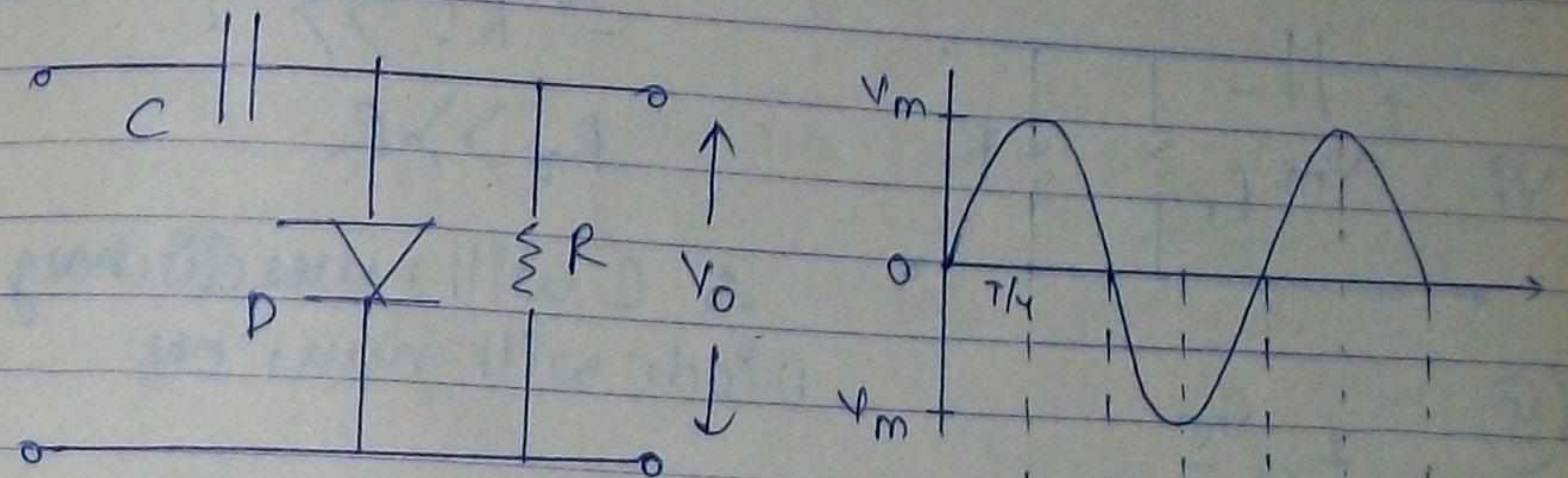
POSITIVE CLAMPER

- These are used to add DC level as per the requirement to AC output signal. These are also called 'DC inserter', 'DC restorer' or 'DC translator'.

- These are of two types :
1. Negative clamper
2. Positive clamper

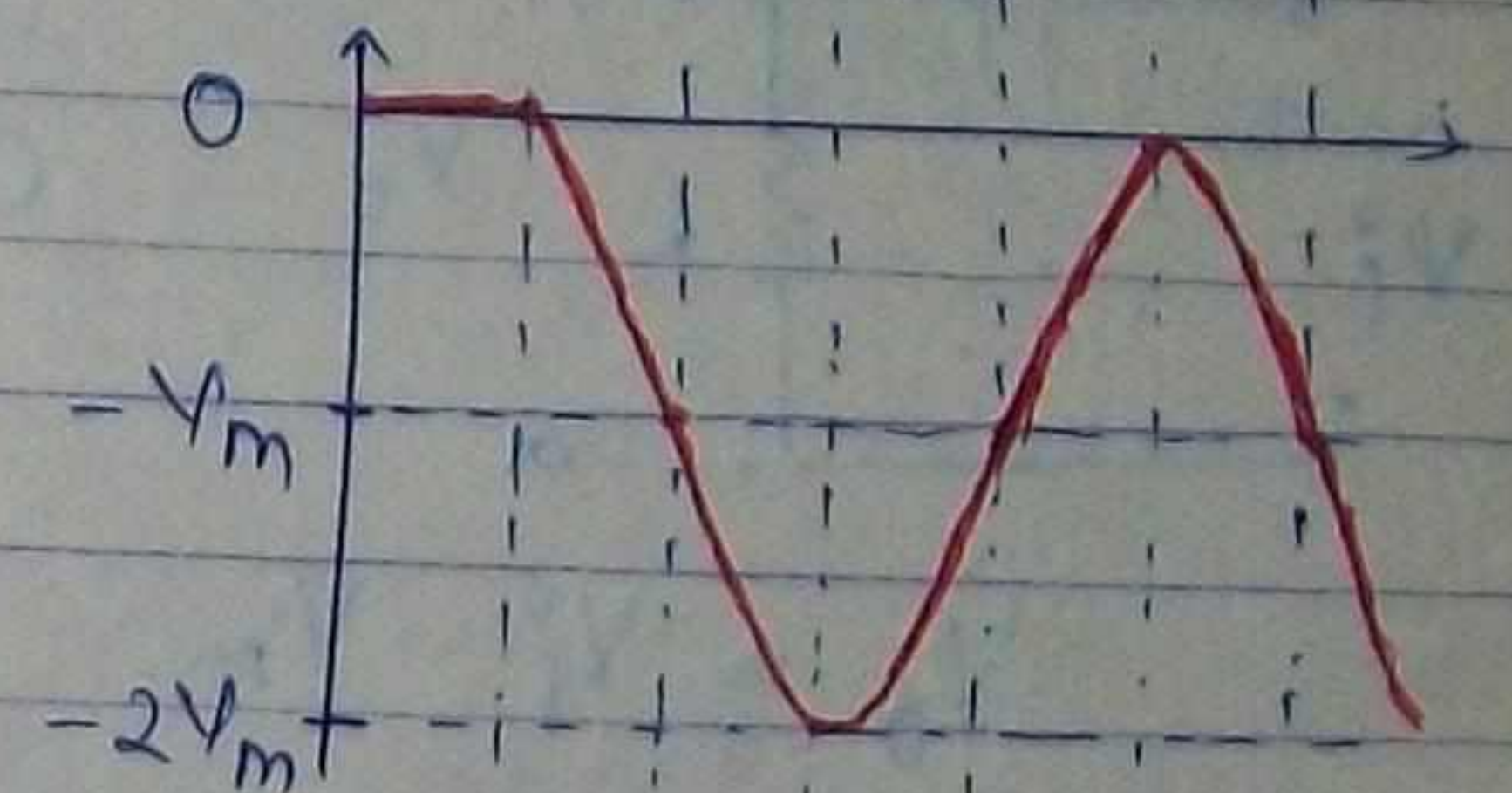
① NEGATIVE CLAMPER -

which add a negative dc level to ac signal.

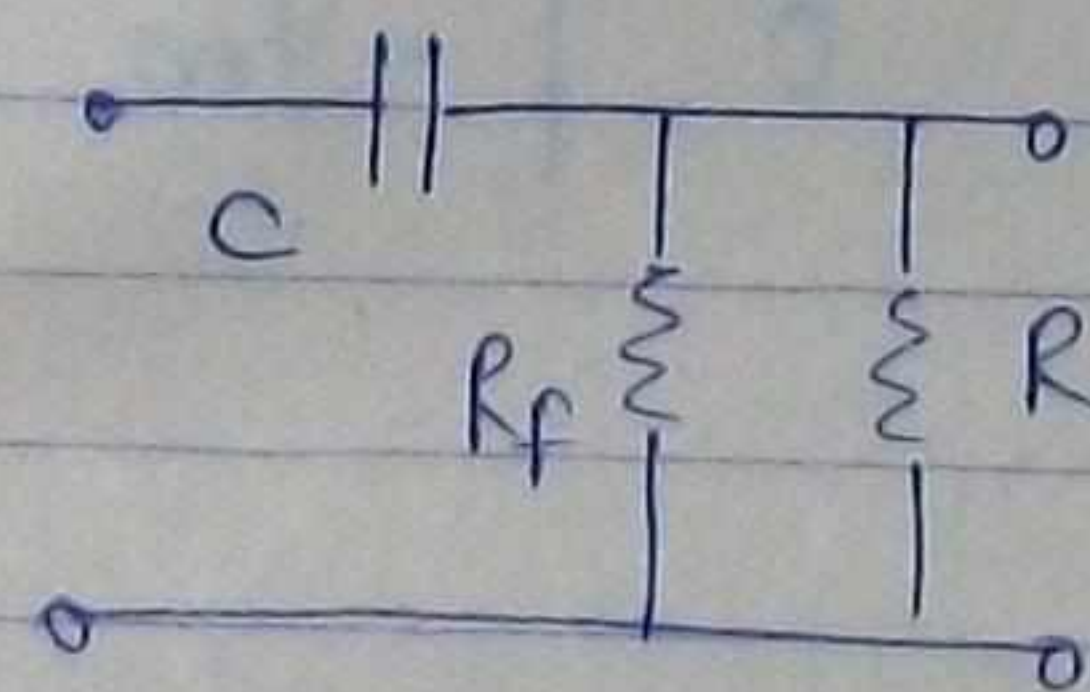


$$V_C(0^-) = V_C(0^+) = 0$$

Initially 'C' acts as short circuit

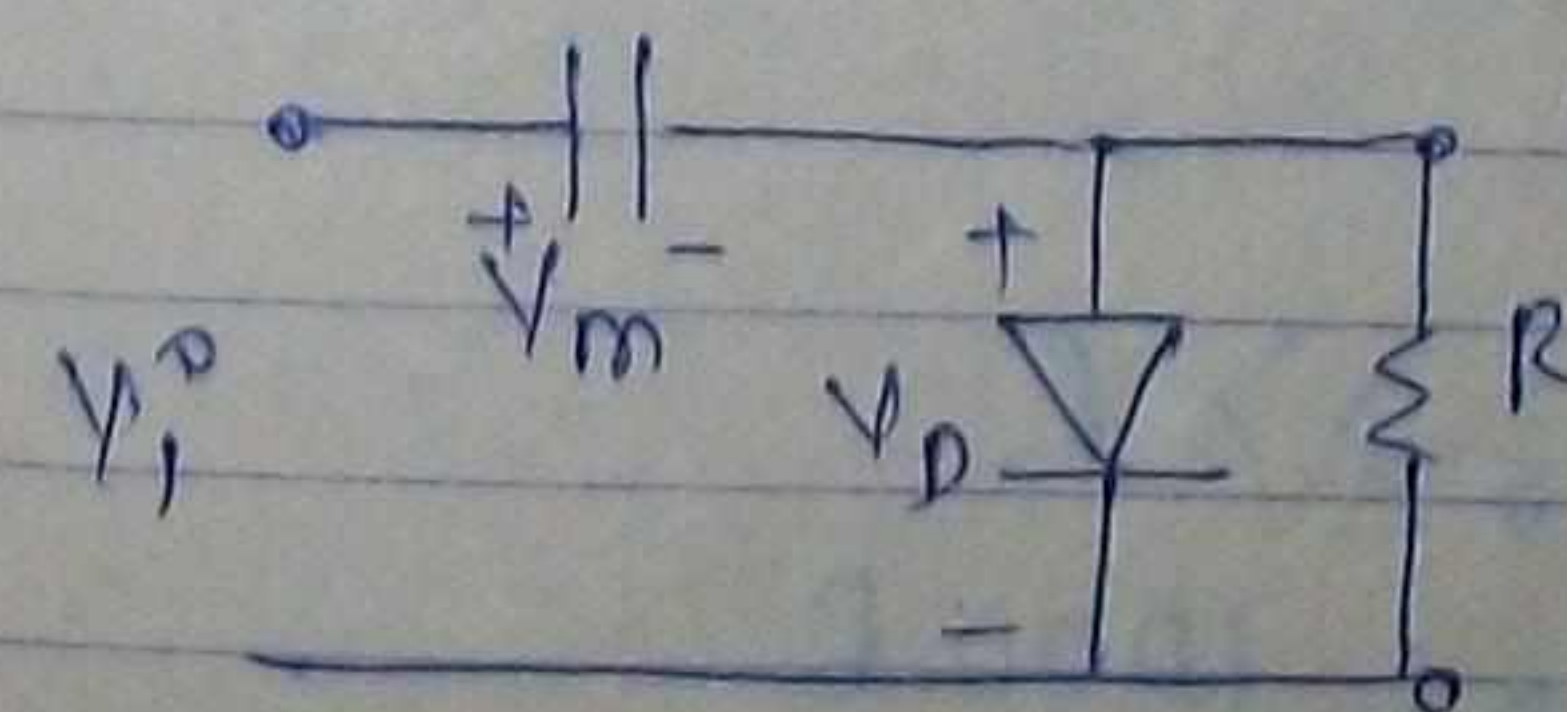


① During +ve half
D → ON



$$\begin{aligned} \tau &= (R_F || R)C \approx R_F C < T \\ R &\gg R_F \end{aligned}$$

② At $t = T/4$ $V_i = V_m$, $V_C = V_m$



$$V_D = V_i - V_m \leq 0$$

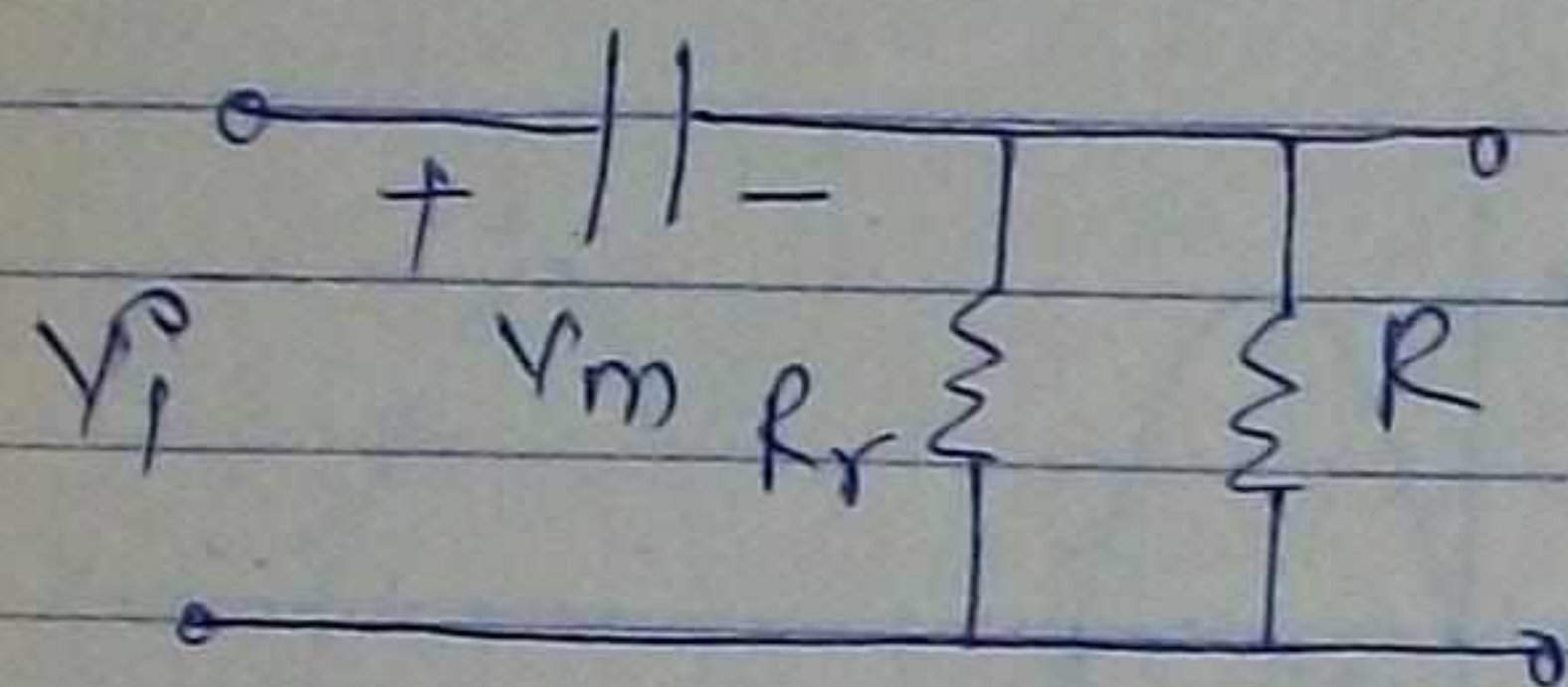
③ $t > T/4$, $V_i \downarrow$, $V_i < V_m$, $V_D < 0$

$D \rightarrow \text{Off}$

$$\tau = (R_f \parallel R) C$$

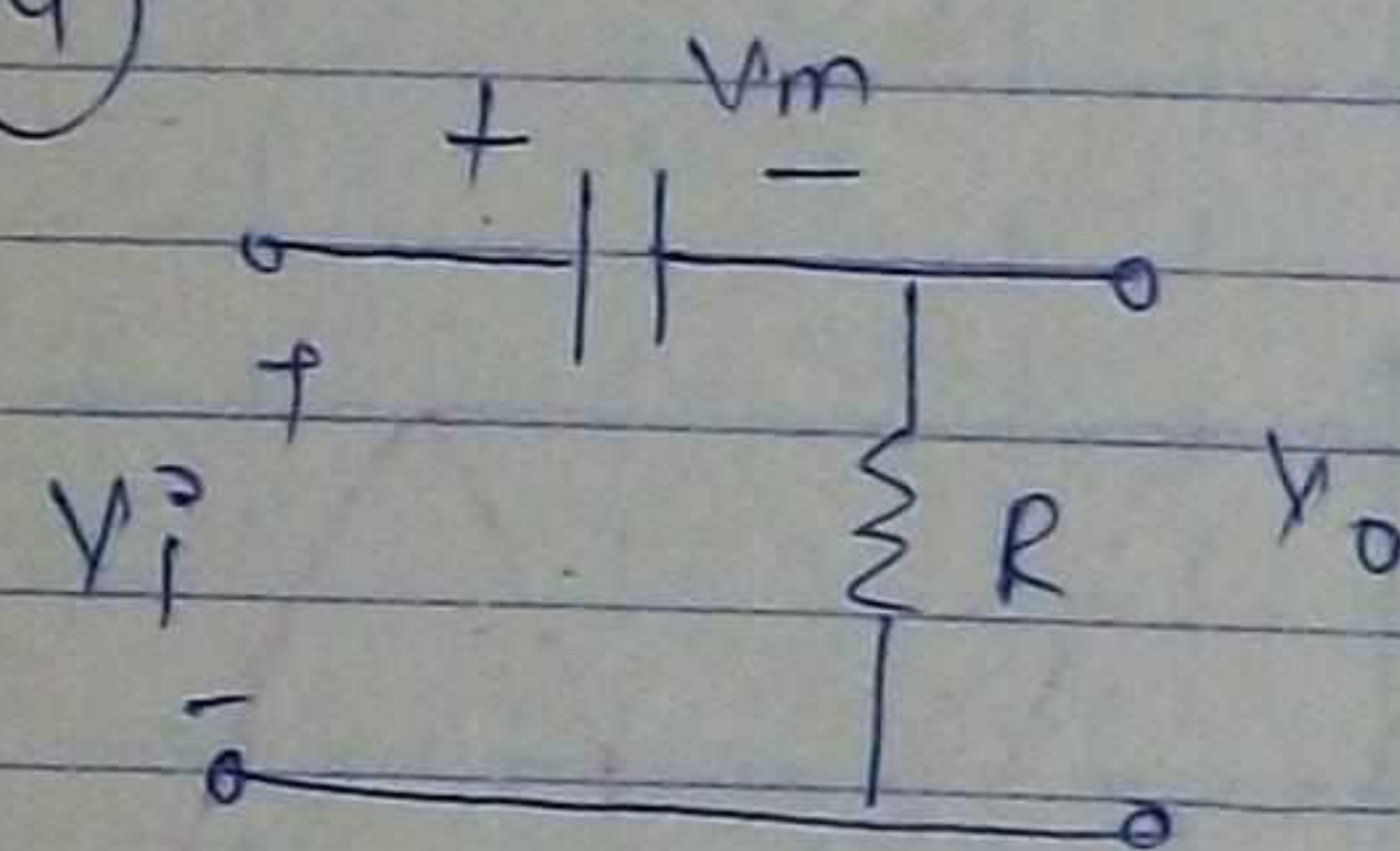
$$\approx RC \gg T$$

$$R_f \gg R$$



$\therefore C$ will never discharge
Diode will never on

④



C will act as a battery

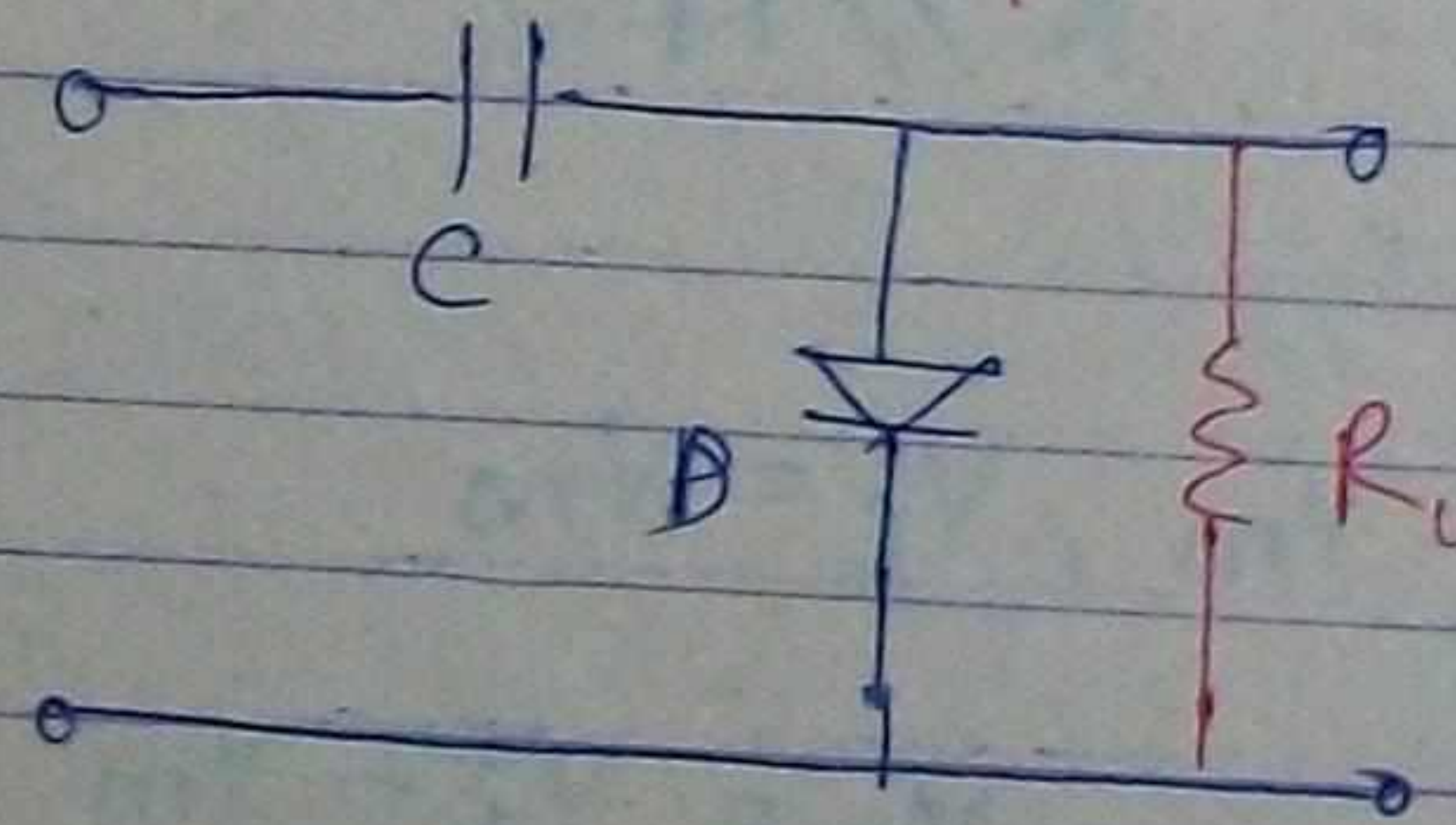
$$V_o = V_i - V_m$$

O/P $\rightarrow$

V_i	V_o
V_m	0
0	$-V_m$
$-V_m$	$-2V_m$
0	$-V_m$

#

Ideally $R = \infty$



$$R_f \ll R$$

$$R_f \gg R$$

$$R_f = \frac{R}{1000}$$

$$R_f = 1000 R$$

$$R_f R_r = R^2$$

$$R = \sqrt{R_f R_r}$$

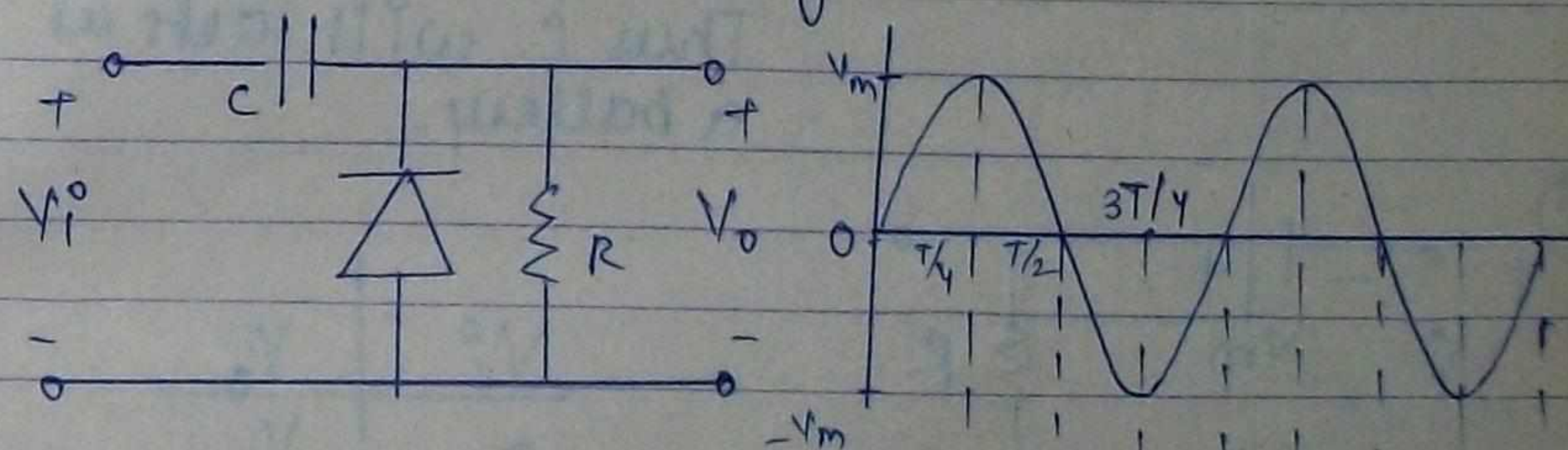
is G.M. of R_f & R_r

— During first positive cycle of I/P voltage V_i ($0 \leq t \leq T/4$) the capacitor get charge to V_m through F.B diode D upto max. value V_m of I/P signal V_i .
The ' C ' once charge to V_m acts as a battery of voltage V_m and therefore

$$V_o = V_i - V_m$$

② POSITIVE CLAMPER —

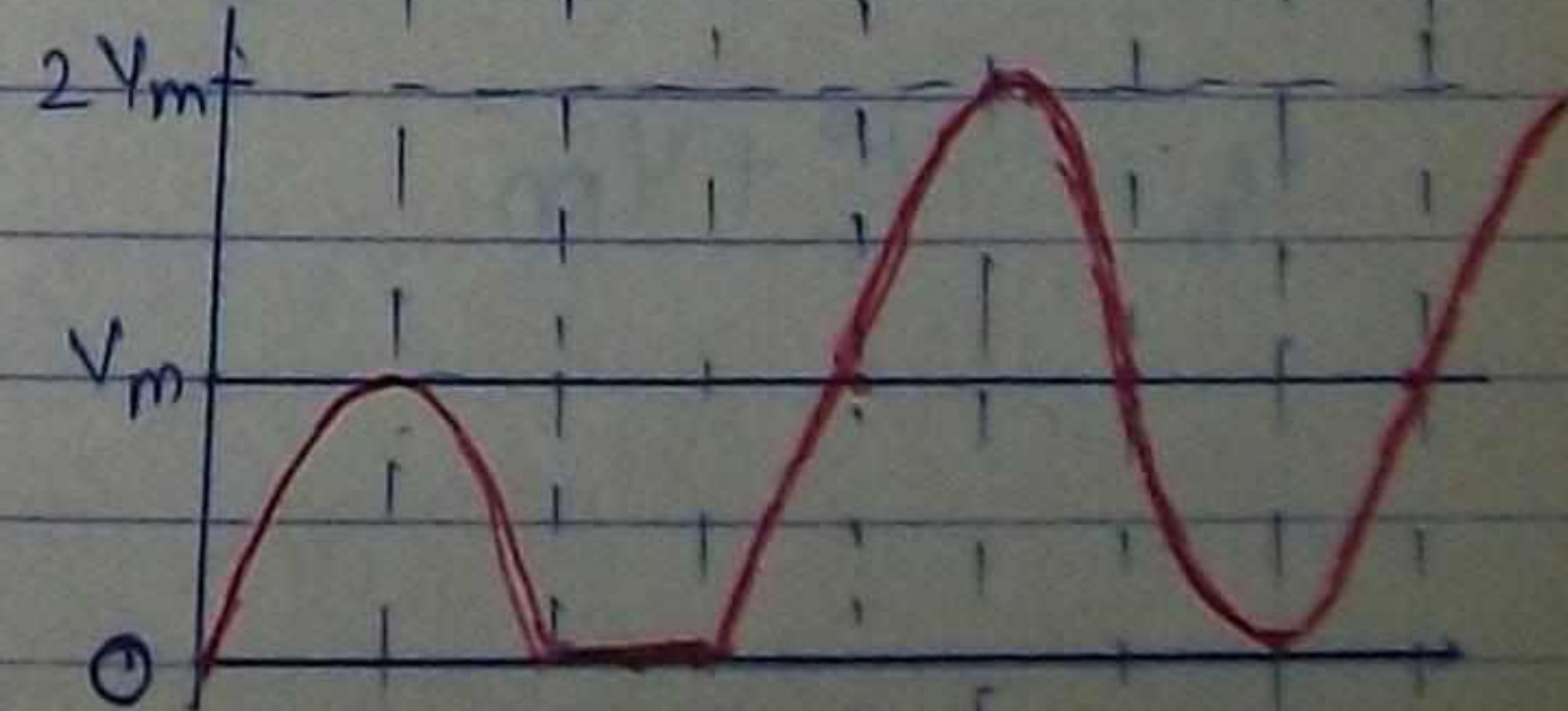
which adds a positive dc level to the a.c. signal.



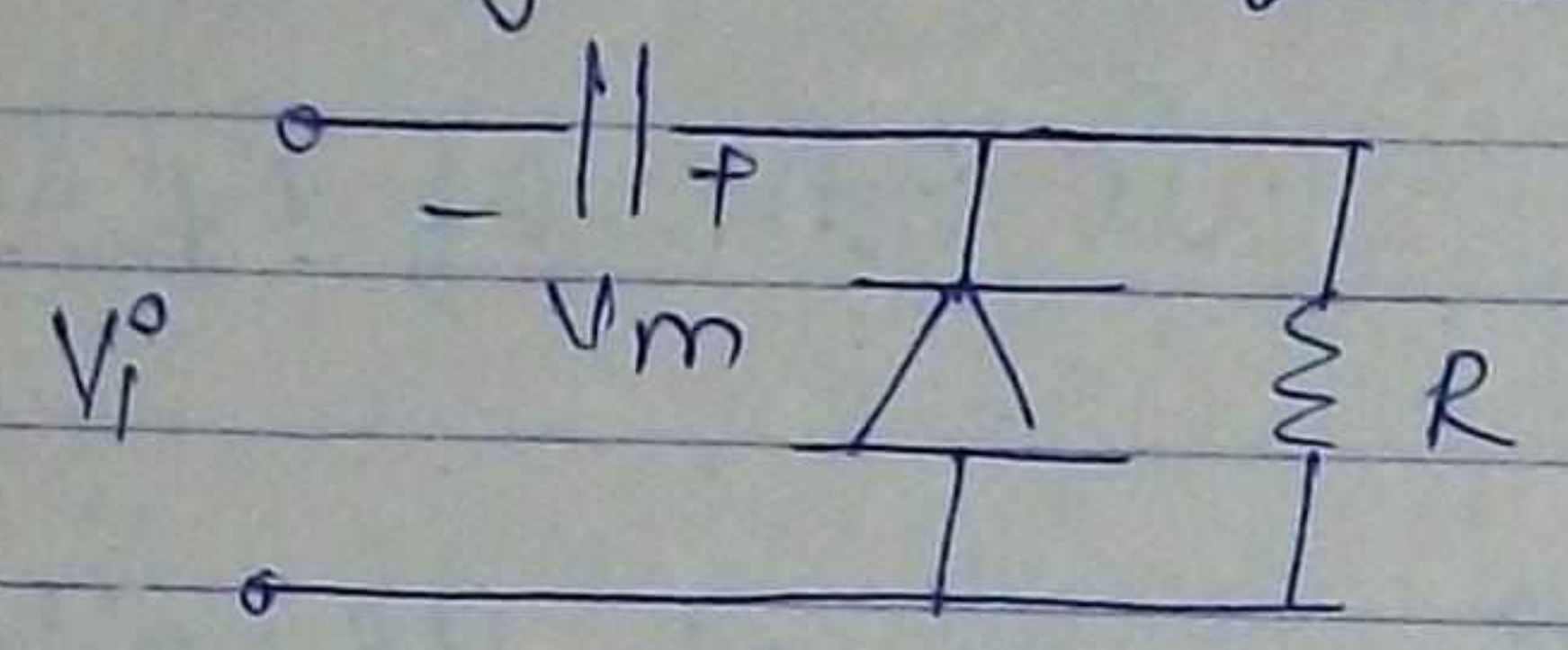
$$V_c(0^-) = 0 = V_c(0^+)$$

Initially ' C ' acts as short circuit

① During +ve half $D \rightarrow \text{off}$ $V_o = V_i$



① During -ve half cycle -



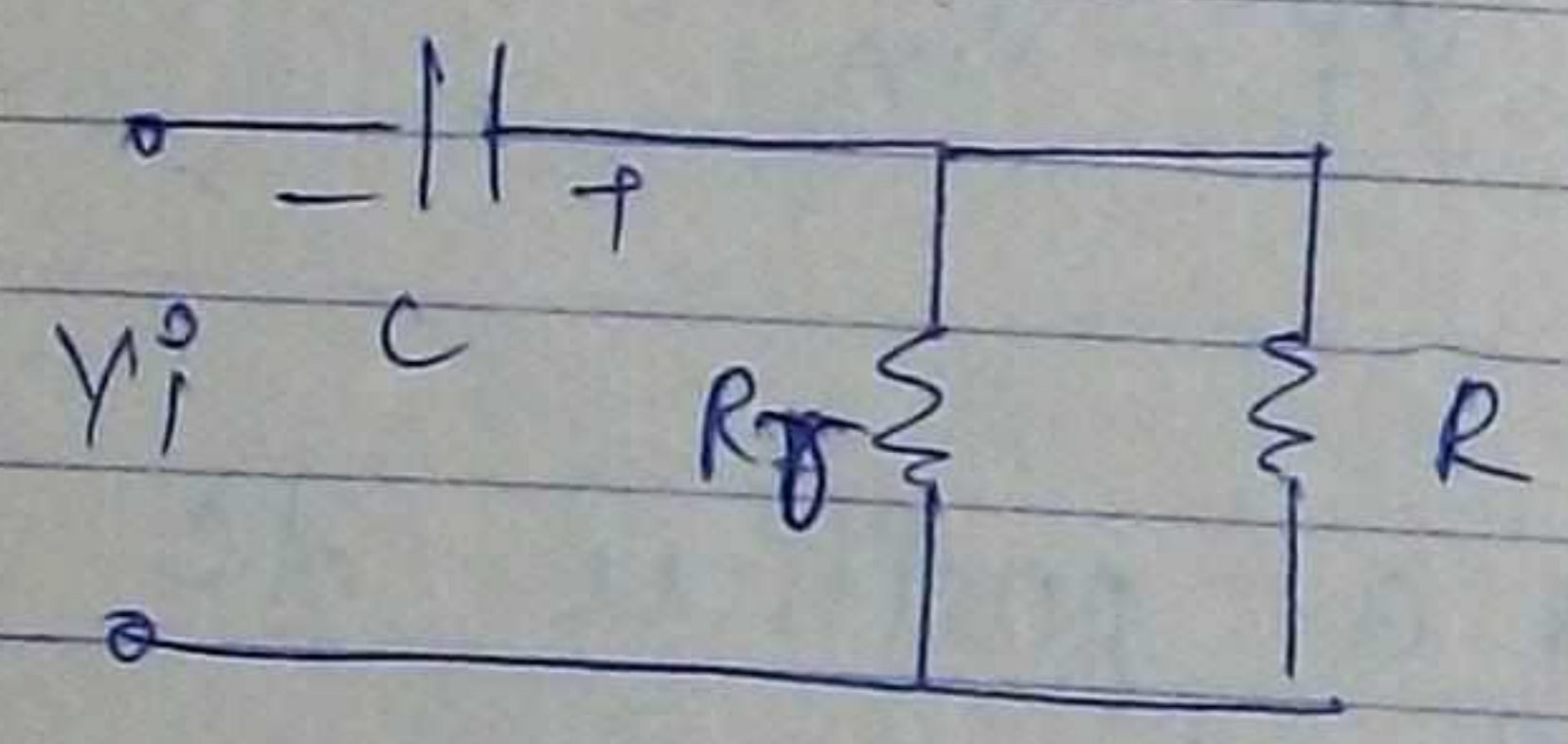
$$\tau = (R_f || R) C$$

$$\tau \approx R_f C$$

$$V_i = -V_m$$

$$V_c = -V_m$$

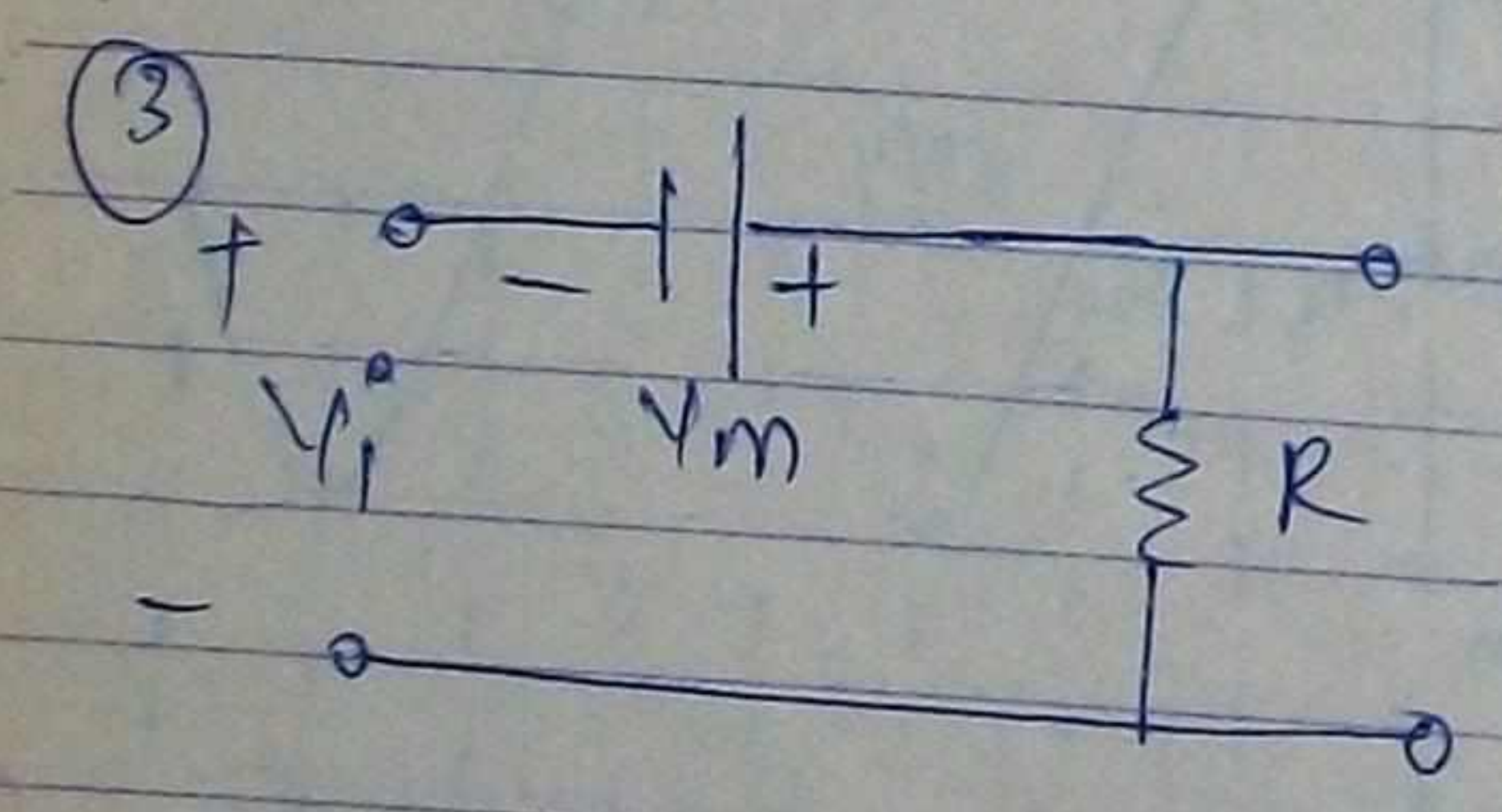
② $t > 3T/4$ $V_i \uparrow$ Diode will remain on
 $V_o = -(V_i + V_m)$
 $V_i + V_m = V_o$
 $V_i > -V_m$
 $V_o > 0$



$$\tau = (R_s || R) C$$

$$= R_s C$$

Thus C will act as a battery.

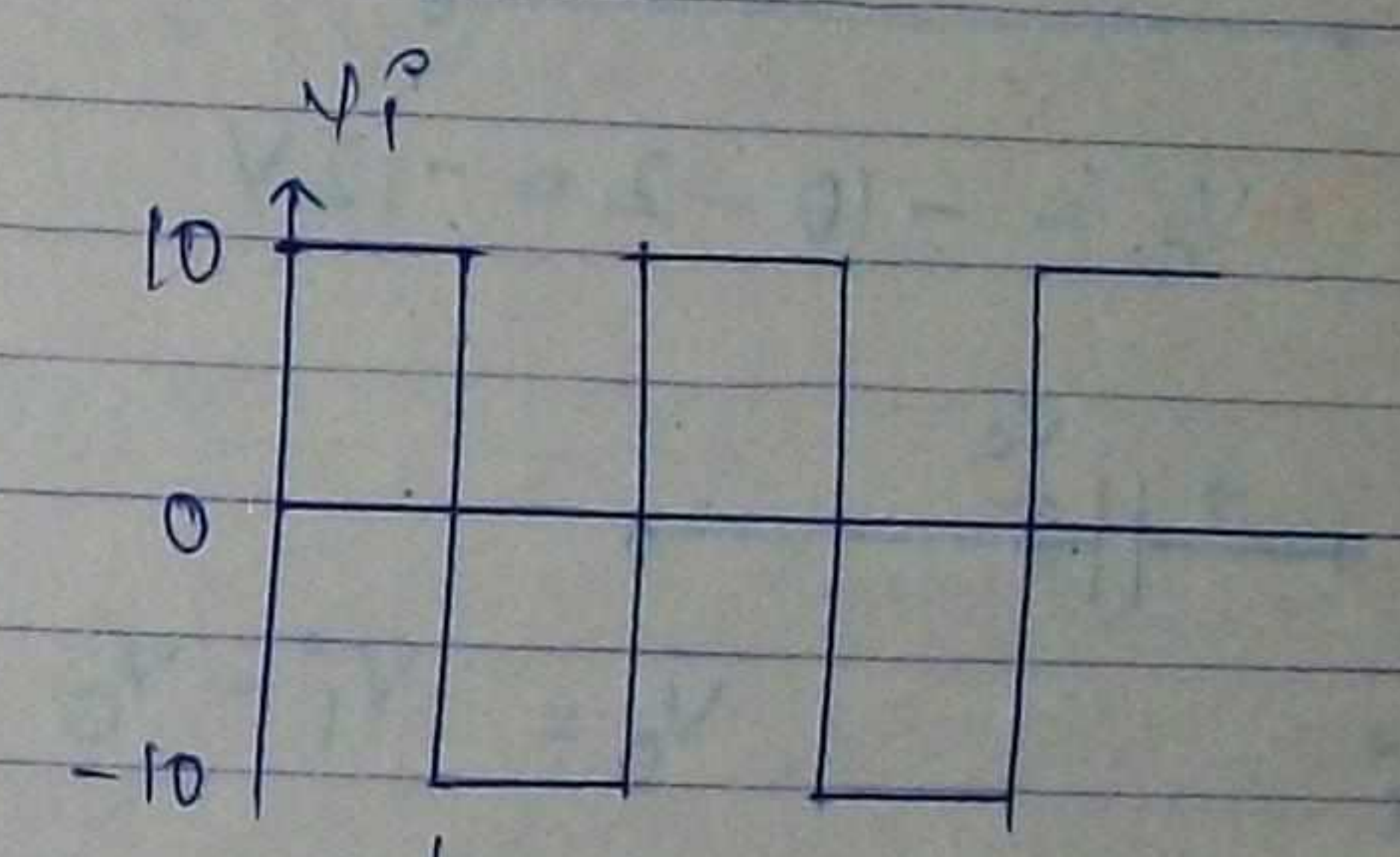
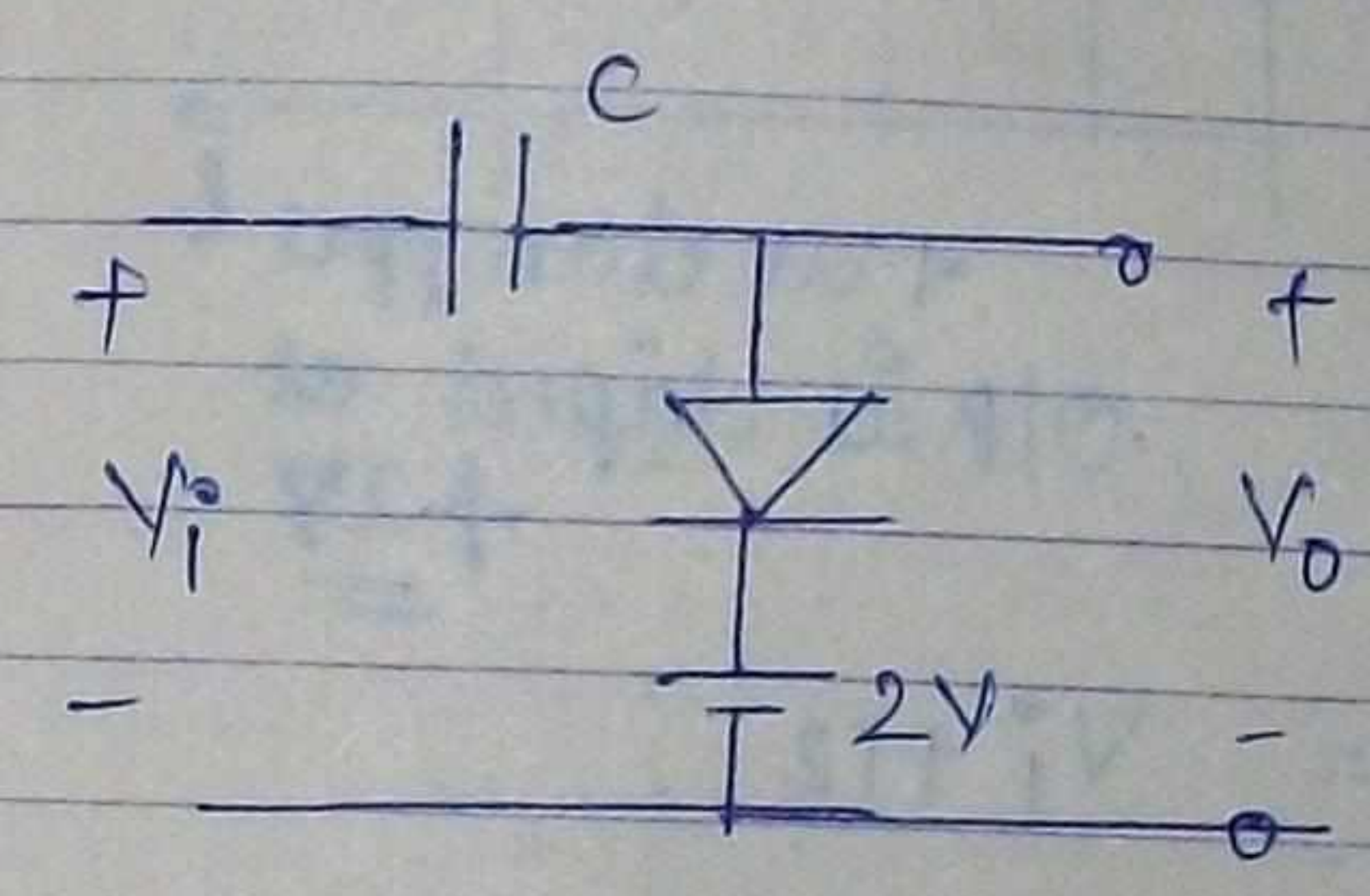


$$V_o = V_i + V_m$$

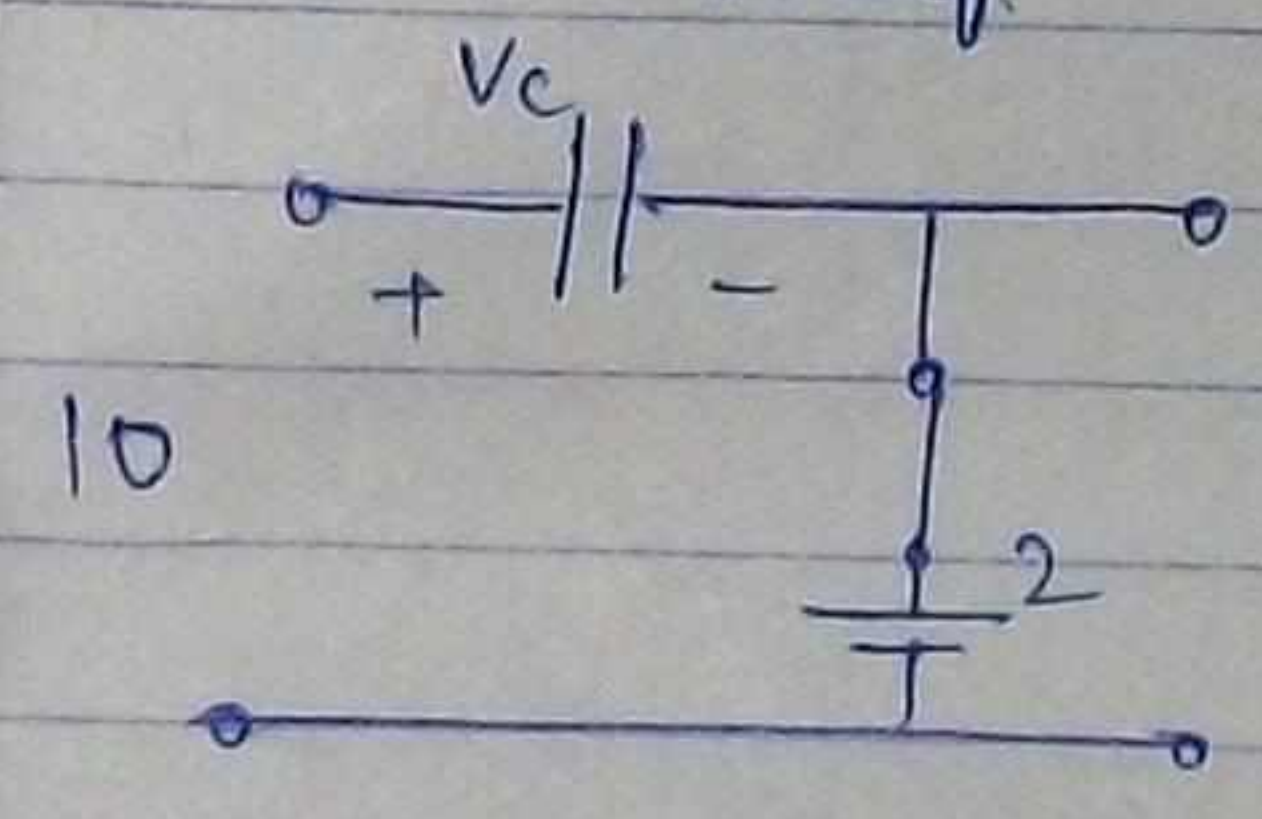
V_i	V_o
0	V_m
V_m	$2V_m$
0	V_m
$-V_m$	0

- During first -ve half of I/P signal the capacitor gets charge upto maximum negative value $-V_m$ through forward biased diode D. Once the capacitor charge upto $-V_m$ it acts as a battery of voltage $-V_m$ and therefore
 $V_o = V_i + V_m$

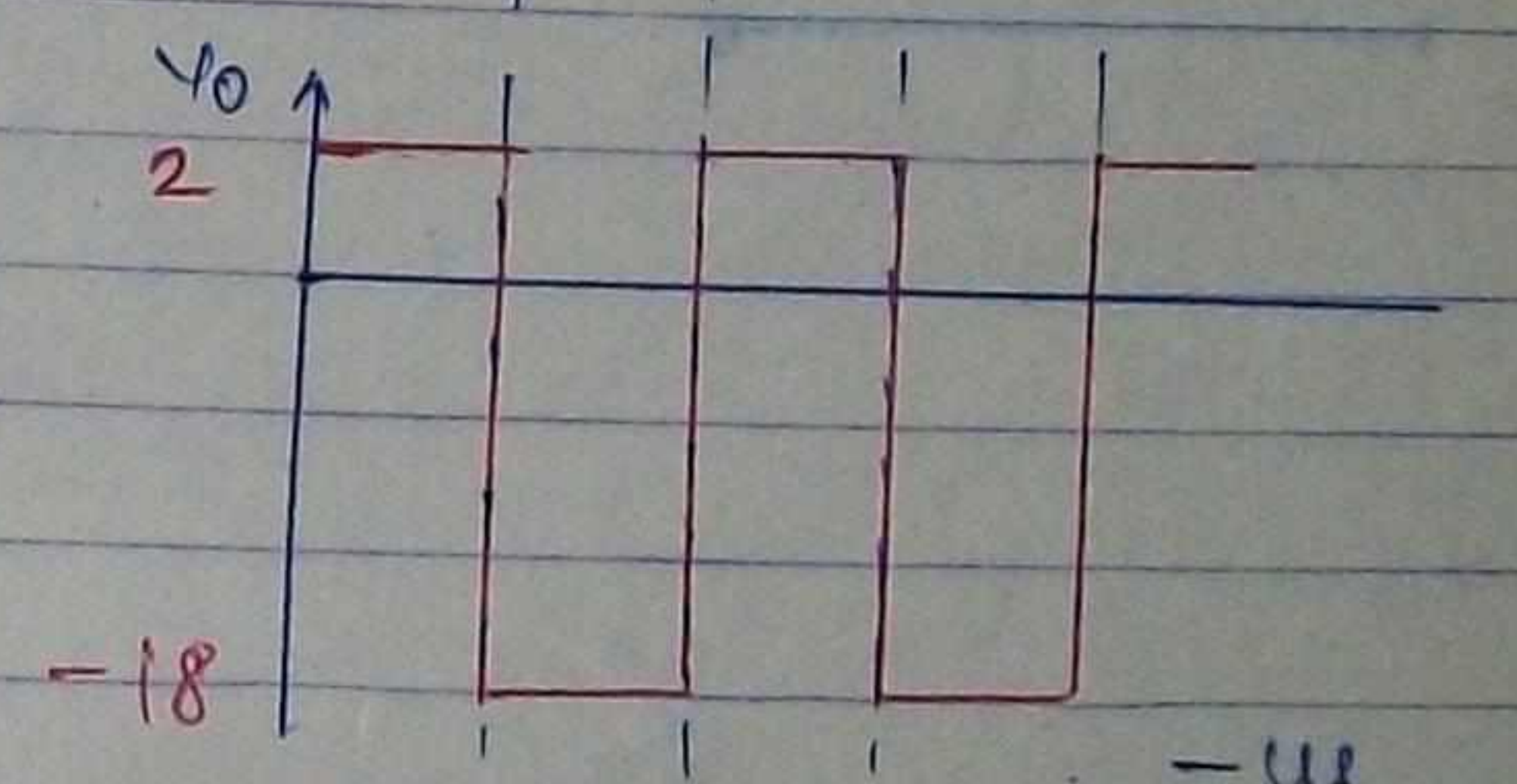
#



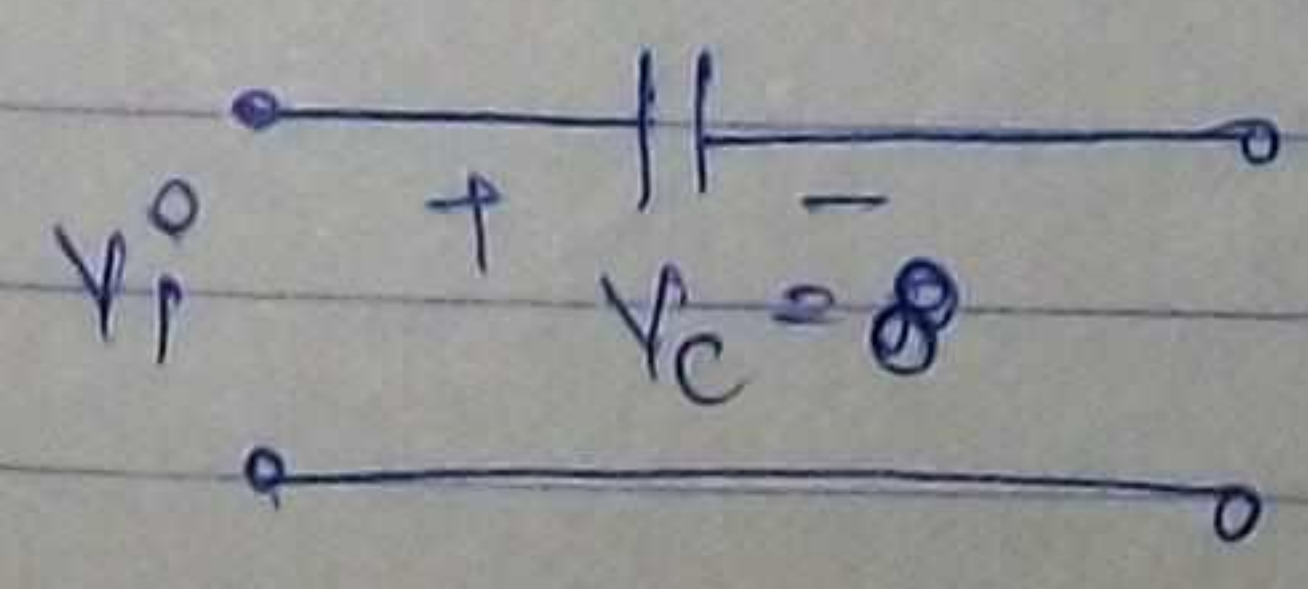
① During +ve half



$$V_c = 10 - 2 = 8V$$

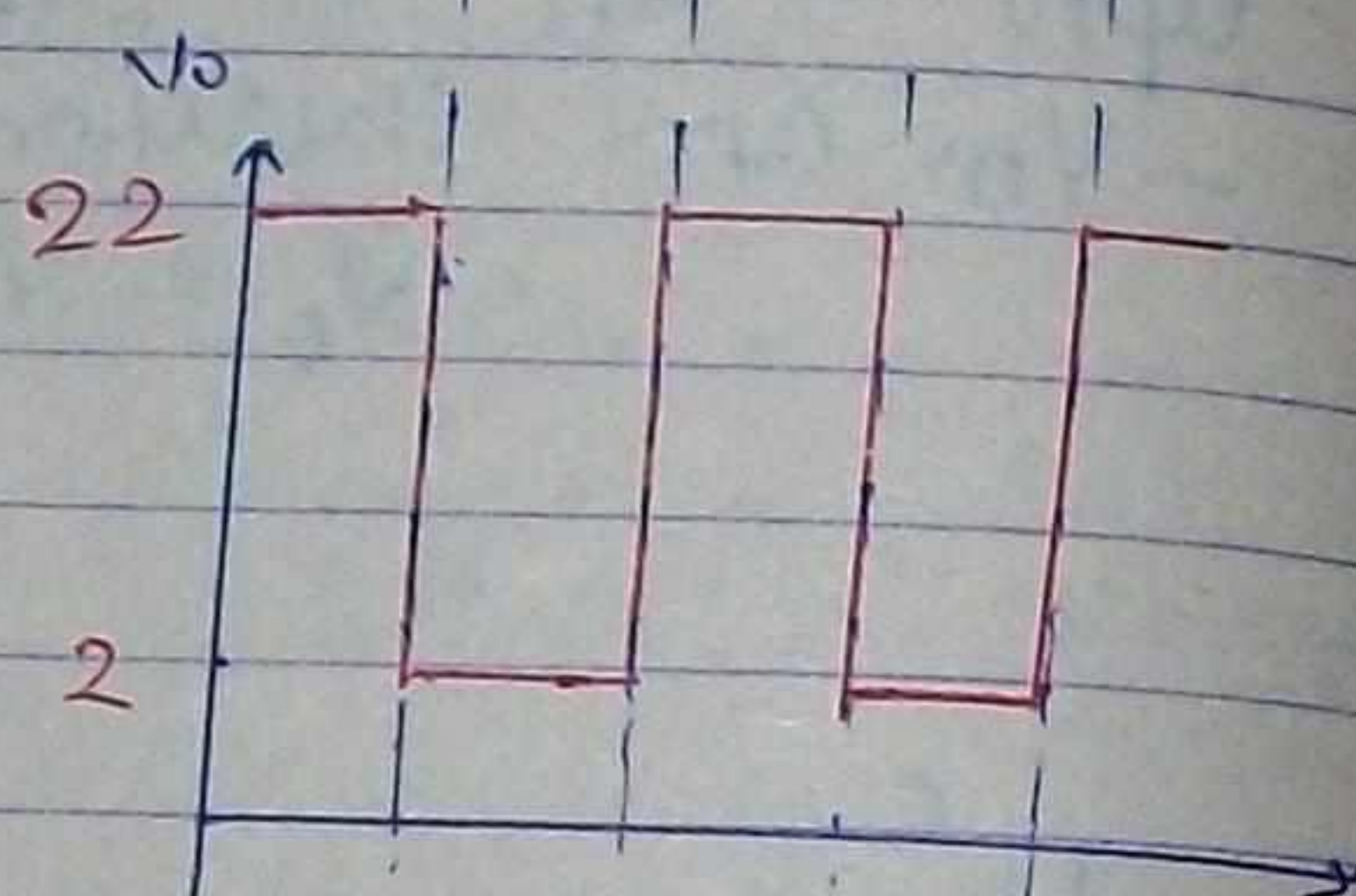
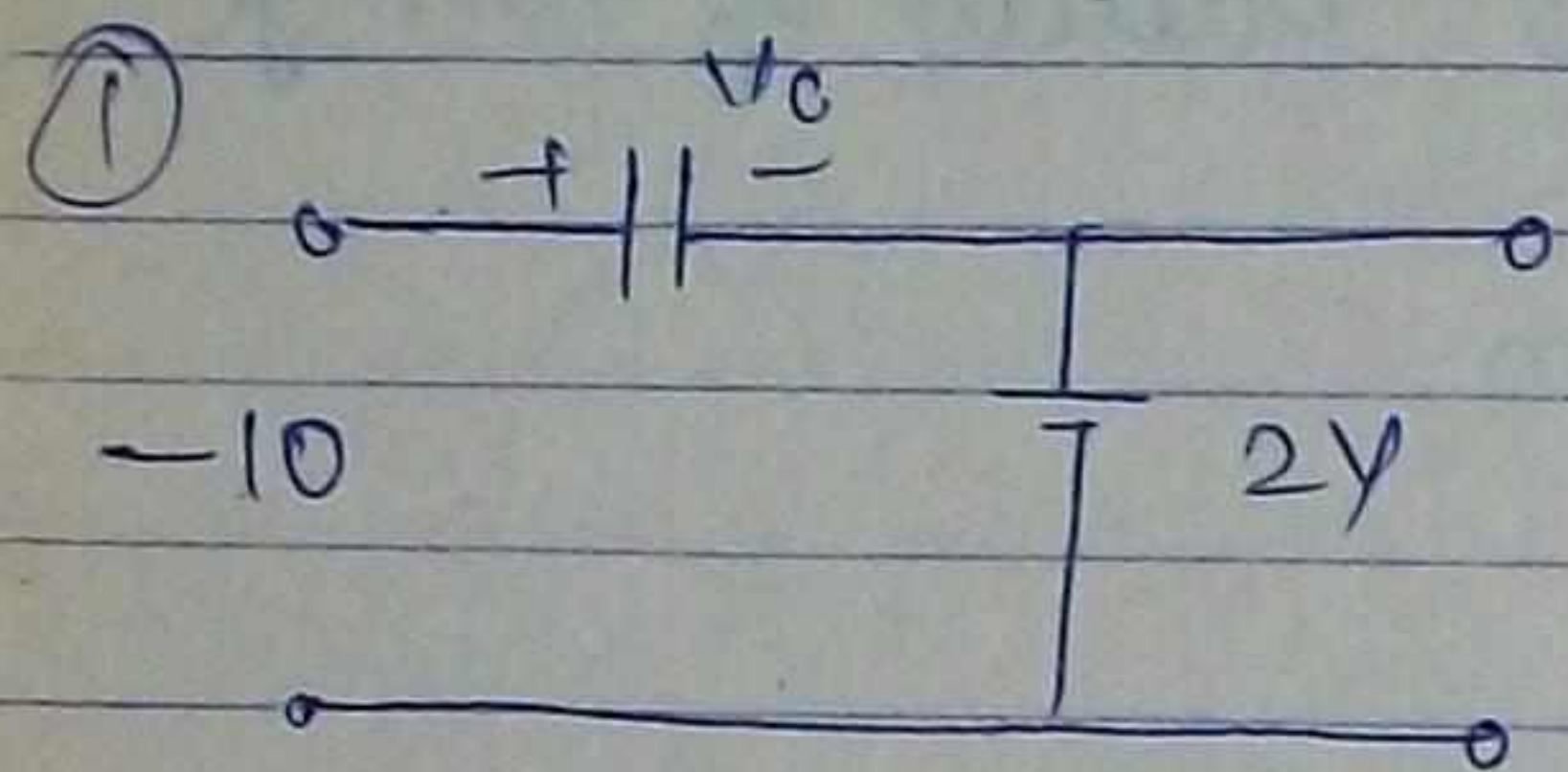
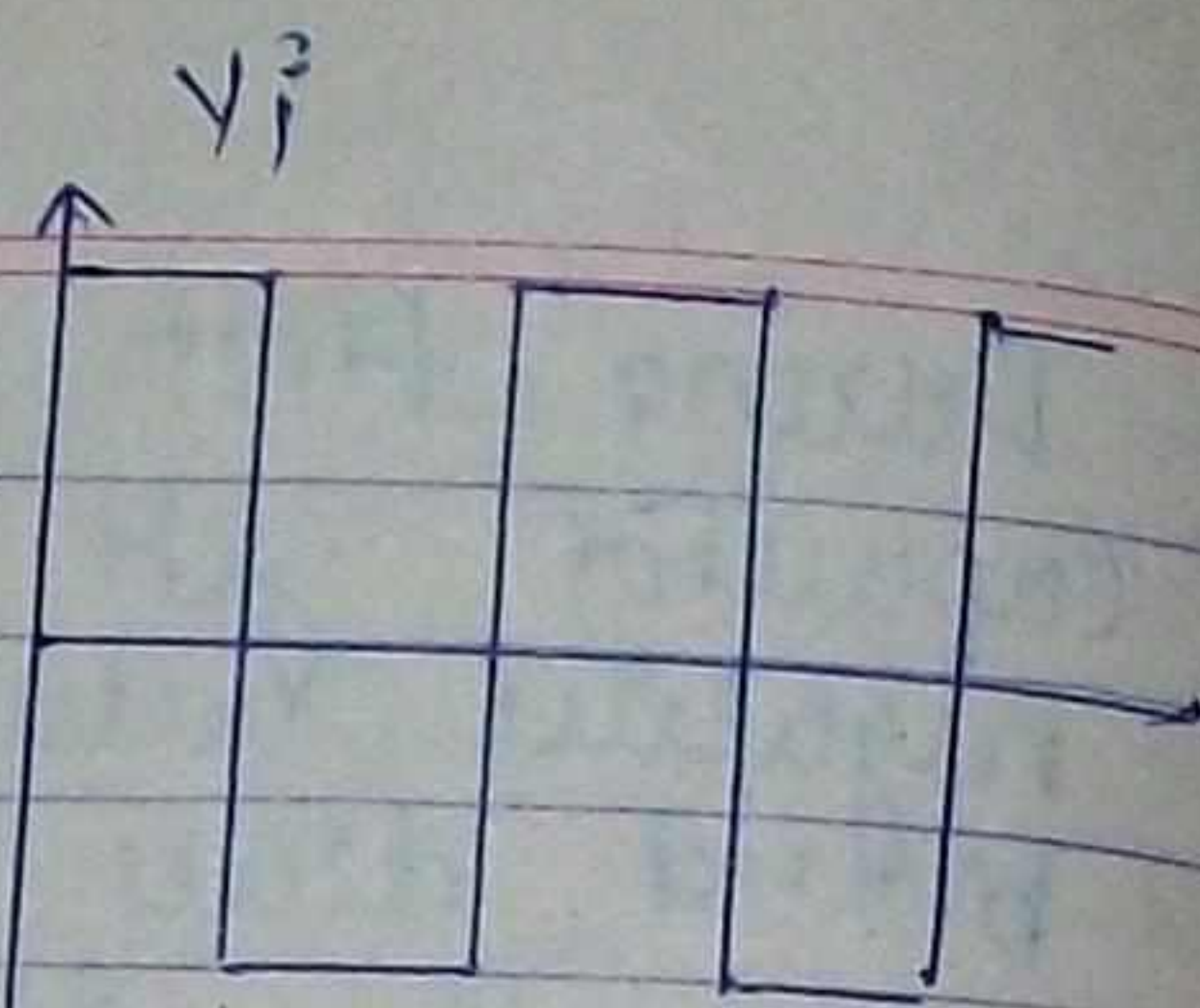
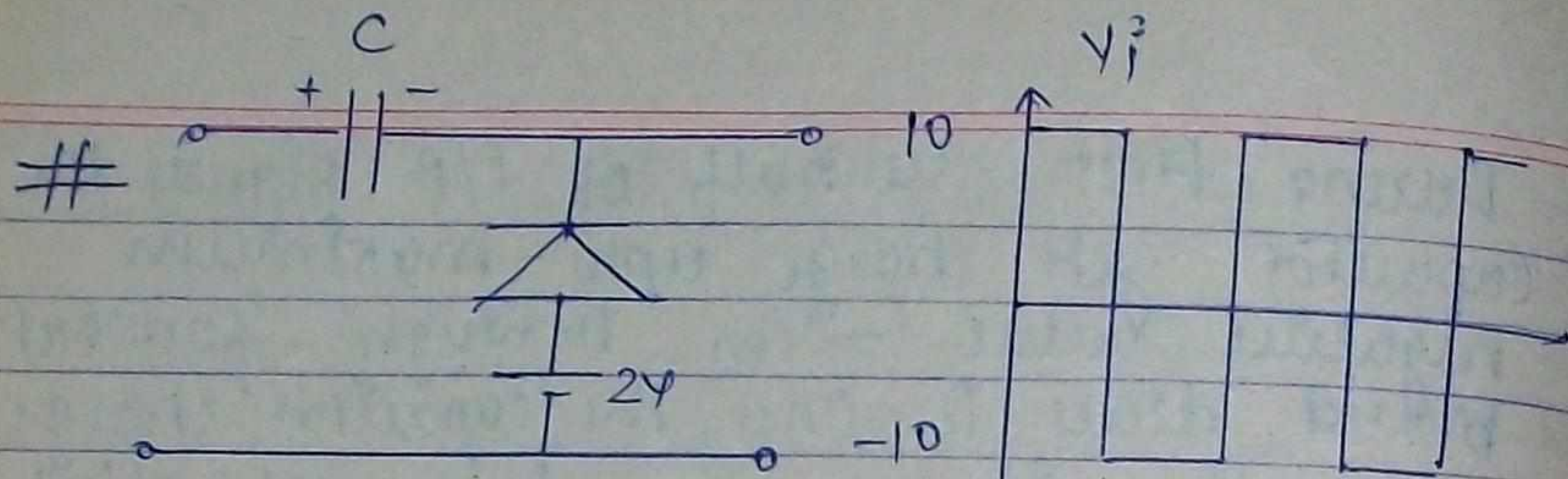


-ve clamped & O/P is clamped at 2V.

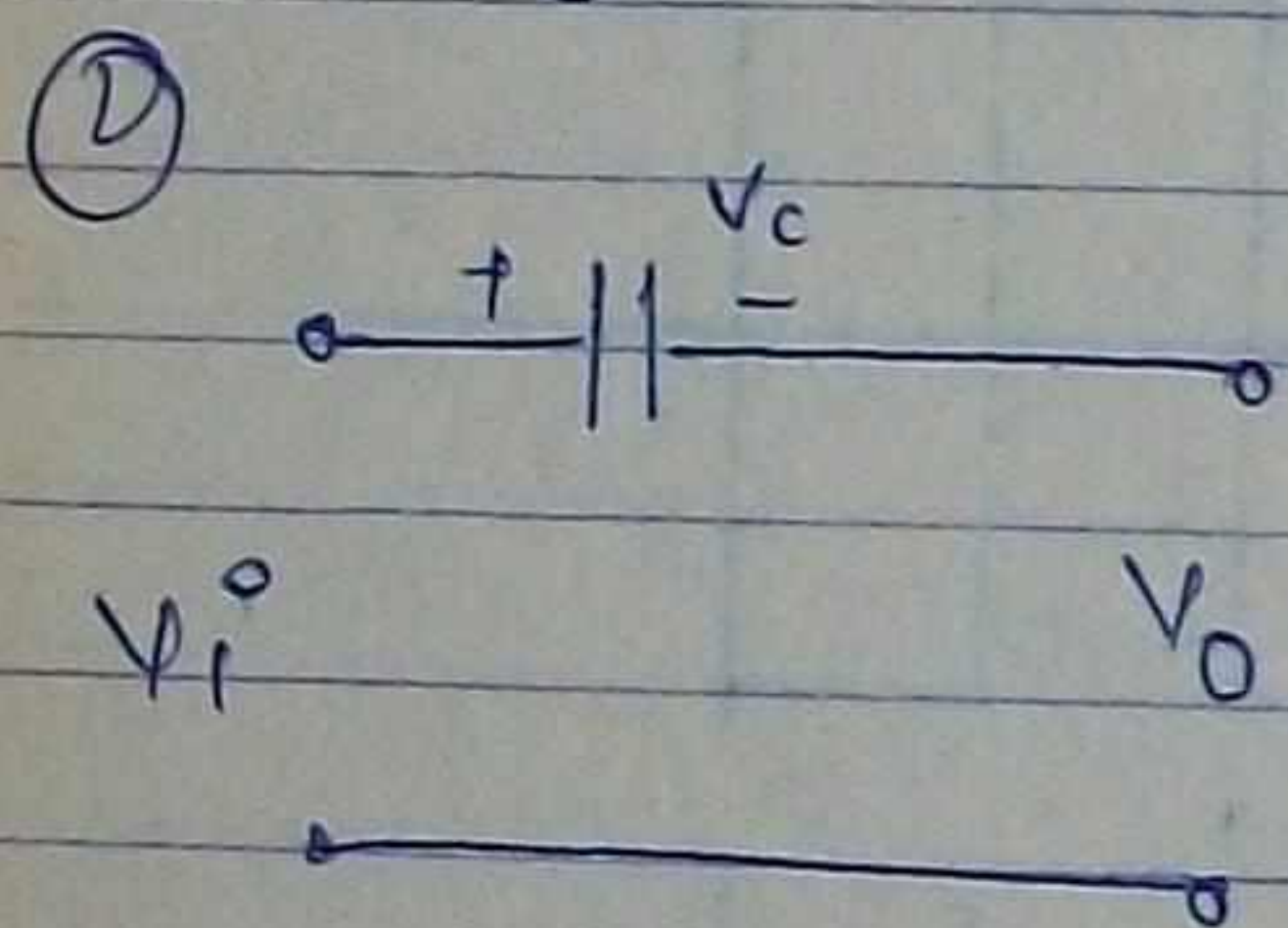


$$V_o = V_i - V_c$$

$$V_o = V_i - 8$$



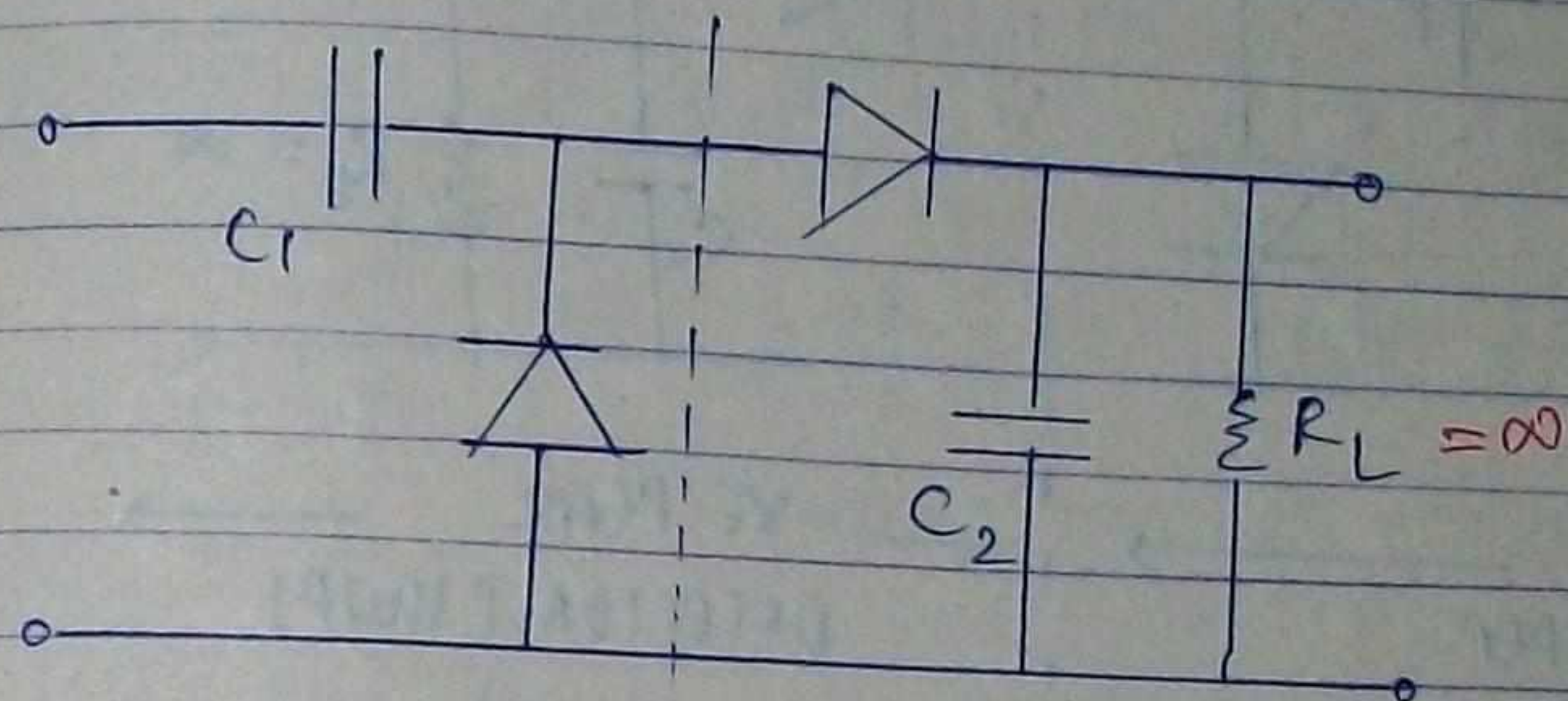
$$V_c = -10 - 2 = -12V$$



$$V_o = V_i - V_c = V_i + 12$$

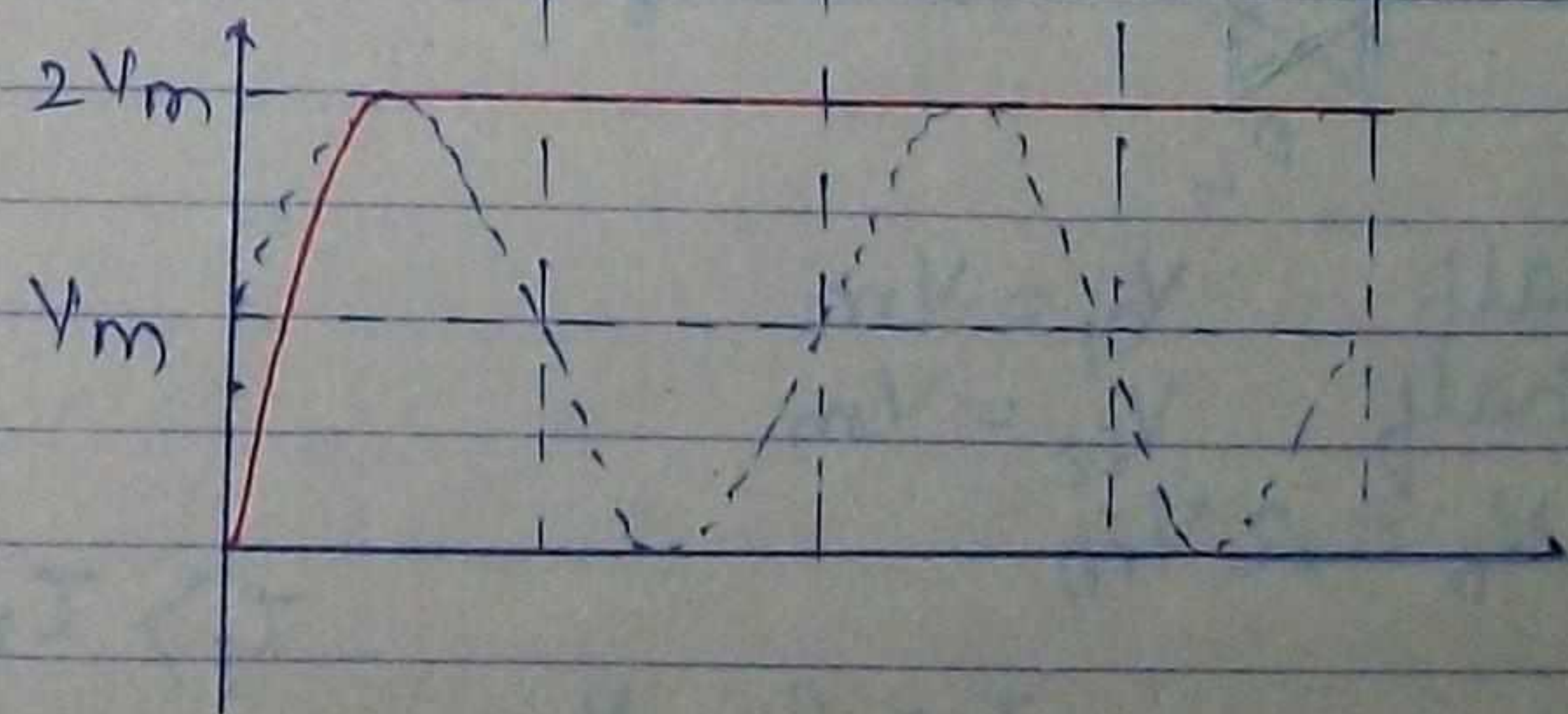
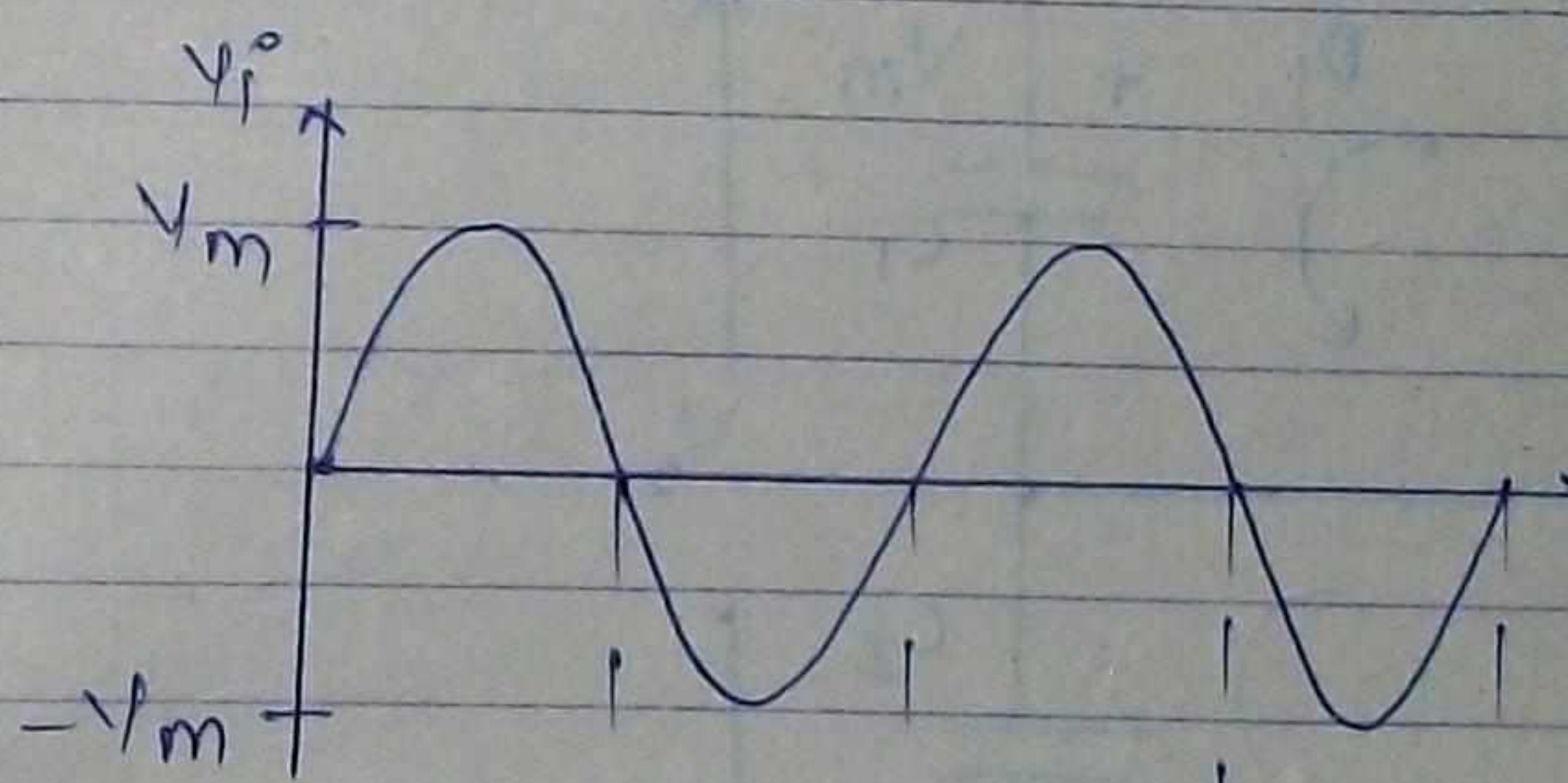
the clamping & O/P is clamped at $\pm 2V$

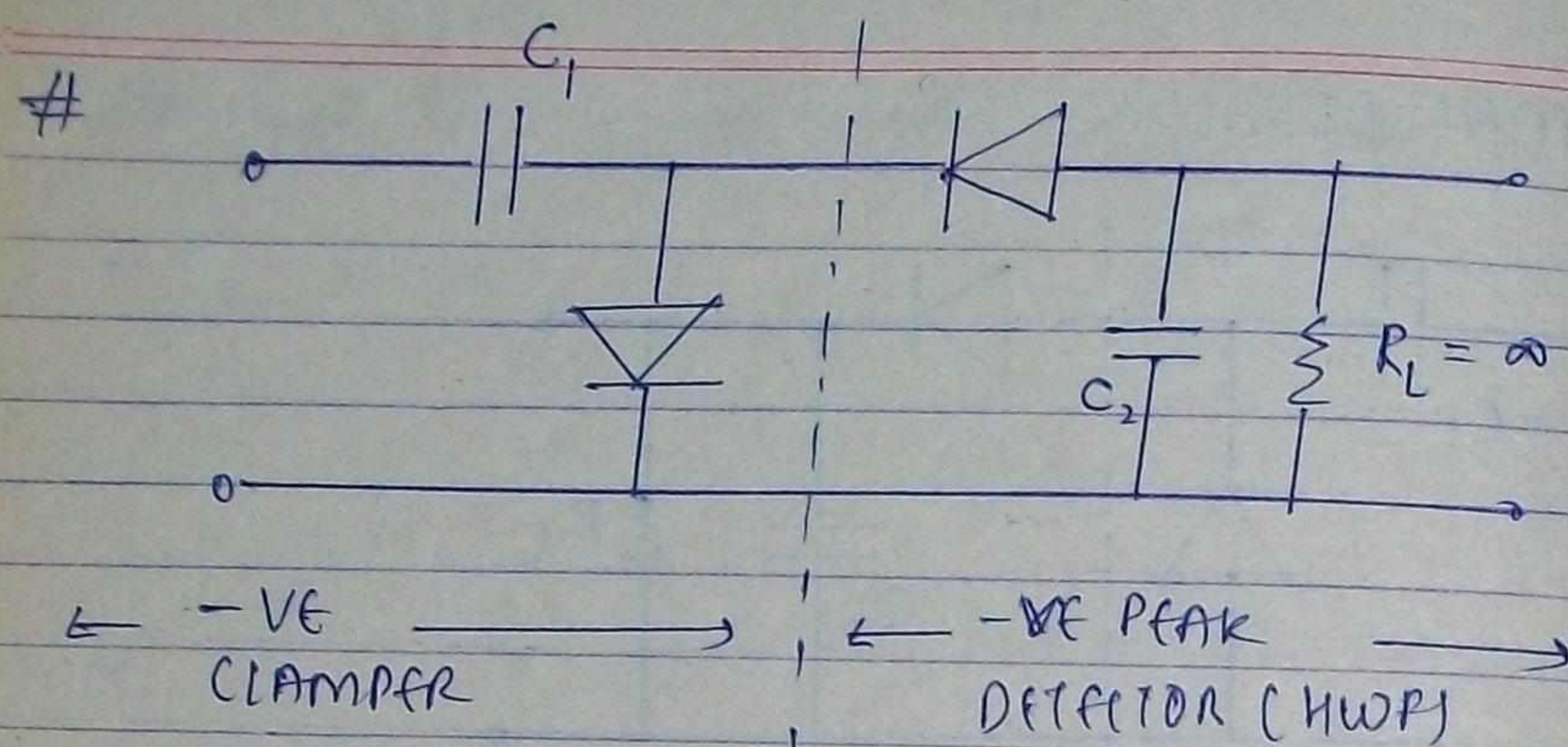
⇒ VOLTAGE DOUBLER -



← POSITIVE CLAMPER → ← +VE PEAK DETECTOR (HWP) →

← VOLTAGE DOUBLER → $V_o = 2V_m$

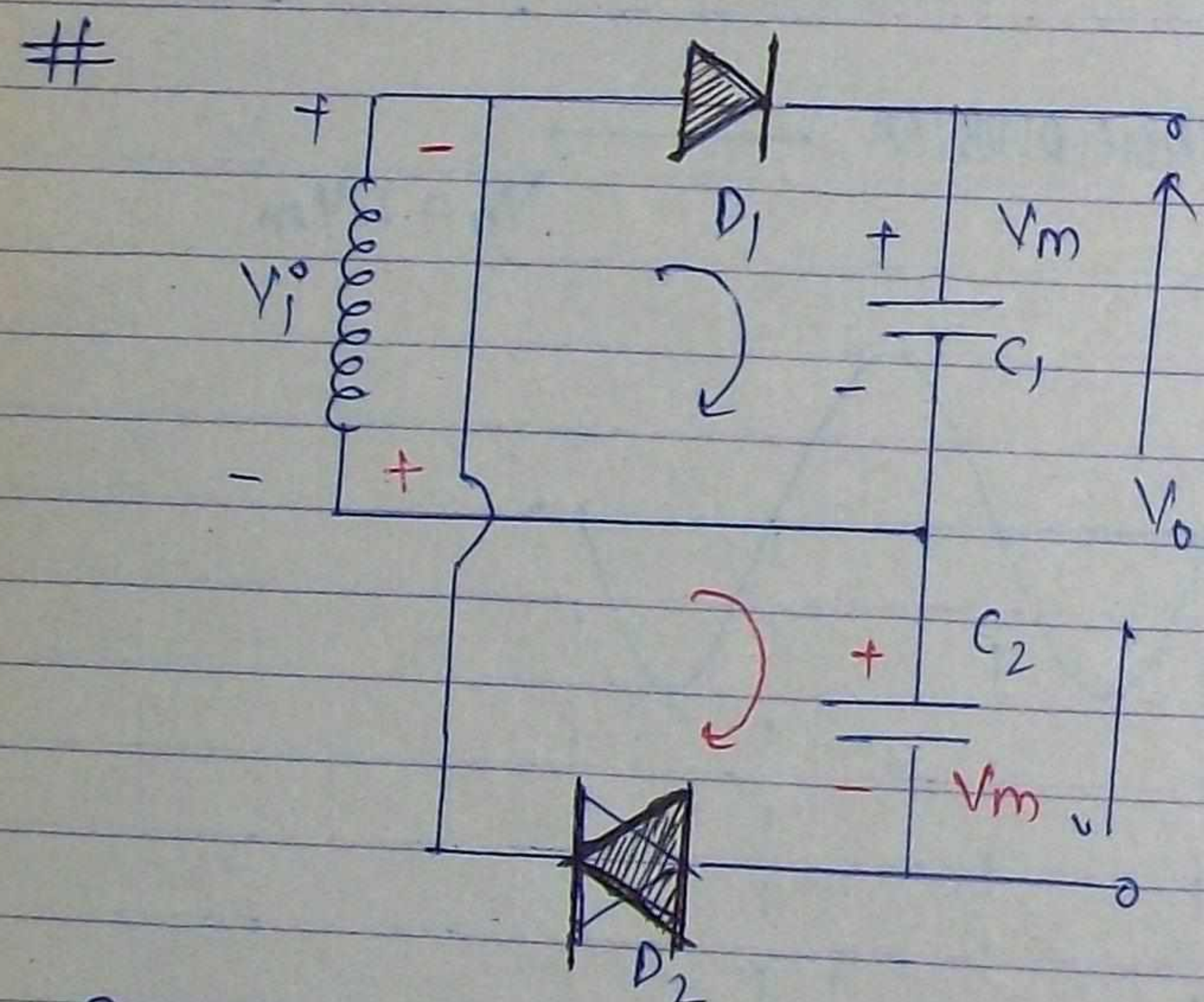




Voltage doubler, $V_o = -2V_m$

$$\tau = R_L C_2$$

↑
better



During +ve half $V_{C1} = V_m$
 During -ve half $V_{C2} = V_m$
 $V_o = 2V_m$

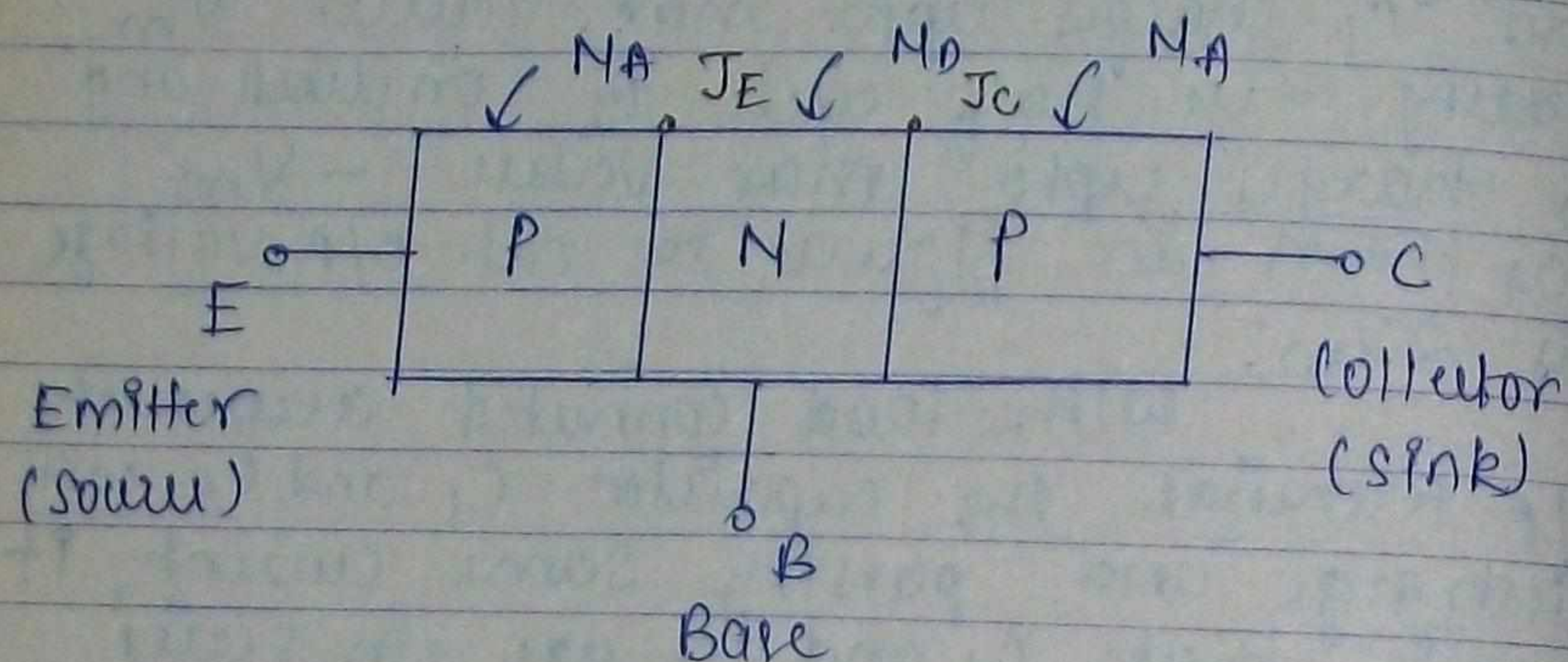
$$C_{eq} = \frac{C_1 C_2}{C_1 + C_2} < C_2 \quad \tau = C_{eq} R_L \quad \tau_1 > \tau_2$$

- During +ve half cycle the diode D_1 conduct and C_1 charges upto max. value V_m .
 During -ve half cycle D_2 conduct and C_2 charges upto max value $-V_m$ as shown in figure so net o/p voltage is $2V_m$.

with load connected across the o/p terminal the capacitor C_1 and C_2 will discharge and passing same current, it implies that C_1 and C_2 are in series with equivalent value $C_1 C_2 / (C_1 + C_2) \leq C_1$ and C_2 . This provides poor filtering action as compared to earlier circuit.

BJT AND TRANSISTORS

- Transfer of resistor.



AREA -

$$C > E < B$$

DOPING CONC -

$$E > C > B$$

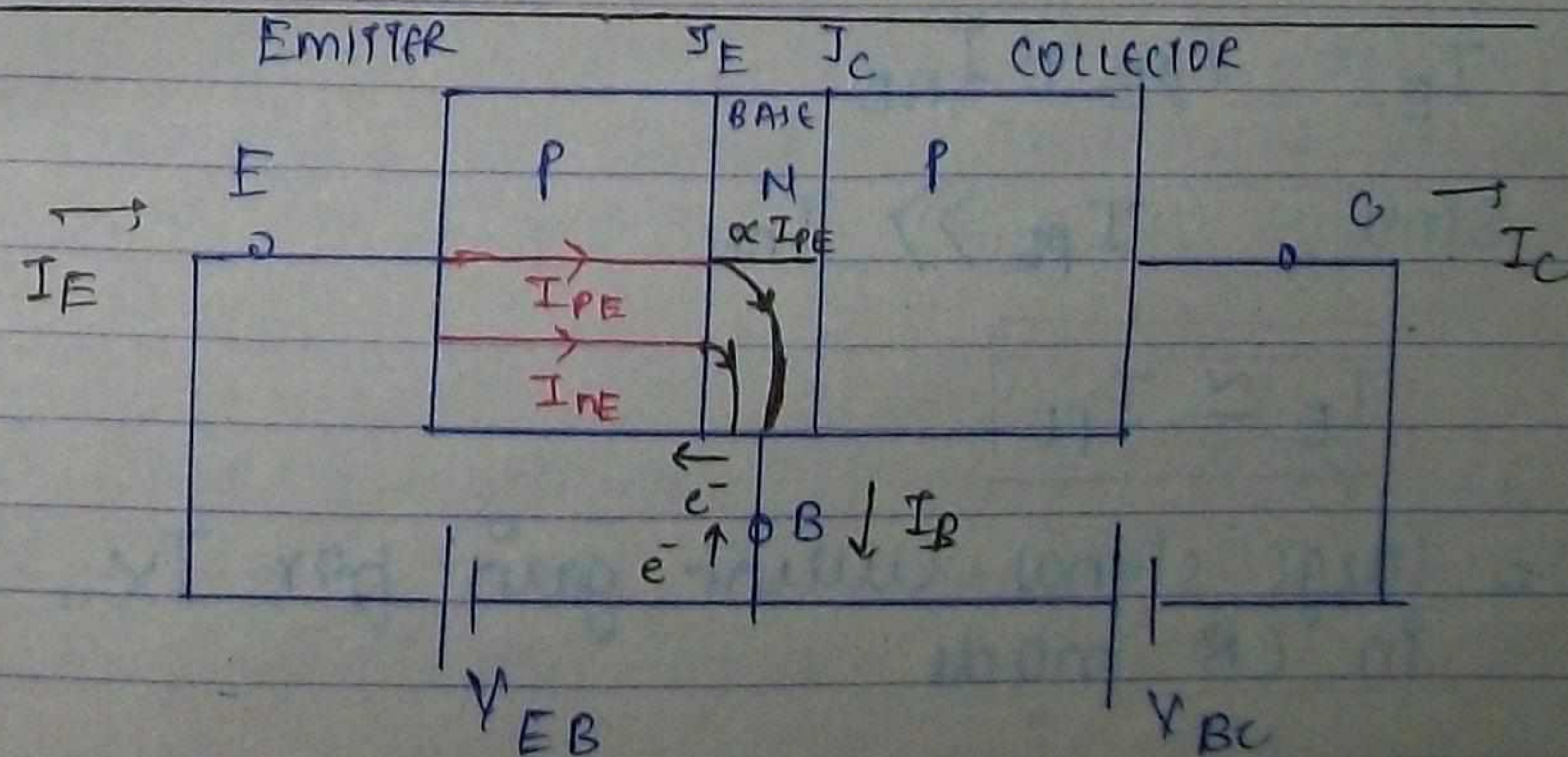
- A high density doping has been used for emitter, so that it is capable of emitting / injecting large number of holes (for pnp) and electrons (for npn) through base into collector, when emitter junction is F.B.
- The surface area of collector is usually more than that of emitter so that collector can handle more power.
- Base region is lightly doped, thin layer

designed to provide electrical isolation between emitter and collector. It also provides a contact point needed for control of emitter-base voltage.

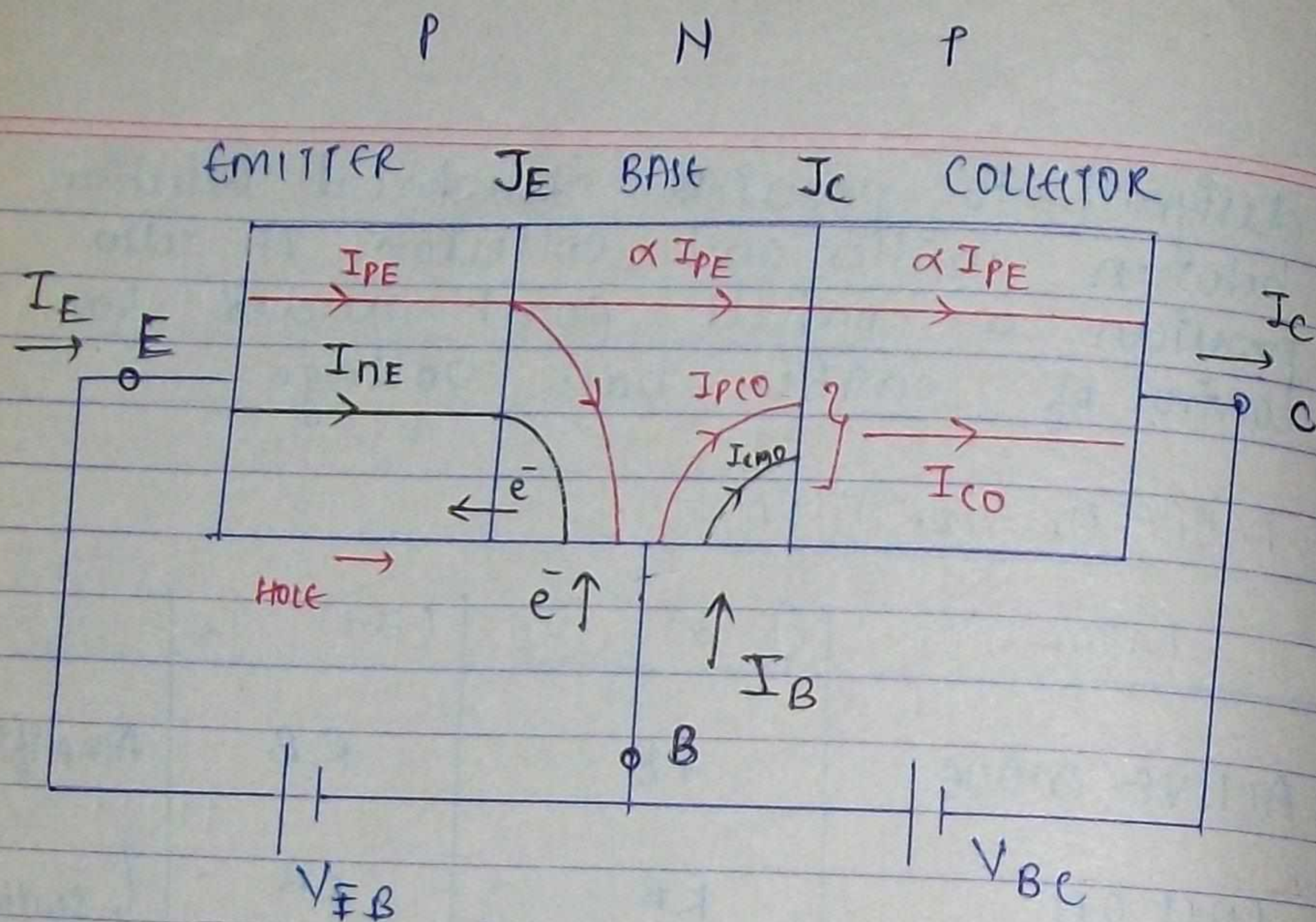
⇒ MODE OF OPERATION -

MODE	(E-B) J_E	(B-C) J_C	VSE
F.B. ACTIVE MODE ↓ NORMAL	FB	RB	Amplifier
CUT OFF	RB	RB	} Switching action
SATURATION	F.B.	FB	
REVERSE ACTIVE MODE ↓ INVERSE	RB	FB	Rarely used as an amp.

⇒ OPERATION OF TRANSISTOR IN ACTIVE MODE -



All current is due to diffusion



under open circuit condition -

$$I_T = I_{\text{DIFFUSION}} + I_{\text{DRIFT}} = 0$$

$$I_{\text{DIFFUSION}} = -I_{\text{DRIFT}}$$

$$I_E = I_{PE} + I_{NE}$$

Since, $I_{PE} \gg I_{NE}$

$$I_E \approx I_{PE}$$

α = large signal current gain for Tx in CB mode

$$\alpha < 1 \quad (0.9 \text{ to } 0.995)$$

$$I_{CO} = I_{PCO} + I_{NCO}$$

I_{CO} = Reverse saturation current

$$I_C = \alpha I_{PE} + I_{CO}$$

$$I_C = \alpha I_E + I_{CO}$$

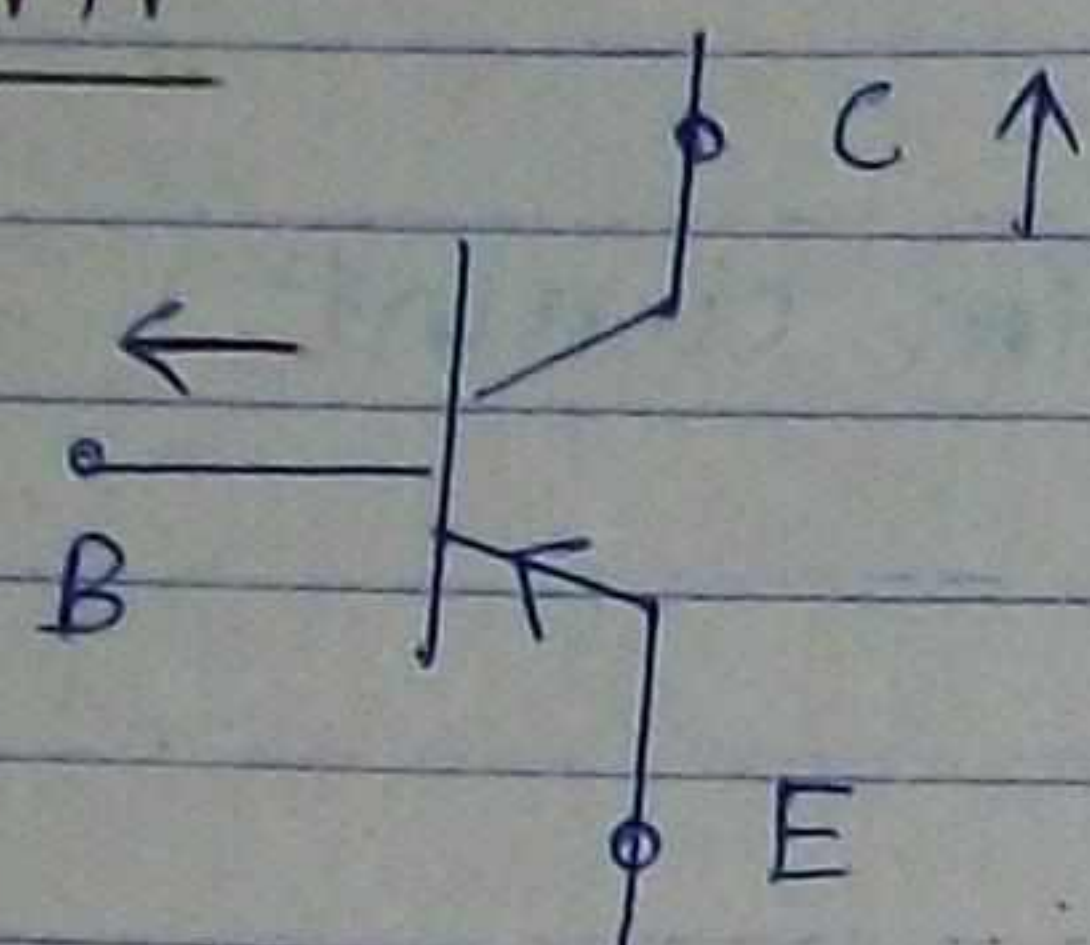
BASIC ASSUMPTION -

- Externally applied voltage V_{EB} & V_{BC} appear across the depletion region in each junction (not shown in fig). Thus potential is constant with in each region. This means that Drift current due to thermally generated carriers have been neglected which in any case is usually very small. Thus all the current component in as shown in figure are due to diffusion current.

- α is not constant, it varies with I_E , V_{BE} and temp. T . Besides the type of Tx. (npn or pnp)

⇒ CIRCUIT SYMBOL -

① PNP -



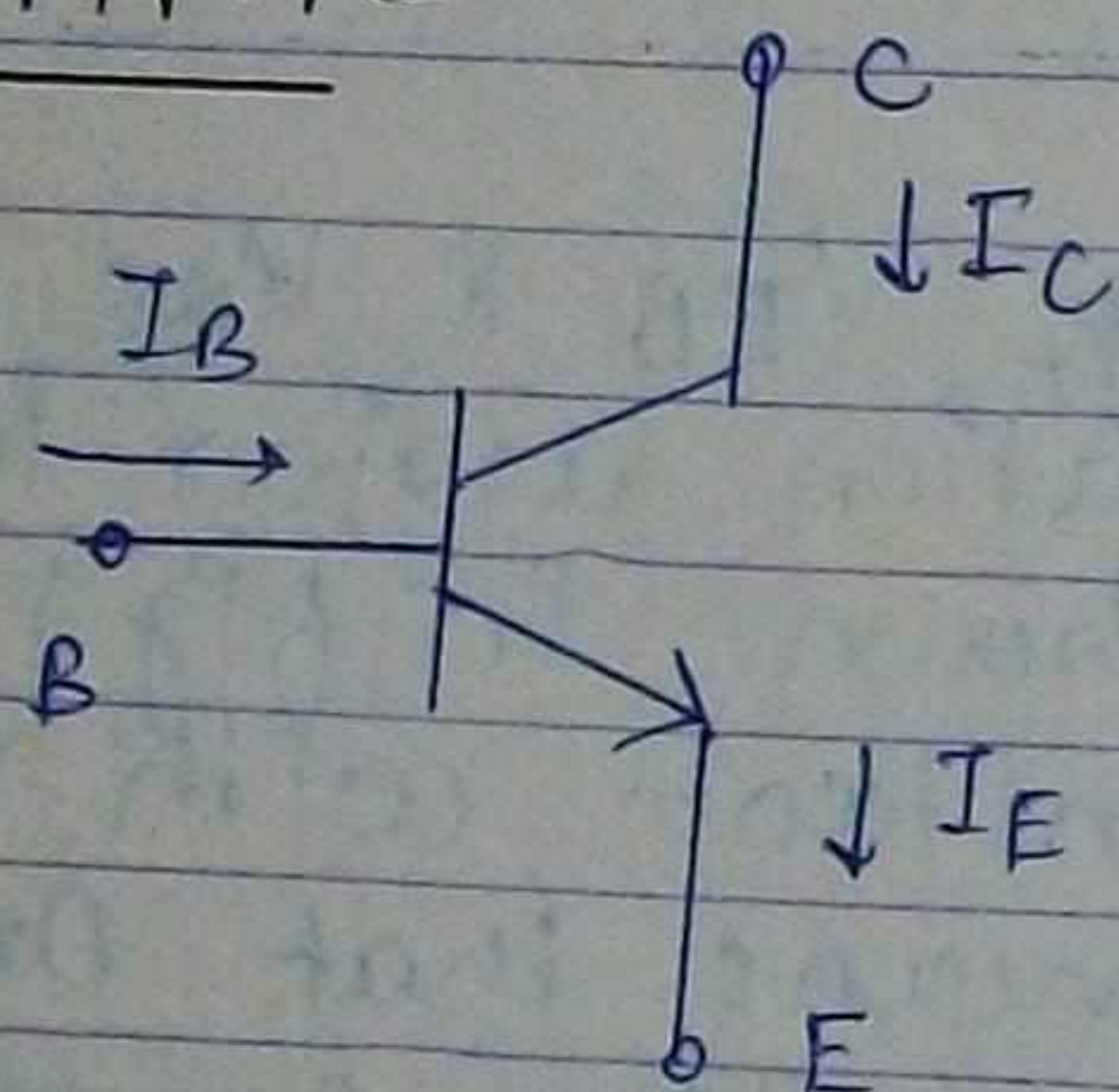
$$I_E = I_C + I_B$$

$$I_E = +ve$$

$$I_B = +ve$$

$$I_C = +ve$$

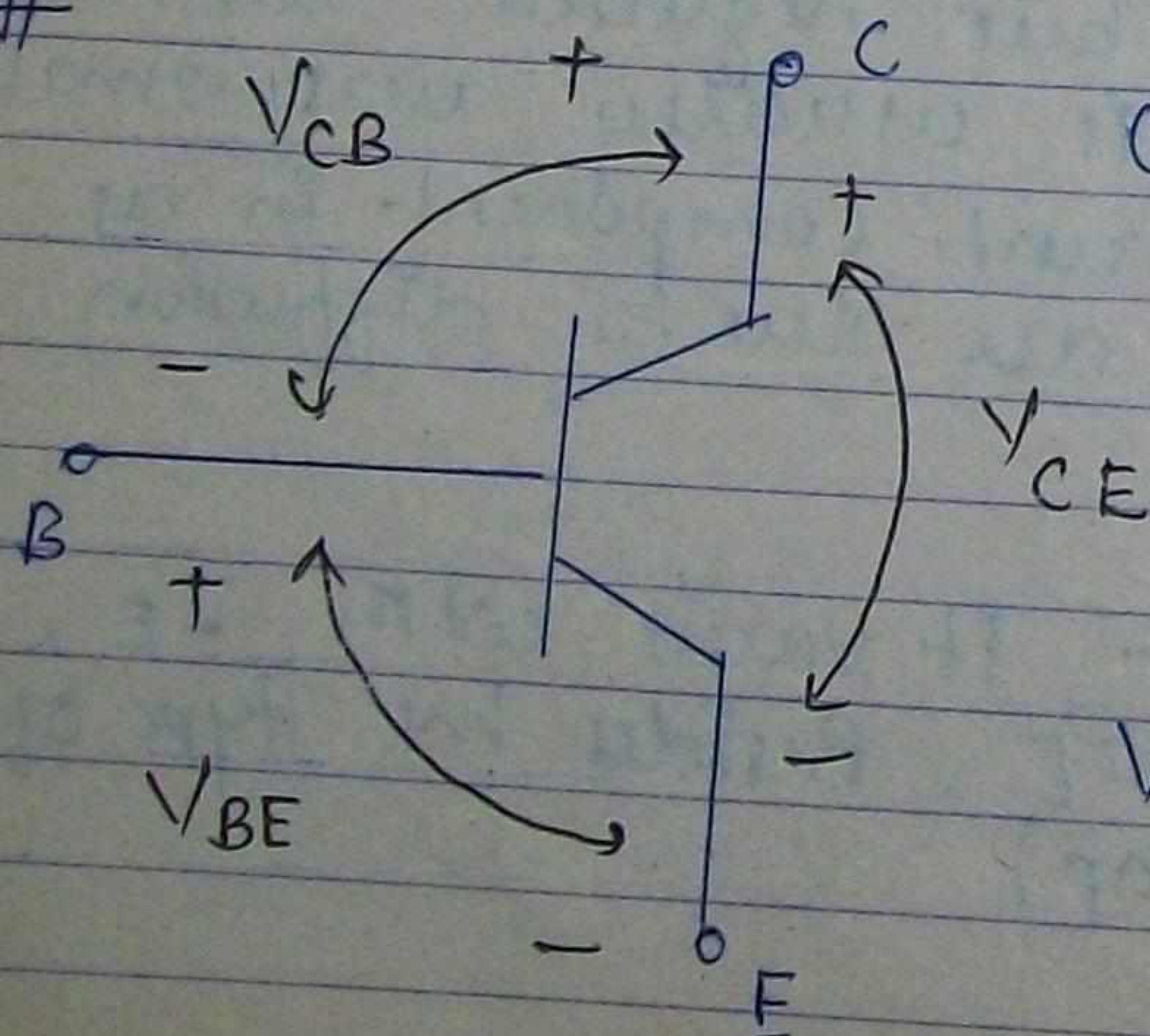
② NPN -



$$I_E = I_C + I_B$$

$$\left. \begin{matrix} I_E \\ I_B \\ I_C \end{matrix} \right\} \begin{matrix} \text{all} \\ +ve \end{matrix}$$

#



also,

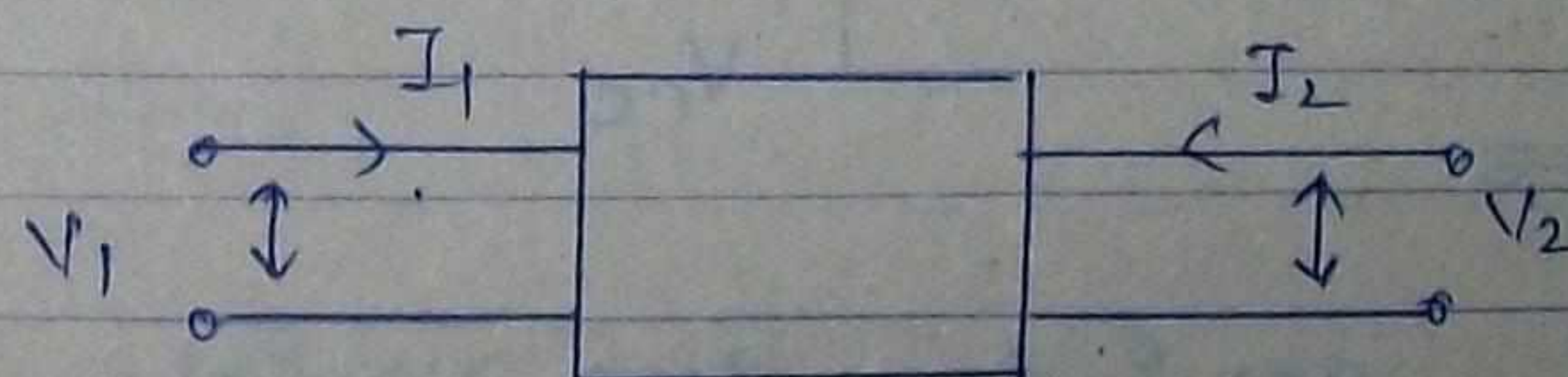
$$V_{CE} + V_{EB} + V_{BC} = 0$$

$$V_{CE} = V_{CB} + V_{BE}$$

⇒ CONFIGURATION -

	CURRENT BUFFER		VOLTAGE BUFFER
	CB	CE	CC
I/P	E	B	B
O/P	C	C	E
COMMON TO I/P & O/P	BASE	EMITTER	COLLECTOR
A_I	LOW ≈ 1	HIGH	V. HIGH
A_V	V. HIGH	HIGH	LOW ≈ 1
R_i	V. LOW	HIGH	V. HIGH
R_o	V. HIGH	HIGH	LOW

⇒ I/P AND O/P CHARACTERISTICS -



$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix}$$

↑
DEPENDENT VARIABLE

↑
INDEPENDENT VARIABLE

EMITTER FOLLOWER

$$V_1 = h_{11} I_1 + h_{12} V_2$$

I/P characteristics

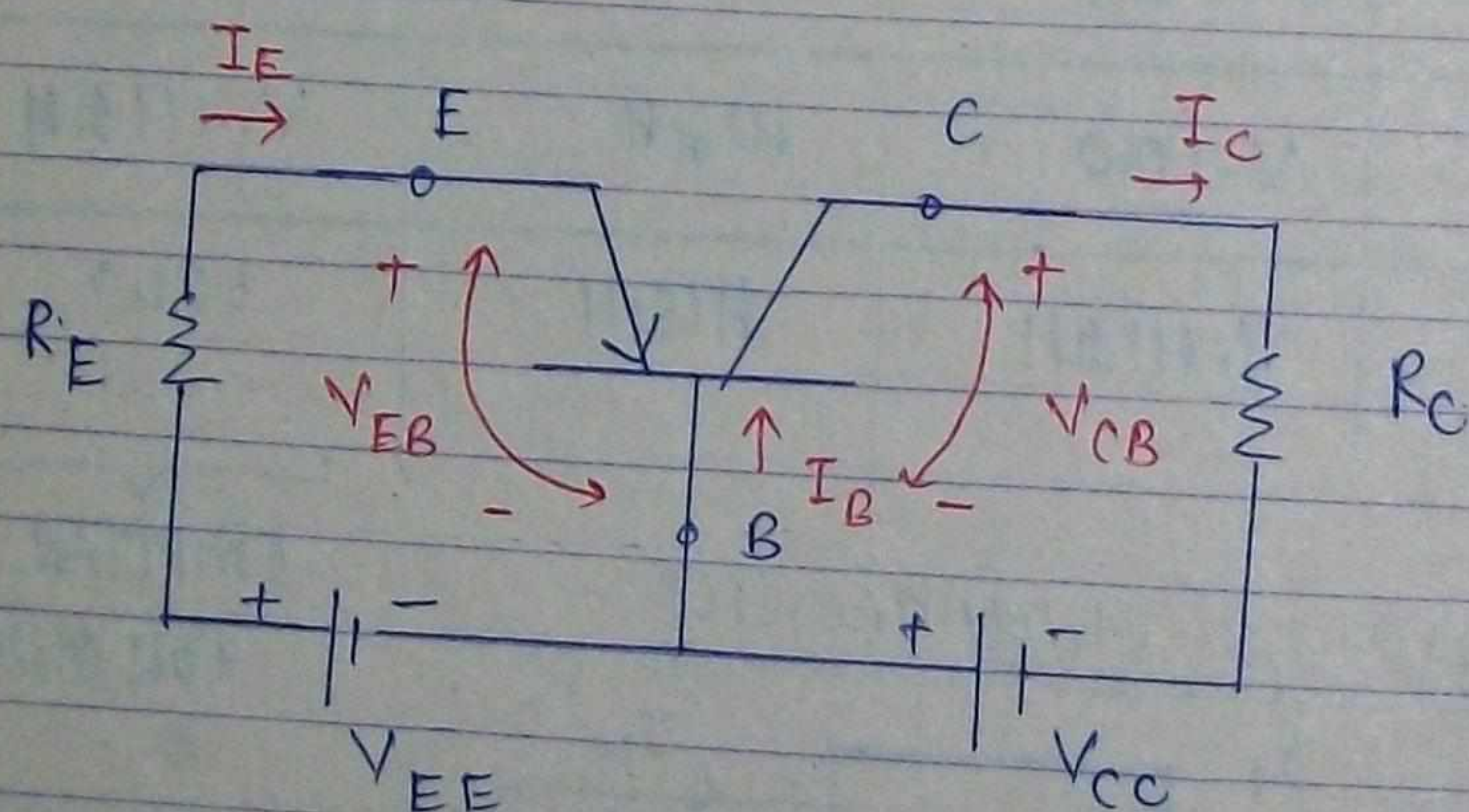
$$V_1 = f[I_1, V_2]$$

$$I_2 = h_{21} I_1 + h_{22} V_2$$

O/P characteristics

$$I_2 = f[I_1, V_2]$$

⇒ COMMON-BASE CONFIGURATION -



I_E and V_{CB} are independent variables

V_{EB} and I_C are dependent variables

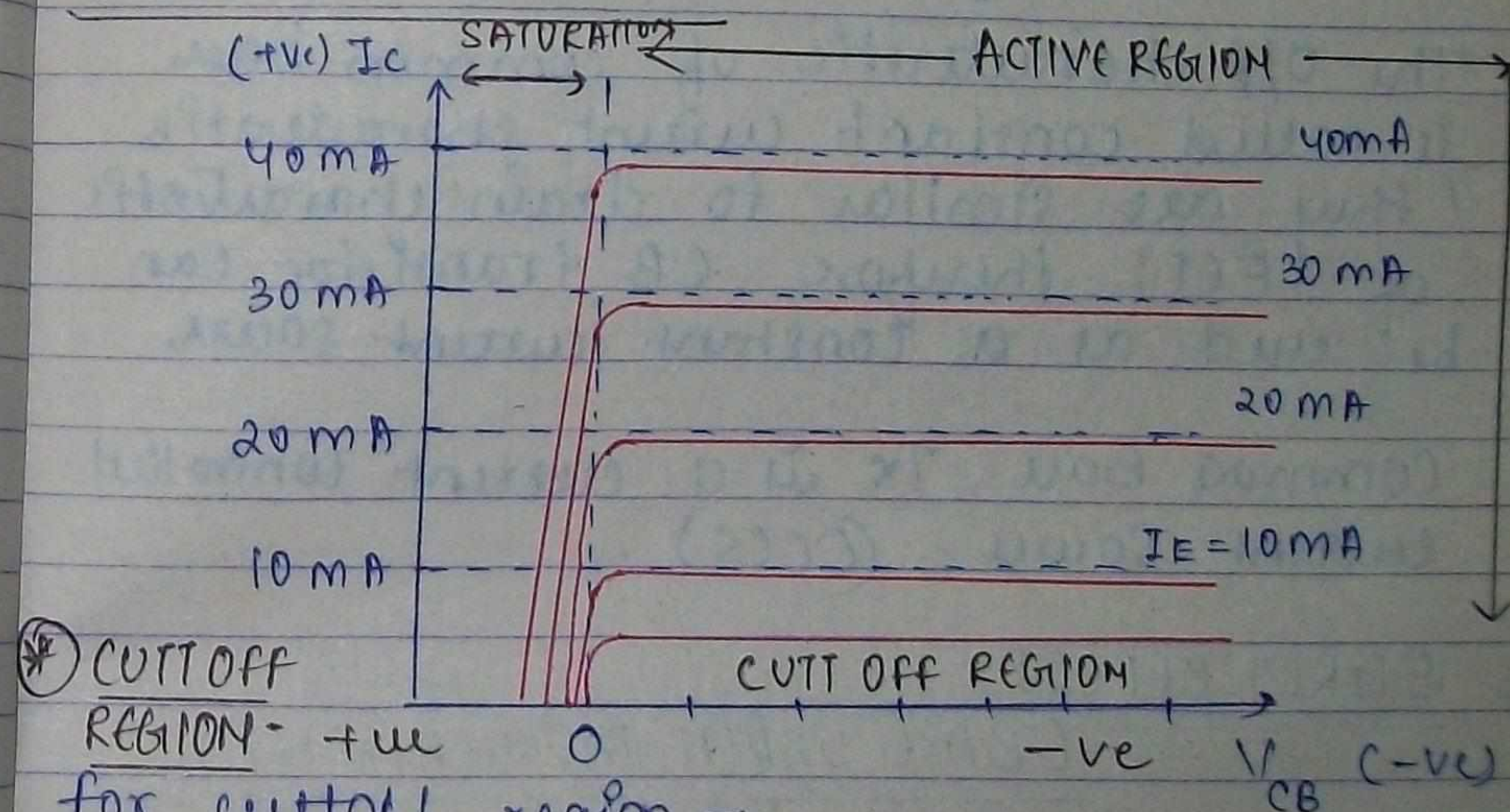
O/P characteristic -

$$I_C = f[I_E, V_{CB}]$$

I/P characteristics -

$$V_{EB} = f[I_E, V_{CB}]$$

OUTPUT CHARACTERISTICS -



* CUT OFF

REGION - +ve
for cutoff region -

$$I_C = \alpha I_E + I_{CO}$$

$$\boxed{I_E = 0} \quad \text{and} \quad \boxed{I_C = I_{CO}}$$

* ACTIVE REGION -

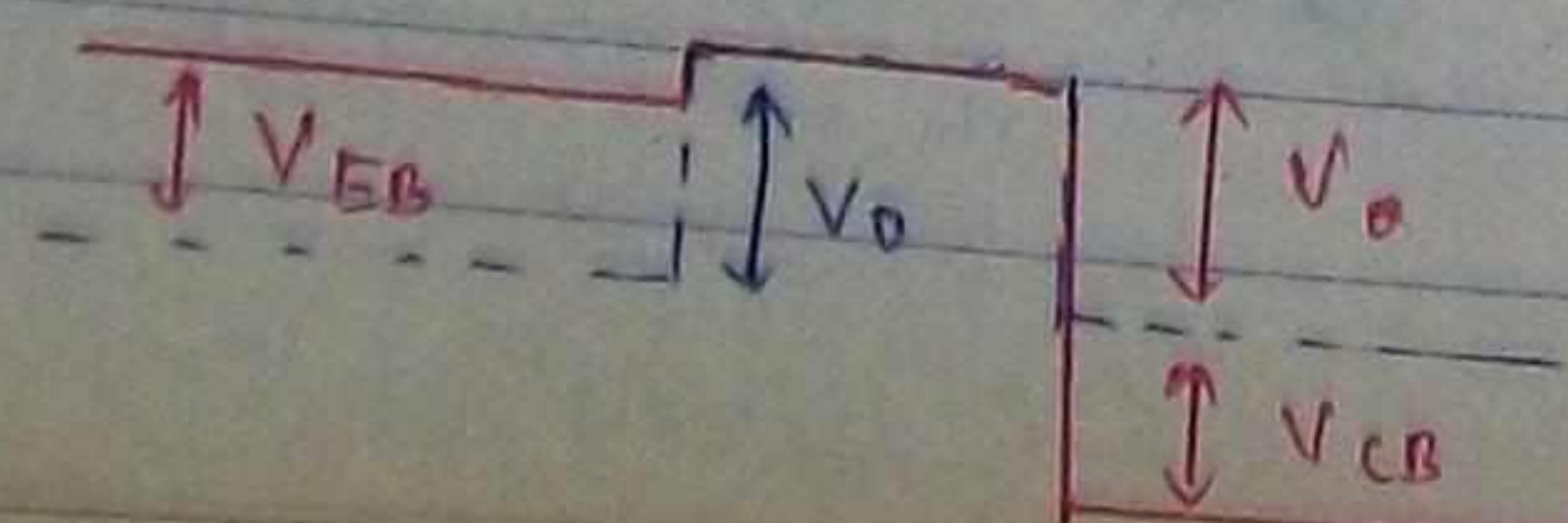
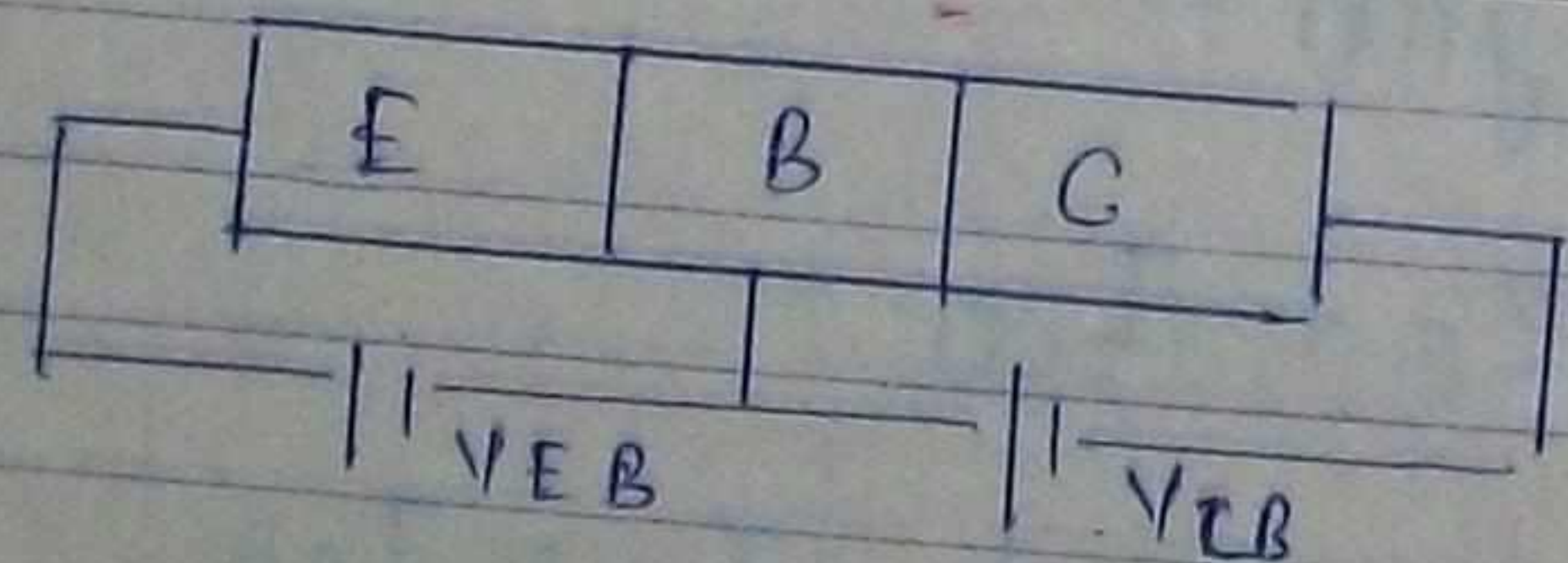
for active region -

$$I_C = \alpha I_E + I_{CO}$$

- I_C is less than I_E as α is always less than 1 ($\alpha < 1$)
- I_C is independent of o/p voltage V_{CB} (by neglecting early effect).
- The o/p characteristic of common base is called constant current characteristic (they are similar to drain characteristics of FET) therefore CB transistor can be used as a constant current source.
- Common Base Tx is a current controlled current source (CCCS).

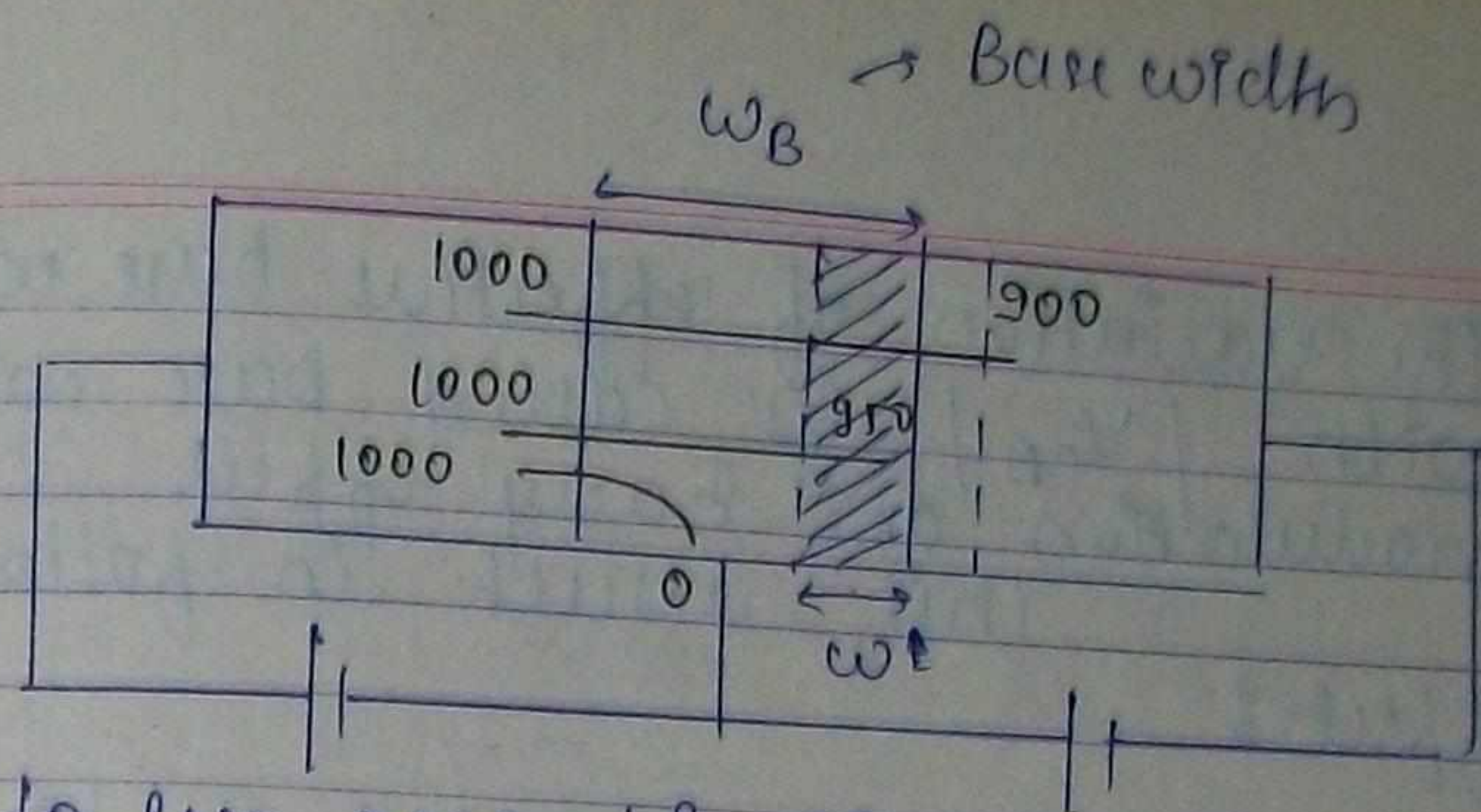
⇒ EARLY EFFECT -

(BASE WIDTH MODULATION)



OPEN CIRCUIT

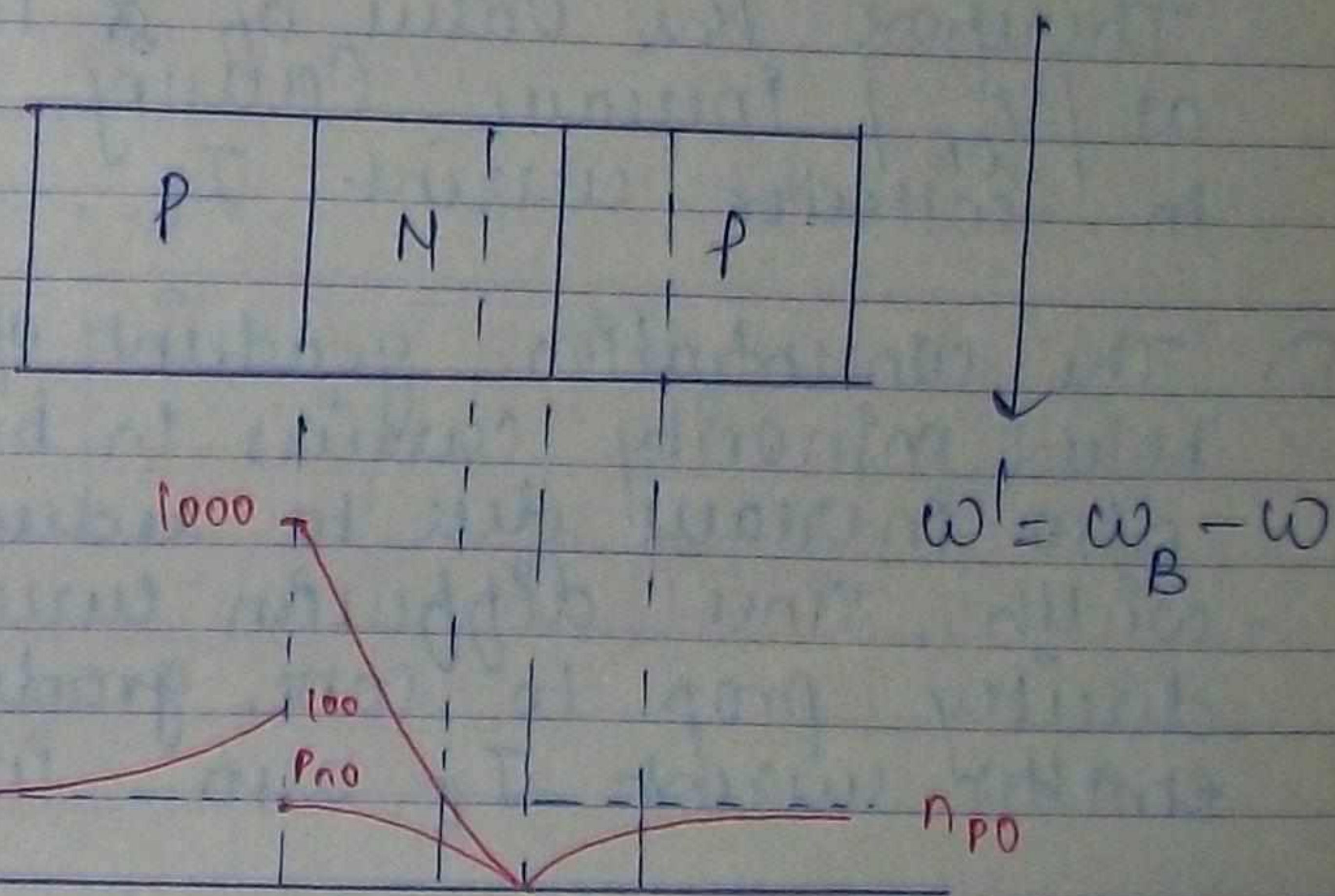
ACTIVE REGION



Due to carrier recombination

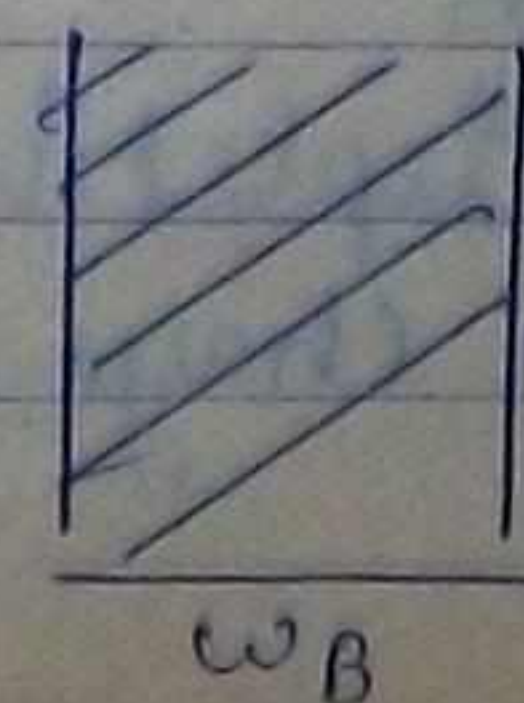
$$\alpha \uparrow$$

$$w' = 0 \text{ (effective base width)}$$



$$I_E = q D_p \frac{dp}{dx}$$

$$\frac{dp}{dx} = \frac{1000 - 0}{w_B} \quad \left| \quad \frac{dp'}{dx} = \frac{1000 - 0}{w'} > \frac{dp}{dx} \right.$$



Punch through or Reach through

- The variation of effective base width with $|V_{CB}|$ is called base width modulation or Early effect. This results in following effect:

① There is less change of recombination in base region as effective base width reduces from w_B to $w_B - w$. Therefore the value of α increases as $|V_{CB}|$ increases. Causing an increase in collector current I_C .

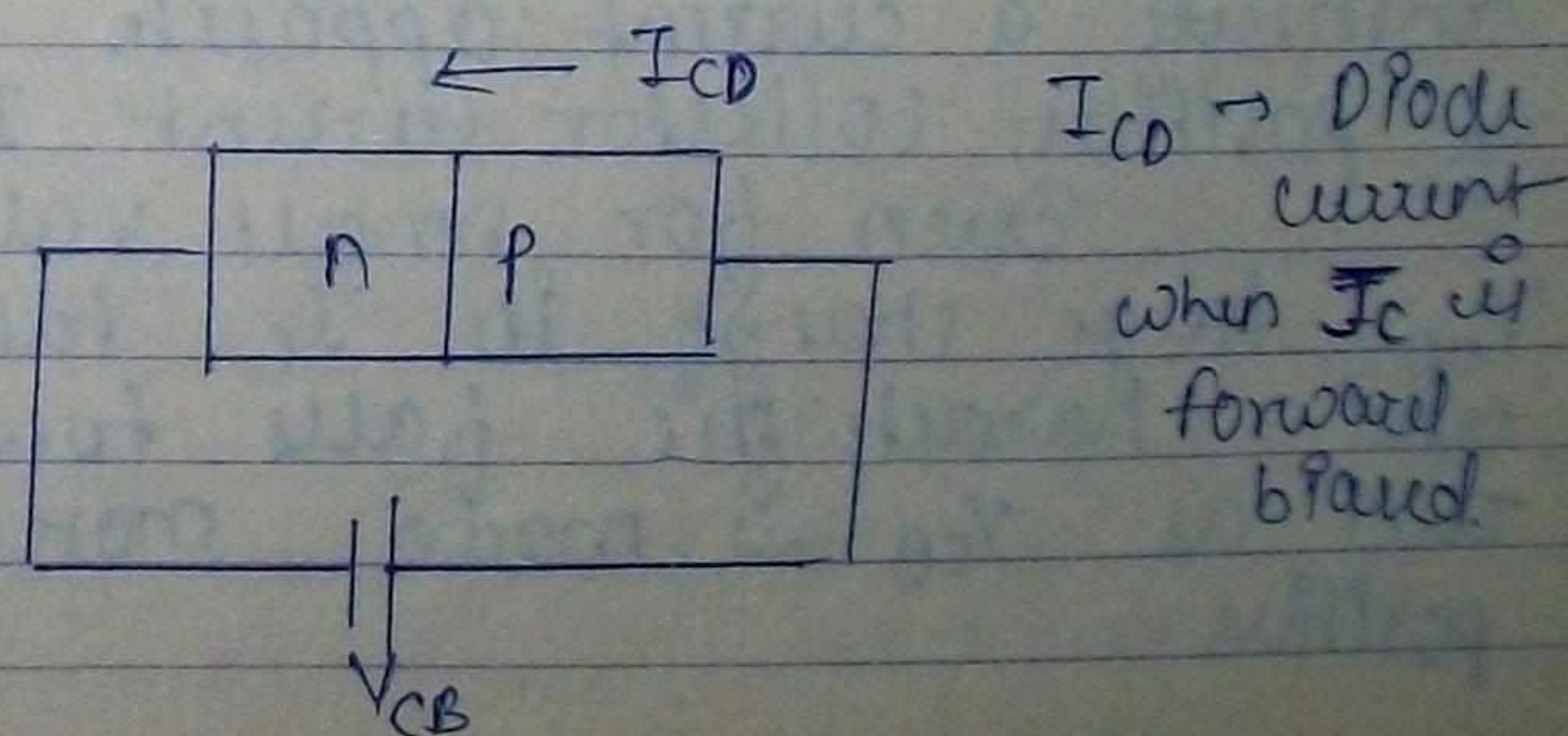
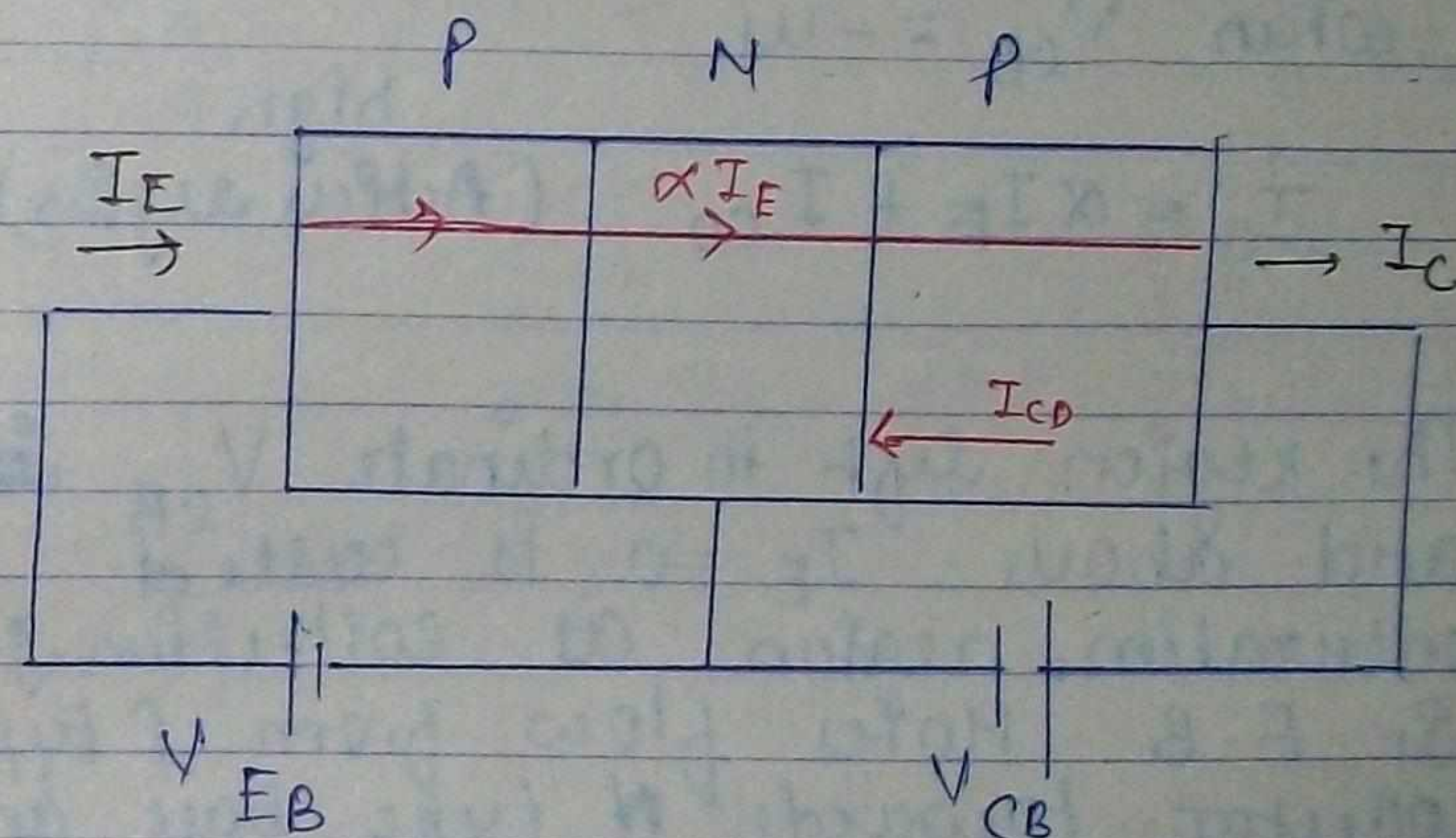
② The concentration gradient of injected holes (minority carriers in base region) also increases due to reduce base-width. Since diffusion current is directly prop. to conc. gradient, emitter current I_E also increases.

③ For large values of $|V_{CB}|$ effective base width may be reduced to zero causing extremely large emitter current. This results in breakdown of transistor and is called punch through or Reach through. Note that the depletion regions have ions (immobile charge carriers)

and hence conduction, the voltage V_{EB} and $|V_{CB}|$ have opposite polarity and therefore short circuit via conduction base when V_{CB} is very large. This causes a very high break-through current.

$$|V_{CB}| \uparrow, \alpha \uparrow, I_E \uparrow, I_C \uparrow$$

⊛ SATURATION REGION -



$$I_C = \alpha I_E - I_{CO}$$

$$I_C = \alpha I_E - I_{CO} [e^{V_{CB}/nV_T} - 1]$$

Assume $\eta = 1$ (for Ge & Si)

$$I_C = \alpha I_E - I_{CO} [e^{V_{EB}/V_T} - 1]$$

(Saturation) region.

$I_{CO} \rightarrow$ diode current

when $V_{CB} = -u$

$$I_C = \alpha I_E + I_{CO} \quad (\text{Active region})$$

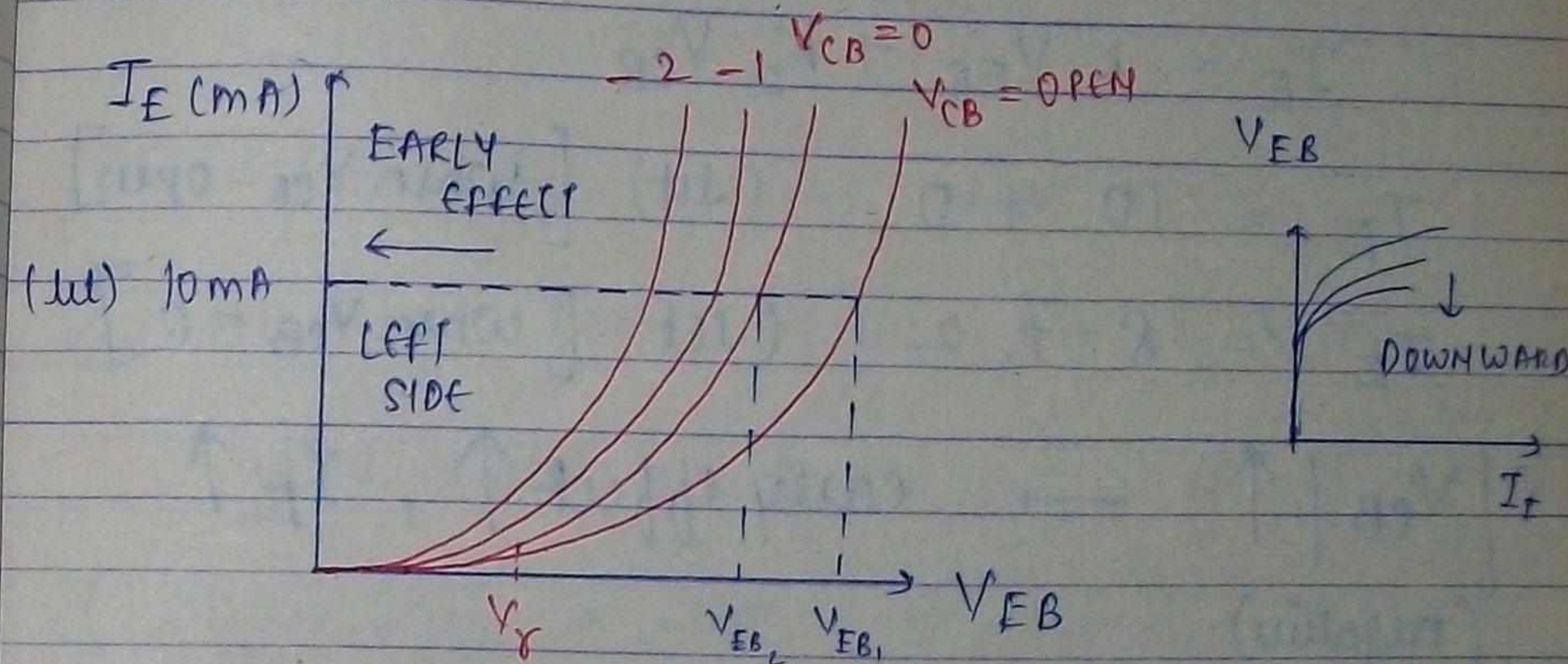
- The region left to ordinate $V_{CB} = 0$ and above $I_E = 0$ is called saturation region as collection junction is F-B. Holes flow from P type collector towards N type base and constitute a current opposite to direction of positive collector current I_C .

Even for small value of V_{CB} large change in I_C take place and characteristic falls towards zero as V_{CB} is made more and more positive.

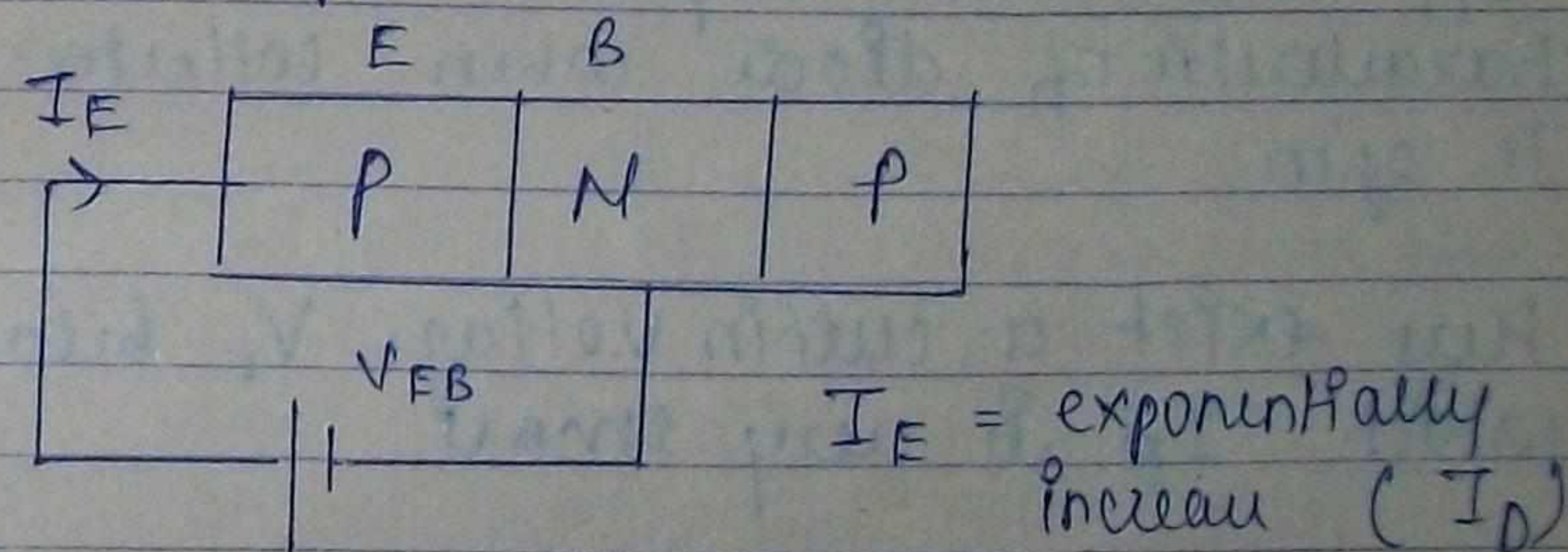
- Since I_{CO} increases exponentially, I_C may even become negative

INPUT CHARACTERISTICS -

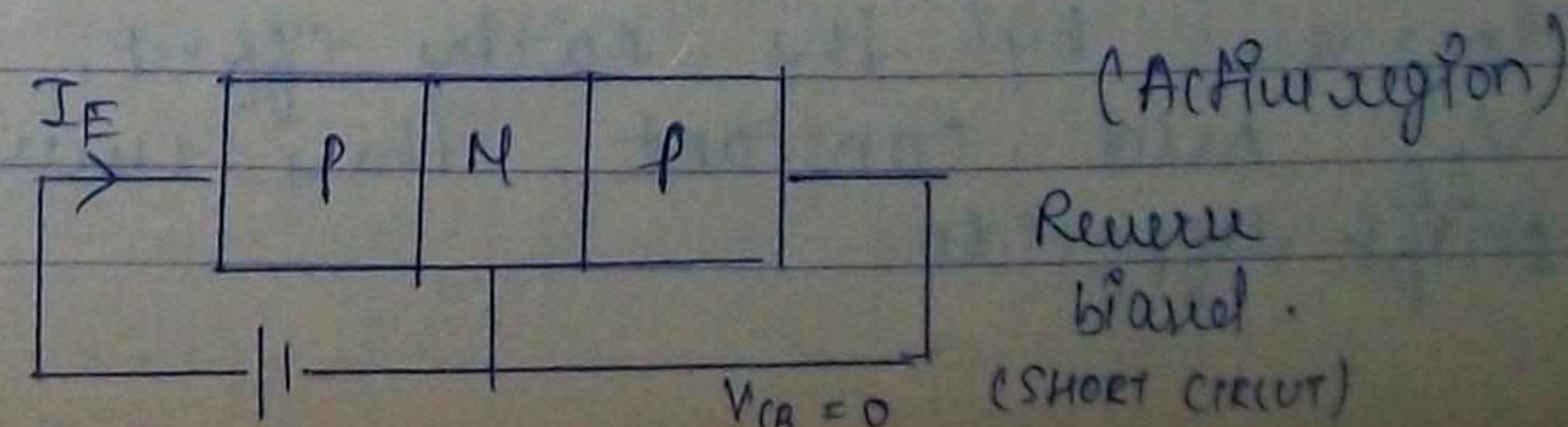
$$V_{EB} = f[I_E, V_{CB}]$$



when $I_C = \text{Open}$



when $I_C = 0$ or $V_{CB} = 0$



$V_{CB} = 0$ Bandwidth change $\uparrow$

$\downarrow$
Early effect $\uparrow$

$\downarrow$
 $\alpha \uparrow$ and $I_E \uparrow$

$$I_E = K_1 V_{EB} + K_2 V_{CB}$$

$$I_E = 10 + 0 \quad (\text{set}) \quad [\text{when } V_{CB} = \text{open}]$$

$$I_E = 8 + 2 \quad (\text{set}) \quad [\text{when } V_{CB} = 0]$$

$|V_{CB}| \uparrow \rightarrow \text{early effect } \uparrow, I_E \uparrow$
(negative)

- I/P characteristic represent forward characteristic of diode when collector is open.

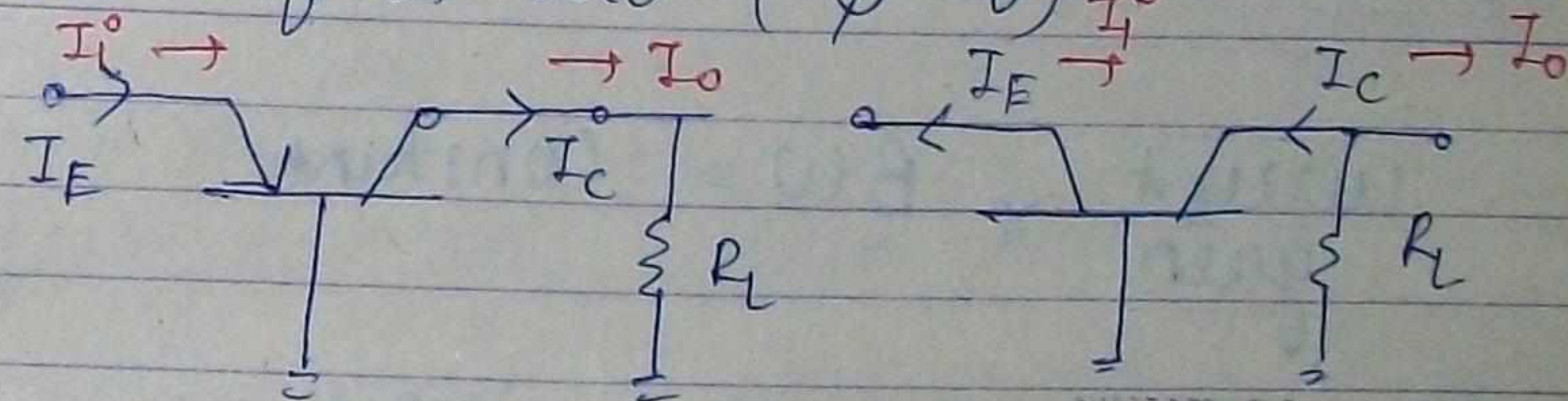
- Here exist a cutoff voltage V_r below which I_E is very small.

- Increase in $|V_{CB}|$ will cause I_E to increase by the early effect, If V_{EB} held constant. This curve shift left side.

PROPERTIES OF COMMON BASE -

- ① Lowest R_i° (less than 100Ω)
- ② Highest R_o (greater than $1M\Omega$)
- ③ Lowest $A_i^\circ = \frac{|I_c|}{|I_E|} = \alpha \leq 1$
- ④ Highest A_v (voltage gain)
- ⑤ moderate power gain ($A_p = A_v \times A_i^\circ$)
(typical value = 68)

- ⑥ Phase shift is zero ($\phi = 0$)



$$A_i^\circ = \frac{I_o}{I_i} = \frac{I_c}{I_E}$$

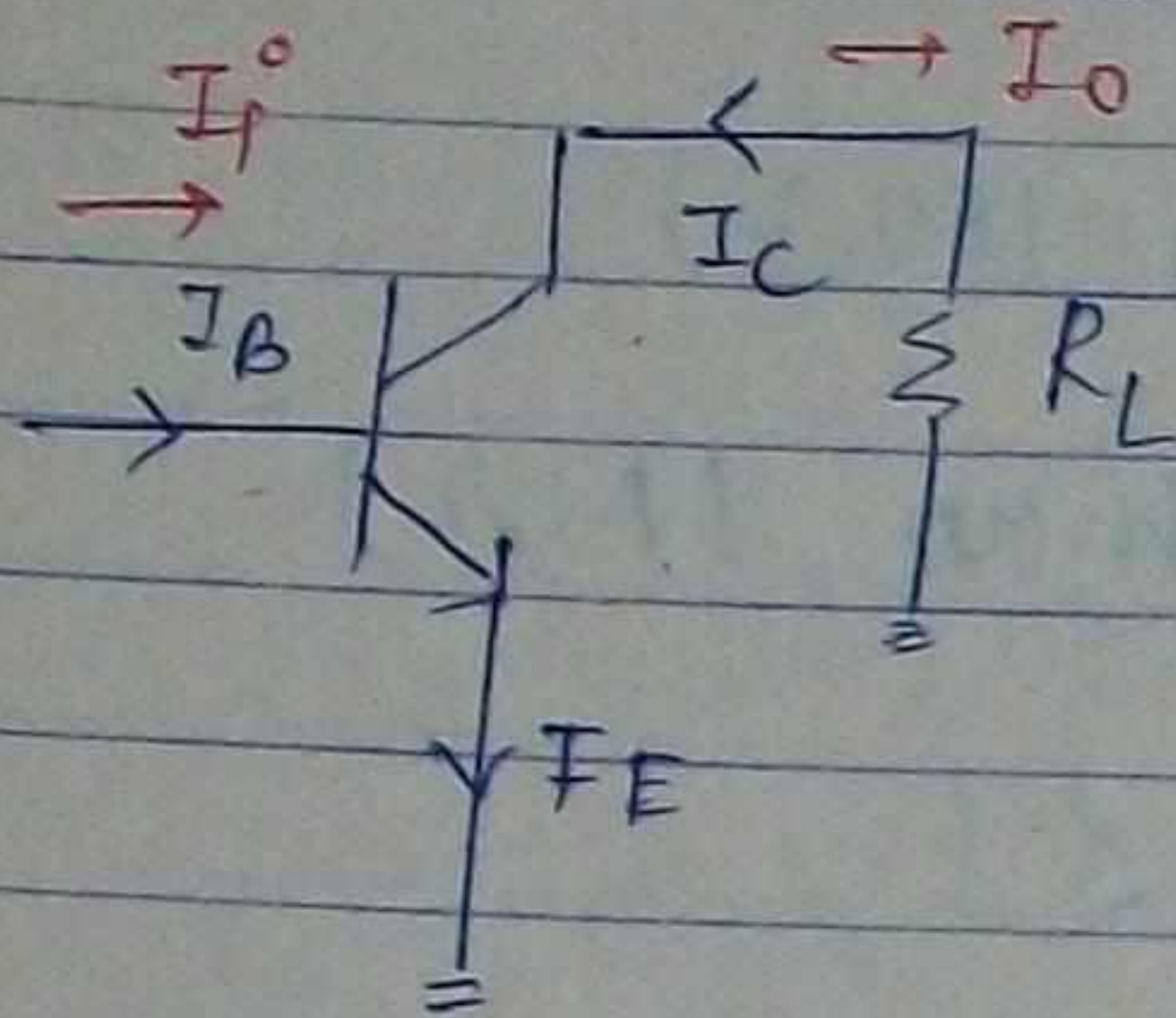
$\phi = 0$

$$A_i^\circ = \frac{I_o}{I_i} = \frac{-I_c}{-I_E}$$

$\phi = 0$

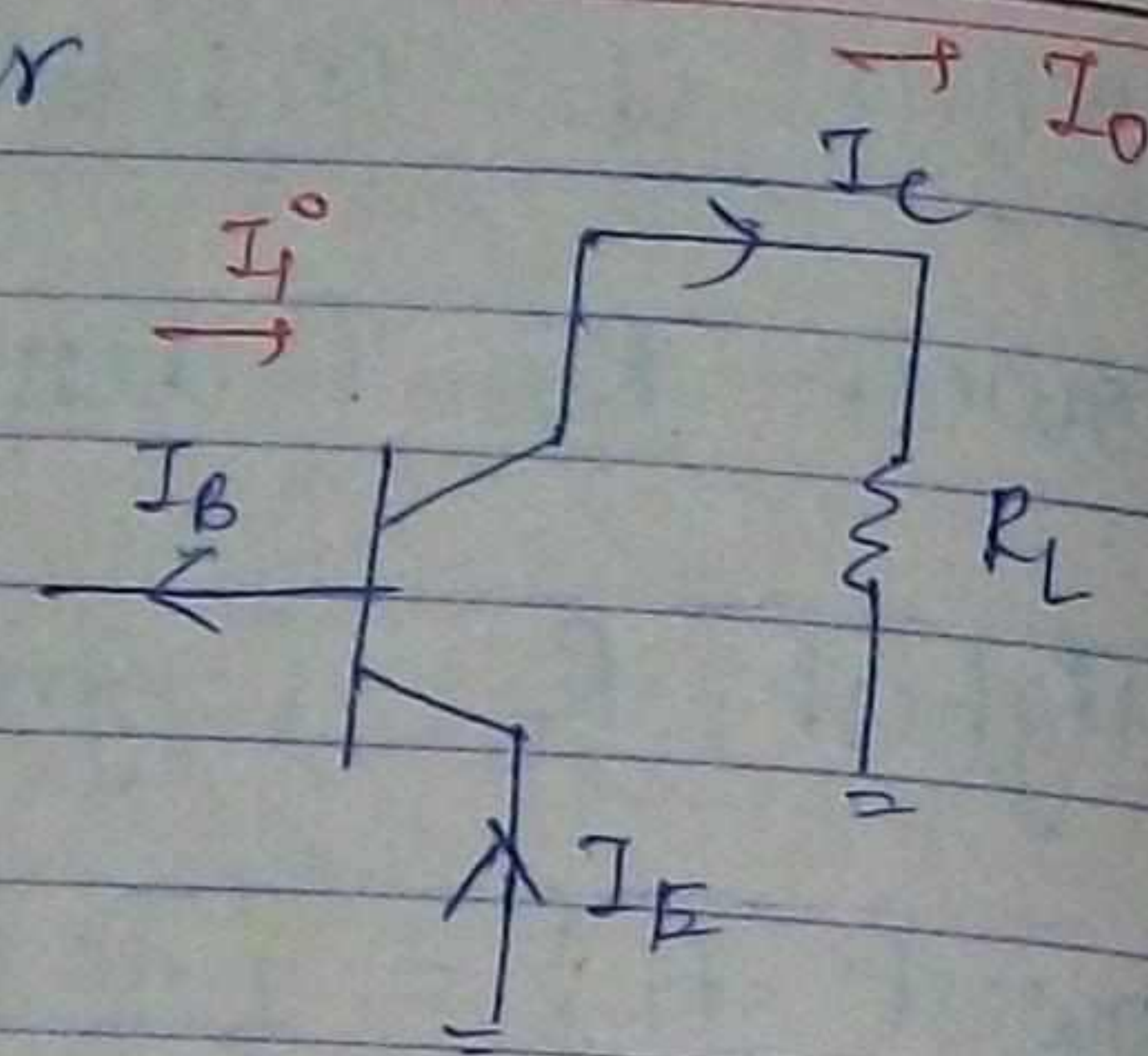
- ⑦ Common base amplifier will offer largest bandwidth and therefore it is more suitable for high frequency application
current gain $\times$ BW = CONSTANT

for Common emitter



$$A_i = \frac{I_o}{I_i} = \frac{-I_C}{I_B}$$

$$\phi = 180^\circ$$



$$A_i = \frac{I_o}{I_i} = \frac{-I_C}{I_B}$$

$$\phi = 180^\circ$$

Similarly for common collector
 $\phi = 0$

$$\text{Current gain} \times BW = \text{Constant}$$

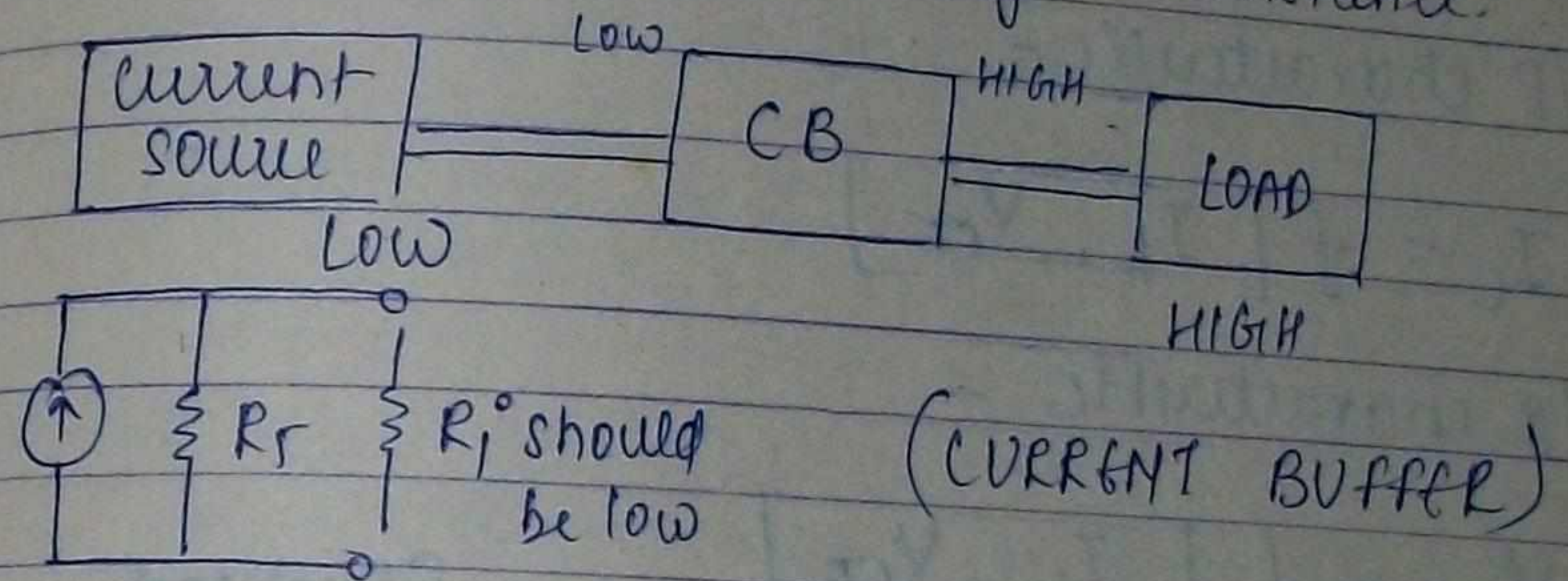
$$\text{Current gain} \approx 1 \quad BW = \text{Highest}$$

⇒ APPLICATION -

- ① It is used as a constant current source
- ② As a non inverting voltage amplifier

③ As a high frequency amplifier.

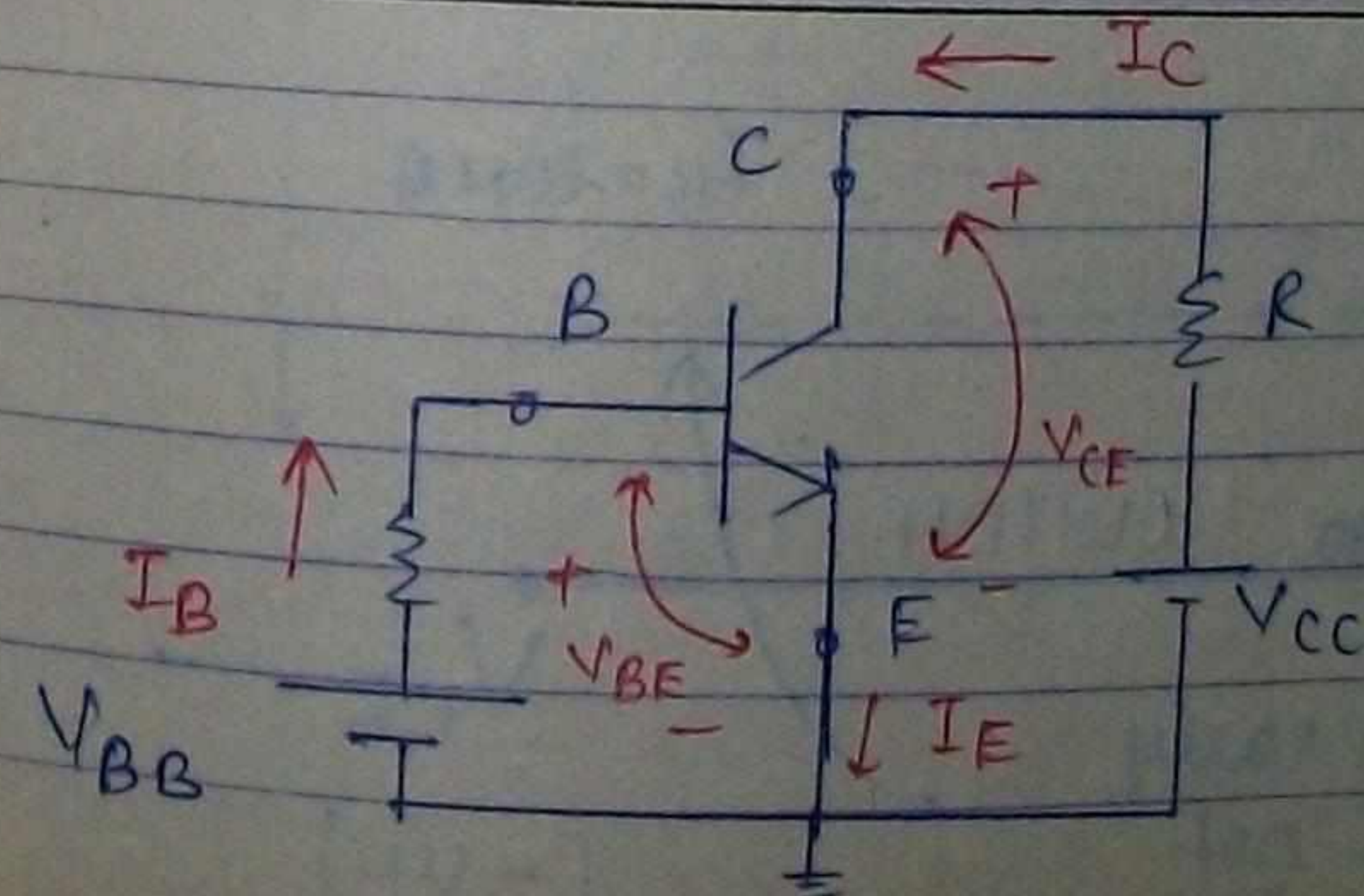
④ As an impedance matching device between low resistance to high resistance.



NOTE -

O/P characteristic should appear as straight horizontal line. However collector current I_C increases as $|V_{CB}|$ increases. This increase in I_C is attributed to a phenomena called Early effect or base width modulation.

⇒ COMMON EMITTER CONFIGURATION -



I_B and V_{CE} independent variables

V_{BE} and I_C are dependent variables

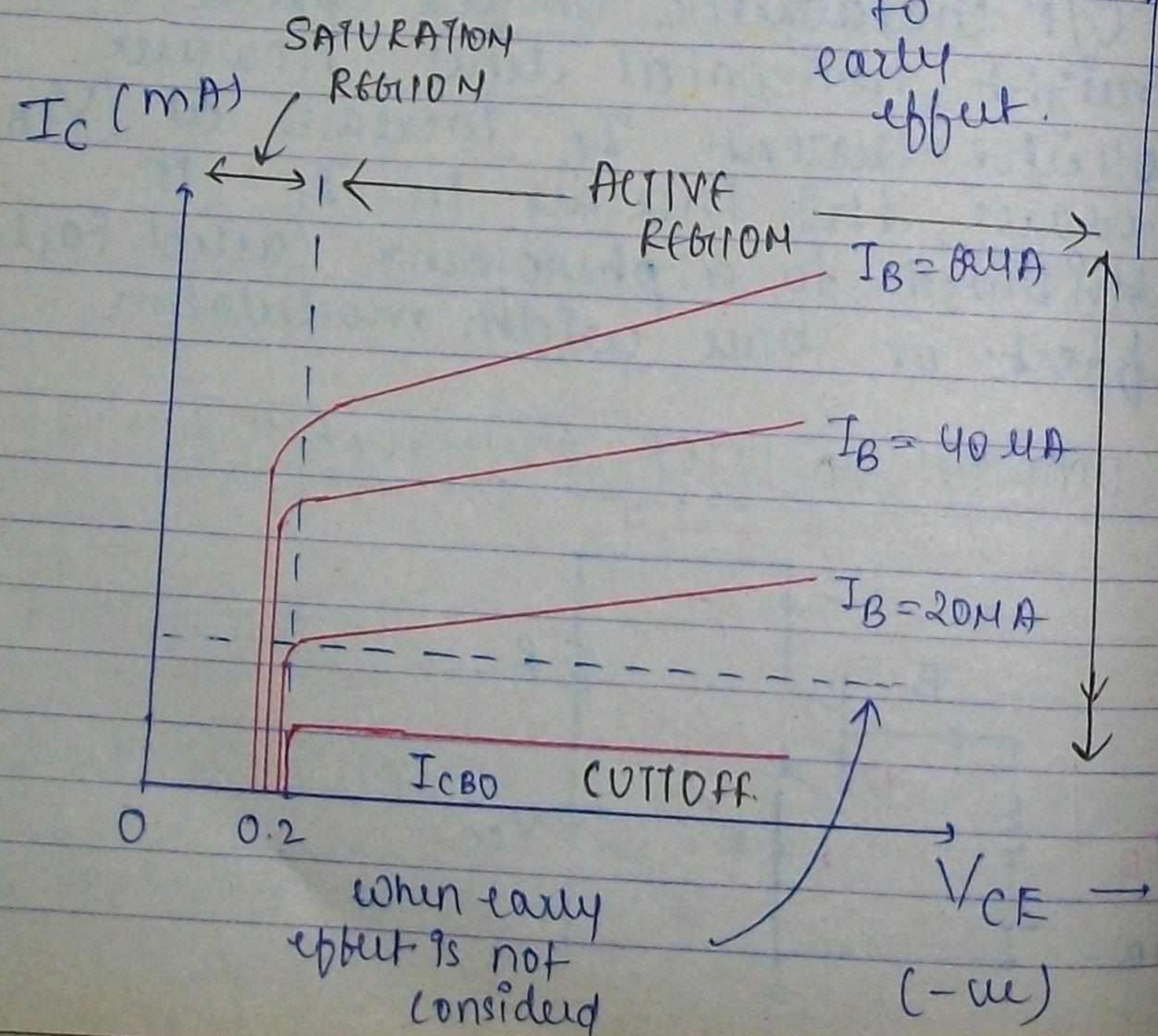
O/P characteristic -

$$I_C = f[I_B, V_{CE}]$$

I/P characteristic -

$$V_{BE} = f[I_B, V_{CE}]$$

⇒ OUTPUT CHARACTERISTICS -



$$I_C = \alpha I_E + I_{CO}$$

$$I_E = I_B + I_C$$

$$I_C = \alpha [I_B + I_C] + I_{CO}$$

$$I_C = \frac{\alpha}{1-\alpha} I_B + \frac{I_{CO}}{1-\alpha}$$

where,

$$\frac{\alpha}{1-\alpha} = \beta = \text{large signal current gain for Tx. in CE configuration}$$

$$\frac{1}{1-\alpha} = 1 + \beta$$

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

ACTIVE REGION

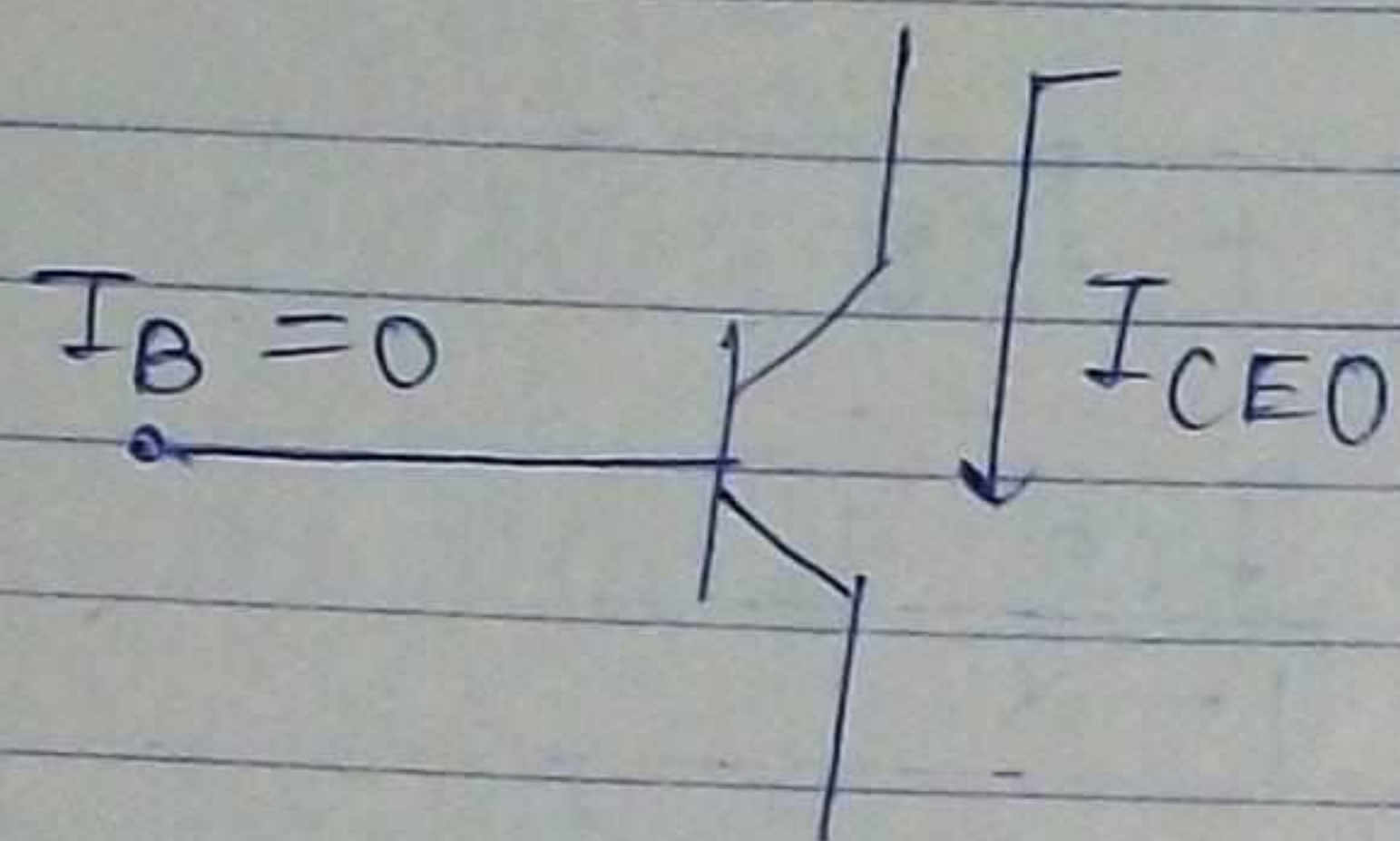
④ CUTOFF REGION -

$$I_E = 0 \quad I_C = I_{CO}$$

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$I_C = \beta I_B + \frac{1}{1-\alpha} I_{CO}$$

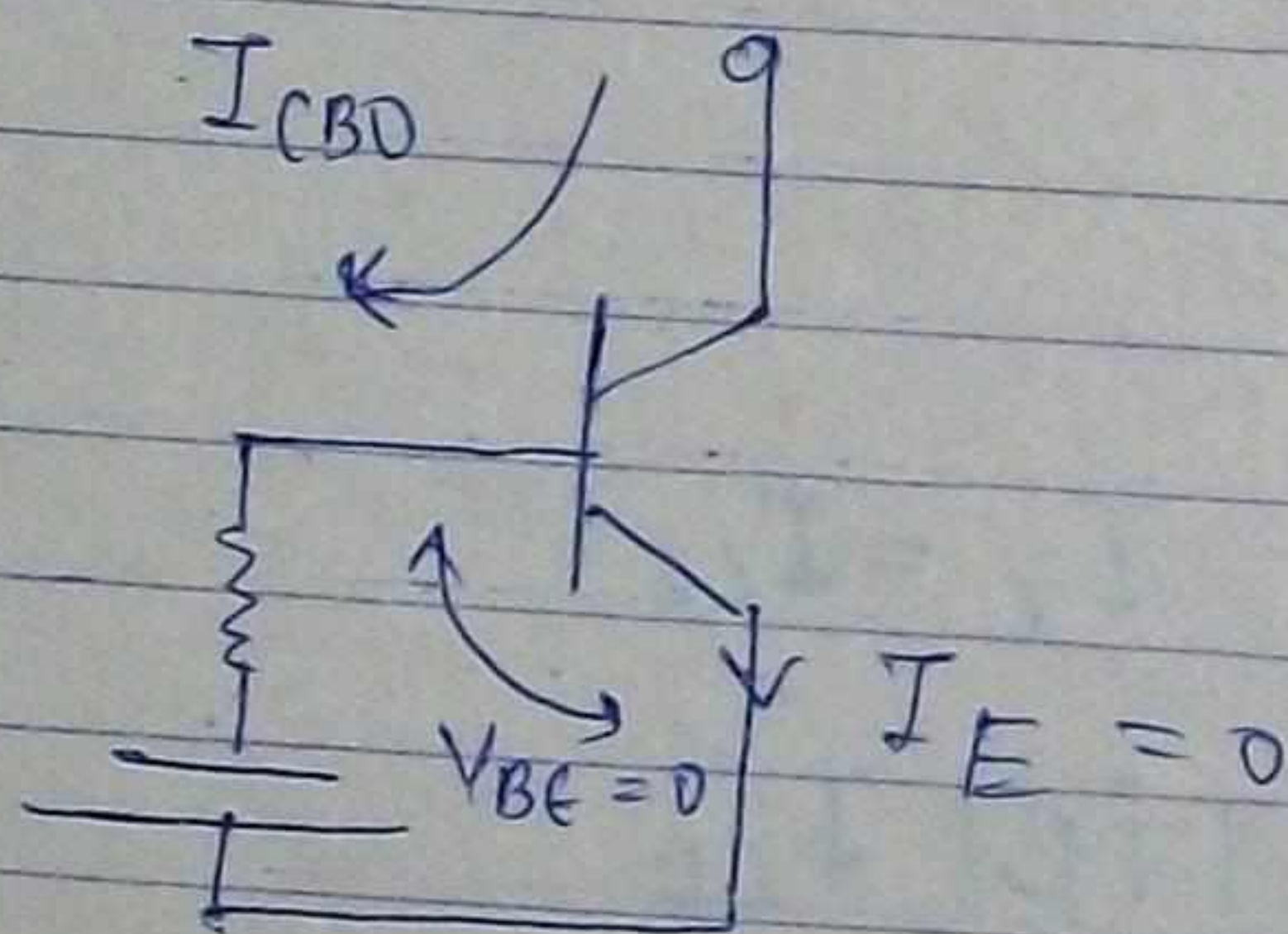
$$I_B = 0, \quad I_C = \frac{I_{CO}}{1-\alpha} \quad \leftarrow \text{very high}$$



I_{CEO} = collector to emitter current when base is open i.e. $I_B = 0$

$$I_{CEO} = \frac{I_{CO}}{1-\alpha} \quad I_{CEO} \gg I_{CO}$$

i.e. reverse saturation current for practical transistor (I_{CEO})



I_{CBO} = collector to base current when emitter is open i.e. $I_E = 0$

I_{CO} = reverse saturation current of ideal diode

$$I_{CEO} > I_{CBO} > I_{CO}$$

- If $I_B = 0$ $I_E = I_B + I_C \approx I_C$ and

$$I_C = I_E = \frac{I_{CO}}{1-\alpha}$$

The actual collector current with collector junction R.B and base open-circuited is designated by symbol I_{CEO} .

- Cutoff is defined as condition when $I_C =$ reverse saturation current I_{CEO} and emitter current $I_E = 0$

In order to cutoff the Tx, it is not enough to reduce I_B to zero instead it is necessary to reverse bias the emitter junction slightly i.e.

$$V_{BE} = -0.1 \text{ Ge.} \\ V_{BE} = 0.0 \text{ Si.}$$

- Reverse collector saturation current (I_{CBO})
The I_C in a physical Tx (a real non idealised or commercial device)

when $I_E = 0$ is designated by symbol I_{CBO} . Two factors cooperate to make I_{CBO} larger than I_{CO}

- ① There exists a leakage current which flows not through the junction but around it and across the surface.

leakage current $\propto$ voltage across the junction.

- ② New carriers may be generated by collision in collector junction transition region leading to avalanche multiplication of current and may cause breakdown.

— I_{CBO} is μA for Ge and is nA for Si

— I_{CBO} approx. double for every $10^\circ C$ rise in temp. for both Ge and Si. Therefore Si can be used upto about $200^\circ C$ and Ge upto $100^\circ C$

* ACTIVE REGION —

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$\text{and } \beta = \frac{\alpha}{1 - \alpha}$$

Early effect $\uparrow$, $\alpha \uparrow$

let, $\alpha = 0.98 \rightarrow 0.985$

$\therefore \alpha$ increases by 0.5%.

$$\beta = \frac{0.98}{1 - 0.98} = 49, \quad \beta' = \frac{0.985}{1 - 0.985} = 63$$

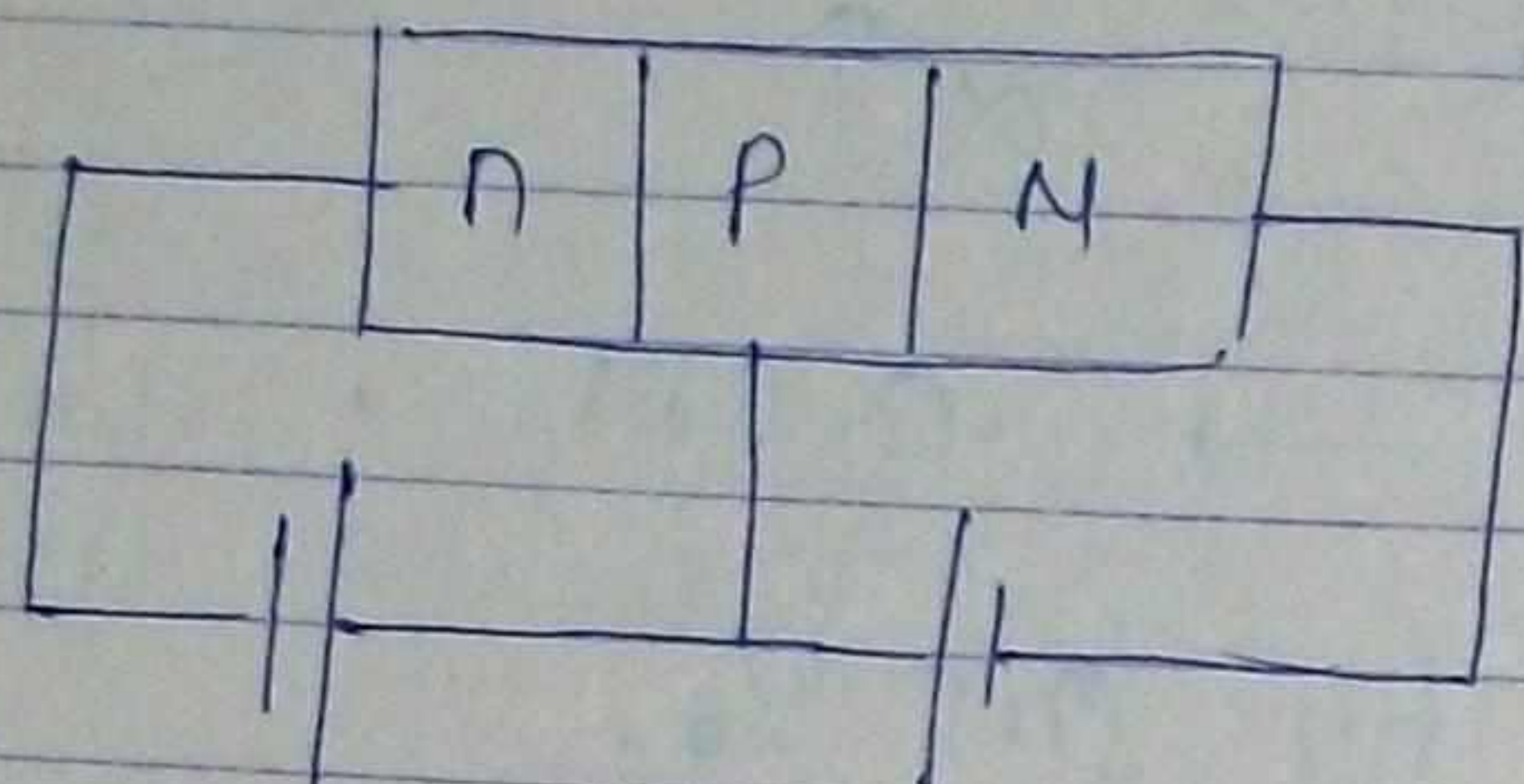
$\therefore \beta$ increases by 34%.

— Due to Early effect an increase in I_C is expected with increase in $|V_{CE}|$ then this influence of early effect is more pronounced in a CE Tx. as can be seen from larger slope of characteristic compared with common base Tx. Therefore CE characteristics are normally subjected through wide variation among Tx of a given type.

* SATURATION REGION —

— Both I_E and I_C are forward biased by

atleast the cuttin voltage V_r



$$V_{BE} = 0.8 \quad V_{BC} = 0.6 = V_B - V_C$$

(saturation)

$$V_{BE} = 0.7 \quad V_B > V_C$$

(active)

CEB

$$V_{CE} + V_{EB} + V_{BC} = 0$$

$$V_{CE} = V_{BE} - V_{BC}$$

$$V_{CE} = 0.8 - 0.6 = 0.2$$

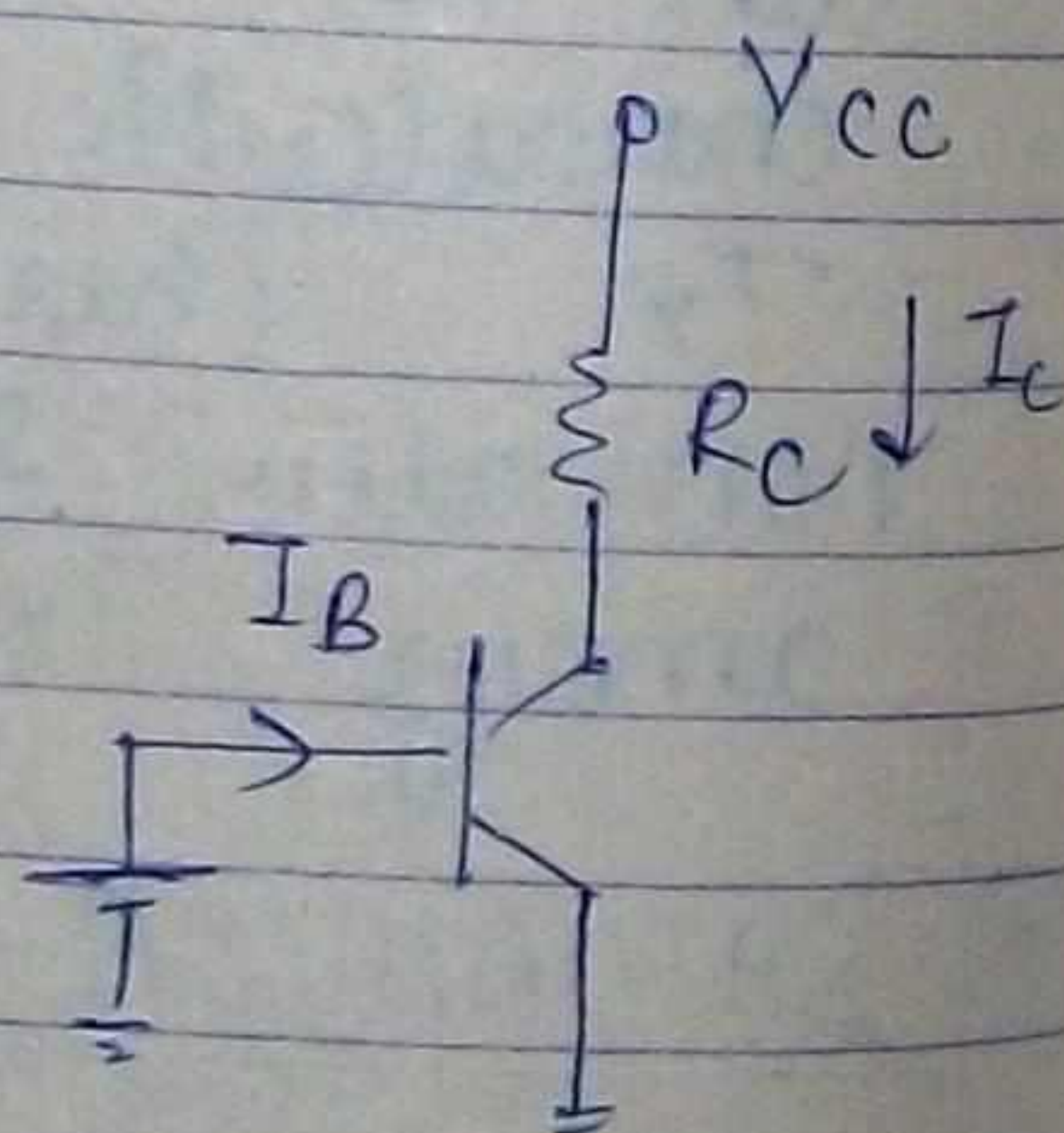
$$V_{CE} \leq 0.2$$

(Tx is in saturation)

$$\# V_{CE} = V_{CC} - I_C R_C$$

$$I_B \uparrow, I_C \uparrow, V_{CE} \downarrow$$

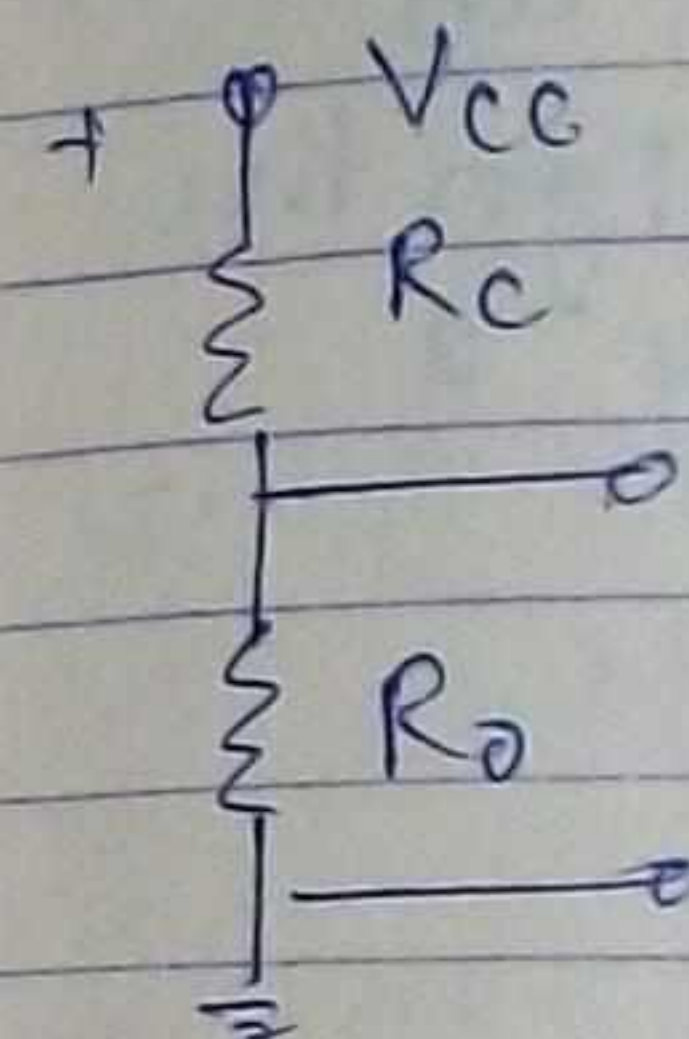
$$V_{CE} = 0.2$$



$$I_C = \beta I_B \rightarrow \text{varying} \uparrow$$

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

$$I_B \uparrow, I_C \uparrow, V_{CE} \downarrow$$



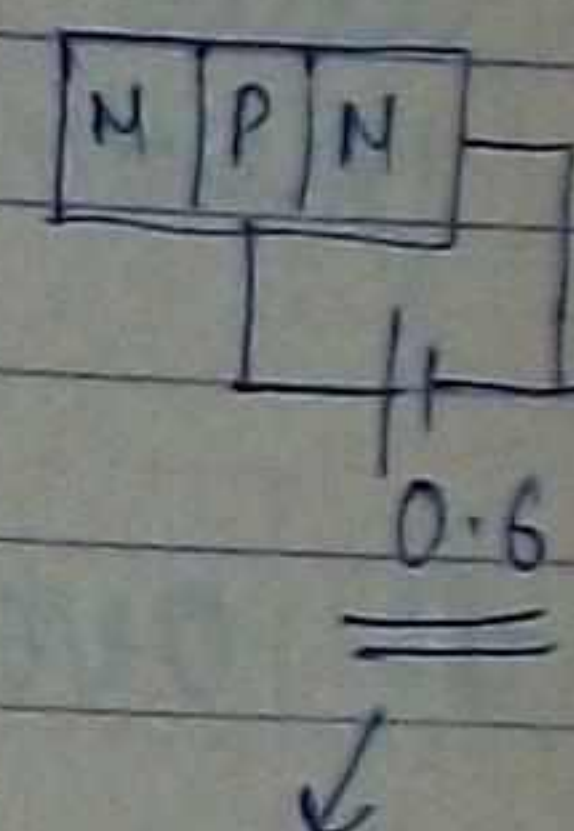
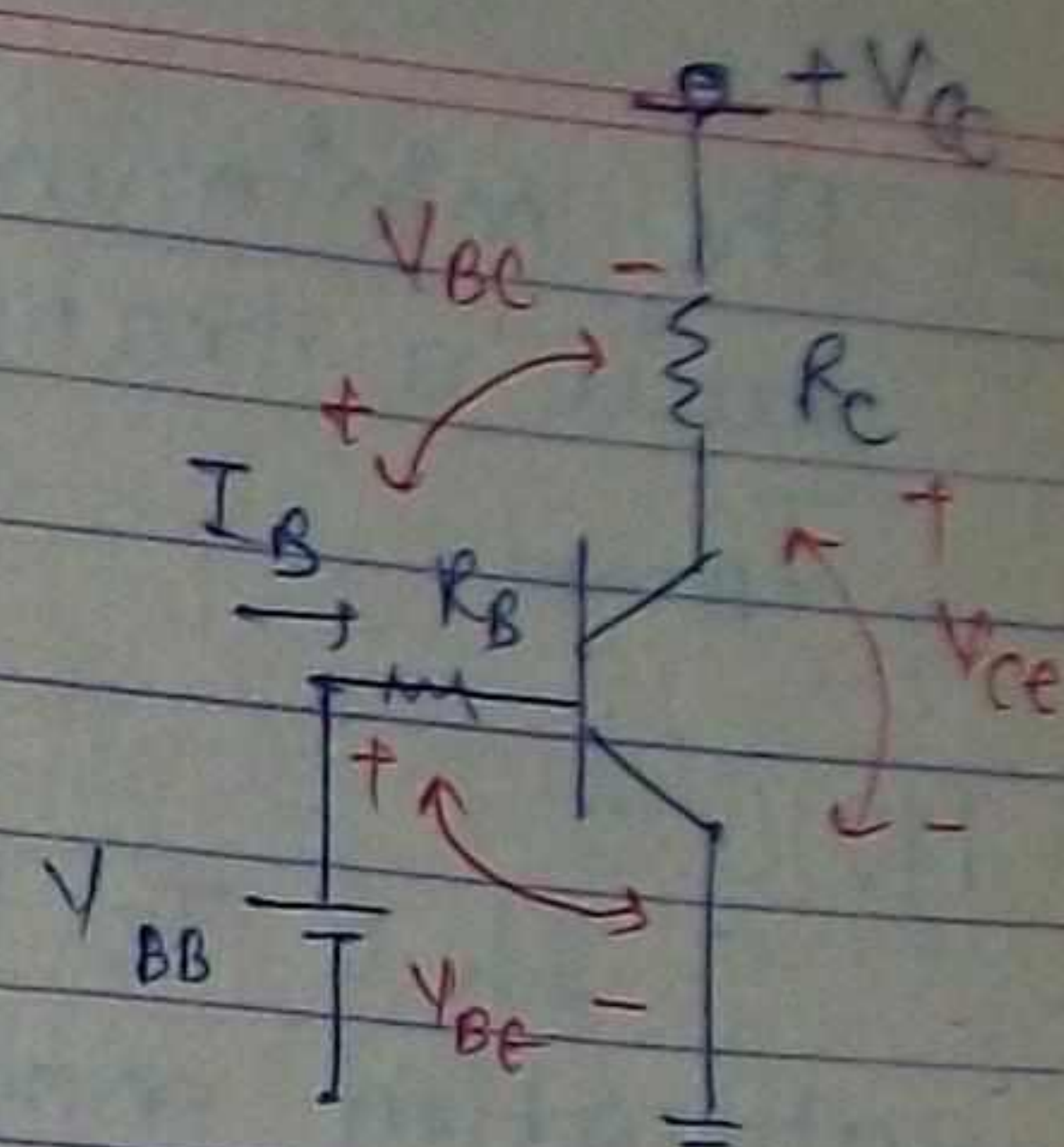
IN SATURATION -

$$V_{CE\text{SAT}} = 0.2 \quad V_{BE} = 0.8 \quad V_{CE} = 0.2$$

$$V_{CE} + V_{EB} + V_{BC} = 0$$

$$V_{BC} = -V_{CE} + V_{BE}$$

$$= -0.2 + 0.8 = 0.6$$



means saturation

$$\# I_{C\text{SAT}} = \frac{V_{CC} - V_{CE\text{SAT}}}{R_C}$$

$$I_B = \frac{V_{BB} - V_{BE}}{R_B}$$

$$I_{B\text{min}} = \frac{I_{C\text{SAT}}}{\beta}$$

$$I_B \geq I_{B\text{min}}$$

then the Transistor is in saturation.

- The minimum base current required to drive a transistor into saturation is I_{Bmin} .

NOTE -

If a Tx have to be operated in saturation mode, we should design the circuit so that $I_B > I_{Bmin}$ by a factor of 2 to 10.

$$\# \text{ OVERDRIVE FACTOR} = \frac{I_B}{I_{Bmin}}$$

⇒ FORCED β (β_{FORCED}) -

The ratio of I_{CSAT} and I_B to ensure saturation is called forced β and is given by:

$$\beta_{FORCED} = \frac{I_{CSAT}}{I_B}$$

⇒ PROPERTIES OF COMMON EMITTER -

- ① R_i is high ($\approx 1k\Omega$)
- ② R_o is high (50k Ω to 500k Ω)

③ A_i is high

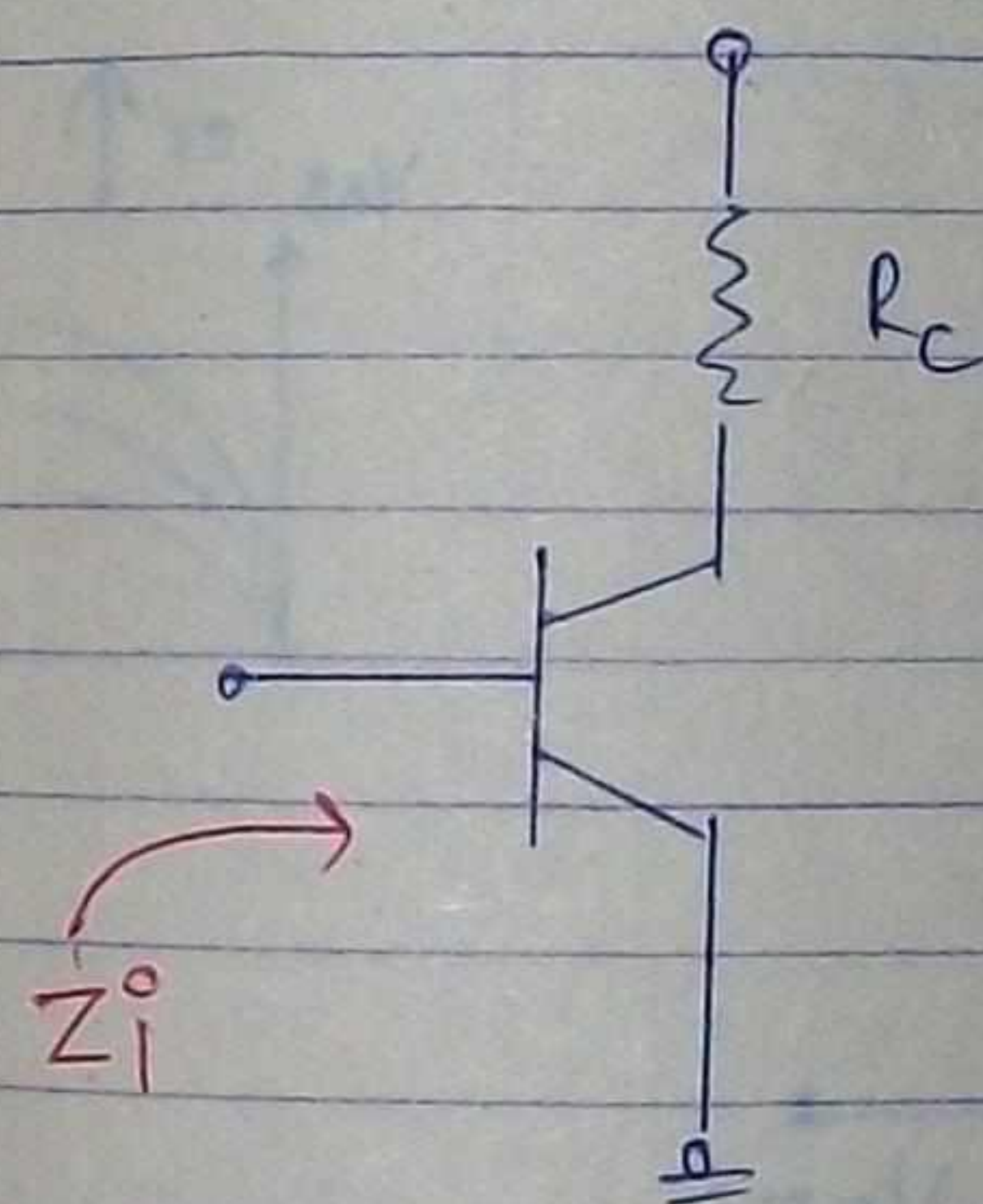
④ A_v is high

⑤ Highest power gain A_p ($A_p = A_i A_v$)

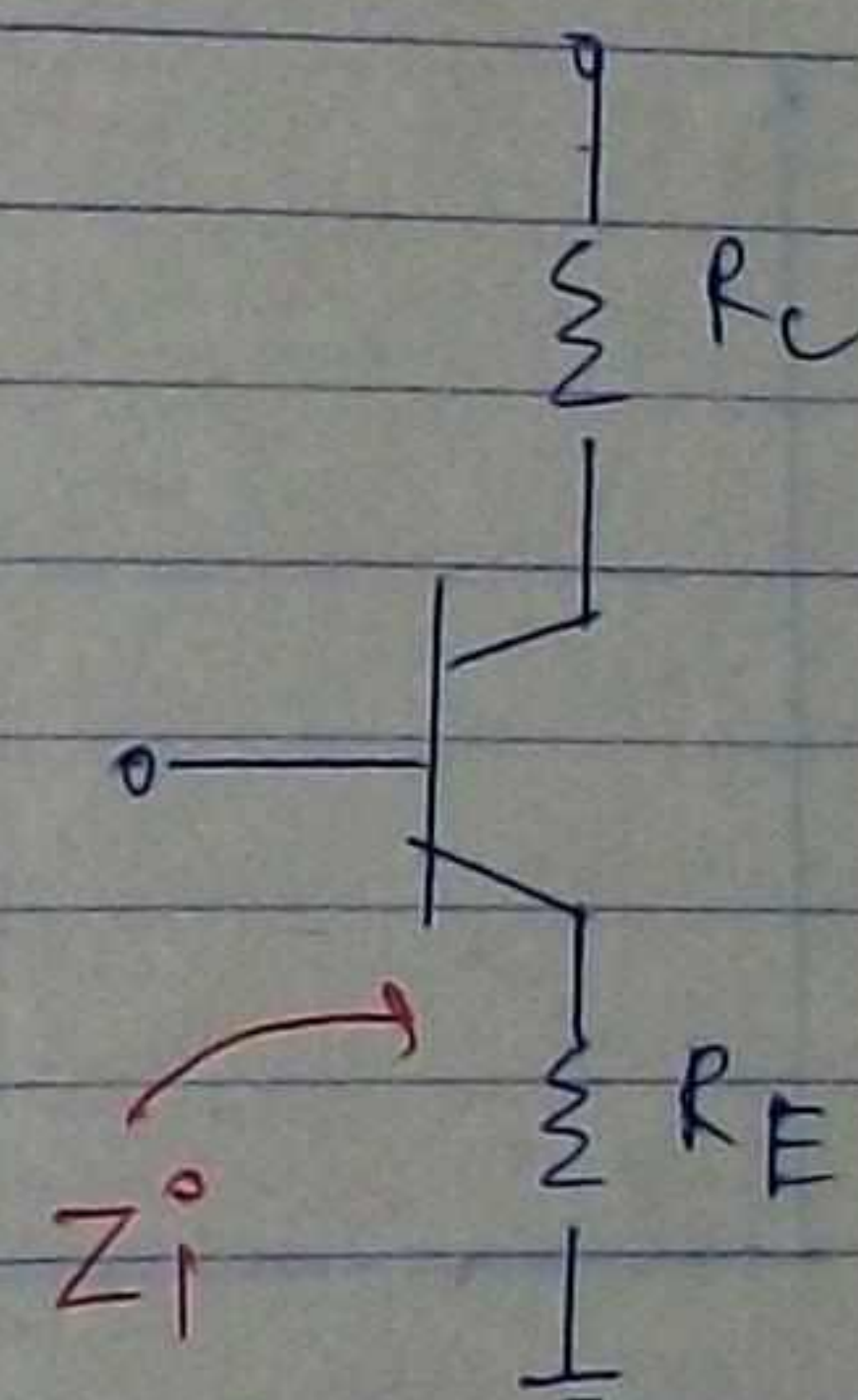
⑥ Phase shift is 180° ($\phi = 180^\circ$)

⑦ Most common and very popularly used configuration.

⇒ COMMON EMITTER WITH EMITTER RESISTANCE (R_E)



$$Z_i = h_{ie} = 1.1k\Omega$$



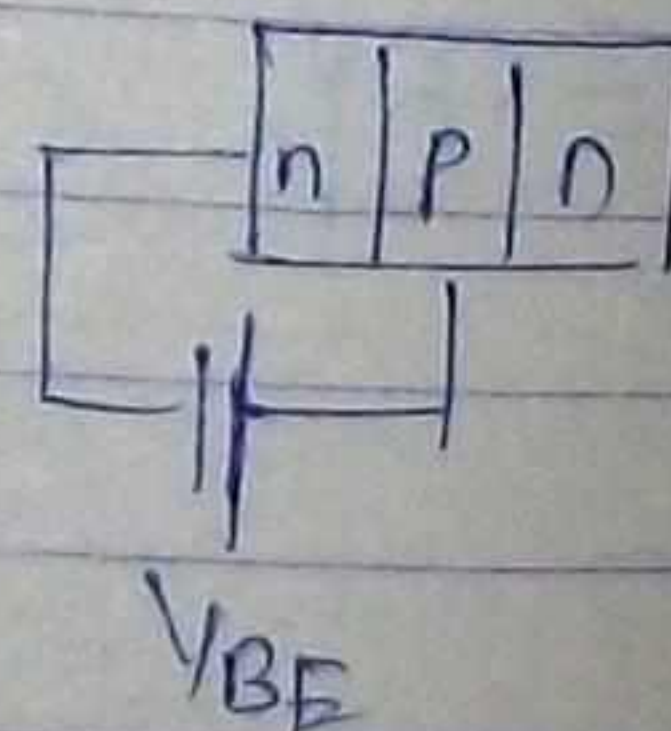
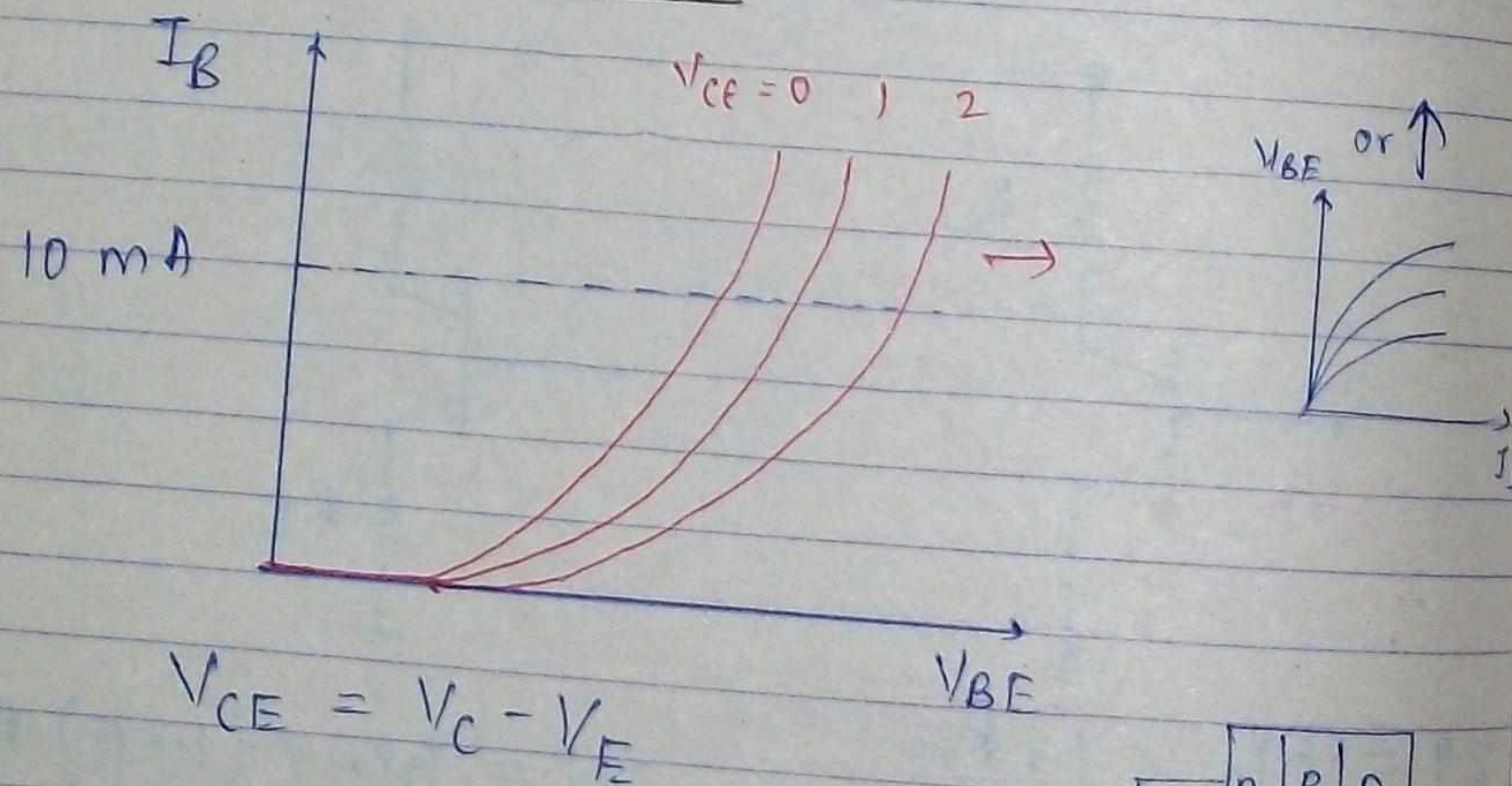
$$\begin{aligned} Z_i &= h_{ie} + (1 + \beta) R_E \\ &= 1.1 + (1 + 99) 1k\Omega \\ &= 101k\Omega \end{aligned}$$

ADVANTAGES -

- ① It increases input resistance.

- ② provides stabilization.
- ③ Increase output resistance.
- ④ It introduces a -ve feedback and therefore reduced the voltage gain A_v of amplifier.
- ⑤ The feedback mechanism is current-series feedback.

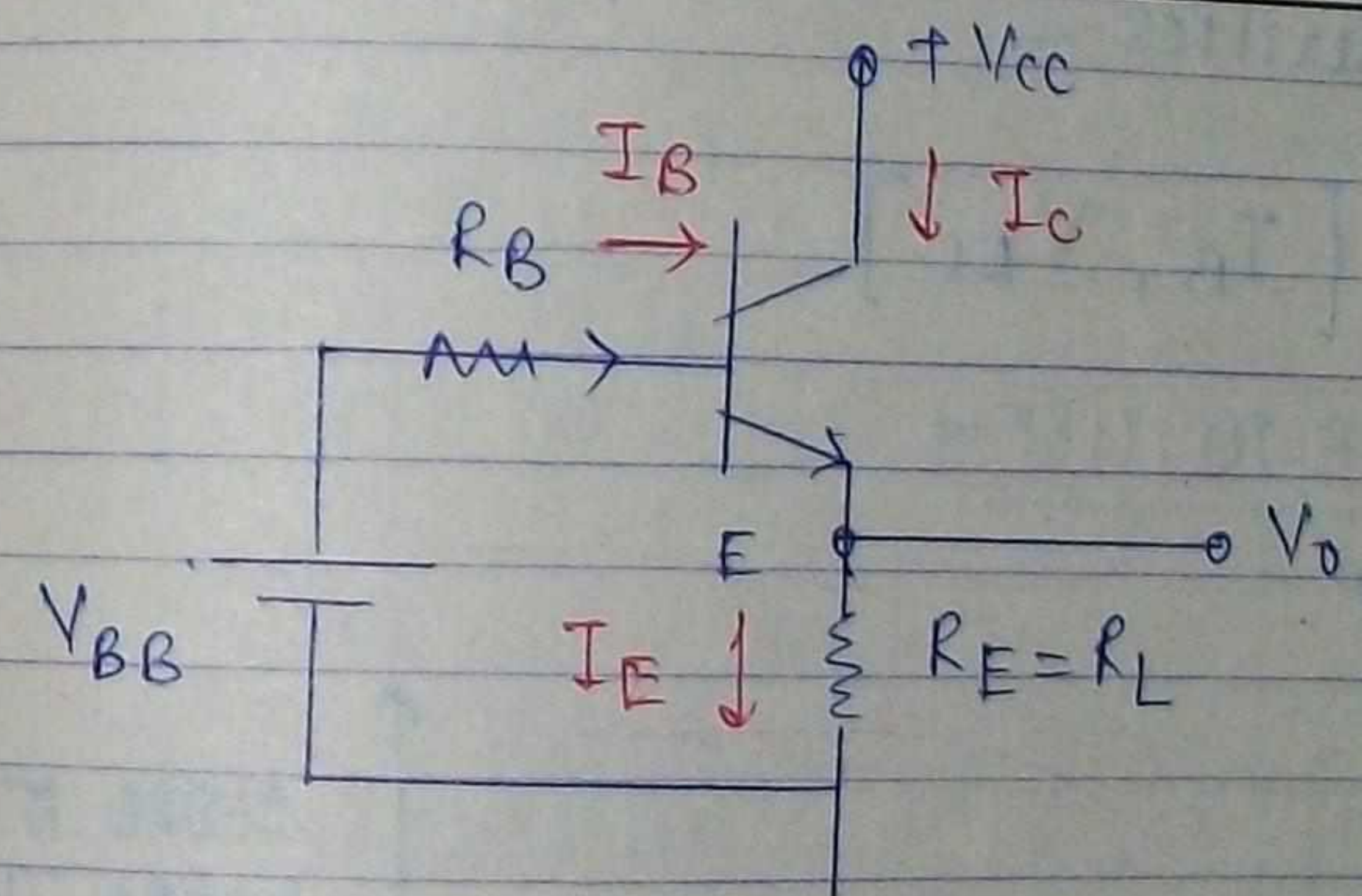
INPUT CHARACTERISTICS -



$V_{CE} \uparrow, V_C \uparrow$
 Early effect $\uparrow, I_E \uparrow, I_C \uparrow, I_B \downarrow$

- Increasing $|V_{CE}|$ with constant V_{BE} cause a decrease in Base width and result in decreasing recombination base current with collector shorted to emitter i.e. $V_{CE} = 0$ and emitter junction J_E is f-b diode. The I/P characteristic is that of a f-b diode.

⇒ COMMON-COLLECTOR CONFIGURATION -

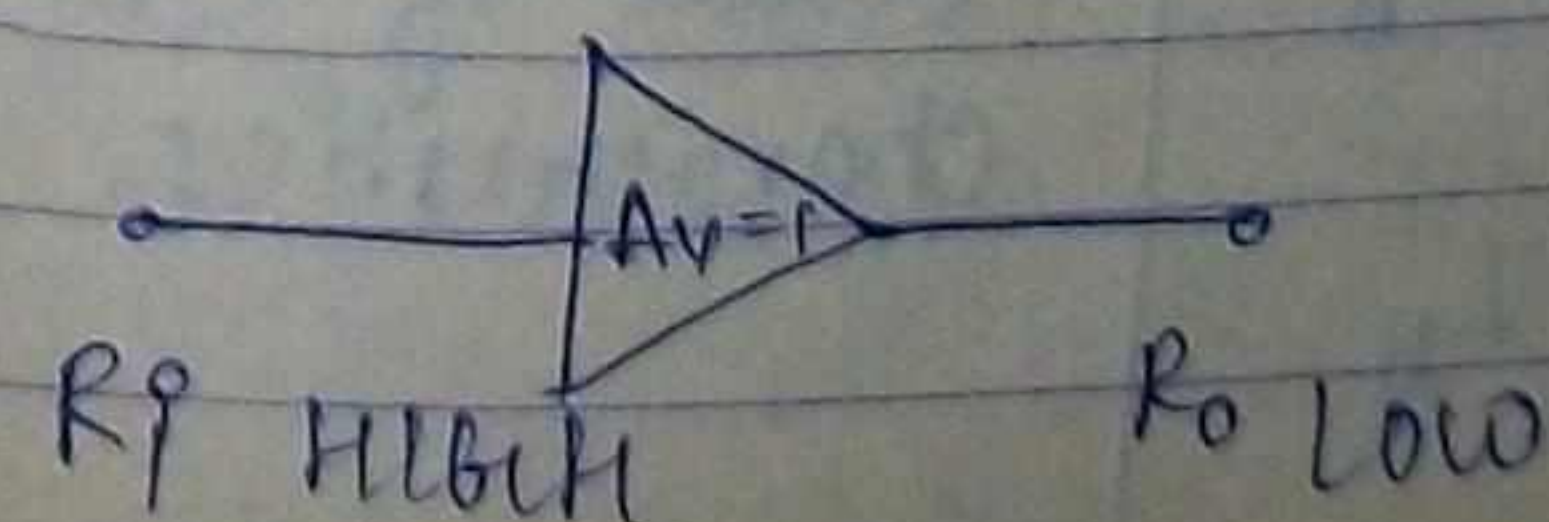


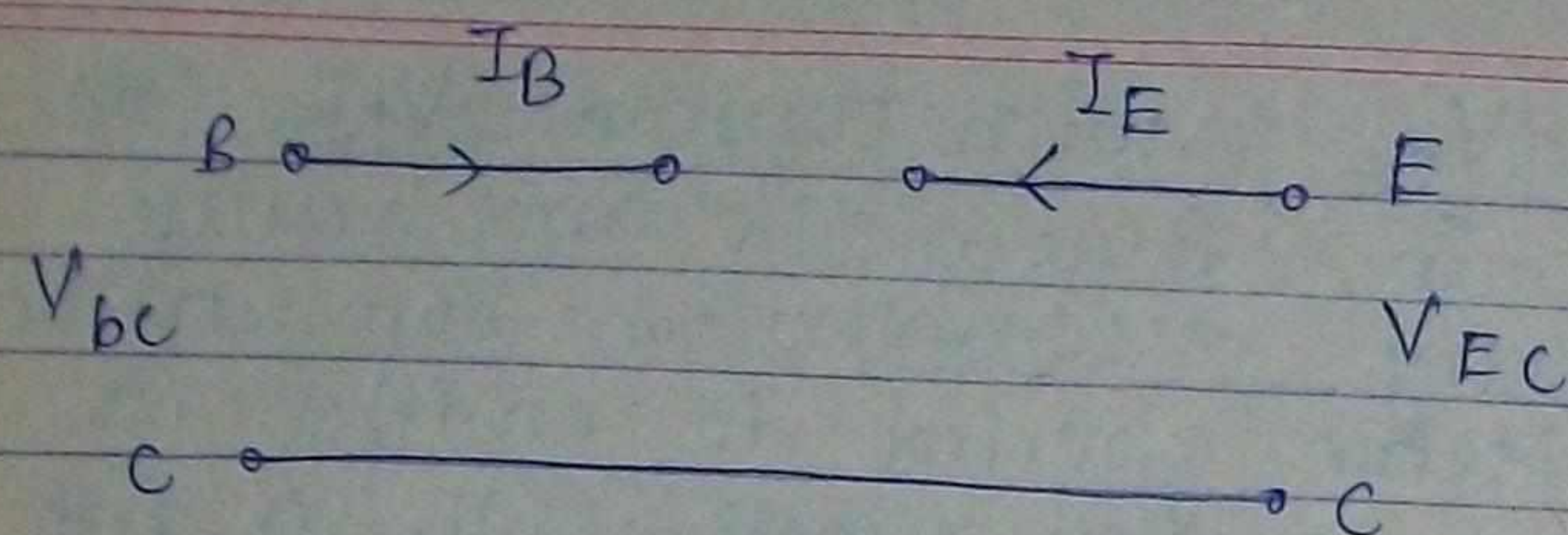
$$A_v = 1, V_o = V_i$$

Emitter follower

$$Z_i = h_{ie} + (1 + \beta)R_E$$

BUFFER





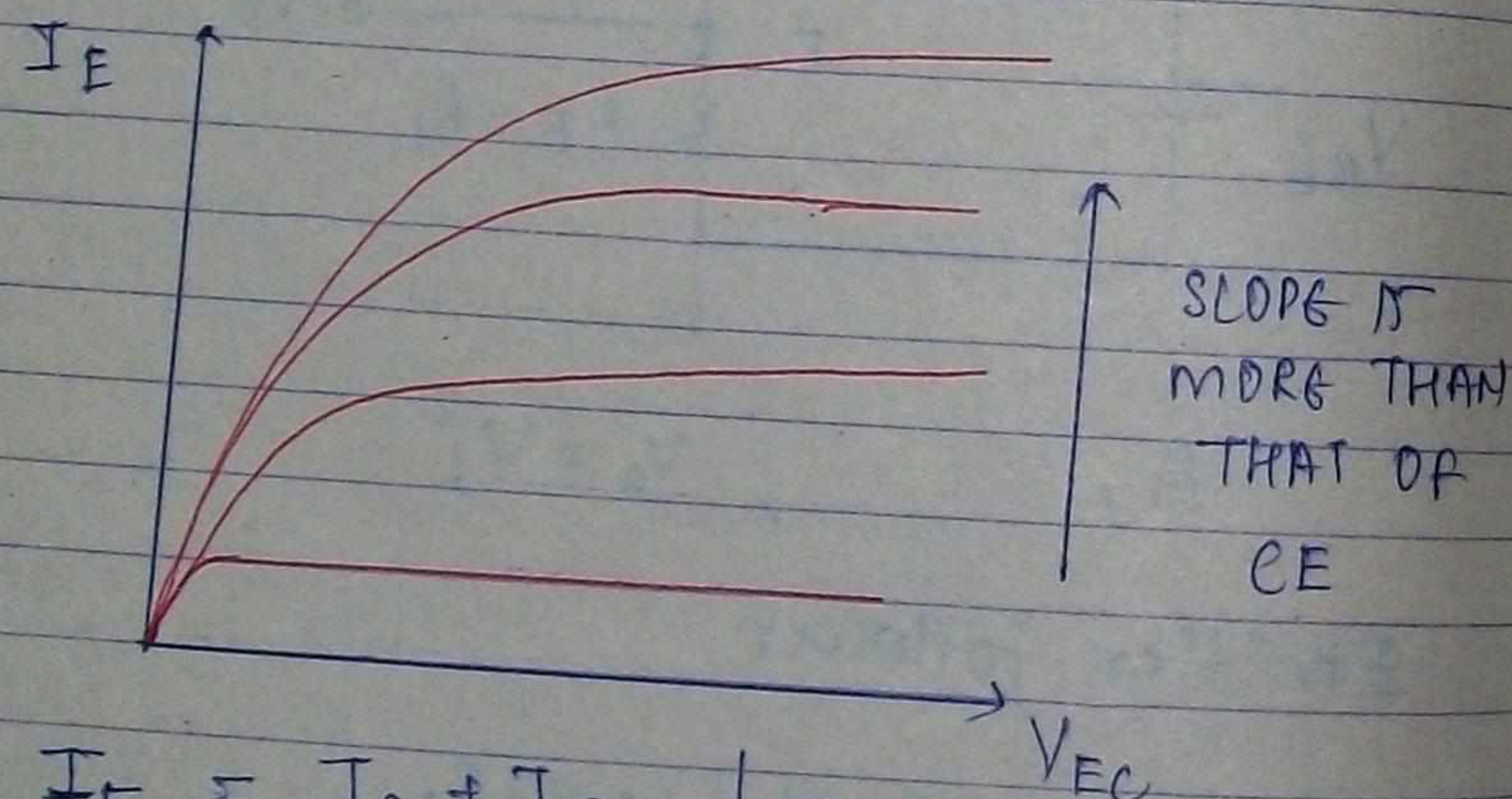
I/P characteristics -

$$V_{BC} = f[I_B, V_{EC}]$$

O/P characteristics -

$$I_E = f[I_B, V_{EC}]$$

OUTPUT CHARACTERISTICS -



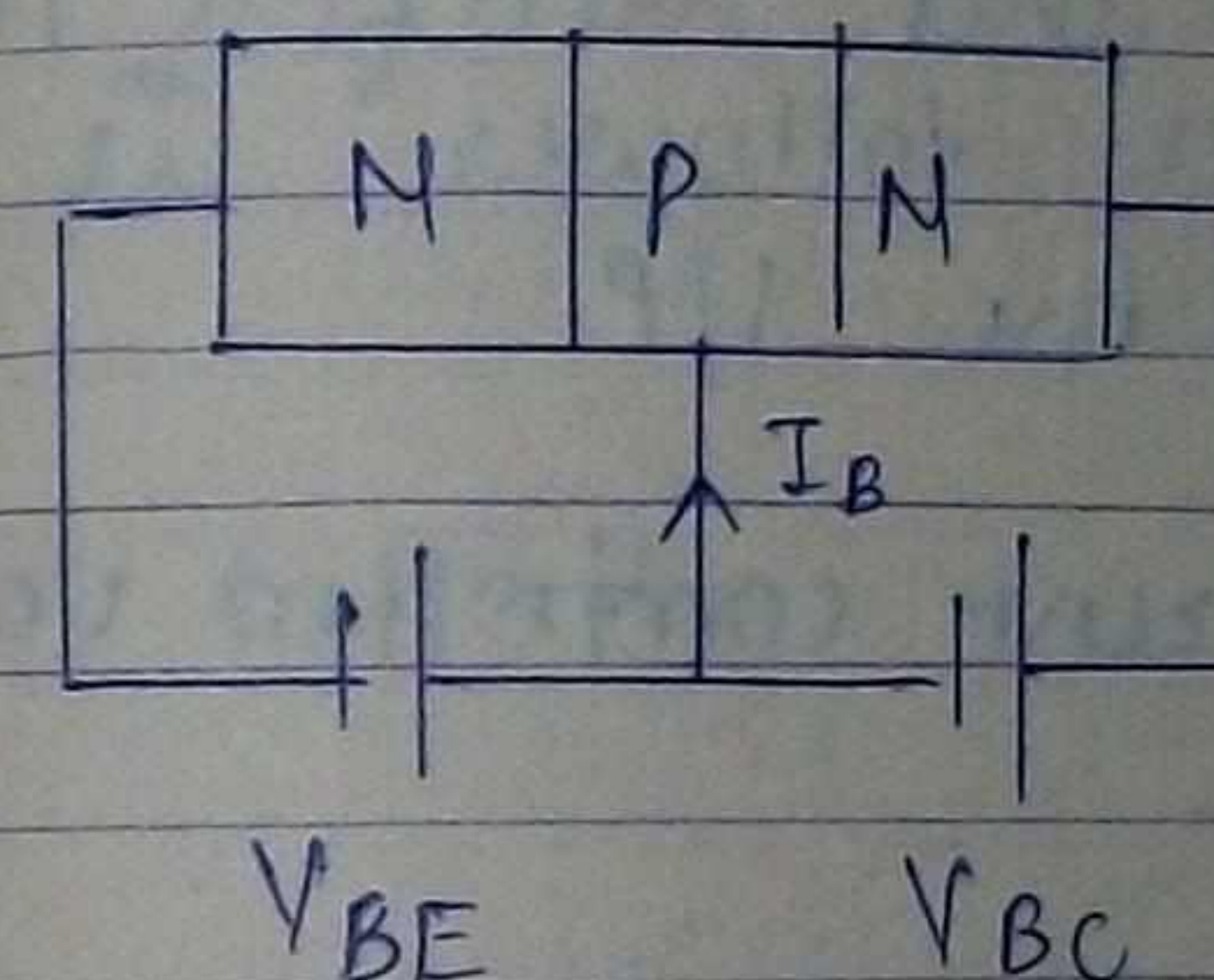
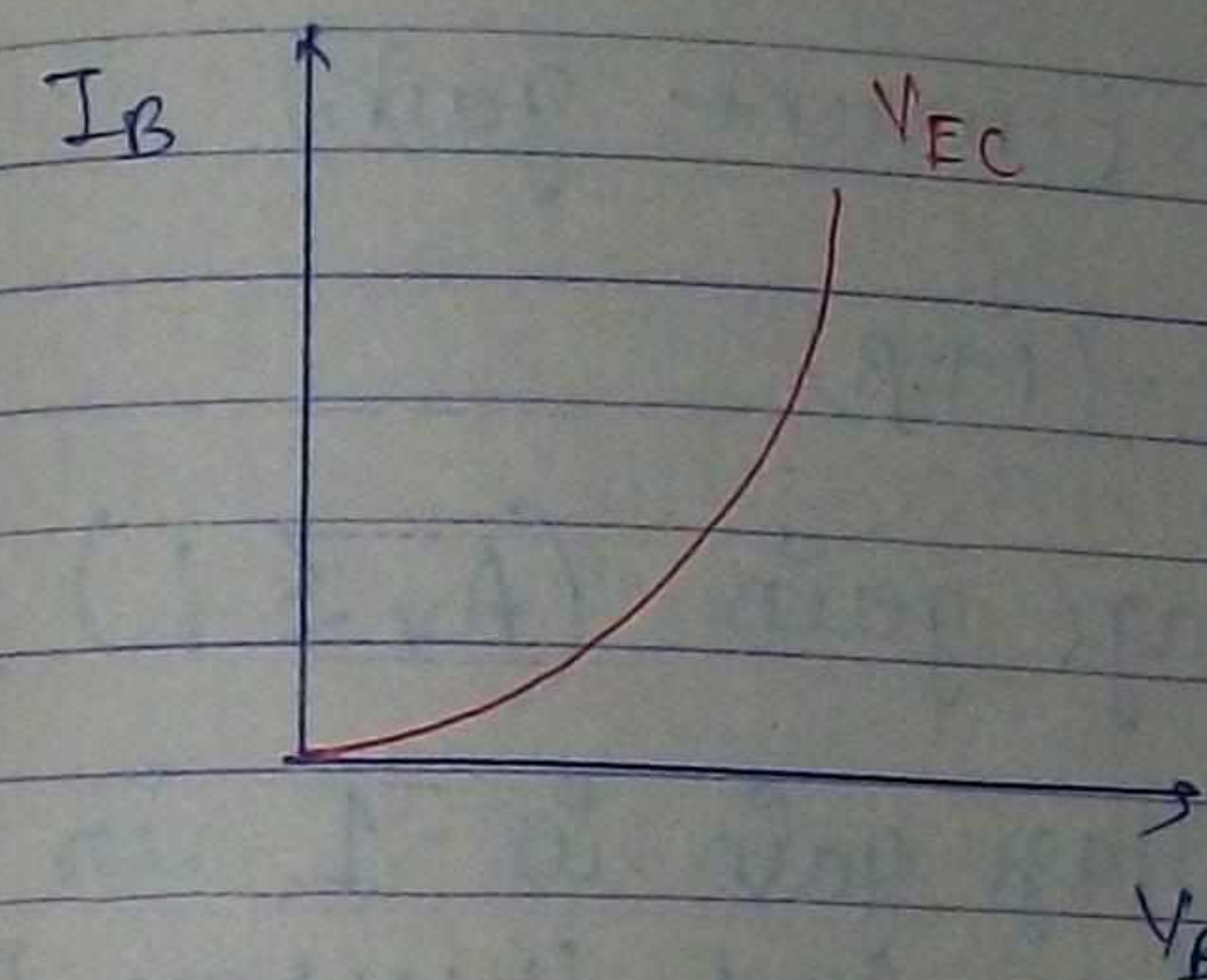
$$I_E = I_C + I_B$$

$$I_E = \beta I_B + I_B$$

$$I_E = (1 + \beta) I_B$$

* Curves are more elevated than the curves of CE o/p characteristics.

INPUT CHARACTERISTICS -



I/P characteristic

- Input characteristic is not similar to forward characteristic of diode.

- CE is popularly known as emitter follower

⇒ PROPERTIES -

① Highest R_i (50 k Ω to 500 k Ω)

② Larger input resistance is mainly due to R_E

② lowest R_o ($R_o > 100\Omega$)

③ Highest gain (current gain)

$$A_i = \frac{I_E}{I_B} = (1 + \beta)$$

④ Lowest voltage gain ($A_v \leq 1$)

The max. voltage gain is 1 in ideal condition and therefore it is called emitter follower or emitter follower or emitter follower the I/P.

⑤ Basically a current controlled voltage source (CCVS).

⑥ Emitter follower is analogous to voltage follower in OPAMP and source follower with FET.

⑦ Voltage follower is (VCVS)
source follower is (VCVS).

⑧ Lowest power gain A_p (typical value = 48)

⑨ phase shift is zero. ($\phi = 0$)

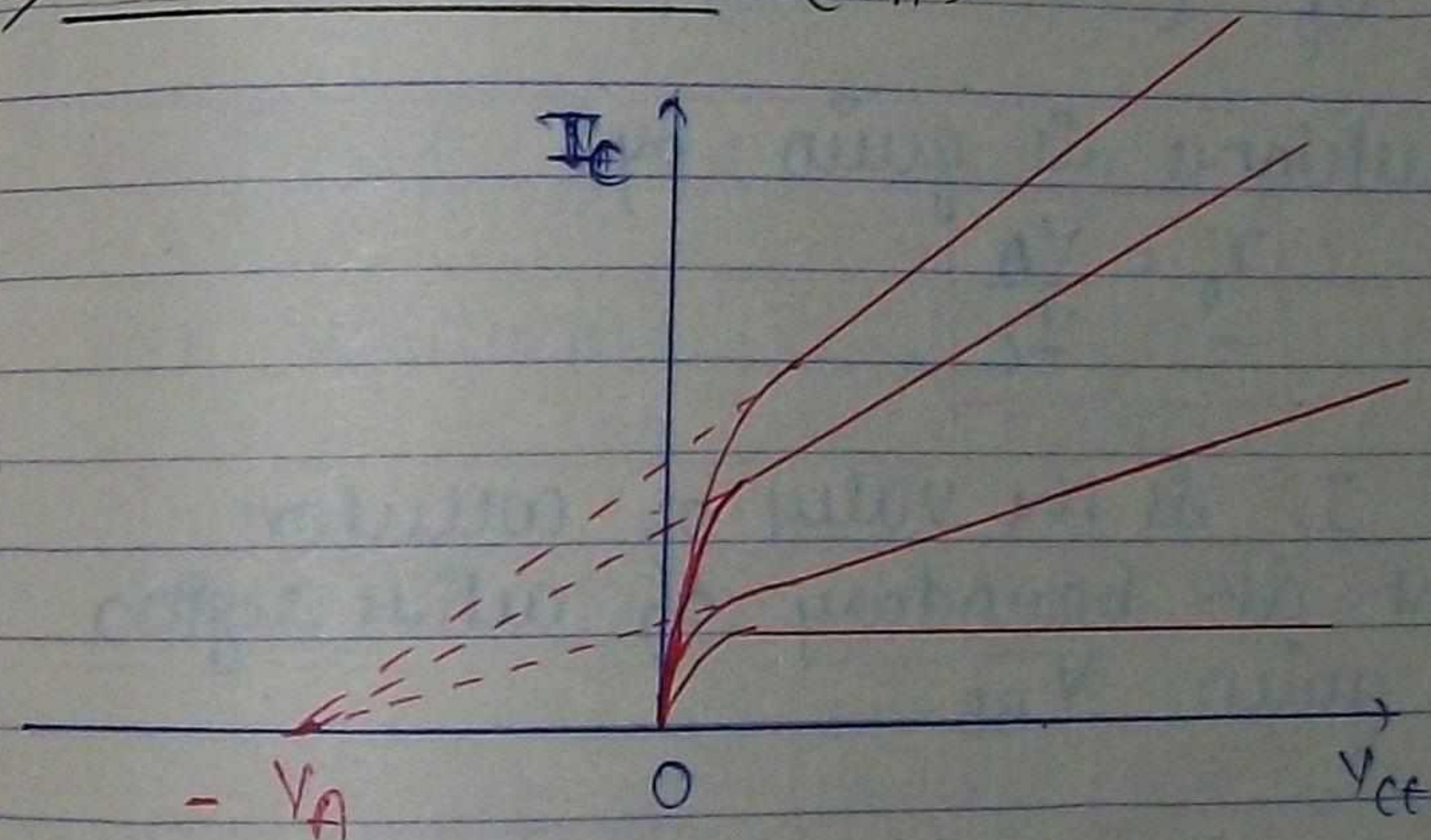
⇒ APPLICATION -

① Highest I/P resistance device

② As a buffer amplifier or as an impedance matching device b/w high resistance to low resistance.

③ As an audio frequency power amplifier.

⇒ EARLY VOLTAGE - (V_A)



$$V_{CE} \uparrow \quad I_E \uparrow \quad I_C \downarrow$$

$$\text{Slope} = \frac{1}{r_o} = \frac{V_2 - V_1}{X_2 - X_1} = \frac{I_{C2} - 0}{V_{CE1} - (-V_A)} \approx \frac{I_{C2}}{V_A}$$

$$r_o = \text{O/P resistance of Transistor} \quad |V_{CE1}| \ll V_A$$

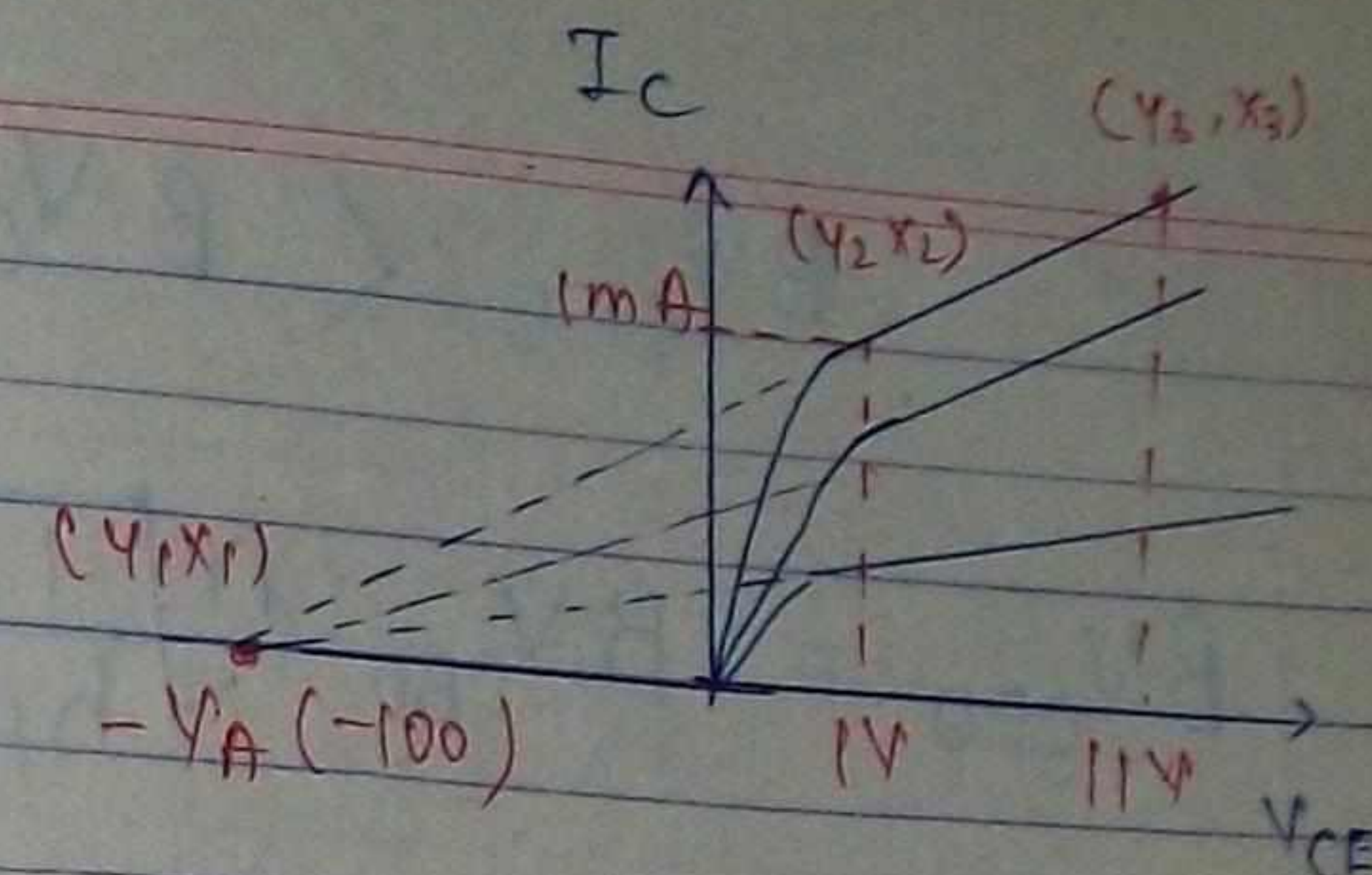
$$r_o = \frac{V_A}{I_{C2}}$$

- It helps in finding the o/p resistance of transistor.
- If linear portion of o/p curve extended backward on negative V_{CE} axis as shown by dashed line then all curve meet at point $-V_A$ ($V_{CE} = -V_A$) the typical value of V_A lies in the range of (50 to 100 V).
- O/P resistance is given by:

$$r_o = \frac{V_A}{I_C}$$

where I_C is the value of collector current at boundary of active region for a given V_{BE} .

Ques. In a CE Tx. at $V_{CE} = 1V$, V_{BE} is adjusted to give a I_C of 1mA. keeping V_{BE} constant V_{CE} is increased to 11V. Find the new value of I_C if $V_A = 100V$.
Soln: and o/p resistance



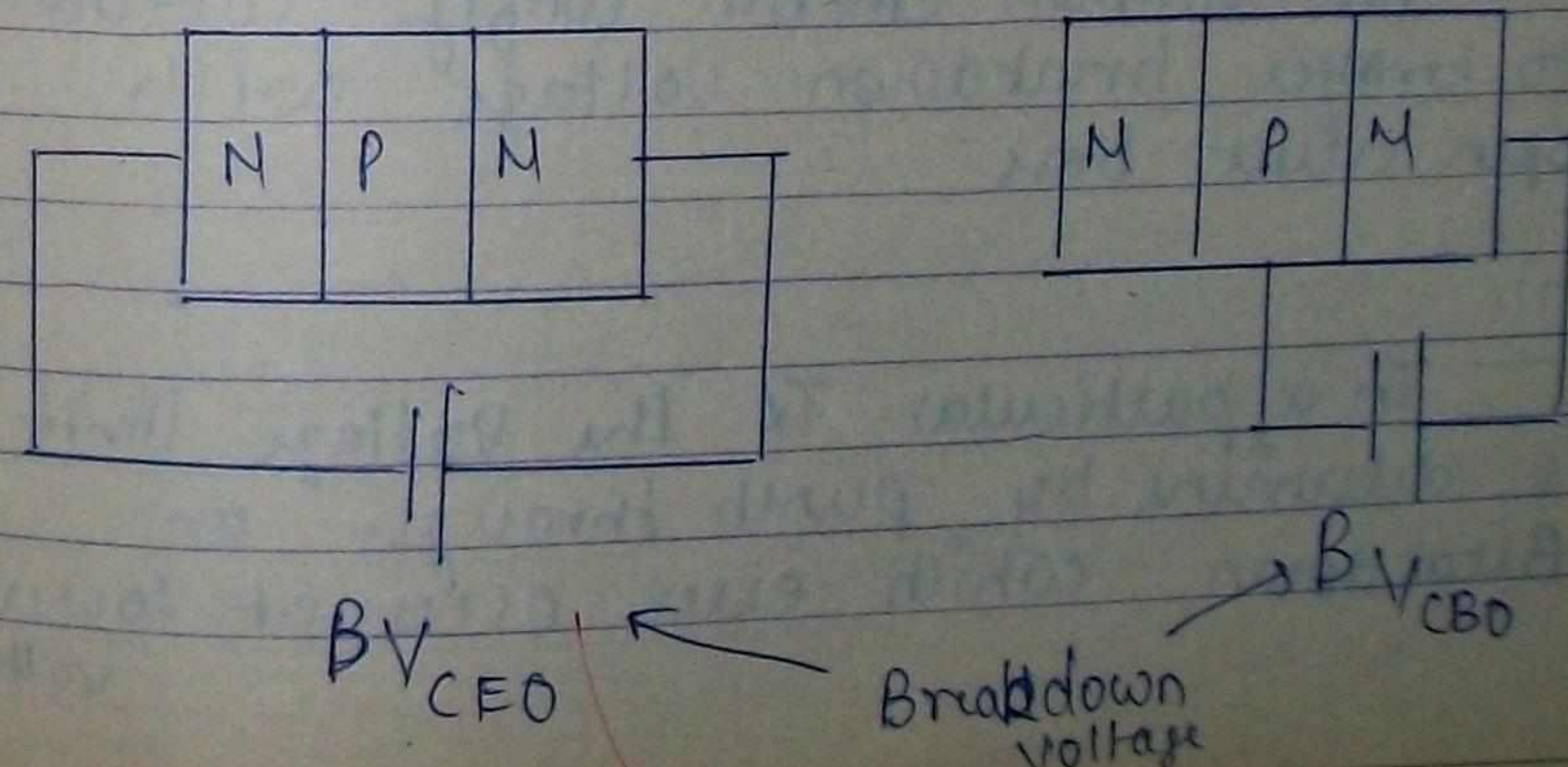
$$\text{slope} = \frac{1}{r_o} = \frac{Y_2 - Y_1}{X_2 - X_1} = \frac{1 - 0}{11 - (-100)} = \frac{1}{101}$$

$$\text{slope} = \frac{1}{101} = \frac{Y_3 - Y_2}{X_3 - X_2} = \frac{I_{C3} - 1}{11 - 1}$$

$$I_{C3} = \frac{10}{11} + 1 = 1.1 \text{ mA}$$

$$\text{o/p resistance } r_o = 101 \text{ k}\Omega$$

⇒ MAXIMUM VOLTAGE RATING -



$$BV_{CEO} < BV_{CBO}$$

$$BV_{CEO} = BV_{CBO} \sqrt[n]{\frac{1}{h_{fe}}}$$

for npn, $n=6$ $h_{fe}=50$

$$BV_{CEO} = 0.52 BV_{CBO}$$

BV_{CBO} -

Maximum reverse biasing voltage which may be applied before breakdown b/w collector and emitter base under condition that emitter lead is open circuit.

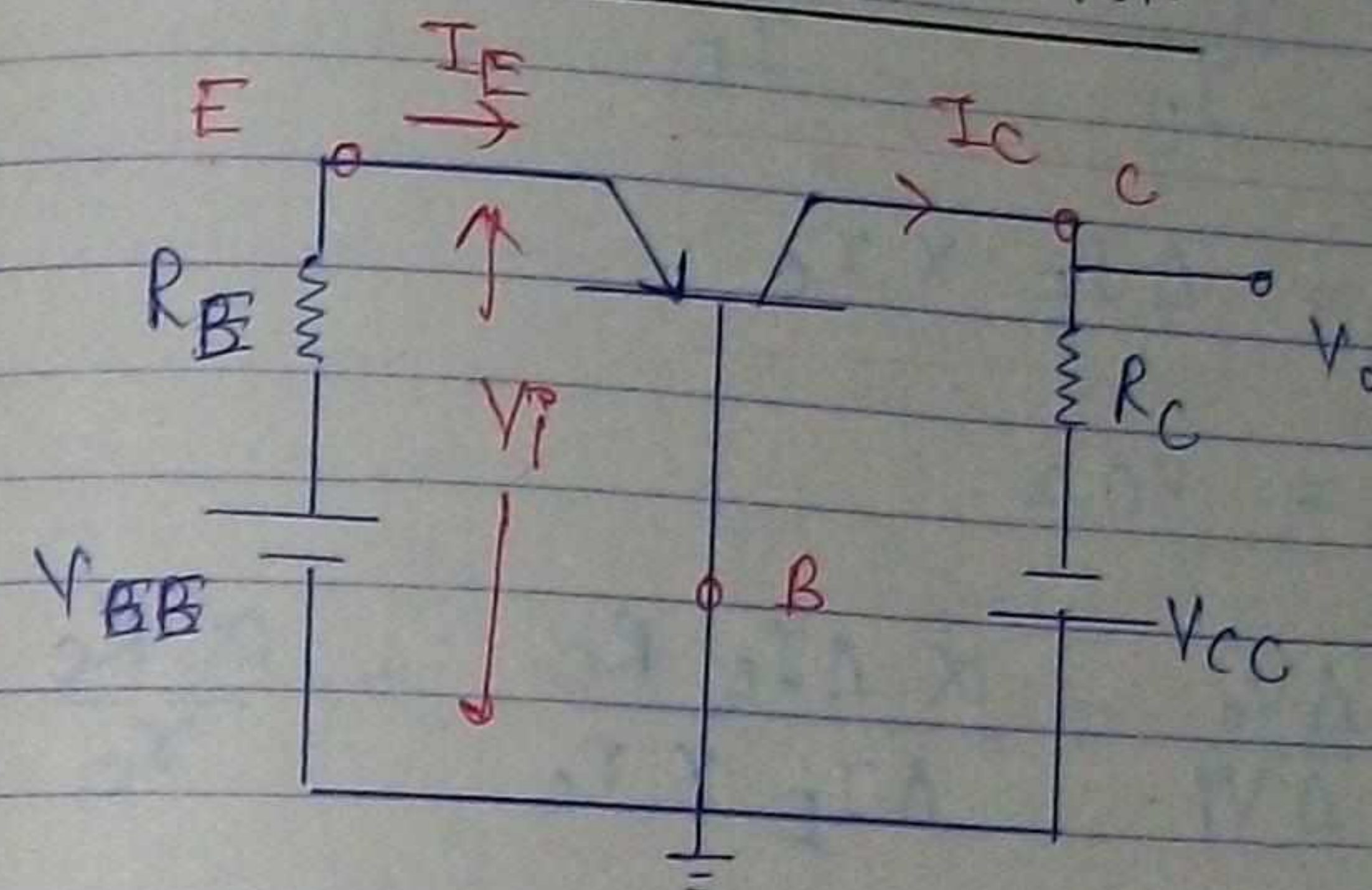
BV_{CEO} -

for common emitter config collector to emitter breakdown voltage with open circuit base

NOTE -

In a particular Tx the voltage limit is determined by punch through or Breakdown which ever occurs at lower voltage.

⇒ TRANSISTOR AS AN AMPLIFIER -



$$I_C = \alpha I_E$$

$V_i \uparrow$ by ΔV_i

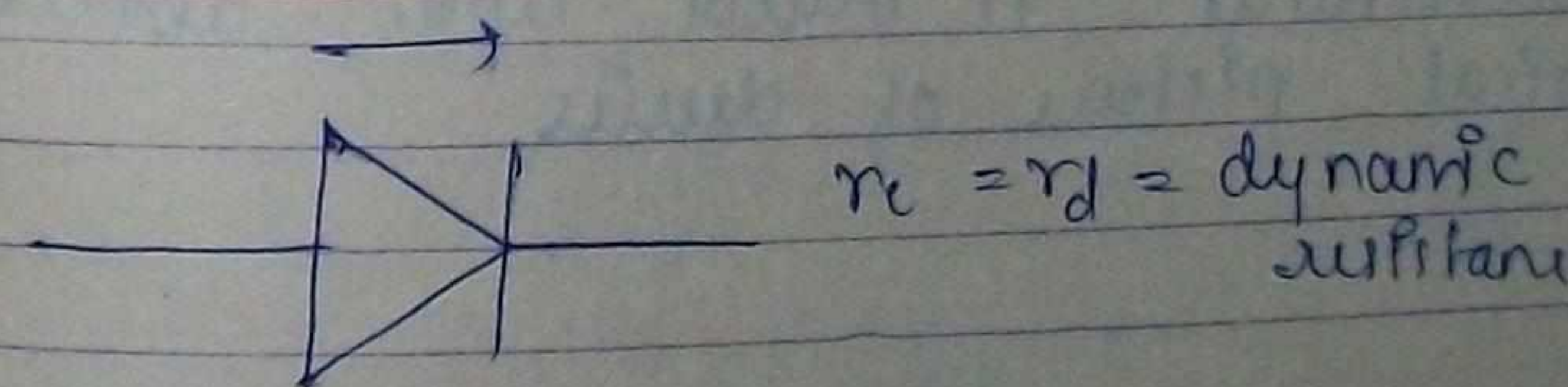
then $I_E \uparrow$ by large amount ΔI_E (exp. fⁿ)

$I_C \uparrow$ by $\Delta I_C = \alpha \Delta I_E$

$$\Delta V_o = \Delta I_C R_C = \alpha \Delta I_E R_C$$

$$A_v = \frac{\Delta V_o}{\Delta V_i} \gg 1$$

#



$$r_d = \frac{n V_T}{I_f} \approx 20 \text{ to } 100 \Omega$$

$$r_e = \frac{\eta V_T}{I_E} = \frac{V_T}{I_E}$$

$$\Delta V_i = \Delta I_E \times r_e$$

$$r_e = 50 \Omega$$

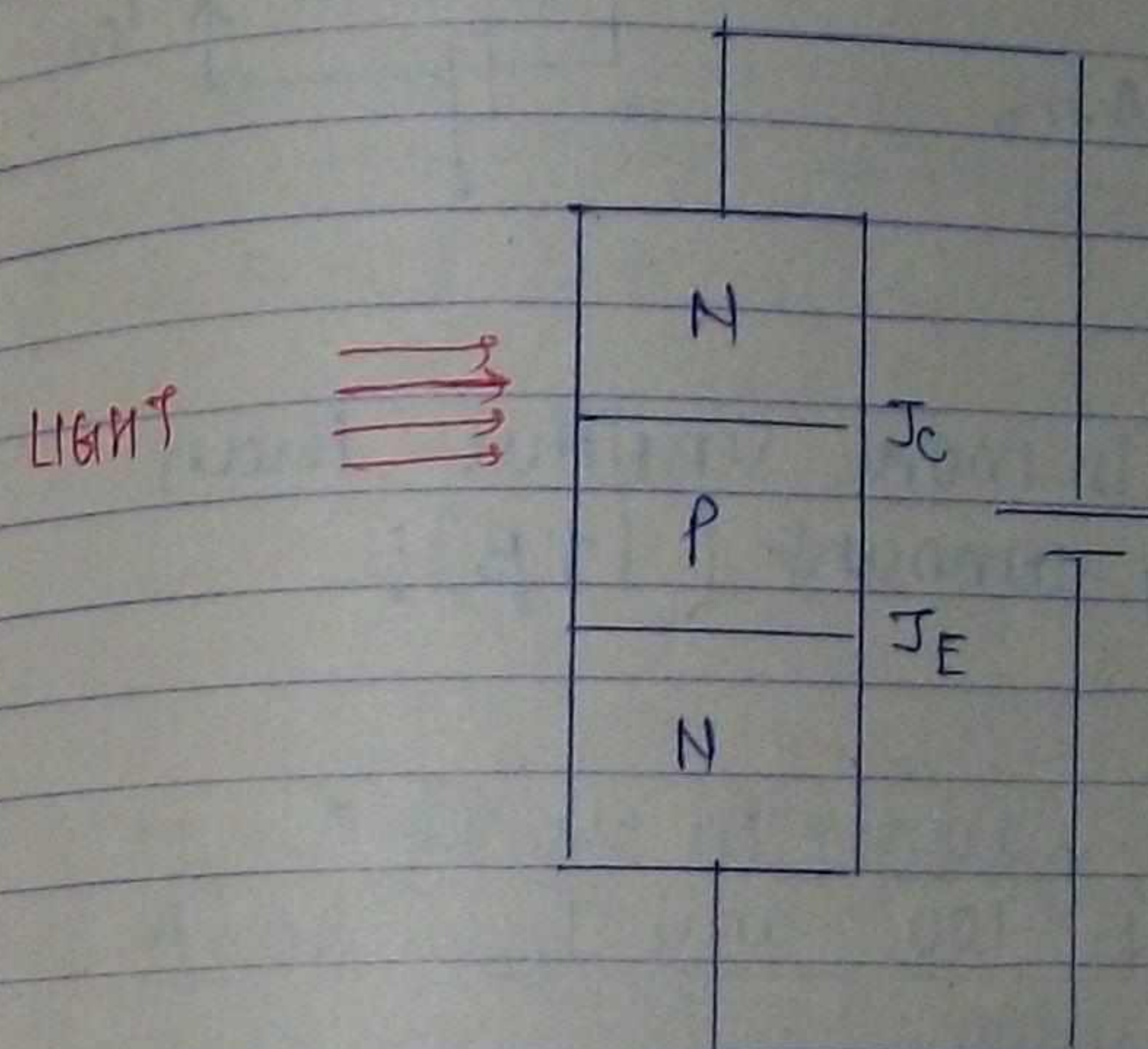
$$A_v = \frac{\Delta V_o}{\Delta V_i} = \frac{\alpha \Delta I_E R_C}{\Delta I_E \times r_e} = \frac{\alpha R_C}{r_e}$$

$$\text{let } \alpha \approx 1 \quad R_C = 3K \quad r_e = 50 \Omega$$

$$A_v = \frac{1 \times 3000}{50} = 60 > 1$$

- Tx. provides power gain as well as voltage or current amplification
- The current in low resistance i/p circuit is transferred to high resistance o/p circuit. The word transistor which originated as a contraction of 'Transfer of resistance' is based upon above physical picture of device.

⇒ PHOTO TRANSISTOR - (PHOTO VOLT DIODE)



$$I_E = I_B$$

$$I_C = \beta I_B$$

Tx is in active region

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$I_B = 0 \quad \text{since Base is open}$$

$$I_C = (1 + \beta) I_{CO}$$

Because of radiation at J_C , $I_{CO} \uparrow$ by ΔI_{CO}

$$I_C + \Delta I_C = (1 + \beta) (I_{CO} + \Delta I_{CO})$$

$$I_C + \Delta I_C = (1 + \beta) I_{CO} + (1 + \beta) \Delta I_{CO}$$

$$\Delta I_C = (1 + \beta) \Delta I_{CO}$$

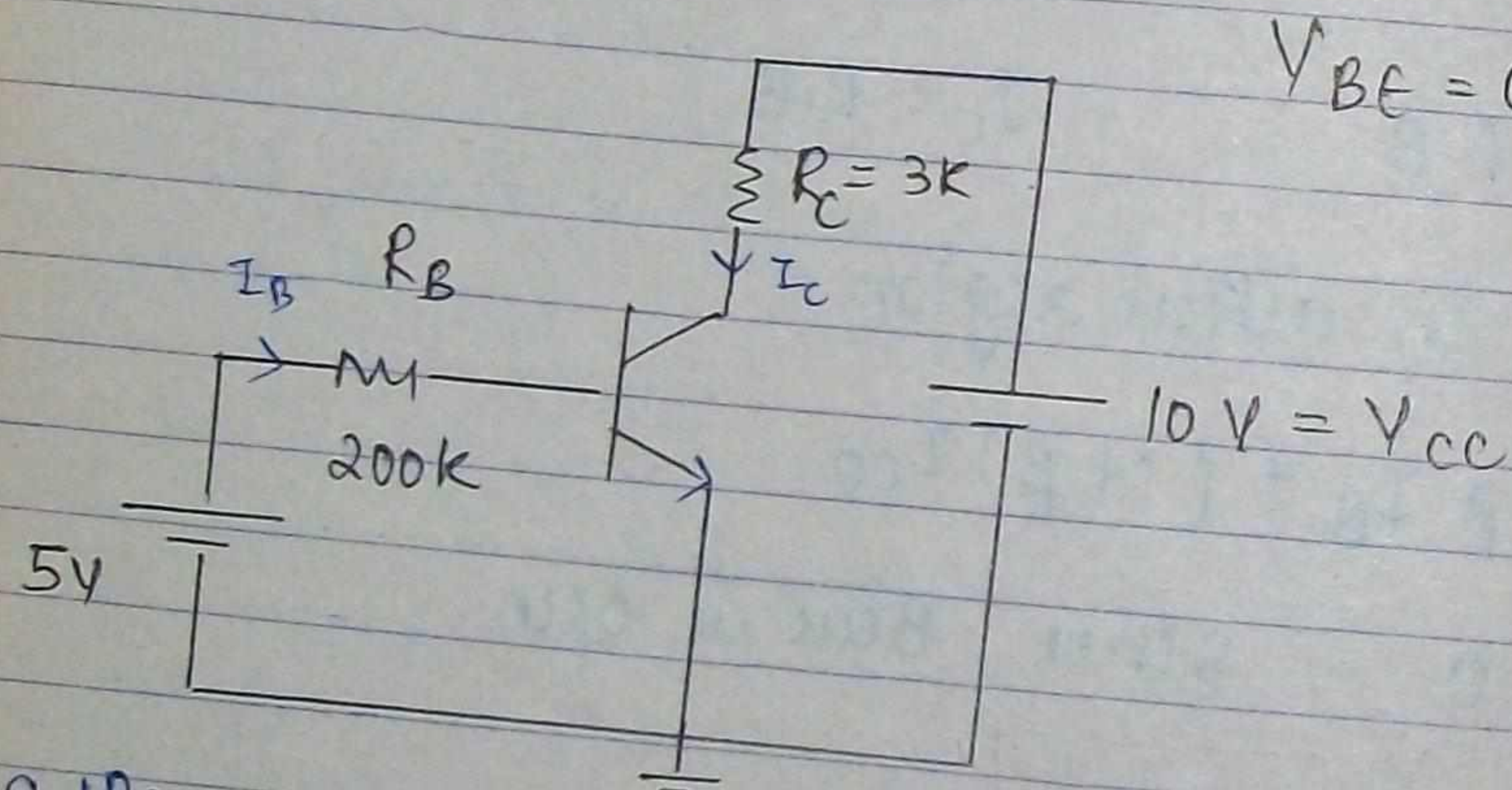
$$I = I_{CO}$$

$$I + \Delta I = I_{CO} + \Delta I_{CO}$$

$$\Delta I = \Delta I_{CO}$$

∴ Phototransistor is more sensitive than photodiode by amount $(1 + \beta)$.

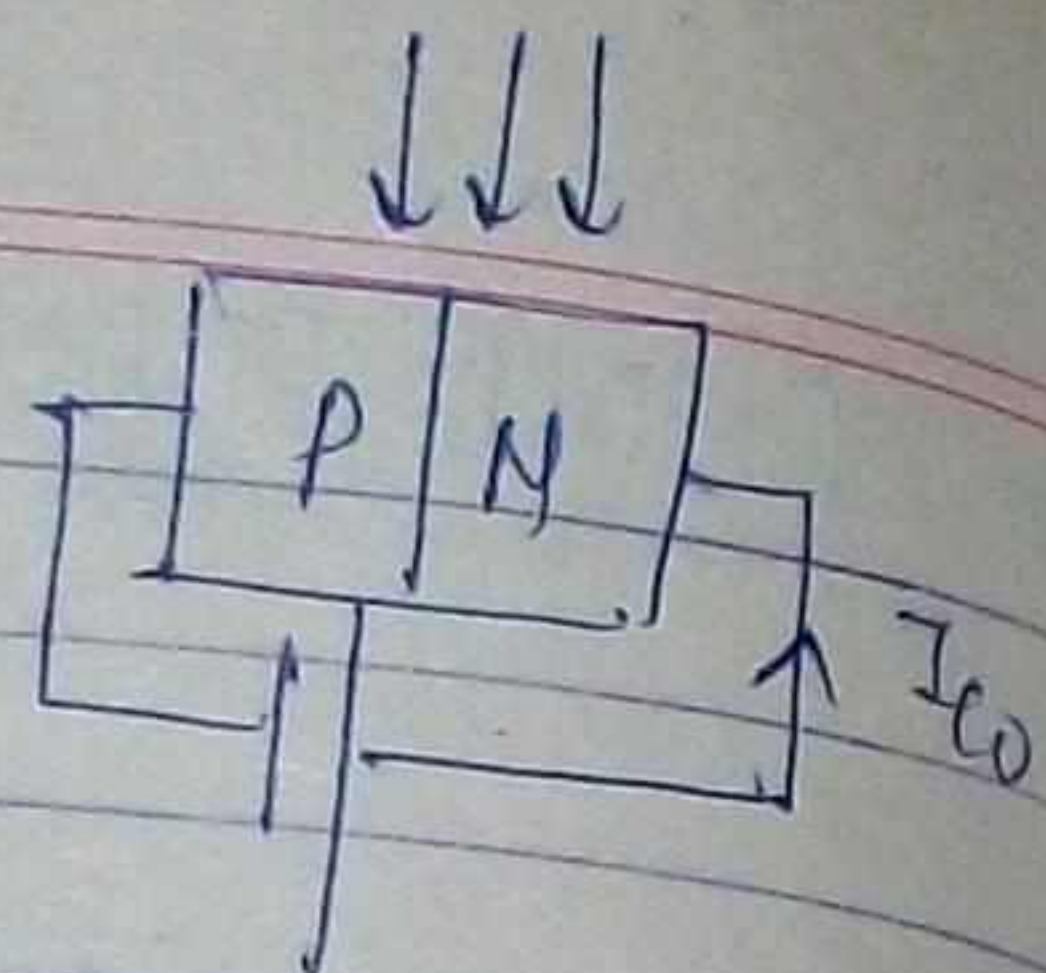
Ques. Find the Tx. current in circuit:
A si Tx with $\beta = 100$ and $I_{CO} = 20 \text{ nA}$ is under consideration.



Soln: $I_E = I_B$

∴ Tx is not in cutoff

Let Tx is in active region



TYPICAL JUNCTION VOLTAGE FOR NPN TX AT 25°C

	$V_{CE SAT}$	$V_{BE SAT} = V_0$	$V_{BE ACTIVE}$	$V_{BE CUTIN} = V_r$	$V_{BE CUTOFF}$	$V_{CE ACTIVE}$
Si	0.2	0.8	0.7	0.5	0.0	> 0.2
Ge	0.1	0.3	0.2	0.1	-0.1	> 0.1

$$I_C = \beta I_B + (1 + \beta) I_{CO} \approx \beta I_B$$

$$I_B = \frac{5 - 0.7}{200k} = \underline{\underline{0.0215 \text{ mA}}}$$

$$I_C = \beta I_B = 100 \times 0.0215 = \underline{\underline{2.15 \text{ mA}}}$$

$$\begin{aligned} V_{CE} &= V_{CC} - I_C R_C \\ &= 10 - 2.15 \times 10^{-3} \times 3 \times 10^3 \\ &= 3.55 \end{aligned}$$

* $V_{CE} > 0.2$ ∴ our assumption is right

② when $R_B = 50k$

$$I_E = I_B$$

Let Tx is in active region

$$I_B = \frac{5 - 0.7}{50k} = 0.086 \text{ mA}$$

$$I_C = \beta I_B = 100 \times 0.086 \text{ mA} = \underline{\underline{8.6 \text{ mA}}}$$

$$\begin{aligned} V_{CE} &= V_{CC} - I_C R_C \\ &= 10 - 8.6 \times 3 \\ &= -15.8 \end{aligned}$$

* $V_{CE} < 0.2$ $\therefore$ Tx is not in active mode but Tx is in saturation mode

$\therefore$ Our assumption is wrong.

Now let Tx is in saturation

$$V_{CE} = V_{CESAT} = 0.2 \quad V_{BE} = V_{BESAT} = 0.8$$

$$I_C = I_{SAT} = \frac{V_{CC} - V_{CESAT}}{R_C} = \frac{10 - 0.2}{3} = \underline{\underline{3.26 \text{ mA}}}$$

$$I_B = \frac{V_{BB} - V_{BESAT}}{R_B} = \frac{5 - 0.8}{50k} = \underline{\underline{0.084 \text{ mA}}}$$

Proof: $I_{Bmin} = \frac{I_{CSAT}}{\beta} = \frac{3.26}{100} = 0.0326 \text{ mA}$

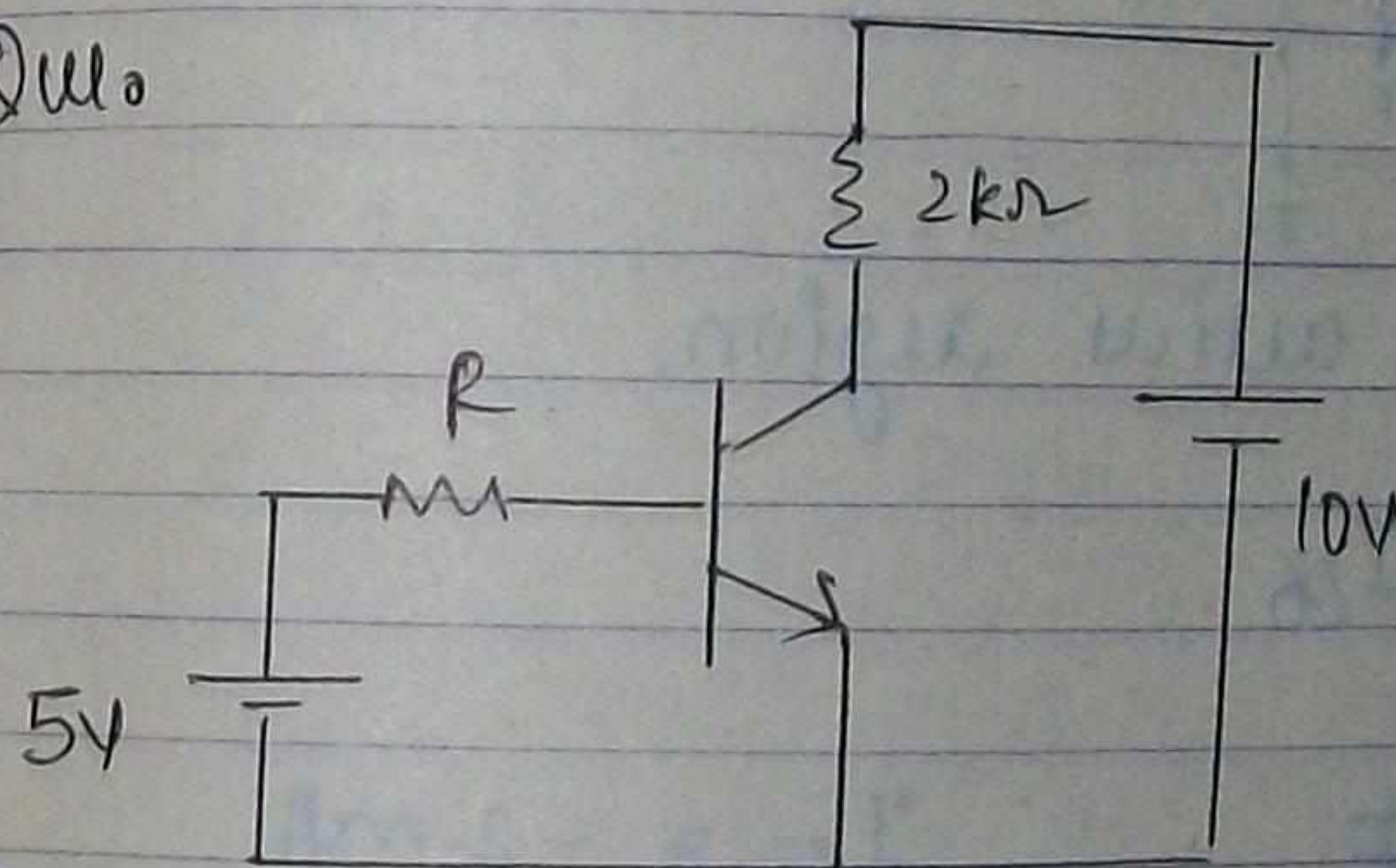
$$I_{Bmin} < I_B$$

$\therefore$ Tx is in saturation

$$\text{Overdrive factor} = \frac{I_B}{I_{Bmin}} = \frac{0.084}{0.0326} = 2.57$$

$\therefore$ Tx is in oversaturation

Ques



find R for Tx to be in saturation

$$\begin{aligned} V_{BESAT} &= 0.8 \\ V_{CESAT} &= 0.2 \\ \beta &= 50 \end{aligned}$$

Soln:

let Tx is in saturation

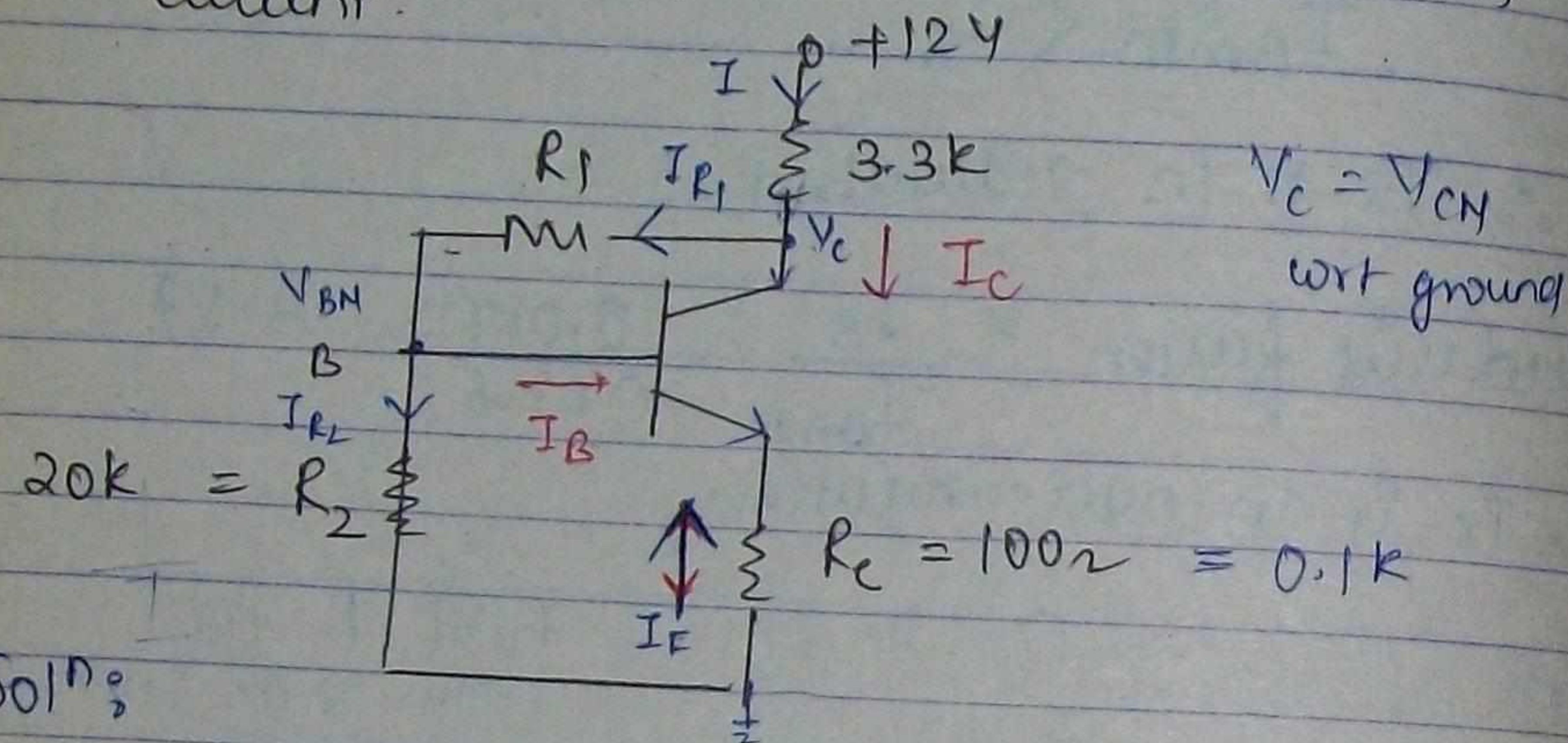
$$I_B = \frac{V_{BB} - V_{BESAT}}{R} = \frac{5 - 0.8}{R}$$

$$\text{for Sat } I_B = I_{Bmin} = \frac{I_{CSAT}}{\beta}$$

$$I_{CSAT} = \frac{V_{CC} - V_{CESAT}}{R_C} = \frac{10 - 0.2}{2} = \frac{9.8}{2} = 4.9 \text{ mA}$$

$$\frac{4.9 \text{ mA}}{50} = \frac{5 - 0.8}{R} \quad R \leq \frac{50 \times 4.2}{4.9} = \underline{\underline{42.86k\Omega}}$$

Ques. If $\alpha = 0.98$ and $V_{BE} = 0.7$ find R_1 in the circuit shown for an $I_E = -2 \text{ mA}$ neglect reverse saturation current.



Soln:

Let Tx is in active region.

$$I_C = \alpha I_E + I_{C0}$$

$$I_C = 0.98 I_E$$

$$I_E = -2 \text{ mA}$$

$$R_1 = \frac{V_{CN} - V_{BN}}{I_{R1}}$$

$$I_C = -\alpha I_E \quad (\text{due to opp dir'n})$$

$$I_C = 0.98 \times 2 \text{ mA}$$

$$I_B + I_C + I_E = 0 \quad (\text{from diagram})$$

$$I_C = 1.96 \text{ mA}$$

$$I_B = -(I_C + I_E) = -(1.96 - 2) = 0.04 \text{ mA}$$

$$V_{BN} = V_{BE} + 2 I_E R_E$$

$$= 0.7 + 2 \times 0.1 = 0.9$$

$$I_{R2} = \frac{V_{BN}}{R_2} = \frac{0.9}{20} = 0.045 \text{ mA}$$

$$I_{R1} = I_B + I_{R2} = 0.085 \text{ mA}$$

$$I = I_{R1} + I_C = 0.085 + 1.96 = 2.045$$

$$V_{CN} = 12 - 2.045 \times 3 = 5.25 \text{ V}$$

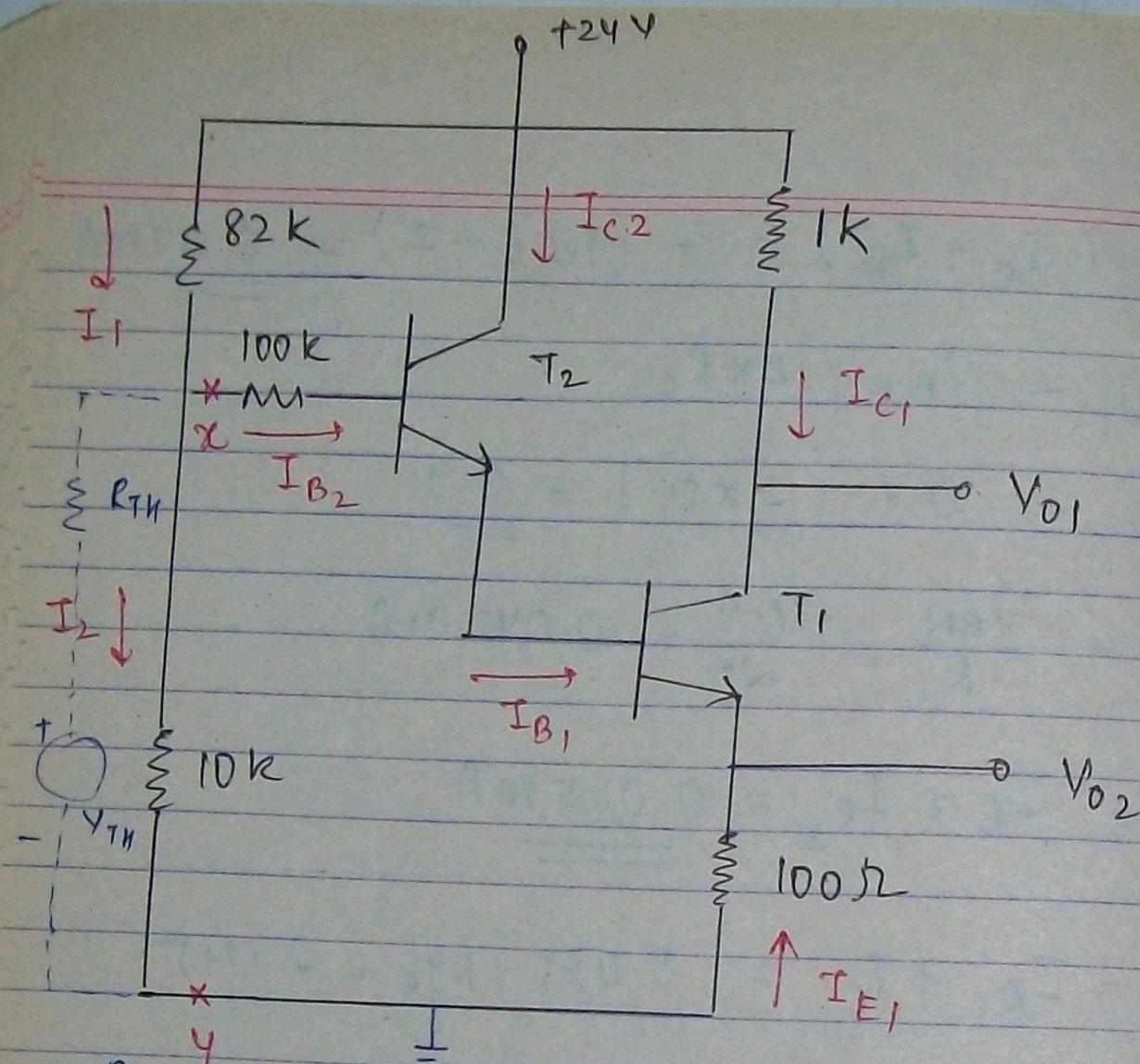
$$R_1 = \frac{5.25 - 0.9}{0.085} = 52 \text{ k}\Omega$$

Ques. for circuit shown Tx Q1 and Q2 operate in Active region with $V_{B1} = V_{B2} = 0.7 \text{ V}$. $\beta_1 = 100$ and $\beta_2 = 50$ the I_{C0} is negligible find

① $I_{B2}, I_{R1}, I_{R2}, I_{C2}, I_{B1}, I_{C1}$ and I_E

② Find V_{O1} and V_{O2} .

$$V_{B1} = V_{B2} = 0.7 \text{ V}$$

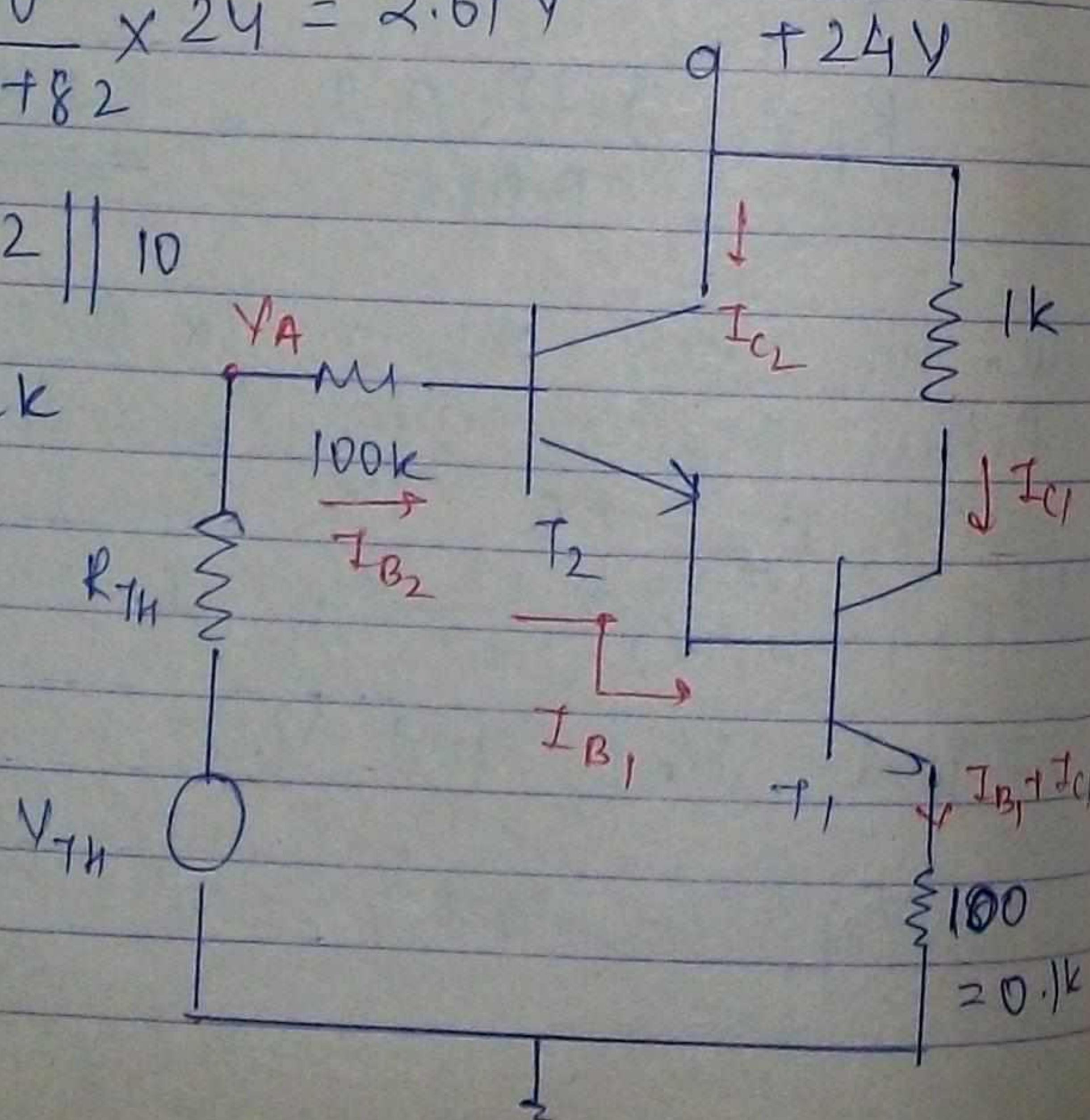


Soln:

$$V_{TH} = \frac{10}{10+82} \times 24 = 2.61 \text{ V}$$

$$R_{TH} = 82 \parallel 10$$

$$\approx 8.92 \text{ k}$$



$$I_{E1} = I_{B1} + I_{C1} = I_{B1} + \beta_1 I_{B1} = (1+\beta) I_{B1} \\ \approx 101 I_{B1}$$

$$I_{B1} = I_{B2} + I_{C2}$$

$$I_{B1} = \beta_2 I_{B2} + I_{B2} = (1+\beta) I_{B2} \\ \approx 51 I_{B2}$$

$$I_{B1} + I_{C1} = 101 I_{B1} = \underline{\underline{5151 I_{B2}}}$$

$$V_{TH} = I_{B2} (R_{TH} + 100) + V_{BE2} + V_{BE1} + 5151 I_{B2} \times 0.1 \text{ k}$$

$$I_{B2} = \underline{\underline{1.94 \mu\text{A}}}$$

$$I_{B1} = 51 I_{B2} = \underline{\underline{99 \mu\text{A}}}$$

$$I_{C1} = \beta_1 I_{B1} = \underline{\underline{9.9 \text{ mA}}}$$

$$I_{C2} = \beta_2 I_{B2} = 50 \times 1.94 = \underline{\underline{97 \mu\text{A}}}$$

$$V_{01} = 24 - I_{C1} \times 1 \text{ k} =$$

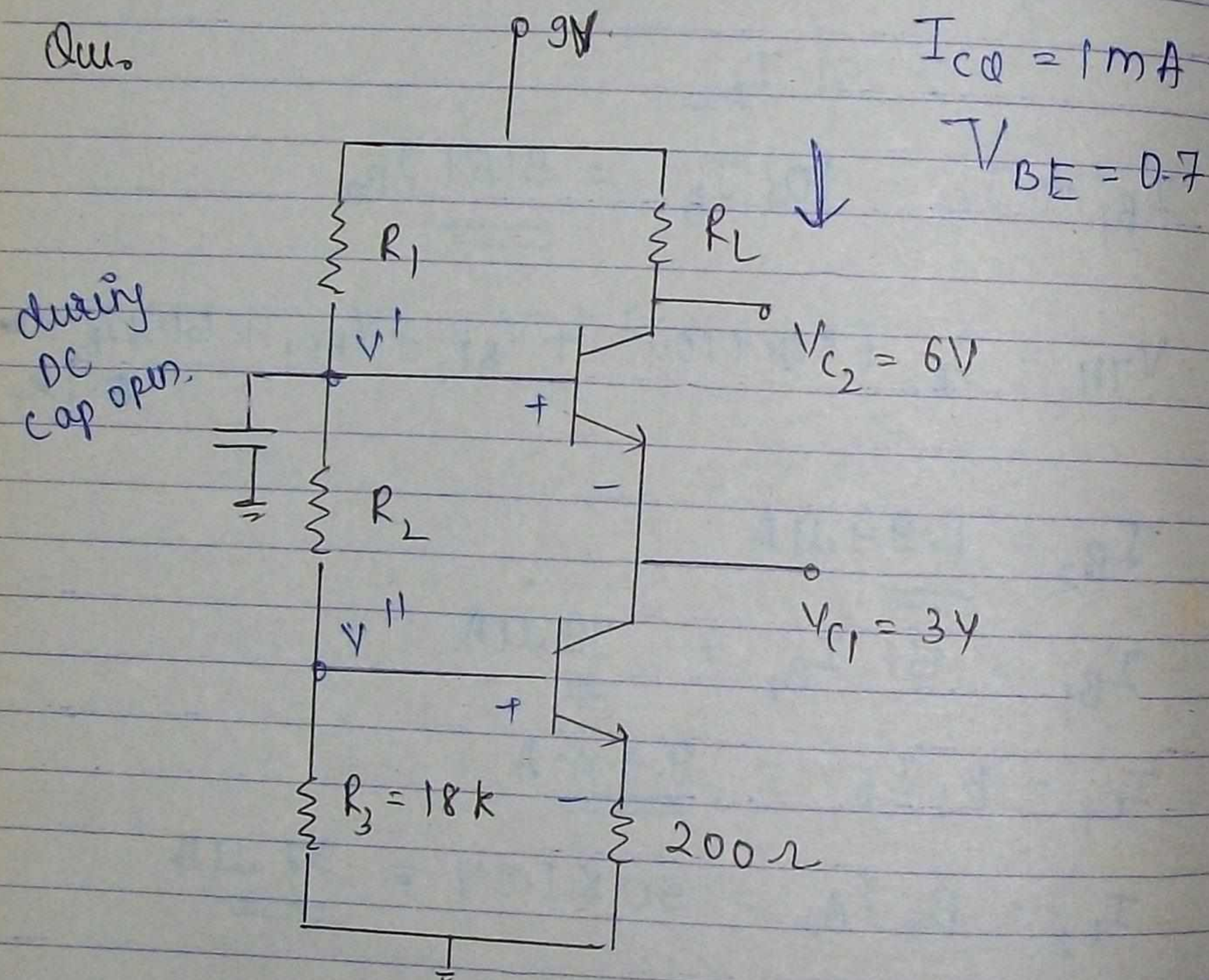
$$V_{02} = (I_{B1} + I_{C1}) \times 0.1 \text{ k} =$$

$$I_E = -(I_{B1} + I_{C1}) =$$

$$V_A = V_{TH} - I_{B2} R_{TH} = 2.59V$$

$$I_1 = \frac{24 - V_A}{82} = \underline{\underline{261 \mu A}}$$

$$I_2 = \frac{V_A}{10} = \underline{\underline{259 \mu A}}$$



Assume β of Tx are very high then calculate R_L , R_1 and R_2

$$I_B = \frac{I_C}{\beta} \approx 0$$

$$R_L = \frac{9 - 6}{I_{CQ}} = \underline{\underline{3k\Omega}}$$

$$V' = 0.7 + 3 = 3.7$$

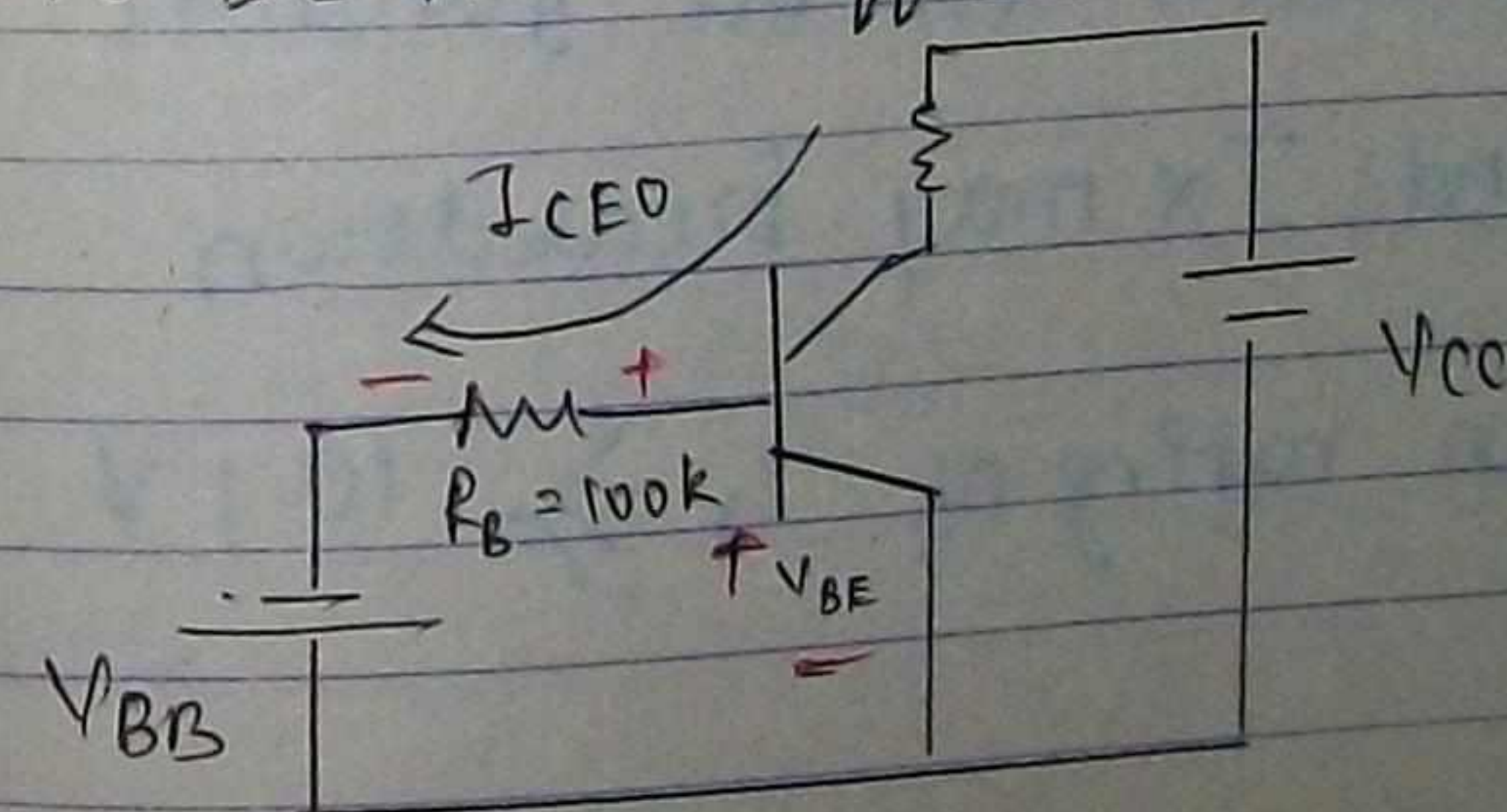
$$V'' = 0.7 + 1 \times 10^{-3} \times 200 = \underline{\underline{0.9}}$$

$$I = \frac{V''}{R_3} = \frac{0.9}{18} = \underline{\underline{0.05mA}}$$

$$R_2 = \frac{V' - V''}{I} = \frac{3.7 - 0.9}{0.05} = \underline{\underline{56k\Omega}}$$

$$R_1 = \frac{9 - V'}{I} = \frac{9 - 3.7}{0.05} = \underline{\underline{106k\Omega}}$$

Ques. If $R_B = 100k$ and I_{CBO} may become as large as $100\mu A$ calculate V_{BB} for Tx to be in cutoff assume Ge. Tx.



for Ge $V_{BE} \leq -0.1$ for cutoff

when $I_{CBO} = 0$ (does not exist)

$$V_{BE} = -V_{BB} \quad (\text{Ideally})$$

when I_{CBO} have exist

$$-V_{BB} = -I_{CBO} R_B + V_{BE}$$

$$V_{BE} = -V_{BB} + I_{CBO} R_B \leq -0.1$$

$$-V_{BB} + 100 \times 10^{-3} \times 100 \leq -0.1$$

$$-V_{BB} \leq -0.1 - 10$$

$$V_{BB} \geq 10.1$$

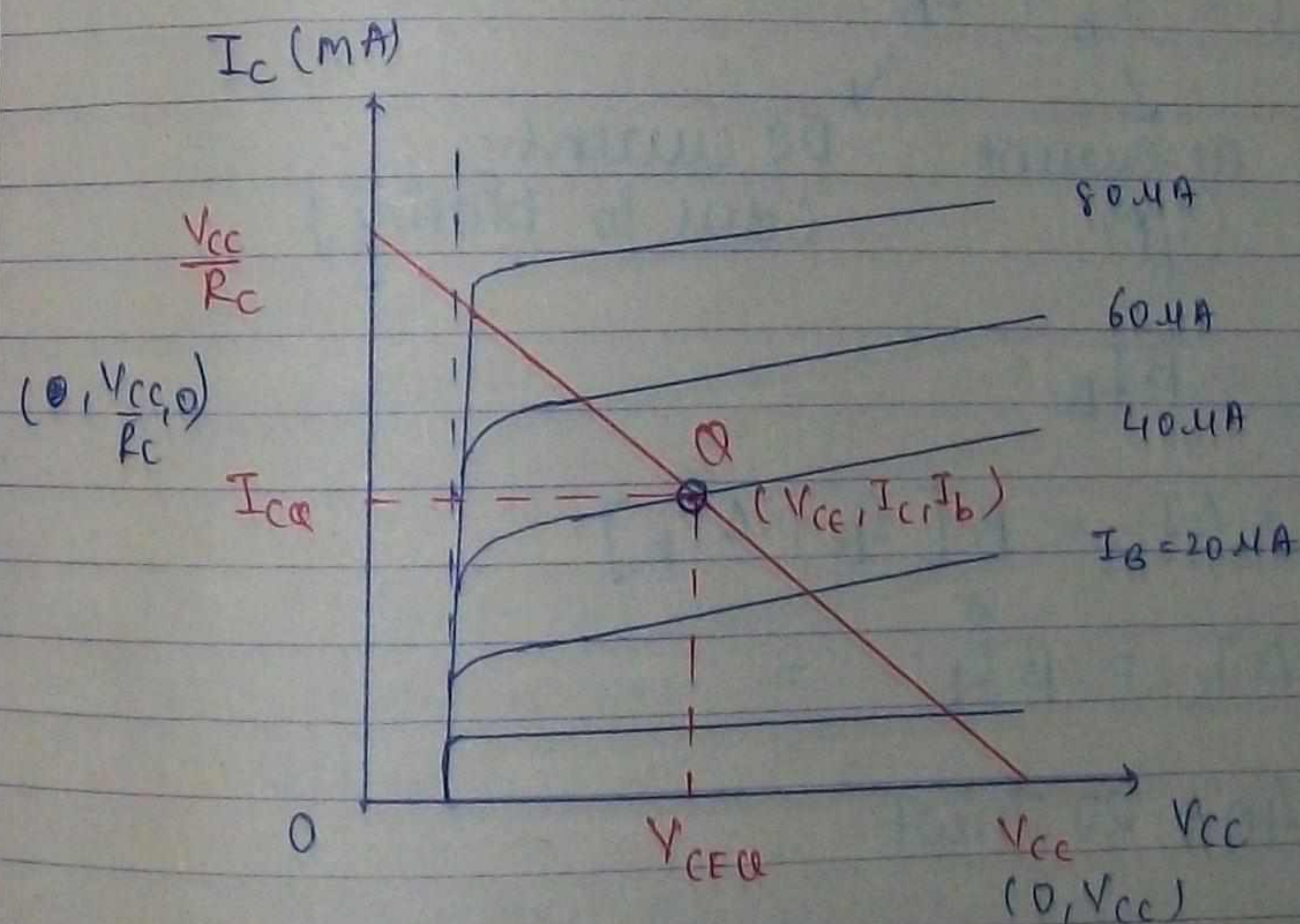
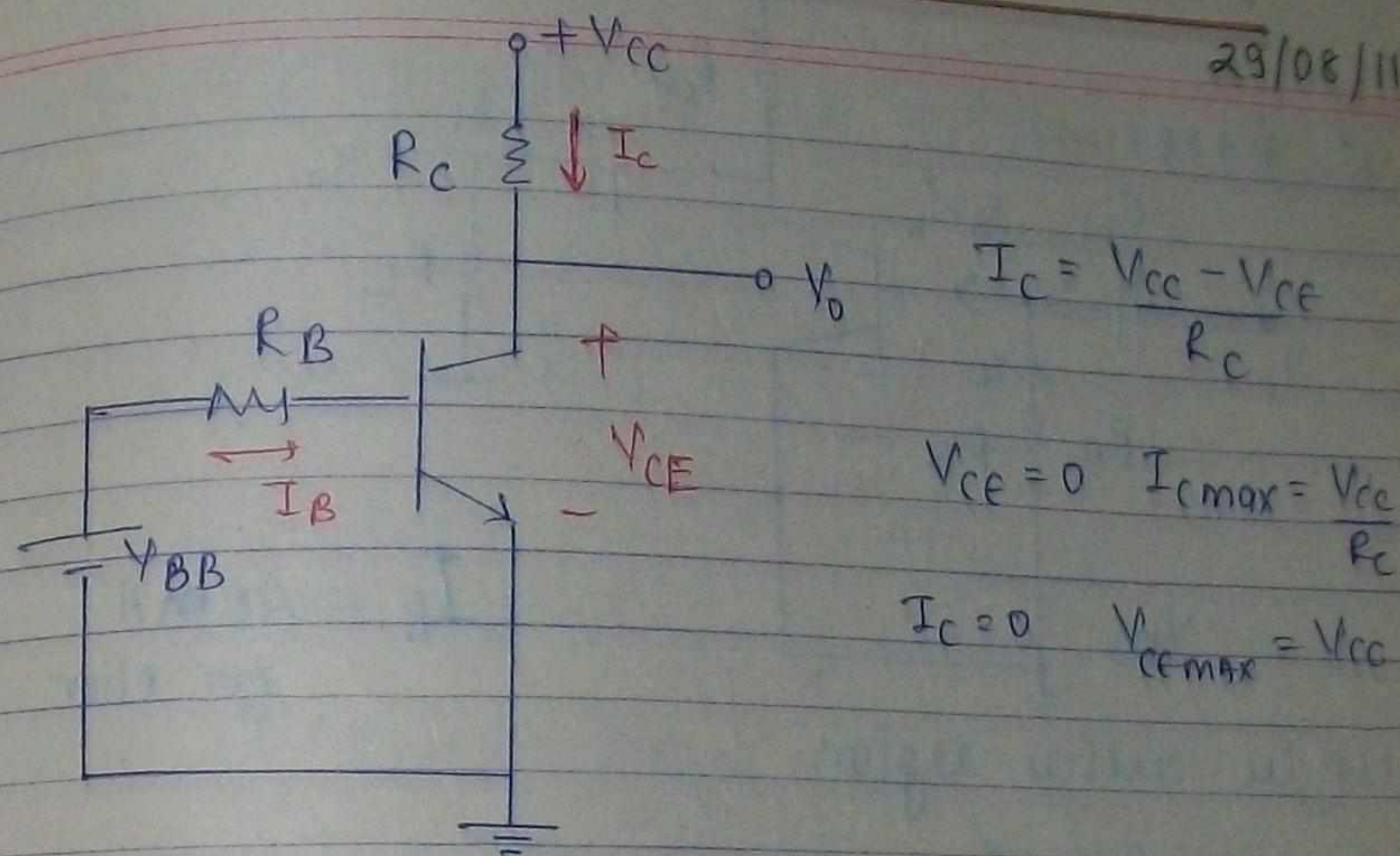
If $I_{CBO} = 0$ then $V_{BE} = -V_{BB} = -10.1$

high voltage across emitter junction
in R.B and Tx may breakdown

So Breakdown rating of Tr ≥ 10.1 V.

TRANSISTOR BIASING & STABILIZATION

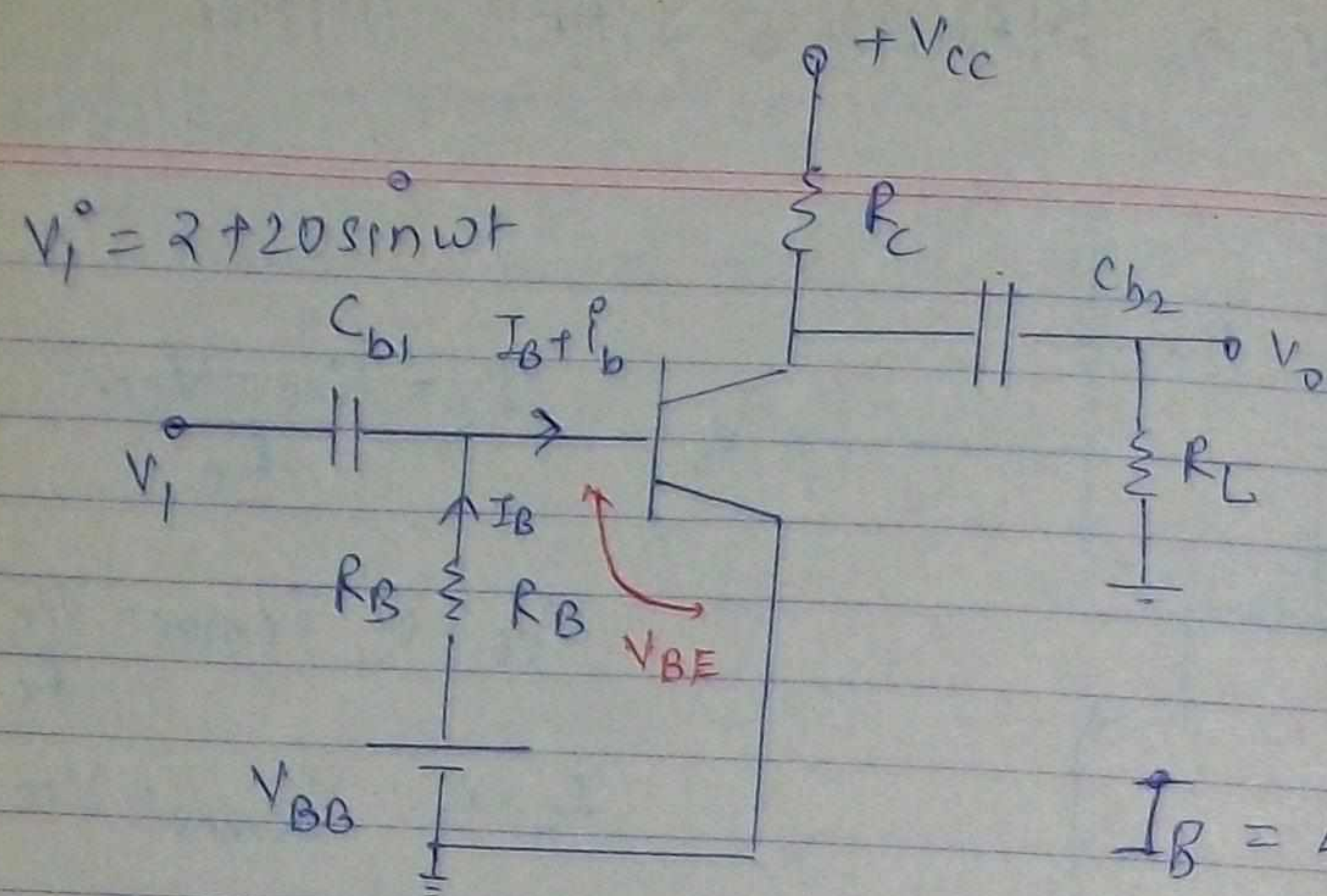
29/08/11



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

$$\text{slope} = \frac{y_2 - y_1}{x_2 - x_1} = \frac{0 - V_{CC}/R_C}{V_{CC} - 0} = -\frac{1}{R_C}$$

$$\text{SLOPE} = -1/R_C$$



under active region

$$i = i_b + I_B$$

ac current
(i/p)

DC current
(due to biasing)

$$I_C = \beta I_B$$

$$I_C + \Delta I_C = \beta [I_B + \Delta i_b]$$

$$\Delta I_C = \beta i_b$$

$$i = 40 + 20 \sin \omega t$$

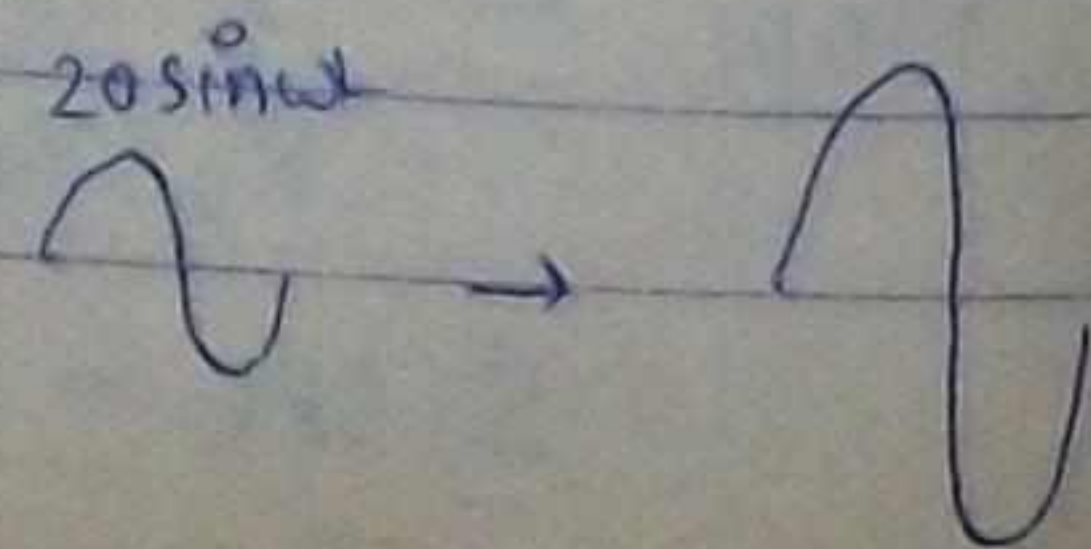
$$I_{B \max} = 60$$

(active)

$$I_{B \min} = 20 \mu A$$

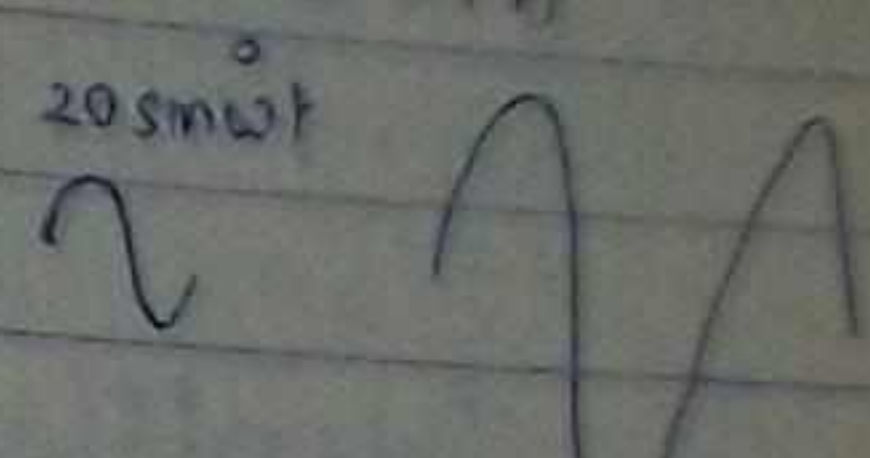
active

both
can be
amplified



but when Q point change to $I_B = 20 \mu A$

$$i = 20 \mu A + 20 \sin \omega t$$



$$I_{B \max} = 40$$

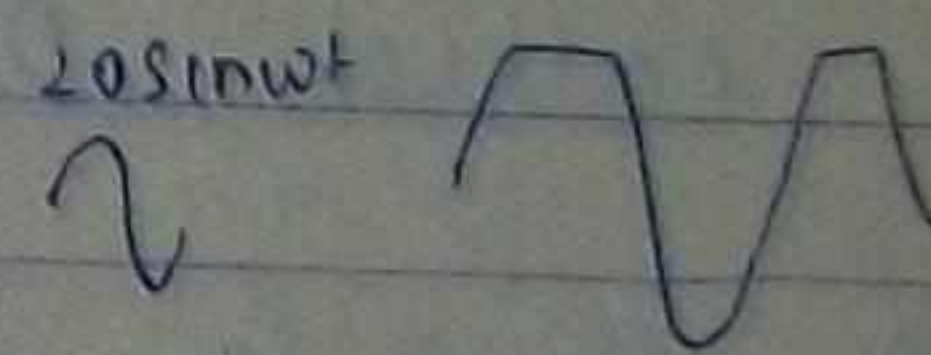
amplified
(active region)

$$I_{B \min} = 0$$

distorted
(cutoff mode)

when Q point change to $I_B = 60 \mu A$

$$i = 60 \mu A + 20 \sin \omega t$$



$$I_{B \max} = 80$$

Distorted
(saturation region)

$$I_{B \min} = 40$$

amplified
(active region)

- The output characteristics of Tx. are divided into three regions (1) saturation (2) Active (3) cutoff.

- Tx can work as an amplifier only when operated in active region. Tx can work as a switch when operated in saturation region and cutoff region. i.e. extreme ends of characteristics.

⇒ PROCEDURE TO PLOT DC LOAD LINE AND Q-POINT -

- ① Identify the value of V_{CC} in given circuit and locate this point on given characteristic.
- ② Find I_{Cmax} and locate this point. Draw a straight line joining I_{Cmax} and V_{CC} , and this straight line is called DC load line.
- ③ Find the operating values (I_B , I_C & V_{CE}) and locate this value on the given characteristics.
- ④ Project the three operating values on the DC load line and intersecting point is called Q-point.
- ⑤ Slope of DC load line is $-1/R_C$. Hence the Tx. is said to be under quiescent condition when zero i/p signal is applied.
- ⑥ Q point is a funn of (I_B, I_C, V_{CE})
- ⑦ The Tx can work as an amplifier when Q point is with in active region. but

Q point is temp. sensitive i.e. when temp. increases I_C increases. $V_{CE} \downarrow$ so that the q-point will be moving towards saturation region & Tx will be temporary stopped working as an amplifier.

- ⑧ The Tx. will provide max. power gain or max. amplification when the Q point is in the middle of DC load line such that $V_{CE} = V_{CC}/2$
- ⑨ for a stable circuit, variation in Q point due to temp. must be small
- ⑩ If the shape of o/p signal differs from i/p signal, it is said to be distorted.

⇒ SOURCE OF INSTABILITY -

Stability will be affected due to:

① THERMAL INSTABILITY -

(A) I_{CO} - Double for every $10^\circ C$ rise in temp.

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

(B) V_{BE} - which decreases at the rate of $2.3 mV/^\circ C$ i.e. $\frac{dV_{BE}}{dT} = -2.3 mV/^\circ C$

(c) β - β increases with increase in temp.

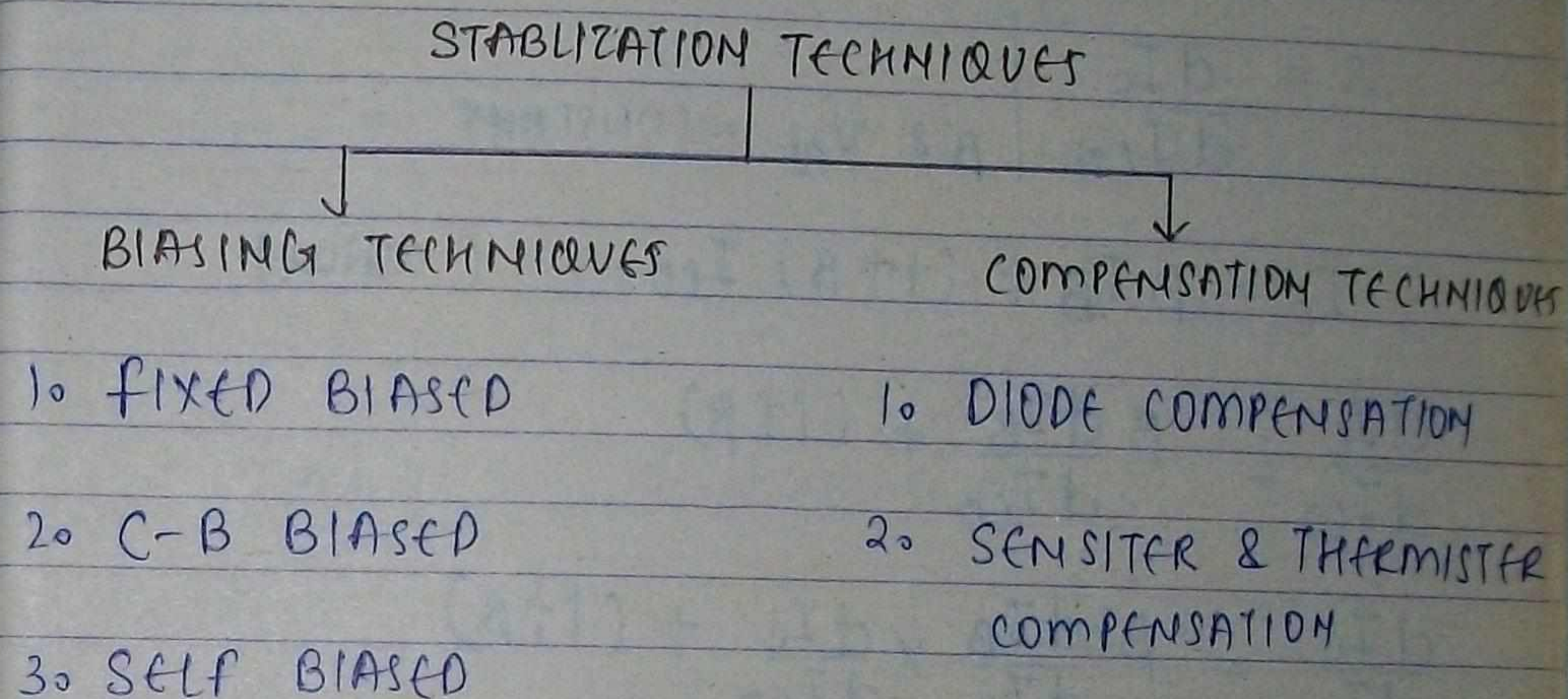
(2) REPLACEMENT OF TRANSISTOR -

β is highly affected due to replacement of Transistor.

spacing of O/P characteristic will increase or decrease for equal change in I_B as β increases or decreases.

29/08/11

⇒ STABILIZATION TECHNIQUES -



⇒ STABILITY FACTOR (S) -

$$S = \left. \frac{dI_c}{dI_{co}} \right|_{B \text{ \& } V_{BE} \text{ CONSTANT}} \quad (+ve)$$

$$S' = S_v = \left. \frac{dI_c}{dV_{BE}} \right|_{B \text{ \& } I_{co} \text{ CONSTANT}} \quad (-ve)$$

$$S'' = S_B = \left. \frac{dI_c}{d\beta} \right|_{I_{co} \text{ \& } V_{BE} \text{ CONSTANT}} \quad (+ve)$$

Since, $I_c = \beta I_B + (1 + \beta) I_{co}$

STABILITY FACTOR (S) -

$$S = \frac{dI_c}{dI_{co}} \bigg|_{\beta \text{ \& } V_{BE} = \text{CONSTANT}}$$

$$I_c = \beta I_B + (1 + \beta) I_{co} \quad (\text{active})$$

$$\frac{dI_c}{dI_{co}} = \beta \frac{dI_B}{dI_{co}} + (1 + \beta)$$

$$\frac{dI_c}{dI_{co}} = \beta \frac{dI_B}{dI_c} \times \frac{dI_c}{dI_{co}} + (1 + \beta)$$

$$\frac{dI_c}{dI_{co}} \left[1 - \beta \frac{dI_B}{dI_c} \right] = (1 + \beta)$$

$$S = \frac{dI_c}{dI_{co}} = \frac{(1 + \beta)}{1 - \beta \frac{dI_B}{dI_c}}$$

$$\# \quad I_c = \beta I_B + (1 + \beta) I_{co}$$

$$T \uparrow, (I_{co} \uparrow, V_{BE} \downarrow, \beta \uparrow), I_c \uparrow, V_{ce} \downarrow$$

$$\uparrow I_B \downarrow I_c \downarrow$$

The Rise in I_c is compensated.

① when $I_B = \text{constant}$, $\frac{dI_B}{dI_c} = 0$

$$S_{\max} = 1 + \beta \quad (\text{circuit is unstable})$$

② when $\frac{dI_B}{dI_c} = -1$, $I_c \uparrow$, $I_B \downarrow$

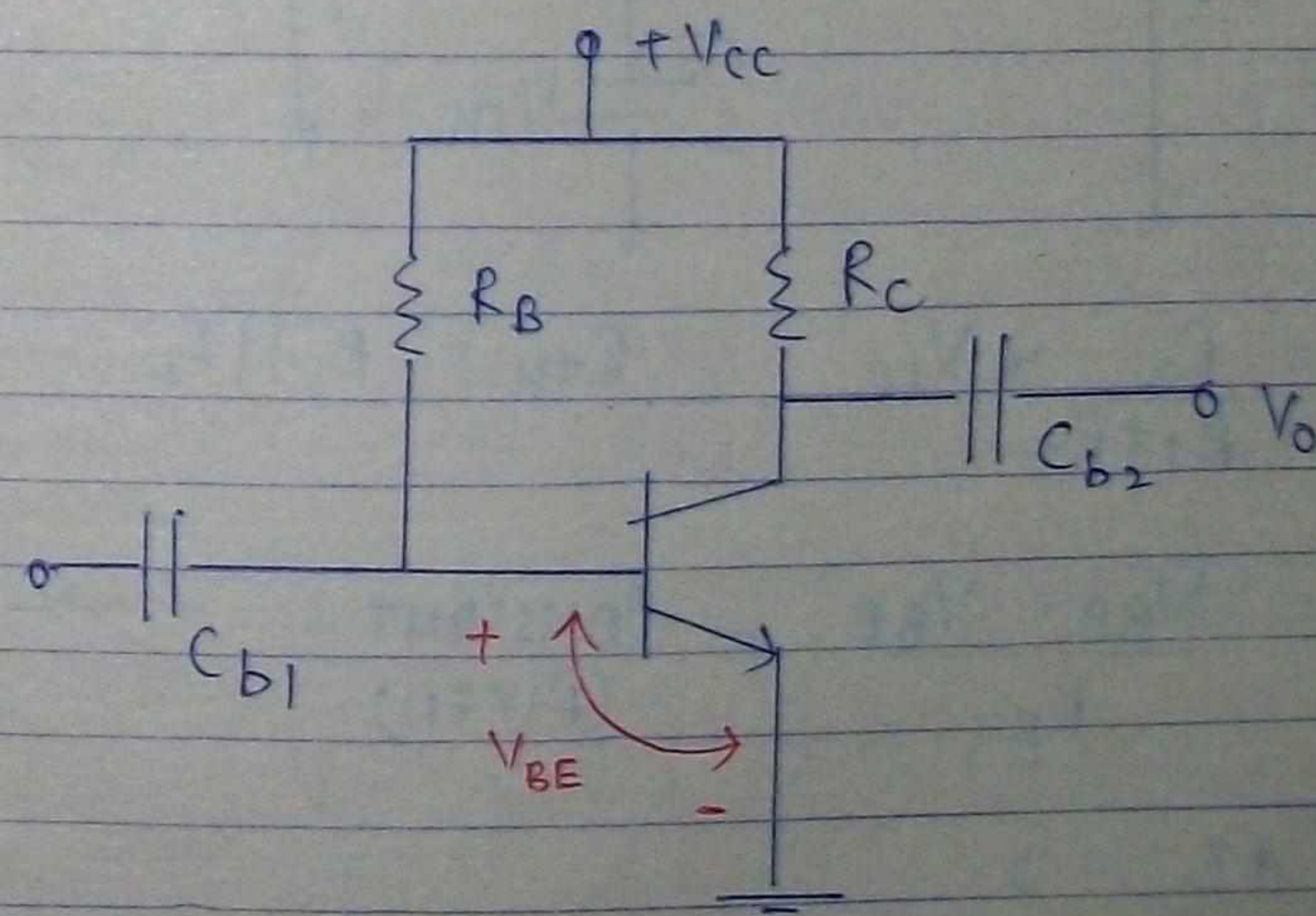
$$S_{\min} = 1 \quad (\text{circuit is stable})$$

Range:

$$S = 1 \text{ to } 1 + \beta$$

$S < 1 + \beta$ for circuit to be stable.

$\Rightarrow$ FIXED BIASED CIRCUIT -

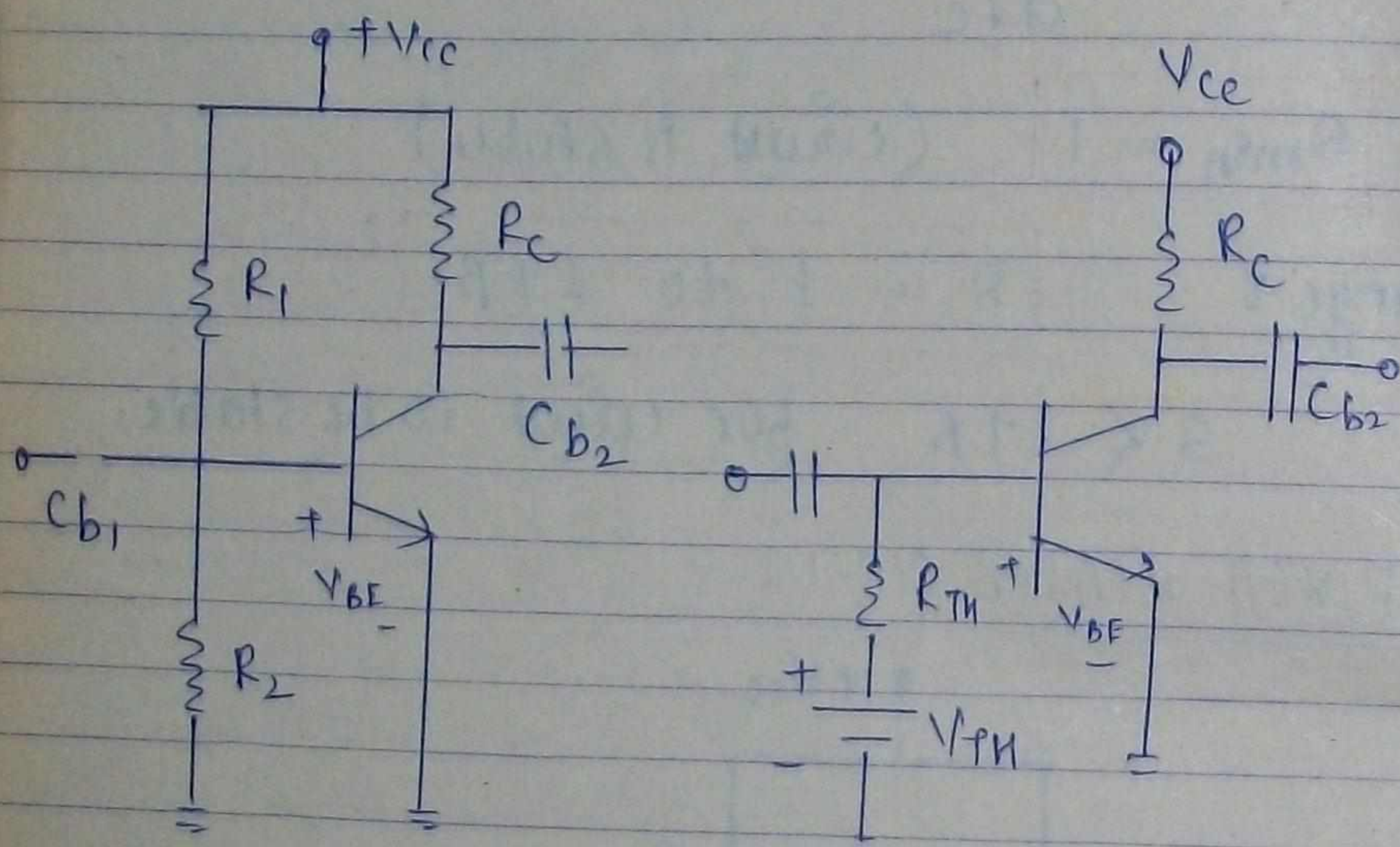


C_{b1} and C_{b2} acts as open circuit for dc
 $Z_c = \frac{1}{2\pi f C}$

$$I_B = \frac{V_{CC} - V_{BE}}{R_B} = \text{CONSTANT (FIXED)}$$

$$\frac{dI_B}{dI_C} = 0$$

$$S = 1 + \beta \quad (\text{unstable}) \quad S_{\text{max}}$$



$$V_{TH} = \frac{R_2}{R_1 + R_2} \times V_{CC}$$

$$R_{TH} = R_1 \parallel R_2$$

$$I_B = \frac{V_{BB} - V_{BE}}{R_B} = \text{CONSTANT (FIXED)}$$

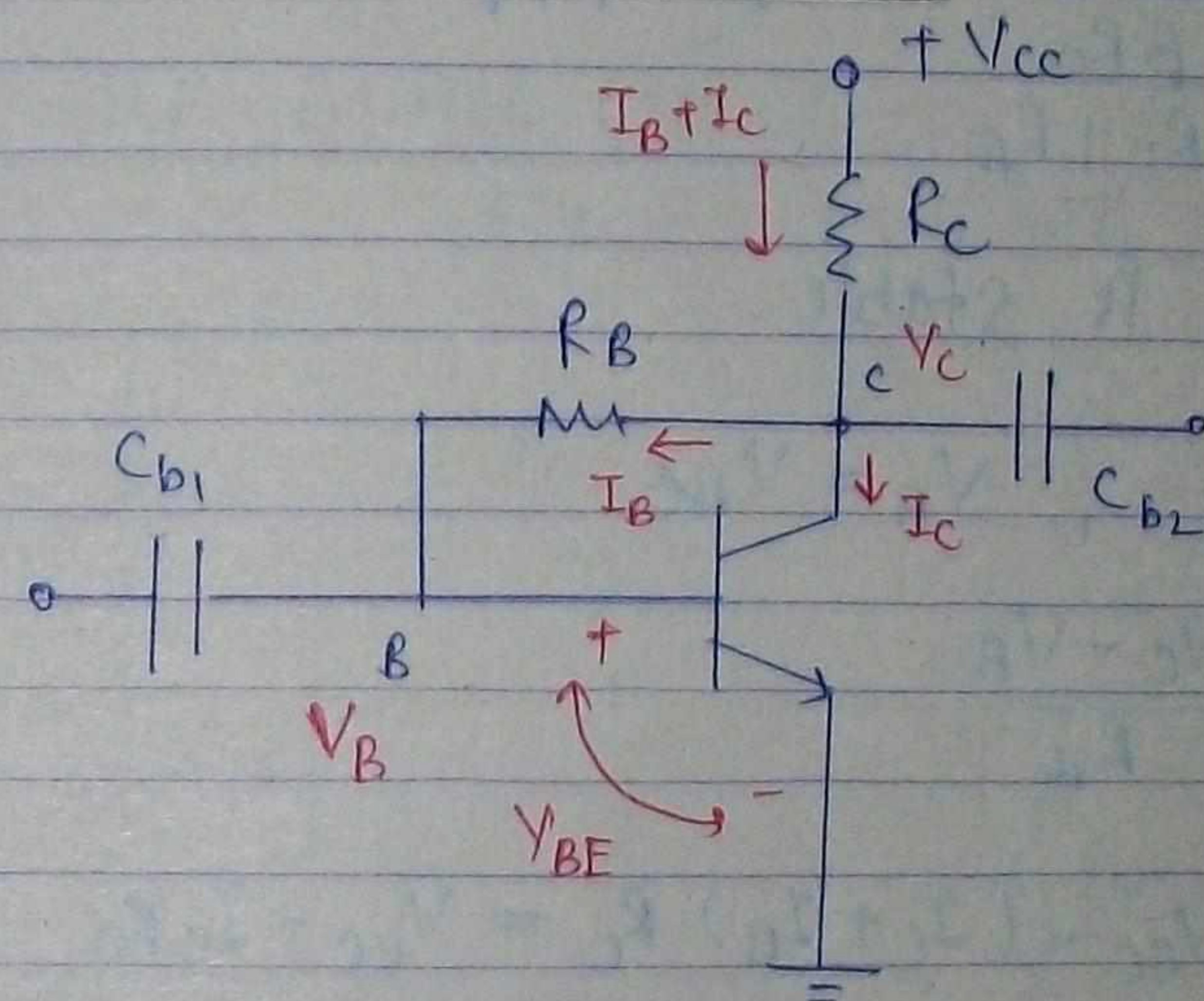
$$dI_B/dI_C = 0$$

$$S = S_{\text{max}} = (1 + \beta)$$

(UNSTABLE)

Since I_B is fixed (constant) hence it is called fixed biased circuit.

⇒ COLLECTOR-BASE BIASED -



$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$S = \frac{(1 + \beta)}{1 - \beta \frac{dI_B}{dI_C}} \quad \text{--- (1)}$$

$$V_{CC} = (I_B + I_C) R_C + I_B R_B + V_{BE}$$

$$V_{CC} = I_B (R_C + R_B) + I_C R_C + V_{BE}$$

diff. wrt I_C

$$0 = \frac{dI_B}{dI_C} (R_C + R_B) + R_C + 0$$

VOLTAGE
SHUNT
FEEDBACK

$$\frac{dI_B}{dI_C} = - \frac{R_C}{R_C + R_B}$$

$$S = \frac{1 + \beta}{1 + \frac{\beta R_C}{R_C + R_B}} < 1 + \beta$$

∴ output is stable.

$$\# \quad V_C = V_{CE}, \quad V_B = V_{BE}$$

$$I_B = \frac{V_C - V_B}{R_B}$$

$$V_C = V_{CE} = V_{CC} - (I_C + I_B) R_C = V_{CC} - I_C R_C$$

$$\text{Since } I_B + I_C = \frac{I_C}{\beta} + I_C = I_C \left(1 + \frac{1}{\beta}\right) \approx I_C$$

$$T \uparrow, [I_{CO} \uparrow, V_{BE} \downarrow, \beta \uparrow] \quad I_C \uparrow$$

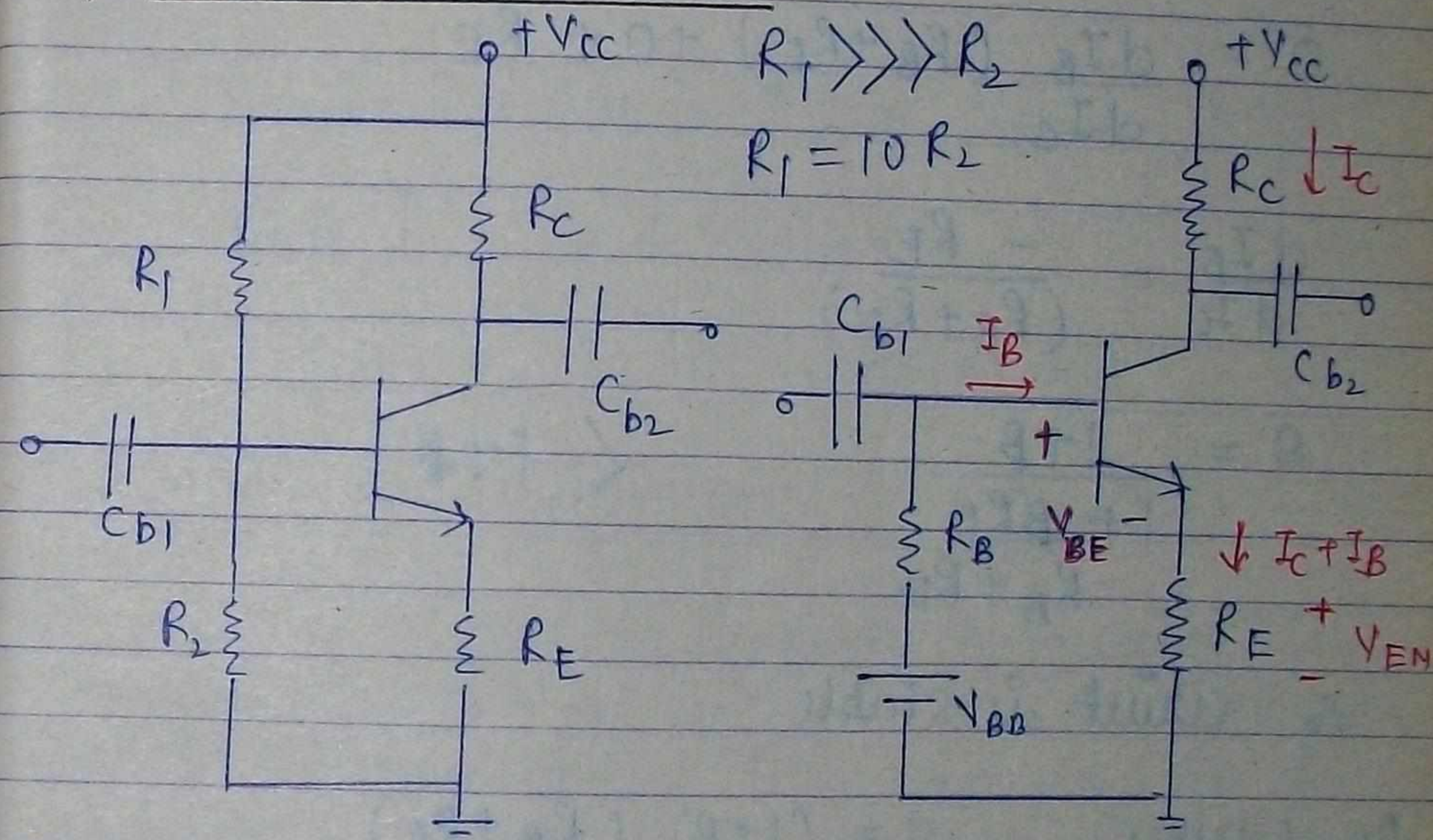
$$\text{If } V_C \downarrow \text{ due to } I_C \uparrow, \quad I_B \downarrow \quad I_C \downarrow$$

Rise in I_C is compensated.

- The feedback mechanism is Voltage Shunt FB therefore R_i and R_o will decrease.

- The function of R_B is to provide base current into Tx. But it introduces unwanted -ve f.B. therefore output is regulated in practice.

⇒ SELF-BIASED CIRCUIT -



$$V_{TH} = V_{BB} = \frac{R_2 V_{CC}}{R_1 + R_2}, \quad R_{TH} = R_B = R_1 || R_2$$

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$S = \frac{dI_C}{dI_{CO}} = \frac{1 + \beta}{1 - \beta \frac{dI_B}{dI_C}} \quad \text{--- (1)}$$

$$V_{BB} = I_B R_B + V_{BE} + (I_C + I_B) R_E$$

$$V_{BB} = I_B (R_B + R_E) + V_{BE} + I_C R_E$$

diff. wrt I_C

$$0 = \frac{dI_B}{dI_C} (R_B + R_E) + 0 + R_E$$

$$\frac{dI_B}{dI_C} = \frac{-R_E}{(R_B + R_E)}$$

$$S = \frac{1 + \beta}{1 + \beta R_E / (R_B + R_E)} < 1 + \beta$$

∴ circuit is stable

On further solving,
$$S = \frac{(1 + \beta) (R_B + R_E)}{R_B + R_E + \beta R_E}$$

$$S = (1 + \beta) \frac{(R_B + R_E)}{R_B + (1 + \beta) R_E}$$

$$(1 + \beta) R_E \gg R_B$$

$$S = \frac{R_B + R_E}{R_E} = 1 + \frac{R_B}{R_E}$$

$$\# V_{EN} = [I_C + I_B] R_E \approx I_C R_E \quad (B \gg 1)$$

$$I_B = \frac{V_{BB} - V_{BE} - V_{EN}}{R_B}$$

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$T \uparrow, I_C \uparrow$ (due to $\uparrow I_{CO}, V_{BE} \downarrow, \beta \uparrow$)

$V_{EN} \uparrow, I_B \downarrow, I_C \downarrow$

∴ Rise in I_C is compensated.

REASON FOR IMPROVEMENT IN STABILITY -

- If I_C tends to increase because of change in I_{CO}, V_{BE} and β , due to variation in temperature, the current in R_E increases. As a consequence of increase in voltage drop across R_E , I_B decreases. Therefore increase in I_C is compensated.

- It is also called as
POTENTIAL DIVIDER BIAS CIRCUIT

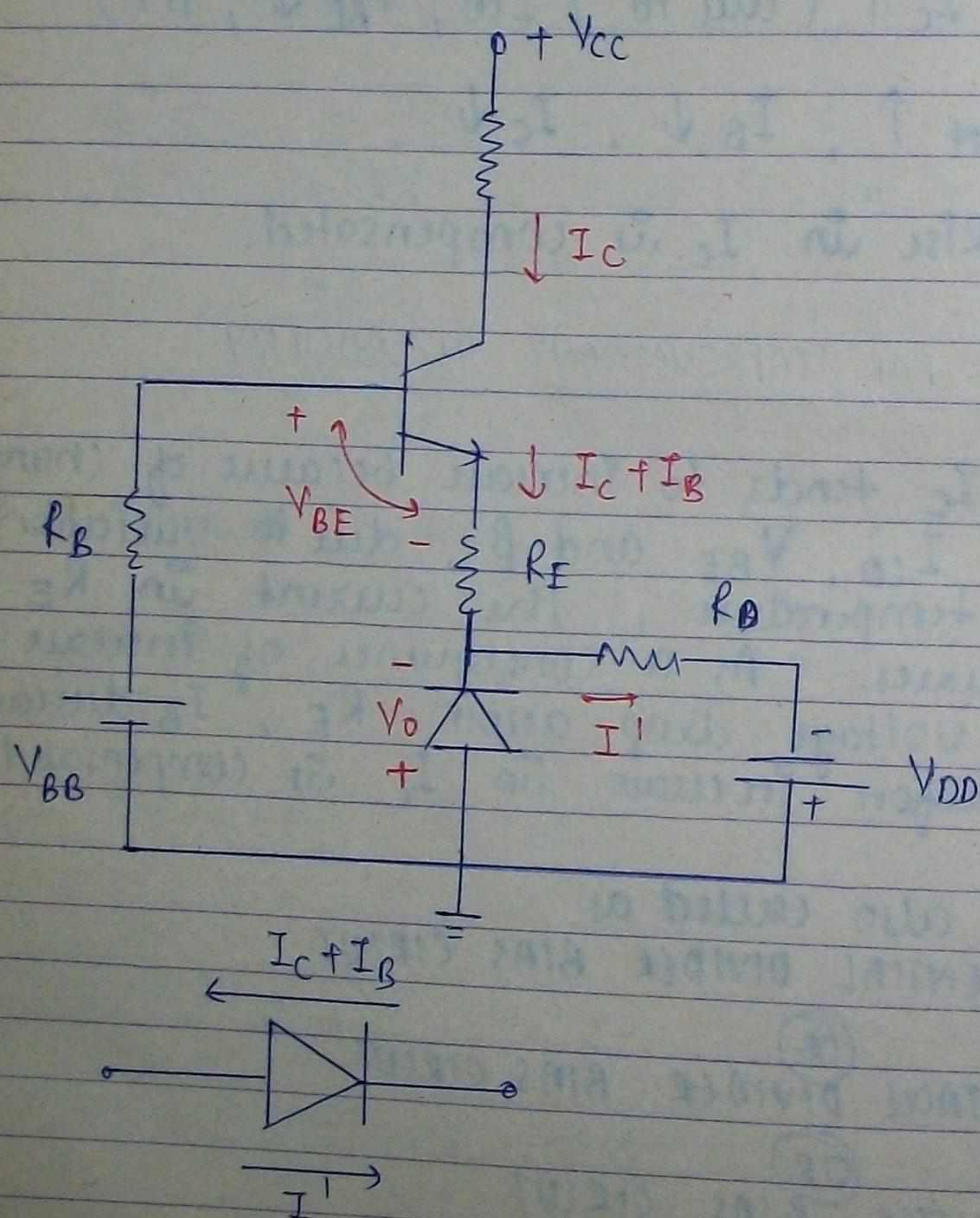
(OR)
VOLTAGE DIVIDER BIAS CIRCUIT

(OR)
EMITTER BIAS CIRCUIT

⇒ BIAS COMPENSATION -

- Compensation techniques refers to use of temp. sensitive device such as diode, transistor, thermistor, sensistors etc.

① DIODE COMPENSATION FOR V_{BE} -



$$I = I' - (I_C + I_B) > 0$$

∴ Diode is forward biased.

$$\frac{dV_{BE}}{dT} = -2.3 \text{ mV}/^\circ\text{C} \quad \checkmark$$

$$\frac{dV_0}{dT} = -2.3 \text{ mV}/^\circ\text{C} \quad \checkmark$$

$$I_C = \beta I_B + (1 + \beta) I_{CO} \quad - (1)$$

$$V_{BB} = I_B R_B + V_{BE} + (I_C + I_B) R_E - V_0$$

$$I_C = \frac{\beta [V_{BB} - (V_{BE} - V_0)] + [R_B + R_E] (1 + \beta) I_{CO}}{R_B + R_E (1 + \beta)}$$

$$V_{BE} - V_0 = \text{CONST}$$

when temp increase by 1°C

$$(V_{BE} - 2.3 \text{ mV}) - (V_0 - 2.3 \text{ mV})$$

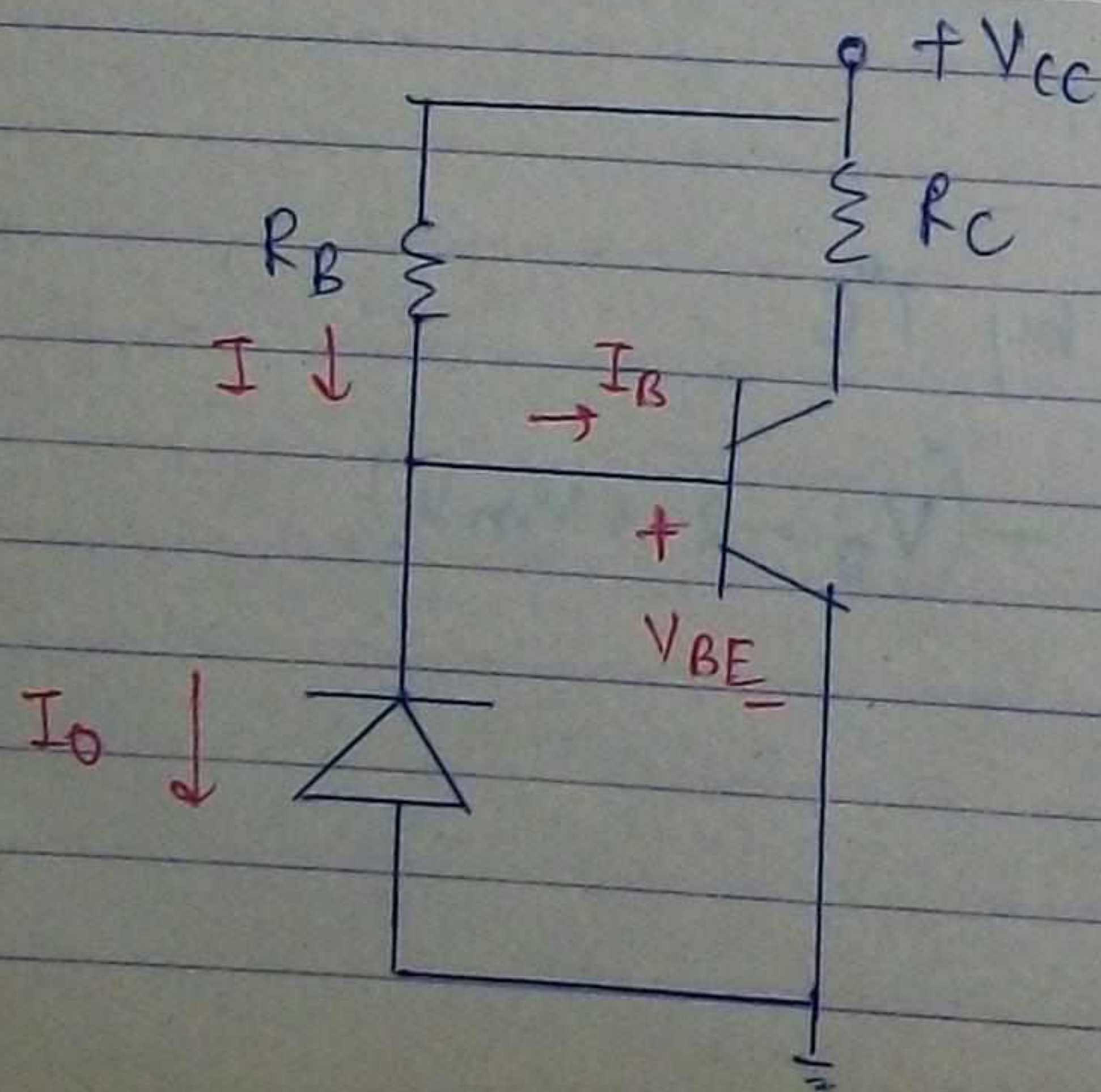
$$= V_{BE} - V_0$$

$$= \text{CONST}$$

- The change in V_{BE} with temp contribute significantly to change in I_C of Si Tx therefore circuit is useful for stabilisation of (Si) Tx.

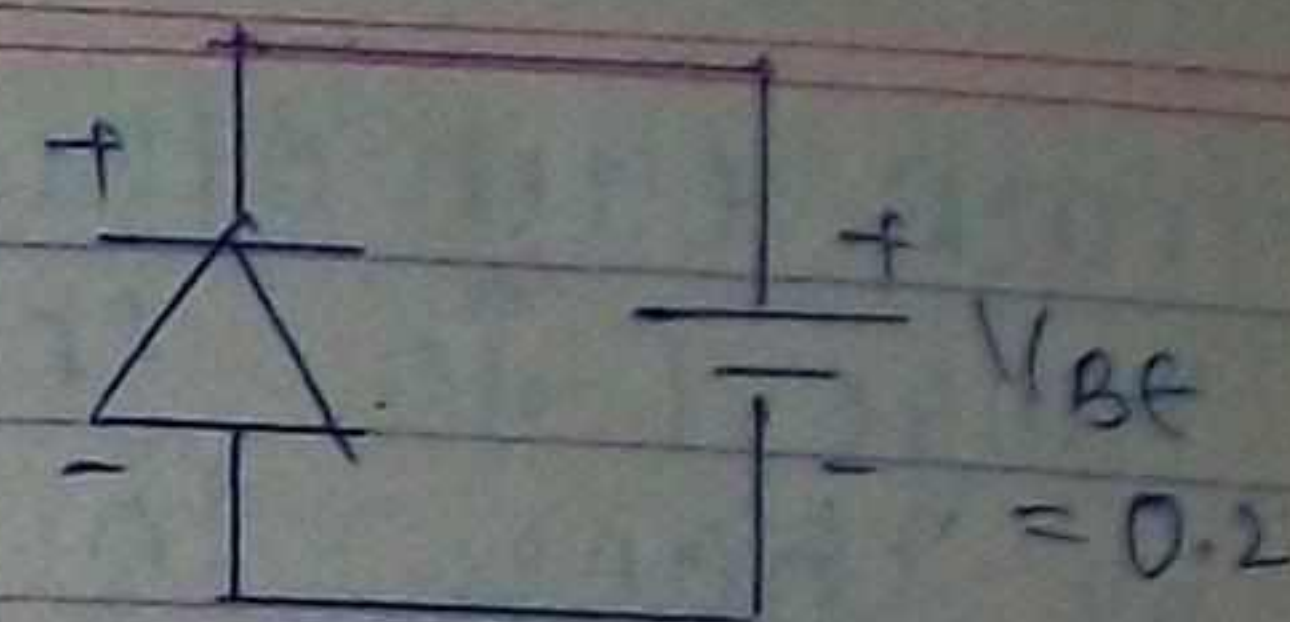
- The diode is kept forward biased by source V_{DD} and Resistor R_D . The diode is of same material and type as Tx. The Voltage V_D across diode will have same temp. coefficient $(-2.3mV/^{\circ}C)$ as V_{BE} , then it is clear from the eqn I_C will be insensitive to variation in V_{BE} (Since change in V_{BE} is cancelled by change in V_D)

② DIODE COMPENSATION FOR I_{CO} -



for Ge $V_{BE} = 0.2$

∴ Diode is off (R.B)



$$I_B = I - I_0$$

$$I = \frac{V_{CC} - V_{BE}}{R_B} \approx \frac{V_{CC}}{R_B} = \text{CONSTANT}$$

$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$I_C = \beta [I - I_0] + \beta I_{CO} \quad \because \beta \gg 1$$

$$I_C = \beta I - \beta I_0 + \beta I_{CO}$$

$$I_C = \underbrace{\beta I}_{\text{CONST.}} + \beta [I_{CO} - I_0]$$

- for Ge Tx, change in I_{CO} with Temp play more important role in collector current stability therefore this circuit is more useful for stabilizing Ge Tx.

- If Diode and Tx are of same type then I_0 of diode will increase with temp at same rate as I_{CO} , then I_0 and I_{CO}

track each other over desired temp. range (ie change in I_{CO} is cancelled by change in I_0) and I_{CO} will be insensitive to variation in I_{CO} .

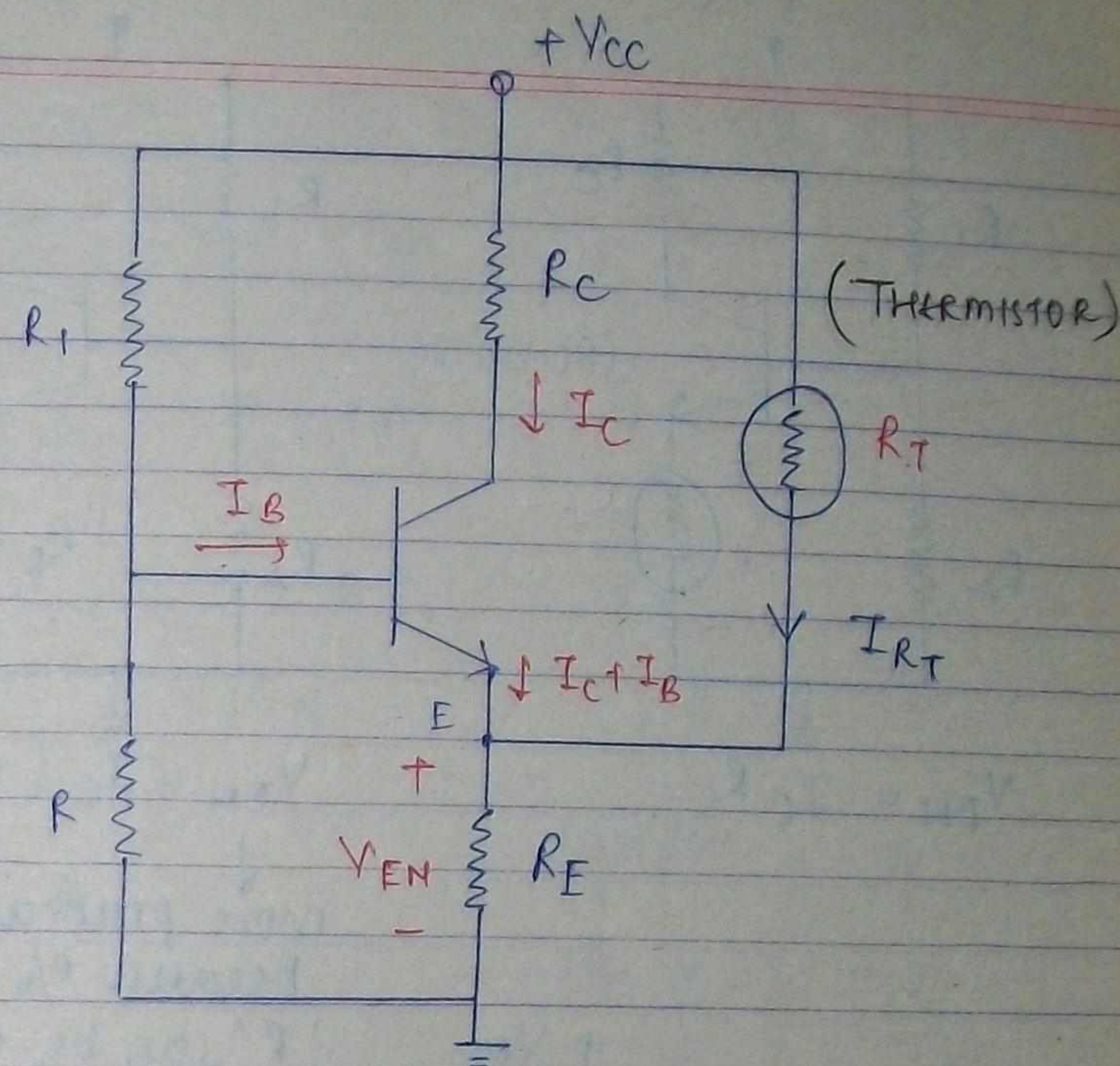
③ SENSISTOR AND THERMISTOR COMPENSATOR -

THERMISTOR - Negative temp. coefficient of resistivity. (lightly doped)

SENSISTOR - Positive temp coefficient of resistivity (highly doped).

$$\sigma = nq\mu$$

	n	
THERMISTOR LIGHTLY DOPED	small	$T \uparrow \mu \downarrow \rightarrow n \uparrow \sigma \uparrow \rho \downarrow -ve$
MEDIUM DOPED	medium	$T \uparrow n \uparrow \mu \downarrow \rightarrow n \mu \rightarrow \sigma \rightarrow \rho \rightarrow \text{const}$
SENSISTOR HIGHLY DOPED	high	$T \uparrow n \uparrow \mu \downarrow \rightarrow n \mu \downarrow \sigma \downarrow \rho \uparrow +ve$



R_T THERMISTOR -

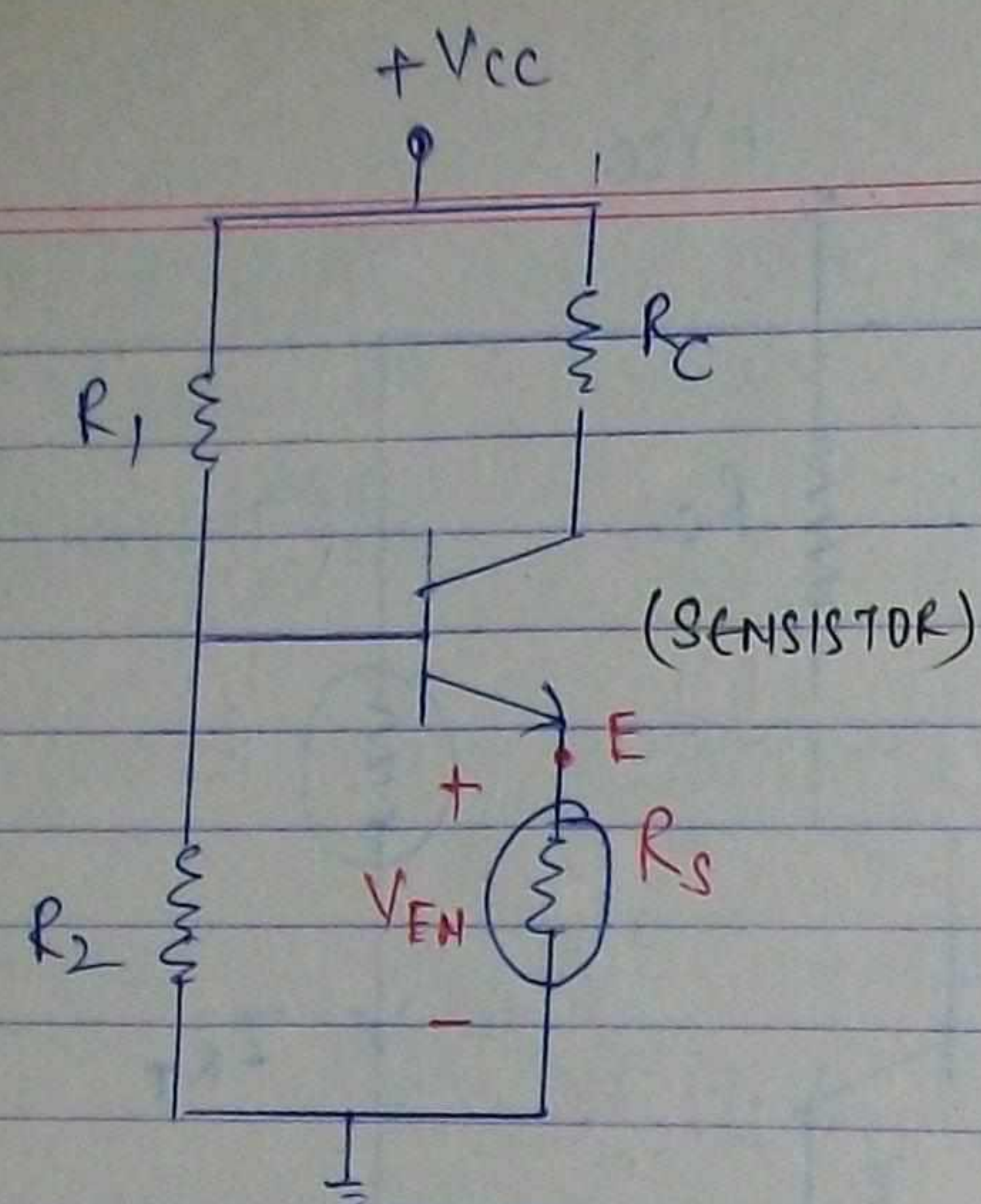
$$V_{EN} = [I_C + I_B + I_{RT}] R_E \approx [I_C + I_{RT}] R_E$$

$$I_B = \frac{V_{BB} - V_{BE} - V_{EN}}{R_{BB}}$$

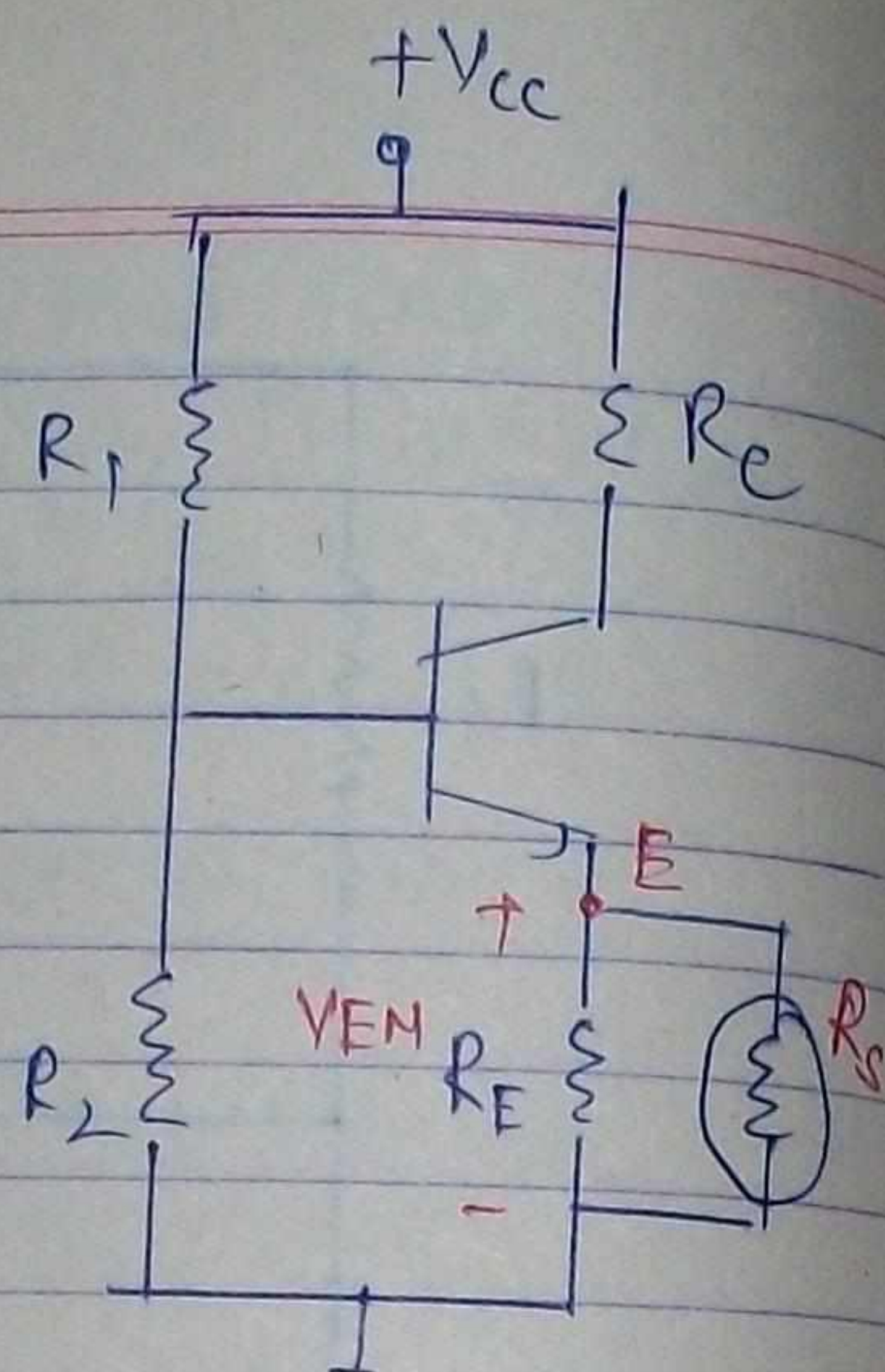
$$I_C = \beta I_B + (1 + \beta) I_{CO}$$

$$T \uparrow, I_C \uparrow, R_T \downarrow, I_{RT} \uparrow, V_{EN} \uparrow, I_B \downarrow, I_C \downarrow$$

Rise in I_C is compensated.

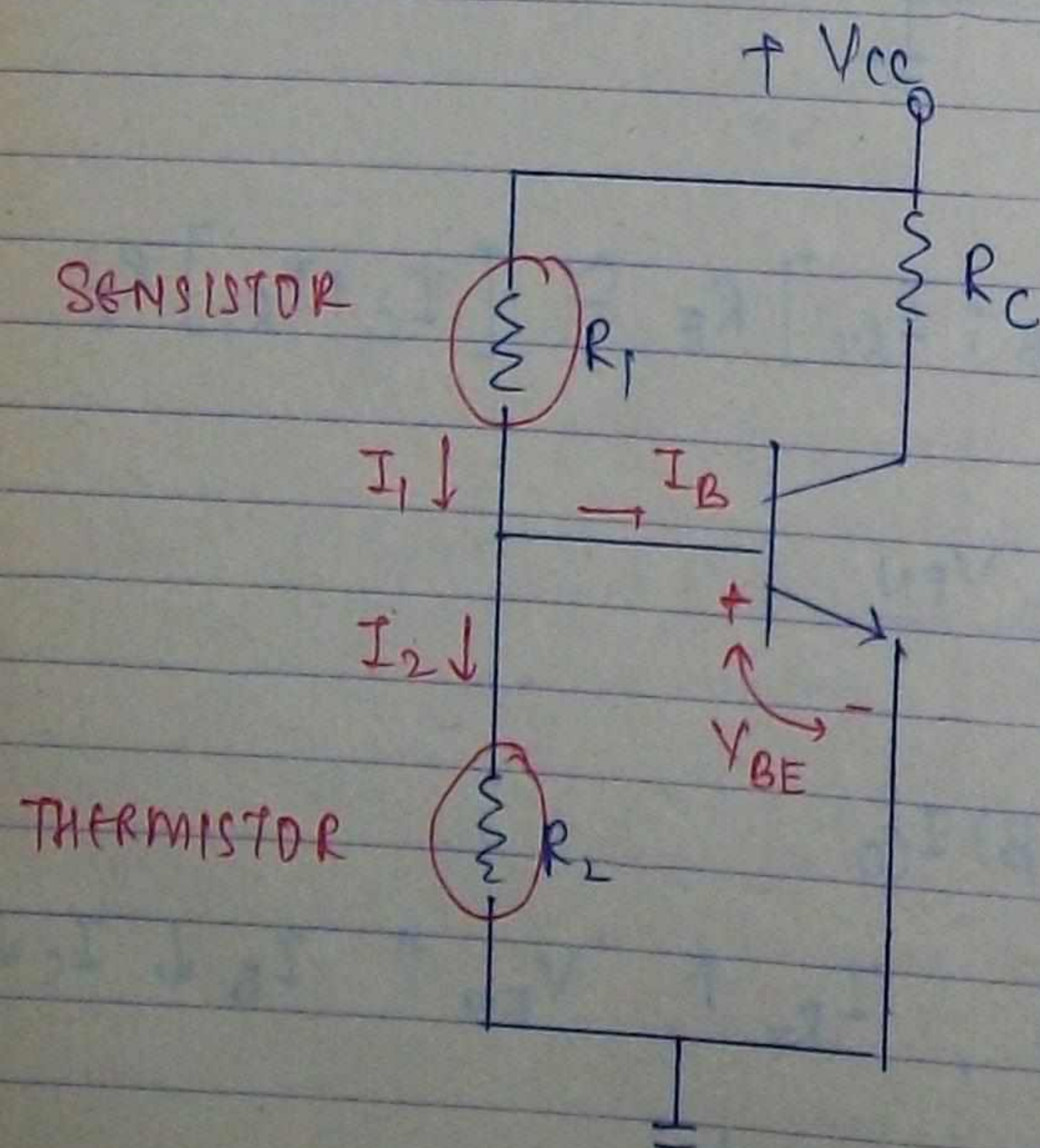


$$V_{EN} = I_C R_S$$



$$V_{EN} = I_C (R_S \parallel R_E)$$

↓
more practically used
because % change in
'R' can be controlled



$$I_B = I_1 - I_2$$

$$I_1 = \frac{V_{CC} - V_{BE}}{R_1}$$

$$I_2 = V_{BE} / R_2$$

As $T \uparrow$, $I_C \uparrow$ $I_B \downarrow$

thus $I_1 \downarrow$ $R_1 \uparrow$ (SENSOR)

and/or

and/or

$I_2 \uparrow$

$R_2 \downarrow$ (THERMISTOR)

Que. In a two stage circuit shown assume

$\beta = 100$ for each Tx.

① Determine R so that quiescent condⁿ are

$V_{CE1} = -4V$, $V_{CE2} = -6V$

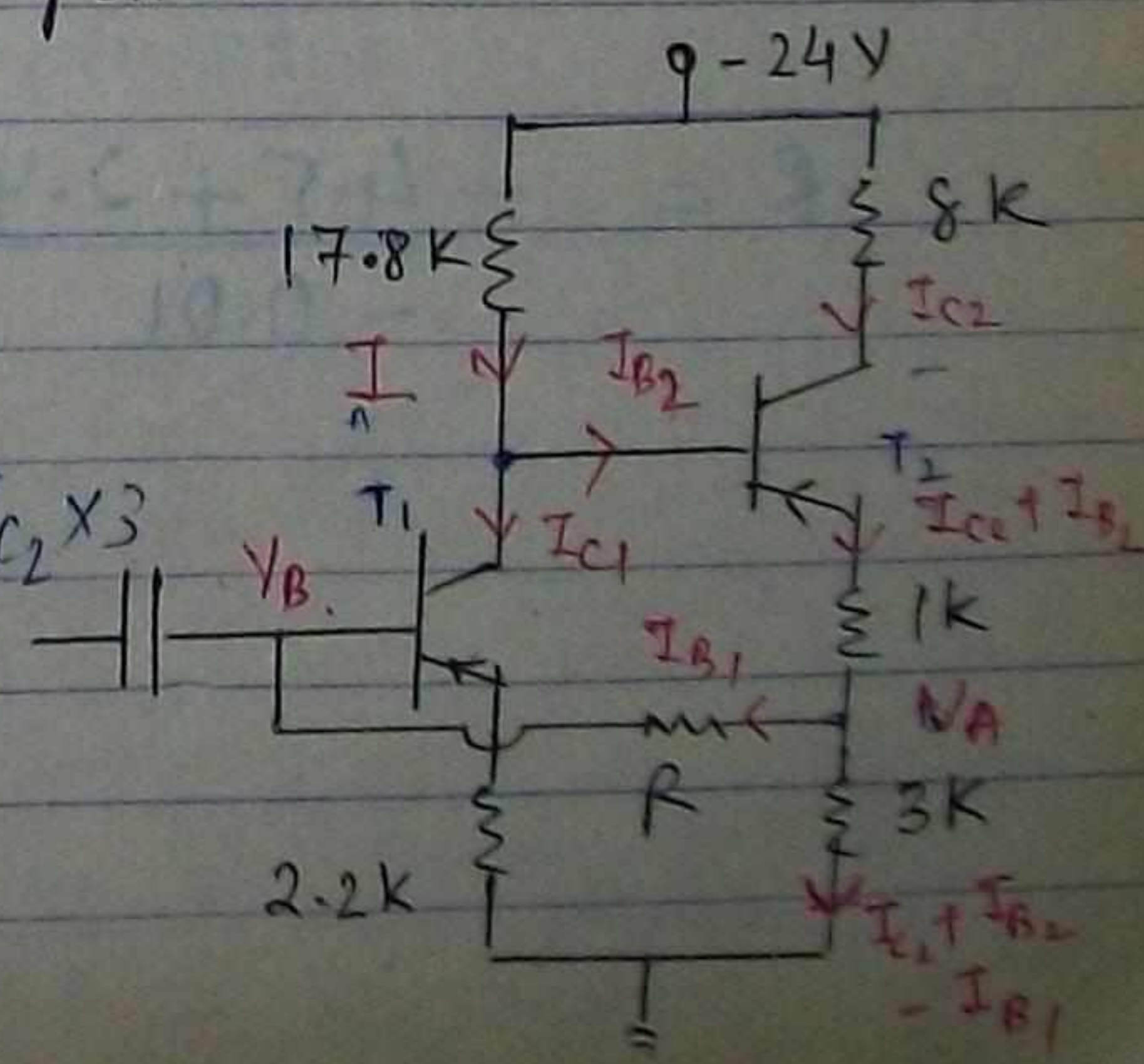
② Explain how q-point stabilization is obtained.

Assume $V_{BE} = 0.2$.

Solⁿ:

$$-24 = 8 \times I_{C2} - 6 + I_{C2} \times 1 + I_{C2} \times 3$$

$$I_{C2} = \underline{\underline{-1.5 \text{ mA}}}$$



$$\text{Sol}^n: -24 = +17.8 I_{C1} + V_{CE1} + 2.2 I_{C1}$$

$$-24 = +17.8 I_{C1} - 4 - 2.2 I_{C1}$$

$$I_{C1} = \frac{-2.4 + 4}{20} = \underline{\underline{-1 \text{ mA}}}$$

$$R = \frac{V_A - V_B}{I_{B1}}$$

$$I_{B1} = \frac{I_{C1}}{\beta} = \frac{-1}{100} = -0.01 \text{ mA}$$

$$I_{B1} \ll I_{C1}$$

∴ our assumptions is right.

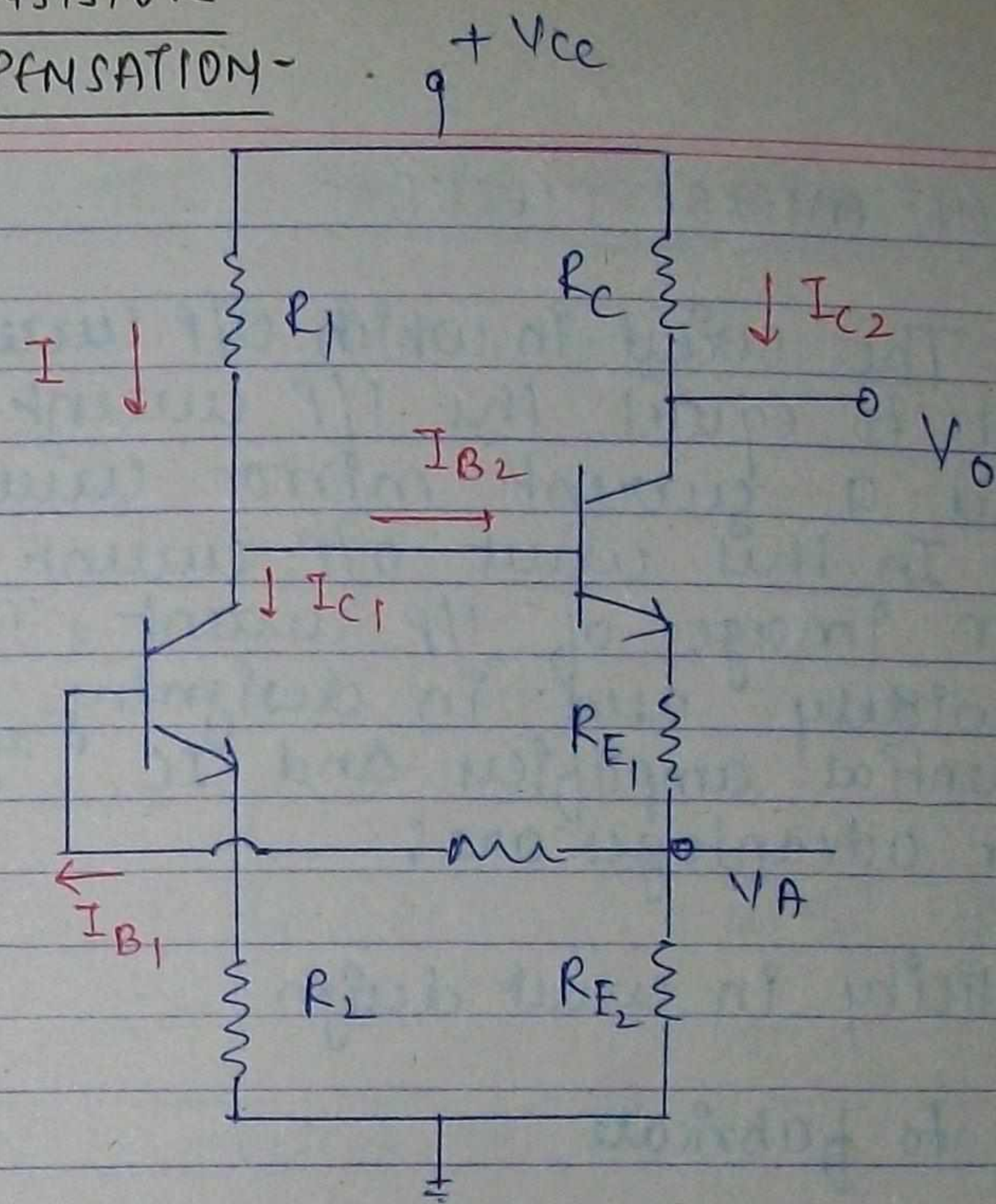
$$V_A = I_{C2} \times 3k = \underline{\underline{-4.5 \text{ V}}}$$

$$V_B = V_{BE1} + 2.2 \times I_{C1}$$

$$= -0.2 + 2.2 \times (-1) = \underline{\underline{-2.4 \text{ V}}}$$

$$R = \frac{-4.5 + 2.4}{-0.01} = \underline{\underline{210 \text{ k}\Omega}}$$

④ TRANSISTOR COMPENSATION -



$$T \uparrow \quad |I_{C2}| \uparrow \quad |V_A| \uparrow \quad |I_{B1}| \uparrow$$

$$I_{C1} \uparrow, I_{B2} \downarrow \quad |I_{C2}| \downarrow$$

Rise in I_{C2} is compensated.



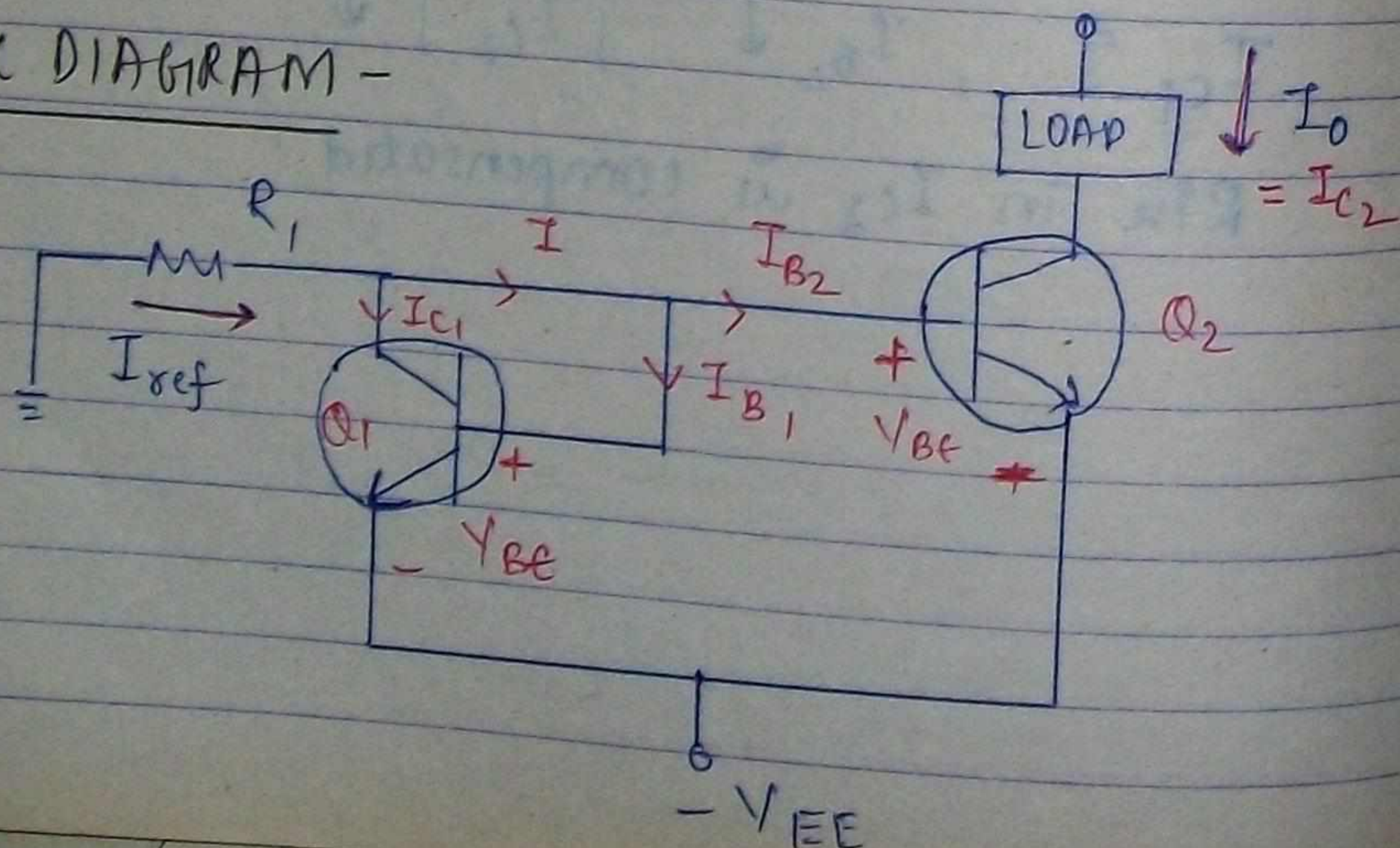
⇒ CURRENT MIRROR CIRCUIT -

The circuit in which O/P current is forced to equal the I/P current is called a current mirror circuit.

In this circuit O/P current is a mirror image of I/P current. They are widely used in designing of differential amplifiers and IC. Their major advantages are:

1. Simplicity in circuit design.
2. Easy to fabricate.
3. Minimum no. of components are required.
4. Low cost.

BASIC DIAGRAM -



Q₁ and Q₂ are identical transistor

$$B_1 = B_2, \quad V_{BE1} = V_{BE2} = V_{BE}$$

$$I_{ref} R_1 + V_{BE} - V_{EE} = 0$$

$$I_{ref} = \frac{V_{EE} - V_{BE}}{R_1} = \text{CONSTANT}$$

$$I_{B1} = I_{B2}, \quad I_{C1} = I_{C2} \quad \because \quad B_1 = B_2 = B$$

$$I_{ref} = I_{C1} + I_{B1} + I_{B2}$$

$$= I_{C1} + 2 I_{B2} = I_{C2} + \frac{2 I_{C2}}{\beta}$$

$$I_{ref} = I_{C2} \left[1 + \frac{2}{\beta} \right]$$

$$I_{C2} = \frac{I_{ref}}{\left[1 + \frac{2}{\beta} \right]} = I_O$$

$$\beta \rightarrow \text{very large} \quad \frac{2}{\beta} \ll 1$$

$$I_O = I_{C2} \simeq I_{ref}$$

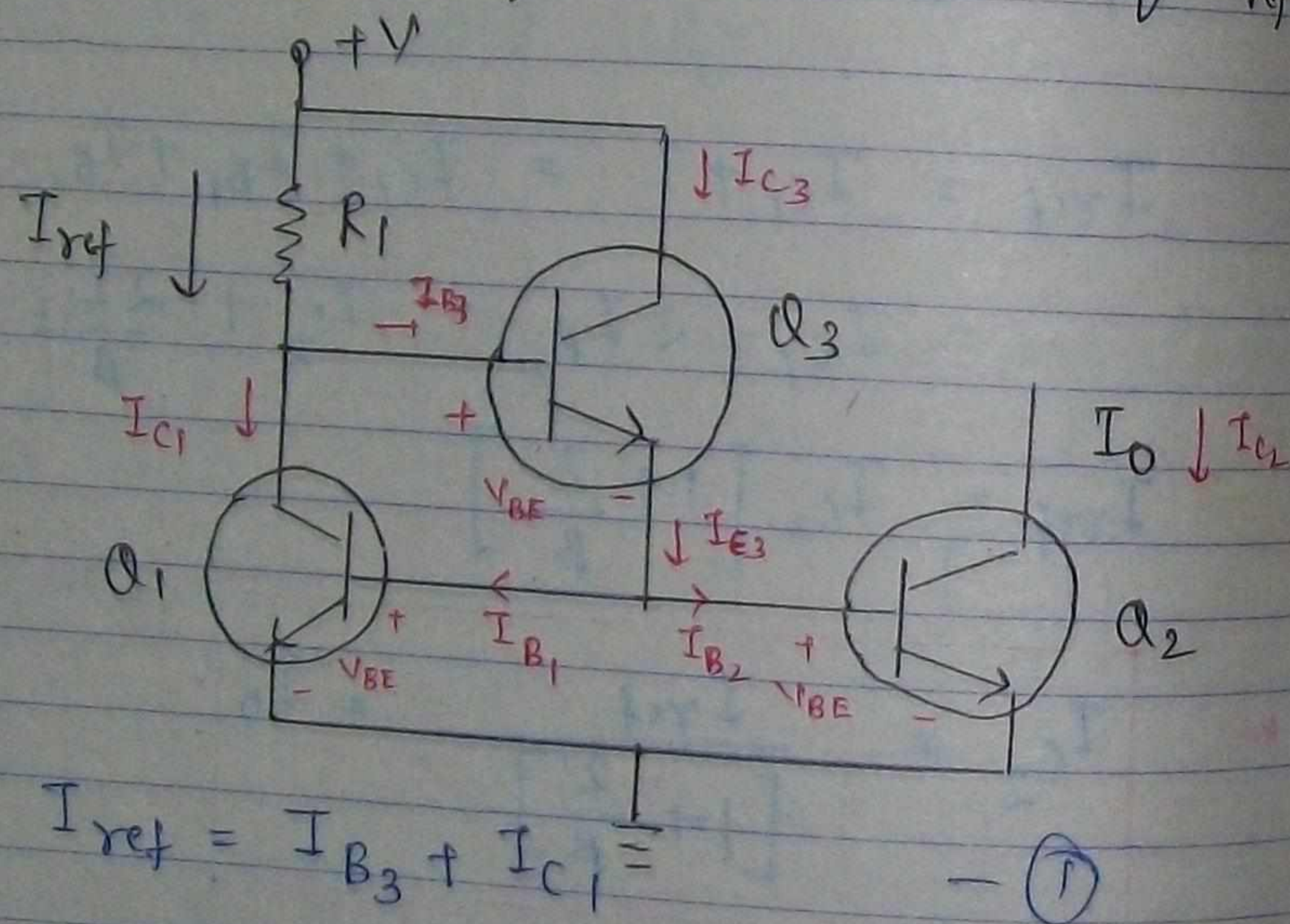
ESSENTIAL CONDITIONS -

- Both the transistor should be identical.
- β should be as large as possible.

Ques. for $I_{ref} = 25 \mu A$ $\beta = 25$ $I_0 = ?$
Soln:

$$I_0 = \frac{I_{ref}}{1 + \frac{2}{\beta}} = \frac{25}{1 + \frac{2}{25}} = 23.14 \mu A$$

Ques. write expression for I_0 in term of I_{ref}



Soln: $I_{ref} = I_{B3} + I_{C1} =$ — (1)

$$I_{ref} = \frac{V - 2V_{BE}}{R_1} = \text{CONST.}$$

$$I_{B1} = I_{B2} \Rightarrow I_{C1} = I_{C2} \quad \beta = \text{same}$$

$$I_{E3} = I_{B3} + I_{C3}$$

$$I_{E3} = I_{B3} + \beta I_{B3}$$

$$I_{B3} = \frac{I_{E3}}{[1 + \beta]}$$

$$I_{E3} = I_{B1} + I_{B2} = 2 I_{B2} = 2 \frac{I_{C2}}{\beta}$$

$$I_{B3} = \frac{2 I_{C2}}{\beta(1 + \beta)}$$

$$\therefore I_{ref} = \frac{2 I_{C2}}{\beta(1 + \beta)} + I_{C2}$$

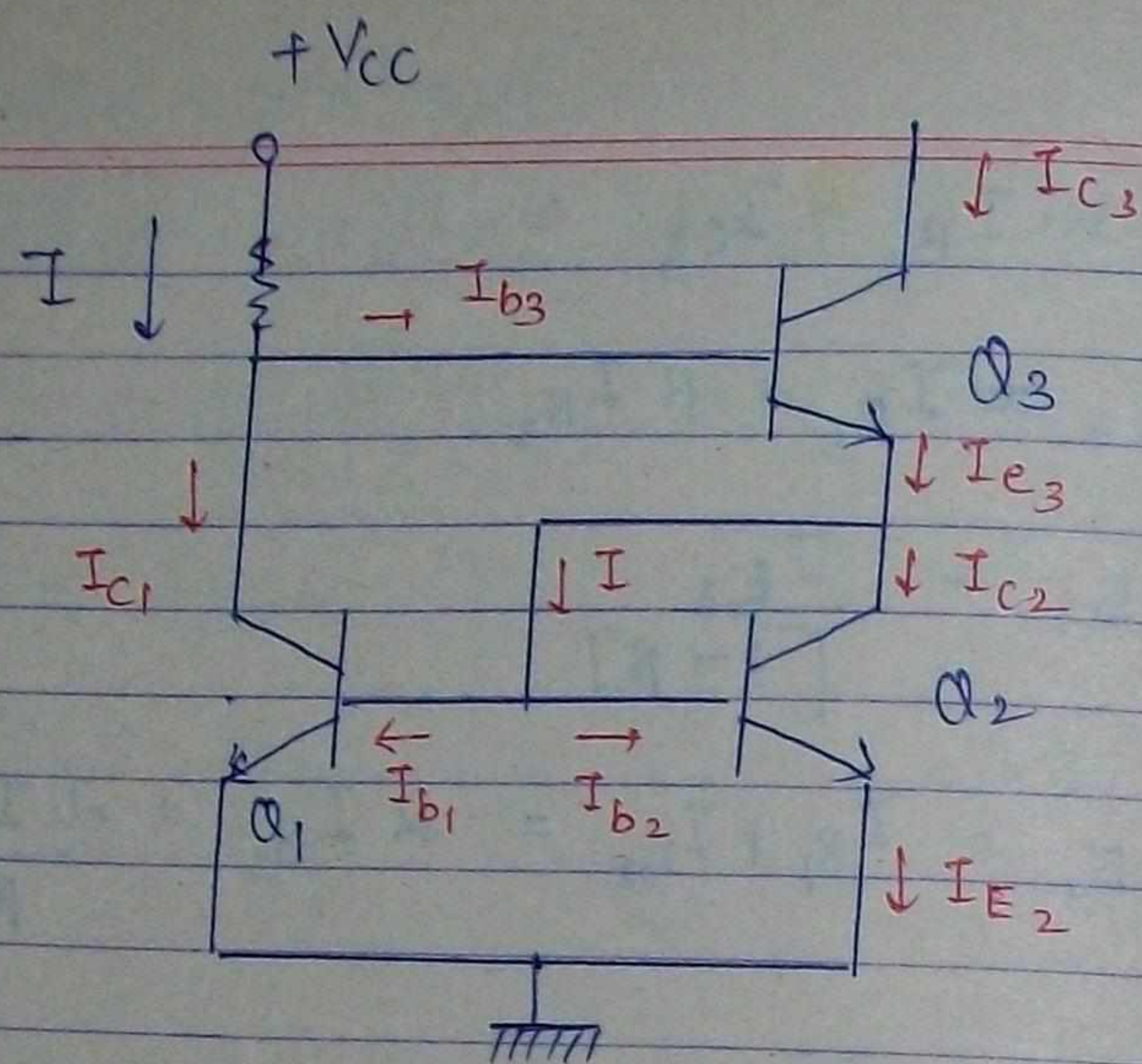
$$I_{ref} = \left[\frac{2}{\beta(1 + \beta)} + 1 \right] I_{C2}$$

$$I_{C2} = \frac{I_{ref}}{\left[1 + \frac{2}{\beta(1 + \beta)} \right]} = I_0$$

Ques P-14 Conv-1.

Soln:





$$I = I_{b3} + I_{c1} \quad - (1)$$

$$I_{b1} = I_{b2} \Rightarrow I_{c1} = I_{c2}$$

$$I_{b3} + \beta I_{b3} = I_{E3}$$

$$I_{b3} = \frac{I_{E3}}{(1+\beta)} \quad - (2)$$

$$I_{E3} = I_{c2} + 2I_{b2}$$

$$I_{E3} = I_{c3} + I_{b3}$$

$$I_{c3} + I_{b3} = I_{c2} + 2I_{b2}$$

$$I_{b3} = I_{c2} + 2I_{b2} - I_{c3}$$

$$I_{E3} = I_{c2} + 2I_{b2}$$

$$I_{E3} = \beta I_{b2} + 2I_{b2} = (\beta + 2) I_{b2}$$

$$I_{c1} = \beta I_{b1}$$

$$I_{c1} = \beta \frac{I_{E3}}{(2+\beta)}$$

$$I_{E3} = \frac{I_{c2} (2+\beta)}{\beta}$$

$$I = I_{c2} \left(\frac{2}{\beta} + 1 + 1 \right)$$

$$I = \left(2 + \frac{2}{\beta} \right) I_{c2} \quad - (3)$$

$$I_{c2} = I_{E3} - 2I_{b2}$$

$$I_{c2} = I_{c3} + I_{b3} - 2I_{b2}$$

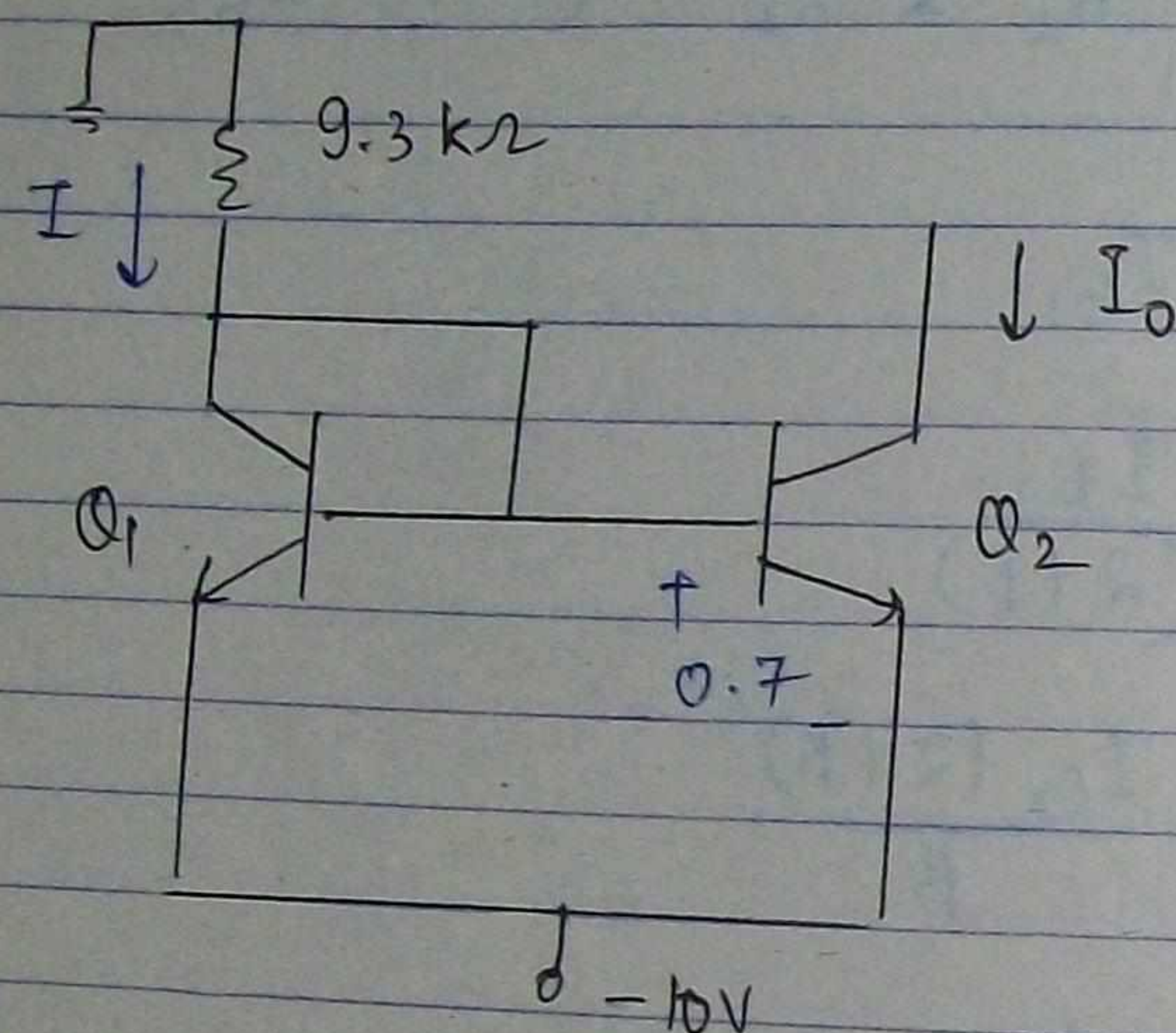
$$= I_{c3} + \frac{I_{c3}}{\beta} - 2 \frac{I_{c2}}{\beta}$$

$$I_{c2} \left(1 + \frac{2}{\beta} \right) = \left(1 + \frac{1}{\beta} \right) I_{c3}$$

$$I = \frac{2(1+\beta)}{\beta} \times \frac{I_{c3} (\beta+1)}{\beta (2+\beta)} \times \beta = \frac{2(1+\beta)^2}{\beta(2+\beta)} I_{c3}$$

$$I_{c3} = I_0 = \frac{\beta(2+\beta) I}{2(1+\beta)^2}$$

Ques. In a Si BJT circuit area of E_1 is half of area of E_2



Soln:

$$I_{C1} = \frac{I_{C2}}{2}$$

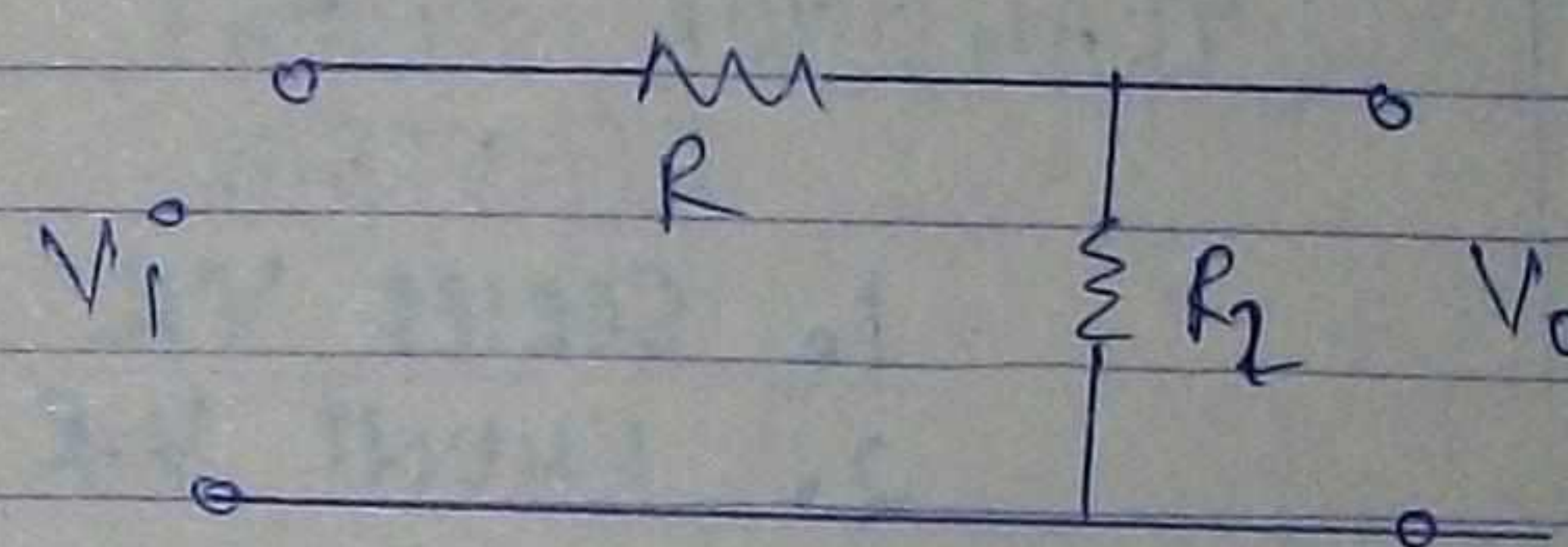
$$I \times 9.3 + V_{BE} - 10 = 0$$

$$I_{C1} = I = \frac{10 - 0.7}{9.3} = 1 \text{ mA}$$

$$I_{C2} = 2 I_{C1} = 2 \text{ mA}$$

⇒ VOLTAGE REGULATOR -

- It regulates the load voltage.
- In the regulator circuit the load voltage will be maintain almost a constant irrespective of load variation and i/p voltage fluctuation.



$$V_o = \frac{R_L}{R + R_L} V_i \rightarrow \text{line regulation}$$

$$\rightarrow \text{load regulation}$$

LINE REG - variation in V_o due to V_i .

LOAD REG - variation in V_o due to R_L .

- The performance of a regulator circuit is analysed by its regulation

$$\% \text{ REGULATION} = \frac{V_{NL} - V_{FL}}{V_{FL}} \times 100 \%$$

V_{NL} = NO LOAD
VOLTAGE

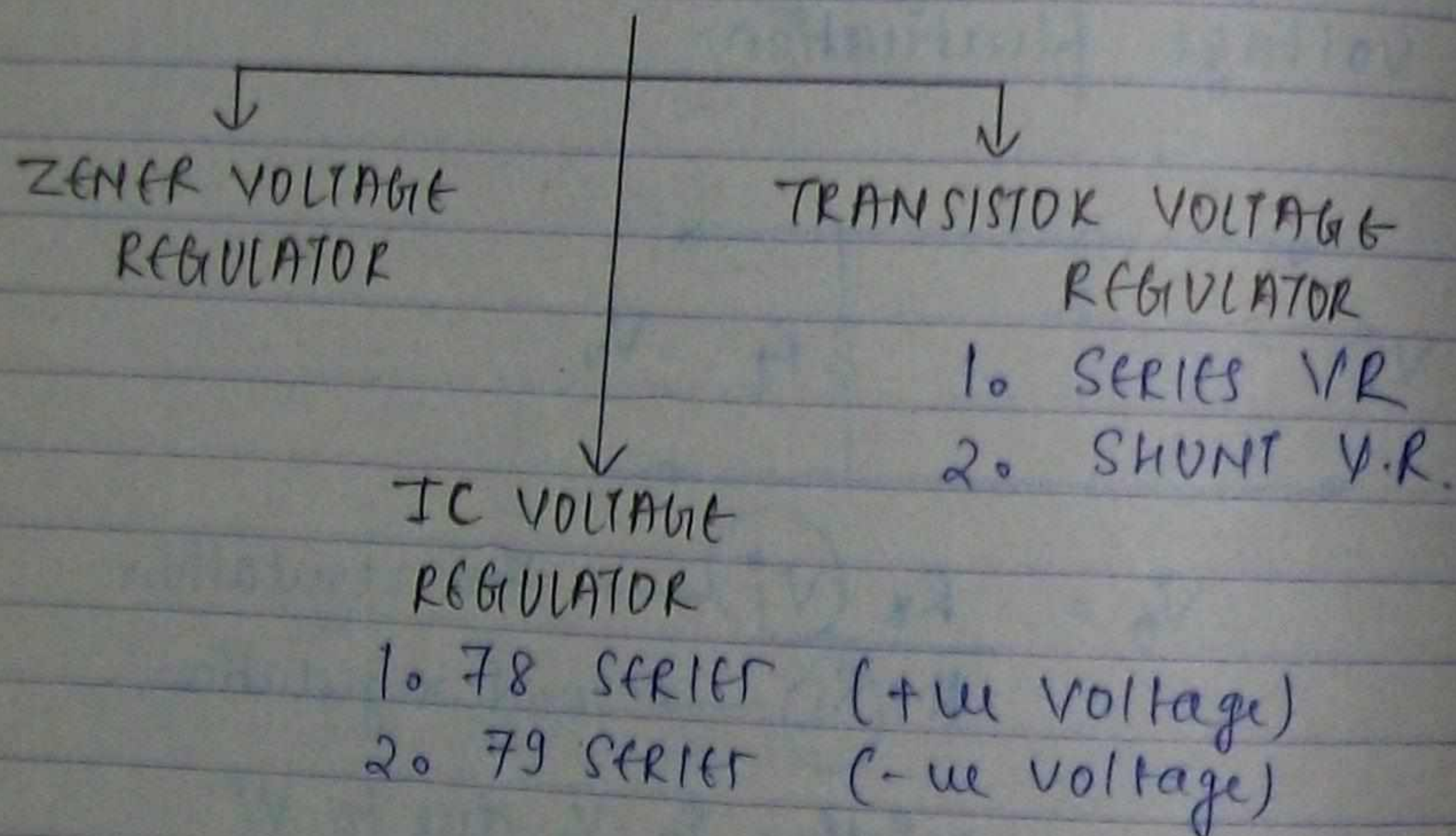
V_{FL} = FULL LOAD
VOLTAGE

Ideally $\% \text{ REG} = 0$

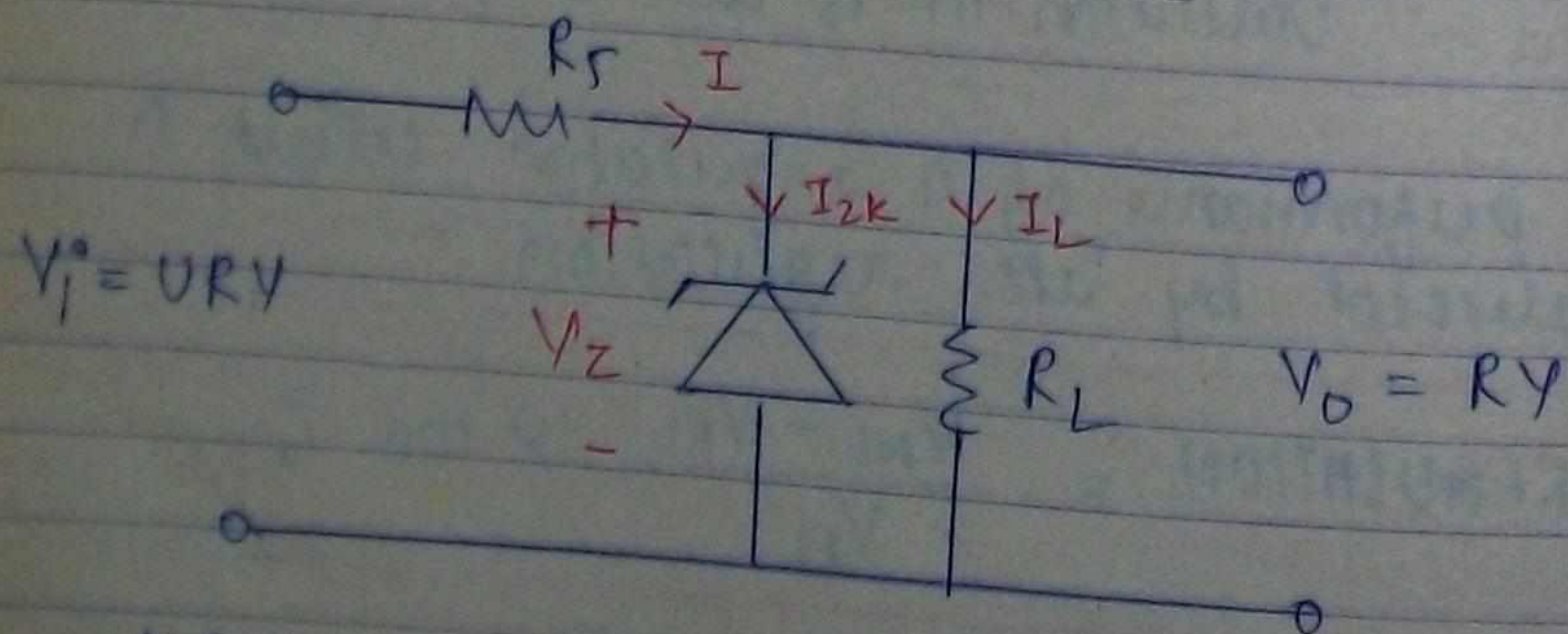
for better performance it should be low.

⇒ REGULATOR CIRCUIT -

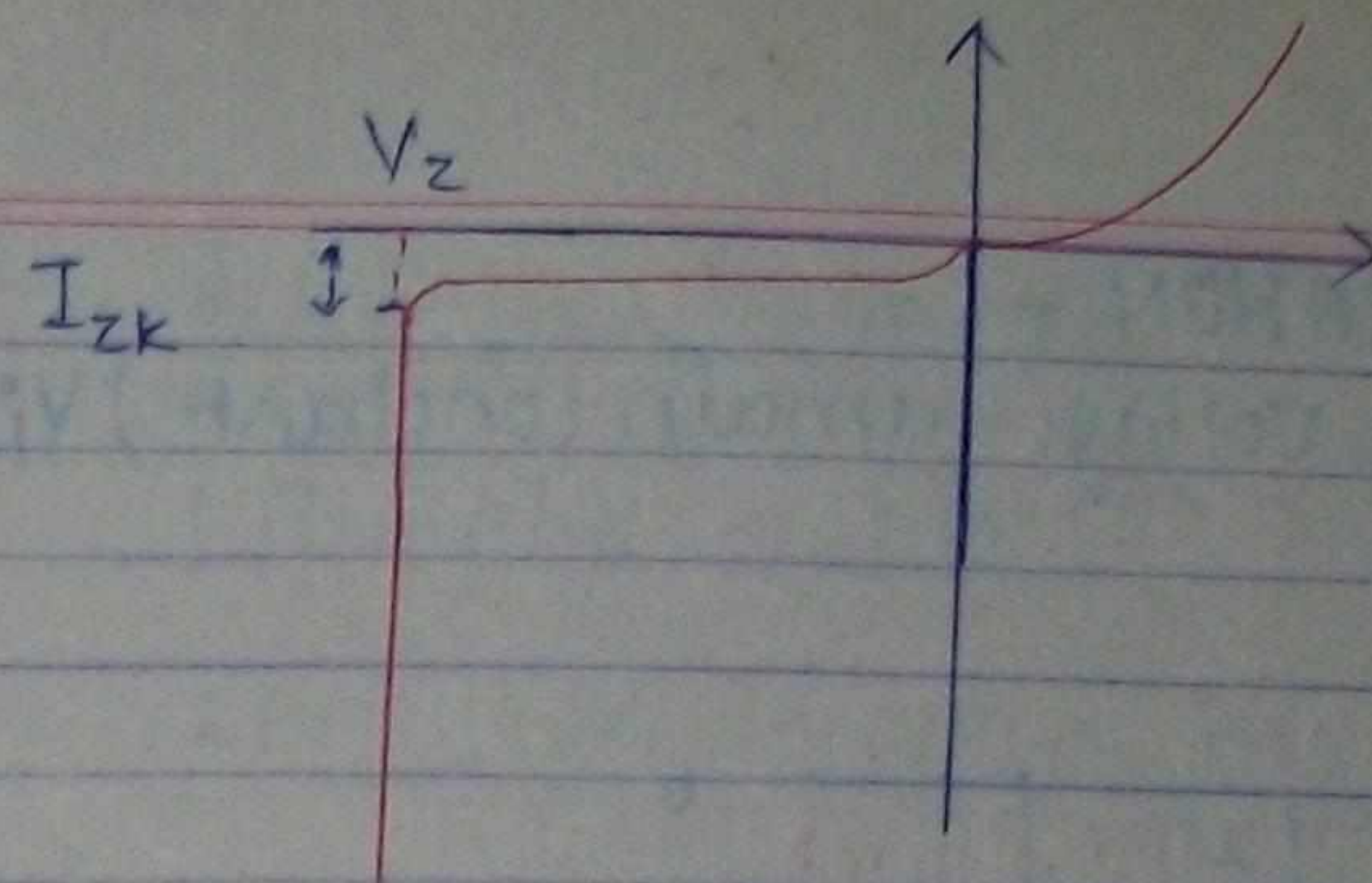
REGULATOR CIRCUIT



⇒ ZENER VOLTAGE REGULATOR -



URV = unregulated voltage
 RV = regulated voltage.



Zener breakdown
 There is a minimum current

$$I_{min} = I_{zk} = \text{knee current}$$

$$I_L = \frac{V_Z}{R_L}$$

for satisfactory performance

$$I \geq I_{zk} + I_L$$

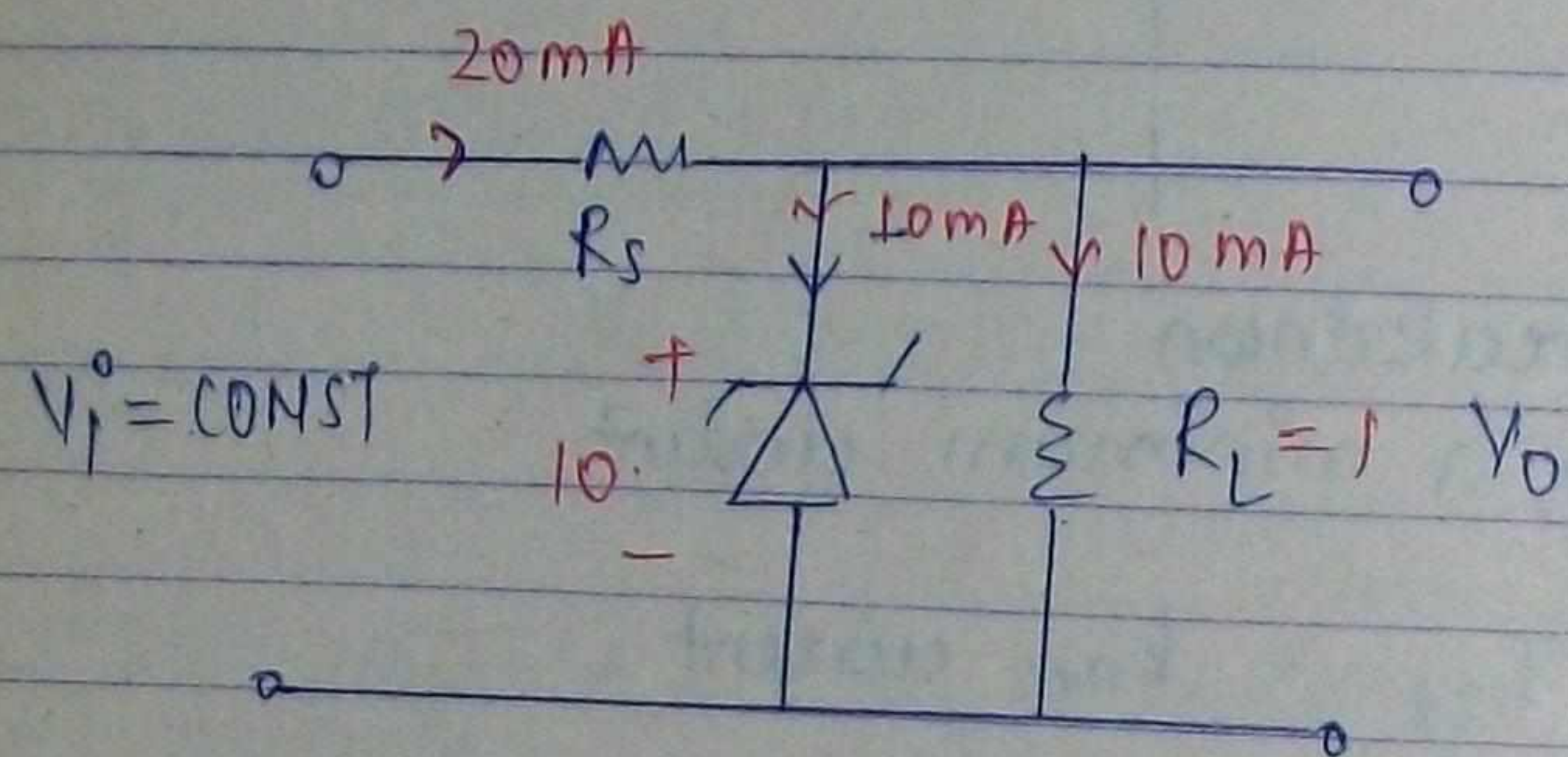
$$I = \frac{V_i - V_o}{R_s} = \frac{URV - RV}{R_s}$$

$V_i \uparrow$, $I \uparrow$, $I_z \uparrow$ but I_L remain constant.

∴ V_o remain constant. (LINE REGULATION)

LOAD REGULATION -

(voltage remain constant) V_i



Now when $R_L = 2k\Omega$ $I_L \downarrow = 5mA$

$R_L \uparrow$, $V_o \uparrow$, $I_Z \uparrow$, $I_L \downarrow$ $V_o = I_L R_L \downarrow$

$\therefore$ Rise in V_o is compensated.

POWER DISSIPATION ACROSS ZENER DIODE -

$$P_Z = I_Z V_Z = I_Z^2 R_Z$$

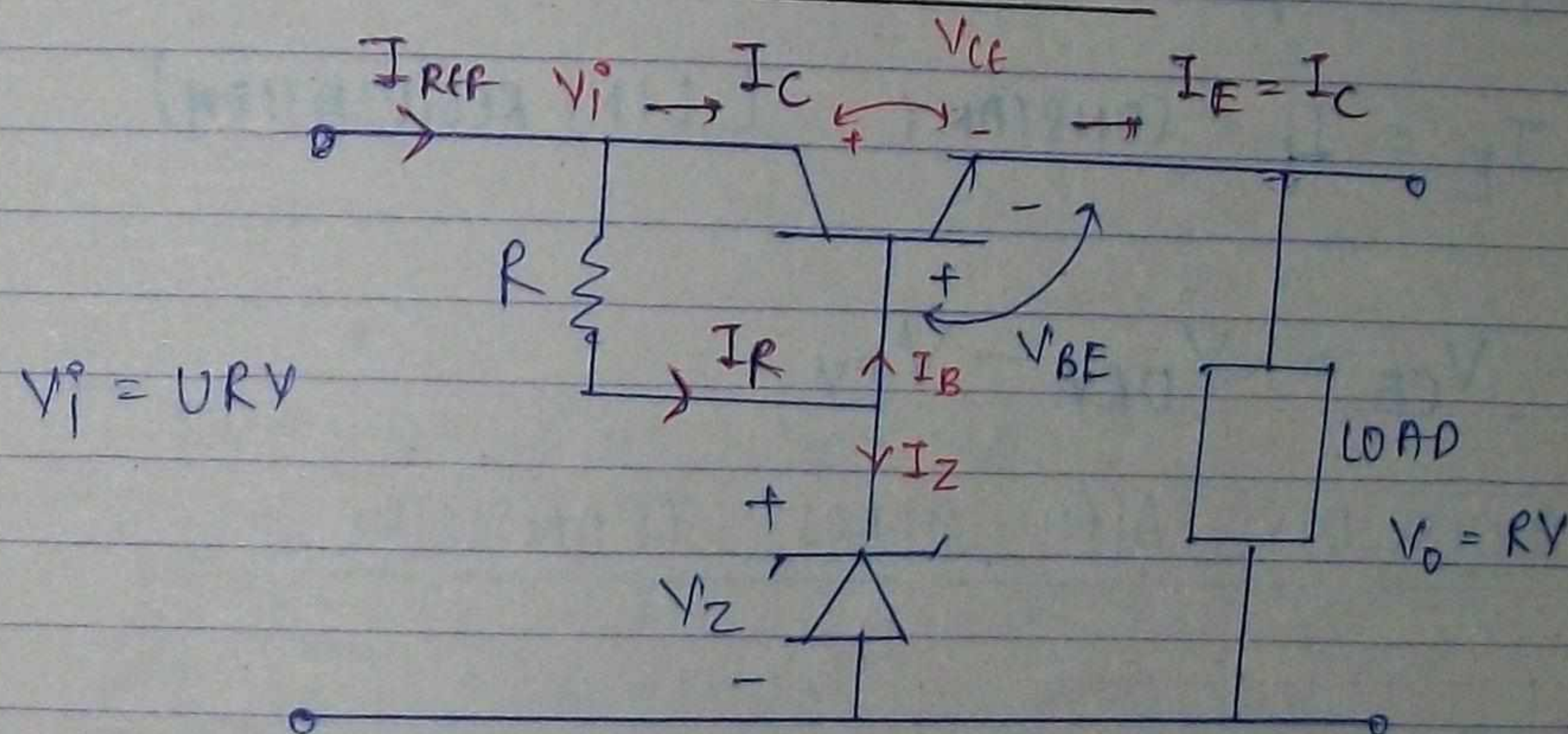
R_Z = dynamic resistance of Zener diode

R_s = limiting resistor.

- If URV is increased in the circuit, the excess power is dissipated by series resistor R_s (limiting resistor)

$\Rightarrow$ TRANSISTOR VOLTAGE REGULATOR -

(1) SERIES VOLTAGE REGULATOR -



$$RV = V_o - V_i = V_Z - V_{BE}$$

$$I_R = \frac{V_i - V_Z}{R}$$

LOAD REGULATION -

$V_i = \text{CONST}$

$\therefore I_R = \text{fixed}$

$R_L \uparrow$ $V_o \uparrow$ $I_Z \uparrow$ $I_B \downarrow$ $I_E \downarrow$ $I_L \downarrow$

$V_o \downarrow$ (load regulation)

Rise in V_o is compensated

LINE REGULATION -

$$R_L = \text{fixed} \quad I_L = \text{fixed}$$

$$V_o = I_L R_L = \text{fixed}$$

$$V_i \uparrow \quad I_R \uparrow \quad I_Z \uparrow \quad I_B \text{ CONSTANT}$$

$$I_E = I_L = \text{CONSTANT} \quad (\text{LINE REGULATION})$$

$$\# \quad V_{CE} = V_{URV} - V_{RV}$$

POWER DISSIPATED ACROSS TRANSISTOR -

$$P_T = V_{CE} \times I_C$$

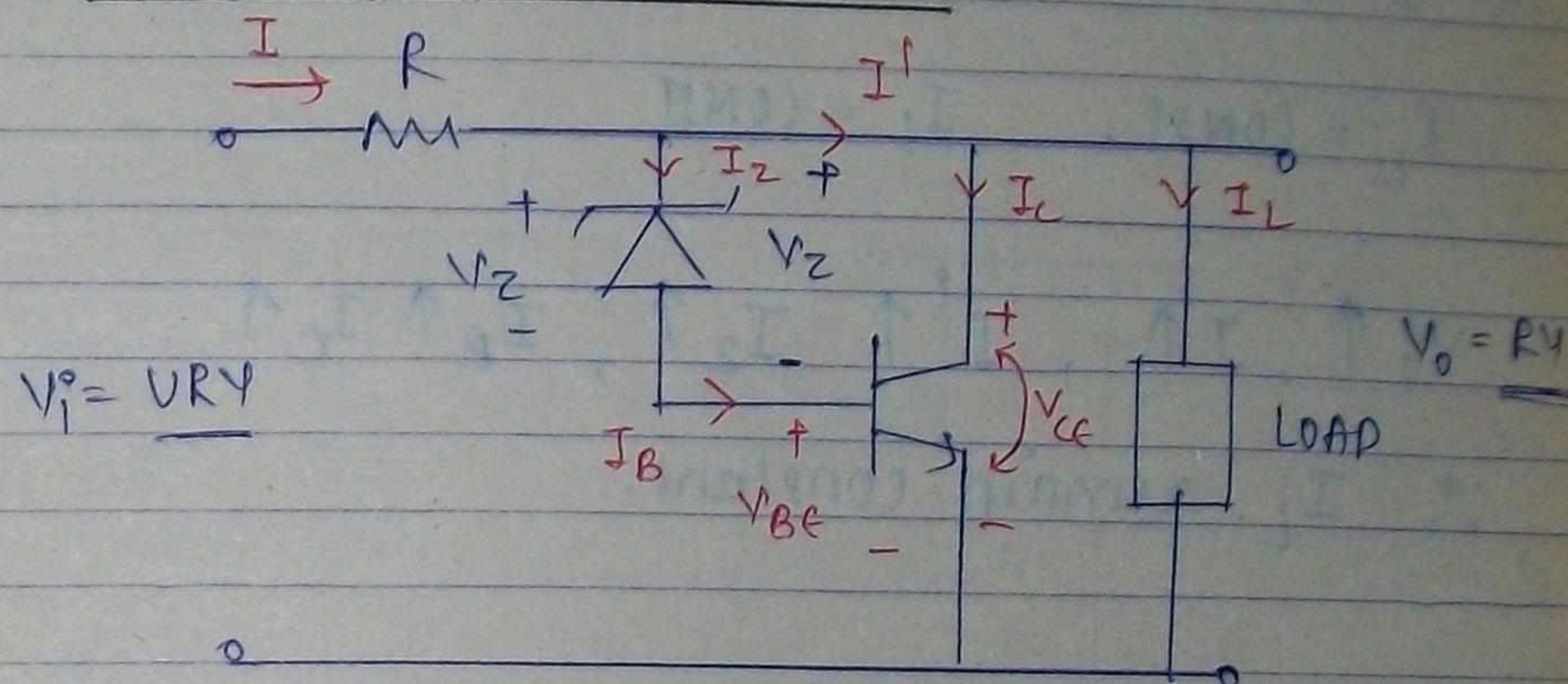
POWER DISSIPATION ACROSS DIODE (ZENER) -

$$P_Z = I_Z V_Z$$

- with the help of T_x , current or power handling capacity of circuit is increased

$$I_C = \beta I_B = \beta I_Z$$

② SHUNT VOLTAGE REGULATOR -



$$RV = V_o = V_{CE} = V_Z + V_{BE}$$

$$I = \frac{URV - RV}{R}$$

LOAD REGULATION -

$$V_i = \text{CONST} \quad I = \text{CONST.}$$

$$V_o \quad R_L \uparrow \quad V_o \uparrow, I_Z \uparrow \quad I_B \uparrow, I_C \uparrow, I_L \downarrow$$

$$\therefore V_o \downarrow$$

$\therefore$ Rise in voltage is (V_o) compensated

LINE REGULATION -

$$R_L = \text{CONST.} \quad I_L = \text{CONST}$$

$$V_i \uparrow, I \uparrow, I' \uparrow, I_2 \uparrow, I_B \uparrow, I_C \uparrow$$

$\therefore I_L$ remain constant.

⇒ FOR SERIES REGULATOR -

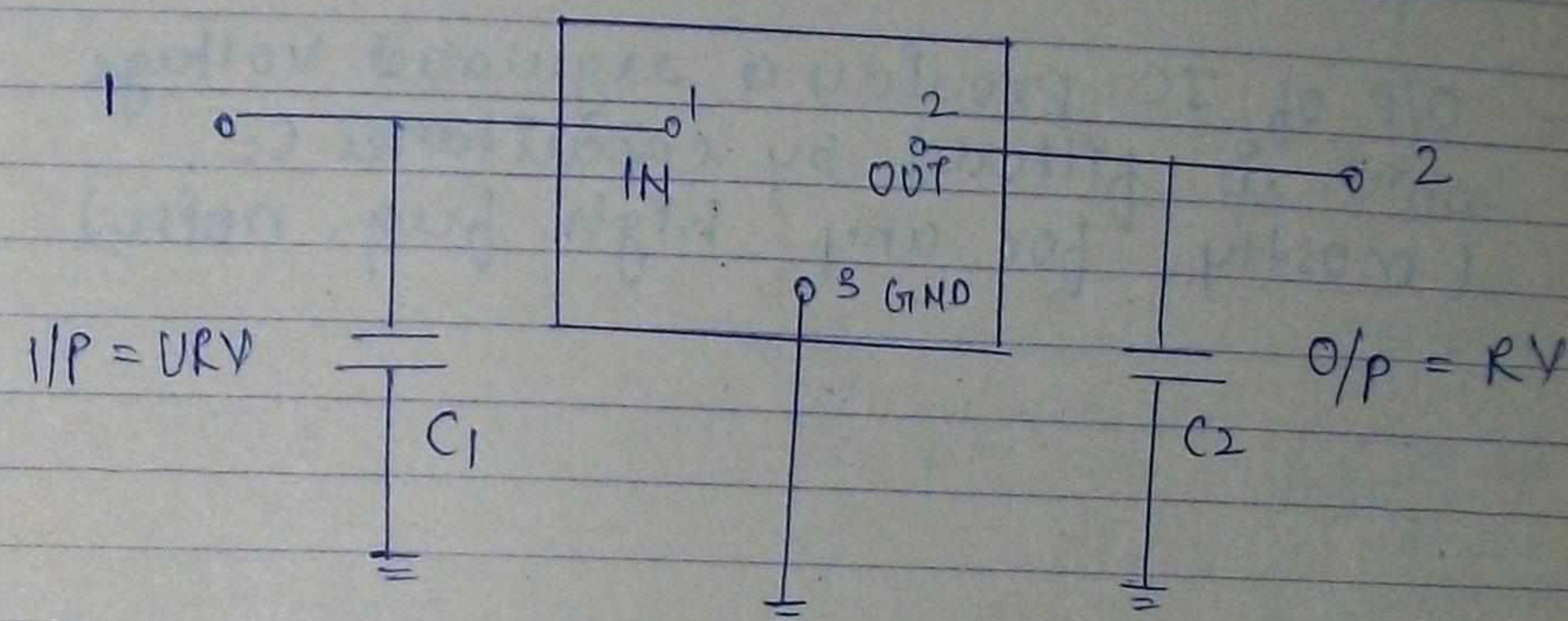
If URV is increased in the circuit extra power is dissipated by the Tx. In the given circuit Tx is in CC configuration. Therefore series voltage regulator is also called emitter follower voltage regulator circuit.

⇒ FOR SHUNT REGULATOR -

- If URV is increased the extra power is dissipated by series resistor R.
- In given circuit Tx is in CE configuration.

30/08/11

⇒ IC VOLTAGE REGULATOR -



78 SERIES (+VE VR)

7805
7810
7812
7815
7824

O/P VOLTAGE

+5V
+10V
+12V
+15V
+24V

79 SERIES (-VE VR)

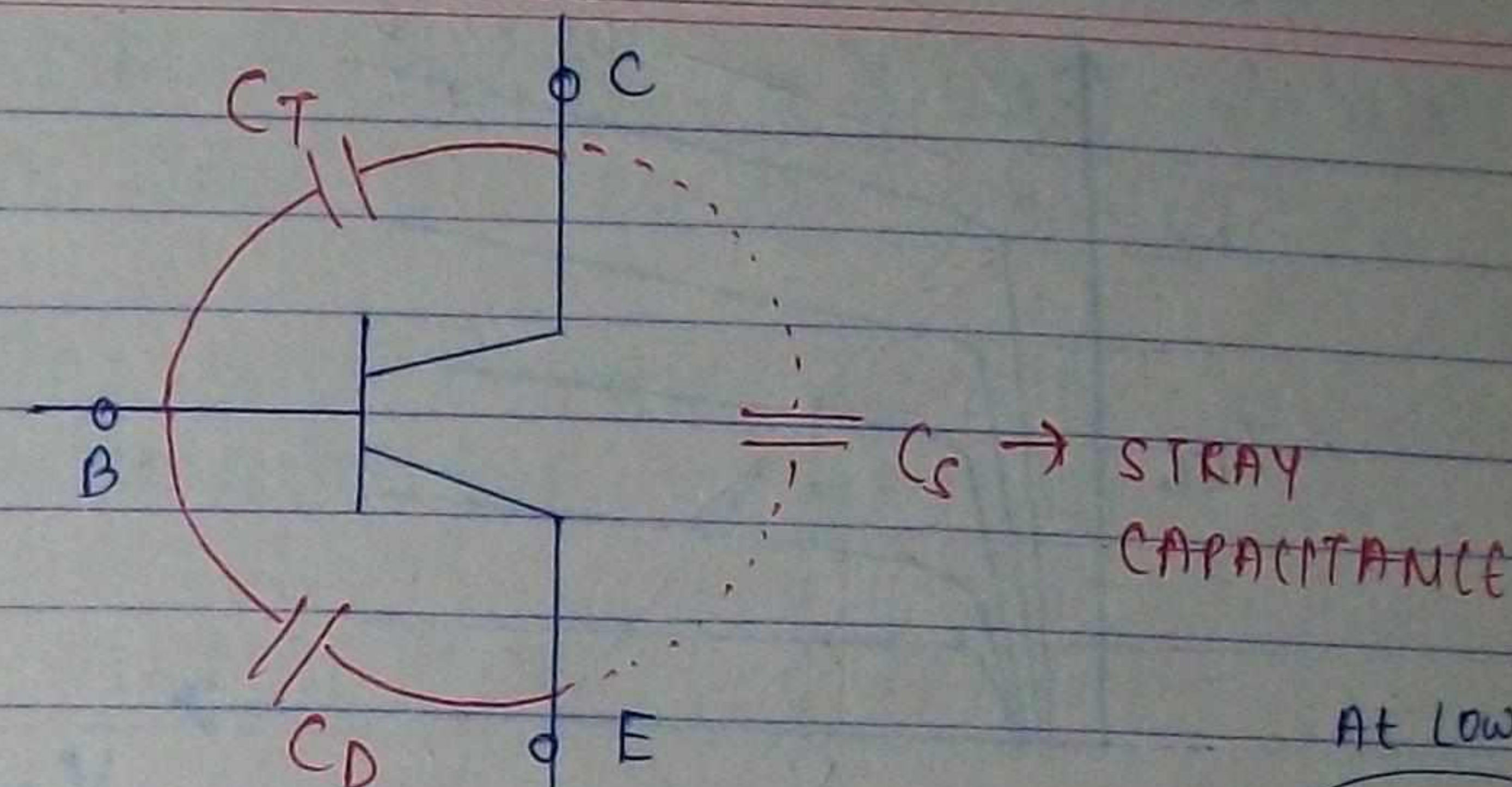
7905
7910
7912
7915
7924

O/P VOLTAGE

-5V
-10V
-12V
-15V
-24V

- An unregulated i/p V_i is filtered by capacitor C_1 .
- O/P of IC provides a regulated voltage which is filtered by capacitance C_2 .
(mostly for any high freq. noise)

LOW FREQUENCY MODEL



$$C_T = C_D = C_S \quad (\text{in pF})$$

$$Z_C = \frac{1}{2\pi f C}$$

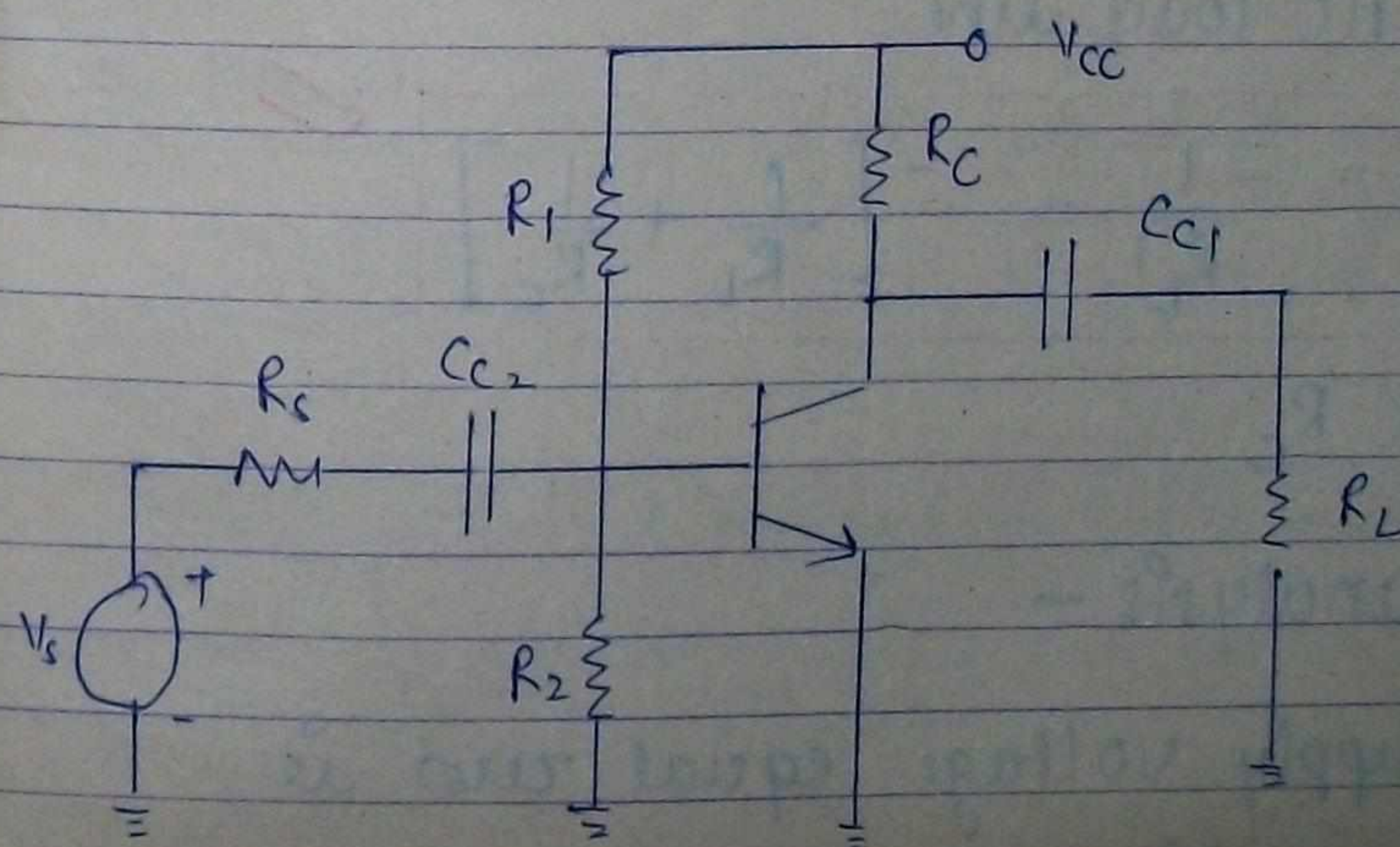
At Low FREQ.

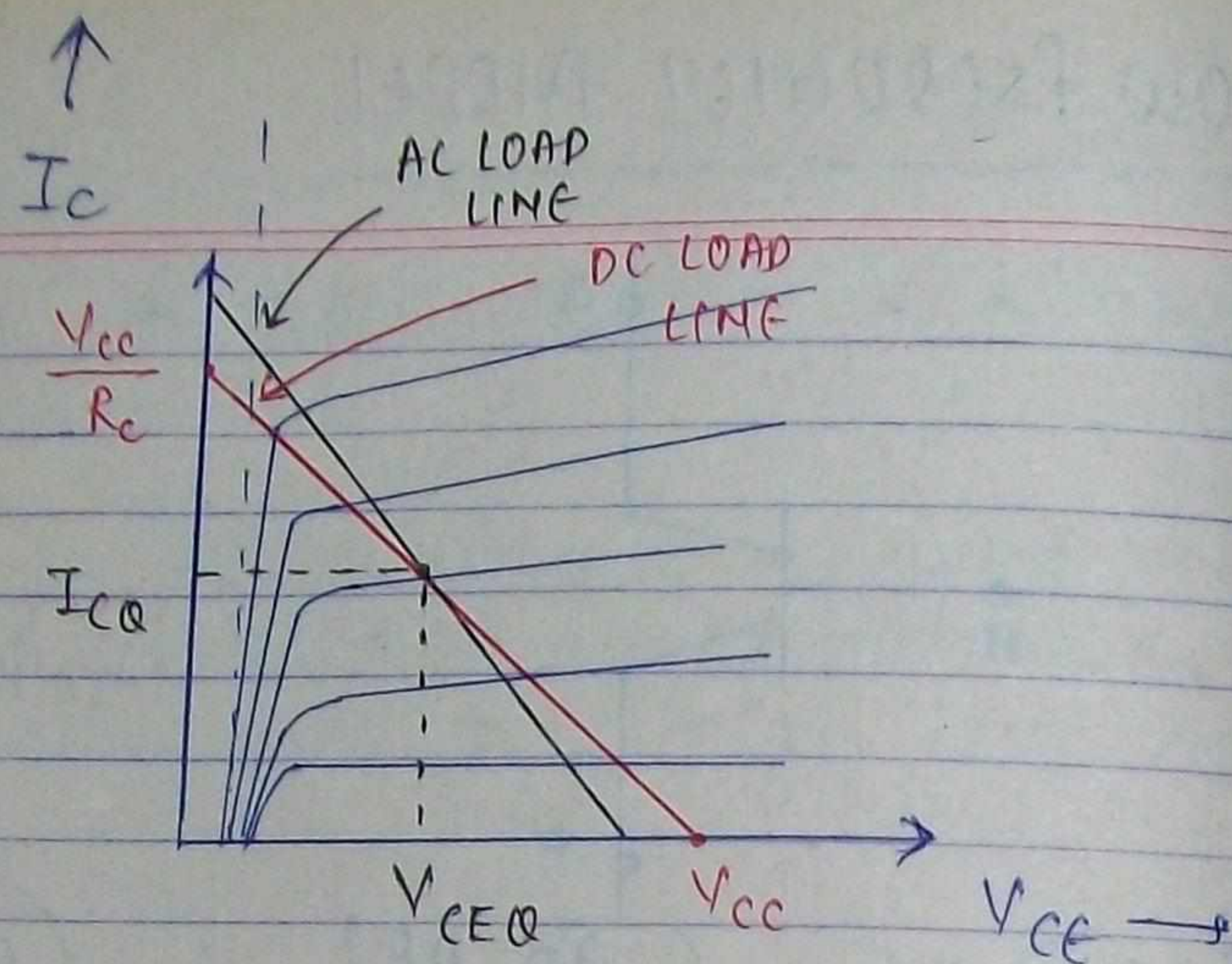
$C_C = \mu F$
create problem

$f \rightarrow \text{low} \quad f \rightarrow \text{low} \quad Z_C = \text{very high.}$

$\therefore$ At low freq $C_T C_D C_S$ acts as open circuit.

$\Rightarrow$ LOW FREQUENCY ANALYSIS OF BJT -





for AC analysis -

$V_{CC} = 0$ C_c and C_e act as short circuit

$$R_L' = R_C \parallel R_L$$

$$\frac{1}{R_L'} = \frac{1}{R_L} + \frac{1}{R_C}$$

Slope of AC load line

$$\text{SLOPE} = -\frac{1}{R_L'} = -\left[\frac{1}{R_L} + \frac{1}{R_C}\right]$$

$R_L' < R_C$

for AC analysis -

① DC supply voltage equal zero i.e. $V_{CC} = 0$

- ② first set Q-point, so that Tx in active region and act as an amplifier.
- ③ Coupling and Bypass capacitor act as short circuit for AC analysis.

⇒ SMALL SIGNAL MODEL OF A BJT -

- A Tx. operates linearly for small i/p signals. So a small signal model of Tx. have been developed in active region of operation.

- There are three small signal models of a BJT is used:

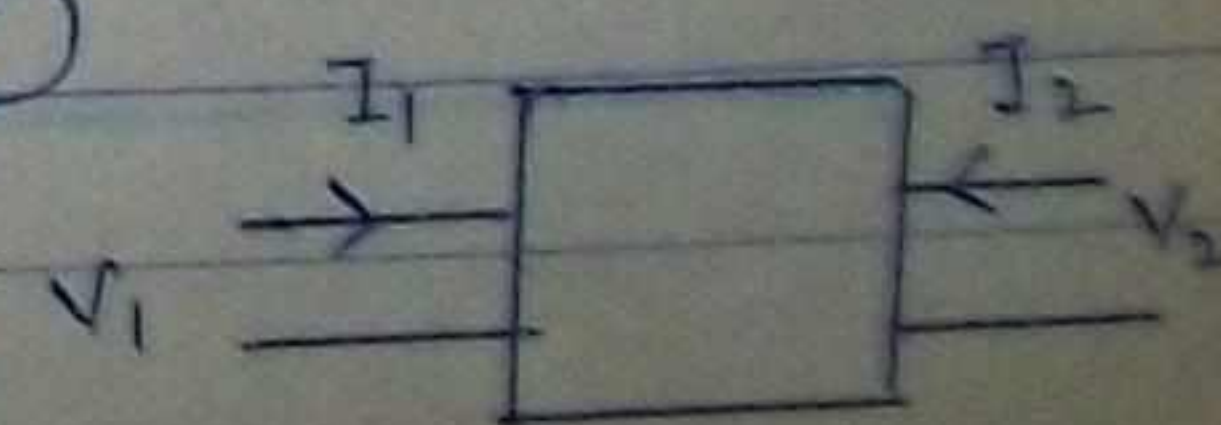
1. H-parameter model (Hybrid model)
2. Hybrid π -model
3. RE r_e model (T-model)

⇒ HYBRID H-PARAMETER MODEL -

$$\begin{bmatrix} V_1 \\ I_2 \end{bmatrix} = \begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \begin{bmatrix} I_1 \\ V_2 \end{bmatrix}$$

$$V_1 = h_{11} I_1 + h_{12} V_2 \quad - (1)$$

$$I_2 = h_{21} I_1 + h_{22} V_2 \quad - (2)$$



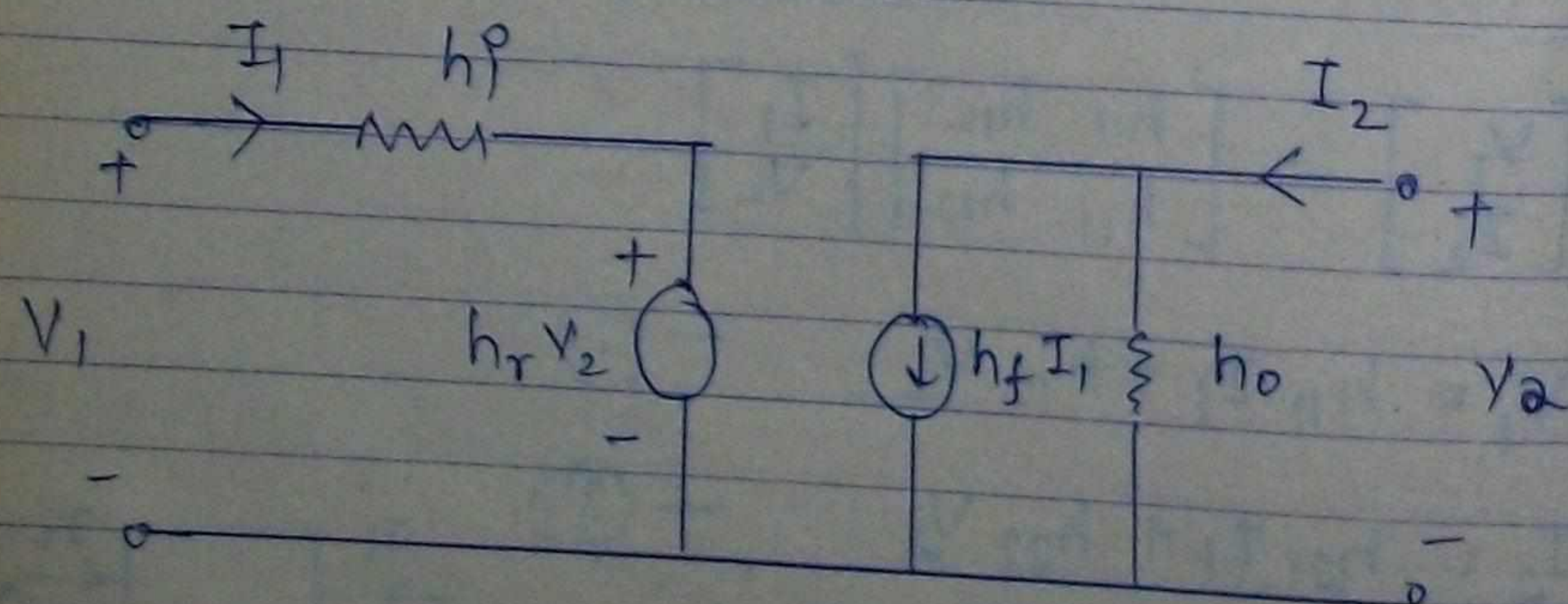
$$h_i^o = h_{11} = \left. \frac{V_1}{I_1} \right|_{V_2=0} = \text{I/P impedance with O/P short circuit (ohm)}$$

$$h_r = h_{12} = \left. \frac{V_1}{V_2} \right|_{I_1=0} = \text{Reverse voltage gain with I/P O.C. (unitless)}$$

$$h_f = h_{21} = \left. \frac{I_2}{I_1} \right|_{V_2=0} = \text{forward current gain with O/P S.C. (unitless)}$$

$$h_o = h_{22} = \left. \frac{I_2}{V_2} \right|_{I_1=0} = \text{Conductance with I/P O.C. (mho)}$$

$$\begin{aligned} V_1 &= h_i^o I_1 + h_r V_2 \\ I_2 &= h_f I_1 + h_o V_2 \end{aligned}$$



$h_{re} V_{ce} = \text{dependent source}$
 $h_{oe} = \text{O/P transconductance}$ } → Due to early effect

The above circuit is called low freq small signal equivalent circuit or h-parameter equivalent circuit of Tx.

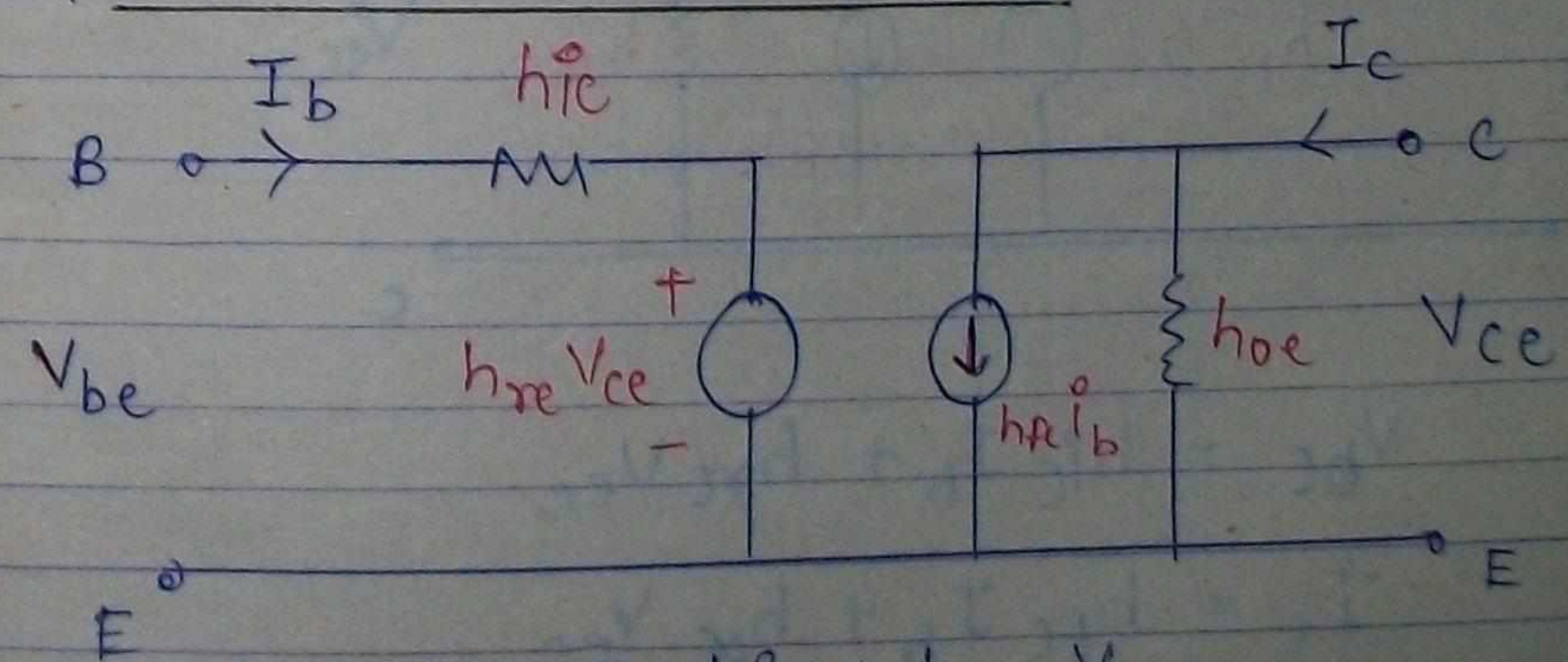
⇒ ADVANTAGE OF h-PARAMETER -

- They are real number at low freq.
- They are graphically obtained from I/P and O/P characteristic of Tx.
- Easy to measured experimentally.

⇒ DISADVANTAGE OF h-PARAMETER -

All the four h-parameters are temperature sensitive.

⇒ FOR COMMON BASE EMITTER -

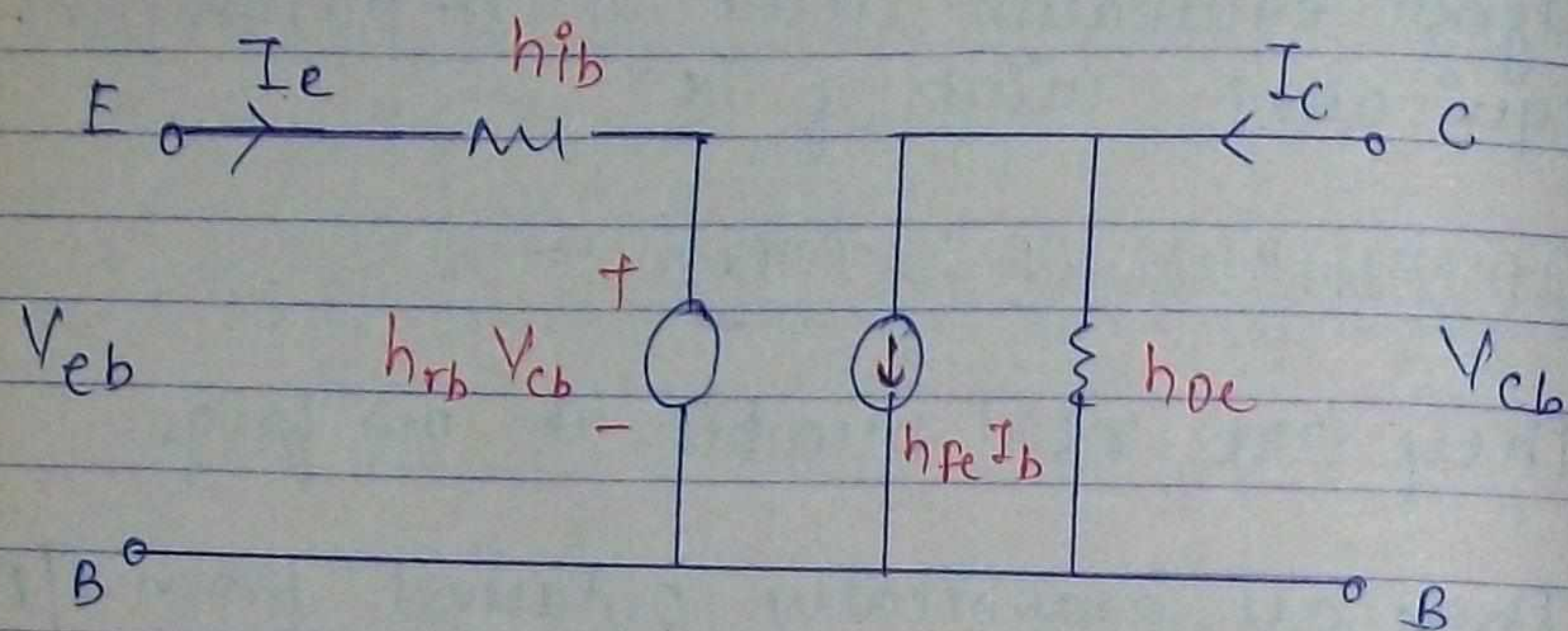


$$V_{be} = I_b h_{ie} + h_{re} V_{ce}$$

$$I_c = h_{fe} I_b + h_{oe} V_{ce}$$

$h_{ob} = \text{least conductance}$

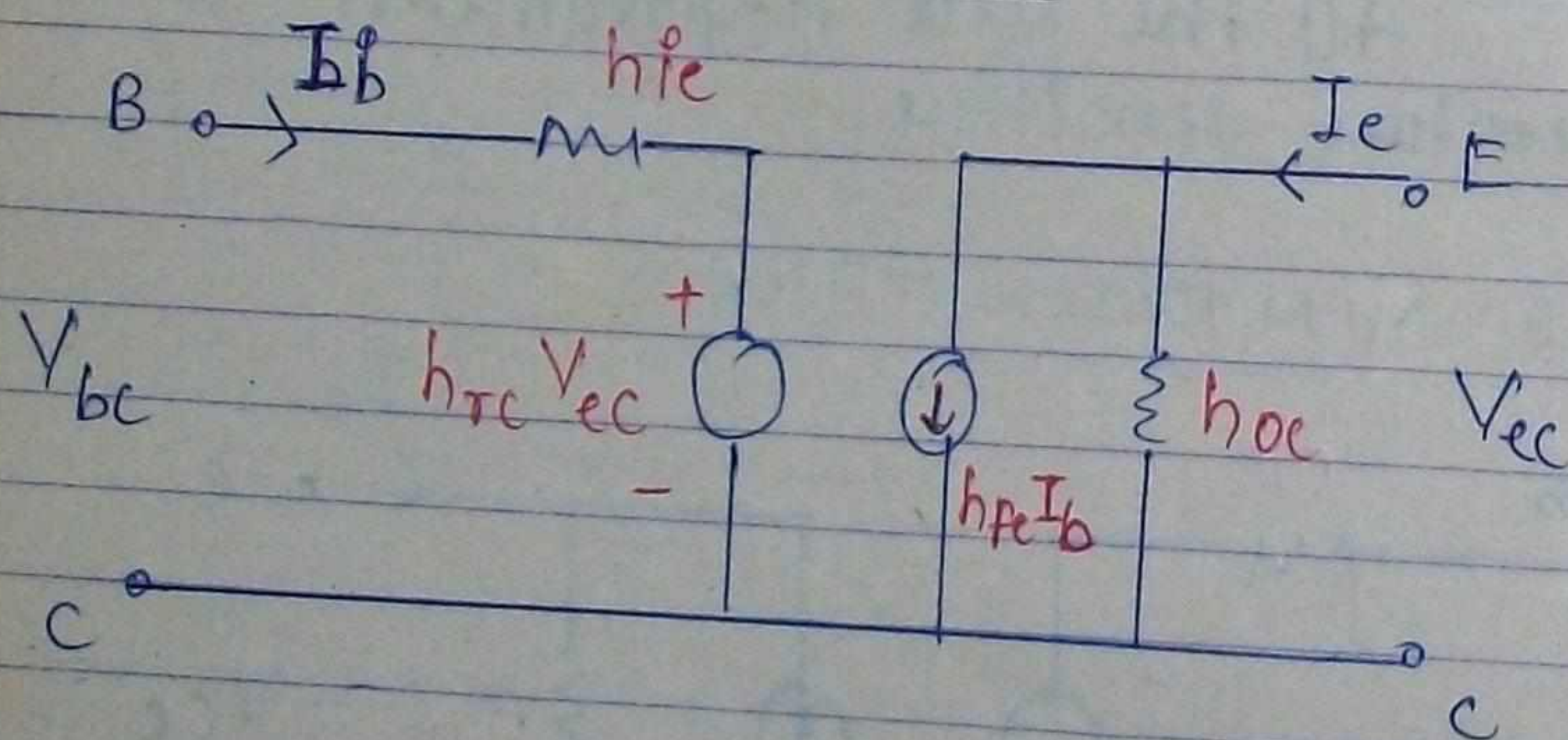
⇒ FOR COMMON BASE -



$$V_{eb} = h_{ib} I_e + h_{rb} V_{cb}$$

$$I_c = h_{fb} I_e + h_{ob} V_{cb}$$

⇒ FOR COMMON COLLECTOR -



$$V_{bc} = h_{ie} I_b + h_{re} V_{ec}$$

$$I_e = h_{fe} I_b + h_{oe} V_{ec}$$

★★

⇒ CONVERSION -

① CC IN TERMS OF CE -

$$h_{ic} = h_{ie}, \quad h_{rc} \approx 1$$

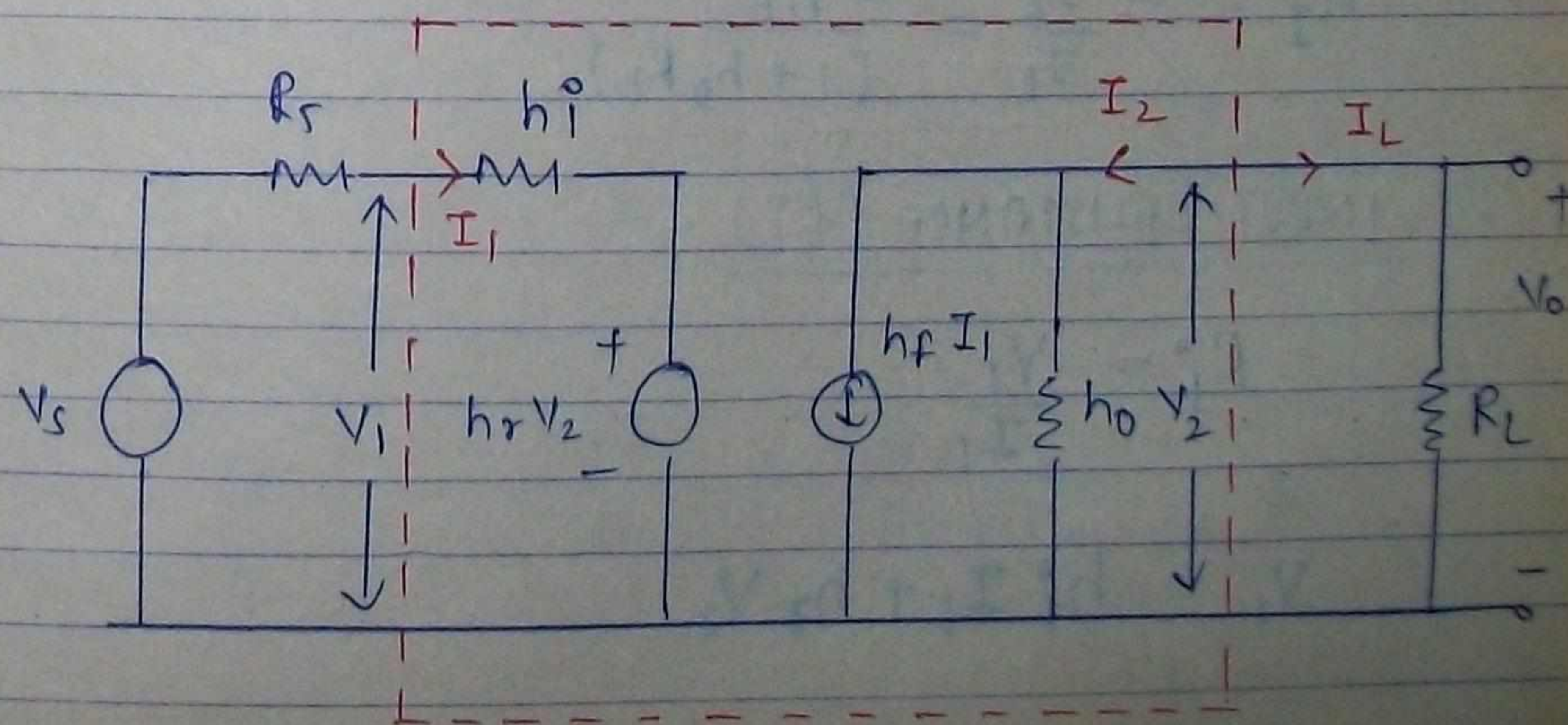
$$h_{fc} = -(1 + h_{fe}), \quad h_{oc} \approx h_{oe}$$

② CB IN TERMS OF CE -

$$h_{ib} = \frac{h_{ie}}{1 + h_{fe}}, \quad h_{rb} = \frac{h_{ie} h_{oe}}{1 + h_{fe}} - h_{re}$$

$$h_{fb} = \frac{h_{fe}}{1 + h_{fe}}, \quad h_{ob} = \frac{h_{oe}}{1 + h_{fe}}$$

⇒ DERIVATION OF $A_I, R_i, A_v, A_{vs}, A_{Is}, R_o$ -



① CURRENT GAIN (A_I) -

$$A_I = \frac{-I_2}{I_1} = \frac{I_L}{I_1}$$

$$I_2 = h_f I_1 + h_o V_2$$

$$V_2 = V_o = I_L R_L = -I_2 R_L$$

$$I_2 = h_f I_1 - h_o R_L I_2$$

$$I_2 = I_1 h_f - h_o R_L I_2$$

$$h_f I_1 = I_2 (1 + h_o R_L)$$

$$\frac{I_2}{I_1} = \frac{h_f}{(1 + h_o R_L)}$$

$$A_I = \frac{-I_2}{I_1} = \frac{-h_f}{(1 + h_o R_L)}$$

② INPUT RESISTANCE (R_i°) -

$$R_i^\circ = \frac{V_1}{I_1}$$

$$V_1 = h_i^\circ I_1 + h_r V_2$$

$$V_o = -I_2 R_L = A_I I_1 R_L$$

$$V_1 = h_i^\circ I_1 + h_r A_I R_L I_1$$

$$R_i^\circ = \frac{V_1}{I_1} = h_i^\circ + h_r A_I R_L$$

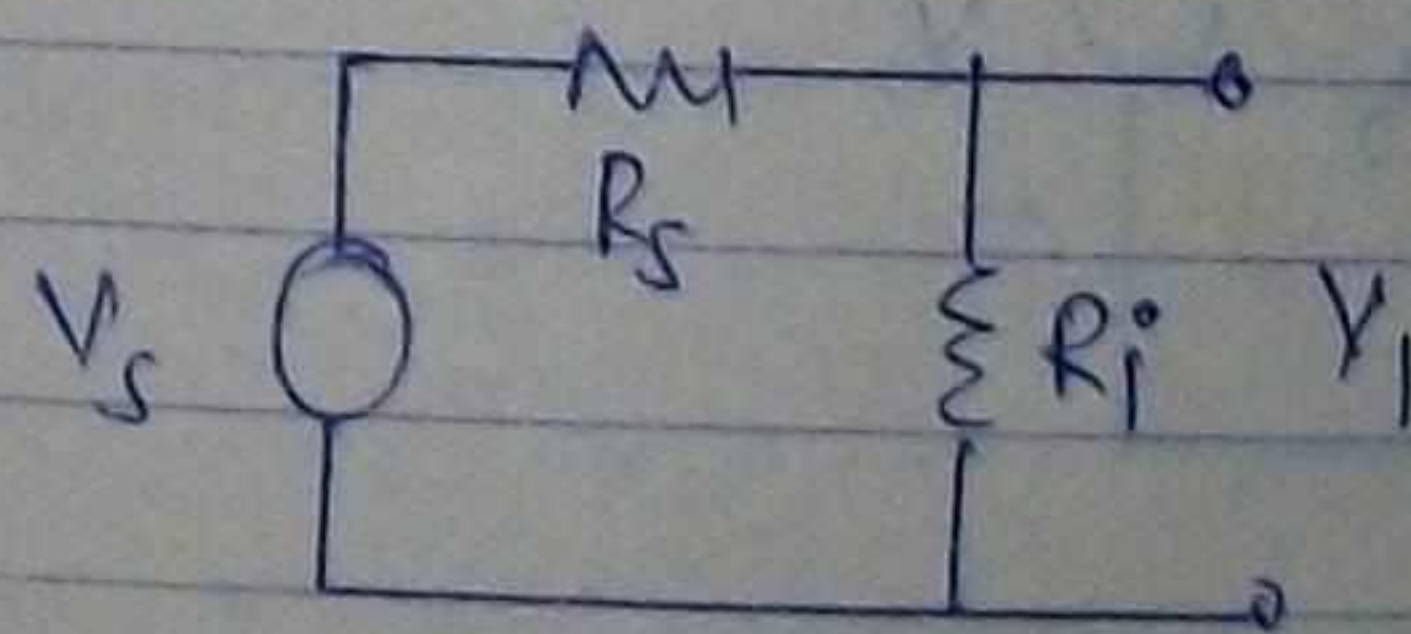
③ VOLTAGE GAIN (A_V) -

$$A_V = \frac{V_2}{V_1} = \frac{-I_2 R_L}{I_1 R_i^\circ} = A_I \frac{R_L}{R_i^\circ}$$

$$A_V R_i^\circ = A_I R_L$$

④ OVERALL VOLTAGE GAIN (A_{Vs}) -

$$A_{Vs} = \frac{V_o}{V_s} = \frac{V_o}{V_1} \times \frac{V_1}{V_s}$$

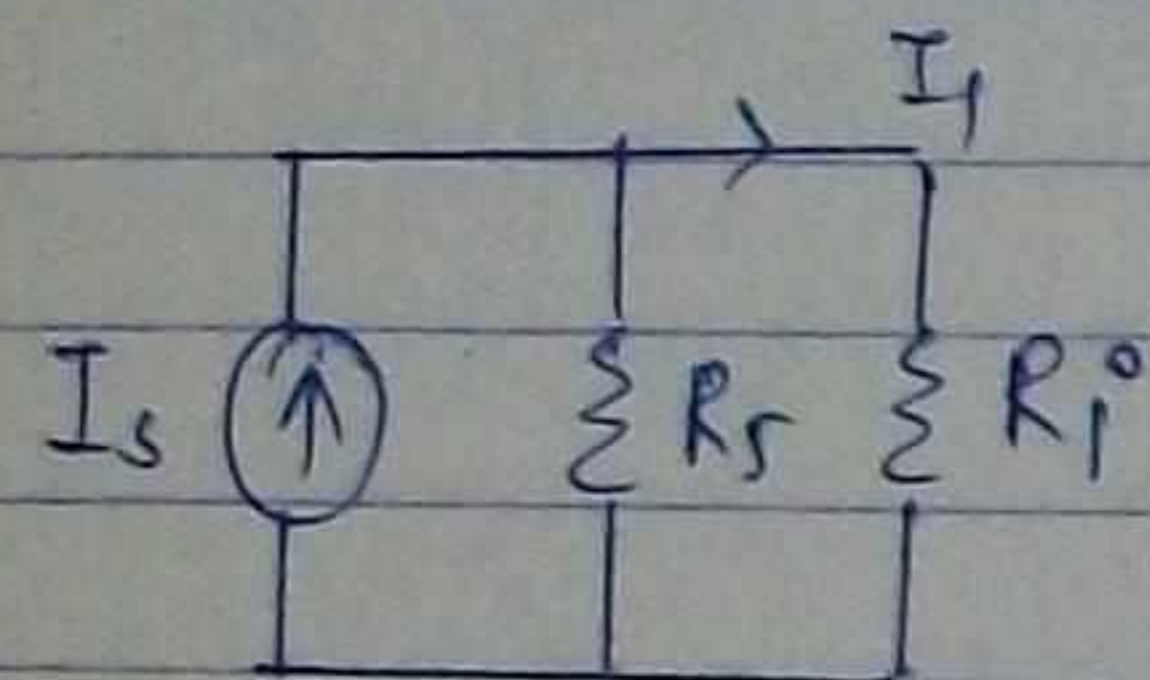


$$\frac{V_1}{V_s} = \frac{R_i^\circ}{R_i^\circ + R_s}$$

$$A_{Vs} = \frac{A_V R_i^\circ}{R_i^\circ + R_s} = \frac{A_I R_L}{R_i^\circ + R_s}$$

⑤ OVERALL CURRENT GAIN (A_{Is}) -

$$A_{Is} = \frac{I_o}{I_s} = -\frac{I_2}{I_s} = -\frac{I_2}{I_1} \times \frac{I_1}{I_s}$$

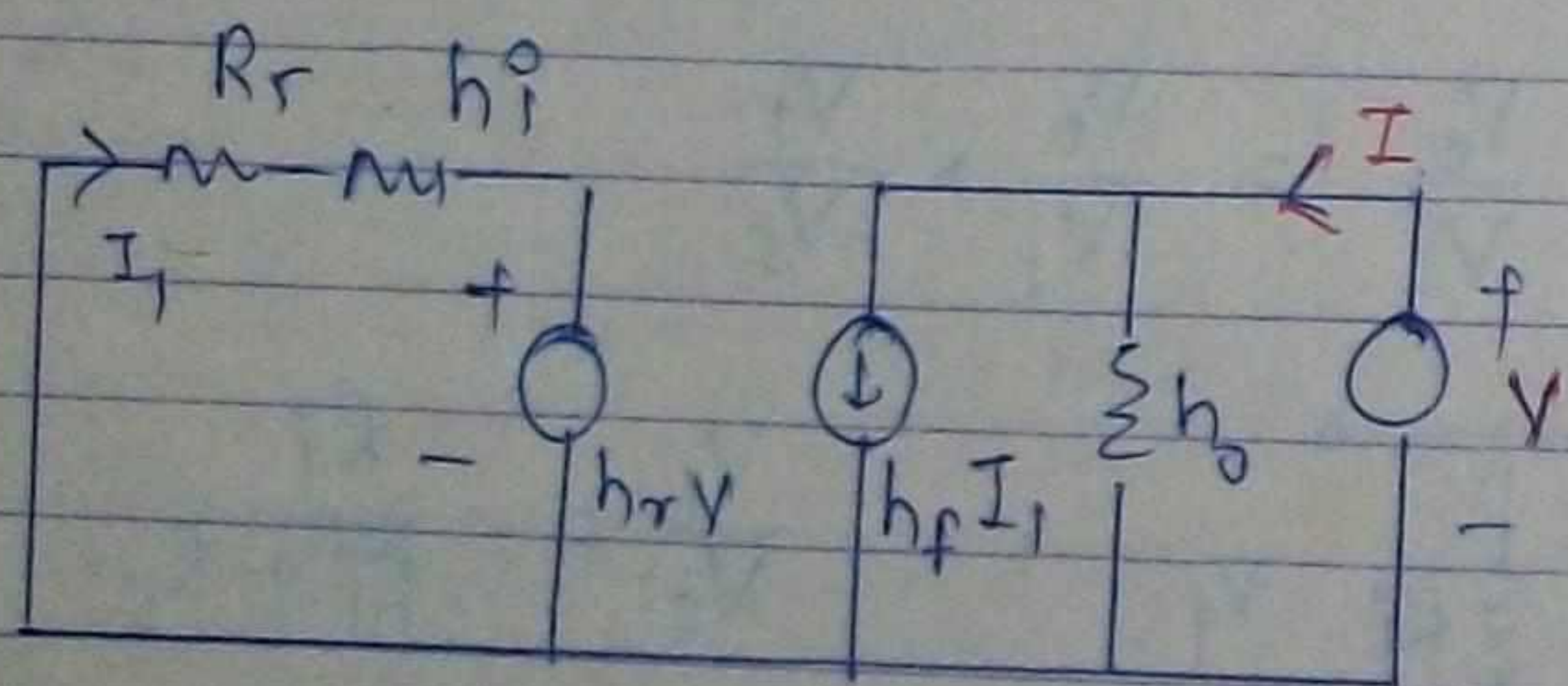


$$\frac{I_1}{I_s} = \frac{R_s}{R_s + R_i}$$

$$A_{Is} = A_I \frac{R_s}{R_s + R_i}$$

⑥ OUTPUT RESISTANCE (R_o) -

$$R_L = \infty, V_s = 0$$



FIRST METHOD - $Y_o = I/V$

$$I = h_f I_1 + h_o V$$

$$I_1 (R_s + h_i) + h_r V = 0$$

$$I_1 = -\frac{h_r V}{R_s + h_i}$$

$$I_1 = -\frac{h_r V}{R_s + h_i}$$

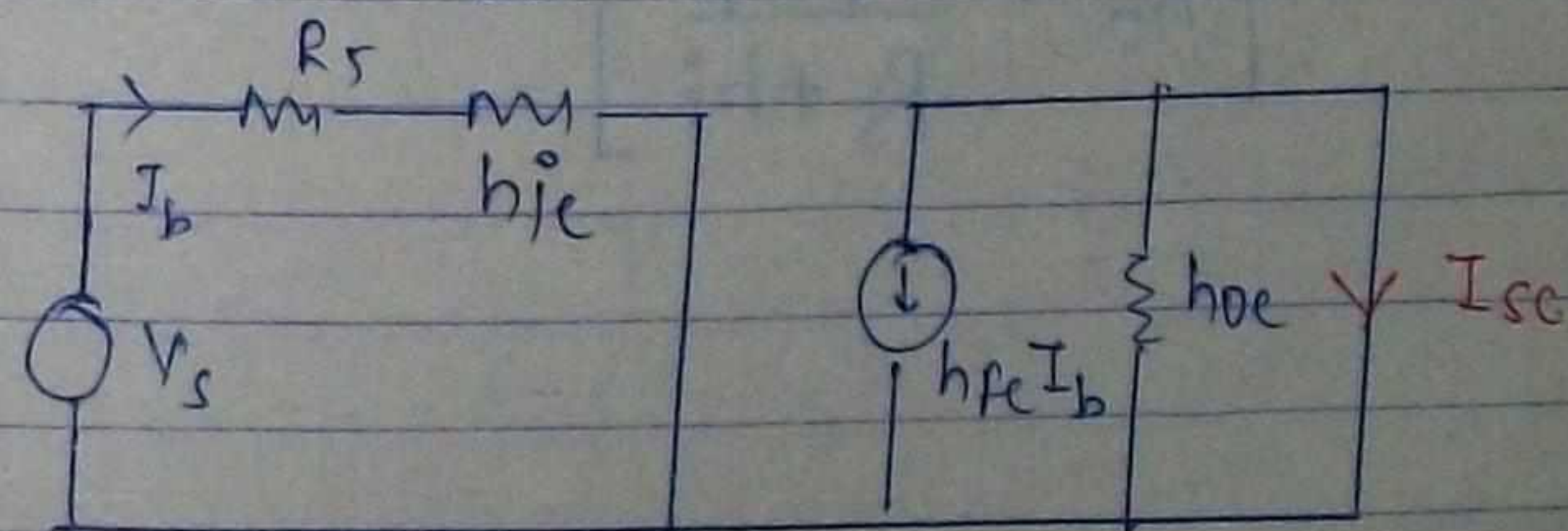
$$I = -\frac{h_f h_r V + h_o V}{R_s + h_i}$$

$$Y_o = \frac{I}{V} = h_o - \frac{h_r h_f}{R_s + h_i}$$

$$R_o = \frac{1}{Y_o}$$

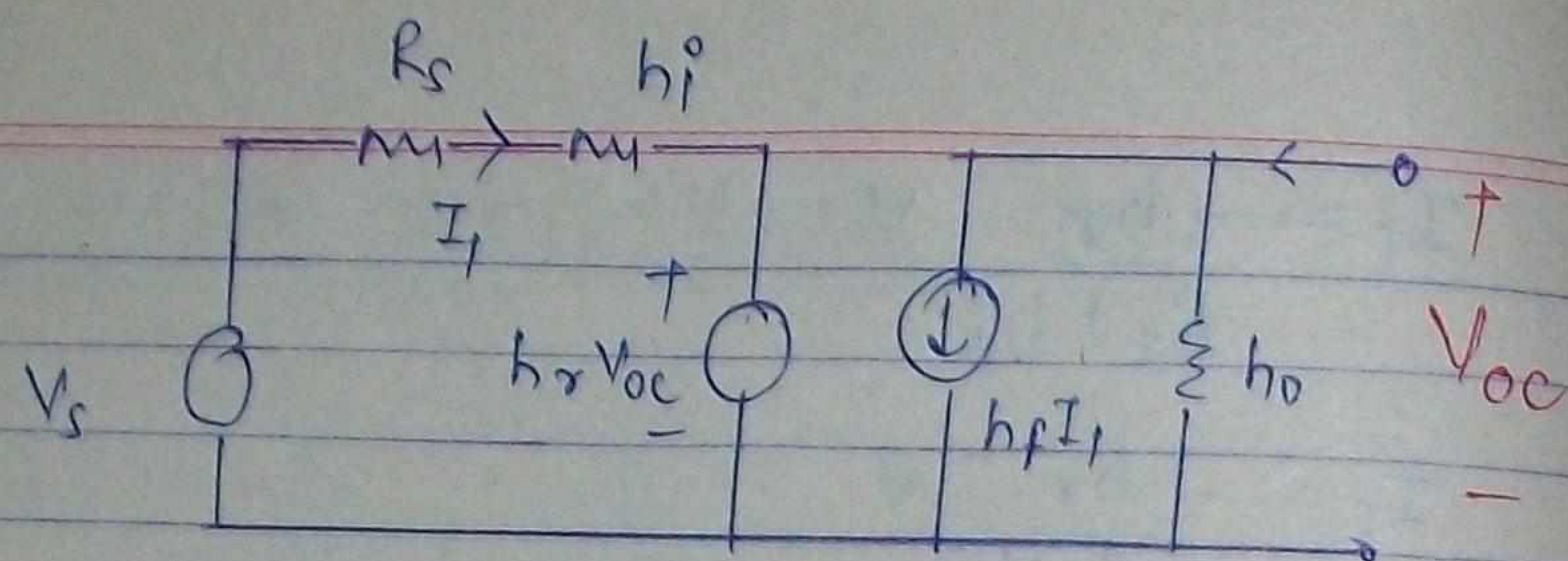
IInd METHOD -

$$Y_o = \frac{I_{sc}}{V_{oc}} \quad R_o = \frac{V_{oc}}{I_{sc}}$$



$$I_{sc} = -h_{fe} I_b = -\frac{h_{fe} V_s}{R_s + h_{ie}}$$

$$I_{sc} = -\frac{h_f}{R_s + h_i} \quad \text{--- (1)}$$



$$V_{oc} = -\frac{h_f I_i}{h_o}$$

$$V_{oc} = -\frac{h_f}{h_o} \left[\frac{V_s - h_r V_{oc}}{R_s + h_i} \right] \quad \text{--- (2)}$$

Since, $I_i = \frac{V_s - h_r V_{oc}}{R_s + h_i}$

$$\textcircled{2}/\textcircled{1} \quad R_o = \frac{V_{oc}}{I_{sc}}$$

$$R_o = \frac{1}{\left[h_o - \frac{h_r h_f}{R_s + h_i} \right]}$$

$\Rightarrow$ SIMPLIFIED HYBRID MODEL - (NEGLECTING EARLY EFFECT)

$$h_{ie} = 1.1k\Omega, \quad h_{re} = 2.5 \times 10^{-4}$$

$$h_{fe} = 50, \quad h_{oe} = 1/40k$$

$$1/h_{oe} = 40k$$

Standard value for CE.

Exact value of CE - configuration -

$$A_I = \frac{-h_{fe}}{1 + h_{oe} R_L}$$

$$R_i = h_{ie} + h_{re} A_I R_L$$

$$A_V = A_I \frac{R_L}{R_i}$$

$$A_{Vs} = \frac{A_V R_i}{R_i + R_s} = \frac{A_I R_L}{R_i + R_s}$$

$$A_{Is} = \frac{A_I R_s}{R_i + R_s}$$

$$Y_o = h_{oe} - \frac{h_{re} h_{fe}}{R_s + h_{ie}}$$

$$R_o = 1/Y_o$$

But here early effect is neglected

s. $h_{oe} = 0$ $h_{re} = 0$

only when,

$$h_{oe} R_L \leq 0.1$$

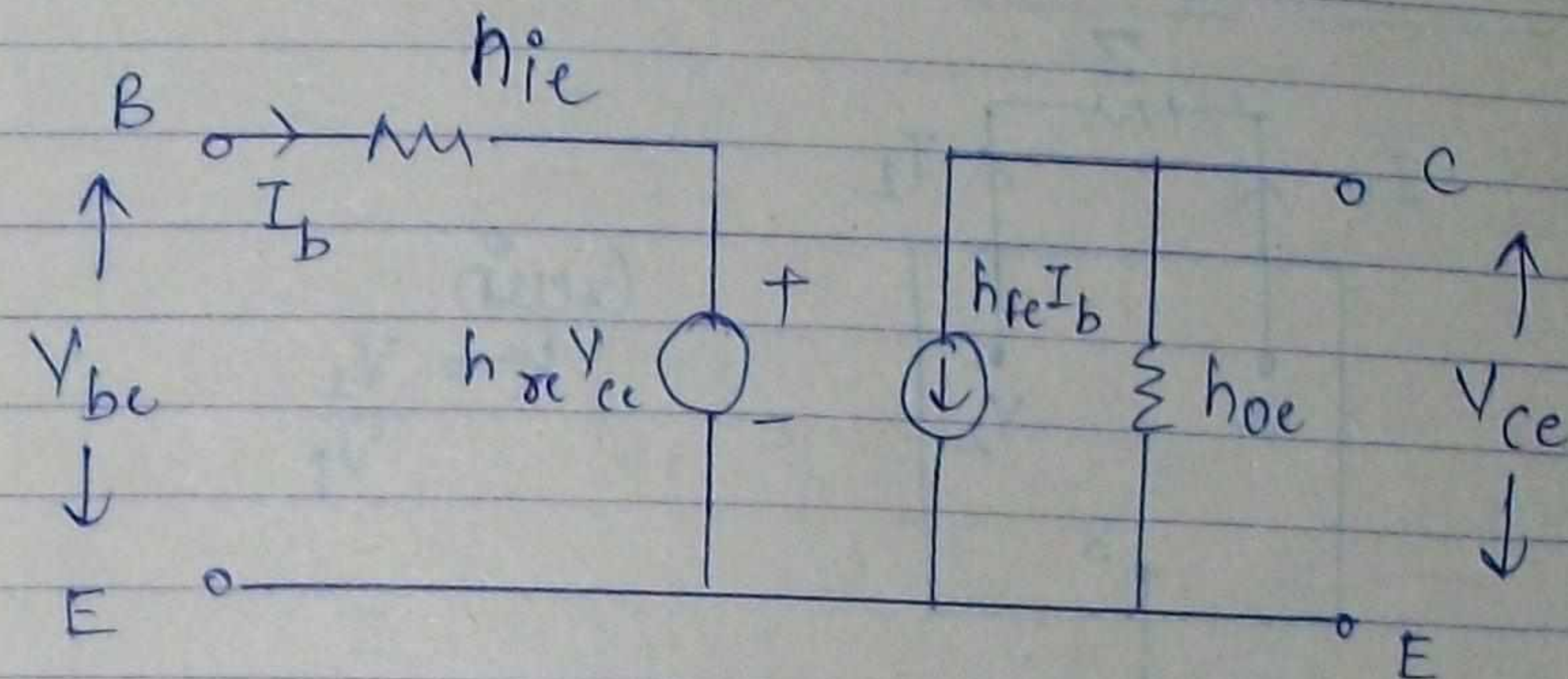
then error is less than 10%.

Now let $h_{oe} R_L = 0.1$

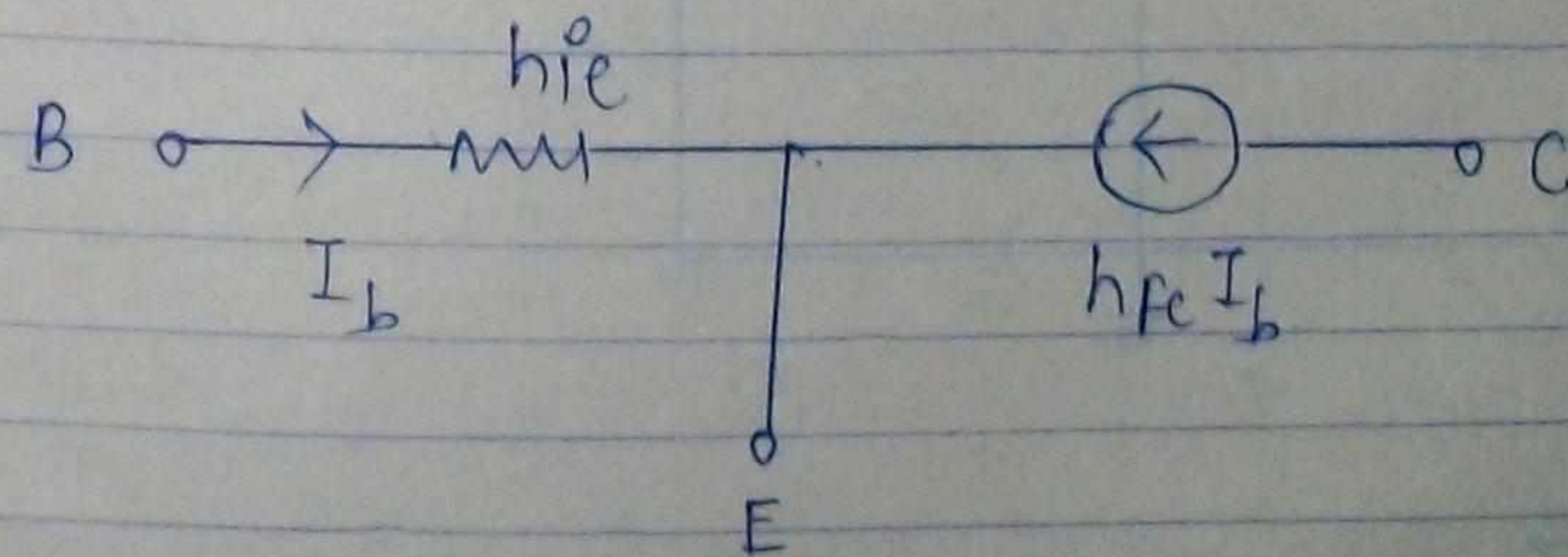
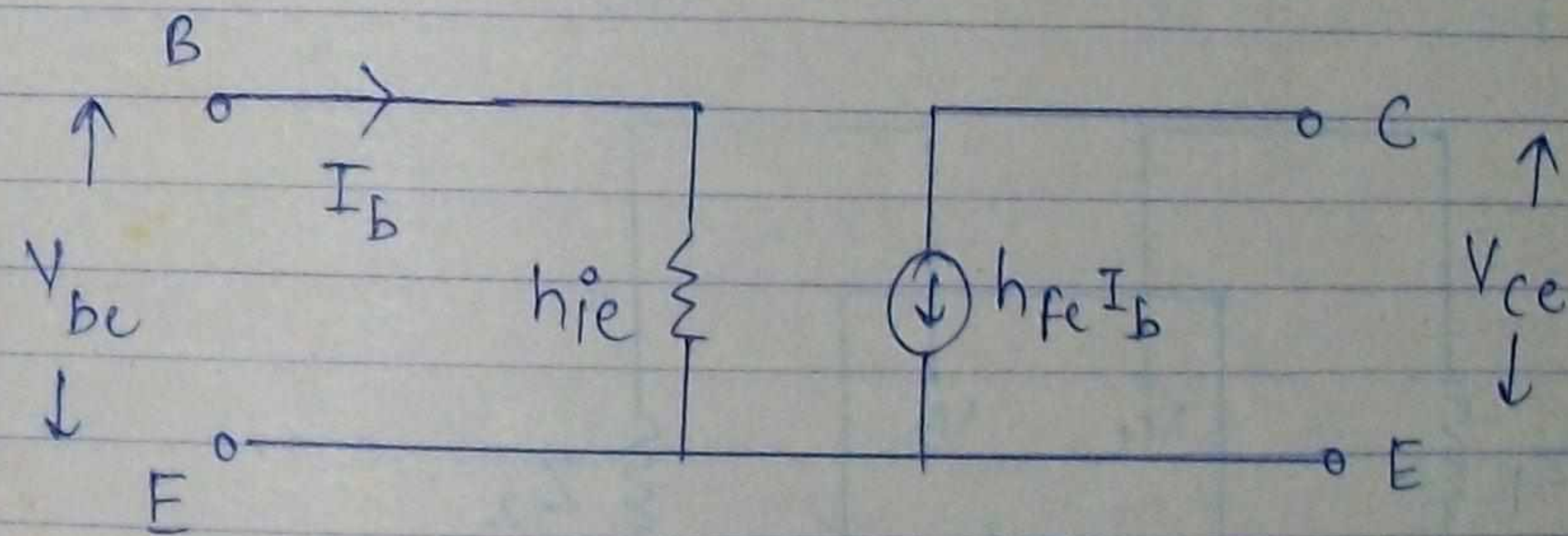
$$A_I = -\frac{h_{fe}}{1+h_{oe}R_L} = -\frac{h_{fe}}{1.1} \quad (\text{exact value})$$

EXACT	APPROXIMATE	RESULT
① $A_I = -\frac{h_{fe}}{1+h_{oe}R_L}$ $= -\frac{h_{fe}}{1.1}$	$A_I = -h_{fe}$	A_I is overestimated by 10%.
② $R_i = h_{ie} + h_{re} A_I R_L$ $= h_{ie} + h_{re} A_I R_L$	$R_i = h_{ie}$	R_i is overestimated by 5%
③ $A_v = \frac{A_I R_L}{R_i}$	$\frac{dA_v}{A_v} = \frac{dA_I}{A_I} + \frac{dR_L}{R_L} - \frac{dR_i}{R_i}$	$= 10 + 10 - 5 = 5\%$ A_v is overestimated by 5%.

⇒ PRACTICAL MODEL -



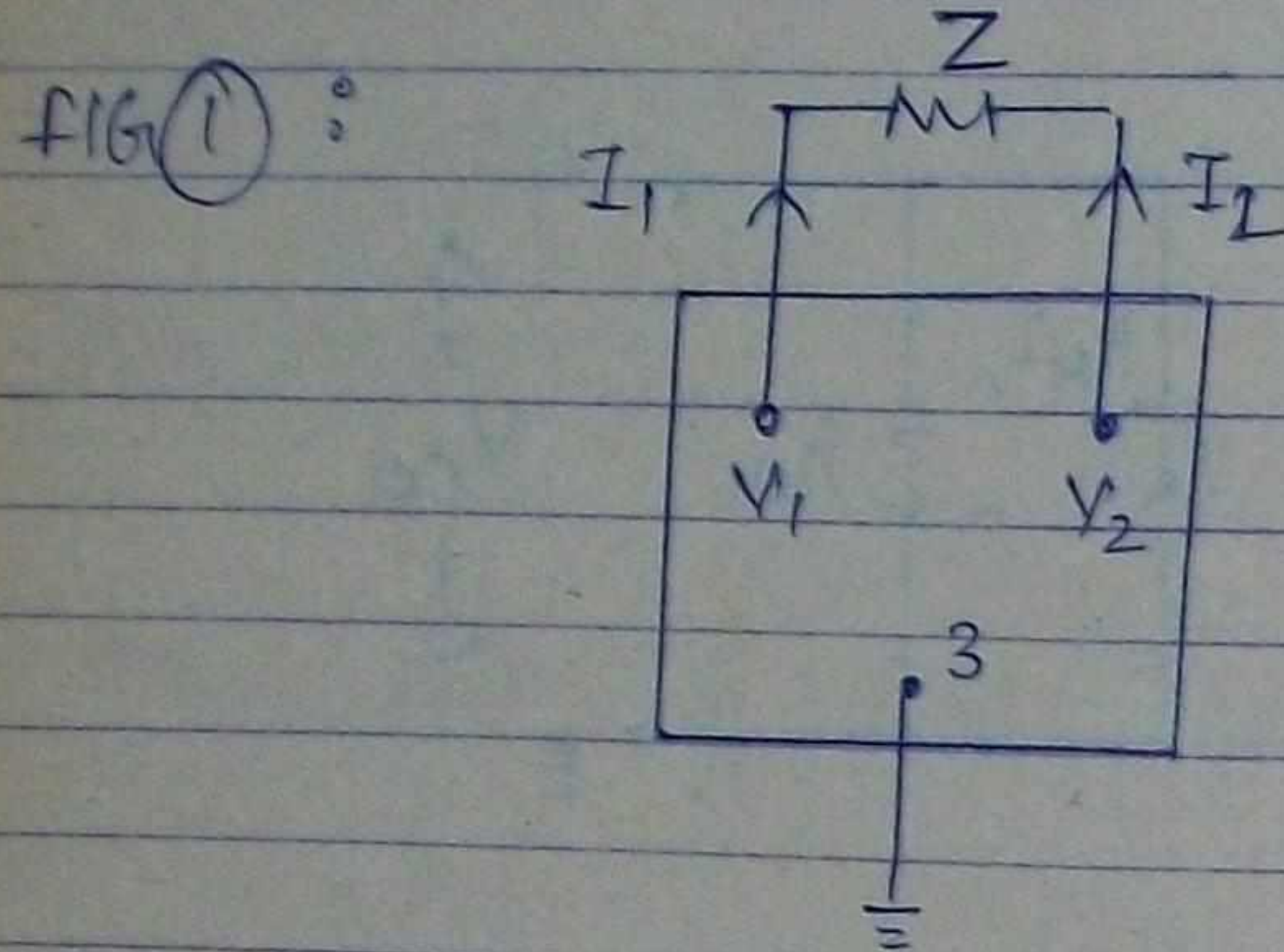
Since $h_{oe} R_L \leq 0.1$ $h_{oe} = h_{re} = 0$



This is valid for CE, CC, CB for both NPN & PNP

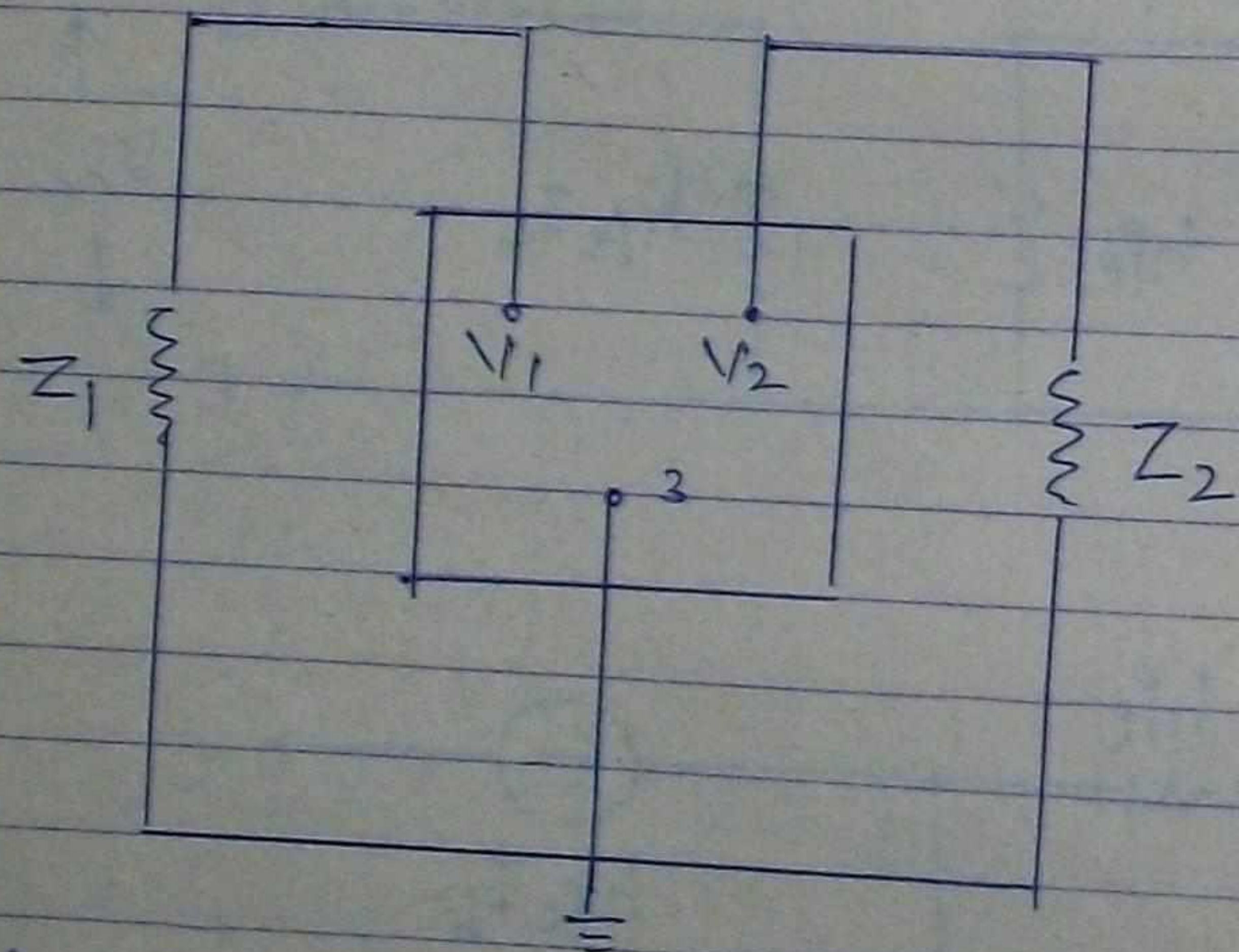
$\left. \begin{array}{l} \text{I/P Resistance decreases} \\ \text{I/P Capacitance increases} \\ \text{I/P Inductance decreases} \end{array} \right\} \text{ due to miller theorem}$

$\Rightarrow$ MILLER'S THEOREM -



Gain
 $K = \frac{V_2}{V_1}$

FIG (2) :



where

$$Z_1 = \frac{Z}{1-K}$$

$$Z_2 = \frac{Z}{1-1/K}$$

FIG (1)

$$I_1 = \frac{V_1 - V_2}{Z} = \frac{V_1}{Z_1}$$

$$Z_1 = \frac{Z V_1}{V_1 - V_2}$$

$$Z_1 = \frac{Z}{1 - V_2/V_1}$$

$$Z_1 = \frac{Z}{1-K}$$

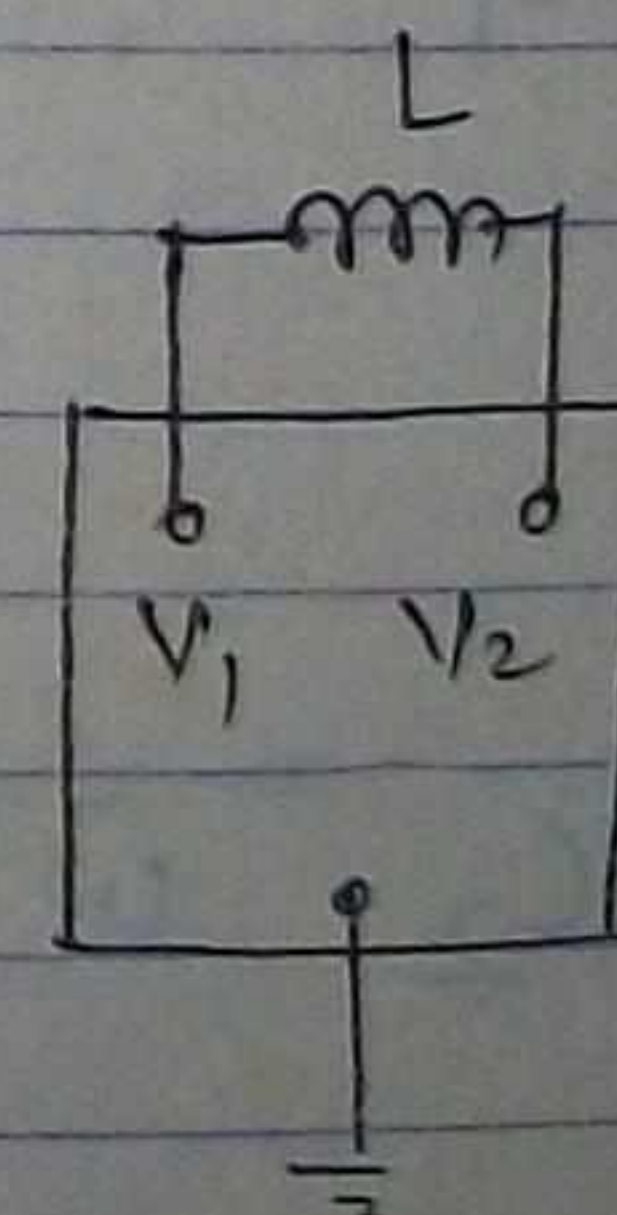
$$I_2 = \frac{V_2 - V_1}{Z} = \frac{V_2}{Z_2}$$

$$Z_2 = \frac{Z V_2}{V_2 - V_1}$$

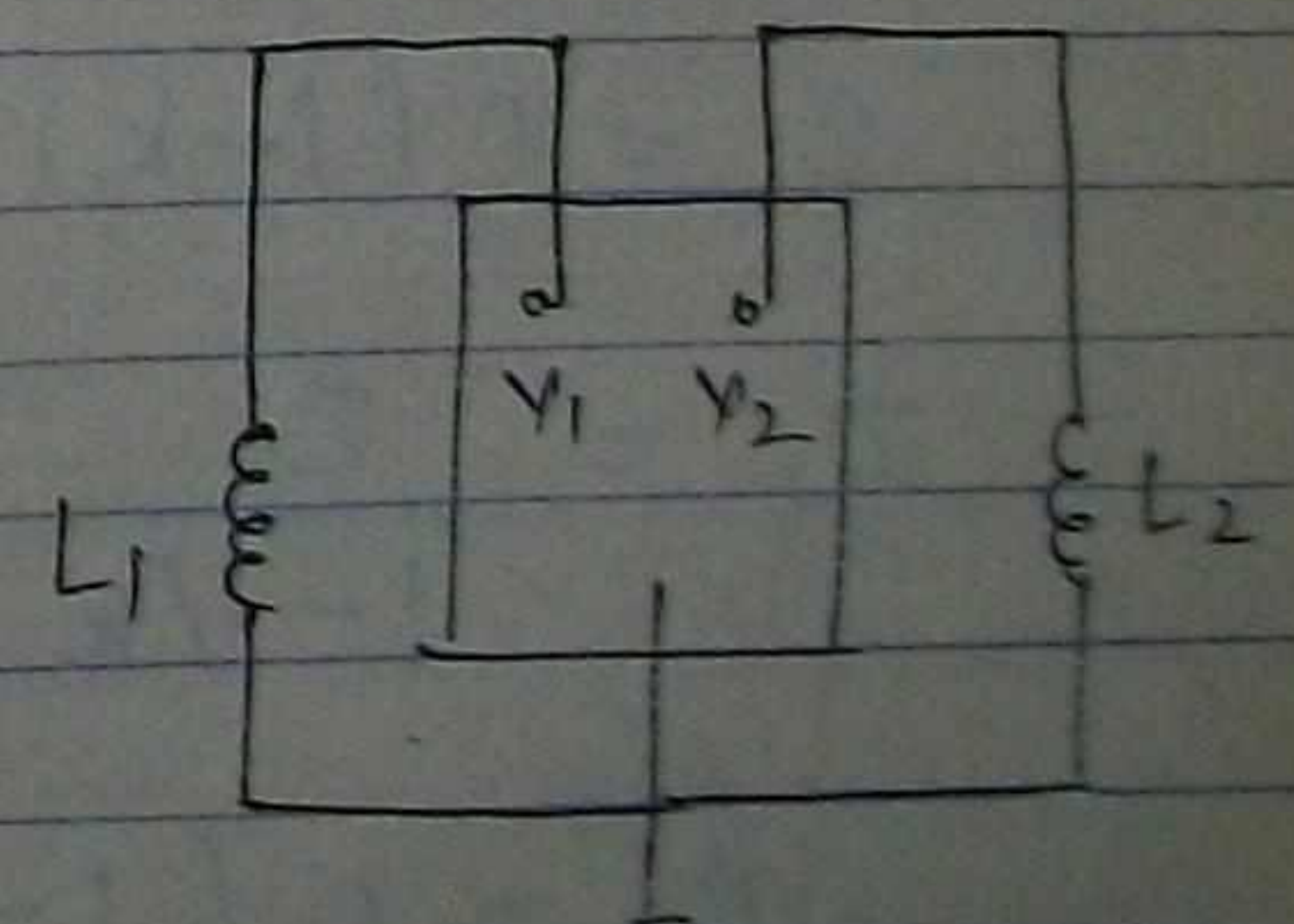
$$Z_2 = \frac{Z}{1 - V_1/V_2}$$

$$Z_2 = \frac{Z}{1 - 1/K}$$

Ques



$\Rightarrow$



find L_1 & L_2 when $K = V_2/V_1$

Solⁿ: $Z_1 = \frac{Z}{1-k}$

$$j\omega L_1 = \frac{j\omega L}{1-k}$$

$$L_1 = \frac{L}{1-k}$$

similarly $Z_2 = \frac{Z}{1-1/k}$

$$j\omega L_2 = \frac{j\omega L}{1-1/k}$$

$$L_2 = \frac{L}{1-1/k}$$

Ques. Previous question when L is replaced by C then find C_1 and C_2

Solⁿ:

$$Z_C = 1/\omega C$$

$$Z_1 = \frac{Z}{1-k}$$

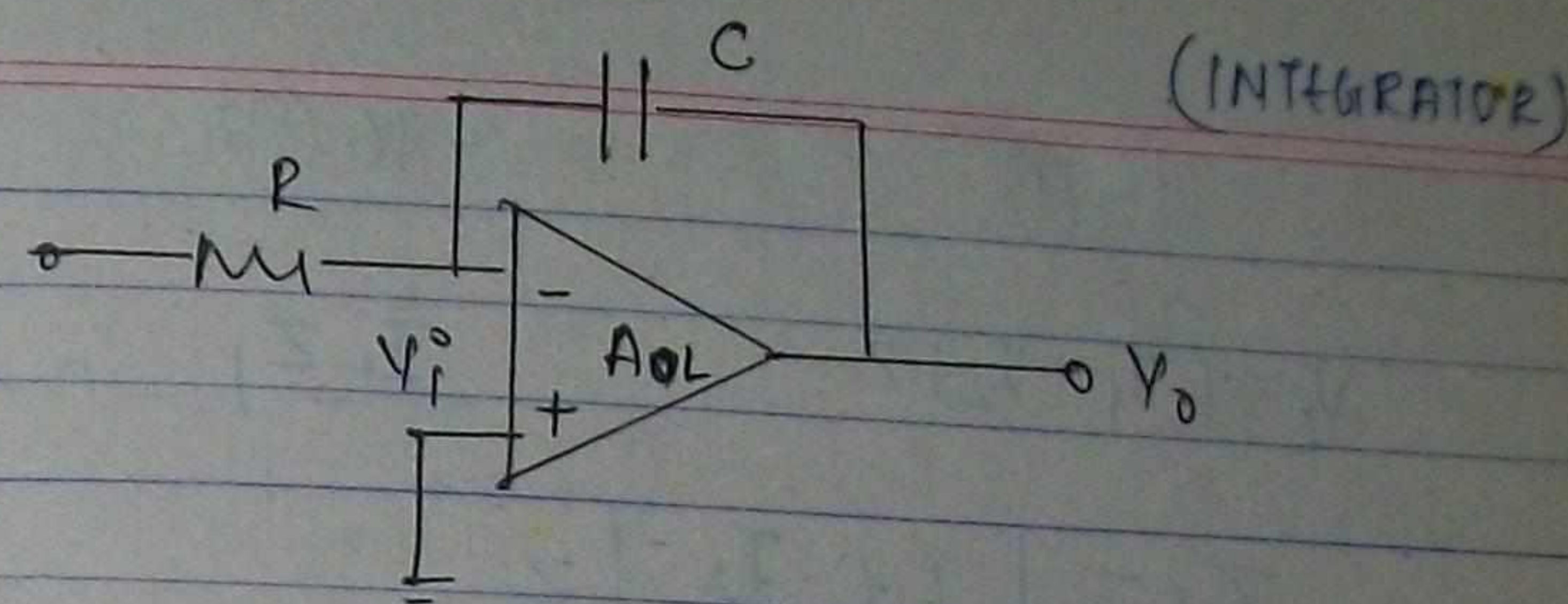
$$\frac{1}{\omega C_1} = \frac{1/\omega C}{1-k} \Rightarrow \frac{1}{C_1} = \frac{1}{C(1-k)}$$

$$C_1 = C(1-k)$$

$$Z_2 = \frac{Z}{1-1/k} = \frac{1}{\omega C_2} = \frac{1}{\omega C(1-1/k)}$$

$$C_2 = C\left(1 - \frac{1}{k}\right)$$

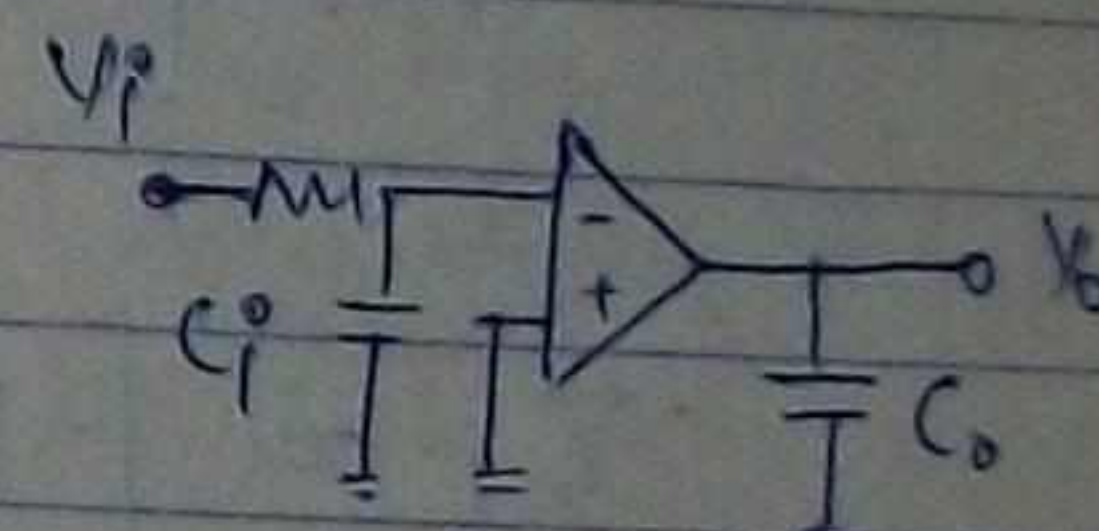
Ques.



Solⁿ:

$$C_i^o = C(1-A_{OL})$$

$$C_{o/p} = C\left(1 - \frac{1}{A_{OL}}\right)$$



$\Rightarrow$ DUAL OF MILLER'S THEOREM -

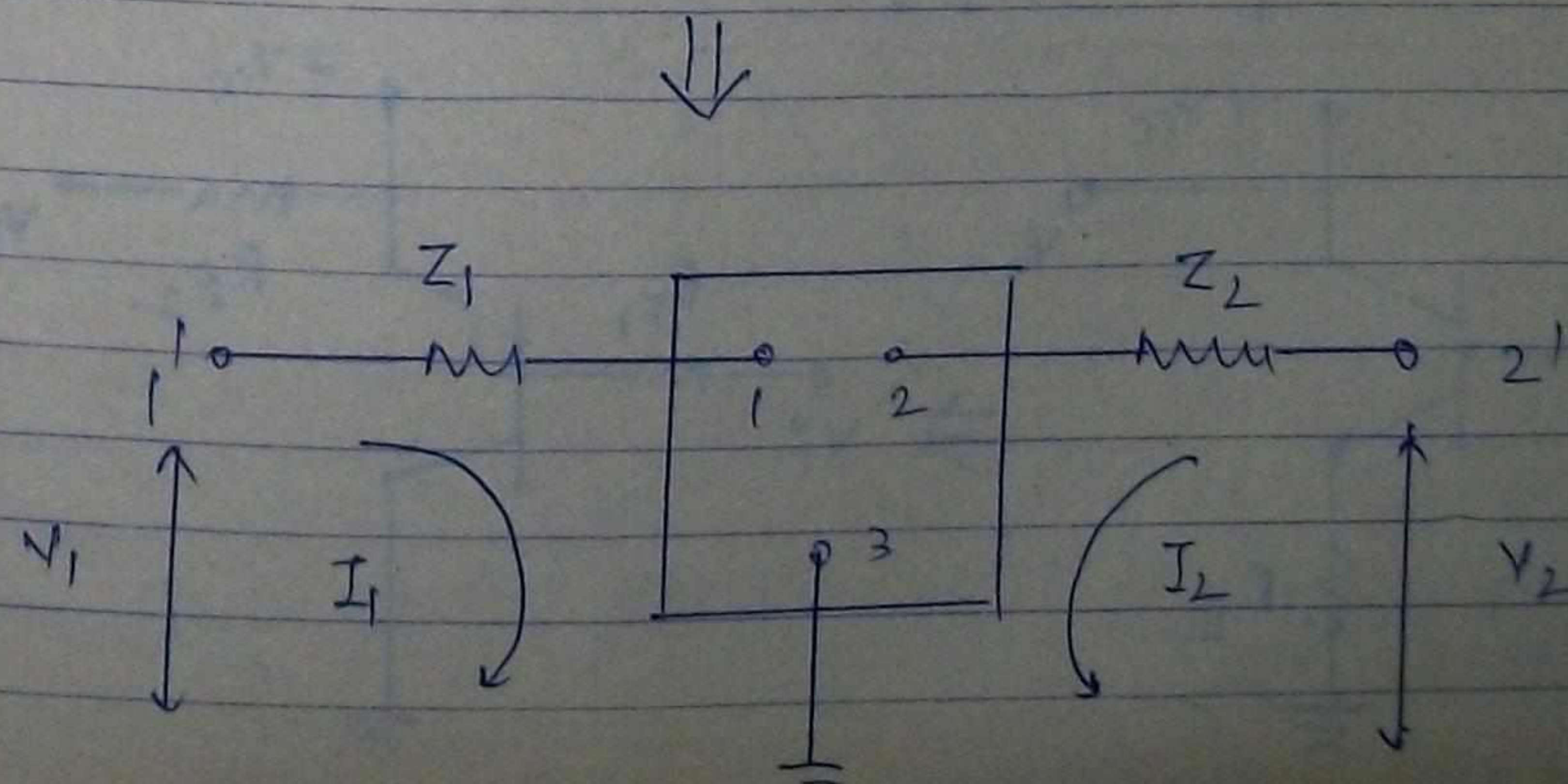
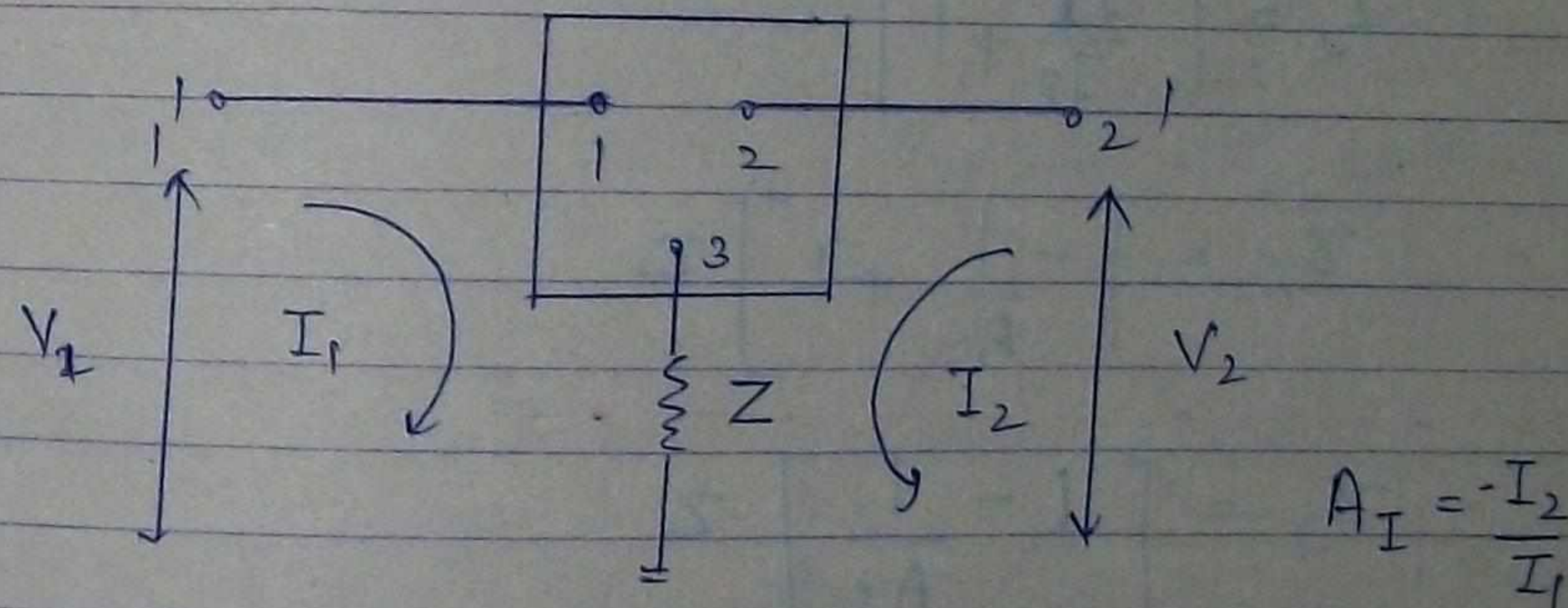


FIG (1)

$$V_1 = (I_1 + I_2)Z = I_1 Z_1$$

$$Z_1 = \left[1 + \frac{I_2}{I_1} \right] Z$$

$$Z_1 = [1 - A_I] Z$$

$$V_2 = (I_1 + I_2)Z = I_2 Z_2$$

$$Z_2 = \left[\frac{I_1}{I_2} + 1 \right] Z$$

$$Z_2 = \left[-\frac{1}{A_2} + 1 \right] Z$$

$$Z_2 = \left[1 - \frac{1}{A_I} \right] Z$$

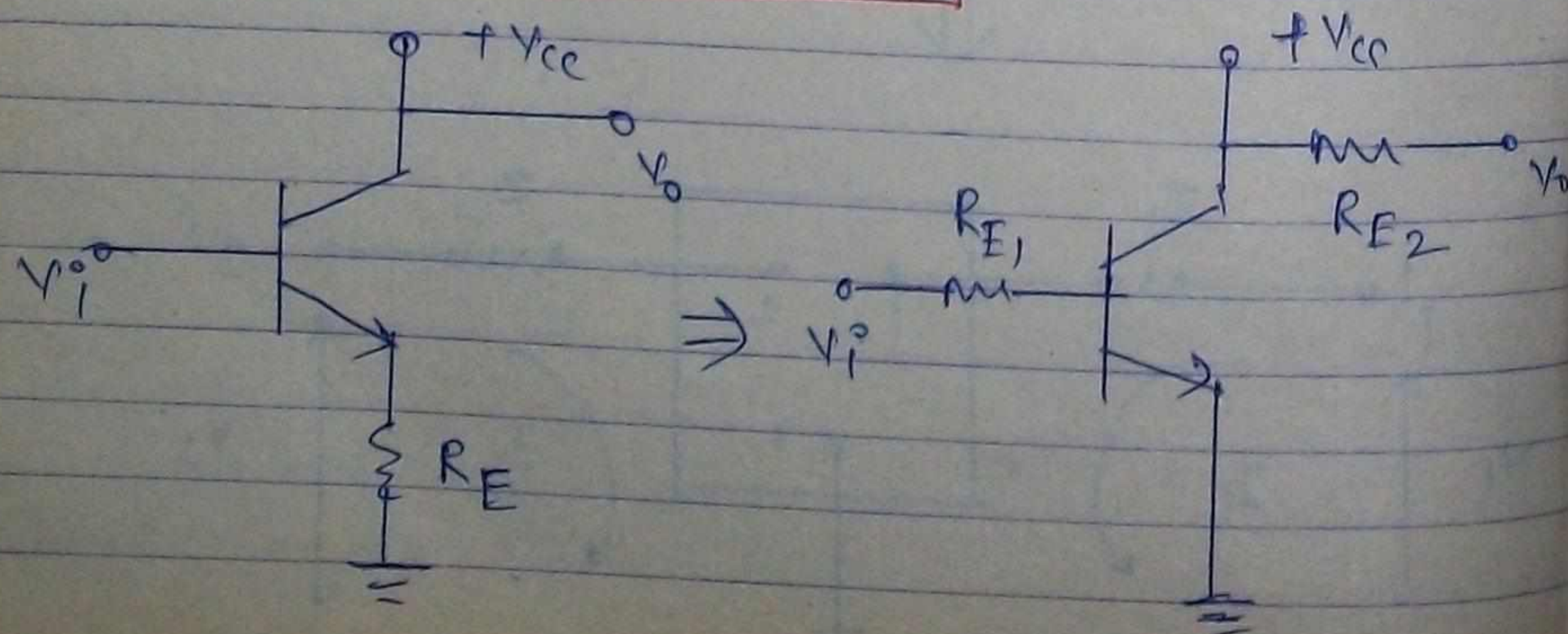
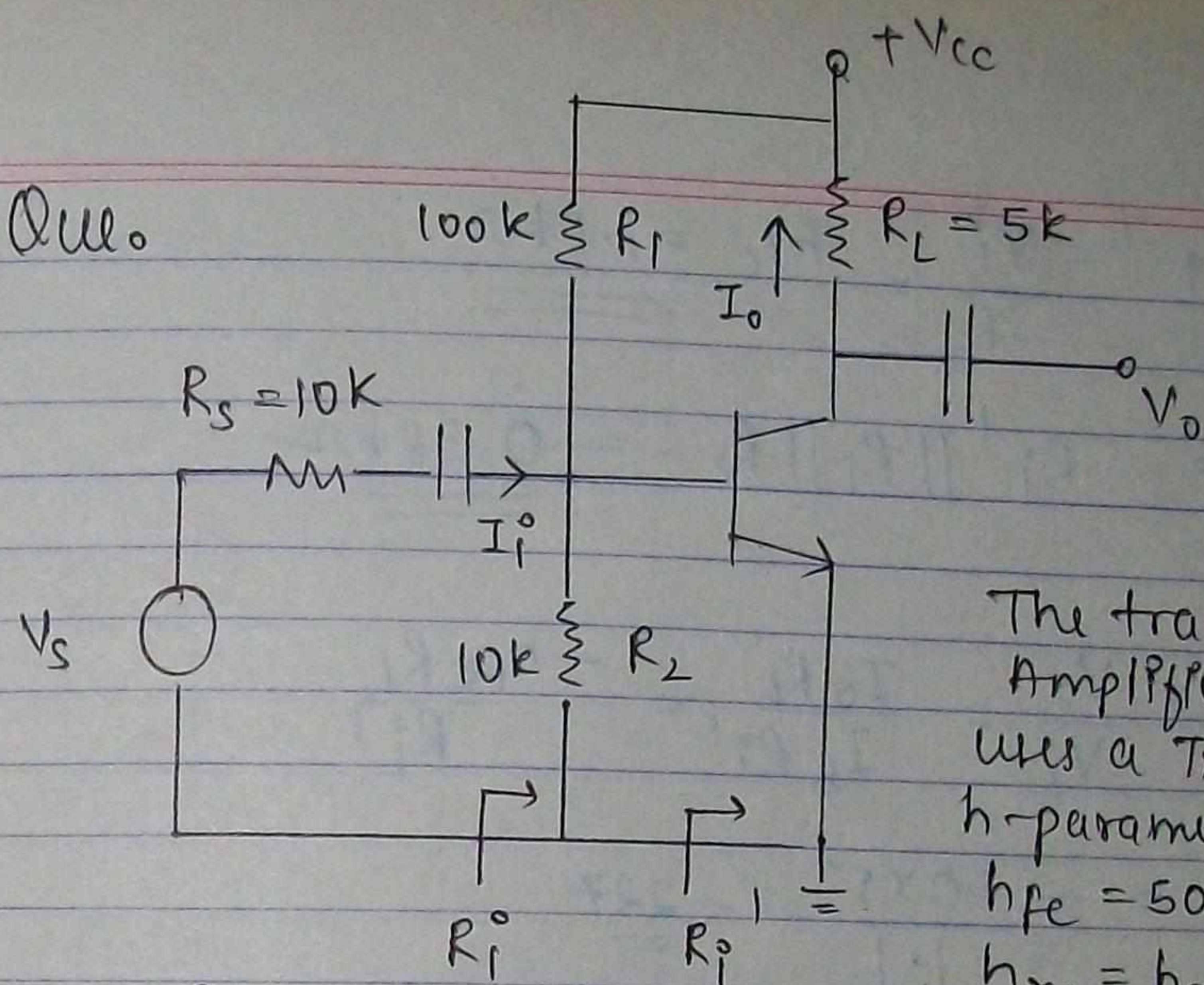


FIG (2)

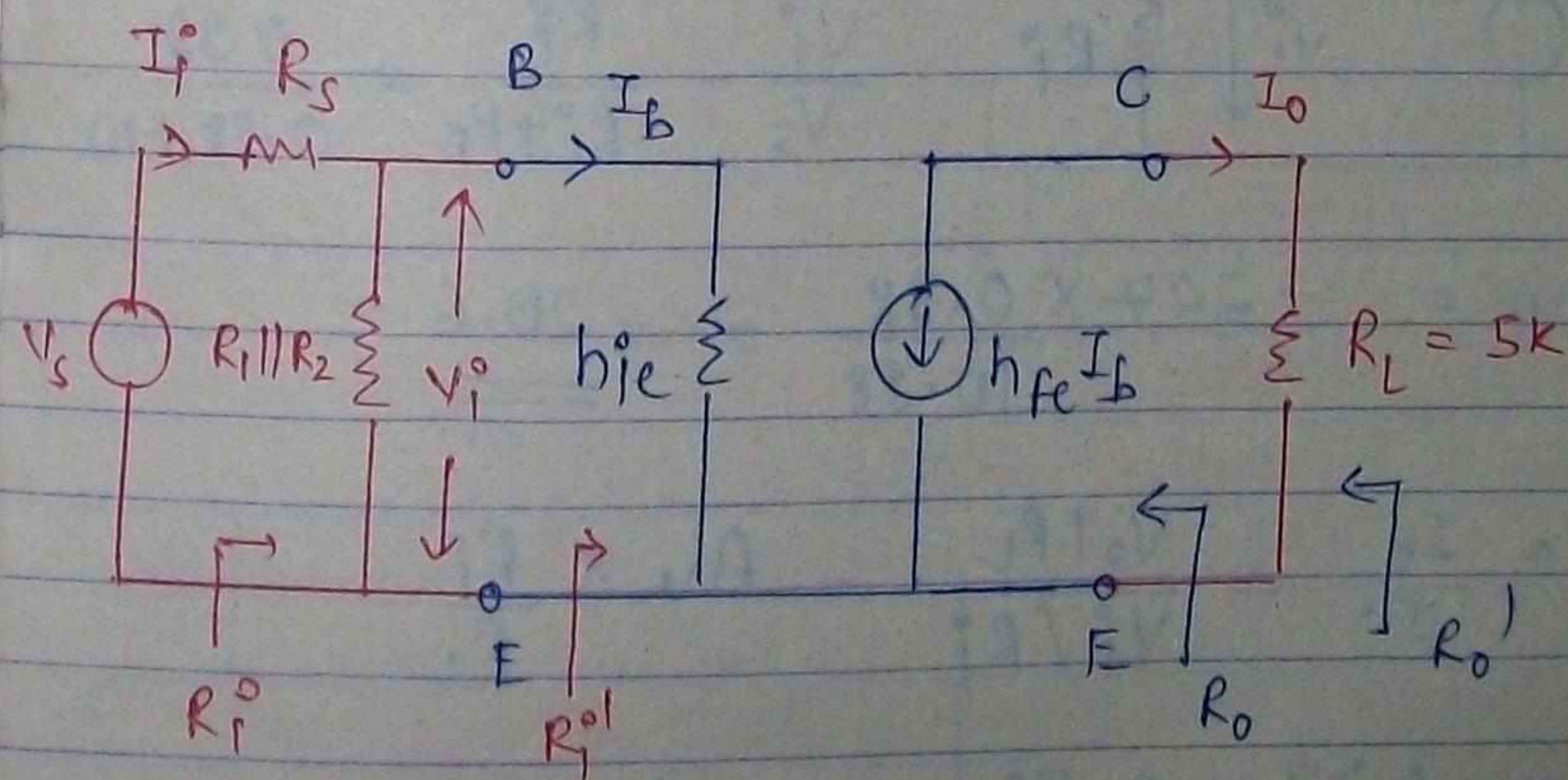
Ques.



The transistor Amplifier shown uses a Tx. whose h-parameters are $h_{fe} = 50$, $h_{ie} = 1.1K$, $h_{re} = h_{oe} = 0$.

Determine, $A_I = \frac{I_o}{I_i}$, A_V , A_{V_S} , R_o , R_i , R_i^o .

Solⁿ: for AC analysis $V_{cc} = 0$, all cap = S.C.



$$I_o = -h_{fe} I_b, \quad \frac{I_o}{I_b} = -h_{fe}$$

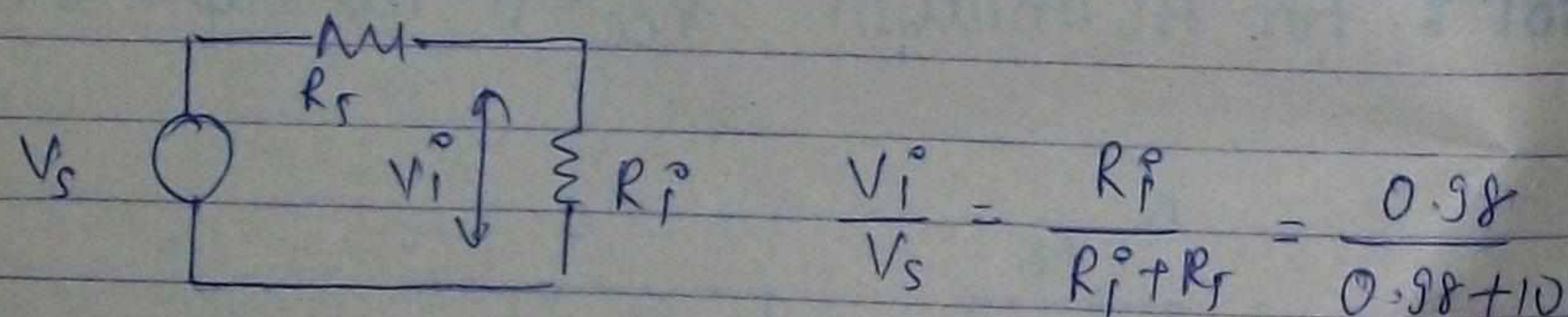
$$R_i^{\circ} = \frac{V_i^{\circ}}{I_b} = h_{ie} = 1.1 \text{ k}\Omega$$

$$R_i^{\circ} = R_i^{\circ} \parallel R_1 \parallel R_2 = 0.98 \text{ k}\Omega$$

$$A_v = \frac{V_o}{V_i} = \frac{I_o R_L}{I_b R_i^{\circ}} = -h_{fe} \frac{R_L}{R_i^{\circ}}$$

$$= -\frac{50 \times 5}{1.1} = -227$$

$$A_{vs} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s} = A_v \times \frac{V_i}{V_s}$$



$$\frac{V_i}{V_s} = \frac{R_i^{\circ}}{R_i^{\circ} + R_s} = \frac{0.98}{0.98 + 10}$$

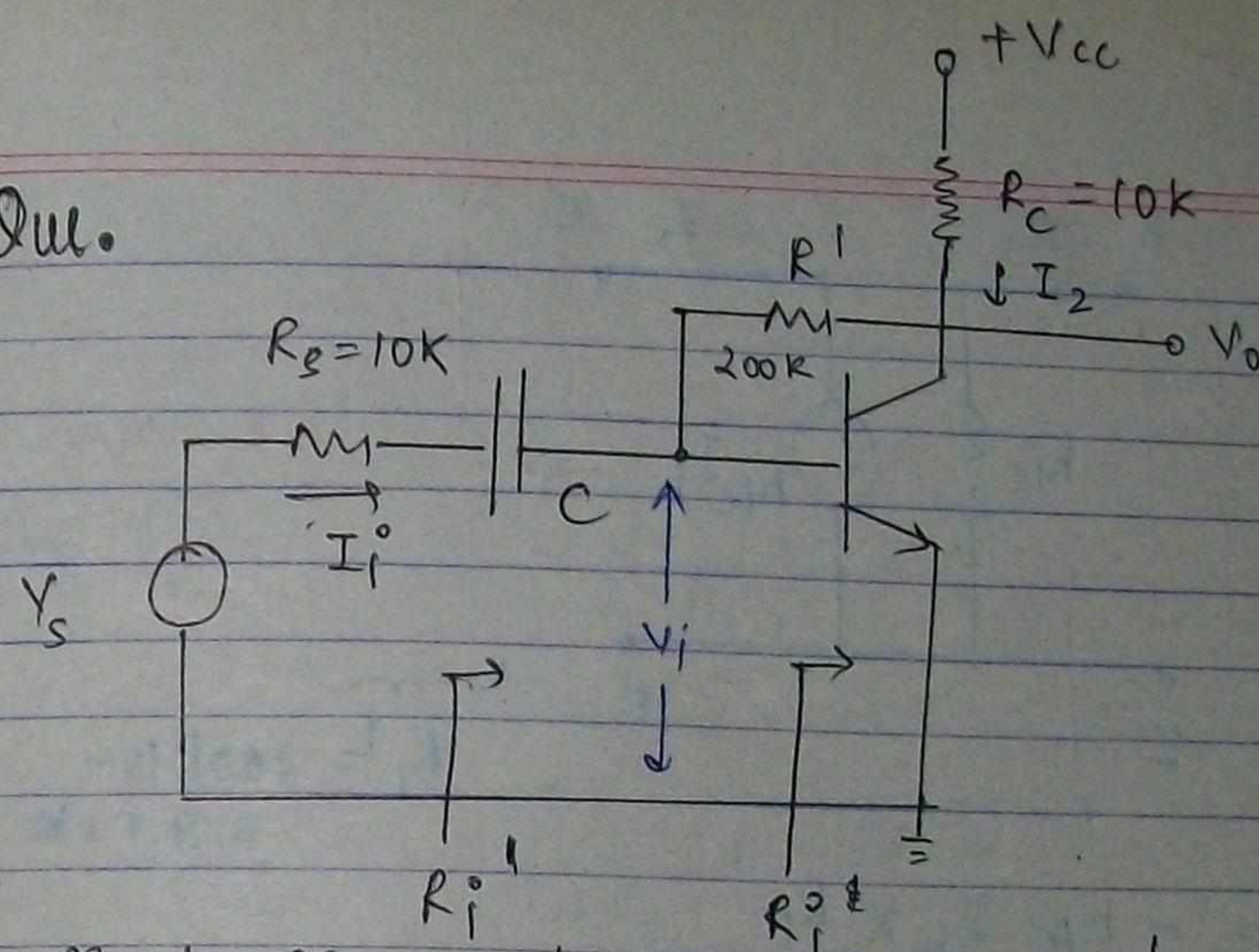
$$A_{vs} = -227 \times \frac{0.98}{10.98} = -20.6$$

$$A_I = \frac{I_o}{I_i} = \frac{V_o / R_L}{V_i / R_i^{\circ}} = A_v \times \frac{R_i^{\circ}}{R_L}$$

$$= -227 \times \frac{0.98}{5} = -44.4$$

$$R_o = \infty \quad R_o^{\circ} = R_o \parallel R_L = 5 \text{ k}\Omega$$

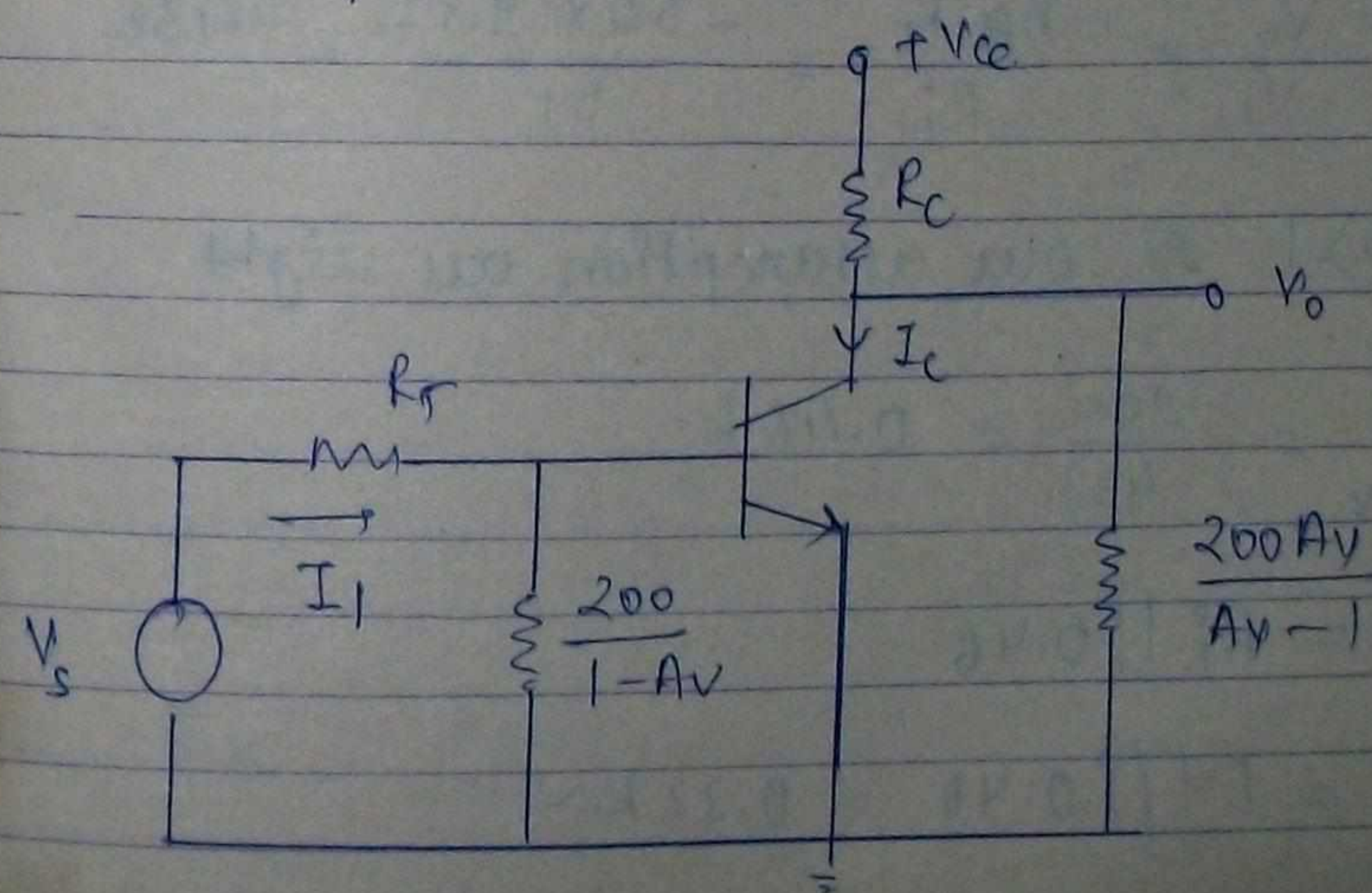
Ques.

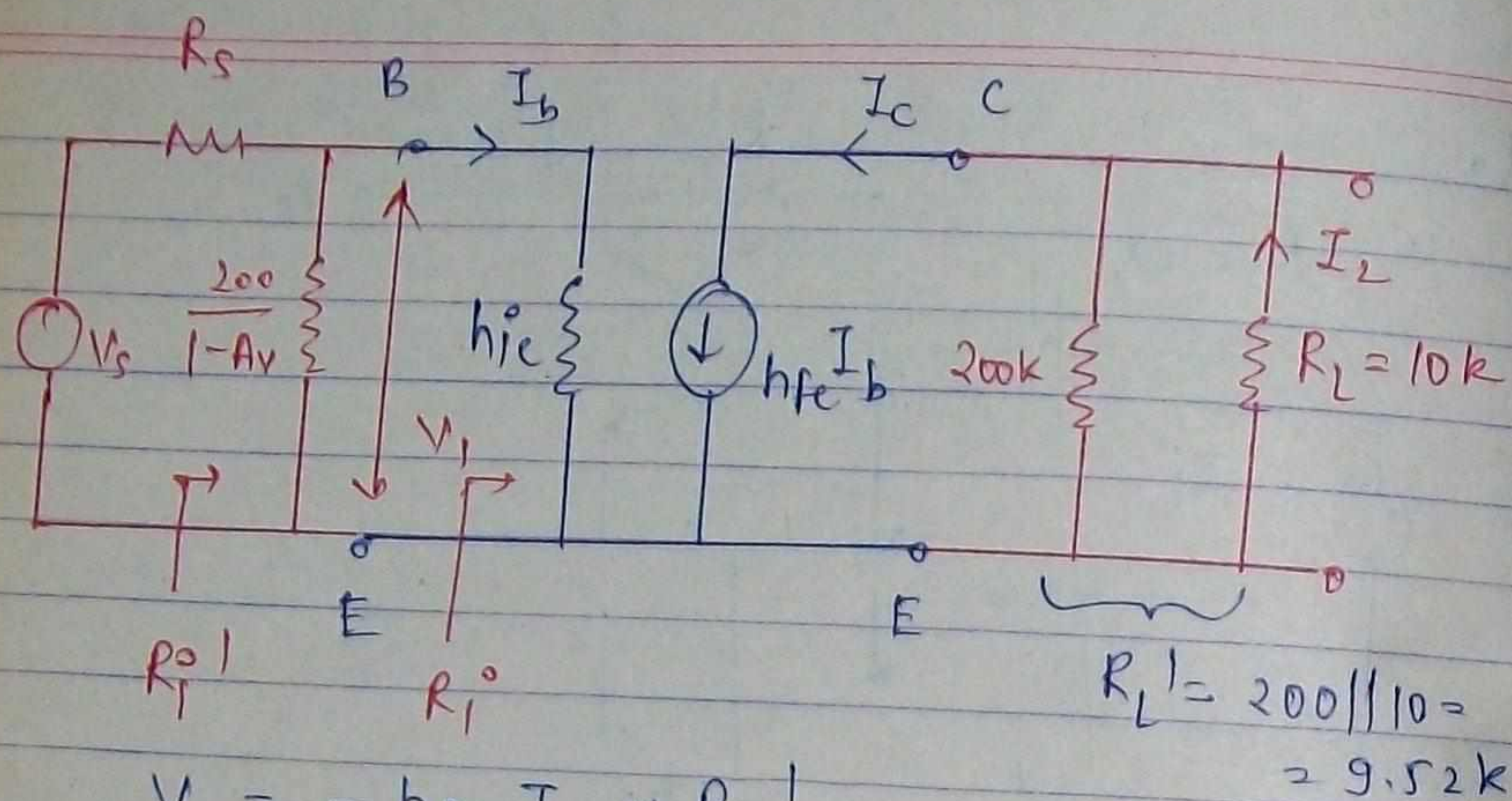


find R_i° , R_i° , A_v , A_{vs} , $A_I^{\circ} = -I_2/I_1$
 $h_{ie} = 1.1$ $h_{fe} = 50$ $h_{re} = h_{oe} = 0$

Solⁿ: $A_v = \frac{V_o}{V_i} \gg 1$ $\therefore$ CE-configuration

Apply miller's theorem





$$V_o = -h_{fe} I_b \times R_L'$$

$$R_i' = \frac{V_i}{I_b} = h_{ie} = 1.1k$$

$$V_i = I_b R_i'$$

$$A_v = \frac{V_o}{V_i} = -\frac{h_{fe} R_L'}{R_i'} = -\frac{50 \times 9.52}{1.1} = -432$$

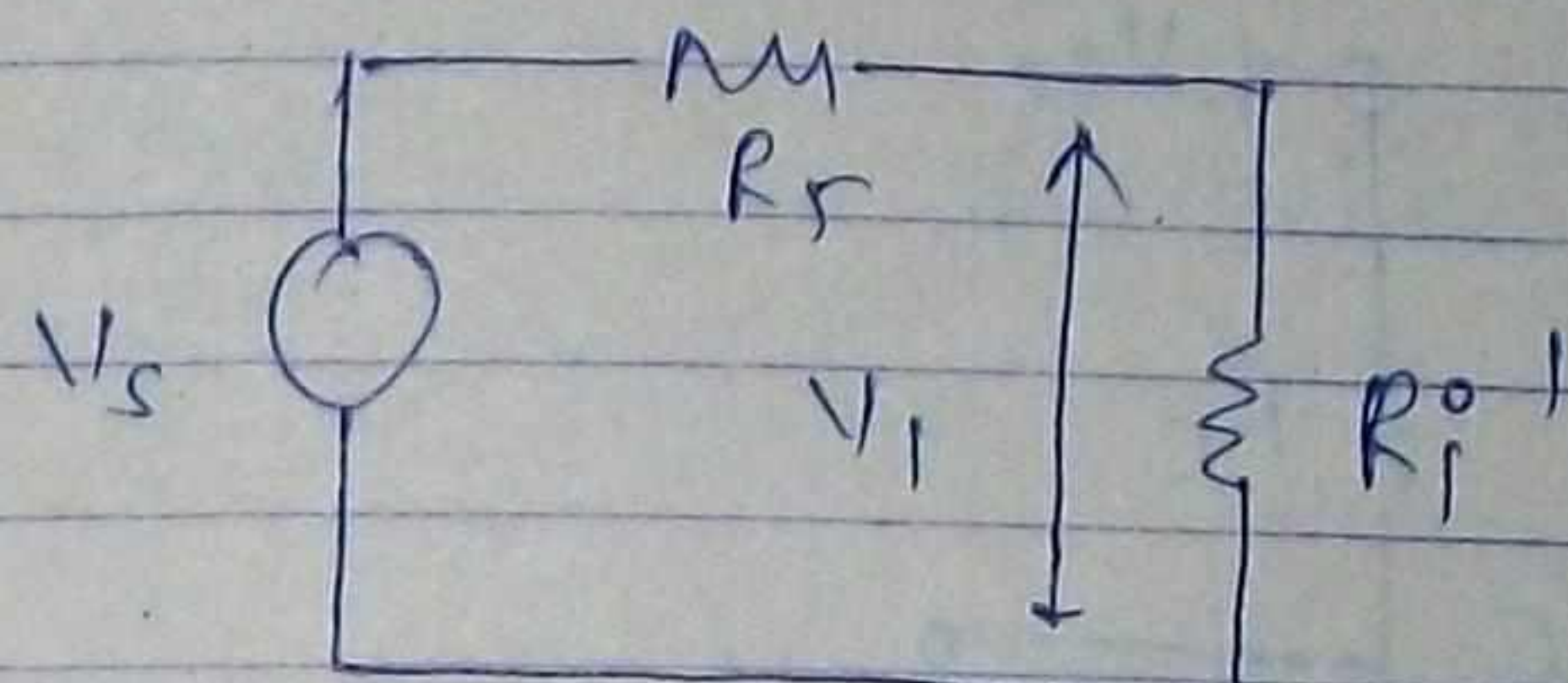
$|A_v| \gg 1$ ∴ our assumption are right

$$\frac{200}{1-A_v} = \frac{200}{433} = 0.46k$$

$$R_i' = R_i || 0.46$$

$$R_i' = 1.1 || 0.46 = 0.32k$$

$$A_{v_s} = \frac{V_o}{V_s} = \frac{V_o}{V_i} \times \frac{V_i}{V_s} = \textcircled{A_v} \times \frac{V_i}{V_s} \quad -432$$



$$\frac{V_i}{V_s} = \frac{R_i'}{R_i' + R_s} = \frac{0.32}{10 + 0.32}$$

$$A_{v_s} = A_v \times \frac{0.32}{10.32} = \underline{\underline{-13.4}}$$

$$A_I' = -\frac{I_2}{I_1} = -\frac{I_2}{I_b}$$

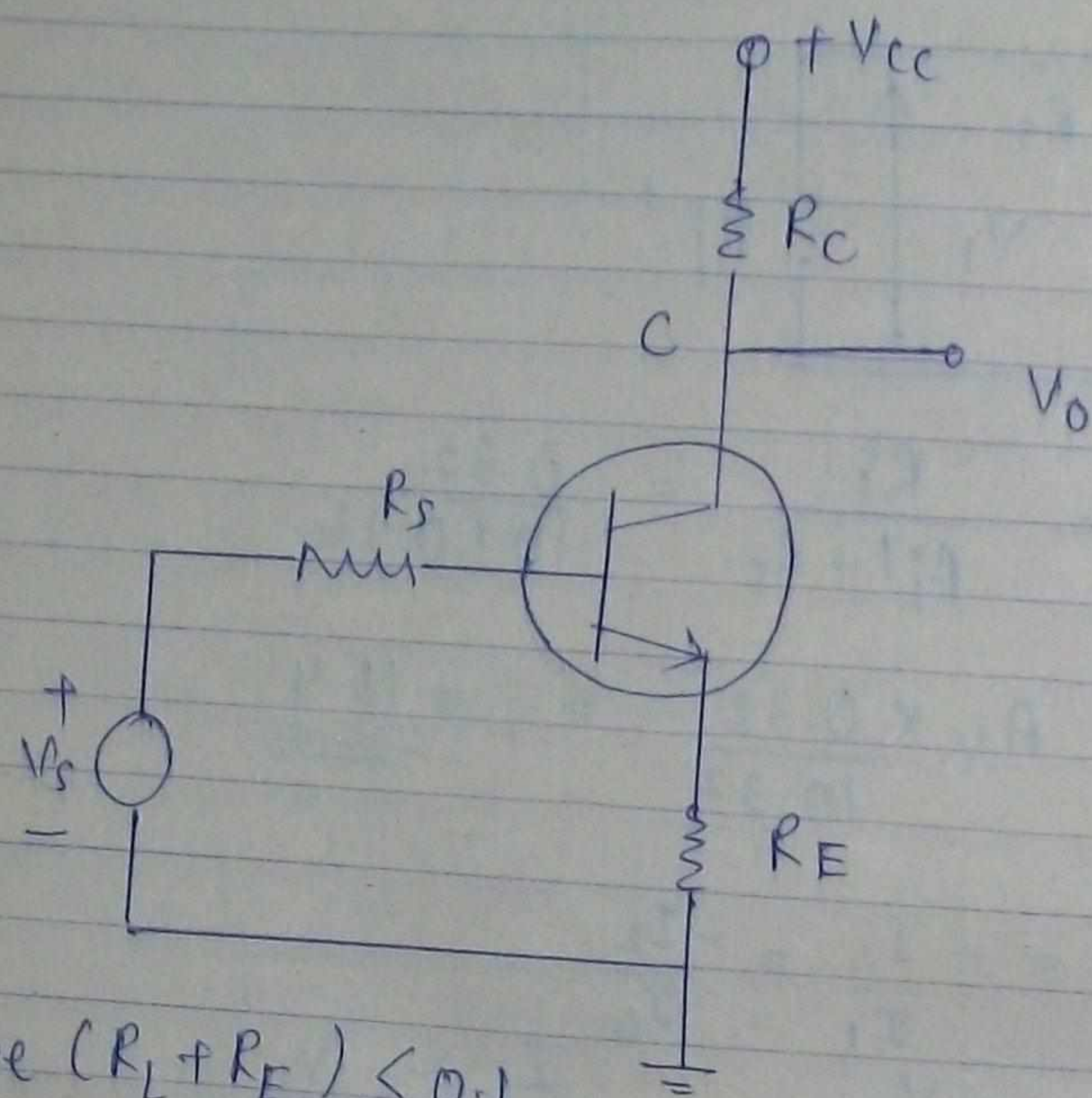
$$I_2 = -\frac{V_o}{R_L} \quad I_1 = \frac{V_i}{R_i'}$$

$$A_I' = \frac{V_o}{R_L} \times \frac{R_i'}{V_i} = \frac{V_o}{V_i} \times \frac{R_i'}{R_L} = A_v \times \frac{R_i'}{R_L}$$

$$A_I' = -\frac{432 \times 0.32}{10k}$$

$$A_I' = \underline{\underline{-13.8}}$$

⇒ COMMON EMITTER WITH UNBYPASSED EMITTER RESISTANCE (R_E) -



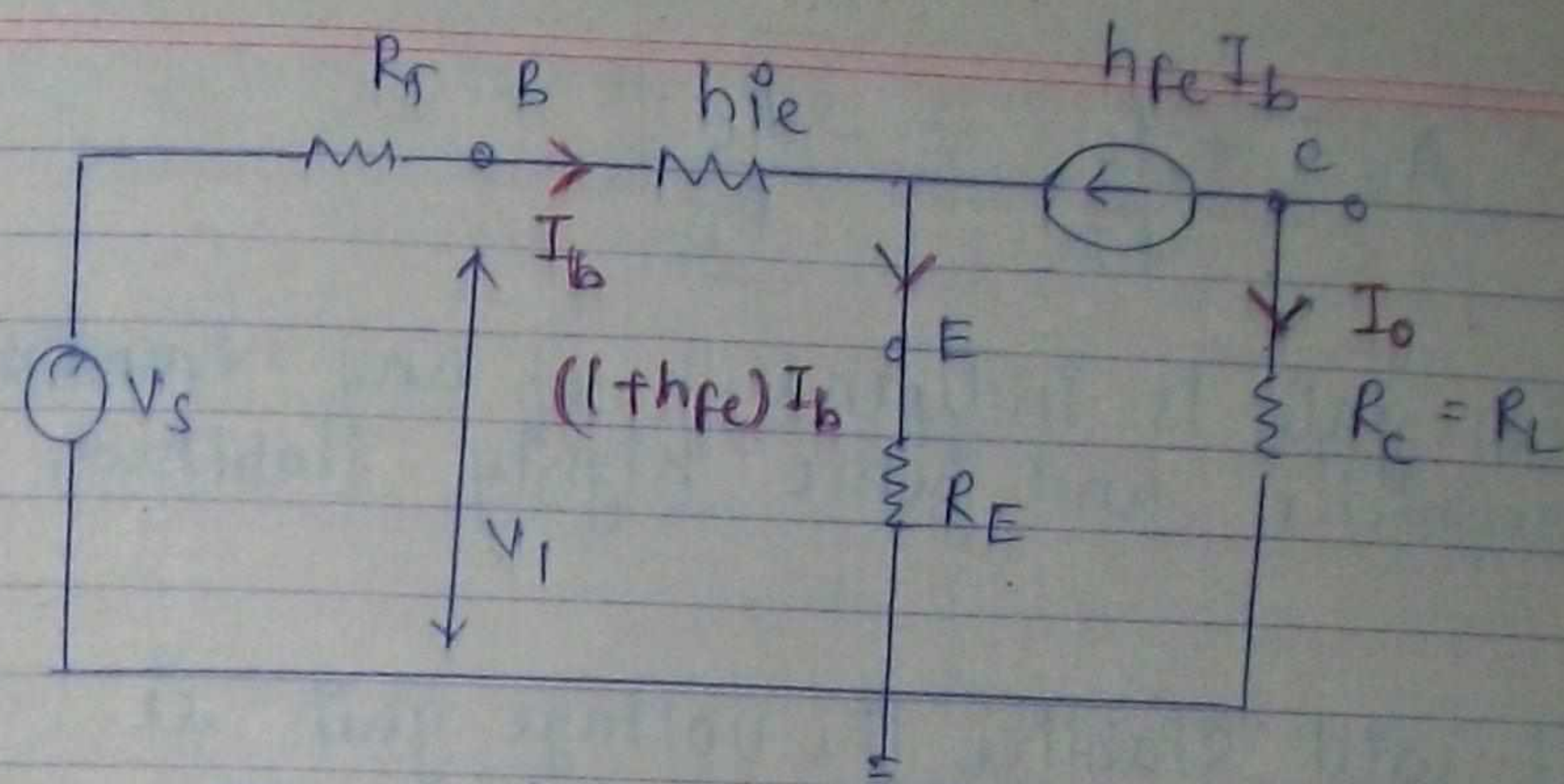
we can use approx. model

CE without R_E -

$$A_I = -h_{fe}$$

$$R_i^o = h_{ie}$$

$$A_v = \frac{A_I R_L}{R_i^o}$$



$$I_o = -h_{fe} I_b$$

$$A_I = \frac{I_o}{I_b} = -h_{fe}$$

$$V_1 = I_b h_{ie} + (1+h_{fe}) R_E I_b$$

$$R_i^o = \frac{V_1}{I_b} = h_{ie} + (1+h_{fe}) R_E$$

$$R_i^o = h_{ie} + (1+h_{fe}) R_E \quad (\text{increased})$$

$$A_v = \frac{V_o}{V_1} = \frac{I_o R_L}{I_b R_i^o} = \frac{A_I R_L}{R_i^o}$$

$$A_v = \frac{-h_{fe} R_L}{h_{ie} + (1+h_{fe}) R_E}$$

$$(1+h_{fe}) R_E \gg h_{ie} \quad h_{ie} \gg 1$$

$$A_v = -\frac{R_L}{R_E} = -\frac{R_C}{R_E}$$

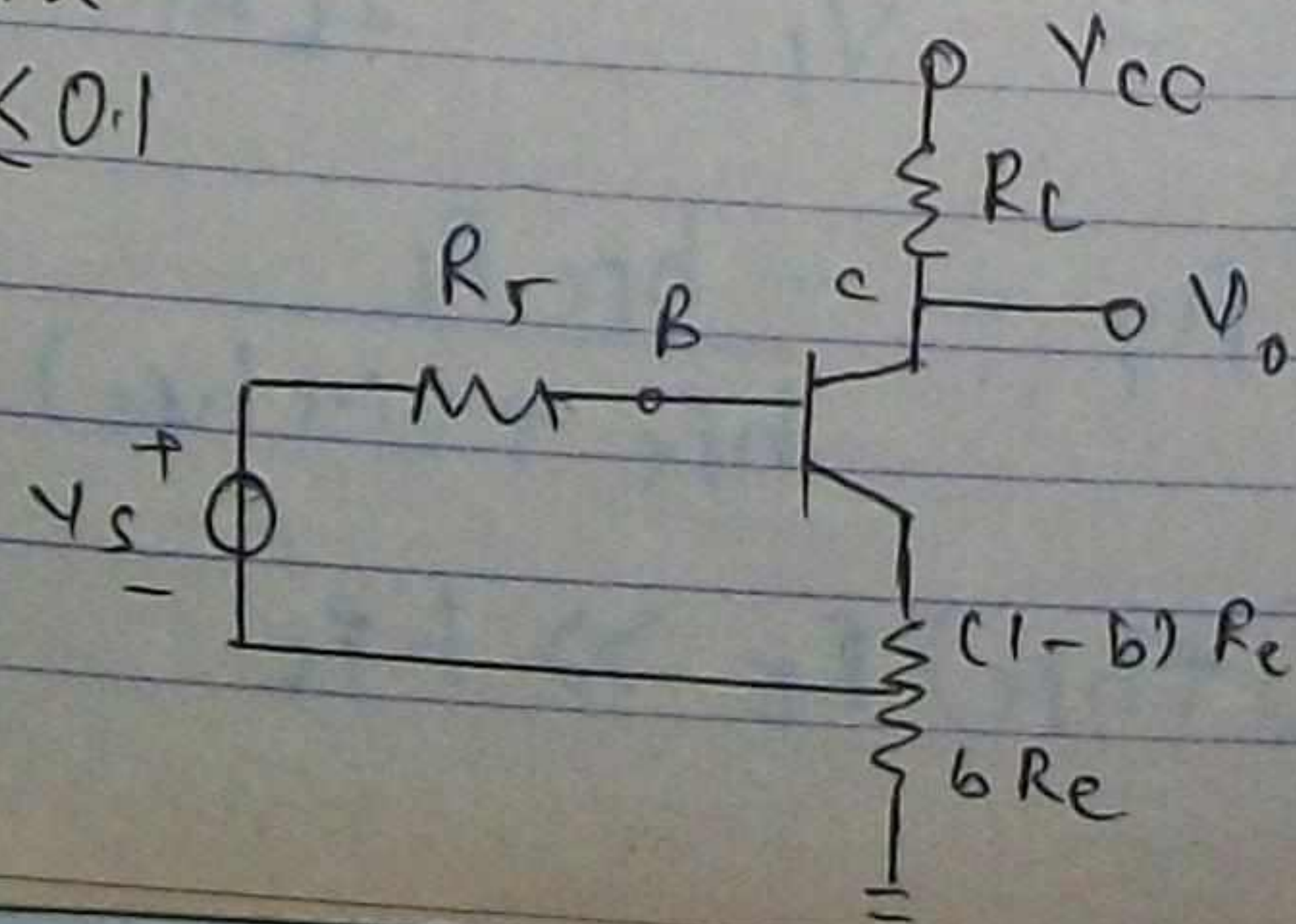
Thus gain is independent of any transistor parameter and hence highly stabilised.

- It will stabilize the voltage gain as voltage gain is independent of any Tx. parameters. (with condition

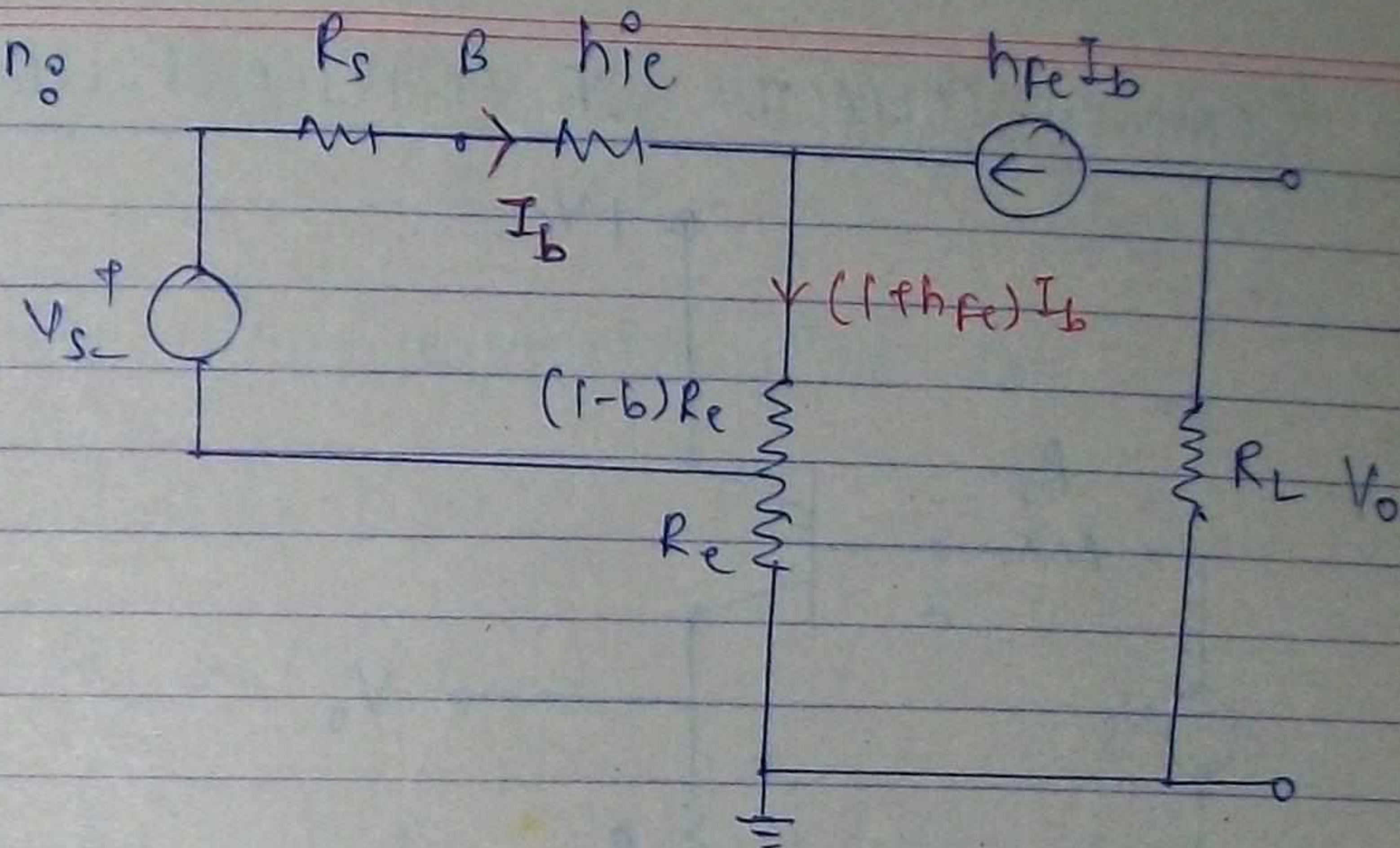
$$(1+h_{fe})R_E \gg h_{fe}$$

- A_i remain unchanged.
- It increases the i/p impedance by $(1+h_{fe})R_E$
- It increases the o/p impedance.

Ques. For circuit show find $A_v = V_o/V_s$ and $R_i = V_s/I_b$ as a funn of R_s, b, R_e and R_L Assume that $h_{oe}(R_L+R_E) \leq 0.1$



Soln:



$$V_o = -h_{fe} I_b R_L$$

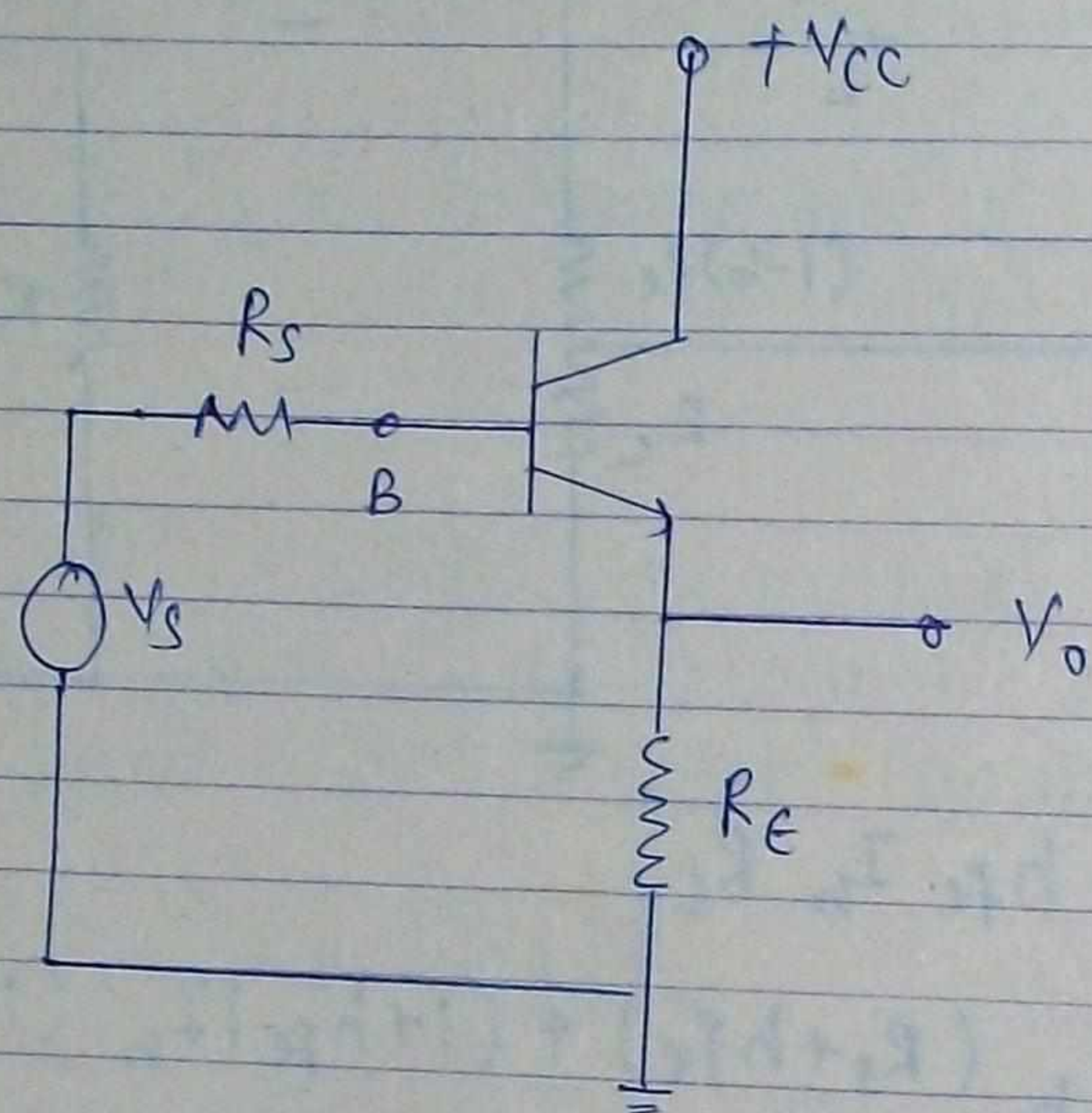
$$V_s = I_b (R_s + h_{ie}) + (1+h_{fe}) I_b (1-b) R_e$$

$$\frac{V_o}{V_s} = \frac{-h_{fe} R_L}{R_s + h_{ie} + (1+h_{fe})(1-b) R_e}$$

$$R_i = \frac{V_s}{I_b} = R_s + h_{ie} + (1+h_{fe})(1-b) R_e$$

31/08/11

⇒ COMMON COLLECTOR OR EMITTER FOLLOWER



CC without R_E -

$$A_I = \frac{-h_{fe}}{1+h_{oe}R_L} = \frac{(1+h_{fe})}{1+h_{oe}R_L}$$

$$R_i^o = h_{ie} + h_{re} A_I R_L = h_{ie} + A_I R_L$$

$$A_v = \frac{A_I R_L}{R_i^o}$$

$$h_{re} = 1, \quad h_{fe} = -(1+h_{fe})$$

$$h_{ie} = h_{ie}, \quad h_{oc} = h_{oe}$$

$$Y_o = h_{oc} - \frac{h_{re} h_{fe}}{R_s + h_{ie}} = h_{oe} + \frac{(1+h_{fe})}{R_s + h_{ie}}$$

By using approximation i.e.

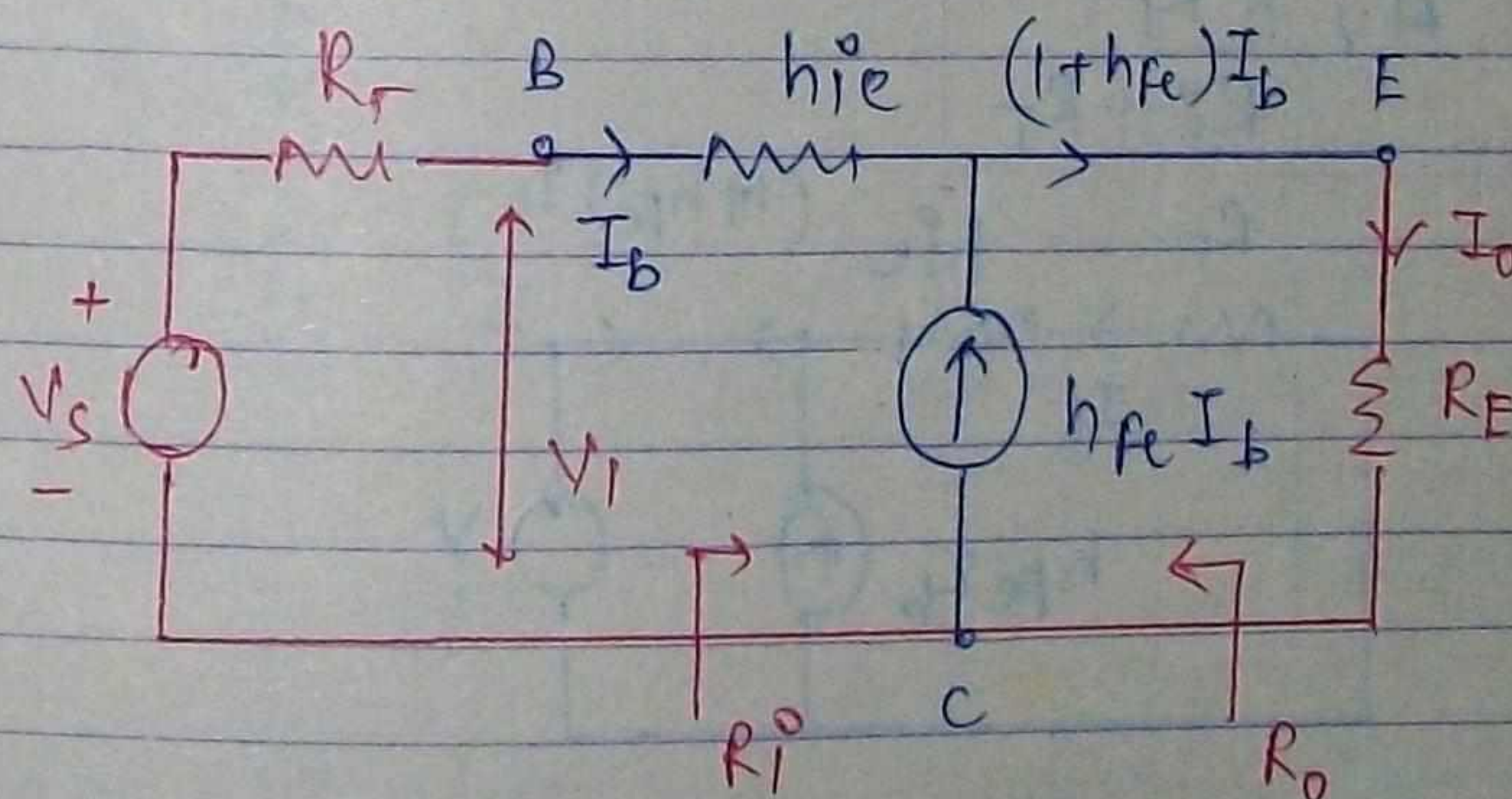
$$h_{oe} R_L \leq 0.1 \quad h_{oe} = h_{re} = 0$$

$$A_I = (1+h_{fe})$$

$$R_i^o = h_{ie} + (1+h_{fe}) R_L$$

$$A_v = \frac{A_I R_L}{R_i^o}$$

$$Y_o = \frac{1+h_{fe}}{R_s + h_{ie}}$$



$$I_o = (1+h_{fe}) I_b$$

$$A_I = \frac{I_o}{I_b} = (1+h_{fe})$$

$$V_i = I_b h_{ie} + (1+h_{fe}) I_b R_E$$

$$R_i^o = \frac{V_i}{I_b} = h_{ie} + (1+h_{fe}) R_E \quad (\text{HIGH})$$

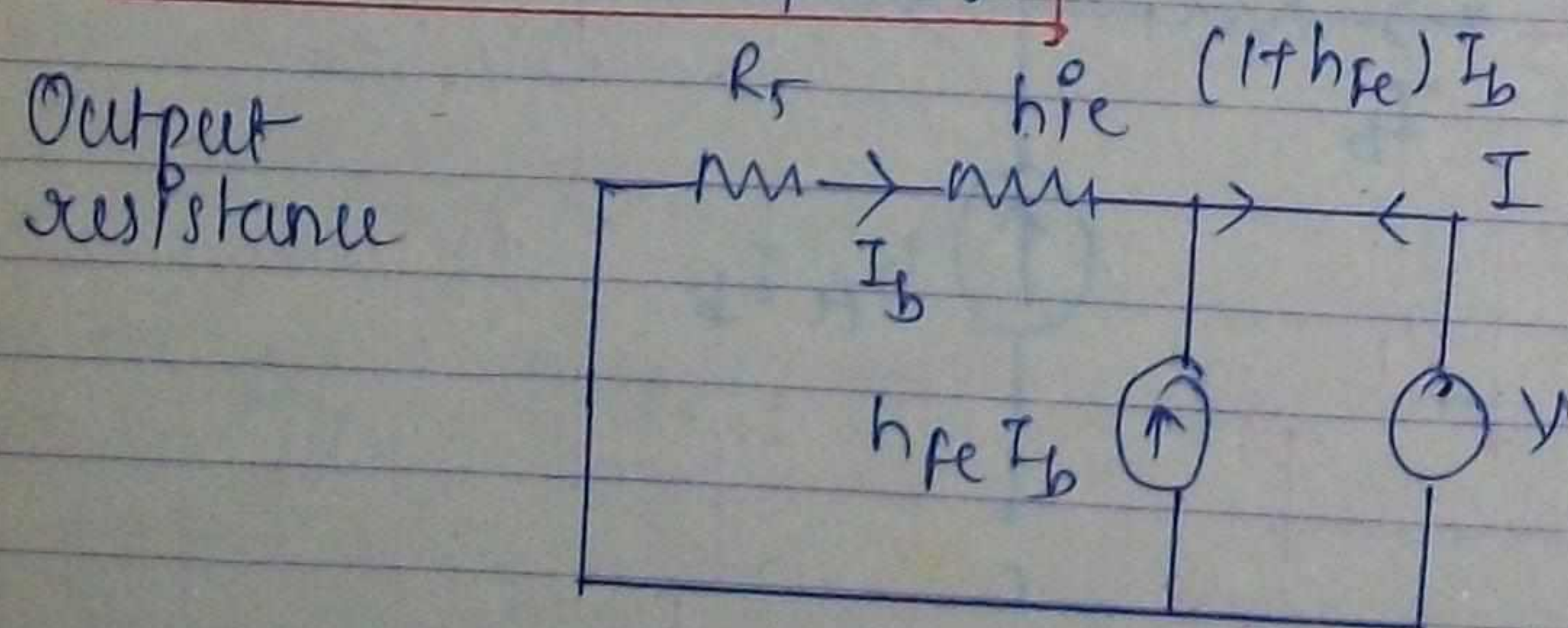
$$A_v = A_i \frac{R_L}{R_i^o}$$

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

when $(1+h_{fe}) R_E \gg h_{ie}$

$$A_v = 1$$

$$A_{vs} = A_v \times \frac{R_i^o}{R_i^o + R_s}$$



$$R_o = V/I$$

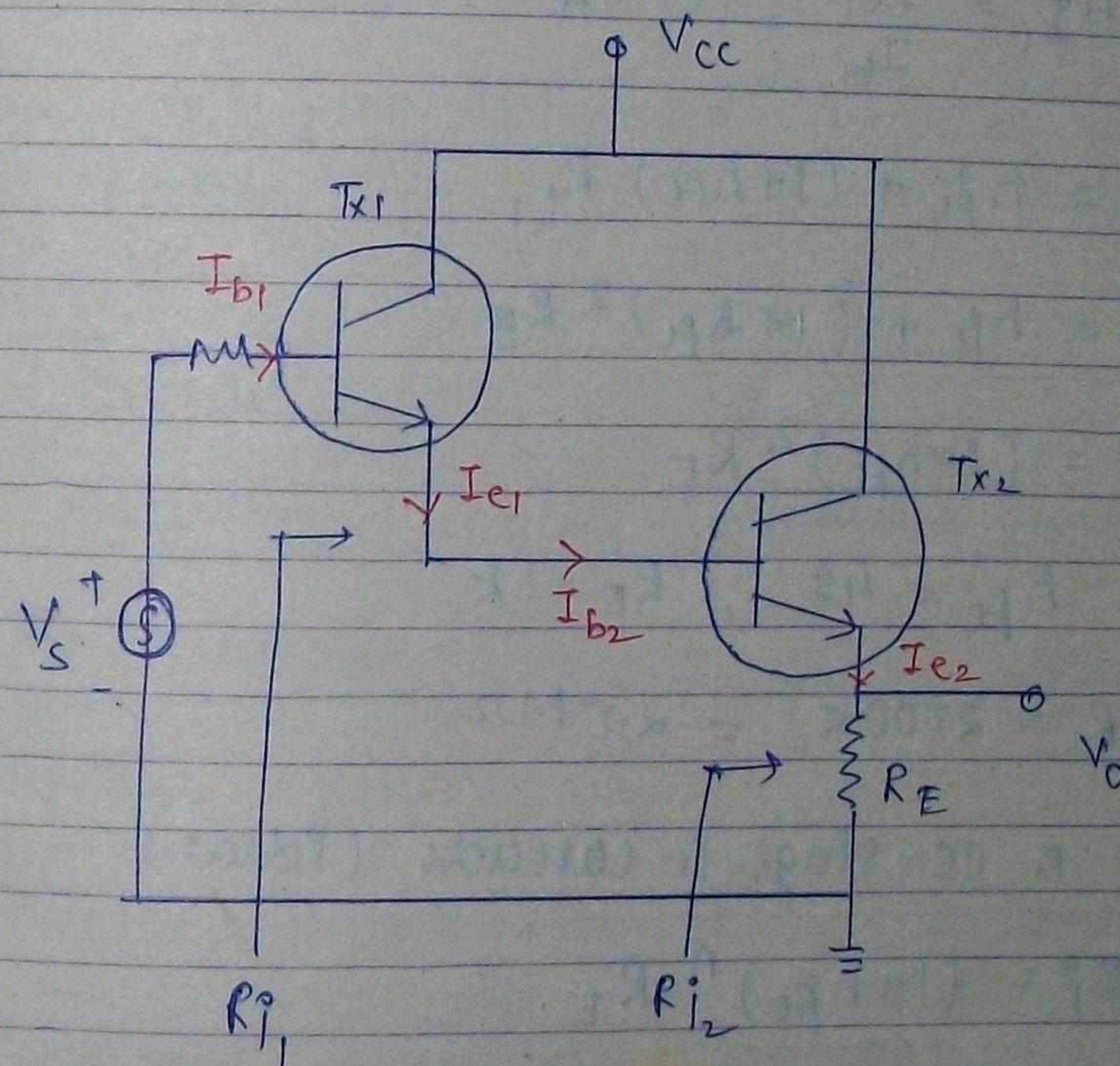
$$I = -(1+h_{fe}) I_b$$

$$V = -I_b (R_s + h_{ie})$$

$$R_o = \frac{V}{I} = \frac{R_s + h_{ie}}{1+h_{fe}} \quad (\text{LOW})$$

⇒ HIGH I/P RESISTANCE CIRCUIT -

⇒ DARLINGTON-PAIR (CC-CC) -



for 2nd stage - $(h_{re} = h_{oe} = 0)$

$$A_{I2} = \frac{I_{e2}}{I_{b2}} = (1+h_{fe})$$

$$R_{i2} = h_{ie} + (1+h_{fe}) R_E \approx (1+h_{fe}) R_E$$

$$A_{V_2} \approx 1$$

for 1st stage -

$$R_{L1} = R_{i2} = (1 + h_{fe}) R_E$$

$$A_{I1} = \frac{I_{e1}}{I_{b1}} = (1 + h_{fe})$$

$$R_i = h_{fe} + (1 + h_{fe}) R_{L1}$$

$$R_i = h_{fe} + (1 + h_{fe})^2 R_E$$

$$\star \quad R_i = (1 + h_{fe})^2 R_E$$

Ex $h_{fe} = 49, R_E = 1k$

$$R_i = 2500k = 2.5 M\Omega$$

for n cc-stage in cascade (ideal)

$$\star \quad R_i = (1 + h_{fe})^n R_E$$

$$A_I = \frac{I_{e2}}{I_{b1}} = \frac{I_{e2}}{I_{b2}} \times \frac{I_{b2}}{I_{e1}} \times \frac{I_{e1}}{I_{b1}}$$

$$A_I = A_{I2} \times A_{I1} \times 1 = A_{I1} \times A_{I2}$$

$$\star \quad A_I = (1 + h_{fe})^2 \approx h_{fe}^2 \approx \beta^2$$

$$h_{fe \text{ eff}} = h_{fe}^2$$

$$\text{or } \beta_{\text{eff}} = \beta^2$$

- Very high current gain.

- Darlington integrated Tx. pair are commercially available with h_{fe} as high as 30,000 therefore this circuit is also called super β Tx or β -multiplication.

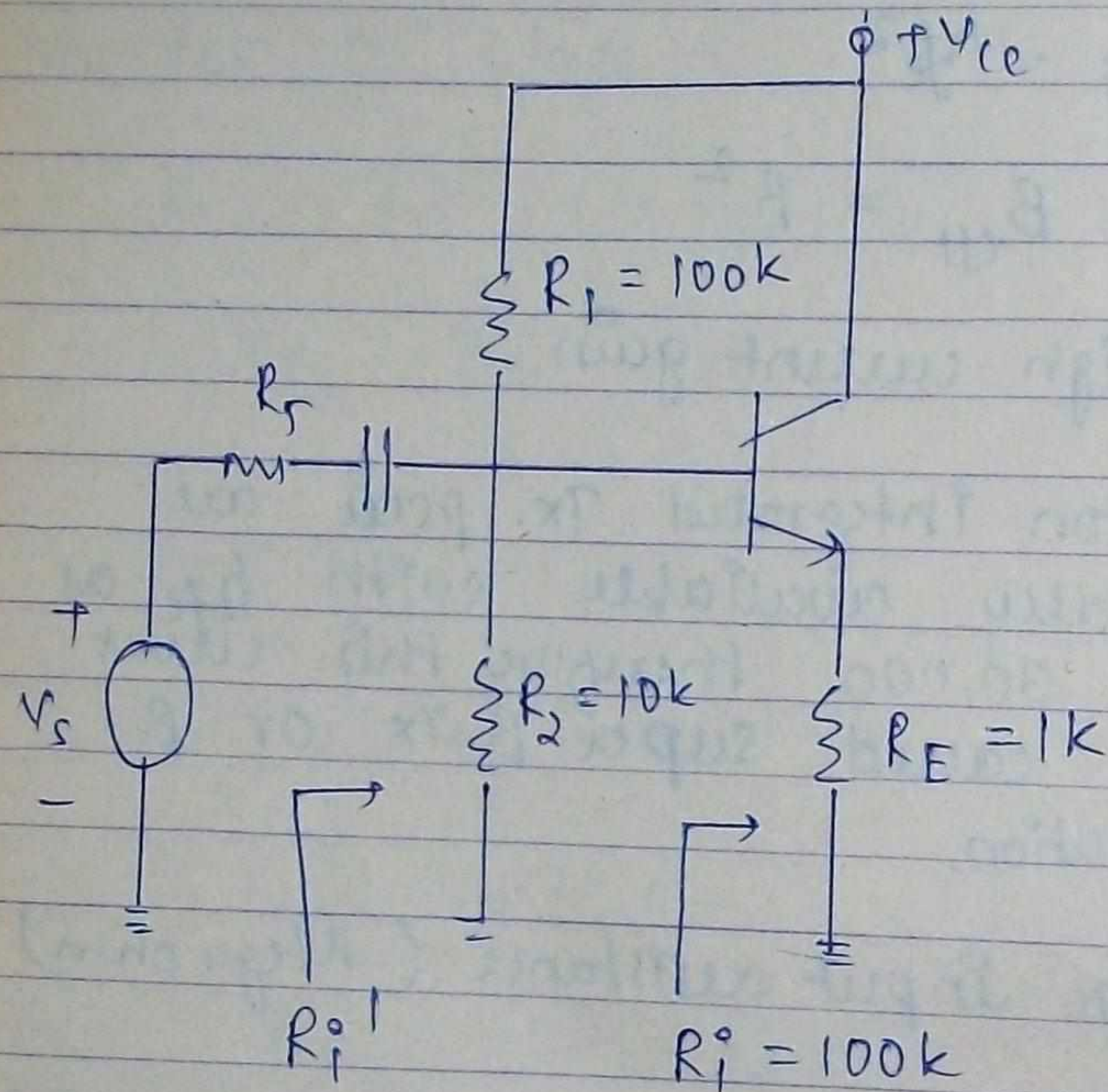
- very large input resistance (Mega ohm)

DISADVANTAGES -

- Highly expensive circuit.

- Leakage current of first Tx is amplified by second and hence overall leakage current may be high and a darlington connection of three or more transistor is usually impractical.

⇒ BIASING PROBLEM AND BOOTSTRAP CIRCUIT -



$$R_E = 1k, h_{fe} = 99$$

$$R_i' = h_{ie} + (1 + h_{fe}) R_E$$

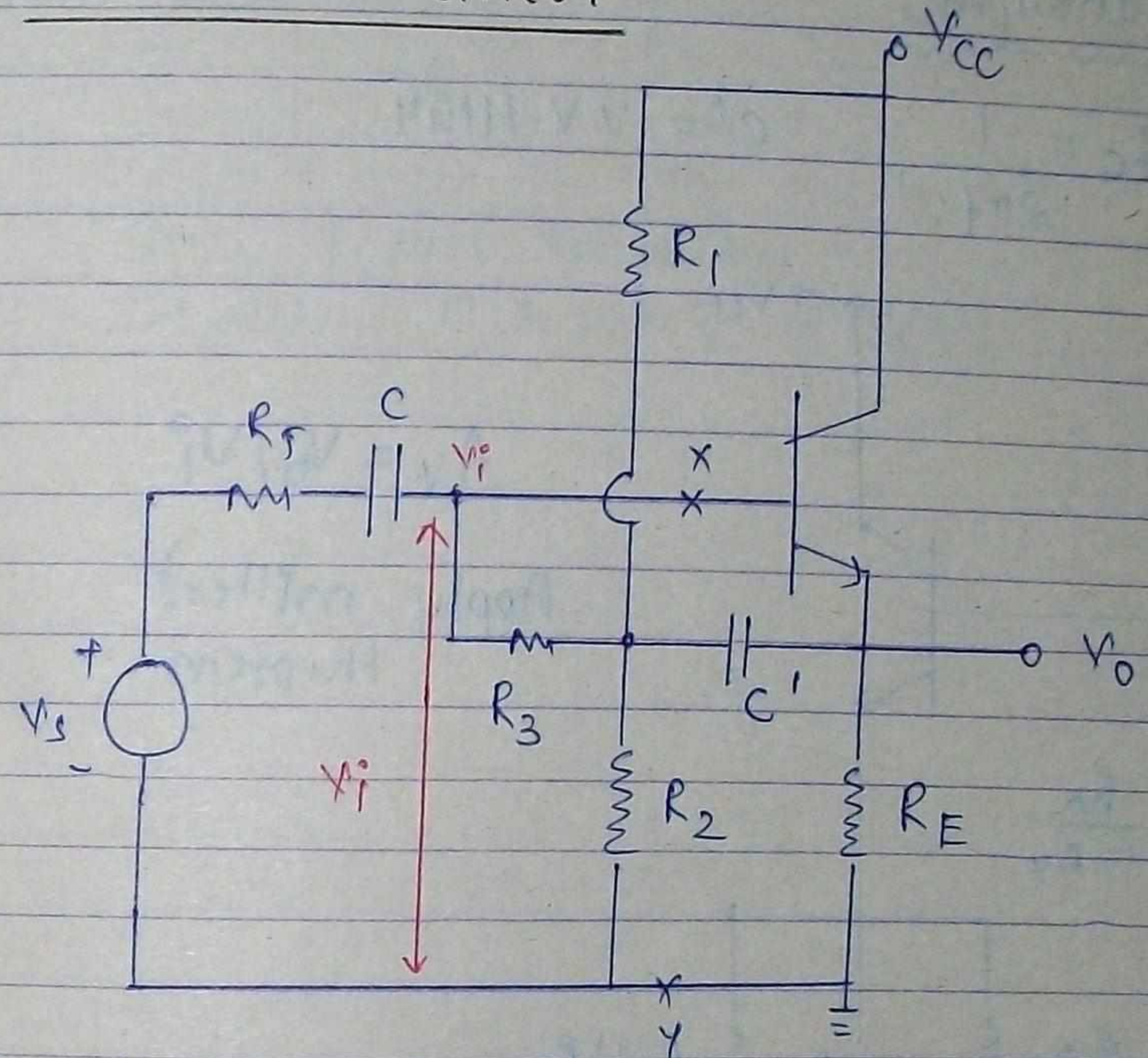
$$R_i' = 100k$$

$$R_i' = 10 \parallel 100 \parallel 100 \approx 10k$$

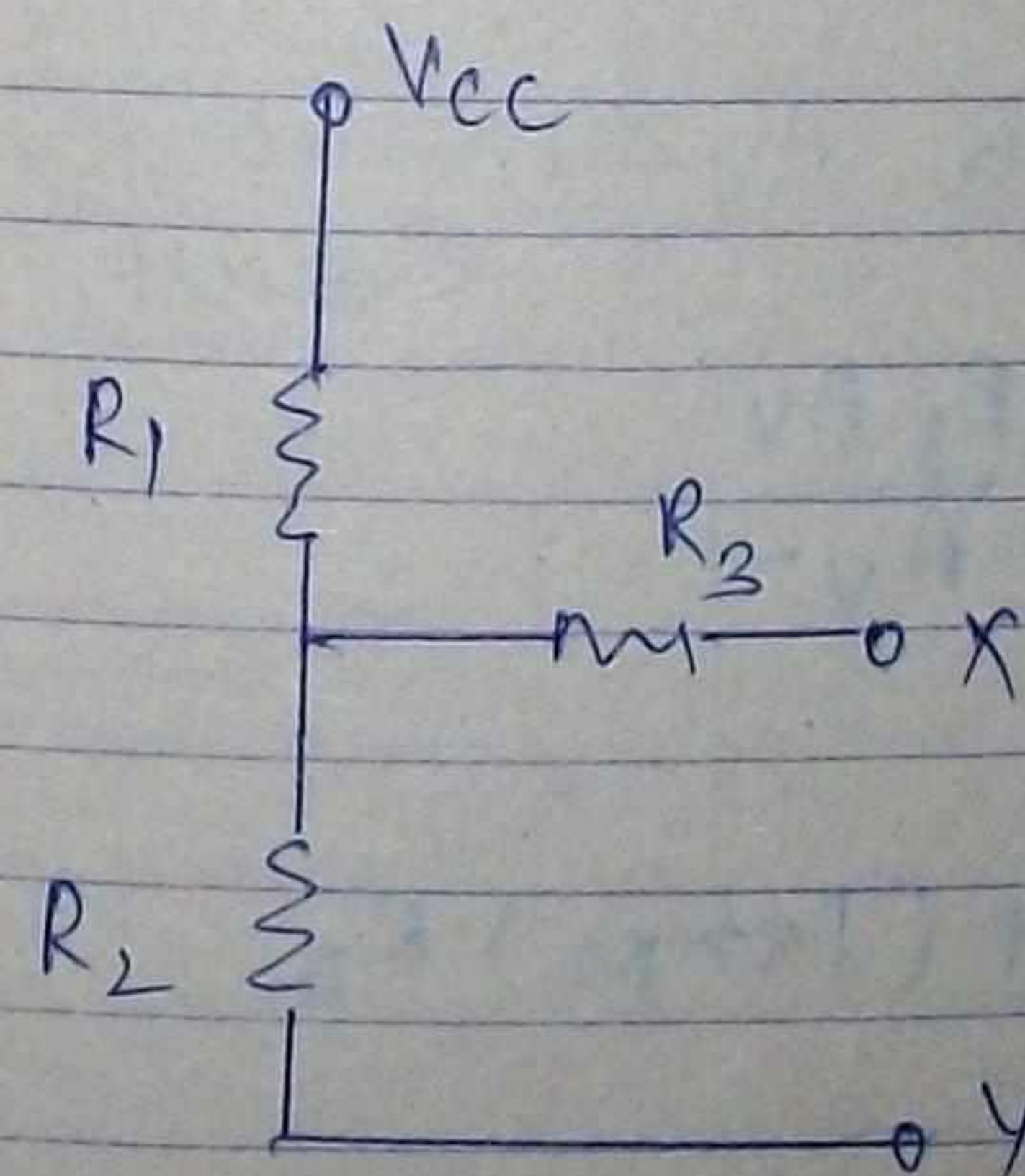
Due to biasing I/P resistance decreases ↓.

★★

BOOTSTRAP CIRCUIT -



for DC analysis, C and C' acts as open circuit



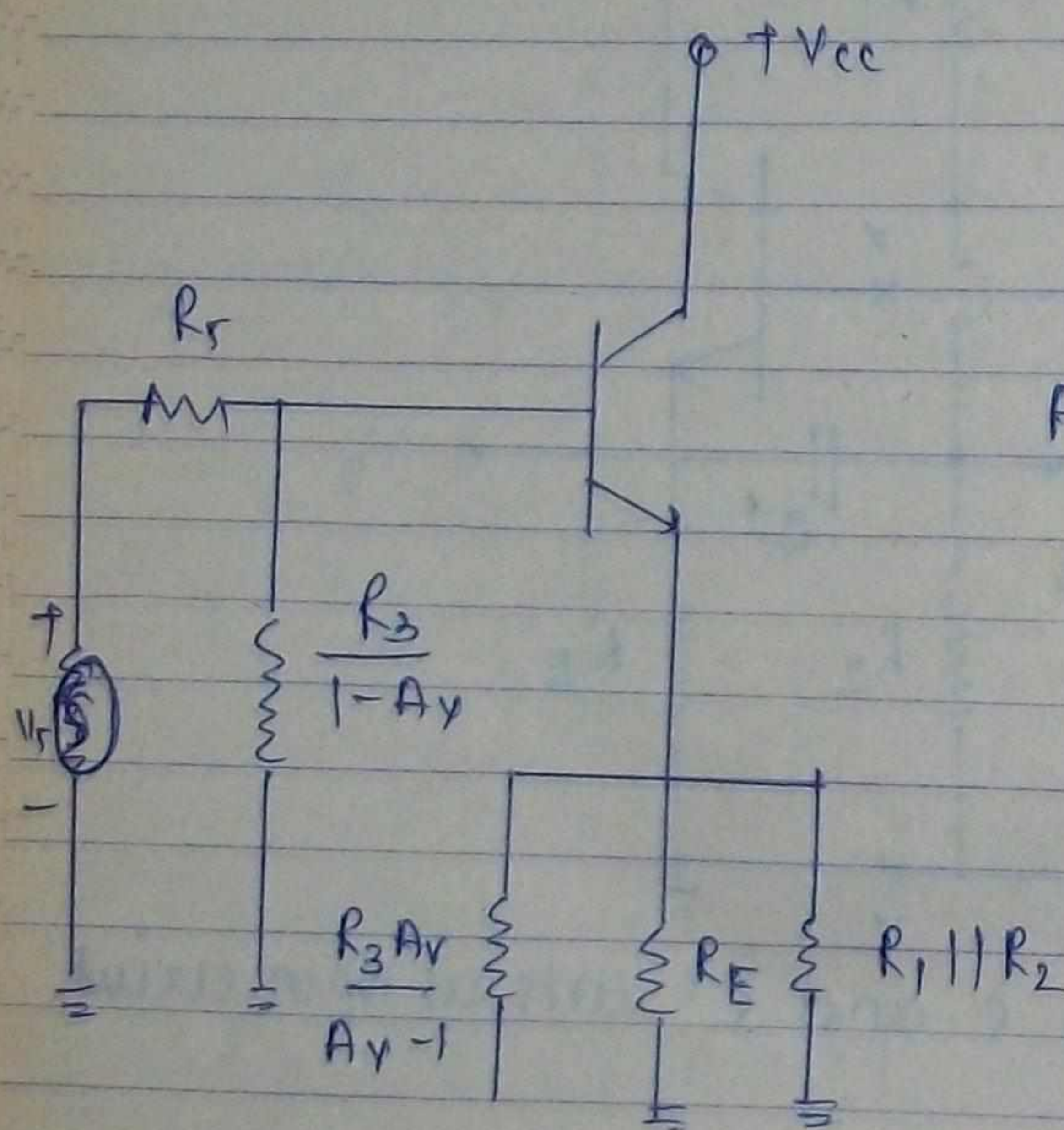
$$V_{TH} = \frac{R_2}{R_1 + R_2} V_{CC}$$

$$R_{TH} = R_1 \parallel R_2 + R_3$$

for AC analysis:

$$Z_C = \frac{1}{2\pi f C'}$$

$$C' = \text{V.V. HIGH}$$



$$A_v = V_o / V_i$$

Apply miller's theorem

$$A_v = 1 \quad (\text{Due to } -ve \text{ f.B. } A_v \approx 1 \text{ (low)})$$

$$R_{eq} = R_E \parallel R_1 \parallel R_2 \parallel \frac{R_3 A_v}{A_v - 1}$$

$$R_{eq} \approx R_E$$

$$R_i^o = 100 \text{ k} = h_{ie} + (1 + h_{fe}) R_E$$

$$R_{ieff} = \frac{R_3}{1 - A_v} \approx \infty$$

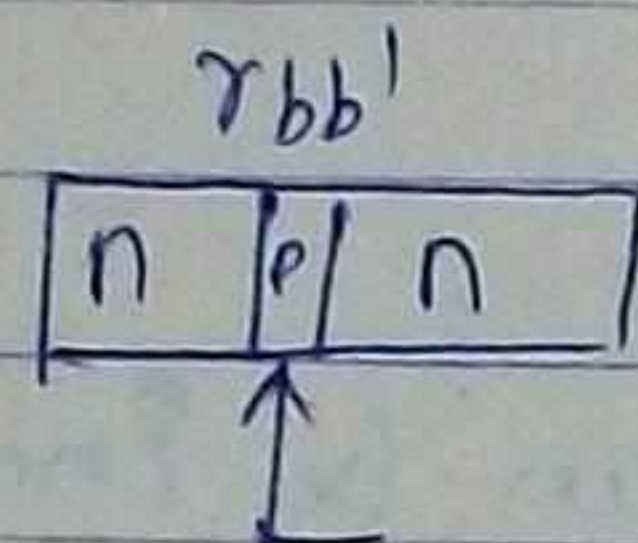
∞

$$R_{ieff} = \frac{R_3}{1 - A_v} \approx \infty$$

Thus input impedance can be increased without disturbing biasing.

TRANSISTOR AT HIGH FREQUENCY

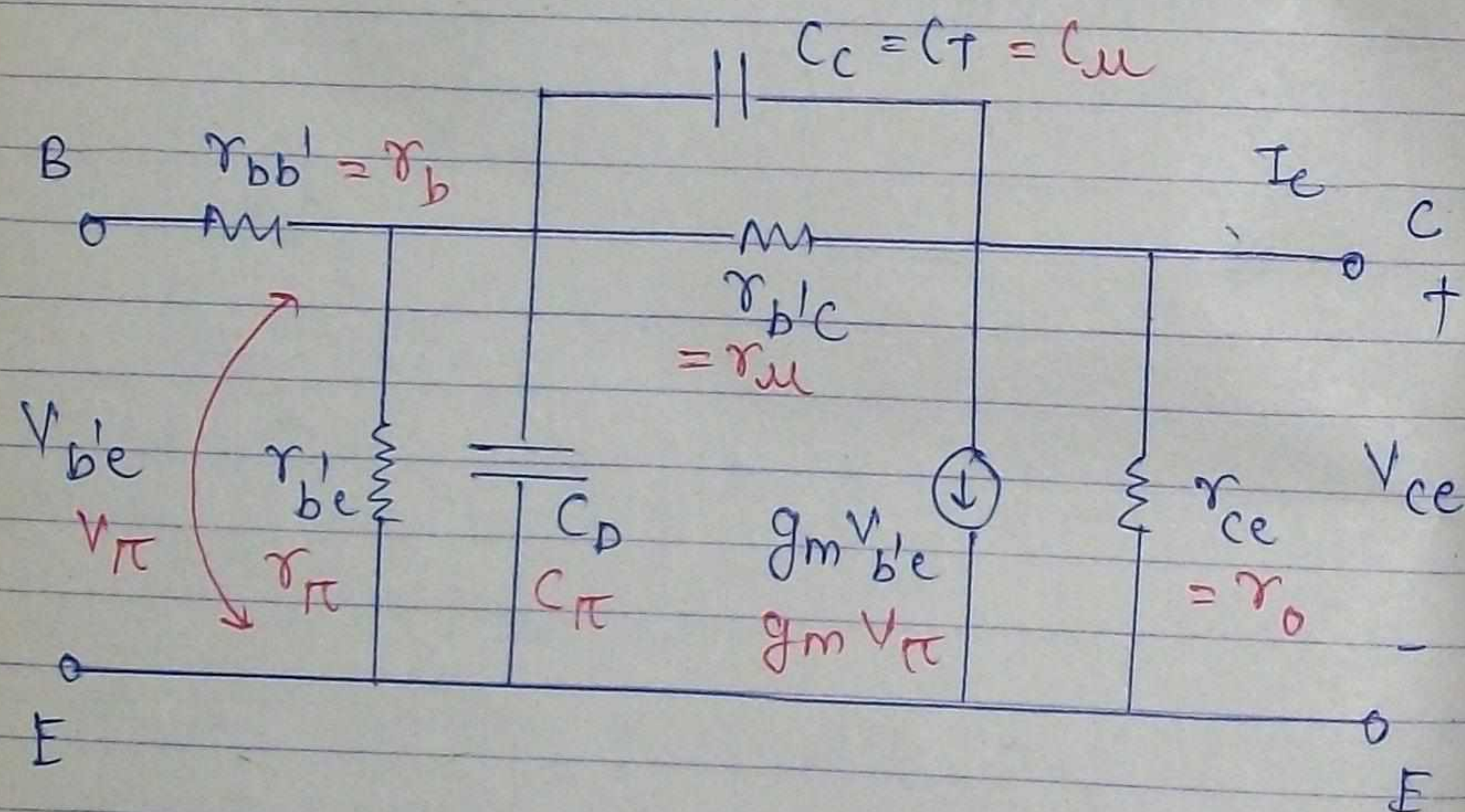
⇒ π - MODEL - (GIAOLETTO MODEL)



At Base area = m^2

$$R = \rho \frac{L}{A} = \text{max. ohmic resistance } (r_{bb'})$$

$r_{bb'}$ = Base spreading resistance.



- The transistor is represented as a voltage controlled source.

- $r_{bb'}$ (BASE SPREADING RESISTANCE) -

Ohmic resistance of base region. The base region of Tx is very thin compare to C and E region

and its resistance lies b/w 40-400Ω. The ohmic resistance of collector is of order of 10Ω and can be neglected in comparison to that of base region.

- r_{π} -

Incremental resistance of emitter-base diode which is F.B. in active region (similar to r_D for diode).

- r_{μ} -

Accounts for feedback from o/p to i/p due to base width modulation or early effect. The value of r_{μ} is usually very high (several MΩ) and will be neglected in analysis.

- r_o -

Output resistance. It is also due to early effect. r_o is ($r_o = V_A / I_C$) where V_A is early voltage and I_C is collector D.C. current.

- $g_m V_{\pi}$ -

Any small signal voltage V_{π} at emitter junction results in a signal collector current $g_m V_{\pi}$. when collector is shorted to emitter i.e. $V_{ce} = 0$

The BJT is represented as (VCC) voltage controlled current source where controlled current is $g_m V_{\pi}$ and controlling voltage is V_{π} and g_m represent the transconductance of T_x .

- C_D - The extra minority carrier storage in base is accounted for by diffusion capacitance C_D .

- $C_C = C_T = C_{\mu} = C_{ob}$ - collector junction barrier capacitance.

TYPICAL VALUES -

$$g_m = 50 \text{ mA/V}$$

$$r_{bb'} = 100 \Omega$$

$$r_{b'e} = 1 \text{ k}\Omega$$

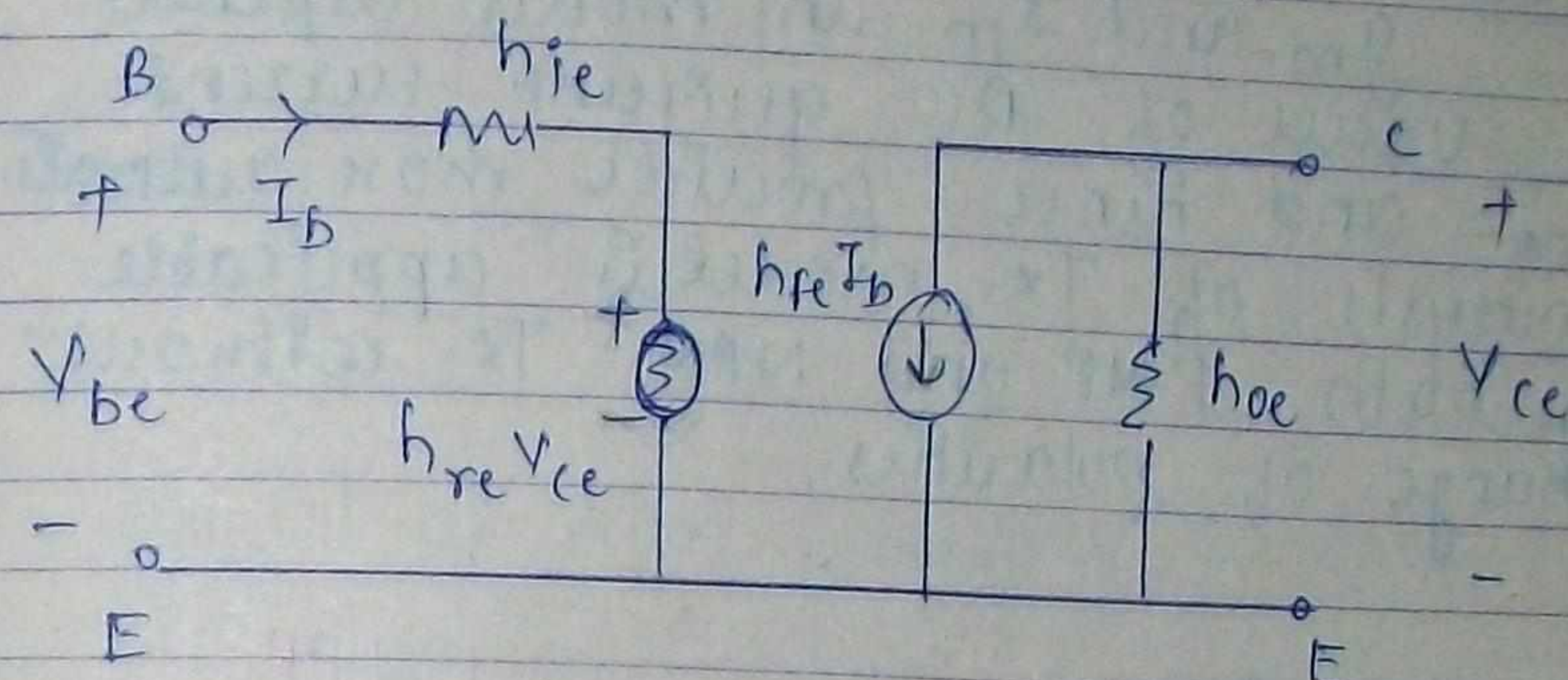
$$r_{b'c} = 4 \text{ M}\Omega$$

$$r_{ce} = 80 \text{ k}\Omega$$

$$C_C = C_T = 3 \text{ pF}$$

$$C_D = 100 \text{ pF}$$

Comparing π -model with h parameter model



FORMULA -

$$g_m = \frac{|I_{CQ}|}{V_T}$$

$$r_{\pi} = r_{b'e} = \frac{h_{fe}}{g_m}$$

$$r_{bb'} = h_{ie} - r_{b'e}$$

$$r_b = h_{ie} - r_{\pi}$$

$$r_{b'c} = \frac{r_{b'e}}{h_{re}}$$

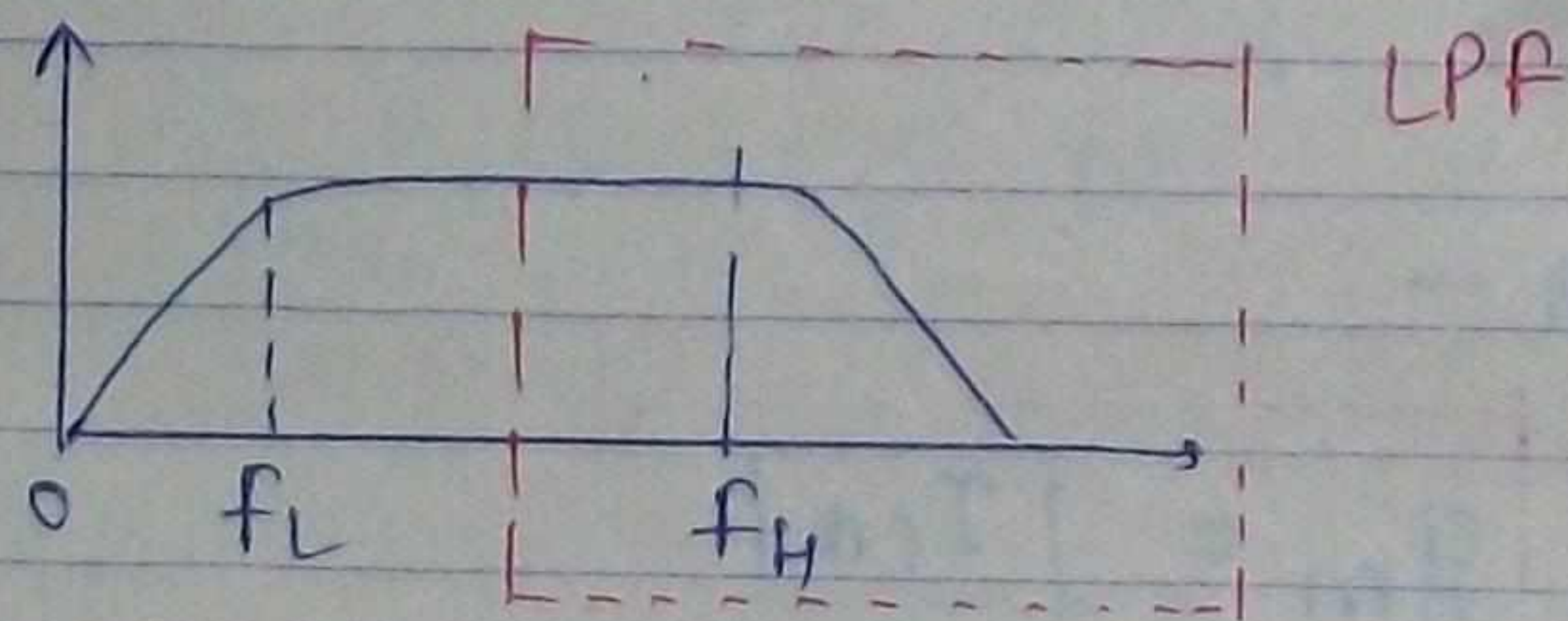
$$r_{\mu} = \frac{r_{\pi}}{h_{re}}$$

$$g_{ce} = h_{oe} - (1 + h_{fe}) g_{b'c}$$

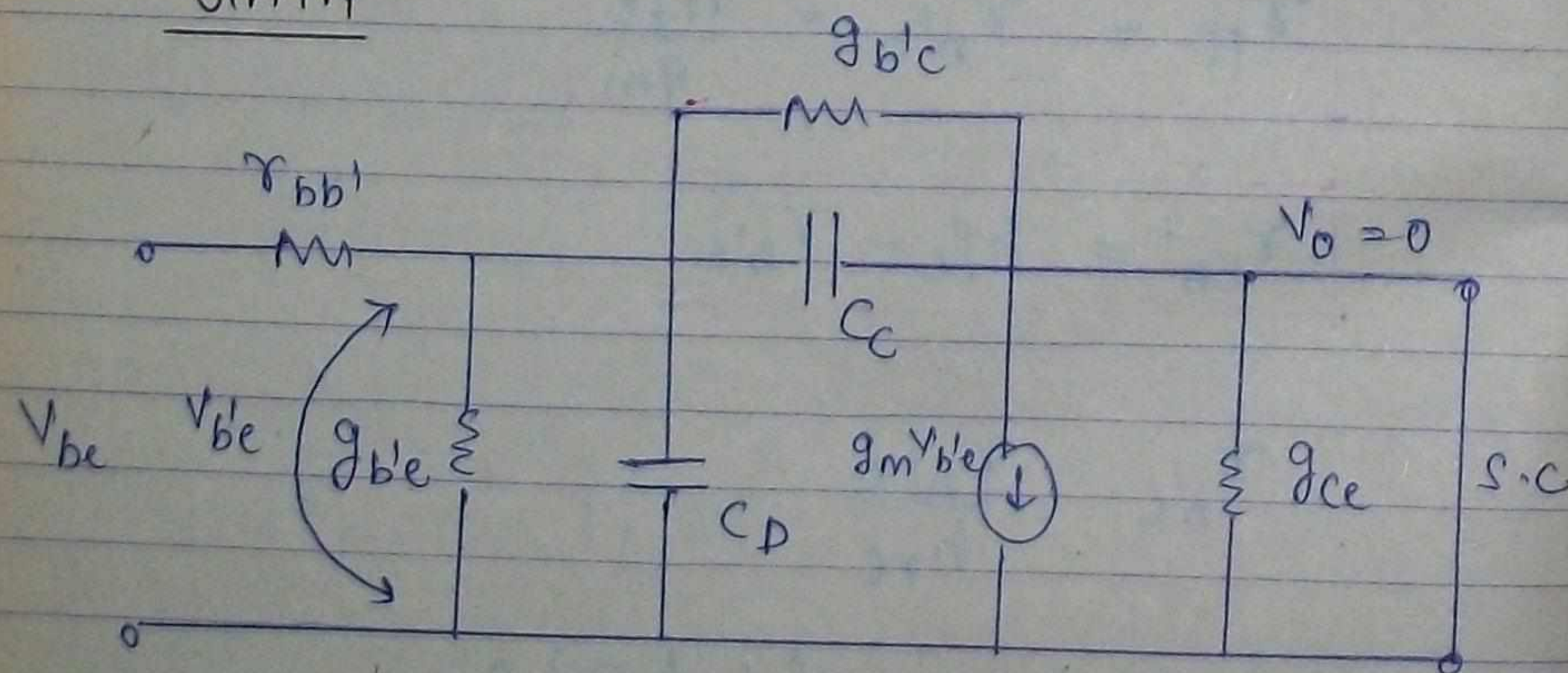
$$r_{b'c} \gg r_{b'e}$$

NOTE -

g_m and r_{π} in model depends on value of DC quiescent current I_{CQ} and hence provide more accurate analysis of Tx. model is applicable to both PNP and NPN Tx without change of polarities.



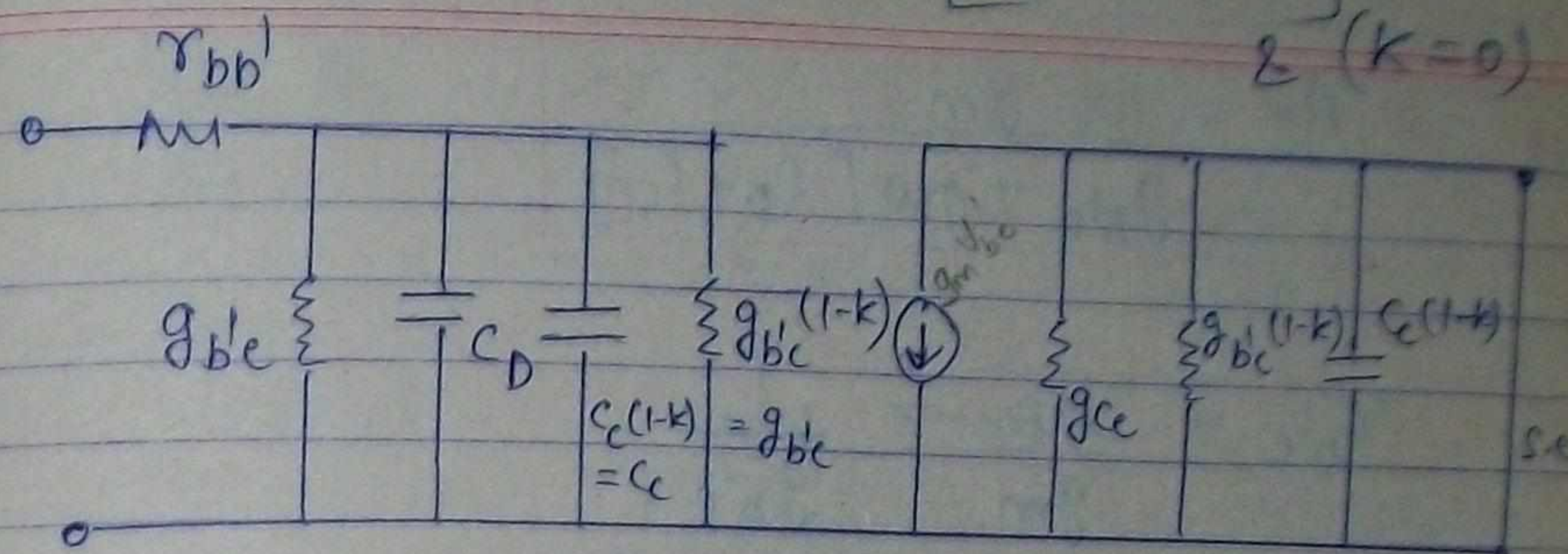
⇒ COMMON EMITTER SHORT CIRCUIT CURRENT GAIN -



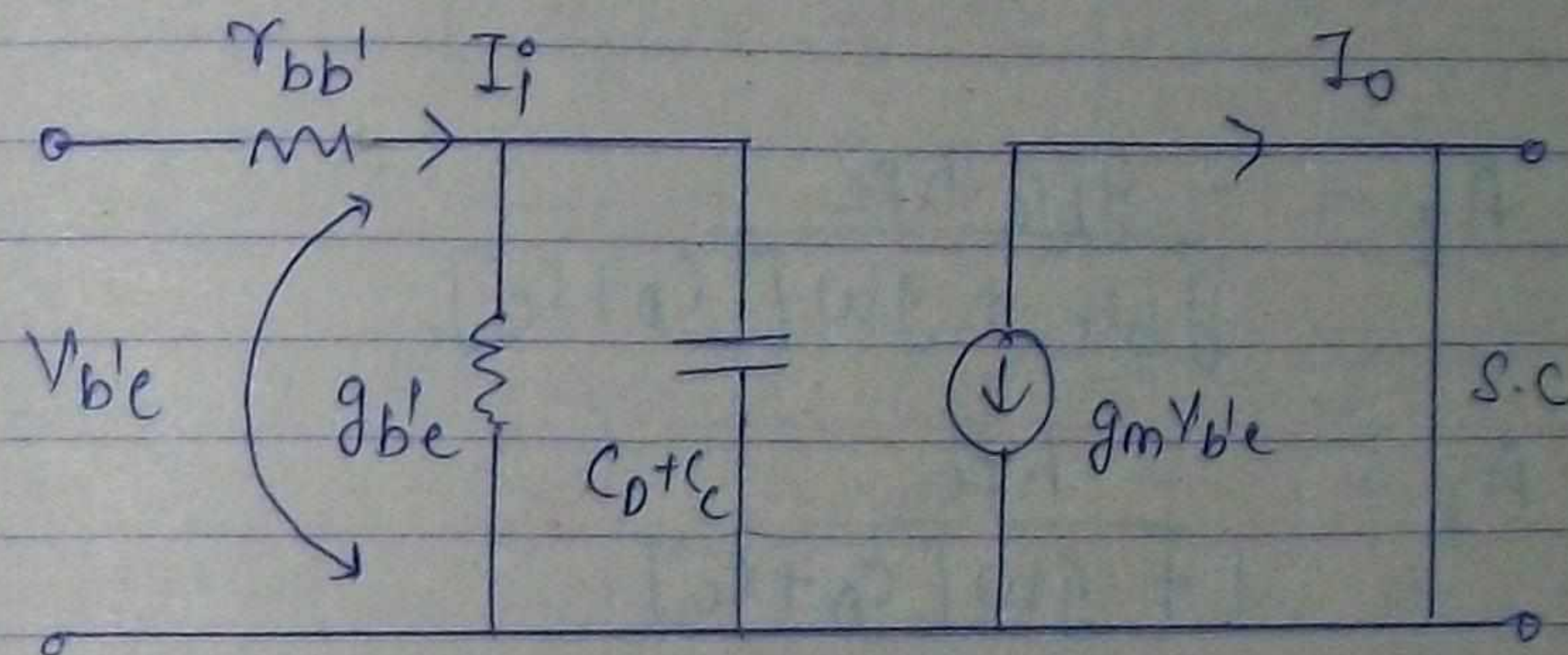
$$K = \frac{V_o}{V_{be}} = \frac{0}{V_{be}} = 0 \quad \text{Apply miller's theorem}$$

(GAIN)

$$[g_{be} > g_{bc}] \quad 2 (K=0)$$



∴ Model at high frequency becomes :



$$A_I = \frac{I_o}{I_i}$$

$$I_o = -g_m V_{be}$$

$$1/p \text{ conductance } Y_i = g_{be} + j\omega [C_D + C_C]$$

$$I_i = V_{be} Y_i$$

$$A_I = \frac{I_o}{I_i} = -\frac{g_m}{Y_i}$$

$$A_I = \frac{-g_m}{g_{b'e} + j\omega[C_D + C_C]}$$

$$r_{b'e} = \frac{h_{fe}}{g_m} = \frac{1}{g_{b'e}}$$

$$g_m = h_{fe} g_{b'e}$$

$$A_I = \frac{-g_{b'e} h_{fe}}{g_{b'e} + j\omega[C_D + C_C]}$$

$$A_I = \frac{-h_{fe}}{1 + j\omega[C_D + C_C] r_{b'e}}$$

A_I for Low pass filter (LPF)

for 3db cutoff freq f_B

$$\text{put } A_I = -h_{fe}/\sqrt{2}$$

$$\omega_B = 2\pi f_B = \frac{g_{b'e}}{C_D + C_C}$$

$$f_B = \frac{g_{b'e}}{2\pi[C_D + C_C]} = \frac{1}{2\pi r_{b'e}[C_D + C_C]}$$

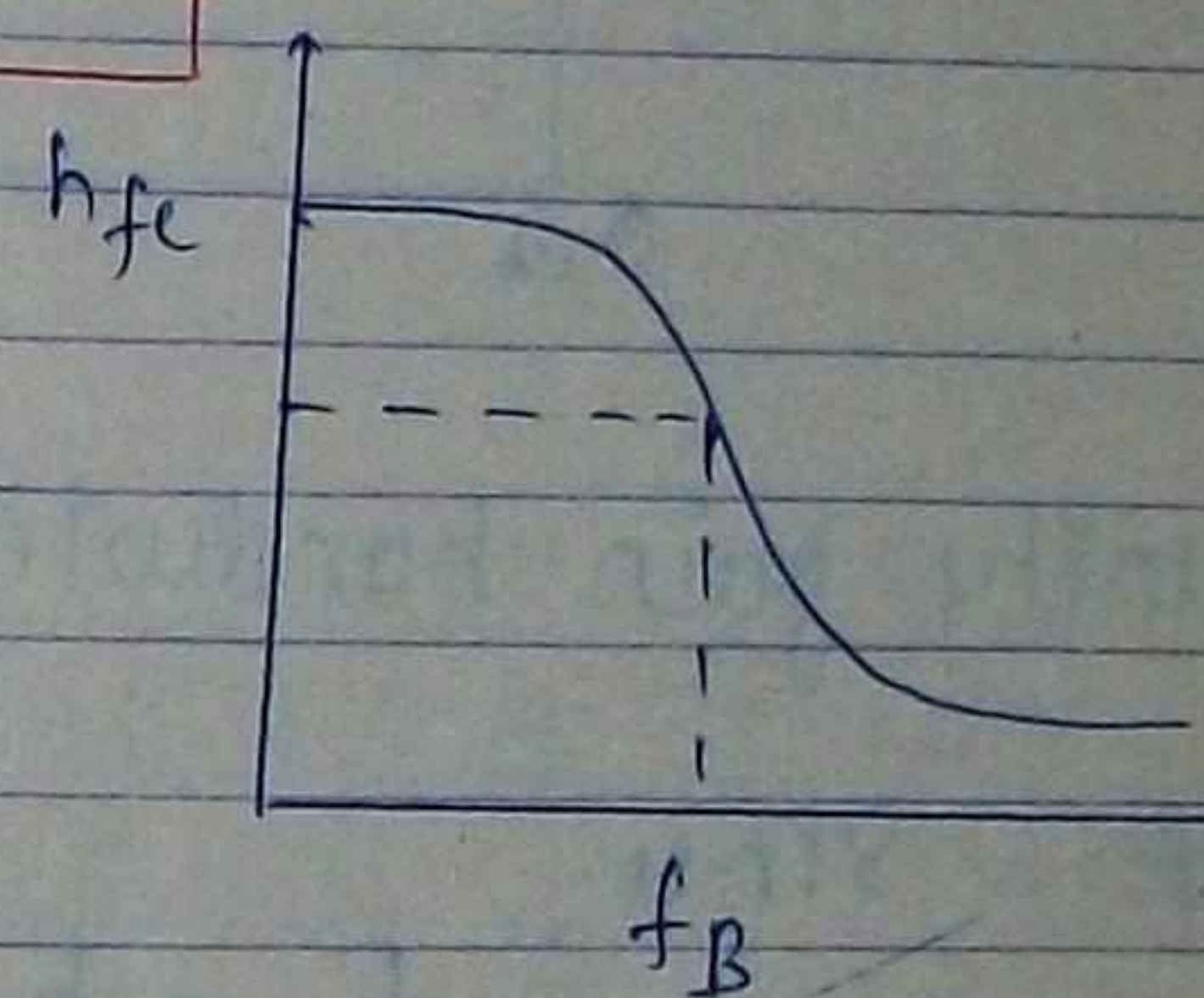
$$A_I = \frac{-h_{fe}}{1 + j\frac{\omega}{\omega_B}} = \frac{-h_{fe}}{1 + j\frac{f}{f_B}}$$

$$|A_I| = \frac{h_{fe}}{\sqrt{1 + (f/f_B)^2}}$$

$$\text{At } f=0 \quad |A| = h_{fe}$$

$$f = f_B \quad |A| = h_{fe}/\sqrt{2}$$

$$f = \infty \quad |A| = 0$$



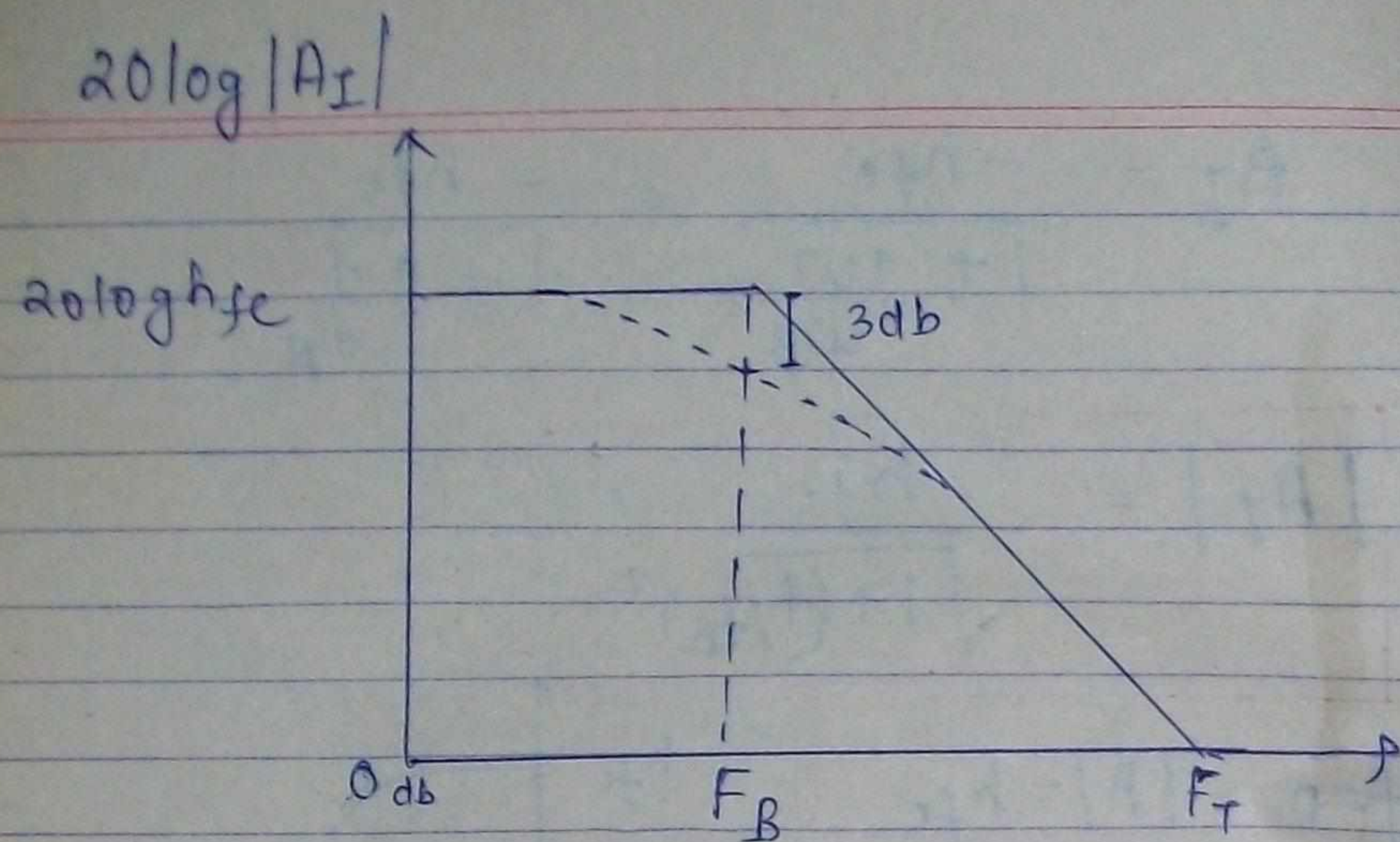
To decide whether $|A_I|$ is amplifier or attenuator

$$\text{let at } f = f_T \quad |A| = 1$$

$$1 = \frac{h_{fe}}{\sqrt{1 + (f/f_B)^2}} \Rightarrow 1 + \left(\frac{f_T}{f_B}\right)^2 = h_{fe}^2$$

$$\left(\frac{f_T}{f_B}\right)^2 = h_{fe}^2 - 1 \approx h_{fe}^2$$

$$f_T = h_{fe} f_B$$



unity gain bandwidth product $= 1 \times f_T = f_T$

Since,

$$f_T = h_{fe} f_B$$

$$f_T = \beta f_B$$

$$f_B = \frac{g_{m/e}}{2\pi [C_c + C_D]}$$

$$f_B = \frac{g_m / h_{fe}}{2\pi [C_c + C_D]}$$

$$h_{fe} f_B = \frac{g_m}{2\pi [C_c + C_D]}$$

$$f_T = \frac{g_m}{2\pi [C_c + C_D]}$$

Since, $g_m = \frac{|I_{cQ}|}{V_T}$ and $C_c \ll C_D$

$$f_T = \frac{g_m}{2\pi C_D}$$

— f_T is defined as frequency where CE short circuit current gain reduces to unity.

— It is also defined as :

1. Highest freq. upto which CE Tx will be working as an amplifier.

2. Where CE Tx β reduces to unity.

3. As unity Gain Bandwidth product of CE Tx. and this GBW product is limited by junction capacitances.

— ALPHA CUTOFF FREQ (f_α) —

$$f_{\alpha} = (1 + h_{fe}) f_{\beta}$$

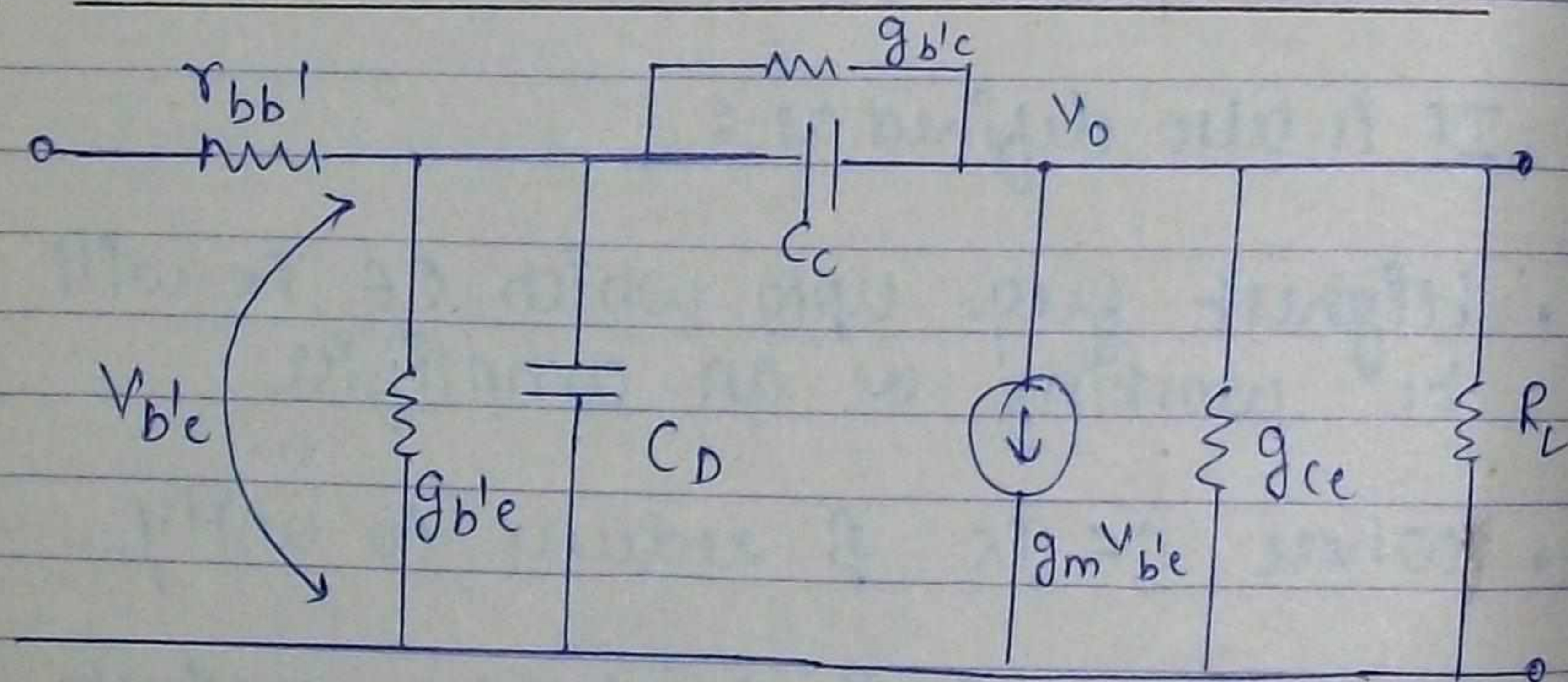
$$f_{\alpha} = (1 + \beta) f_{\beta}$$

also $f_{\beta} \times f_T \leftarrow f_{\alpha} \quad (A_T \downarrow)$

- It is called (f_{α}) Bandwidth of CB Tx. at high frequency.

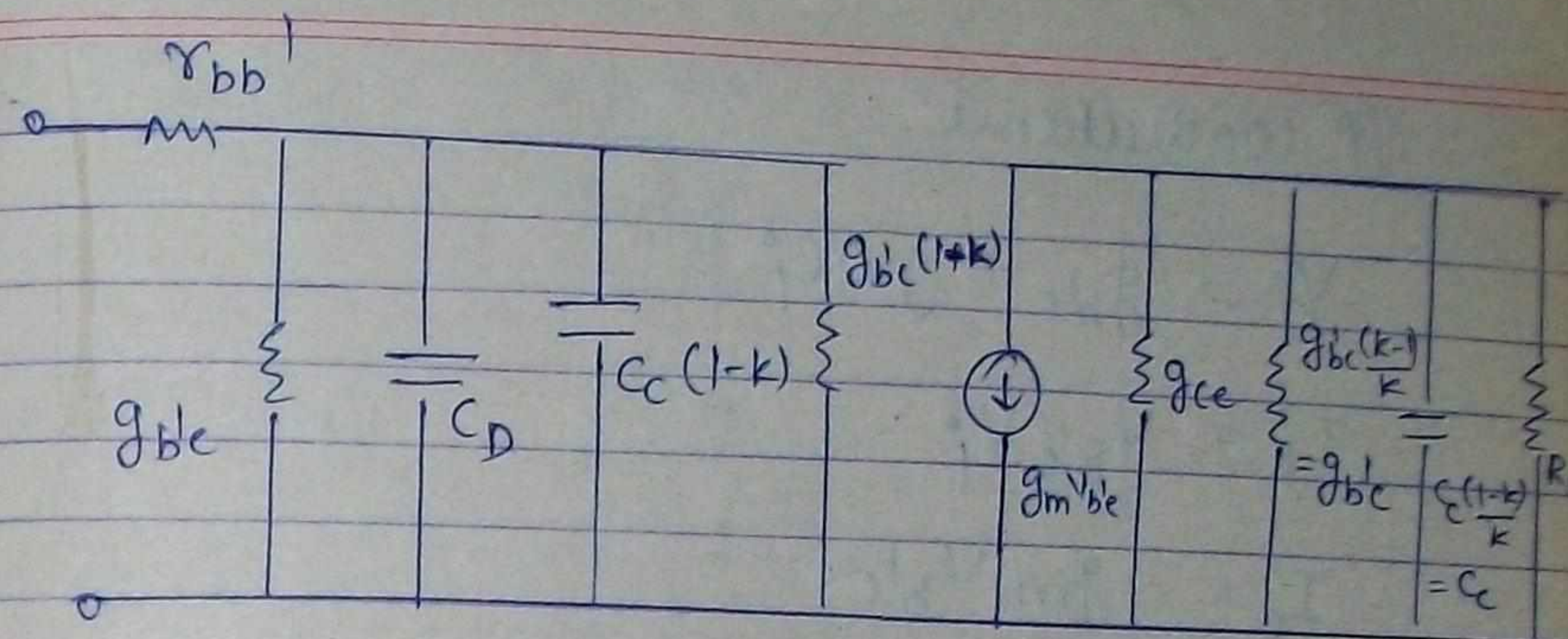
- The BW of CB amplifier is always greater than BW of CE Tx or CC Tx.

⇒ COMMON EMITTER WITH RESISTIVE LOAD -

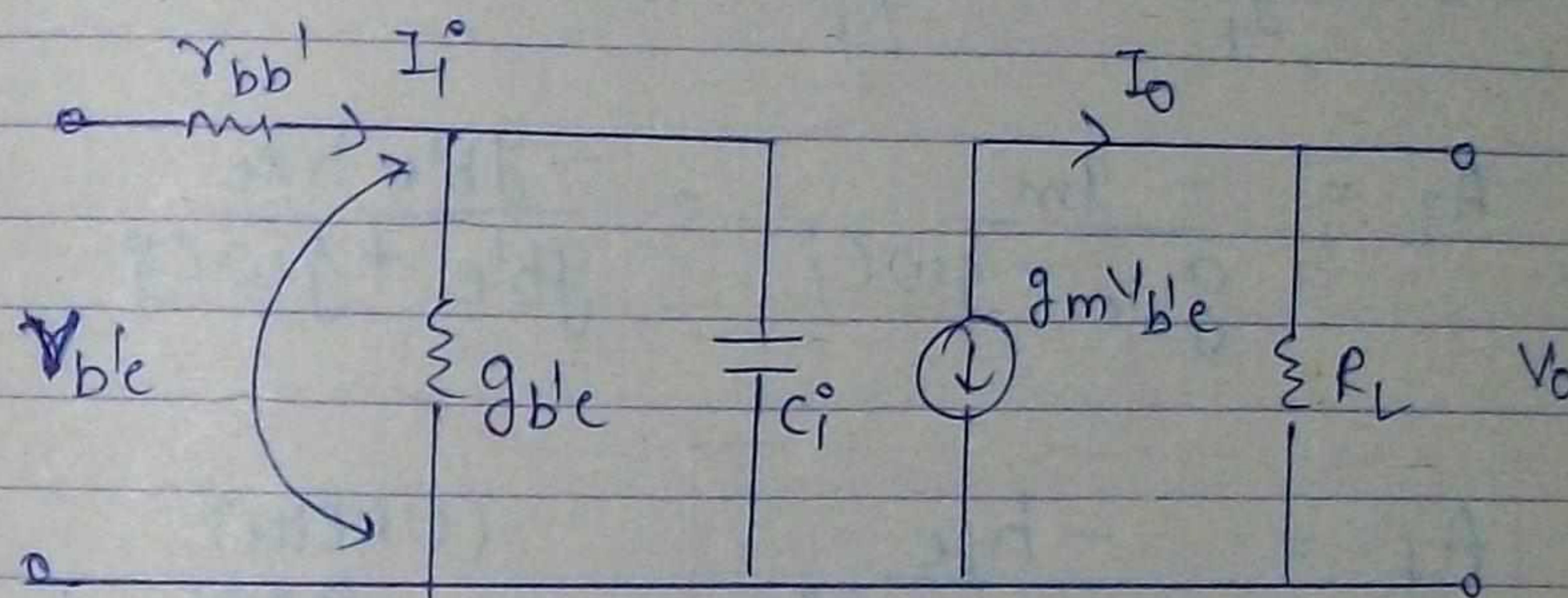


here $k = \frac{V_o}{V_{be}}$ voltage gain $|k| \gg 1$

∴ By millers theorem:



by approximation



$$V_o = -g_m V_{be} R_L$$

$$k = \frac{V_o}{V_{be}} = -g_m R_L$$

VOLTAGE GAIN

$$C_i^o = C_D + C_C [1 + g_m R_L]$$

i/p capacitance

$C_i^o \uparrow$ due to miller effect

1/p conductance

$$Y_i = g_{b'e} + j\omega C_i$$

$$I_i = V_{b'e} Y_i$$

$$I_o = -g_m V_{b'e}$$

$$A_I = \frac{I_o}{I_i} = \frac{-g_m}{Y_i}$$

$$A_I = \frac{-g_m}{g_{b'e} + j\omega C_i} = \frac{-g_{b'e} h_{fe}}{g_{b'e} + j\omega C_i}$$

$A_I = \frac{-h_{fe}}{1 + j\omega \frac{C_i}{g_{b'e}}}$

CURRENT
GAIN

for 3db bandwidth put $A_I = h_{fe}/\sqrt{2}$

$$\omega_H = 2\pi f_H = \frac{g_{b'e}}{C_i} = \frac{g_{b'e}}{C_D + C_C [1 + g_m R_L]}$$

$f_H = \frac{g_{b'e}}{2\pi [C_D + C_C (1 + g_m R_L)]}$

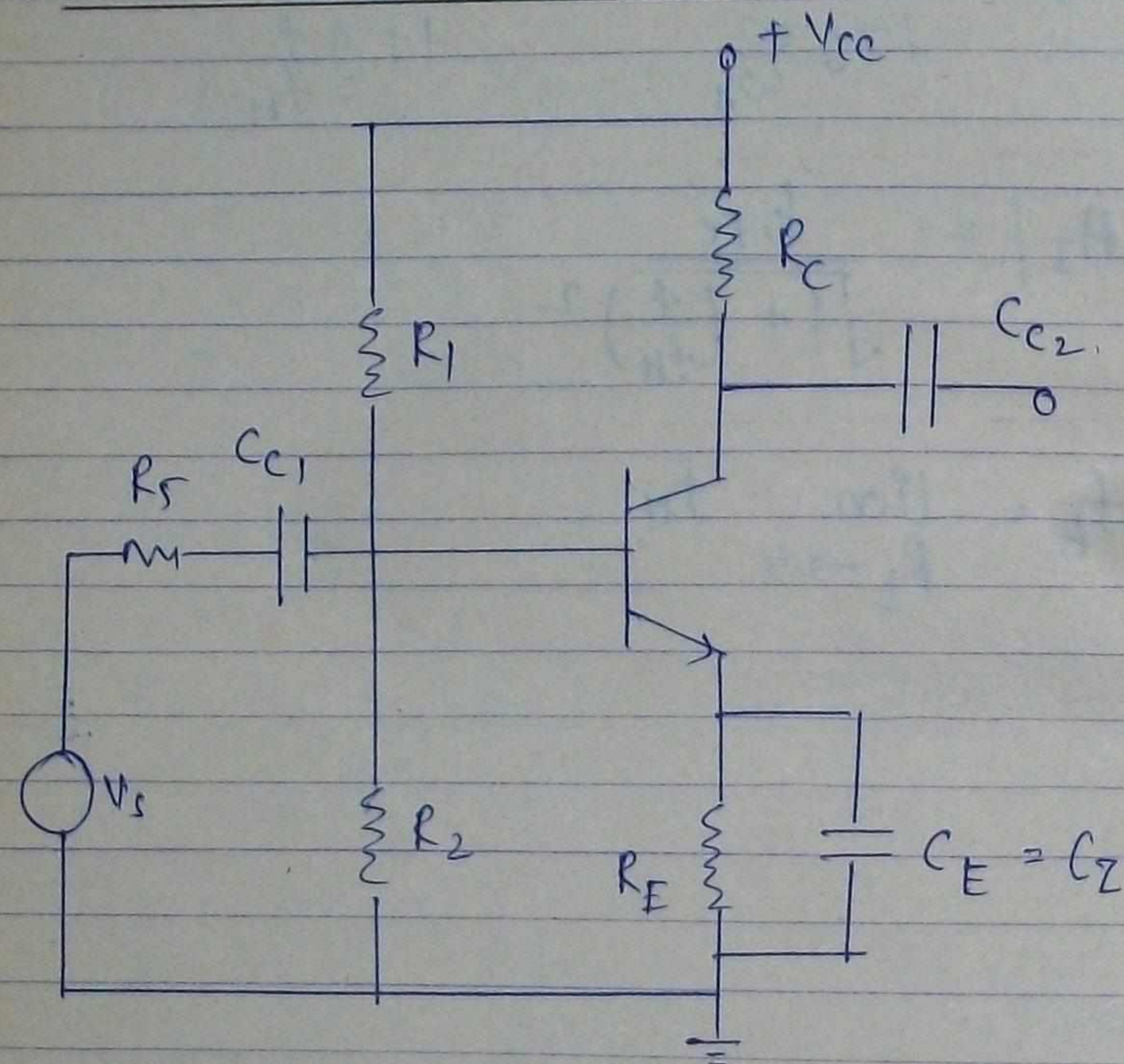
$$A_I = \frac{-h_{fe}}{1 + j\frac{\omega}{\omega_H}} = \frac{-h_{fe}}{1 + j\frac{f}{f_H}}$$

$|A_I| = \frac{h_{fe}}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}}$

$f_B = \lim_{R_L \rightarrow 0} f_H$

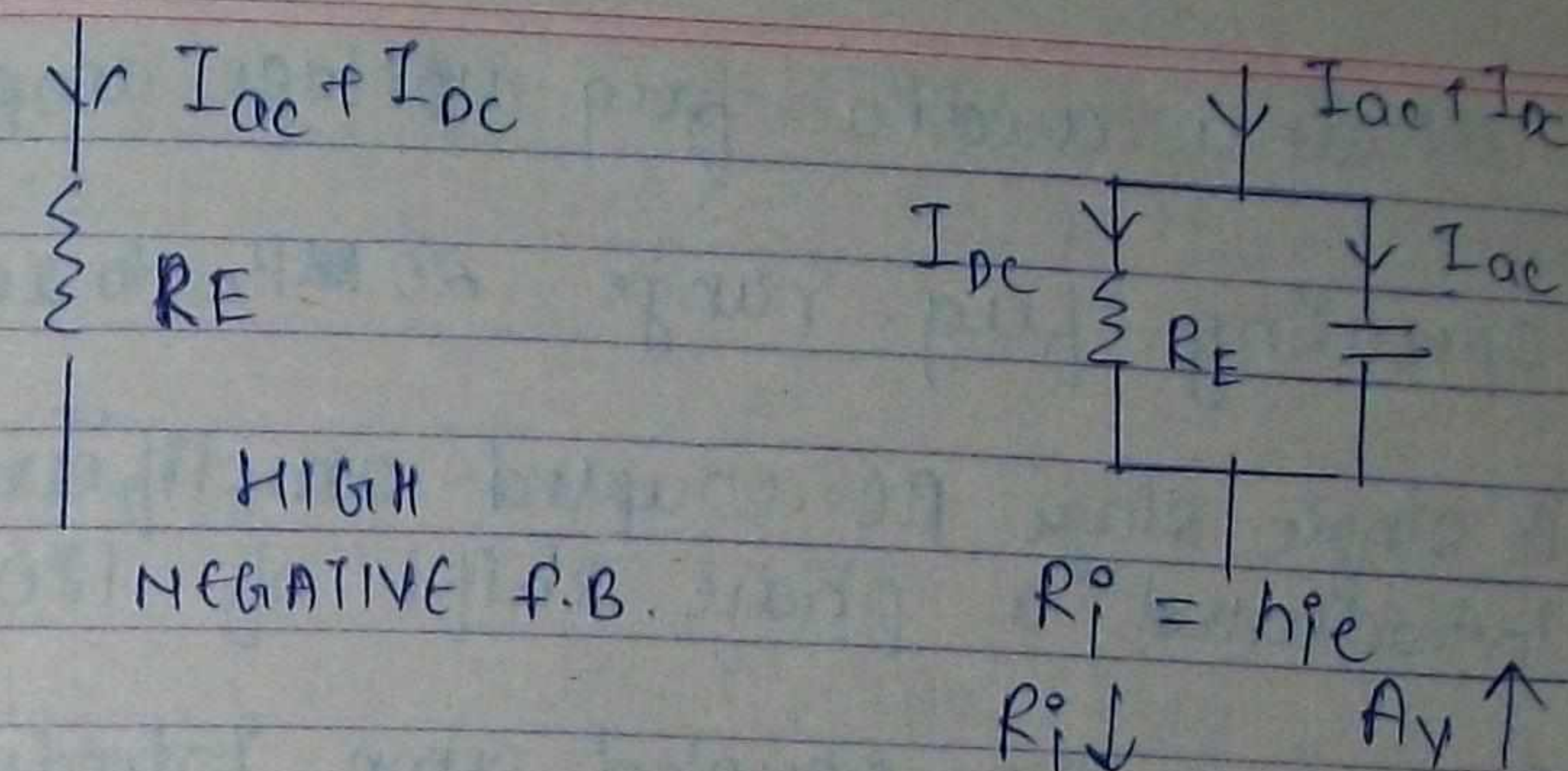
MULTISTAGE AMPLIFIERS

⇒ RC COUPLED AMPLIFIER -

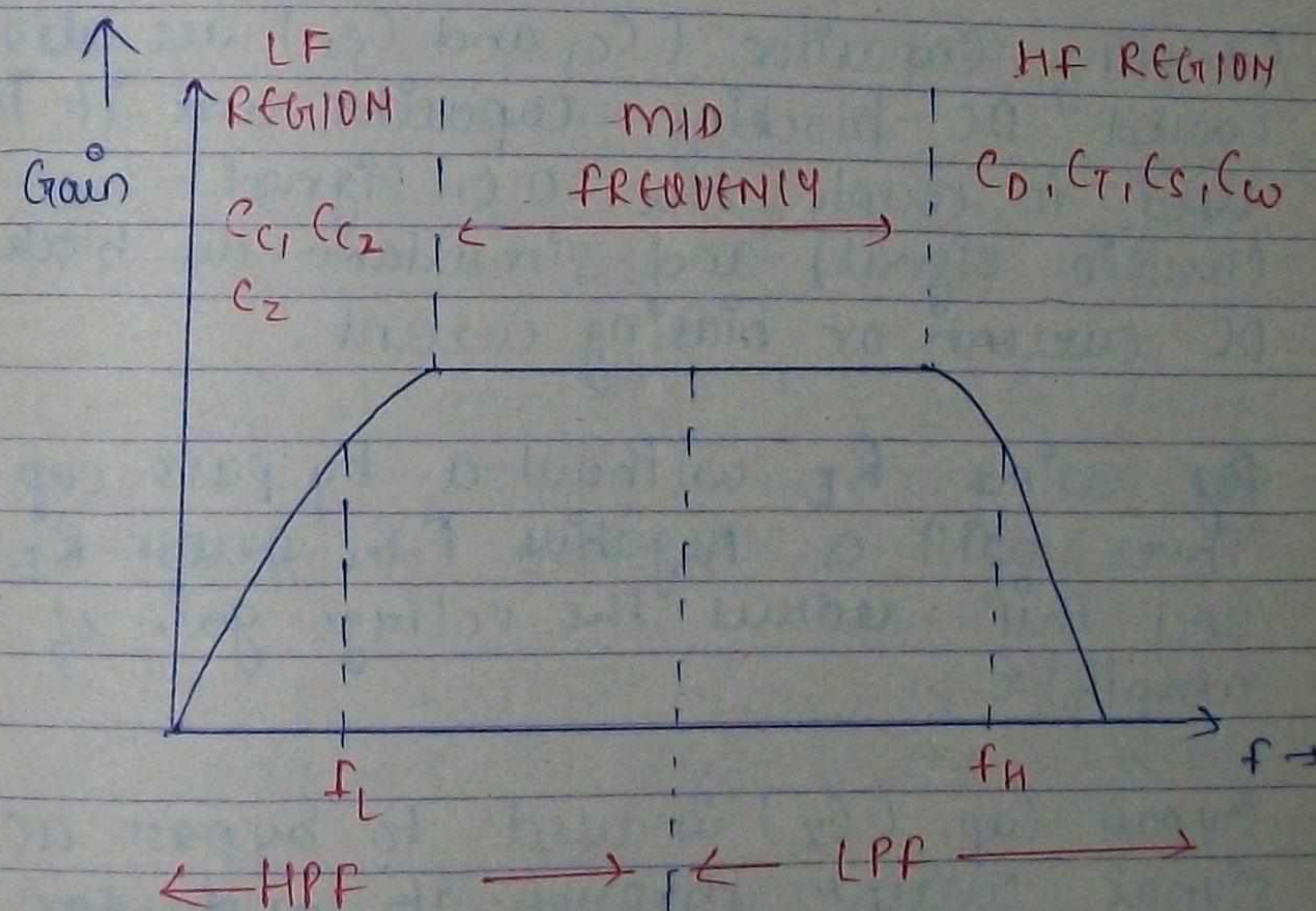


- C_{c1} and C_{c2} are called DC blocking capacitor
- C_{c1} and C_{c2} are very high
- $C_E = C_Z$ is bypass capacitor.

$$Z_{C_Z} = \frac{1}{2\pi f C_Z} \text{ very high.}$$



⇒ FREQUENCY RESPONSE CURVE -



at $f \rightarrow$ mid frequency range

$$C_{c1}, C_{c2}, C_Z = \text{V. HIGH}$$

$$Z_C = \frac{1}{2\pi f C}$$

SHORT CIRCUIT

$$C_D, C_T, C_S, C_W = \text{V. LOW}$$

$$Z_C = 1/2\pi f C$$

OPEN CIRCUIT

- It is a audio freq voltage amplifier
- Operating freq. range 20 Hz to 20 kHz .
- A single stage RC coupled amplifier introduced a phase shift of 180° .
- Two stage RC coupled amp. introduced a phase shift of 360° or 0° .
- Coupling Capacitor (C_c and C_c) are also called DC blocking capacitor and it is used to couple the a.c. signal (audio signal) and simultaneously block DC current or biasing current.
- ~~By~~ using R_E without a by pass cap. There will a negative F.B. across R_E and this reduces the voltage gain of amplifier.
- Bypass Cap. (C_c) is used to bypass a.c. signal current through it and for DC current or biasing current by pass cap. is open.
- By using Bypass cap. -ve f.B. across R_E is eliminated therefore voltage gain of amp. is increased.

- In an amplifier for better performance always R_E and C_c (C_c) combination is used.

FREQUENCY RESPONSE CURVE -

- The fall of gain in low freq region is due to effect of coupling cap. (C_c & C_c) and bypass cap. (C_c).
- In mid freq region all coupling and bypass cap. will be treated as a.c. short and all junction cap. (C_D (C_T (C_w)) will be treated as a.c. open.
- The gain of amplifier is max at mid frequency. Therefore amplifier analysis is generally done at mid frequency.
- The fall of gain at high frequency in a BJT amplifier is due to the effect of:
 1. Junction capacitance (C_D and C_T)
 2. wiring capacitance (C_w)
 3. Stray capacitance (C_s)

- High freq. fall is mainly due to junction capacitor.

- Cutoff freq. (f_H and f_L) is also called 3 dB freq or half power frequencies.

- At cutoff freq. gain of amplifier reduces to 70.7% of peak value.

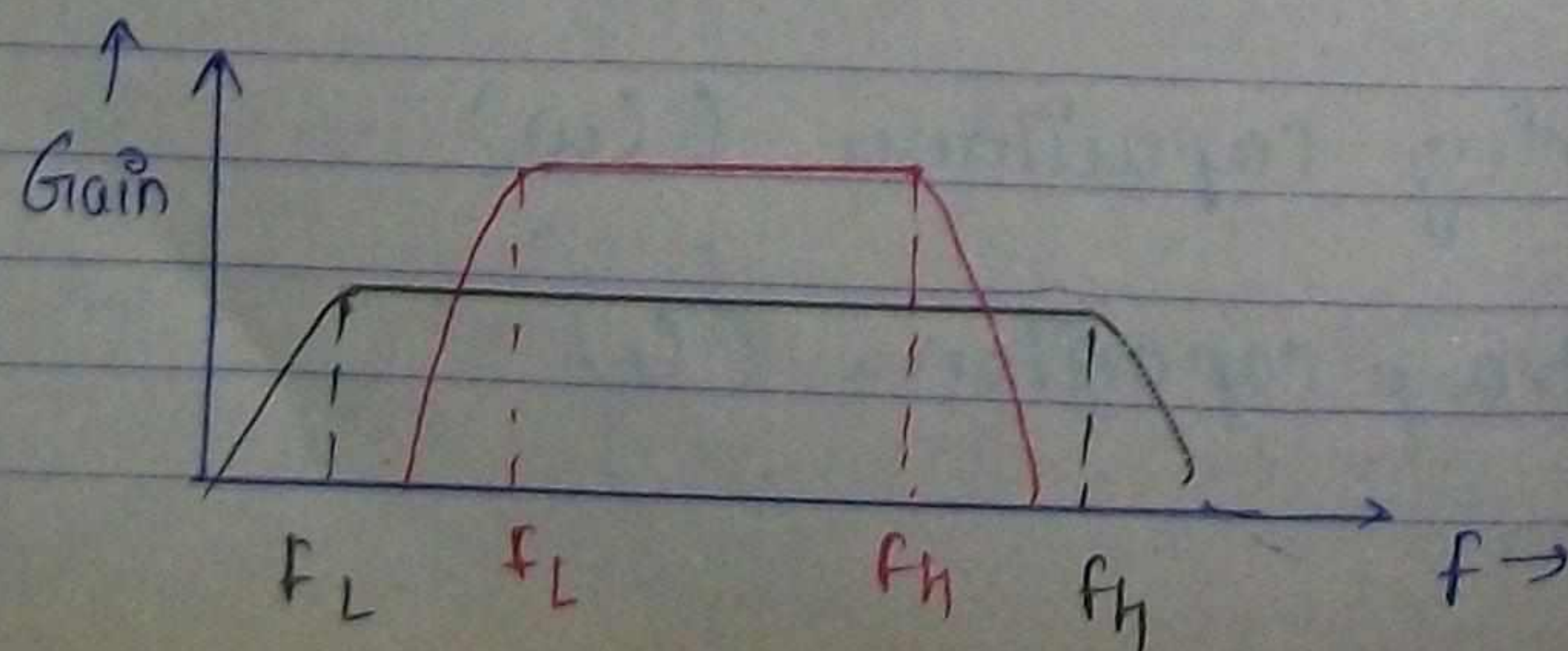
- BANDWIDTH -

It is the band of input signal freqs where the gain of amplifier is almost a constant

$$BW = f_H - f_L$$

- A large BW indicates better reproduction of I/P signal.

- In an amplifier the gain BW product is always a constant. i.e. when one increases the other decreases.

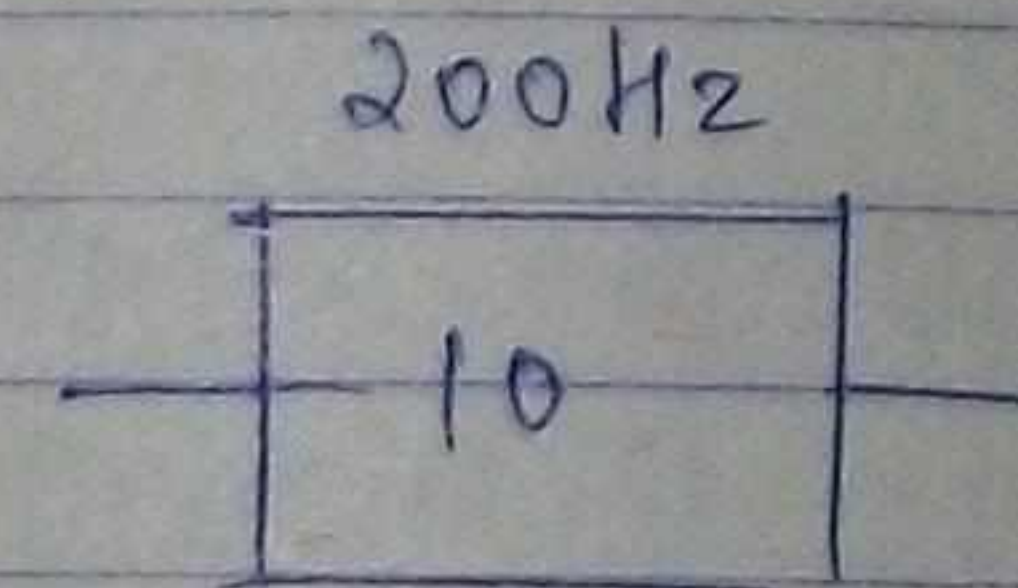


⇒ DISADVANTAGE OF SINGLE STAGE RC COUPLED AMPLIFIER -

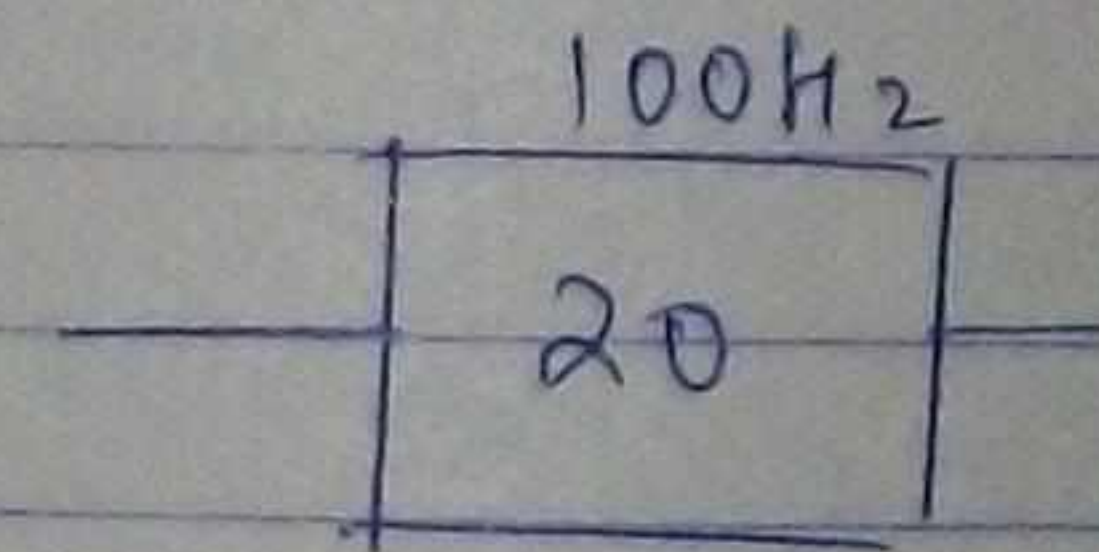
- Smaller gain bandwidth product.

NOTE -

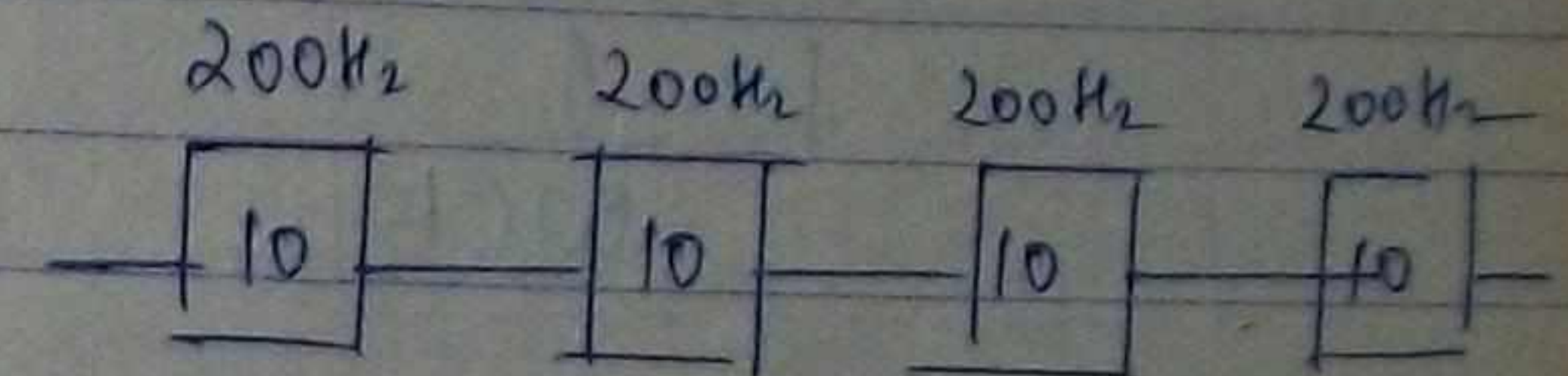
Amplifiers are connected in cascade to get large gain bandwidth product.



$$G \times BW = \text{CONST} \\ = 2000 \text{ Hz}$$



still greater than single stage $\leftarrow$ $G \times BW$



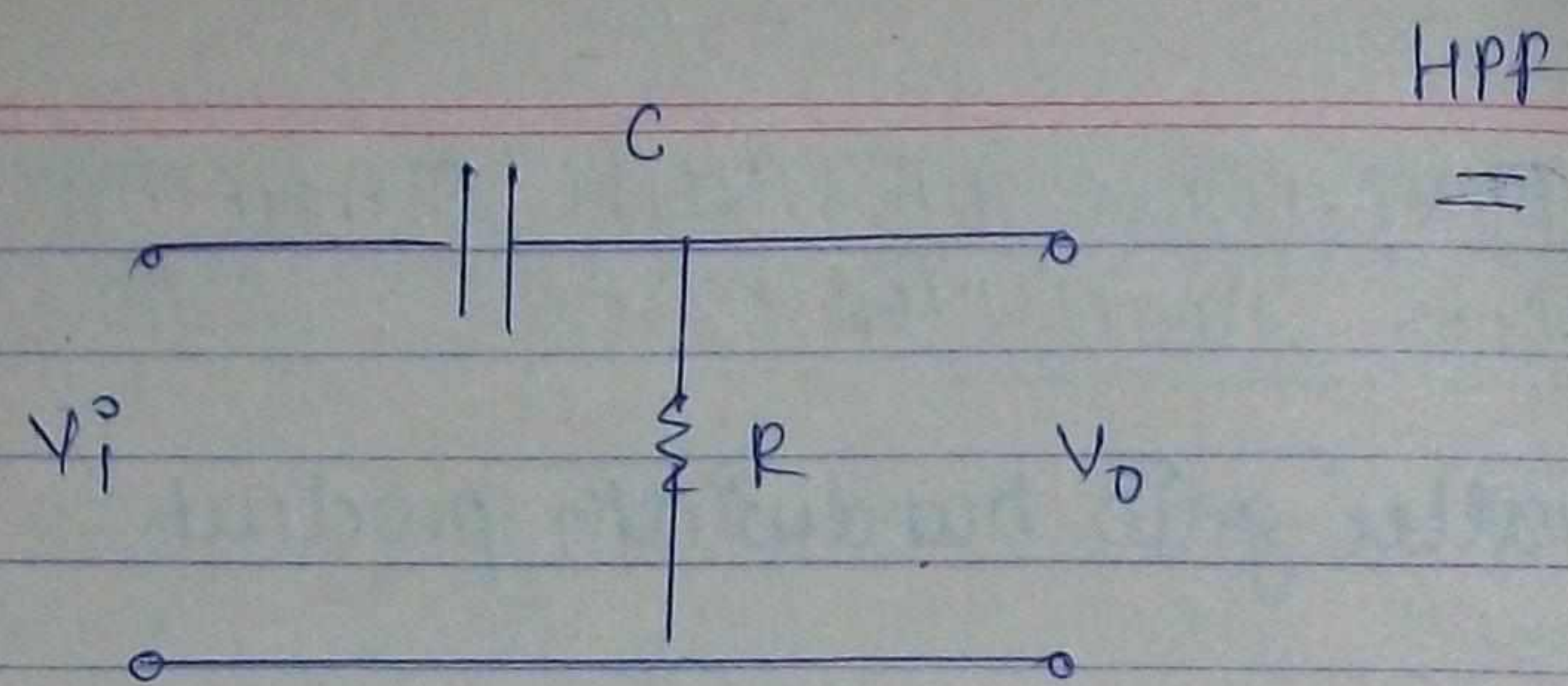
$$10^4 \times 200 \text{ Hz}$$

put practically BW of each stage decreased let 150 Hz

$$\therefore G \times BW = 10^4 \times 150 \text{ Hz}$$

⇒ LOW FREQ ANALYSIS OF RC COUPLED AMPLIFIER -

- Amplifier will act as a high pass filter in low frequency.



$$A = \frac{V_o}{V_i} = \frac{R}{R + \frac{1}{j\omega C}}$$

$$A = \frac{1}{1 - \frac{j}{\omega CR}}$$

$$\omega_L = 2\pi f_L = 1/RC$$

$$A = \frac{1}{1 - \frac{j\omega_L}{\omega}} = \frac{1}{1 - \frac{j f_L}{f}}$$

$$|A| = \frac{1}{\sqrt{1 + \left(\frac{f_L}{f}\right)^2}}$$

$$\phi = \tan^{-1}\left(\frac{f_L}{f}\right)$$

Total phase shift:

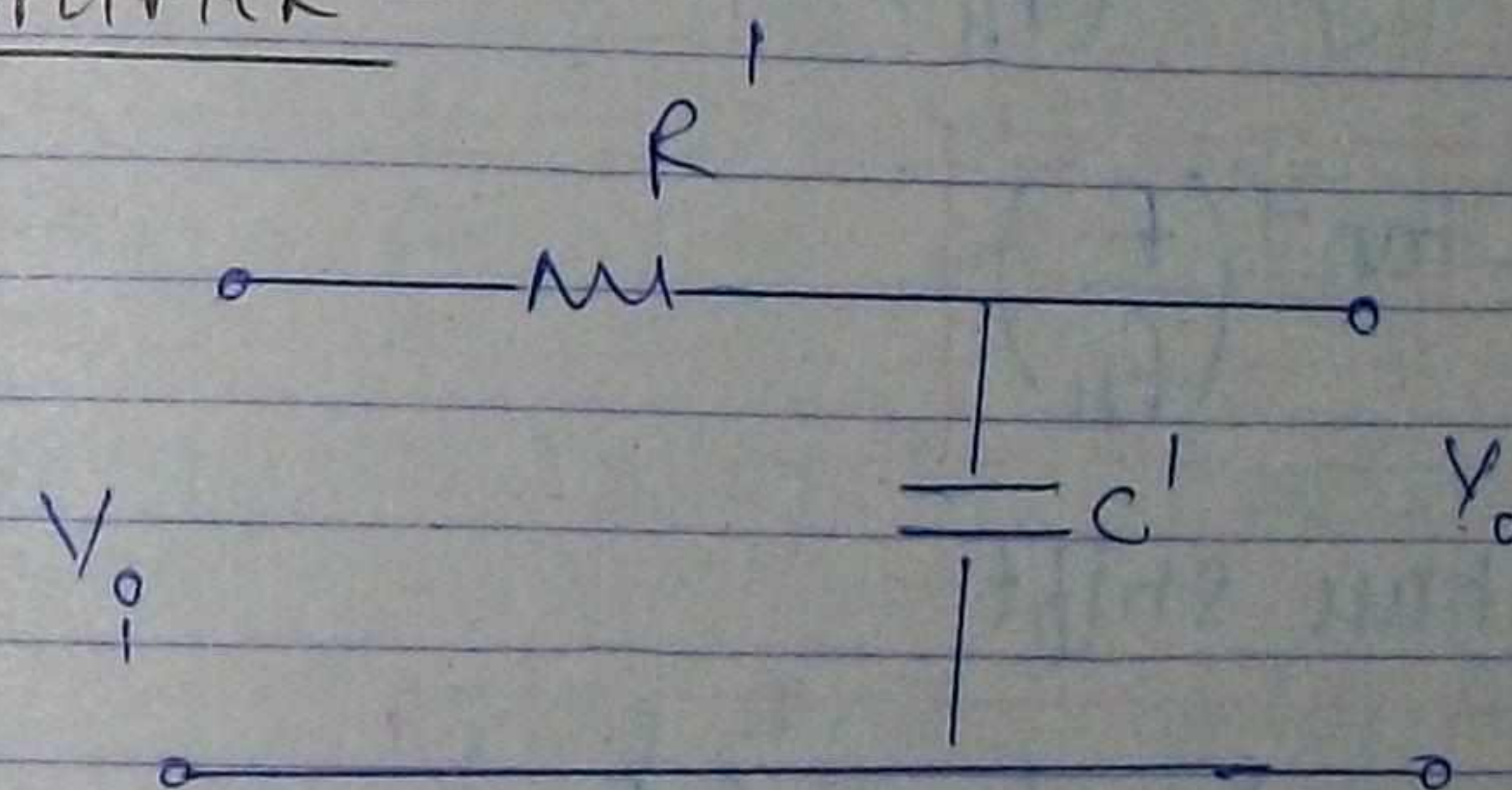
$$\phi_T = 180 + \tan^{-1} \frac{f_L}{f}$$

180 due
to CE config.

at $f = f_L$ = low 3db cutoff freq.

$$\phi = 180 + 45 = 225^\circ$$

⇒ HIGH FREQUENCY ANALYSIS OF RC COUPLED AMPLIFIER -



$$A = \frac{1/j\omega C'}{R' + 1/j\omega C'}$$

$$A = \frac{1}{1 + j\omega R' C'}$$

$$\omega_H = 2\pi f_H = 1/R'C'$$

$$f_H = \frac{1}{2\pi R' C}$$

$$A = \frac{1}{1 + j \frac{\omega}{\omega_H}} = \frac{1}{1 + j \frac{f}{f_H}}$$

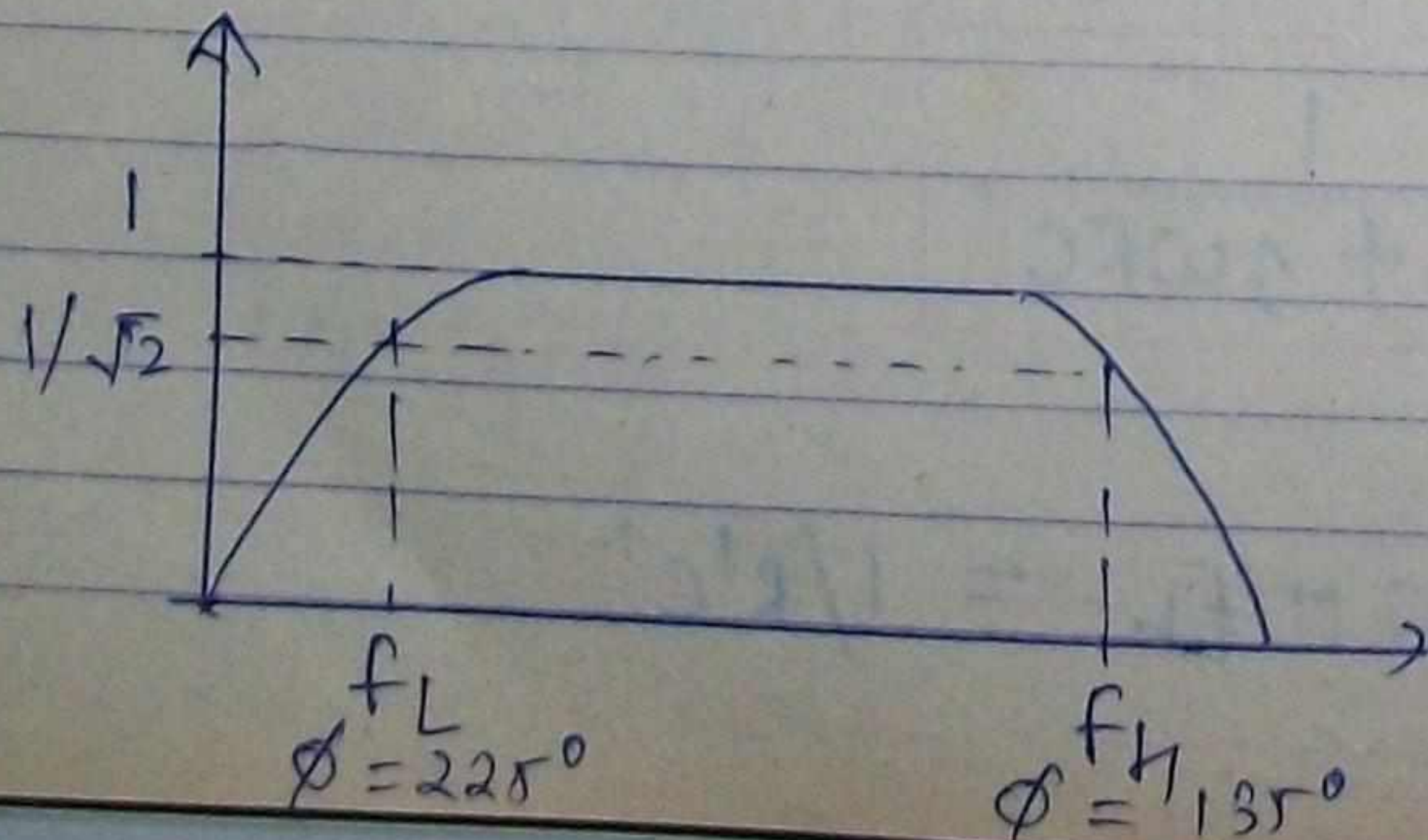
$$|A| = \frac{1}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}}$$

$$\phi = -\tan^{-1}\left(\frac{f}{f_H}\right)$$

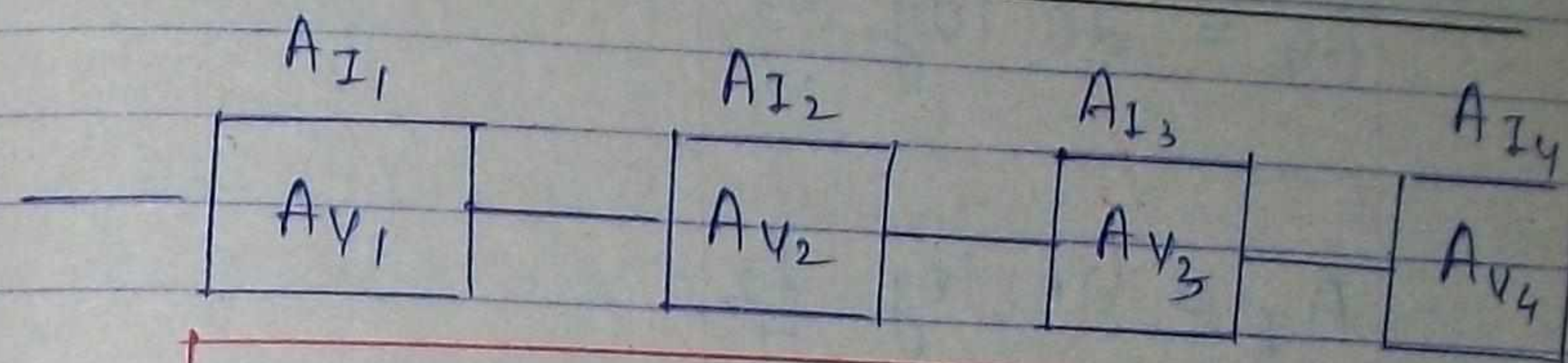
Total phase shift

$$\phi_T = 180^\circ - \tan^{-1}\left(\frac{f}{f_H}\right)$$

at $f = f_H$, $\phi_T = 180 - 45 = 135^\circ$



⇒ CASCADE OR MULTISTAGE AMPLIFIER-



$$A_V = A_{V1} A_{V2} A_{V3} A_{V4}$$

$$20 \log A_V = 20 \log A_{V1} + 20 \log A_{V2} + 20 \log A_{V3} + 20 \log A_{V4}$$

$$A_I = A_{I1} A_{I2} A_{I3} A_{I4}$$

$$20 \log A_I = 20 \log A_{I1} + 20 \log A_{I2} + 20 \log A_{I3} + 20 \log A_{I4}$$

$$A_P = A_V A_I$$

$$(A_P)_{dB} = 10 \log \frac{P_2}{P_1}$$

$$P_2 = V_2^2 \quad P_1 = V_1^2$$

$$(A_P)_{dB} = 10 \log \frac{V_2^2}{V_1^2}$$

$$A_V = 20 \log \frac{V_2}{V_1}$$

$$A_I = 20 \log \frac{I_2}{I_1}$$

- Amplifiers are connected in cascade to get large gain bandwidth product.
- when Amplifiers are connected such that the o/p of one is given to i/p of another. They are said to be cascaded.
- when amplifiers are cascaded proper impedance matching must be provided in b/w stages so that:
 1. Max power will be transferred from one to another stage.
 2. O/P will not be distorted.
- Three different types of coupling
 1. RC coupling (audio voltage amp)
 2. Direct coupling (DC amp.)
 3. Transformer coupling (Power amp.)

- In multistage amp. Gain BW product is always a constant.

- The Gain BW product of two stage amp. is greater than that of single stage amp.

- In multistage amp. BW is reduced.

⇒ BANDWIDTH OF MULTISTAGE AMPLIFIER -

$BW^* =$ BW of multistage amplifier.

$$BW^* = f_H^* - f_L^*$$

$f_H^* =$ High 3db cutoff freq of multistage amp

$f_L^* =$ Low 3db ——— || ——— || ———

CASE (I) : N IDENTICAL STAGES IN CASCADE -

$$BW = f_H - f_L$$

$BW =$ BW of individual stage

$f_H =$ High 3db cutoff freq of individual amp

$f_L =$ Low ——— || ——— || ———

(A) f_H^* -

for individual stage $A_I = \frac{1}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}}$

$$|A_I^*| = \left[\frac{1}{\sqrt{1 + \left(\frac{f}{f_H}\right)^2}} \right]^n$$

at $f = f_H^*$ $|A_I^*| = 1/\sqrt{2}$

$$\frac{1}{\sqrt{2}} = \left[\frac{1}{\sqrt{1 + \left(\frac{f_H^*}{f_H}\right)^2}} \right]^n$$

★ $f_H^* = f_H \sqrt{2^{1/n} - 1}$

• $f_H^* < f_H$

$n = 2$ $f_H^* = 0.64 f_H$ ✓

$n = 3$ $f_H^* = 0.51 f_H$ ✓

(B) f_L^* -

for individual stage $A_I = \frac{1}{\sqrt{1 + \left(\frac{f_L}{f}\right)^2}}$

$$|A_I^*| = \left[\frac{1}{\sqrt{1 + \left(\frac{f_L}{f}\right)^2}} \right]^n$$

at $f = f_L^*$ $|A_I^*| = 1/\sqrt{2}$

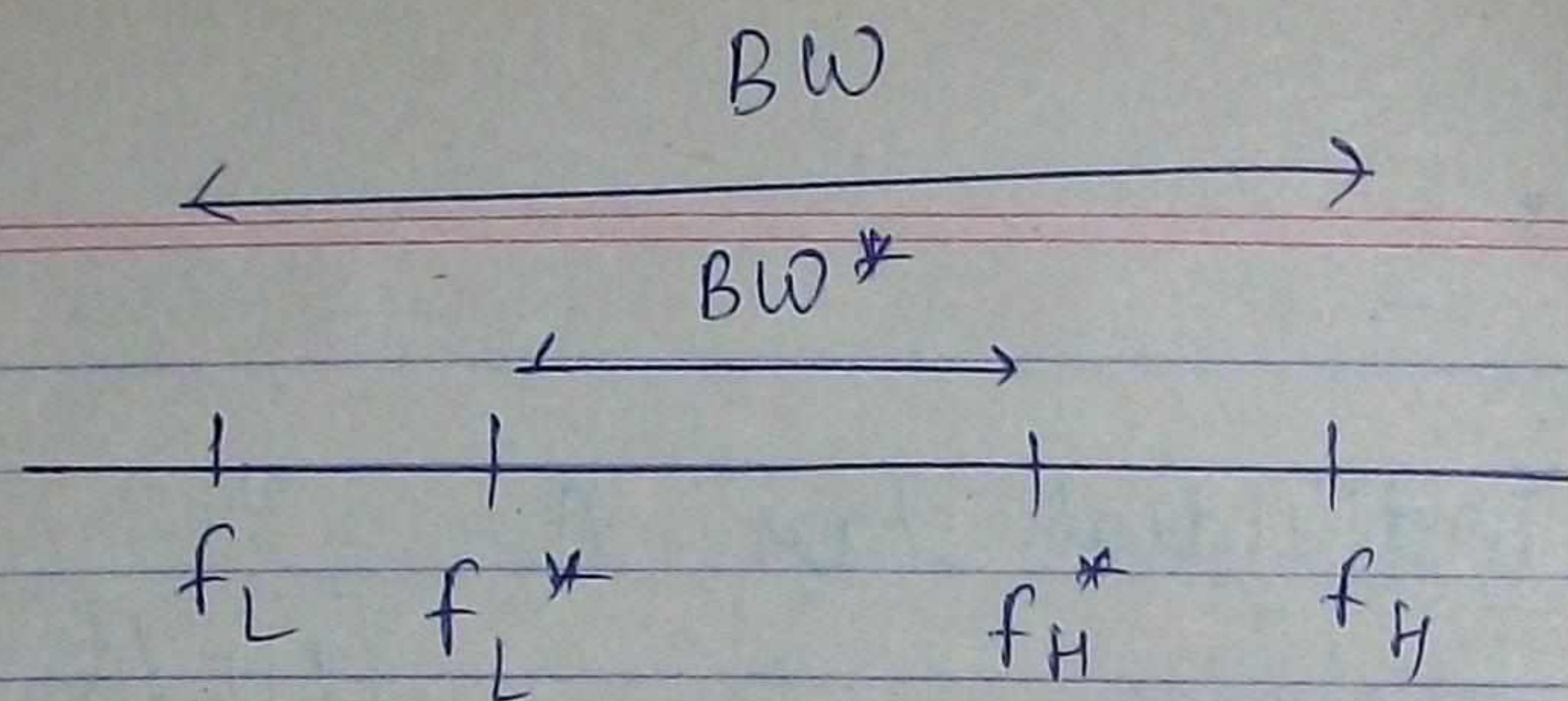
$$\frac{1}{\sqrt{2}} = \left[\frac{1}{\sqrt{1 + \left(\frac{f_L}{f_L^*}\right)^2}} \right]^n$$

★ $f_L^* = \frac{f_L}{\sqrt{2^{1/n} - 1}}$

• $f_L^* > f_L$

$n = 2$ $f_L^* = 1.5625 f_L$ ✓

$n = 3$ $f_L^* = 1.98 f_L$ ✓



$$BW^* < BW$$

— There is reduction in bandwidth by using multistage

CASE (2) : n - NON IDENTICAL INTERACTING STAGE (NO PROPER IMPEDANCE MATCHING) IN CASCADE

$$\frac{1}{F_H^*} = 1.01 \times \sqrt{\frac{1}{f_{H1}^2} + \frac{1}{f_{H2}^2} + \dots + \frac{1}{f_{Hn}^2}}$$

$$f_L^* = 1.01 \times \sqrt{f_{L1}^2 + f_{L2}^2 + \dots + f_{Ln}^2}$$

RISE TIME -

$$t_r^* = 1.01 \times \sqrt{t_{r1}^2 + t_{r2}^2 + \dots + t_{rn}^2}$$

TOTAL RISE TIME -

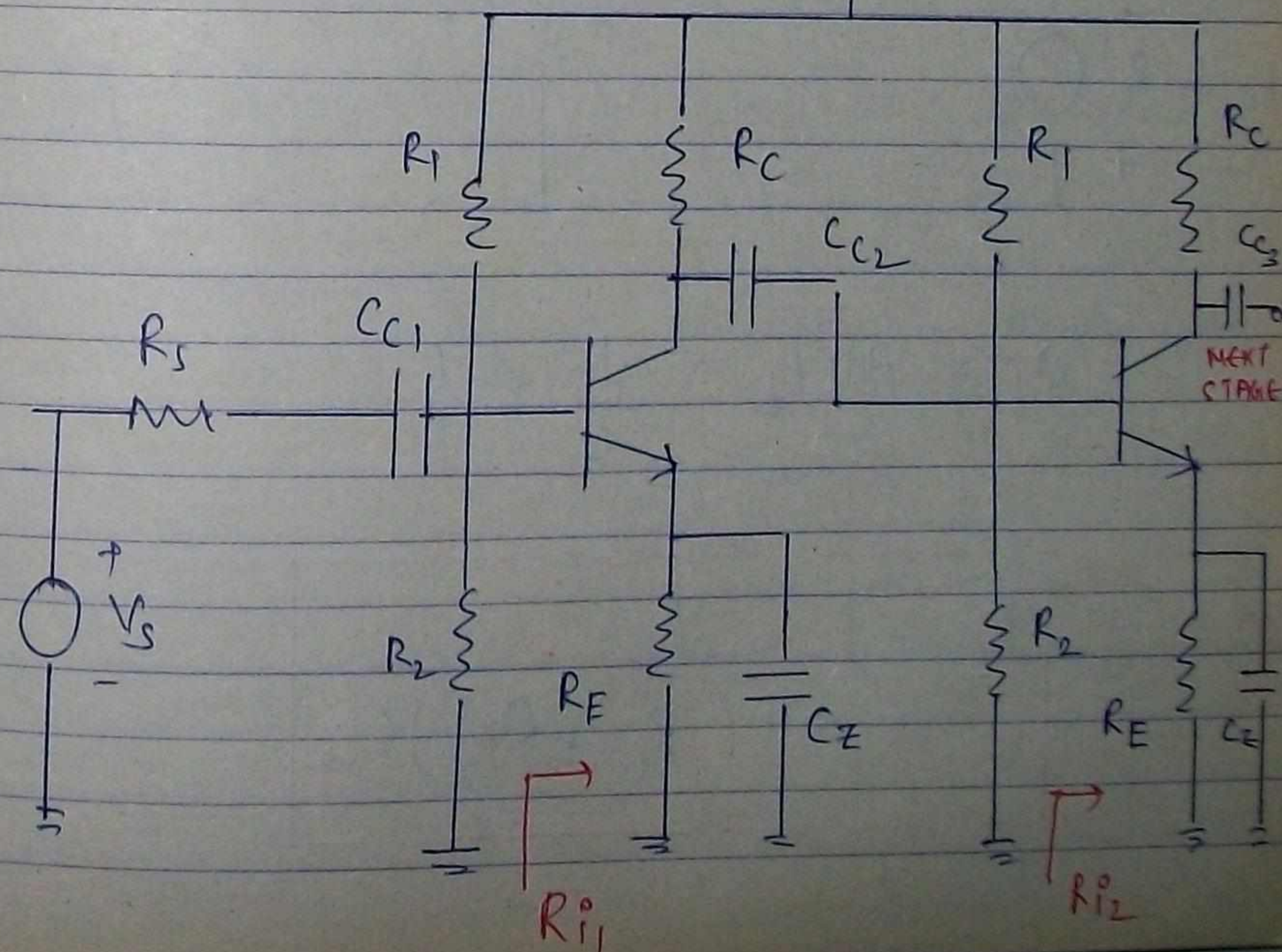
$$t_r^* = 1.01 \times \sqrt{t_{r0}^2 + t_{r1}^2 + \dots + t_{rn}^2}$$

t_{r0} = rise time of I/P signal.

t_{rn} = rise time of individual stage.

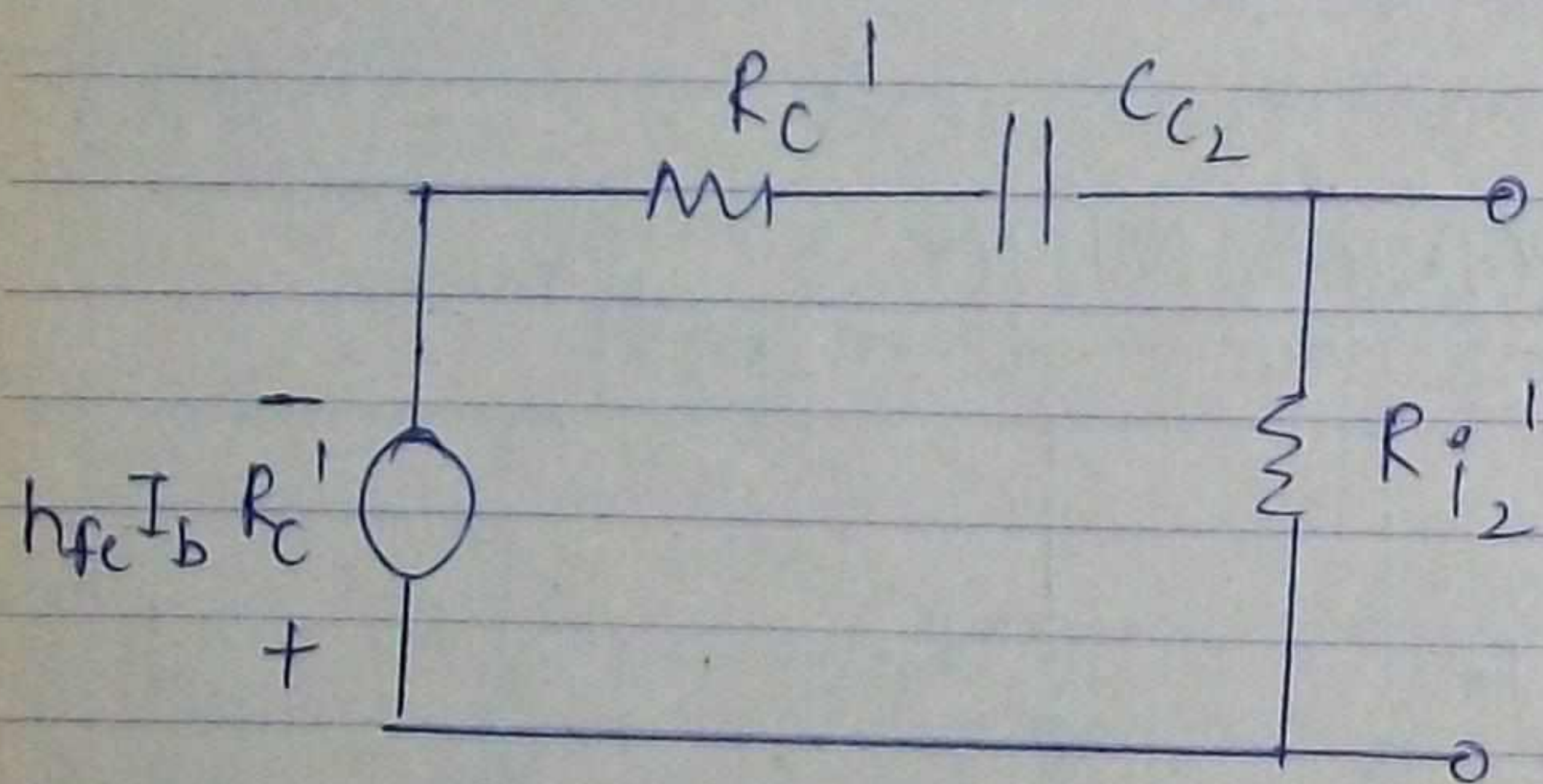
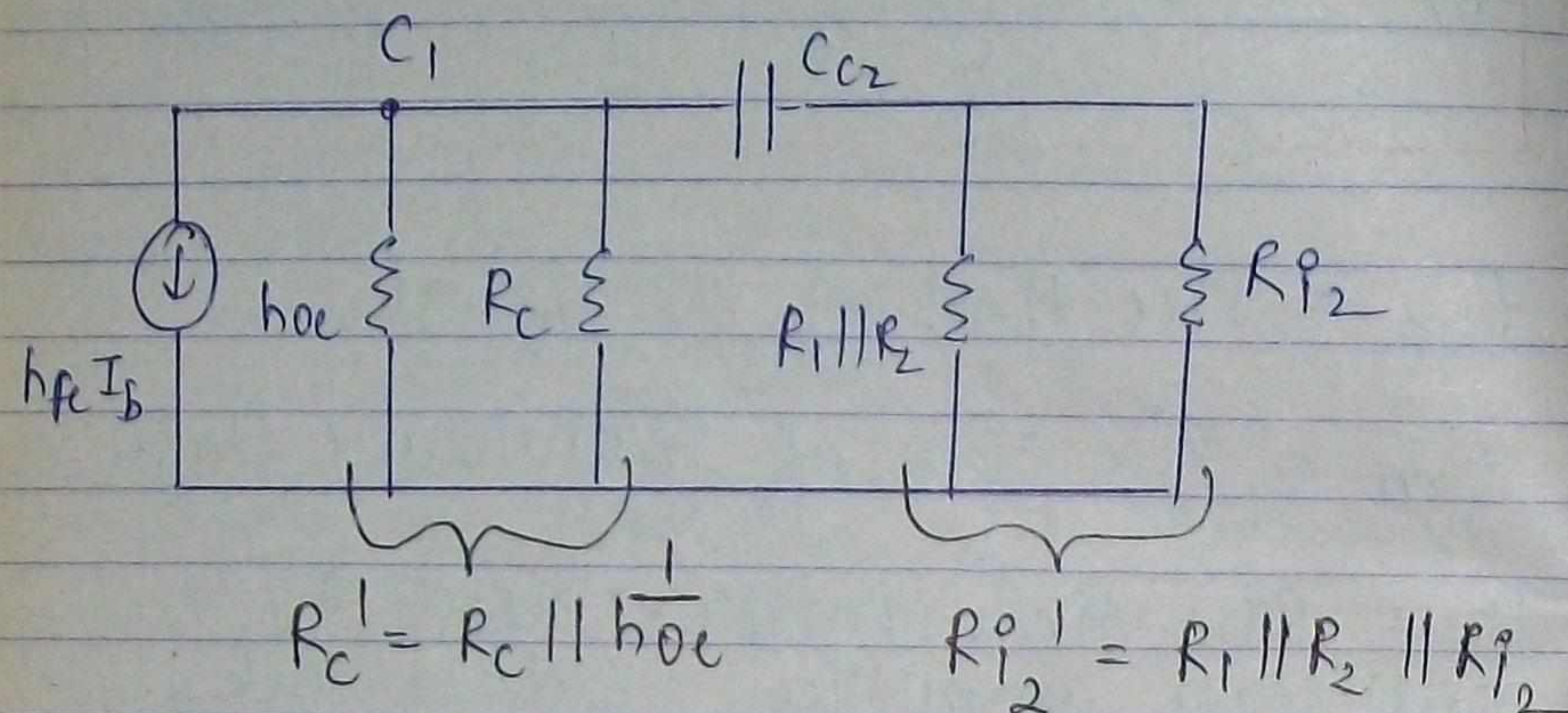
Thus Rise time increase due to multistage amplifier (draw back)

$\Rightarrow$ RC COUPLED AMPLIFIER - $+V_{CC}$



① DUE TO C_{C2} —

let $C_{C1} = C_{C3} = C_Z = \infty$

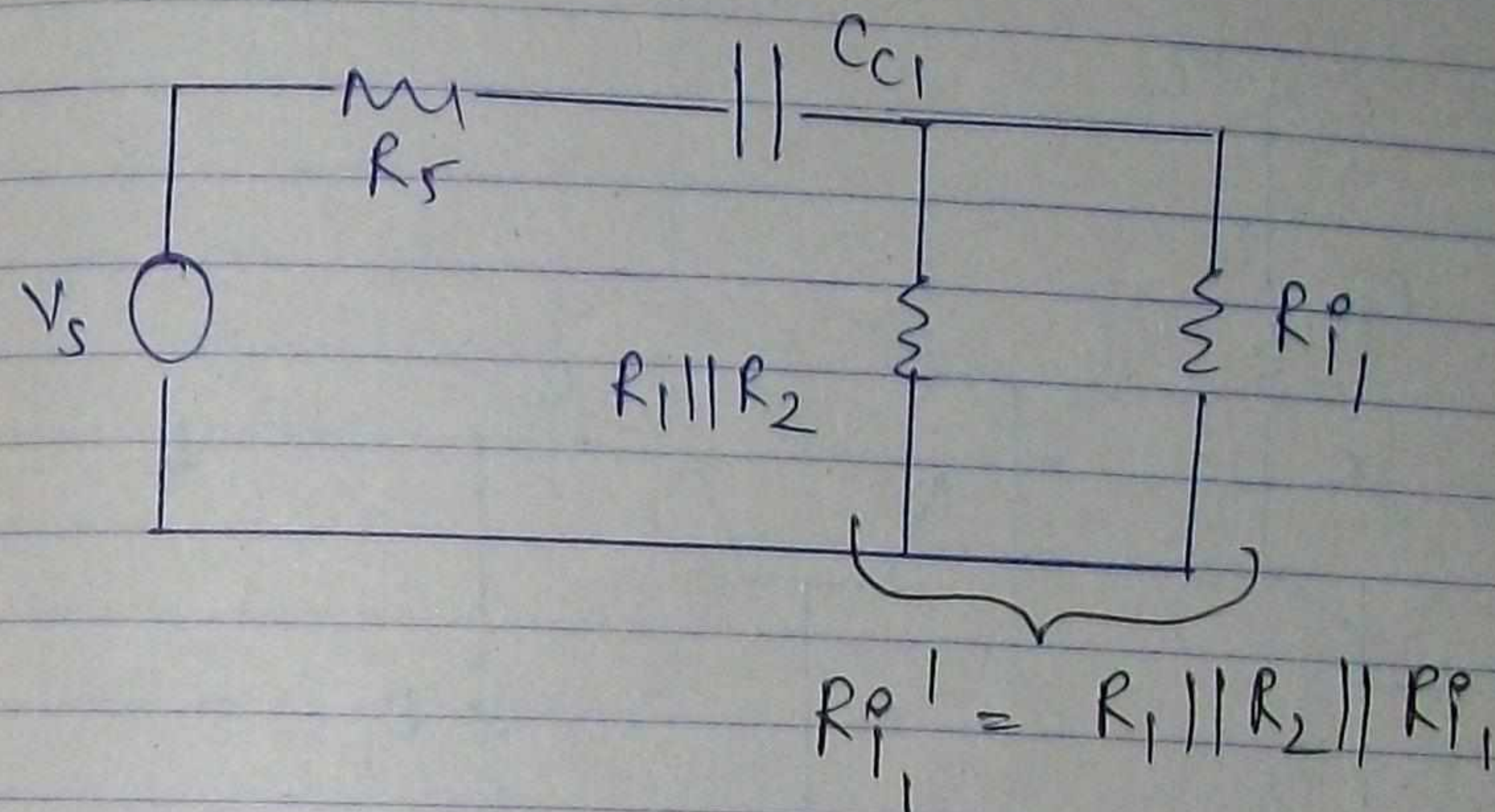


$$\omega_L = 2\pi f_L = \frac{1}{\tau} = \frac{1}{R_C}$$

$$f_L = \frac{1}{2\pi (R_C' + R_{i_2}') C_{C2}}$$

② DUE TO C_{C1} —

let $C_{C2} = C_{C3} = C_Z = \infty$

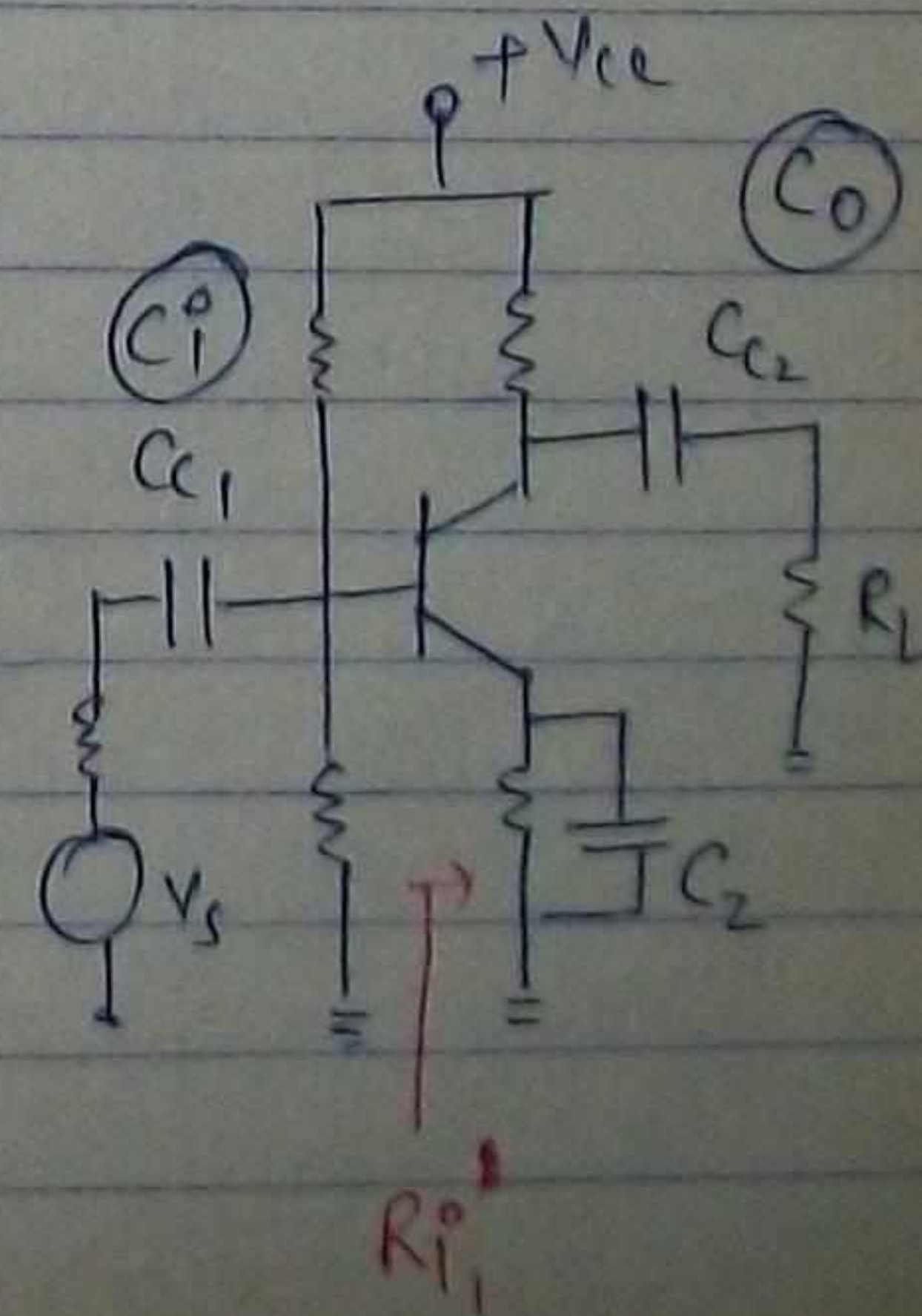


$$f_L = \frac{1}{2\pi (R_S + R_{i_1}') C_{C1}}$$

for single stage RC coupled amplifier.

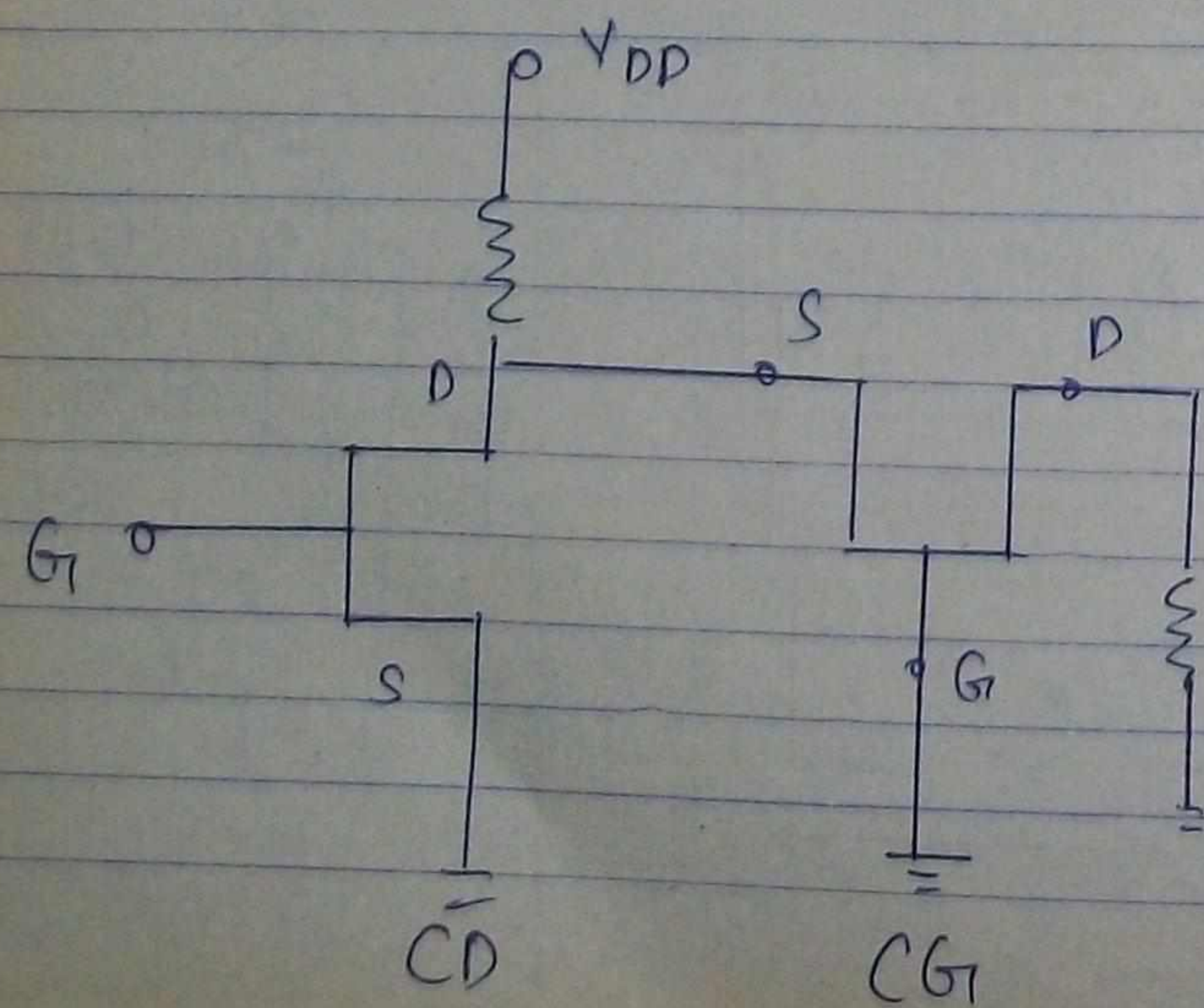
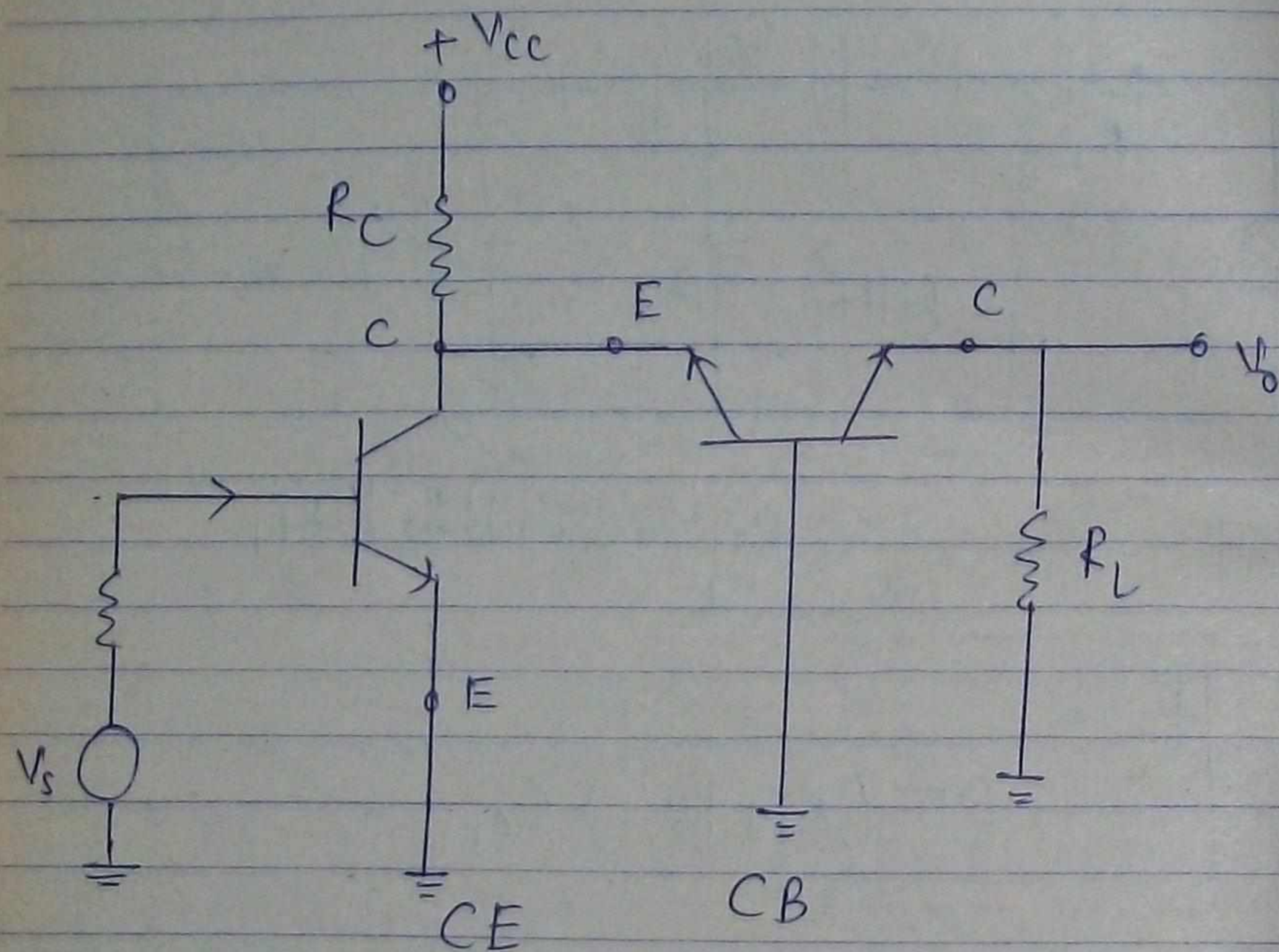
$$f_{L_i} = \frac{1}{2\pi (R_S + R_{i_1}') C_{C1}}$$

$$f_{L_o} = \frac{1}{2\pi (R_C' + R_L) C_{C2}}$$

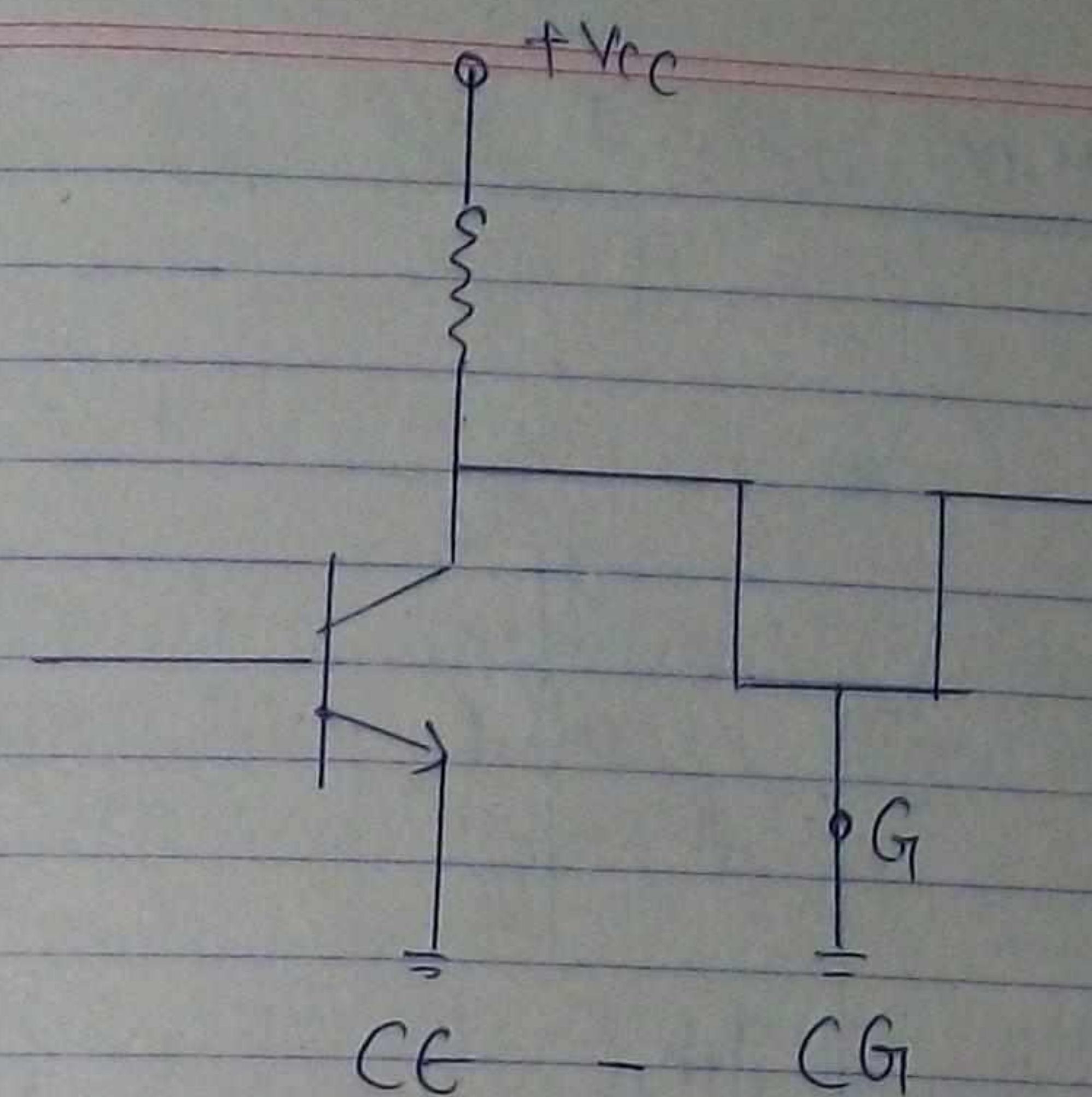


AA

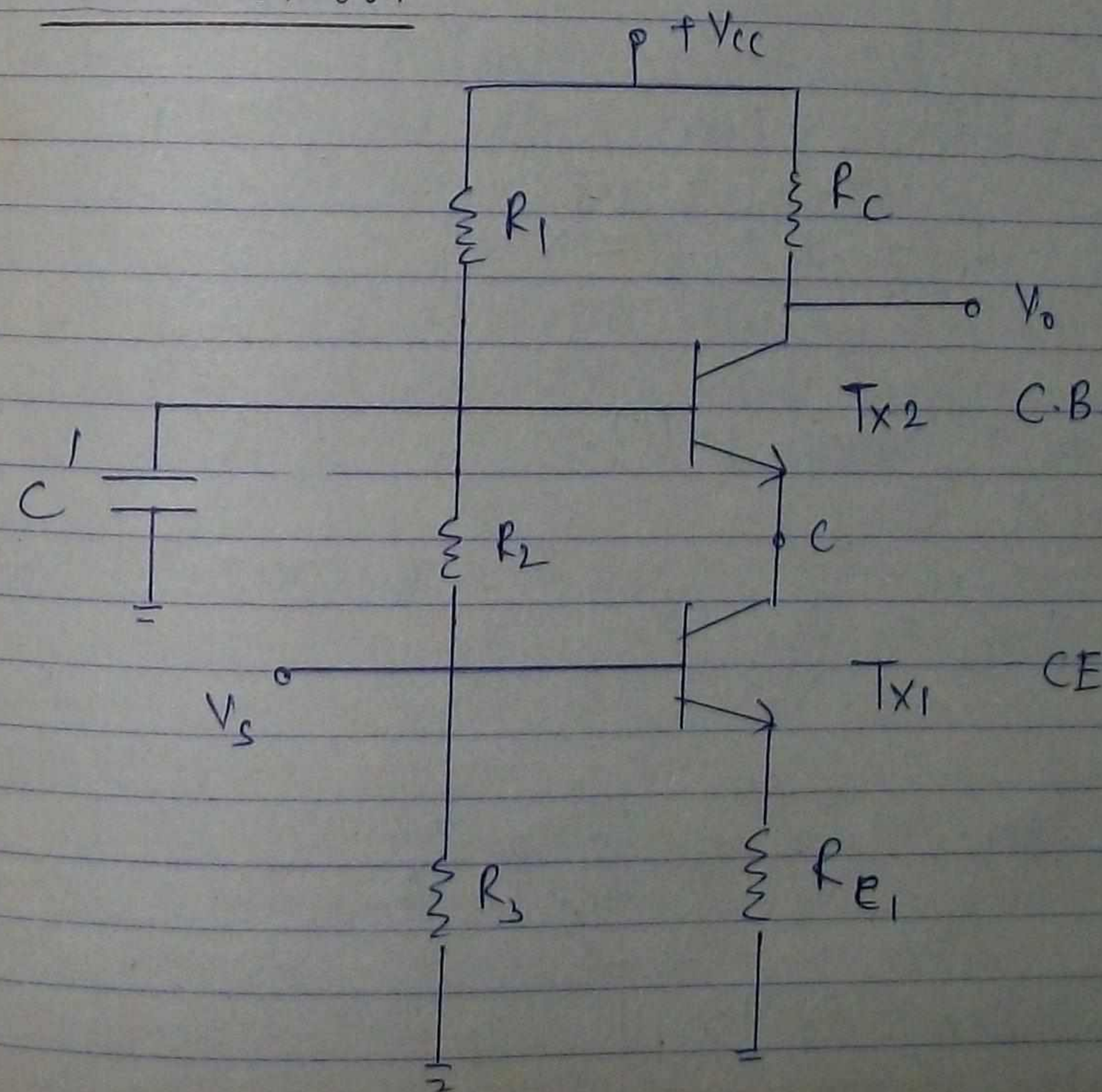
⇒ CASCODE AMPLIFIER - (CE-CB)



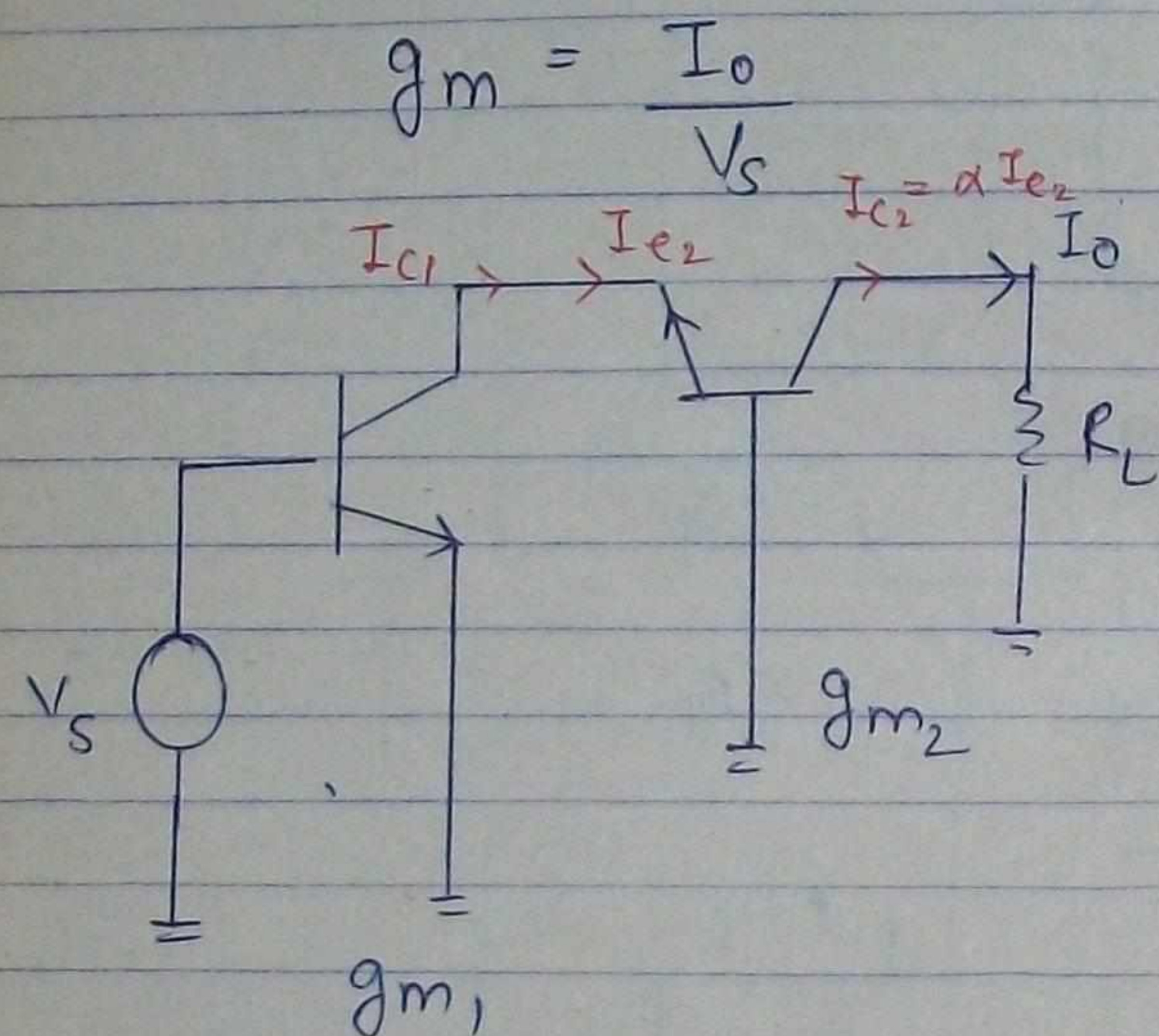
(CS-CB)
also.



BASIC CIRCUIT -



⇒ TRANSCONDUCTANCE-



$$I_O = I_{C2} = \alpha I_{E2} = \alpha I_{C1}$$

$$I_O = \alpha I_{C1} \quad - (1)$$

$$g_{m1} = \frac{I_{C1}}{V_s}$$

$$g_m = \frac{I_O}{V_s} = \frac{I_O}{I_{C1}} \times \frac{I_{C1}}{V_s}$$

$$g_m = \alpha g_{m1}$$

$$g_m = \frac{\beta}{1+\beta} g_{m1}$$

$$\beta \gg 1$$

$$g_m = g_{m1}$$

- It is a specially designed multistage amplifier where common emitter is followed by CB.

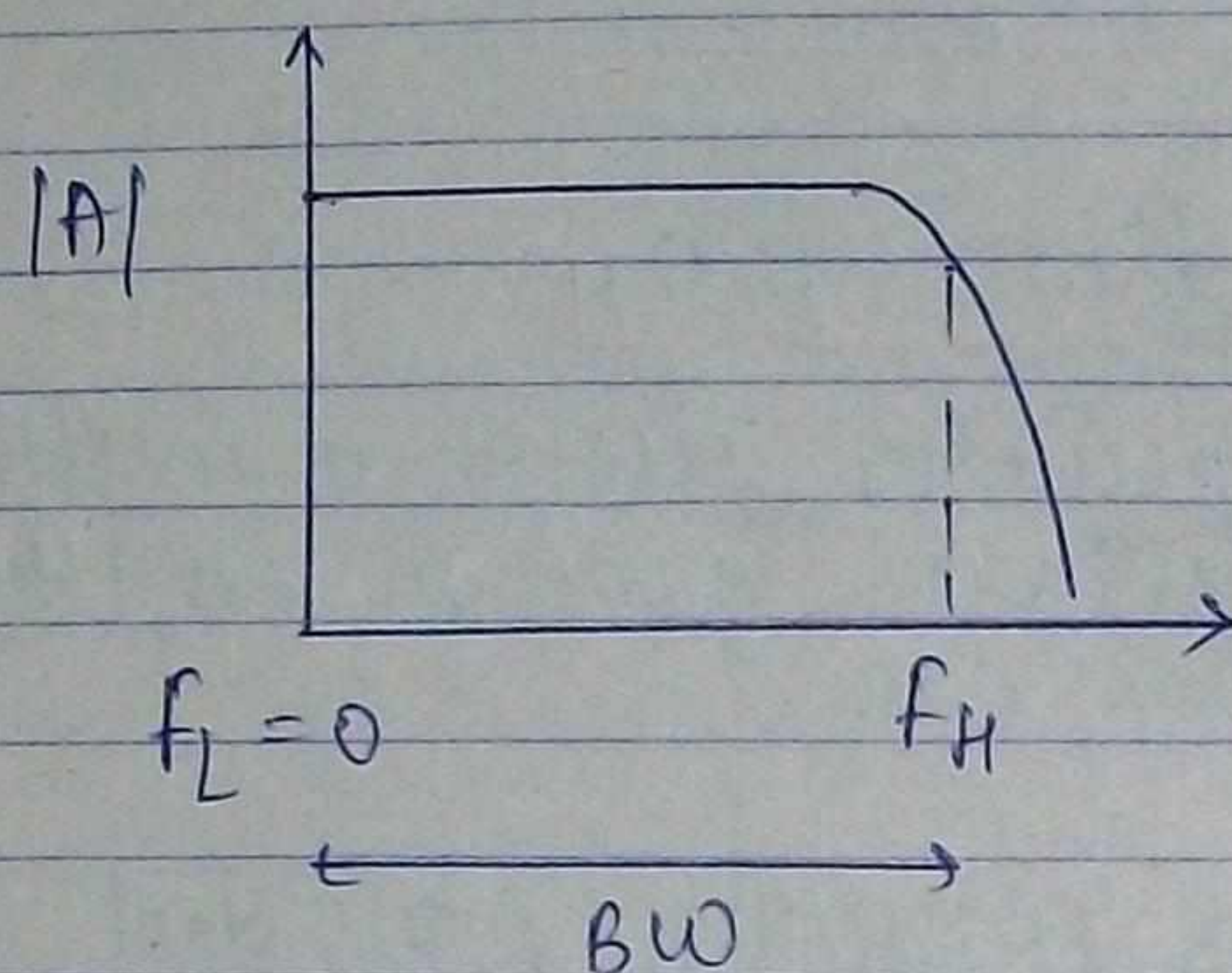
- The type of coupling provided is direct coupled.

- It is suitable to amplify AC and DC signal but major application is as a high freq amplifier.

- The R_i of amplifier is equal to R_i of CE and O/P resistance is decided by O/P resistance of CB.

⇒ DR/COUPLED

⇒ DIRECT COUPLED AMPLIFIER -

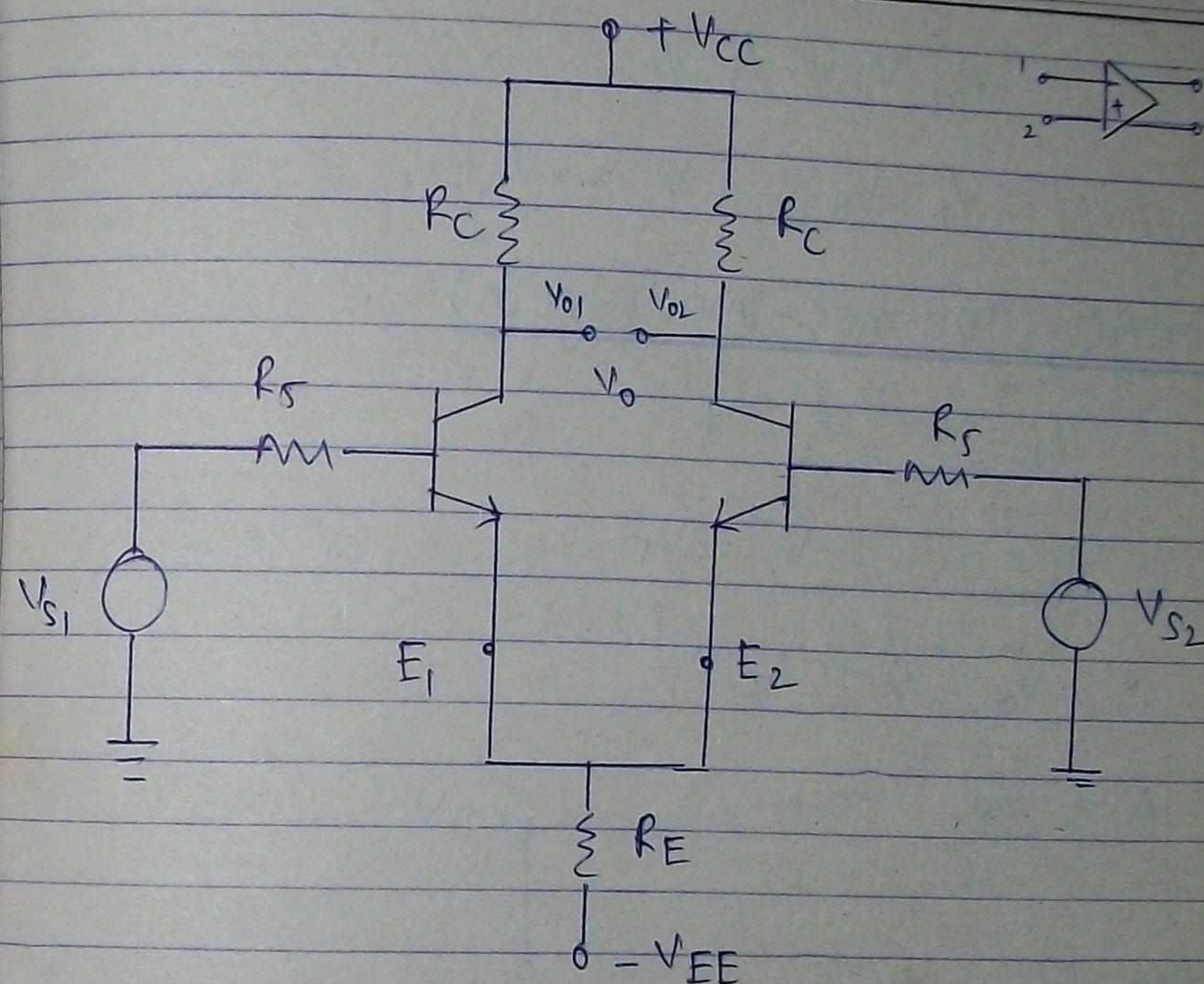


★

$$BW = f_H - f_L = f_H$$

- Suitable to amplify DC signal along with a wide band of AC signal
- widely used as instrumentation amp.
- In a DC amplifier there is no proper DC isolation in between the stages and therefore stability is less.
- Any DC amplifier suffers from Drift problem
($T \uparrow$ $B \uparrow$)
- Popularly used DC amplifier is
EMITTER COUPLED DIFFERENTIAL AMPLIFIER

⇒ EMITTER COUPLED DIFFERENTIAL AMPLIFIER -



MODE OF OPERATION -

1. Dual I/P balanced output.
2. Dual I/P unbalanced O/P.
3. Single I/P balanced O/P.
4. Single I/P unbalanced O/P.

⇒ DERIVATION OF A_c , A_d & CMRR -

$$V_o = A_d V_d + A_c V_c$$

$$V_d = V_1 - V_2 \quad V_c = \frac{V_1 + V_2}{2}$$

for A_c (common-mode gain) -

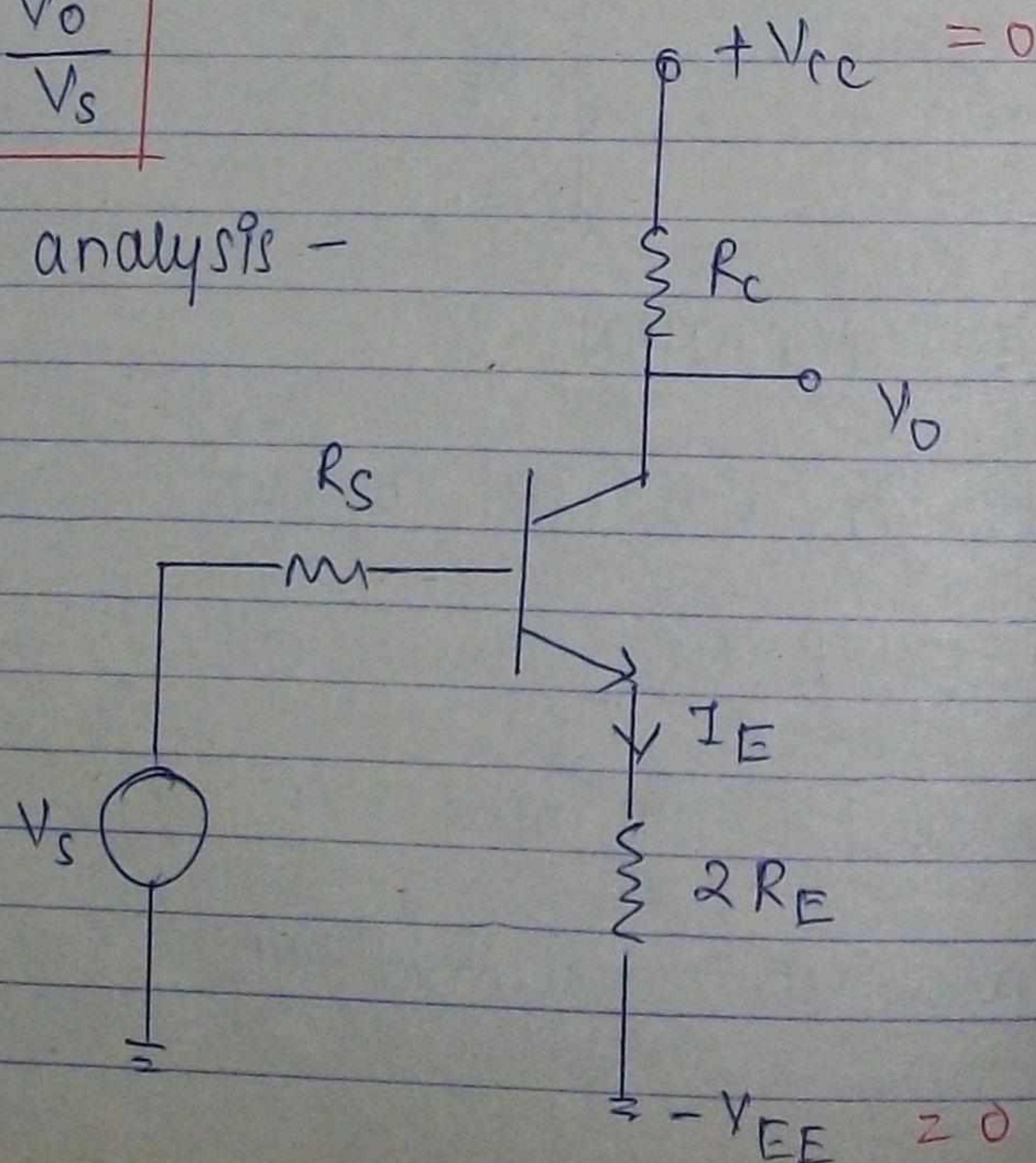
$$V_1 = V_s \quad V_2 = V_s$$

$$V_d = 0 \quad V_c = V_s$$

$$\therefore V_o = 0 + A_c V_s$$

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

for AC analysis -



CE with emitter resistance $2R_E$

$$A_I = -h_{fe}$$

$$R_i^o = h_{fe} + (1 + h_{fe}) 2R_E$$

$$A_{V_s} = \frac{V_o}{V_s} = \frac{A_I R_i^o}{R_i^o + R_s} = \frac{A_I R_L}{R_i^o + R_s}$$

$$A_c = \frac{V_o}{V_s} = \frac{-h_{fe} R_c}{R_s + h_{ie} + (1 + h_{fe}) 2R_E}$$

$$R_E \uparrow \quad A_c \downarrow \quad R_E = \infty \quad A_c = 0$$

$$A_c \propto \frac{1}{R_E}$$

for A_d (Differential gain) -

$$V_1 = V_s/2 \quad V_2 = -V_s/2$$

$$V_c = 0 \quad V_d = V_s$$

$$V_o = A_d V_s + 0$$

$$A_d = V_o/V_s$$

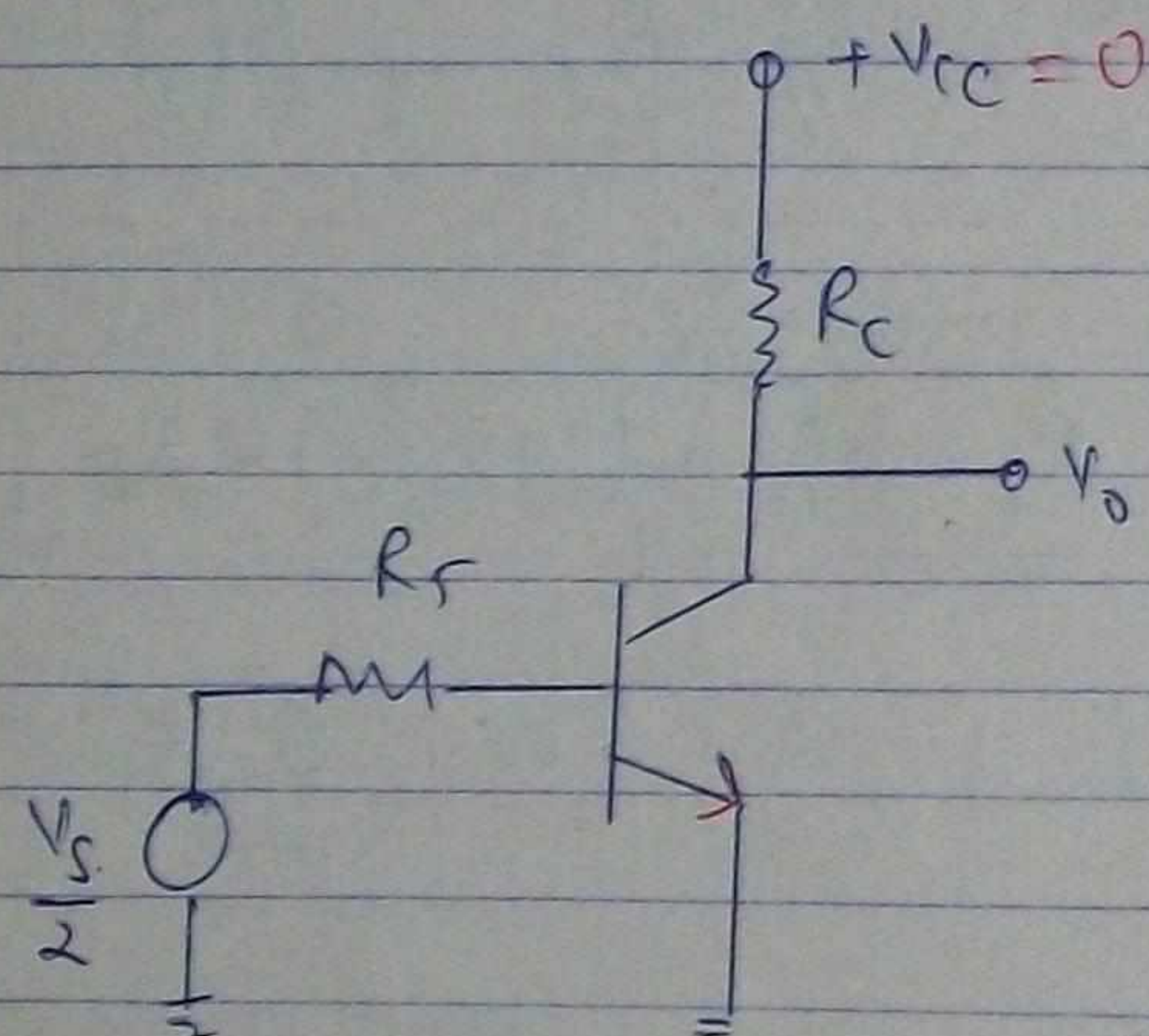
AC analysis -

Current through R_E

$$I_e = I_{e1} - I_{e2} = 0$$

∴ drop across R_E is zero

∴ we can remove R_E .



$$A_I = -h_{fe}, \quad R_i = h_{ie}$$

$$A_{V_s} = \frac{A_I R_L}{R_s + R_i}$$

$$A_{V_s} = \frac{V_o}{V_s/2} = -\frac{h_{fe} R_c}{R_s + h_{ie}}$$

$$A_d = \frac{V_o}{V_s} = -\frac{1}{2} \frac{h_{fe} R_c}{R_s + h_{ie}}$$

for CMRR -

$$CMRR = \frac{|A_d|}{|A_c|}$$

$$CMRR = \frac{\frac{1}{2} \frac{h_{fe} R_c}{R_s + h_{ie}}}{\frac{h_{fe} R_c}{R_s + h_{ie} + (1 + h_{fe}) 2 R_E}}$$

$$CMRR = \frac{R_s + h_{ie} + (1 + h_{fe}) 2 R_E}{2 (R_s + h_{ie})}$$

$$R_E \uparrow \quad A_c \downarrow \quad CMRR \uparrow$$

for $(1 + h_{fe}) 2 R_E \gg R_s + h_{ie}$

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

- Effect of increasing R_E -

- ① R_i increases
- ② $CMRR \uparrow$
- ③ $A_c \downarrow$
- ④ Gain $\downarrow$
- ⑤ $\uparrow$ -ve f.B across R_E