Analog Circuit DESIGN

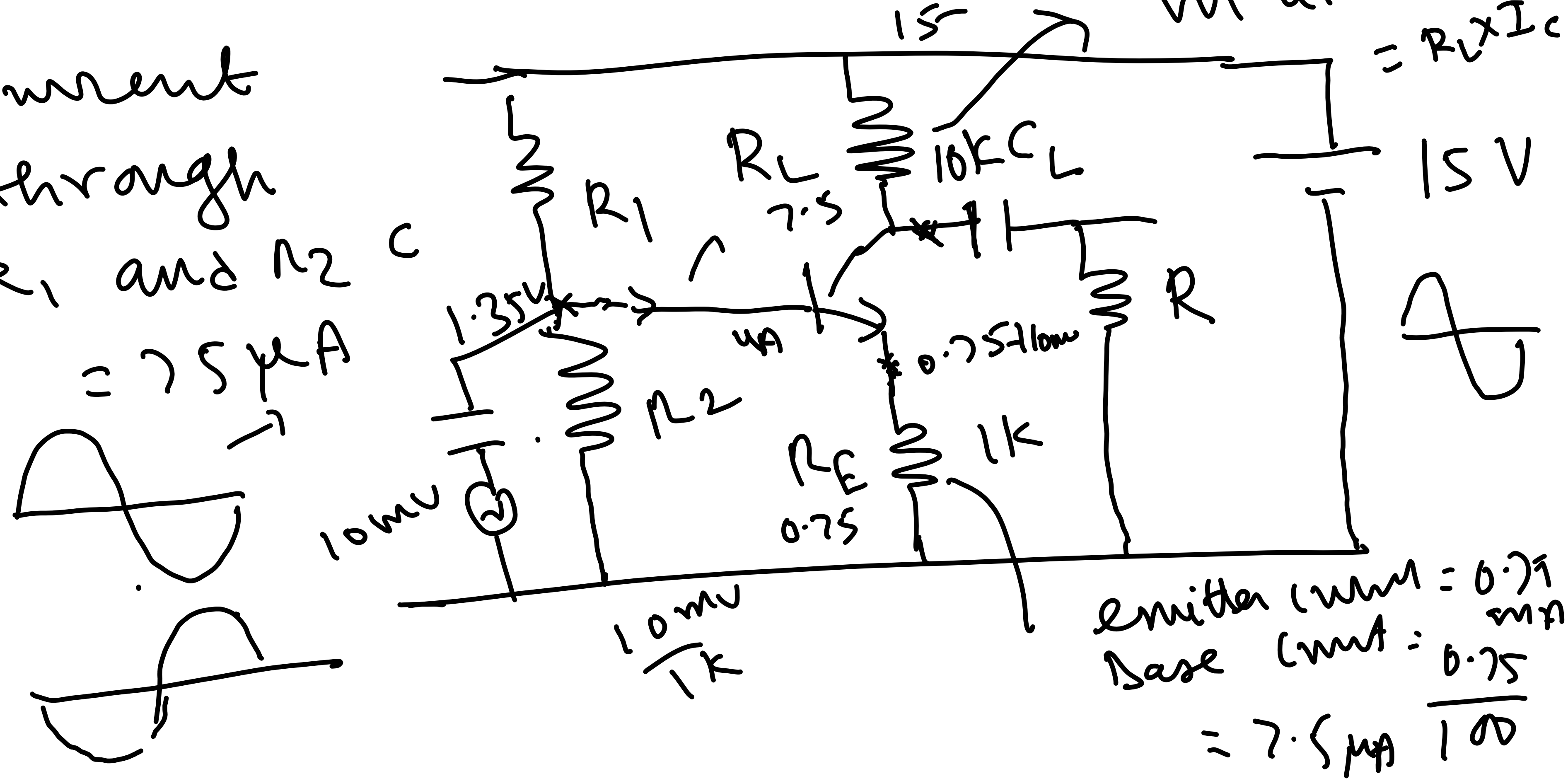
1. How to design a analog circuit?
2. Error budgeting
3. Techniques involved in analog circuits.

1. Designing with operational amplifiers and analog integrated circuits.
   Franco S.
2. Paul Horowitz and Winfield Hill
   Art of Electronics

Transistor amplifier

current through R_1 and R_2 C $= 75\,\mu A$

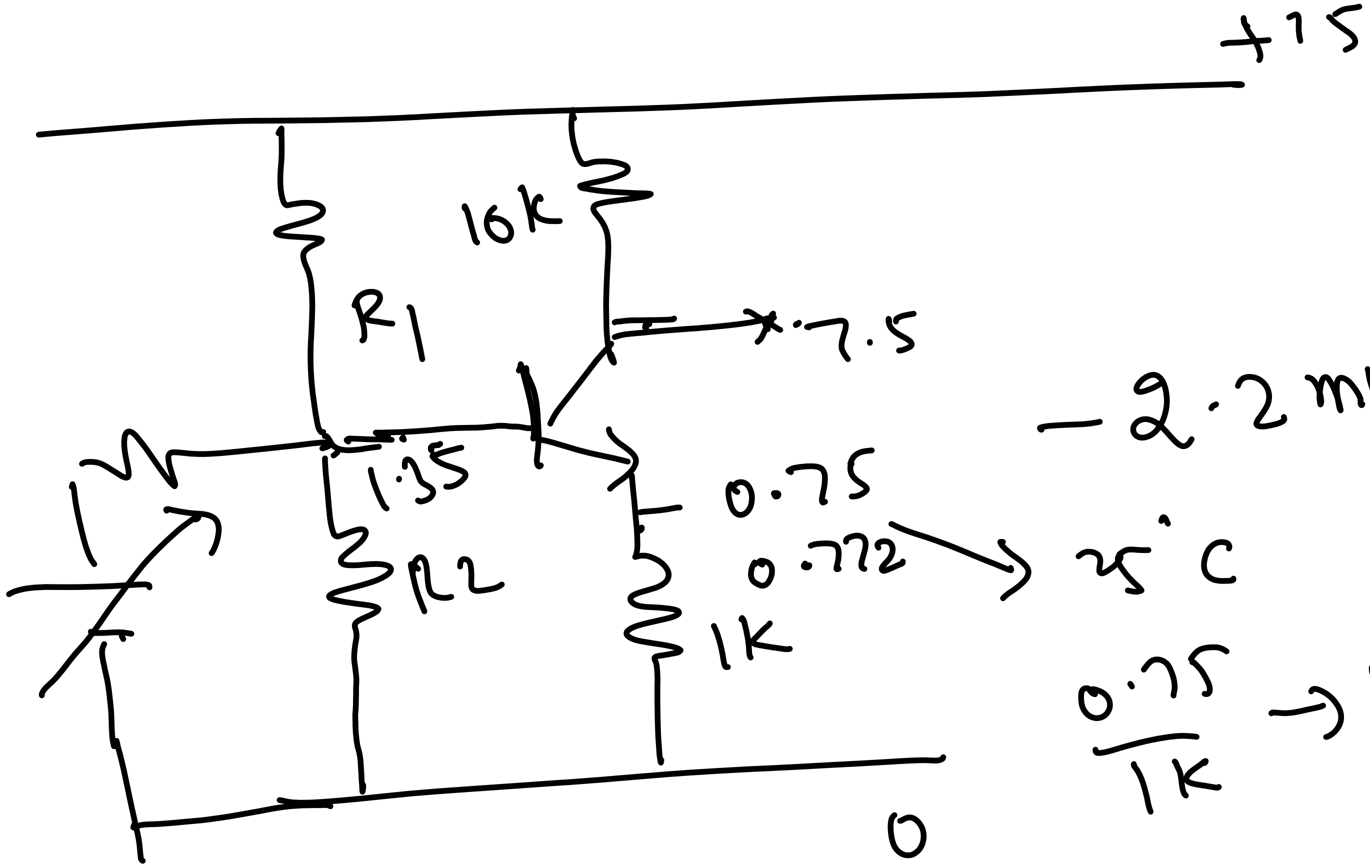
vol arc R_L $= R_L \times I_c$



emitter current $= 0.75$ mA

Base current $= \frac{0.75}{100}$

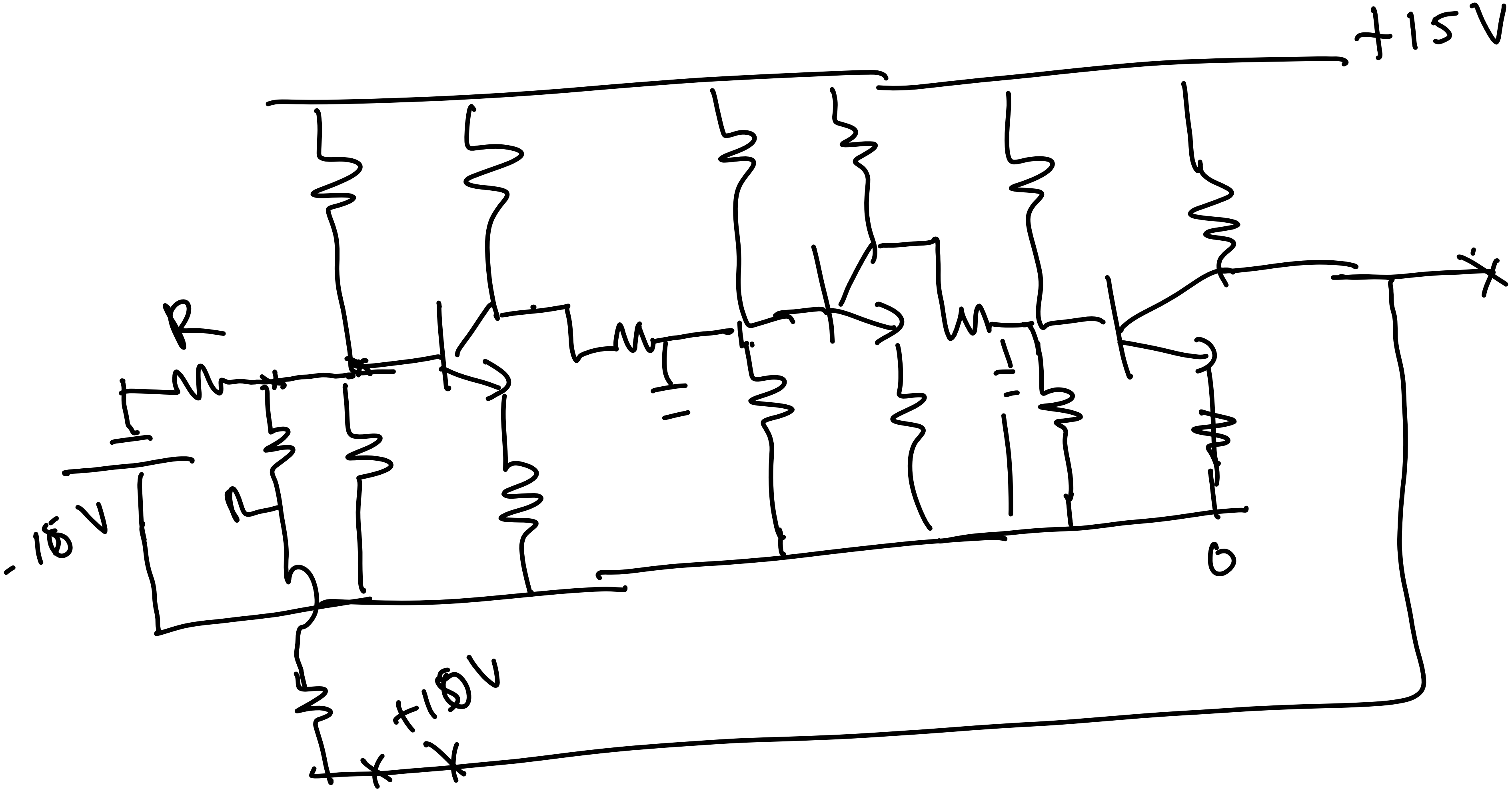
$= 7.5\,\mu A$



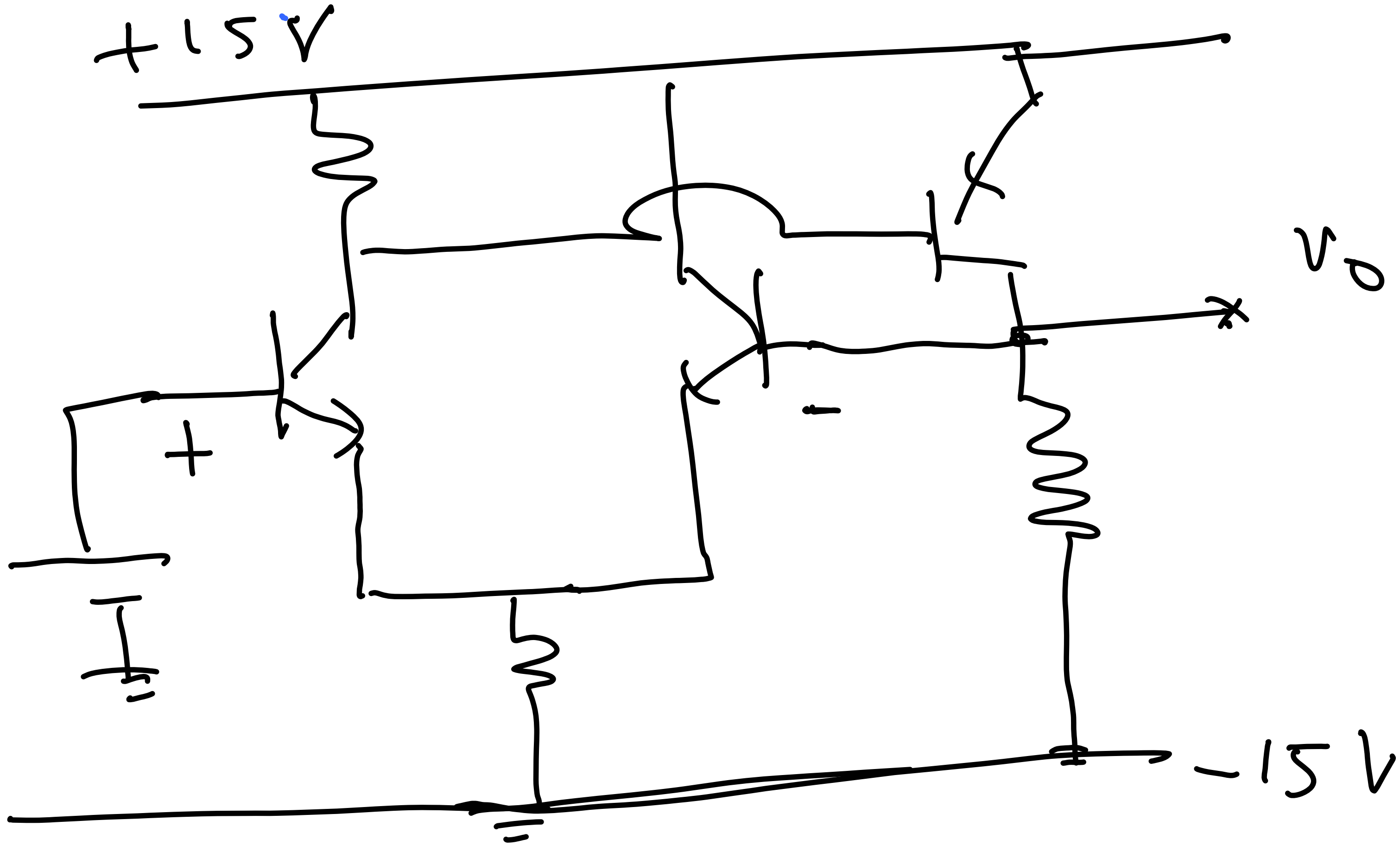
$-2.2\ mV/^{\circ}C$

$\rightarrow 25^{\circ}C$

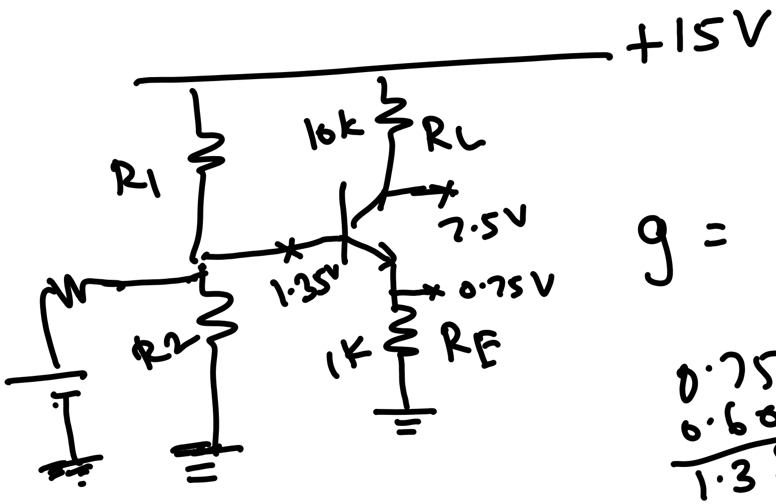
$$\frac{0.75}{1k} \rightarrow \frac{0.772}{1k}$$



+15V
R
-15V
+15V
0

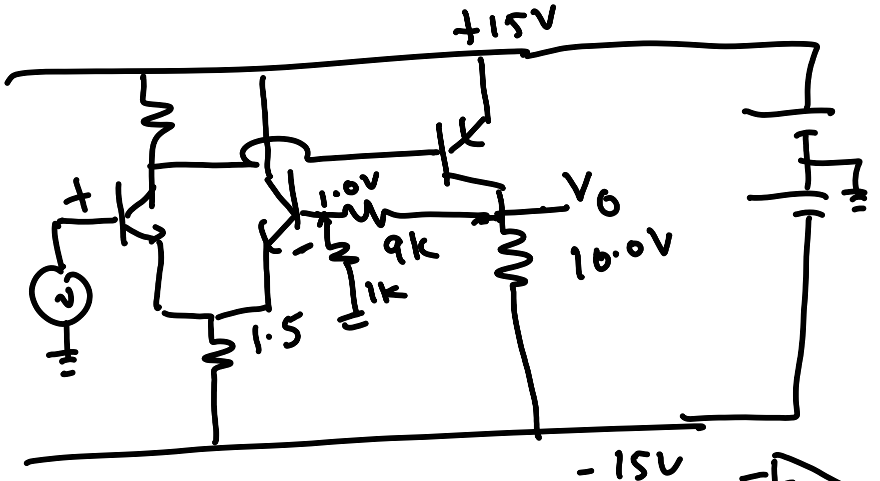


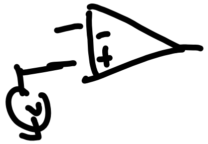
+15V
+
-
v_o
-15V

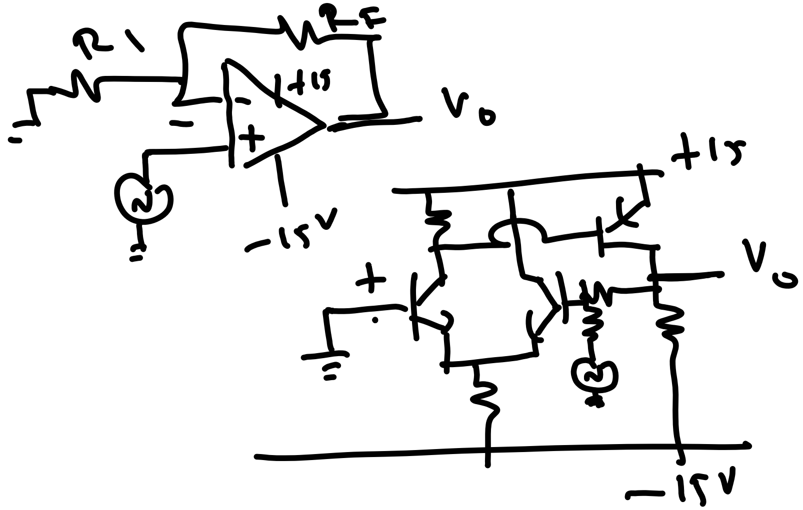


$$g = \frac{R_L}{R_E}$$

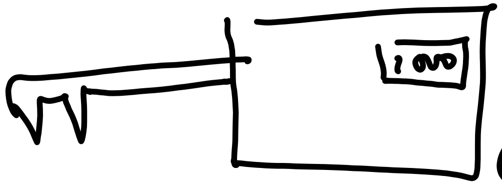
$$\begin{array}{r} 0.75 \\ 0.60 \\ \hline 1.35 \end{array}$$


+15V
+
1.0V
V0
9k
-
1k
1.5
10.0V
-15V


-
+



R1
RF
+15
-15V
Vo
+15
Vo
+
-15V



1. error budgeting
2. circuit design
3. Techniques

+15V

$V = 14.4V$

$I = 0.1\,mA$

$R_1 = \frac{14.4 \times 10^3}{0.1}$

$= 144K$

$15 - 0.6$

$= 14.4V$

$R_1 \rightarrow 100K$

$\rightarrow 1.5K$

R

.6

15V

15V

0.1 mA

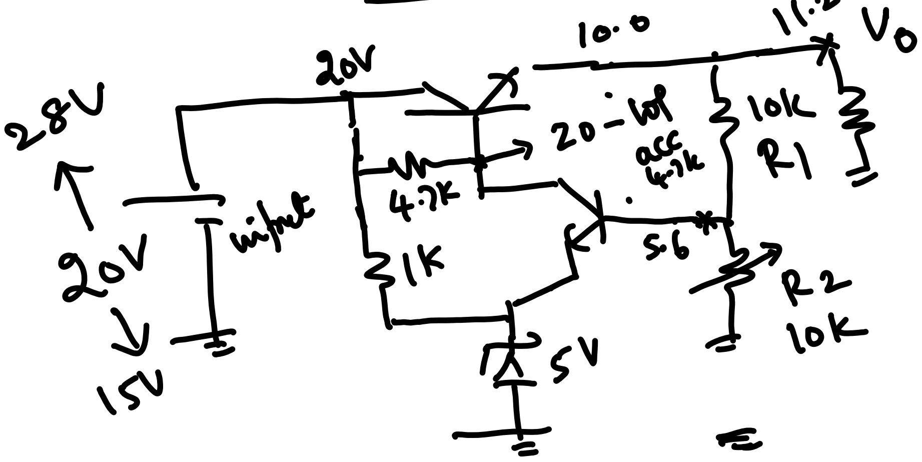
3019 $I_b = ?$

$\beta = 100$

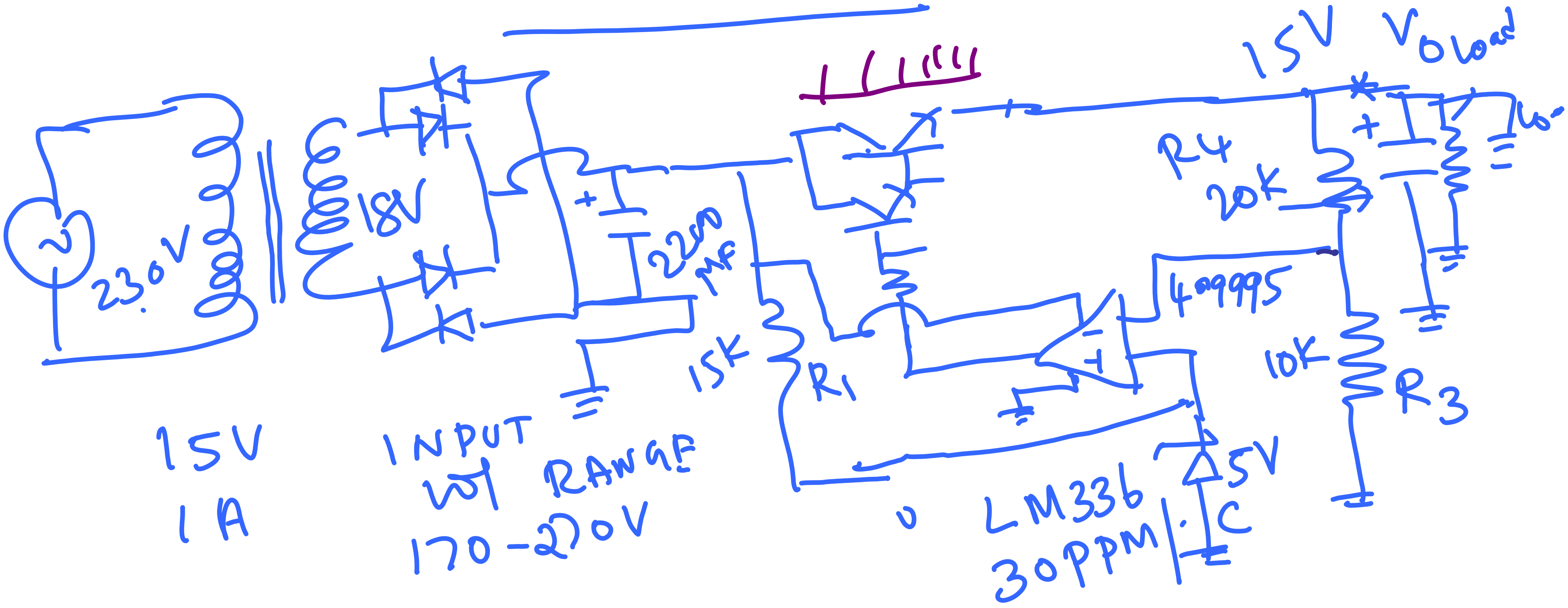
$I_c = 10\,mA$

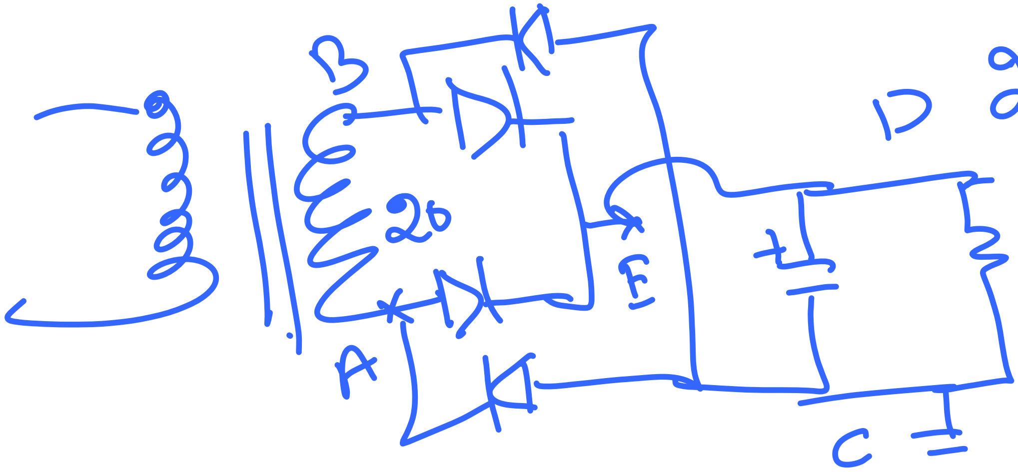
$I_b = \frac{10 \times 10^{-3}}{100}$

Voltage regulator



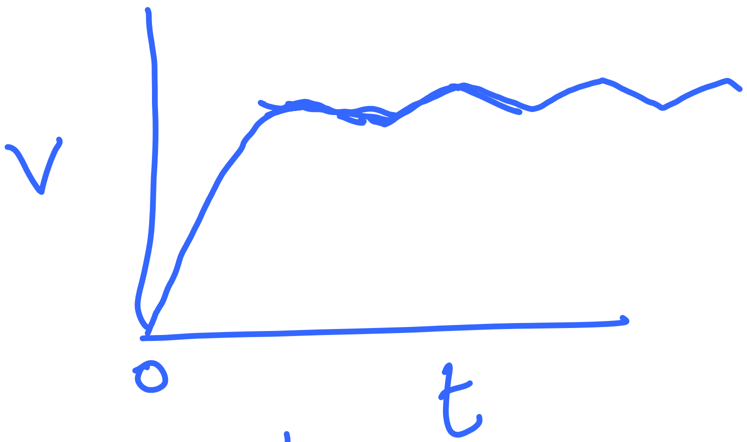
APPLICATIONS OF OP AMP

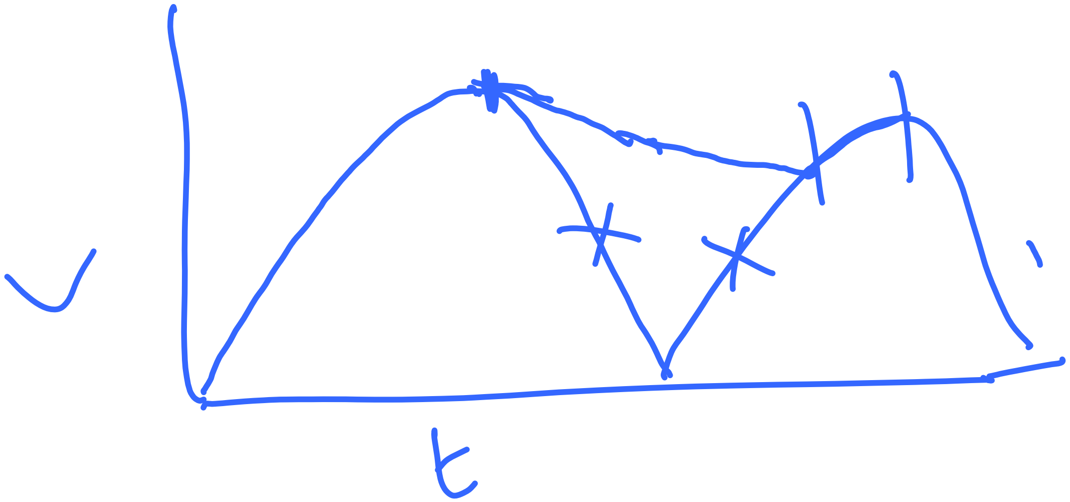




D 20x1.414
28V

wrt acc
C and D





At 230V AC input

the sec vol = $\dfrac{13 \times 23\cancel{0}}{17\cancel{0}}$

129
6

$$= \begin{array}{r} 39 \\ 26 \\ \hline 299 \\ \hline 17 \end{array} \approx 18V$$

Pri : 0-230V

Sec = 0-18V

Case 1

Continuous

1 A

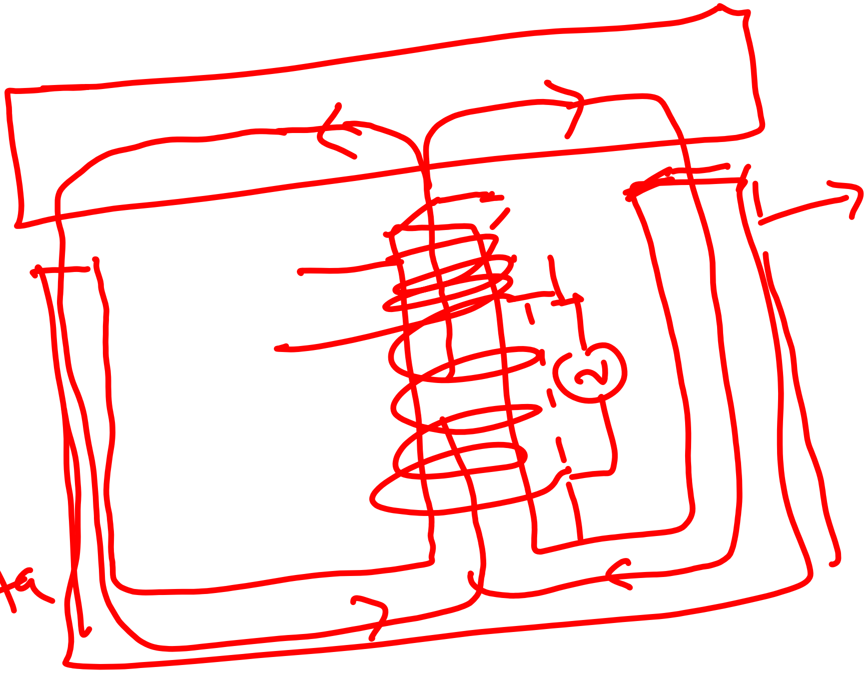
$I^2 R = 1^2 \times 1 = 1W$

Case 2

3 A for $\frac{1}{3}t$

$$\frac{3^2 \times R}{3} = \frac{9 \times 1}{3}$$

$= 3W$



— cold rolled grain oriente

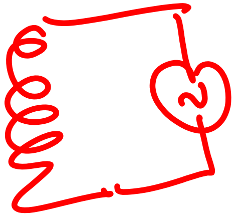
1) CRGO
2) CRNGO
3) Dyanamo

Saturation level = 1 Tesla

How many no. of turns?

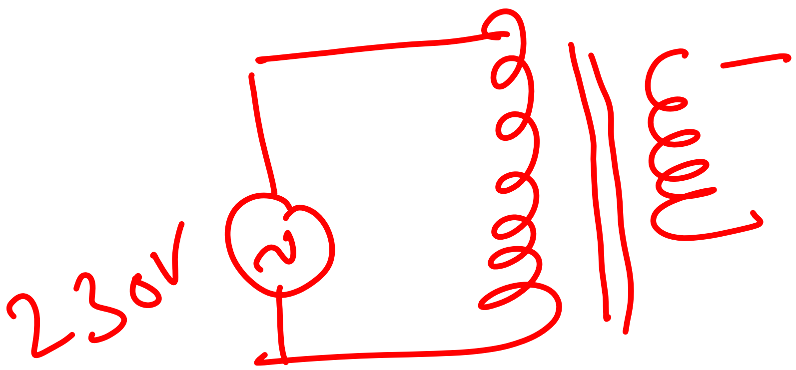
B = flux density = 1 Tesla

L



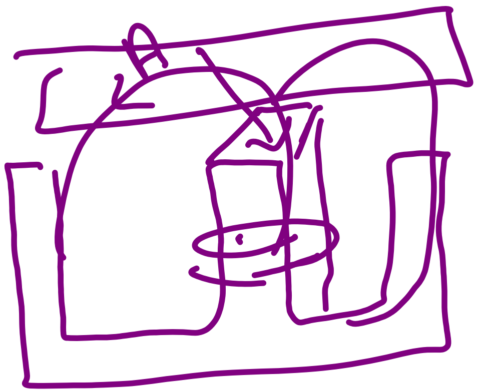
f

$I = \frac{V}{jL\omega}$



Basic Concept

Fix $B = 1$ Tesla



$$BAN\omega = V$$

$$1 \times A \times N \times 2\pi \times 50 = 230$$

$$N = \frac{230}{A \times 2\pi \times 50 \times 1}$$

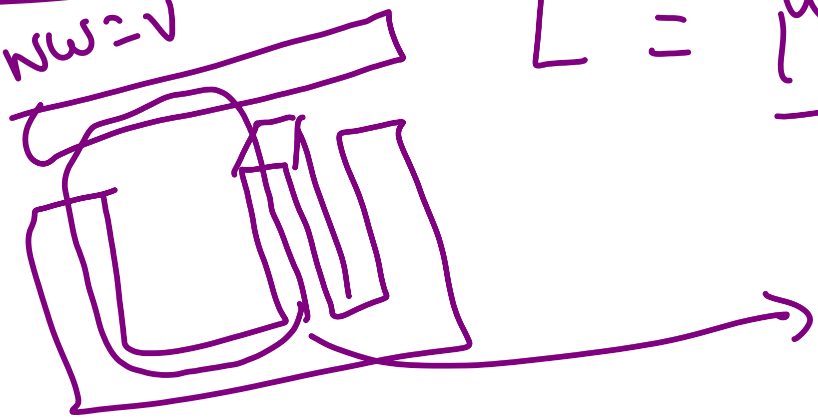
Thickness of the wire

$4\pi \times 10^{-7}$ H/m

3000

$$L = \frac{\mu_0 \mu_r n^2 A}{l}$$

BANω = V



Primary current $= \frac{V}{L\omega}$

① Mag. current

$$\frac{V}{jL\omega} = I$$

② Reflected sec current

For 1 : 1

Reflected current = Secondary current = 1

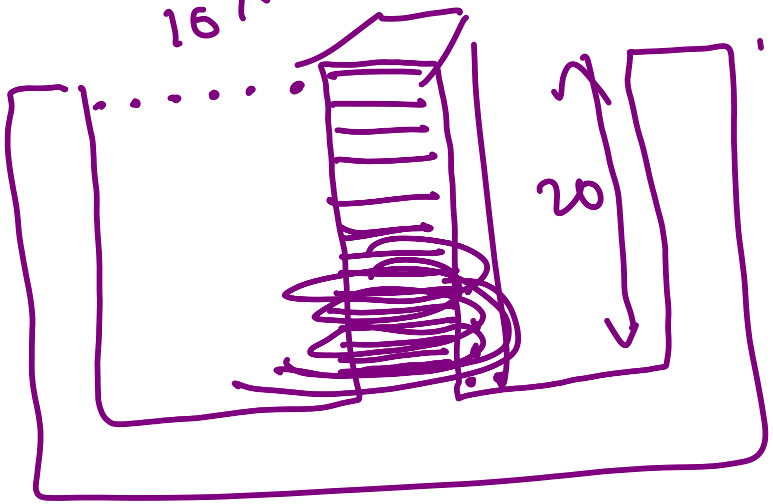
$3A / 1\,mm^2$ $\quad \pi r^2 = 1\,mm^2$ $\quad A = 0.3\,mm^2$

BANW = 230

$3\,A/mm^2$

$16 \times 20 = 200$

of $\phi = 1\,mm$



Heat Sink Design

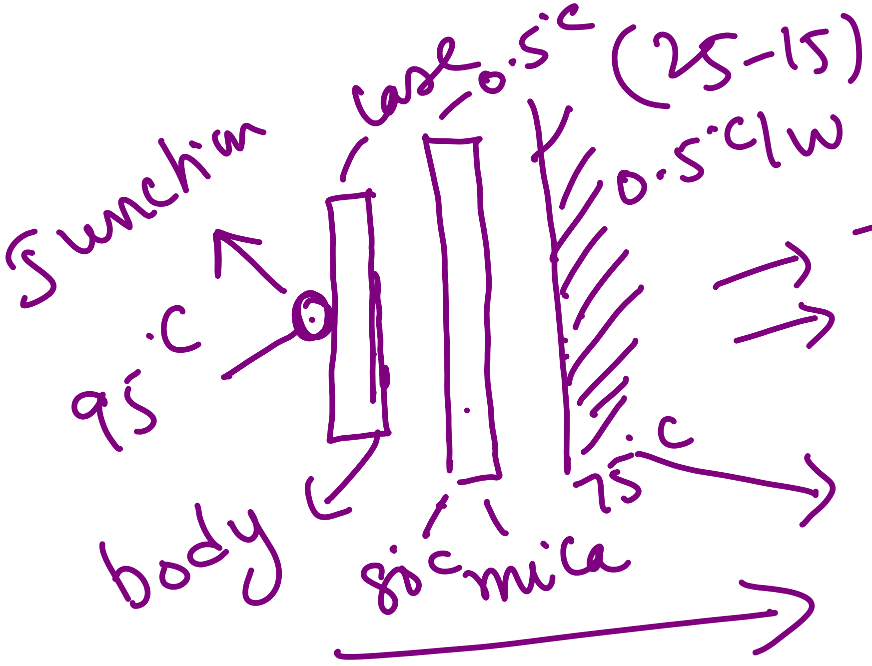
1. Fix the Junction temp
   Let it be 100°C

2. What is the max ambient temp?
   ✓ 0 – 70°C — Lab equipments
   – 20°C – 85°C — industrial grade
   – 40°C – 105°C — mil grade

$J_T = 100^\circ C$

$T_{amb} = 70^\circ C$

$P_D = V_{acc} \times I = 10 \times 1 = 10W$

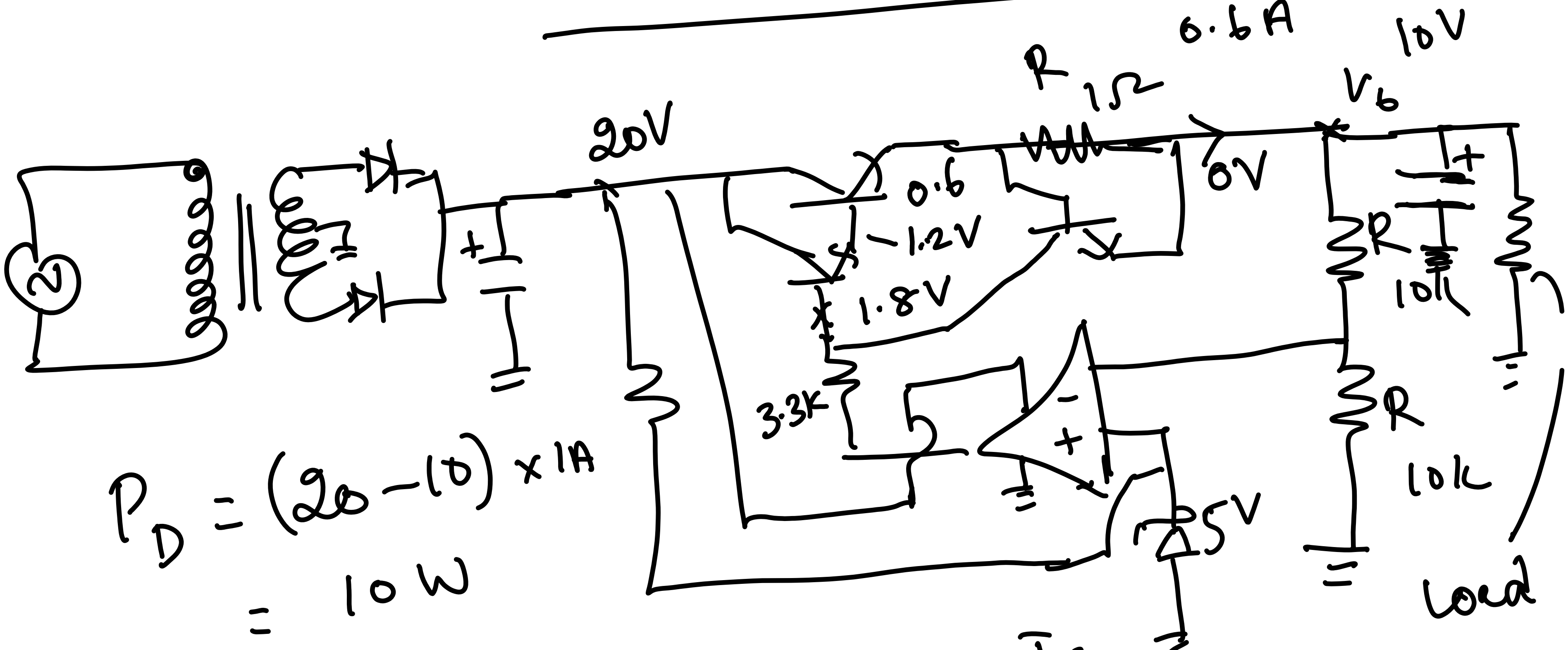


2N3055

Thermal resistence Heat sink from Jun to case $= 1.5^\circ c/w$

mica $= 0.5^\circ c/w$

Short circuit protection

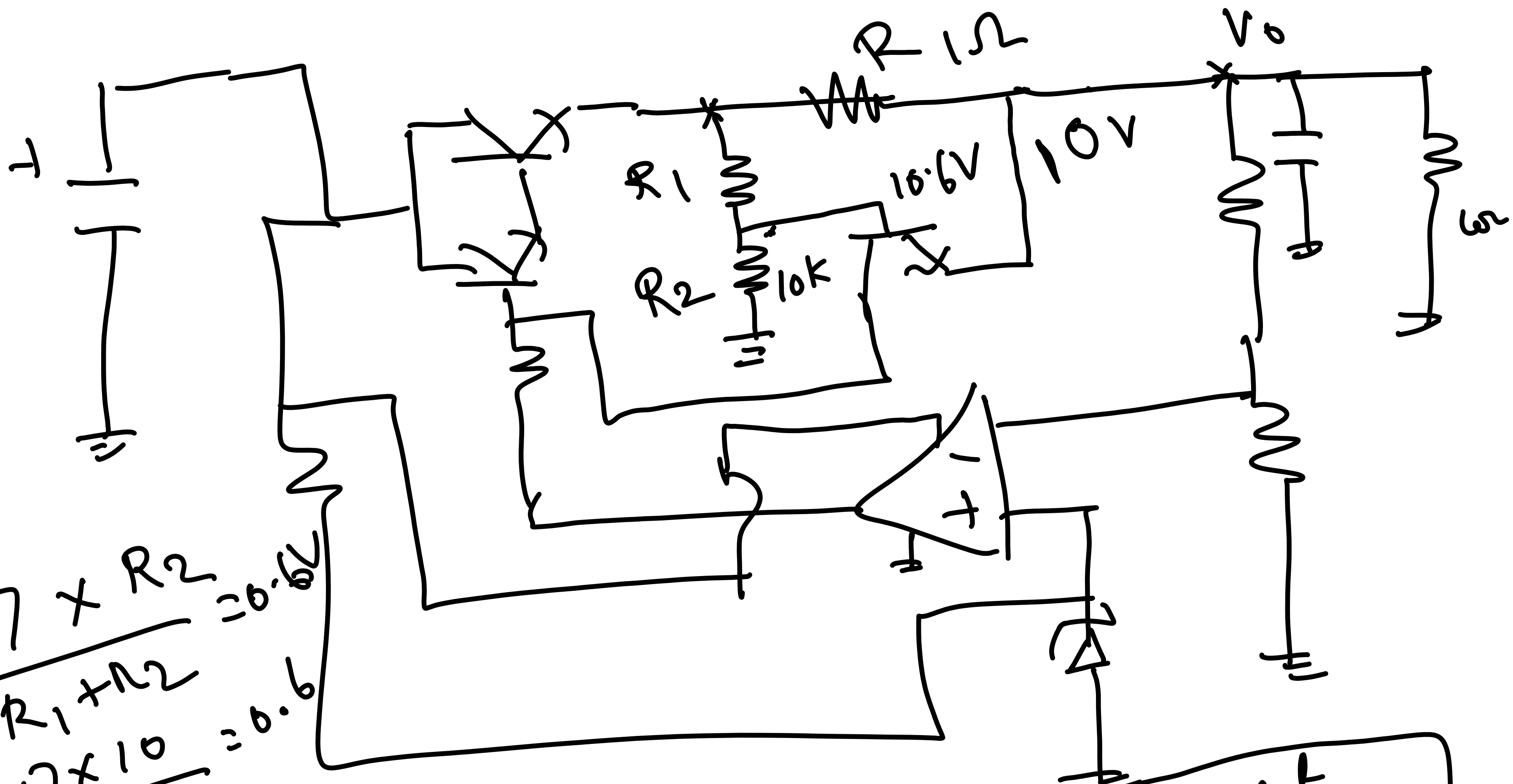


$$P_D = (20-10) \times 1A$$
$$= 10W$$

$$0.6V = R \times I_S$$
$$0.6 = 1\Omega \times 0.6A$$

UNDER NORMAL CON $P_W = (20-10) \times 0.6 = 6W$

Under Short circuit $P_W = 20 \times 0.6 = 12W$



$$\frac{0.7 \times R_2}{R_1+R_2} = 0.6V$$

$$\frac{0.7\times 10}{R_1+10} = 0.6$$

$\underline{7}$

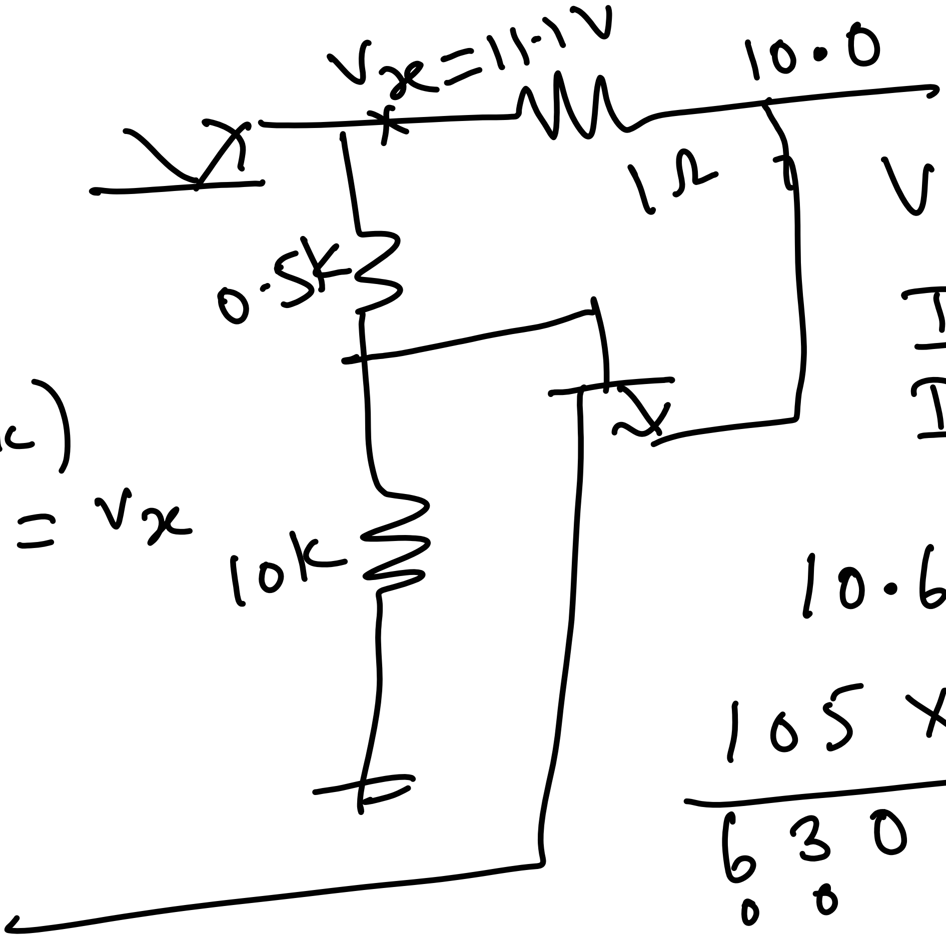
$\boxed{\text{Fold Back current limiting}}$

$$\frac{7}{R_1+10} = 0.6$$

$$7 = 0.6R_1 + 6$$

$$1 = 0.6R_1$$

$$R_1 = \frac{1}{0.6} = \frac{1000}{600} = 0.5k$$



$V_D = 1.1V$

$I = 1.1A$

$I_s = 0.7A$

$$\frac{10.6\,(10.5k)}{10k} = V_x$$

$$\frac{10.6 \times 10.5}{10}$$

10.6×1.05

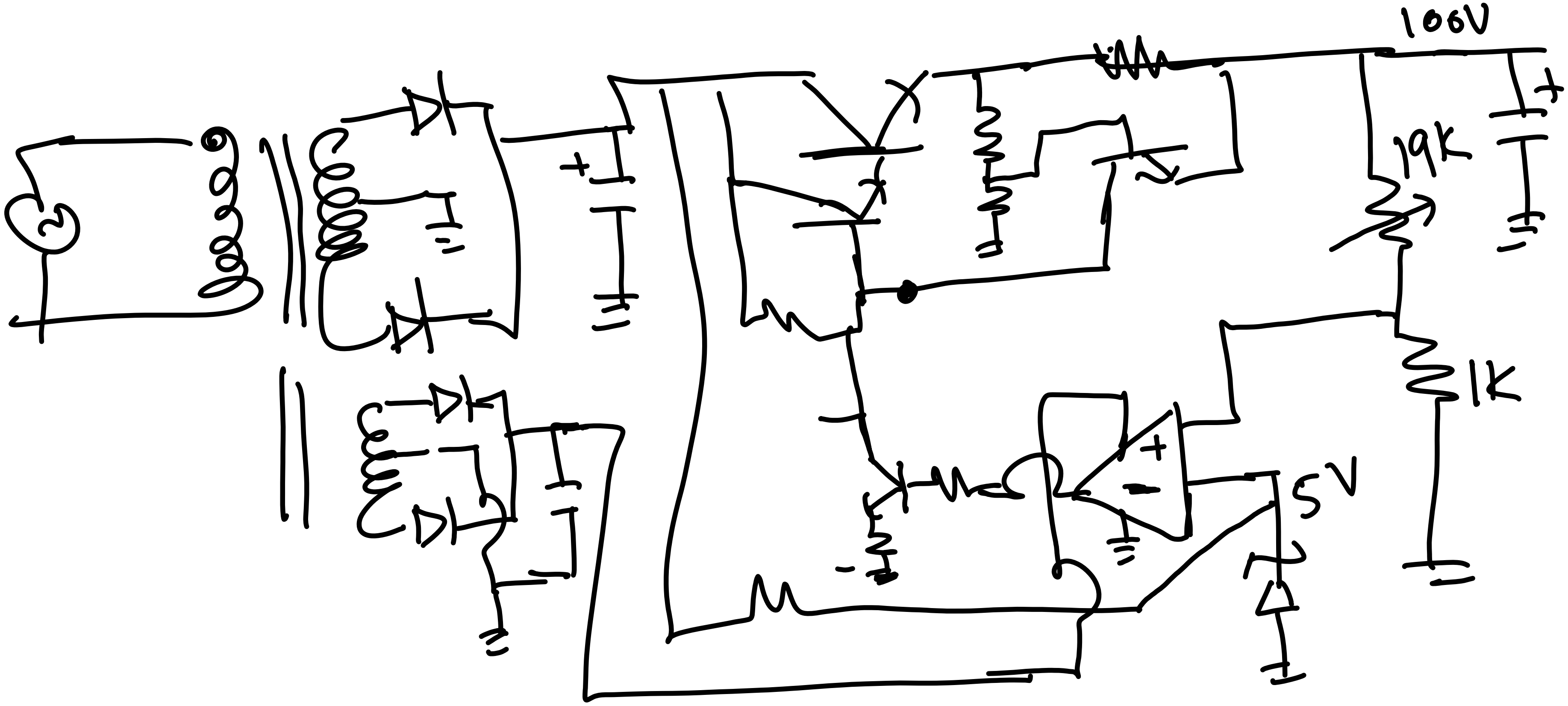
105 × 106 3

630

0 0

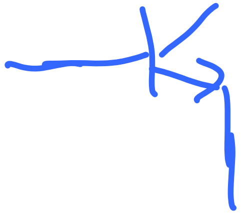
105

11.136


100V
19K
1K
5V
+
-

Temperature indicator

IC Sensor – LM335



$-2.2\ \text{mV}/^\circ C$

at $25^\circ C$ $V_{BE} = 0.6V$

$35^\circ C$ $V_{BE} = 0.578$

$R = ?$

$T_H = 100^\circ C$

$T_C = 0^\circ C$

At $0^\circ C$ vol acc the sensor $= 2.73V$

At $100^{\circ}C$

vol acc the sensor

$= 3.73V \rightarrow 1.00V$

$V_0 = (273 + T_c) \times 10\,mV$

At $0^{\circ}C$ $= 2.73V$ ✓ $\rightarrow 0.00V$

At $1^{\circ}C = 2.74V$

$2^{\circ}C = 2.75V$

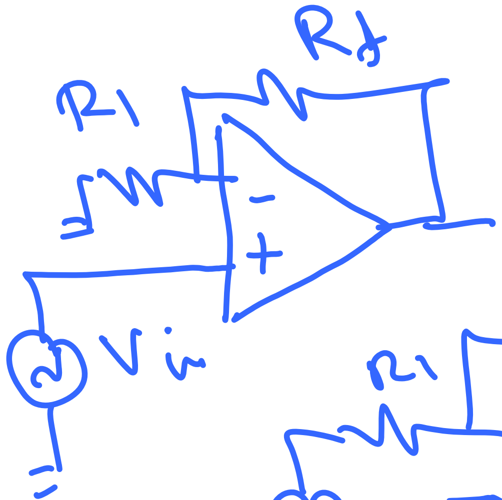
$25^{\circ}C = 2.73 + 0.25$

$2.98V$

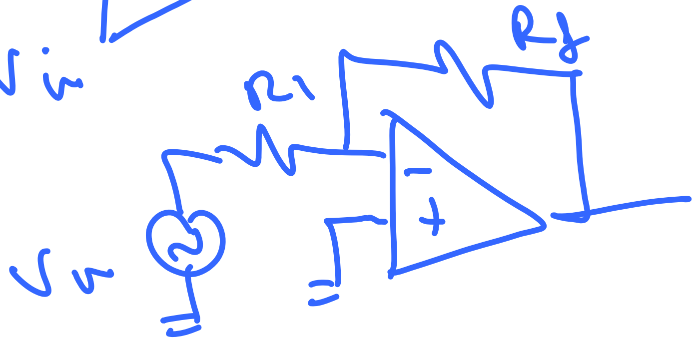
At 0°C → 0 V

at 100°C → 1.006 V

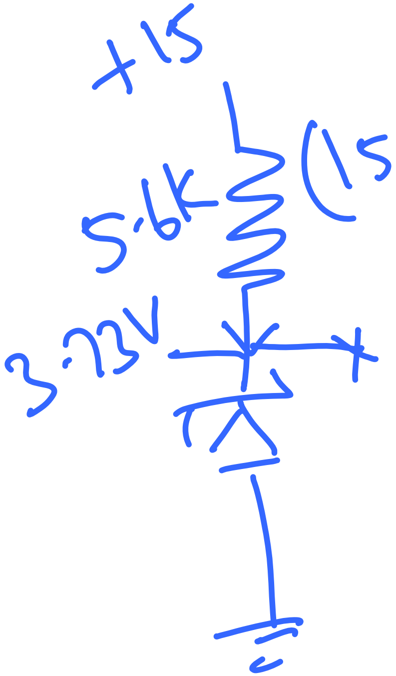
50°C → 0.500 V



$$V_o = V_{in}\left(1+\frac{R_f}{R_1}\right)$$



$$V_o = -V_{in}\frac{R_f}{R_1}$$



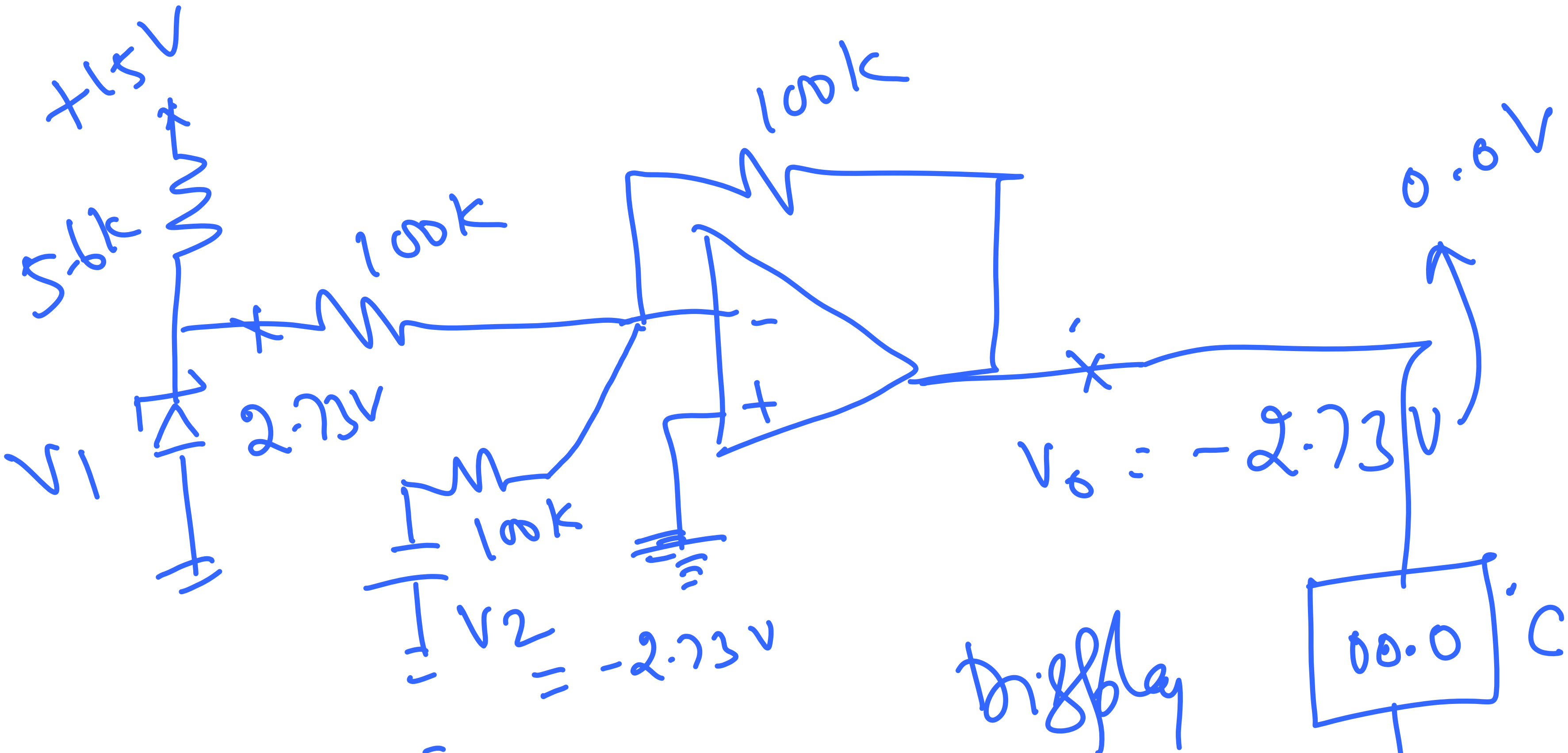
$(15 - 3.73)$

$$\begin{array}{r} = 15.000 \\ - \; 3.730 \\ \hline 11.270 \\ \hline \end{array}$$

$$R = \frac{11.270}{\text{current}} = \frac{11.270}{2 \times 10^{-3}}$$

$$= 5.56\,K$$

$$= 5.6\,K$$



$$-V_1 + V_2 = 0$$

At 100°C

Band gap
LM336

At $100^{\circ}C$

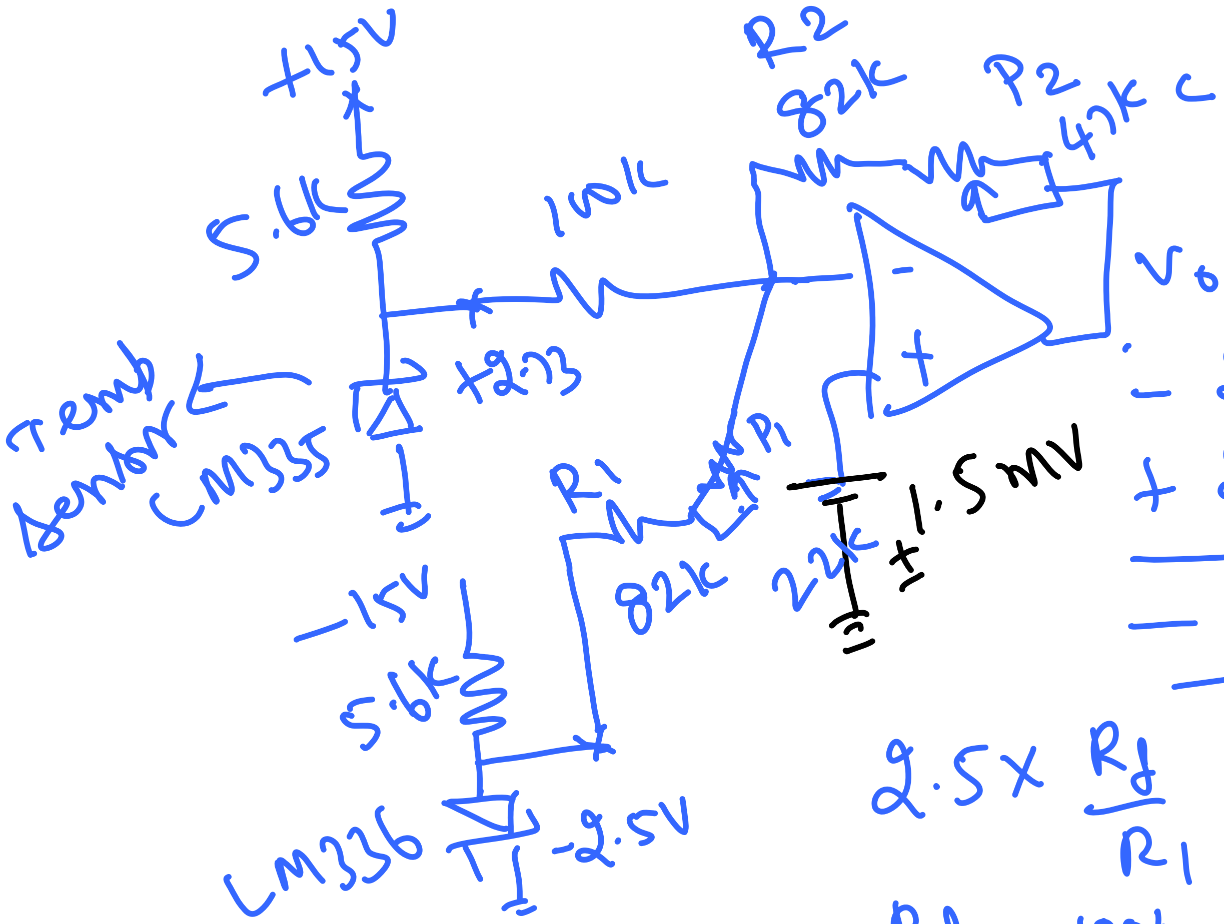
$V_s = 3.73V$

$V_0 = \underbrace{-3.73}_{V_{01}} + \underbrace{2.73}_{V_2}$

$= -1.0 V$

At $0^{\circ}C = 0 V$

$100^{\circ}C = -1.0 V$



3.72×9

$= 3.73$

3.74×9

$= 3.73$

$$\begin{array}{r} -\ 2.73 \\ +\ 2.50 \\ \hline -\ 0.23 \\ \hline \end{array}$$

$$2.5 \times \frac{R_f}{R_1} = 2.73$$

$R_f = 100k$

$$2.5 \times \frac{100k}{R_1} = 2.73$$

$$\frac{2.5 \times 100k}{2.73} = R_1$$

1. Keep the sensor in ice

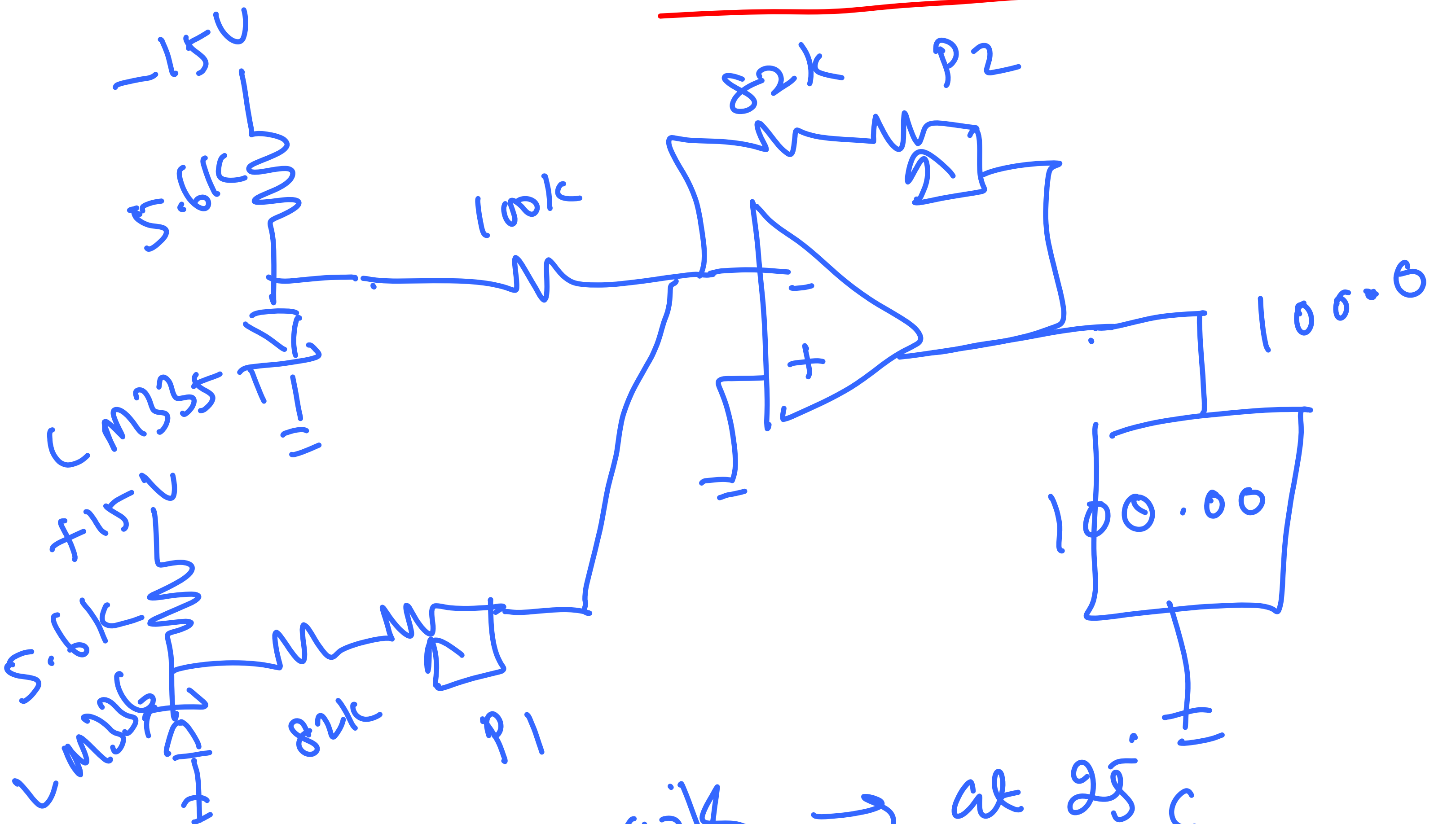
   Adj P_1 to set $V_o = 0$

2. Keep the sensor at $100°C$

   Adj P_2 to get $V_o = 1.00\,V$

3. Repeat step 1 and 2 several times

Error calculations



100k → at 25°C

? → at 50°C

LM336 temp drift ±30ppm/°C max

Working temp range

Ambient temp

range → 100°C

Temp Drift = 30 ppm/°C

Total drift = 30×100 ppm

$$= \frac{30\times100}{10^6}$$

original wt = 2.5

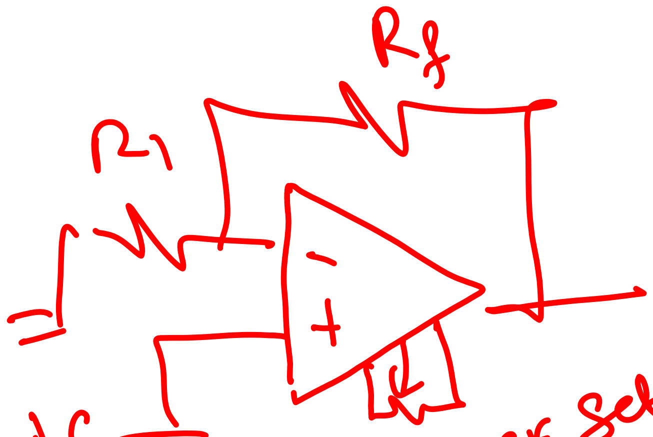
Total change for 2.5 V
for $\Delta T = 100^\circ C$

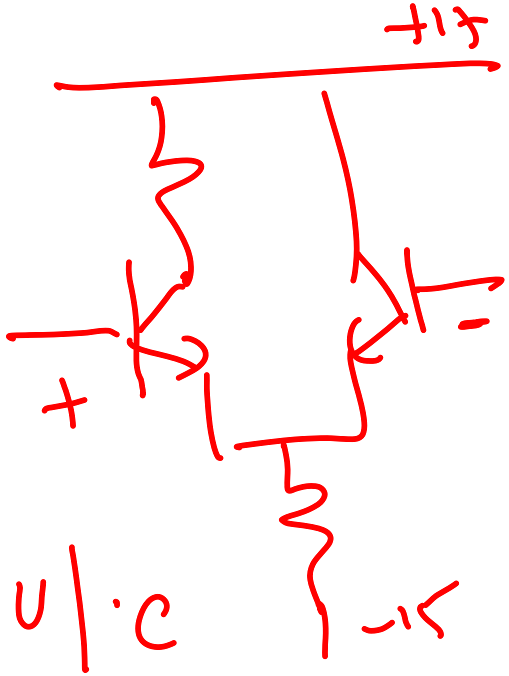
$$= \frac{2.5 \times 30 \times 100}{10^6} \ V$$

$$= \frac{2.5 \times 3 \times 10^3}{10^6}$$

$$= 7.5 \times 10^{-3}$$

$$= 7.5 \ mV$$





$15 \times 10^{-6} \times 100$

$1.5\ mV$

$15\ \mu V/^\circ C$

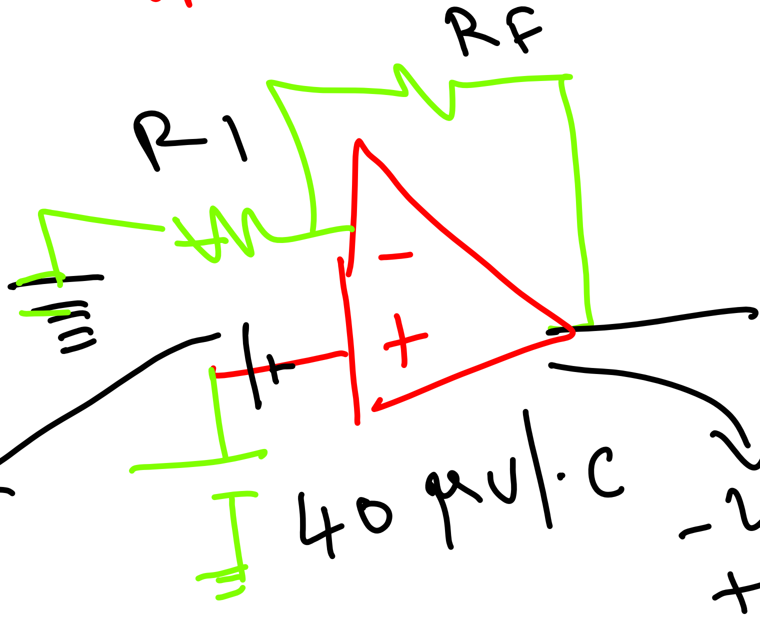
V_{off} off set

V_{off} Drift $= \pm 15\ \mu V/^\circ C$

$15\ \mu V \times 100$

For 741

$1.5\ mV$



$40\ \mu V/^\circ C$

$-20^\circ C$

$+88^\circ C$

$V_0 = 400\ \mu V$

$= 400\ \mu V \pm 1.5\ mV$

$15 \times \frac{1000}{400} = \frac{150}{4}$

$$\text{Error} = \frac{15\times 10\cancel{00}}{4\cancel{00}} = \frac{150}{4}$$

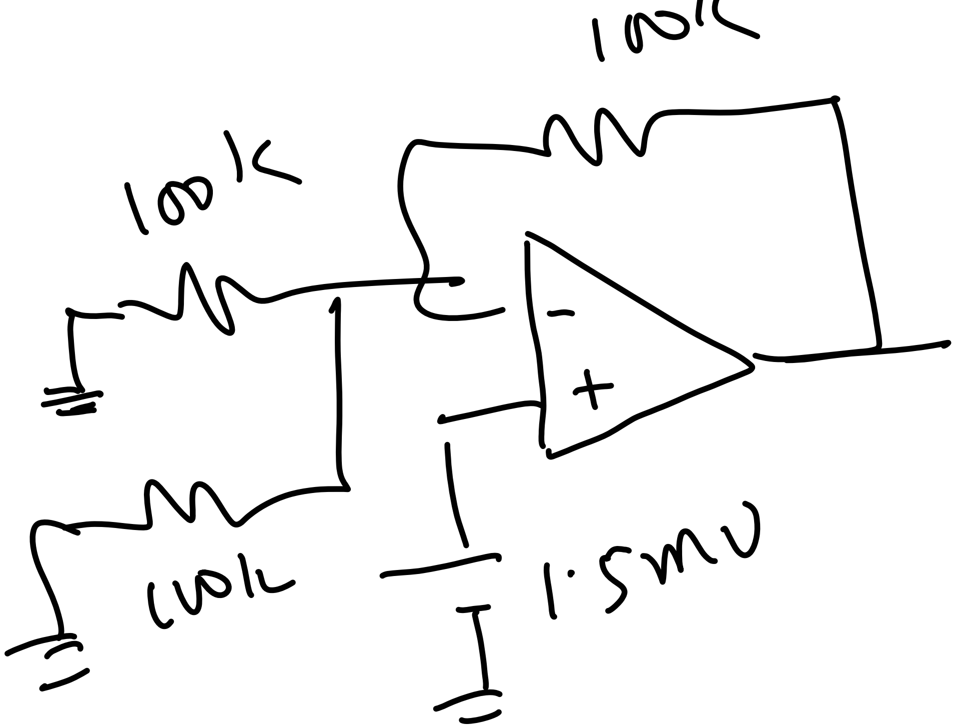
$$= \underline{37.5^\circ C}$$

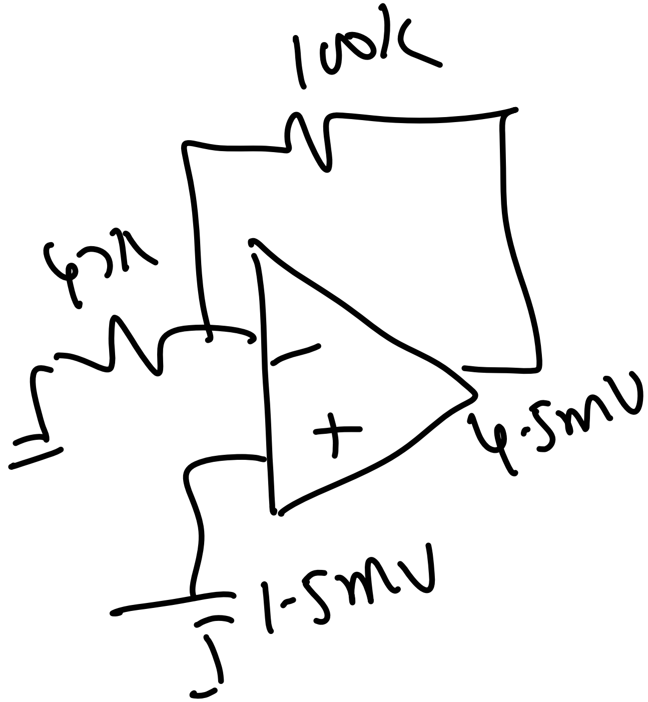
Actual temp $= 1^\circ C$

What is shown is $= 1^\circ C + 37.5^\circ C$

$= 38.5$

$= -36.5^\circ C$





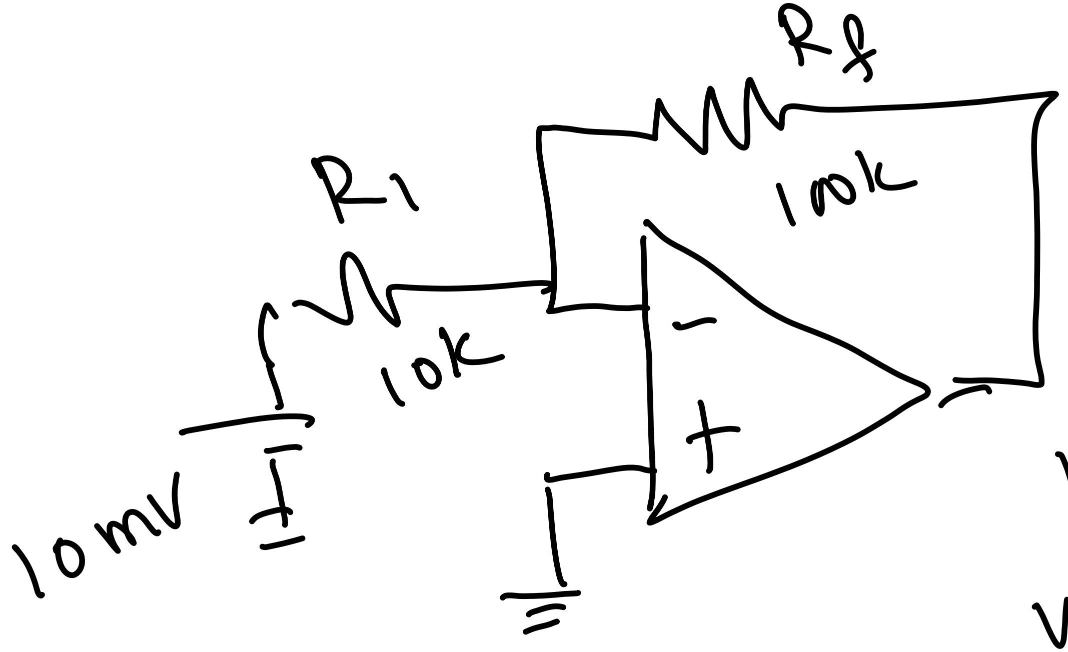
error due to $g = 3$

V_{OFF} drift $= 4.5\,mV = 0.45^\circ C$

error due to zener drift $= 7.5\,mV$ =

$= 0.75^\circ C$

Resistance drift

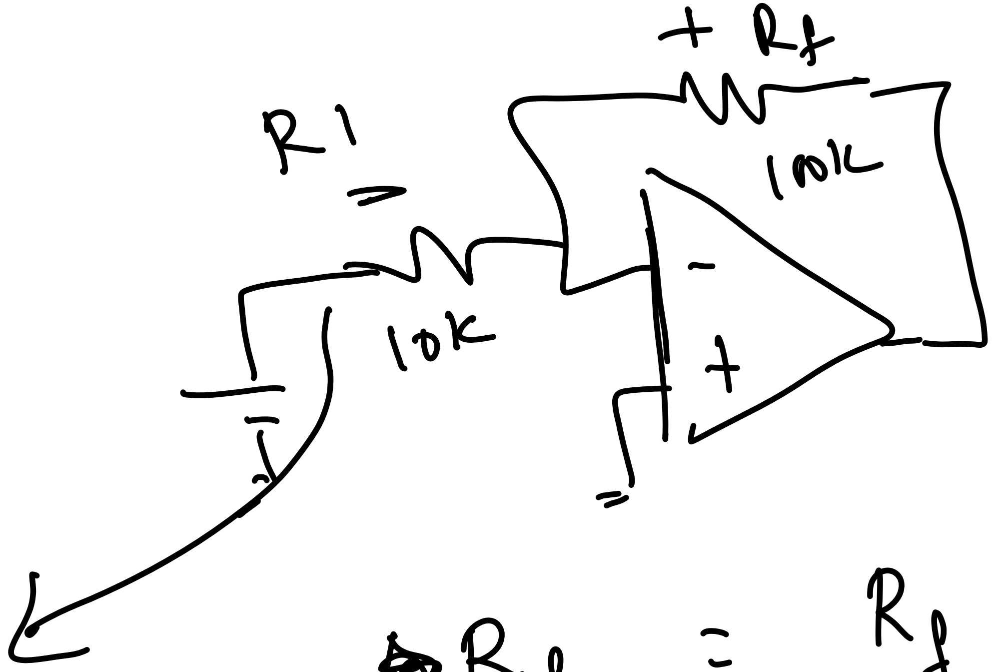


$g = 10$

$V_{in} = 10\,mV$

$V_o = -100\,mV$

Temp co-~~eff~~ff = ± 300 PPM/°C



At

$$100°C = \frac{100k + 3k}{10k - 300\Omega} = \frac{103}{9.7}$$

$$R_f \text{ at } 100°C = R_f + \frac{R_f \times 300 \times 100}{10^6}$$

$$= 10^5 + \frac{10^5 \times 3 \times 10^4}{10^6}$$

$$= 10^5 + 10^5 \times 3 \times 10^{-2}$$

$$= 10^5 + 3 \times 1000 = 100k + 3k$$

R_1 at 100°C

$$= 10^4 - \frac{10^4 \times 300 \times 100}{10^6}$$

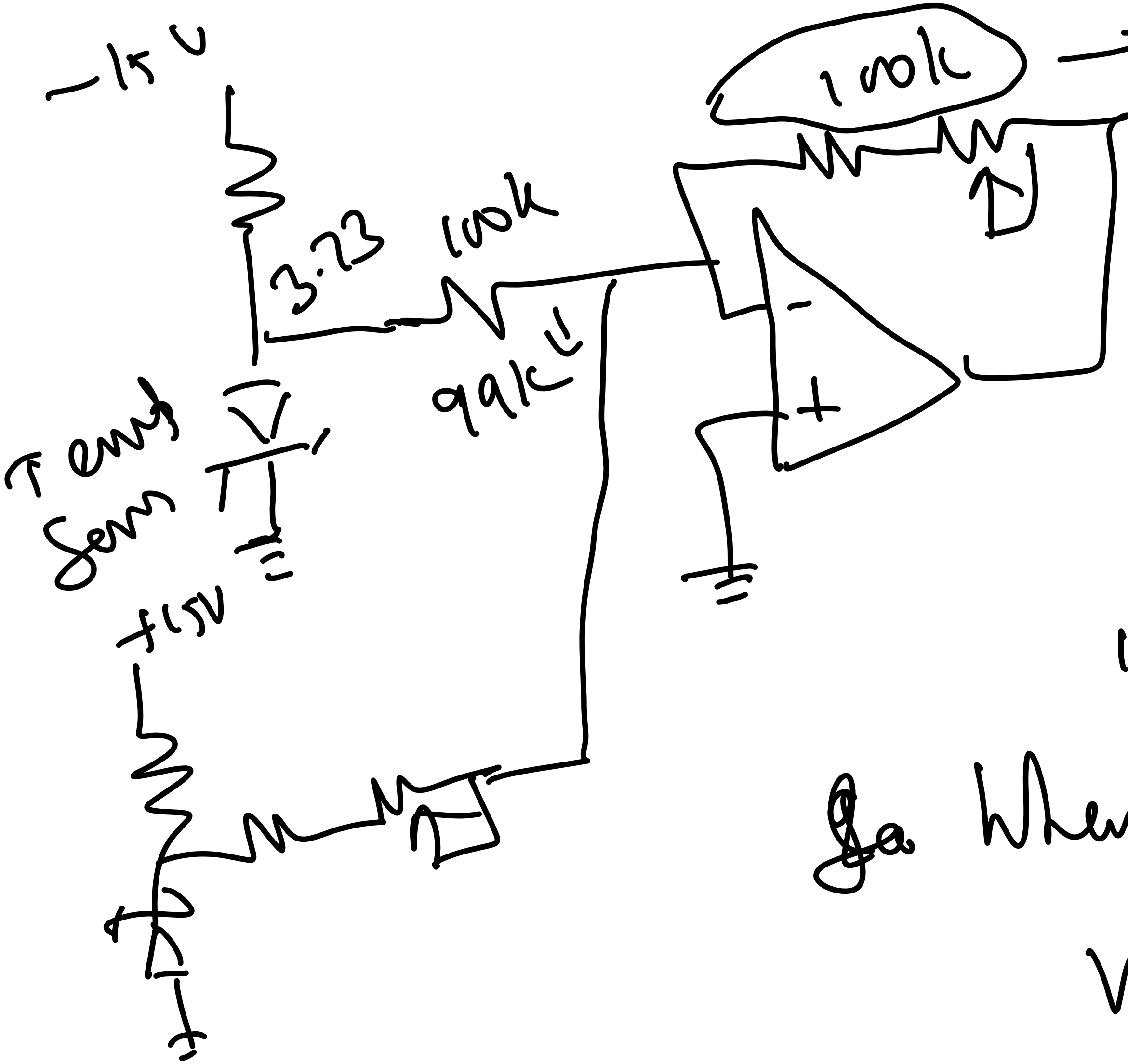
$$= 10^4 - \frac{10^4 \times 3 \times 10^4}{10^6}$$

$$= 10k - 300$$

New gain $= \dfrac{103}{9.7k}$

$V_0 = 10mv \times 11$

$= 110\ mv$



$\rightarrow 101k$

$$100k + 10^5 \times \frac{100}{10^6} \times 100$$

$$10^5 + 10^3 = 101K$$

When gain $= \frac{100k}{100k} = 1$

V_o due to Sensor $= 3.73V$

When gain $= \frac{101}{99}$

$$V_o = \frac{3.73 \times 101}{99} - 3.$$

$$\text{error} = \frac{3.73^v \times 101}{99} - 3.73$$

$$\simeq 37\ mv$$

(1) Zener drift error $= 7.5\ mV = 0.75^\circ c$

Resistance drift error $= 37\ mv = 3.7^\circ c$

Voff drift error $= 4.5\ mv = 0.45^\circ c$

Resistors → Carbon Comp
→ ± 300 ppm/°C

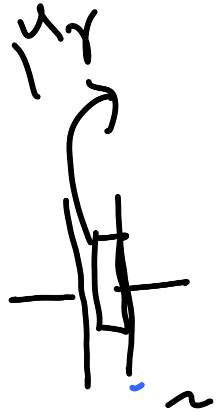
→ Metal film resistors
→ ± 10 ppm/°C → 100 ppm/°C

→ Wire wound resistors
→ +10 ppm/°C
+3000 ppm/°C

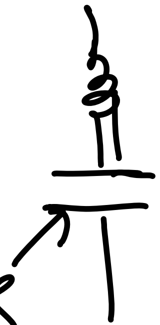
Capacitors

~~no~~ less inductance

1. Ceramic capacitors



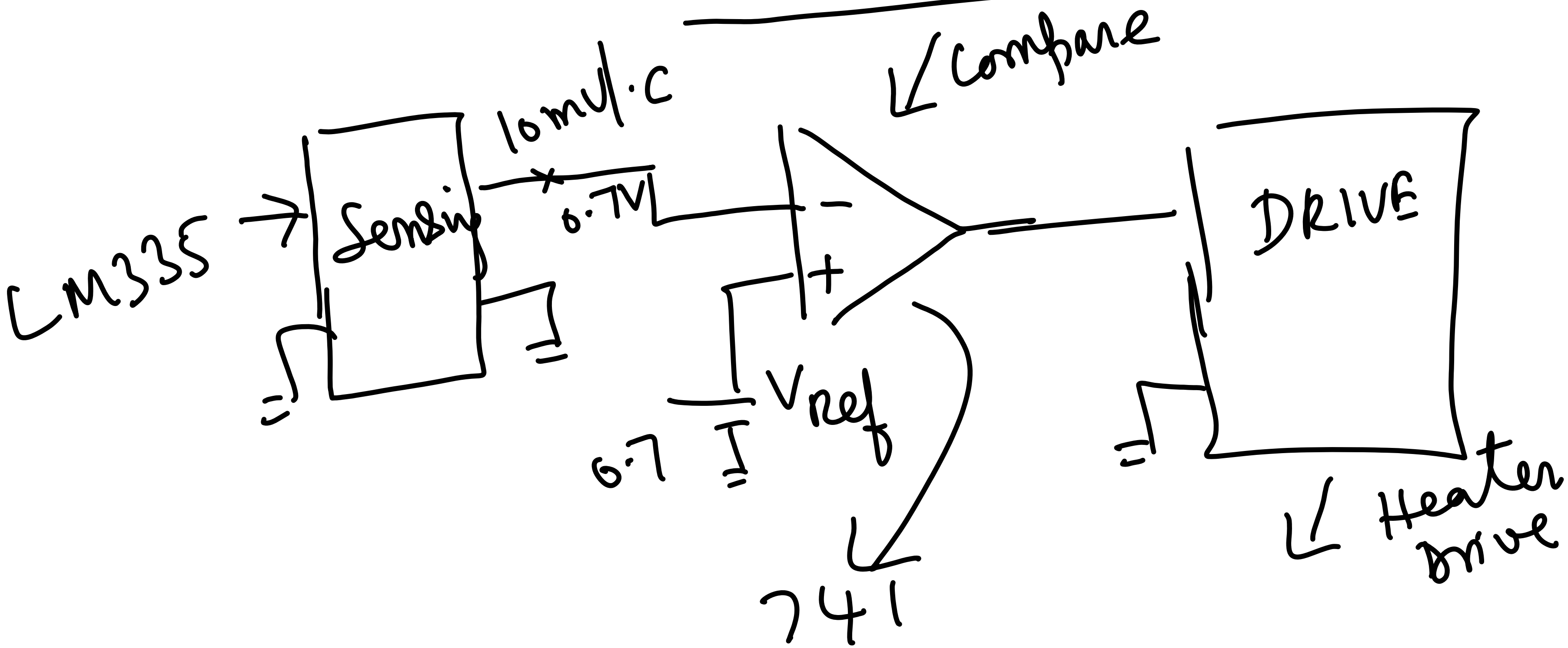
2. Electrolytic capacitors

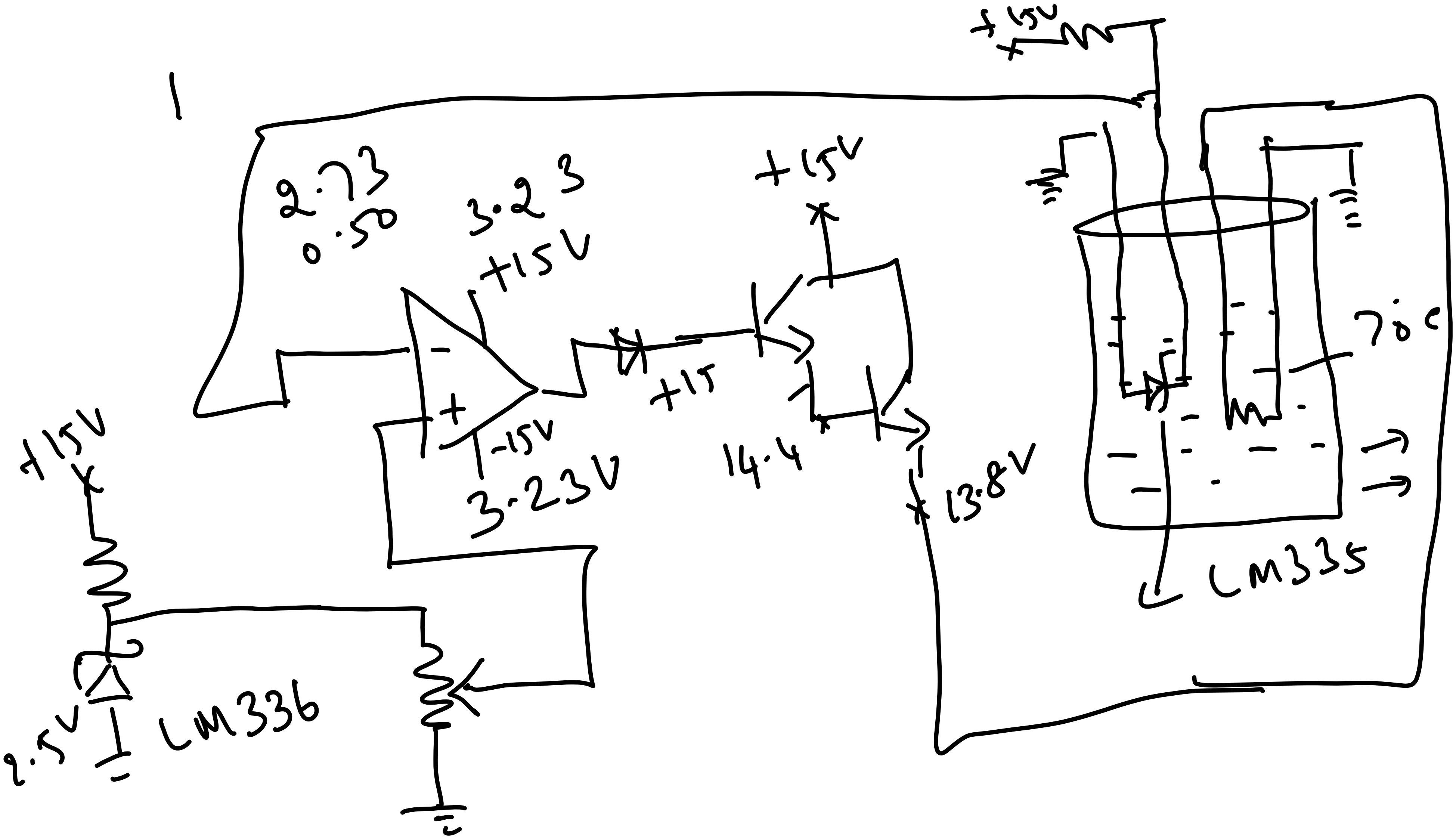


3. plastic film capacitors

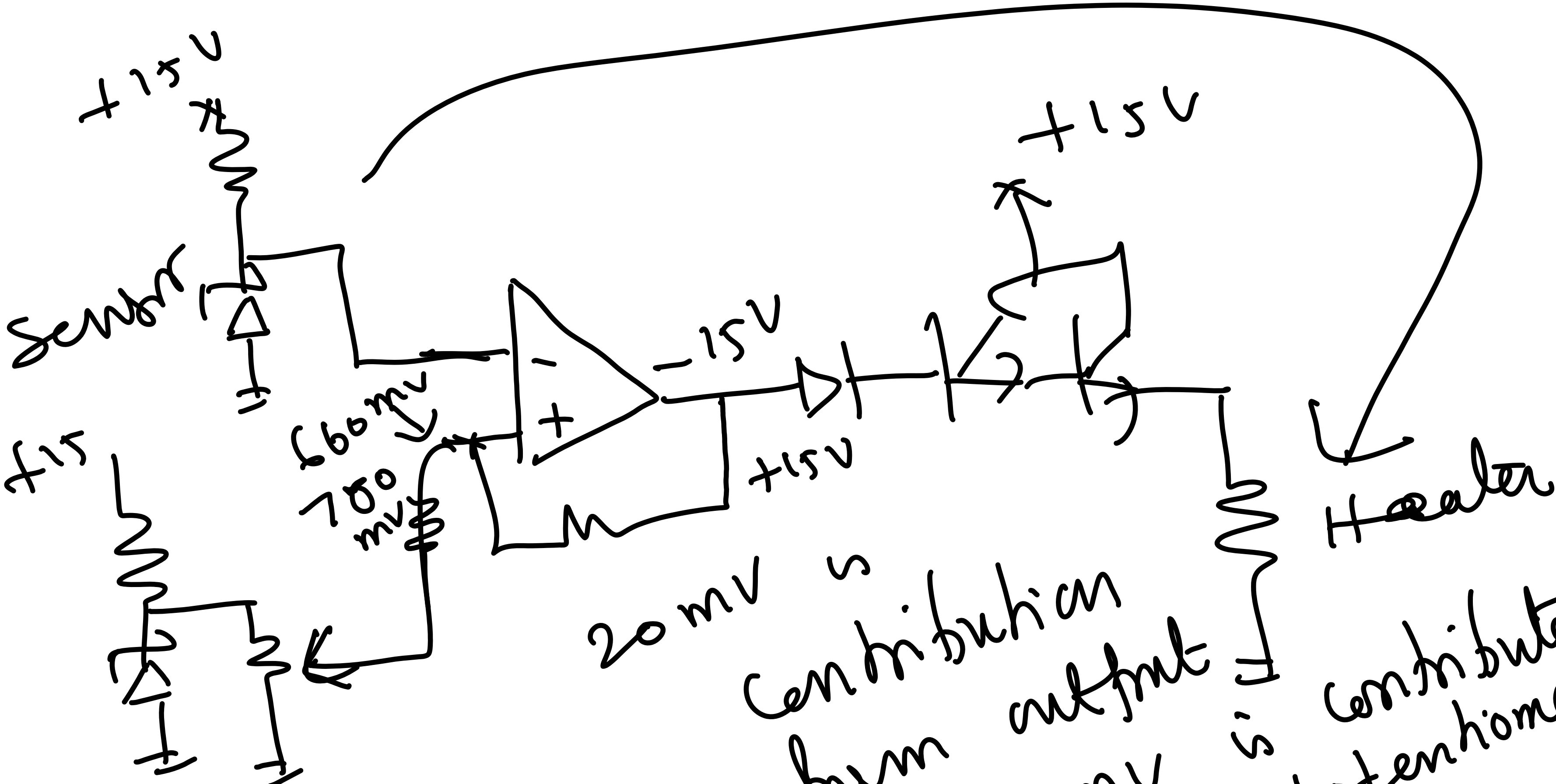
Temperature Controller

ON/OFF CONTROLLER

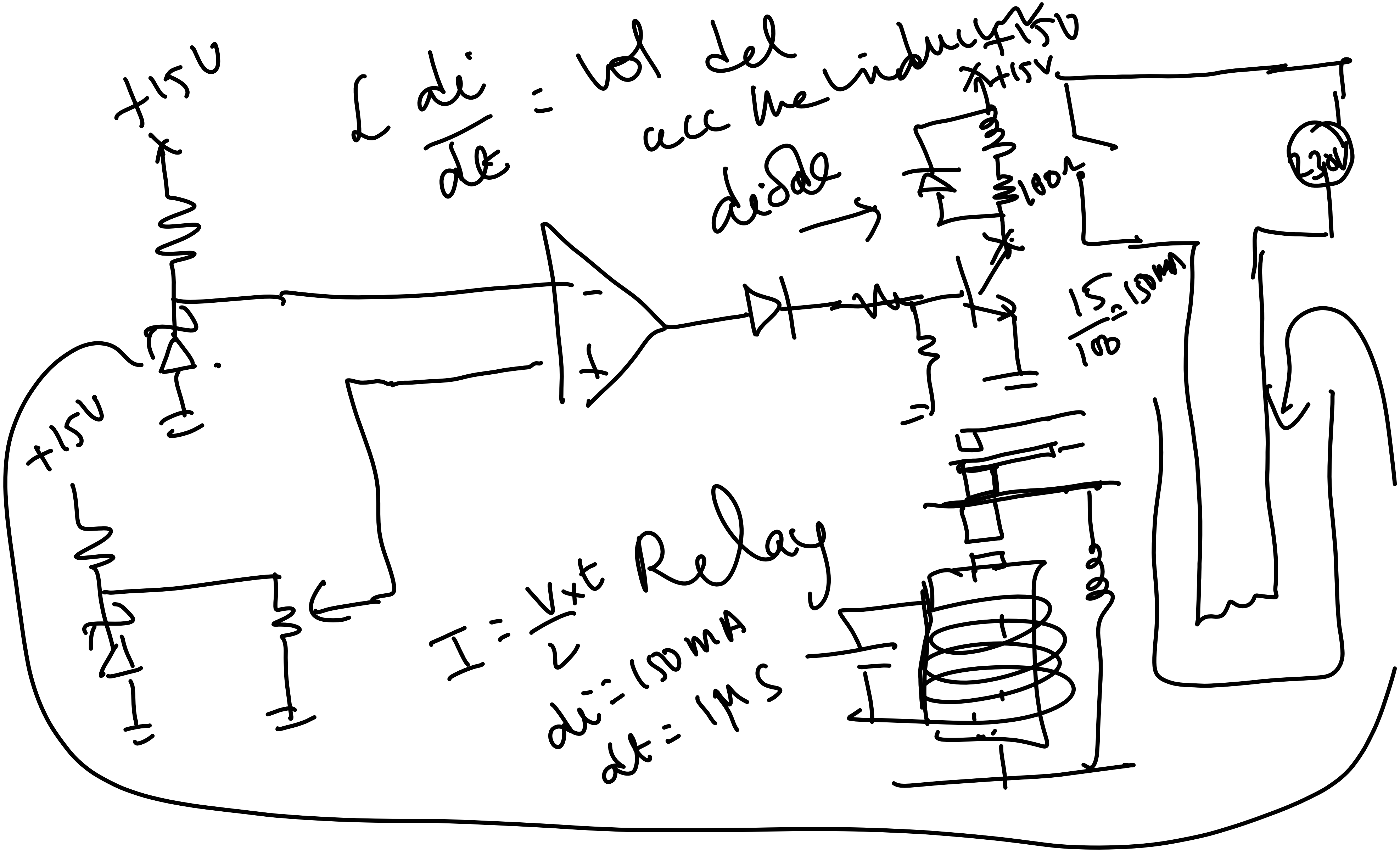




+15V
2.73
0.50
3.23
+15V
-15V
3.23V
+15
+15V
14.4
13.8V
+15V
70°C
LM335
+15V
2.5V
LM336



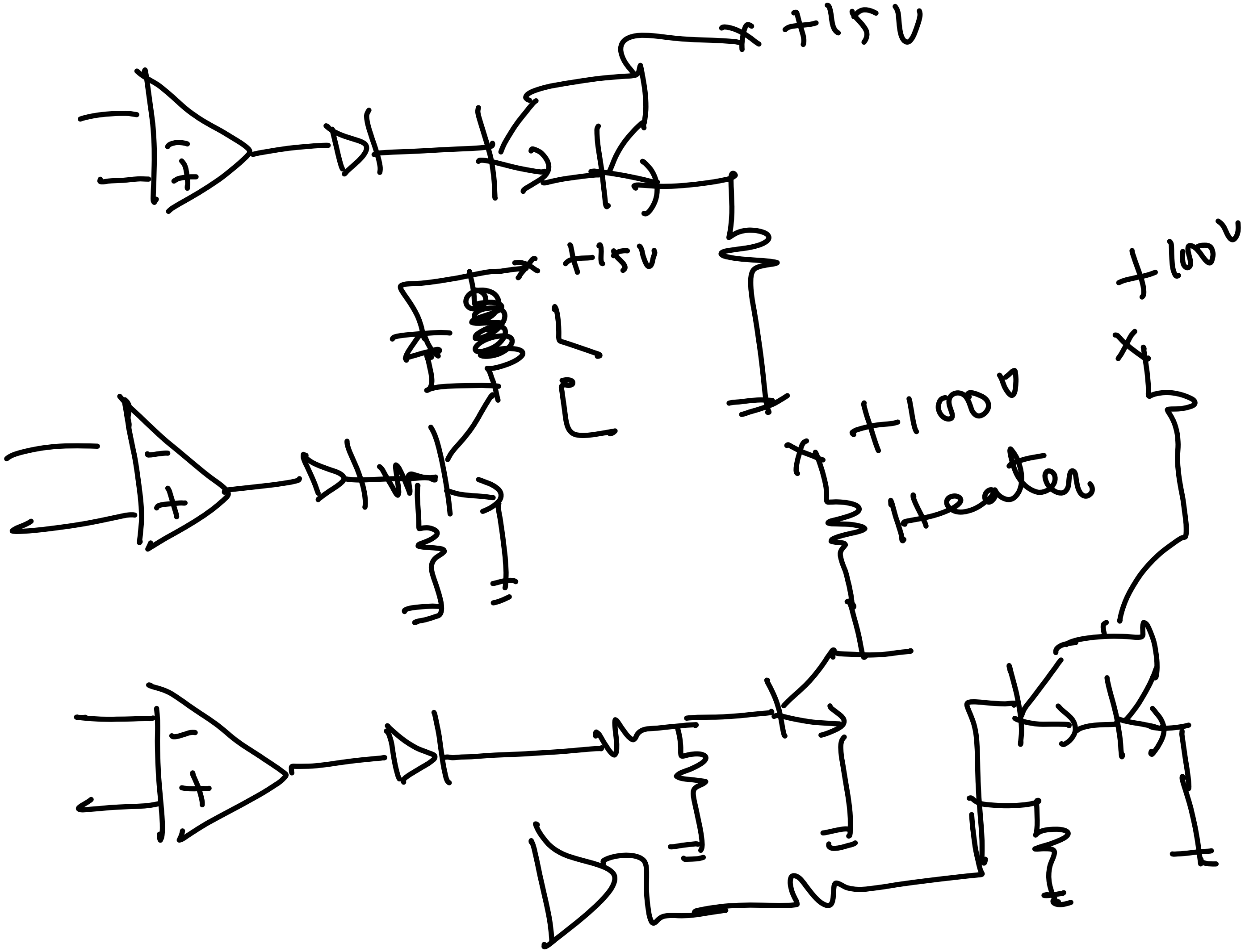
20mV is contribution from output
680mV is contributed by the potentiometer

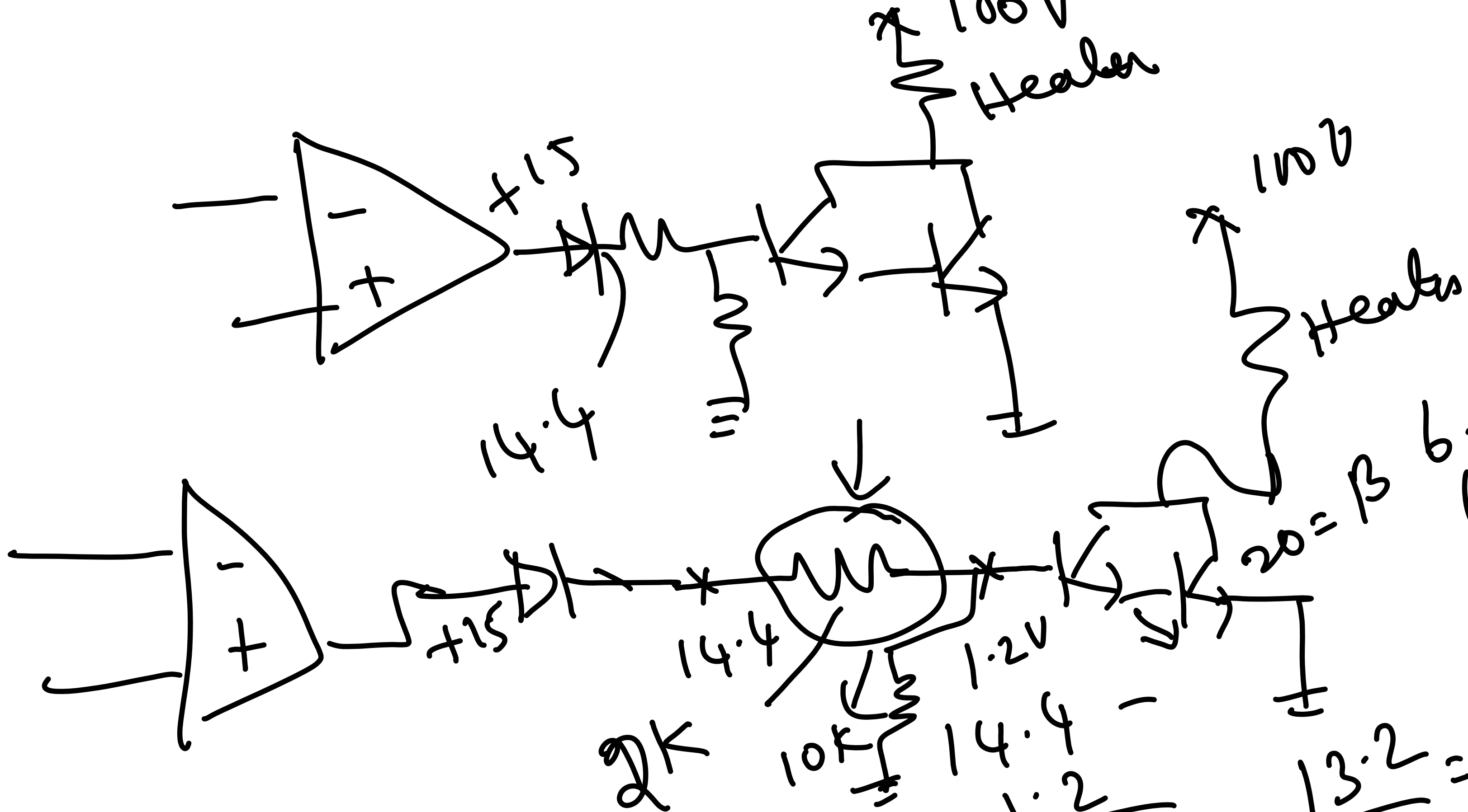


$L \frac{di}{dt}$ = vol del
acc the inductor
diode
+15V
+15V
+15V
+15V
100Ω
$\frac{15}{100}$ = 150mA
230V
Relay
$I = \frac{V \times t}{L}$
di = 150mA
dt = 1μs

Sol $\quad acc = L\frac{di}{dt} = \frac{1 \times 150 \times 10^{-3}}{10^{-6}}$

$= 150 \times 1000$





$$\begin{array}{r} 14.4 - \\ 1.2 \\ \hline 13.2 \end{array}$$

$$\frac{13.2}{2} = 6.6\ \text{mA}$$

$$6.6 \times 100$$

660 mA

Proportional Temperature Controller

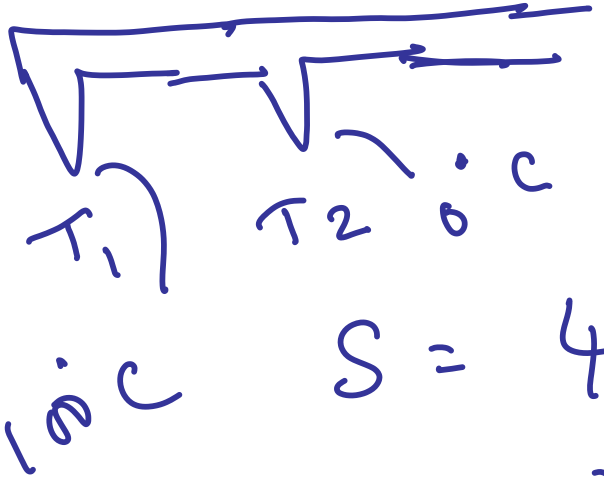
1) Thermo Couple as temp Sensor
② Heater vol as 100 V DC / 5 A

ON/OFF { Temp is low → Heater is ON
Temp is high → Heater is OFF

Proportional Controller
~~h~~eater vol $\propto$ error
heater vol $\propto (T_S - T_A)$
(50 − 30)

Heater o/p = Difference between set temp and actual temp (sensor

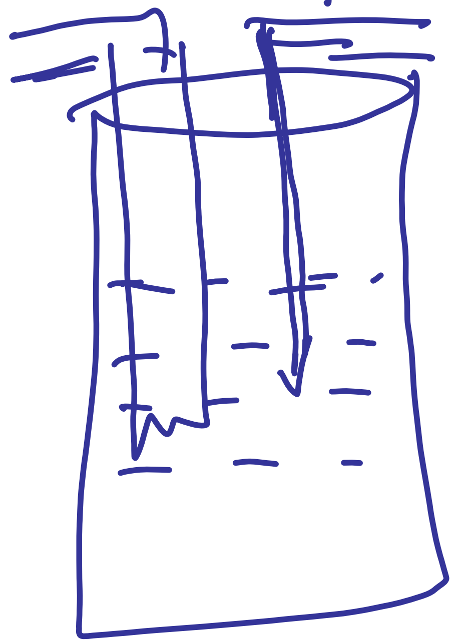
$(T_1 - T_2)S = V_0$



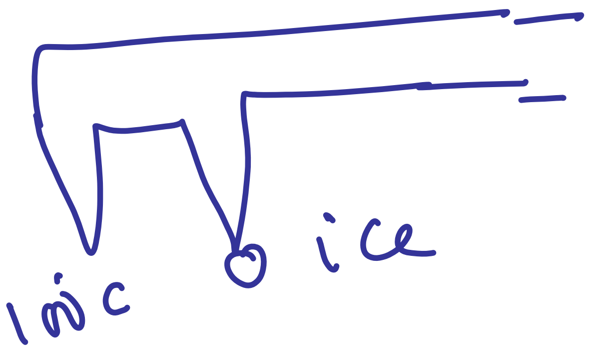
$S = 40\,\mu V/°C$

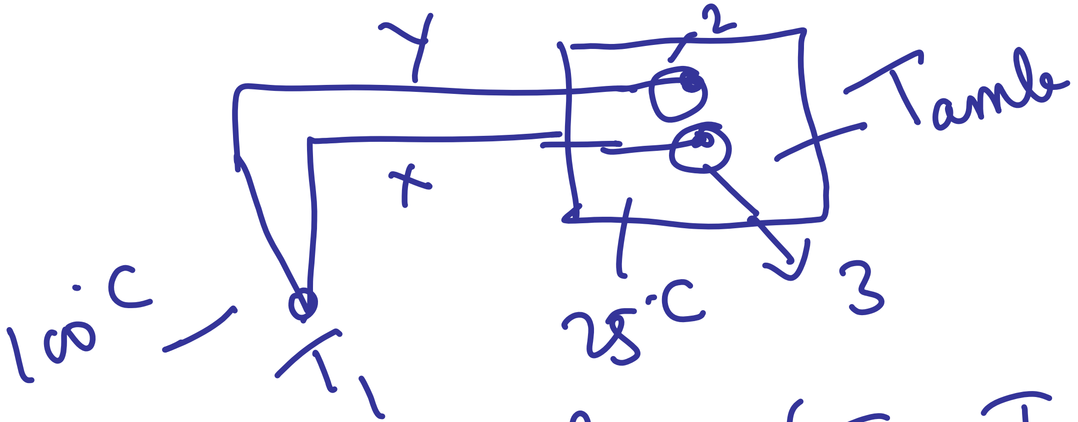
$(T_1 - T_2) = 100°C$

$= 40\,\mu V \times 100 = 4\,mV$



Single junction thermocouple





Vol at 2, and 3 $= (T_1 - T_{amb})S$

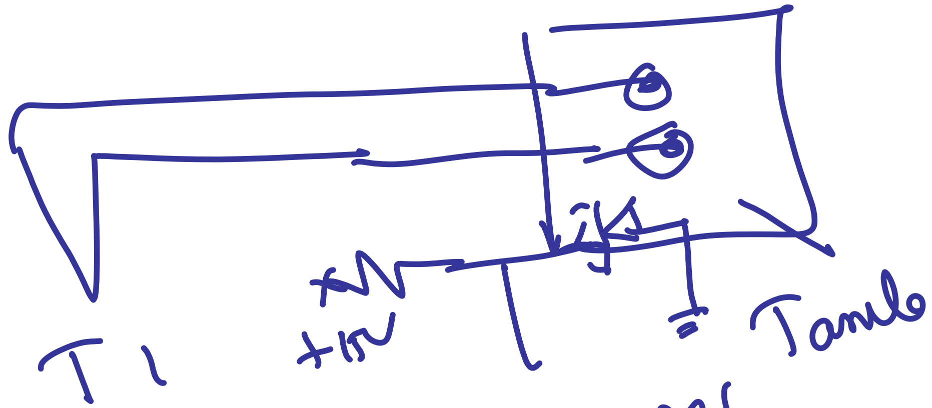
$S = 40\,\mu V/°C$

$(100-25)\times 40\mu V = 75 \times 40\,\mu V$

$= 3.000\ mV$

$(100-50)\times 40\mu V = 2.000\ mV$

ambient temp compensation



LM336

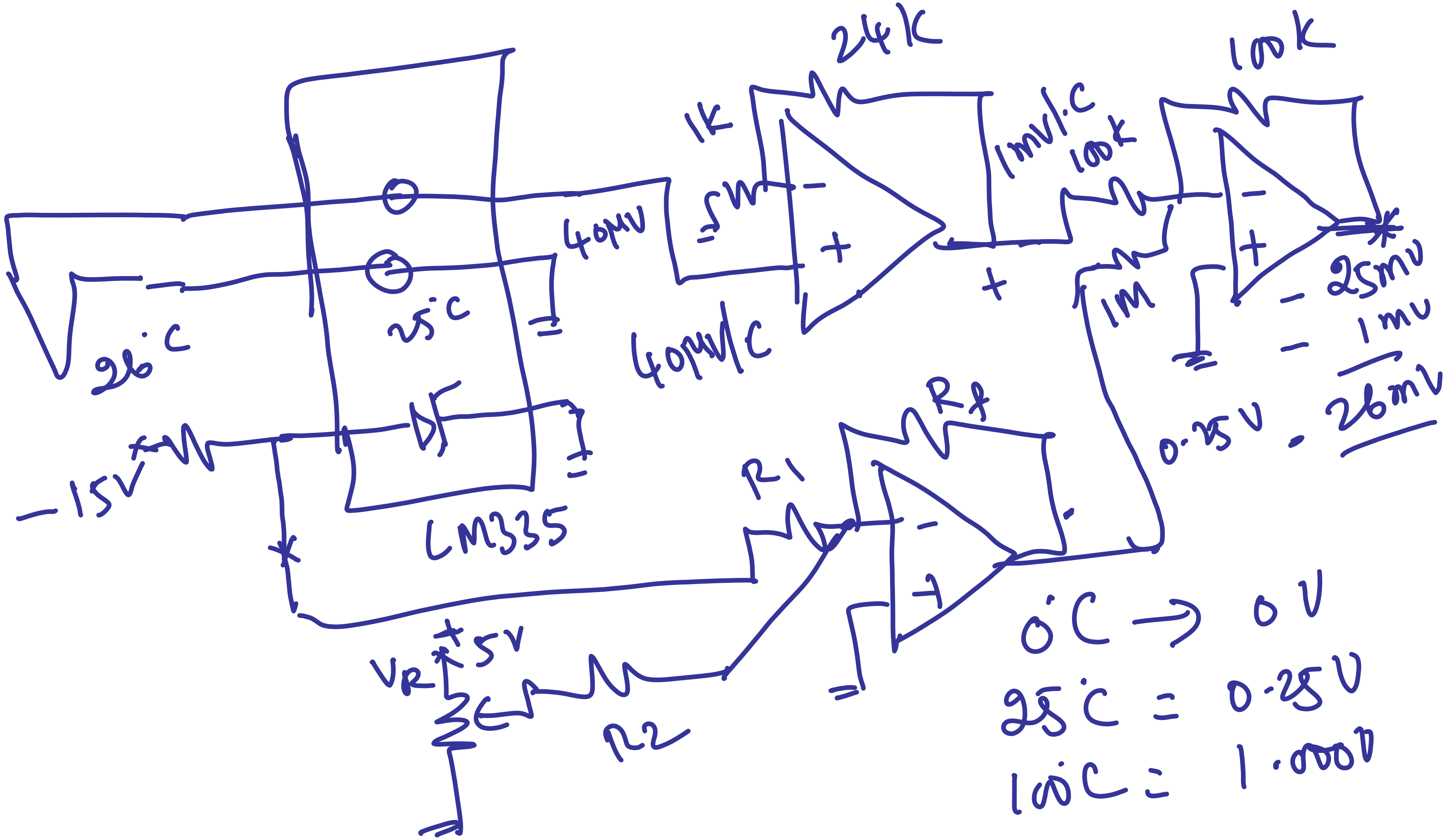
ambient temp must be measured. Use another sensor

Thermocouple w/

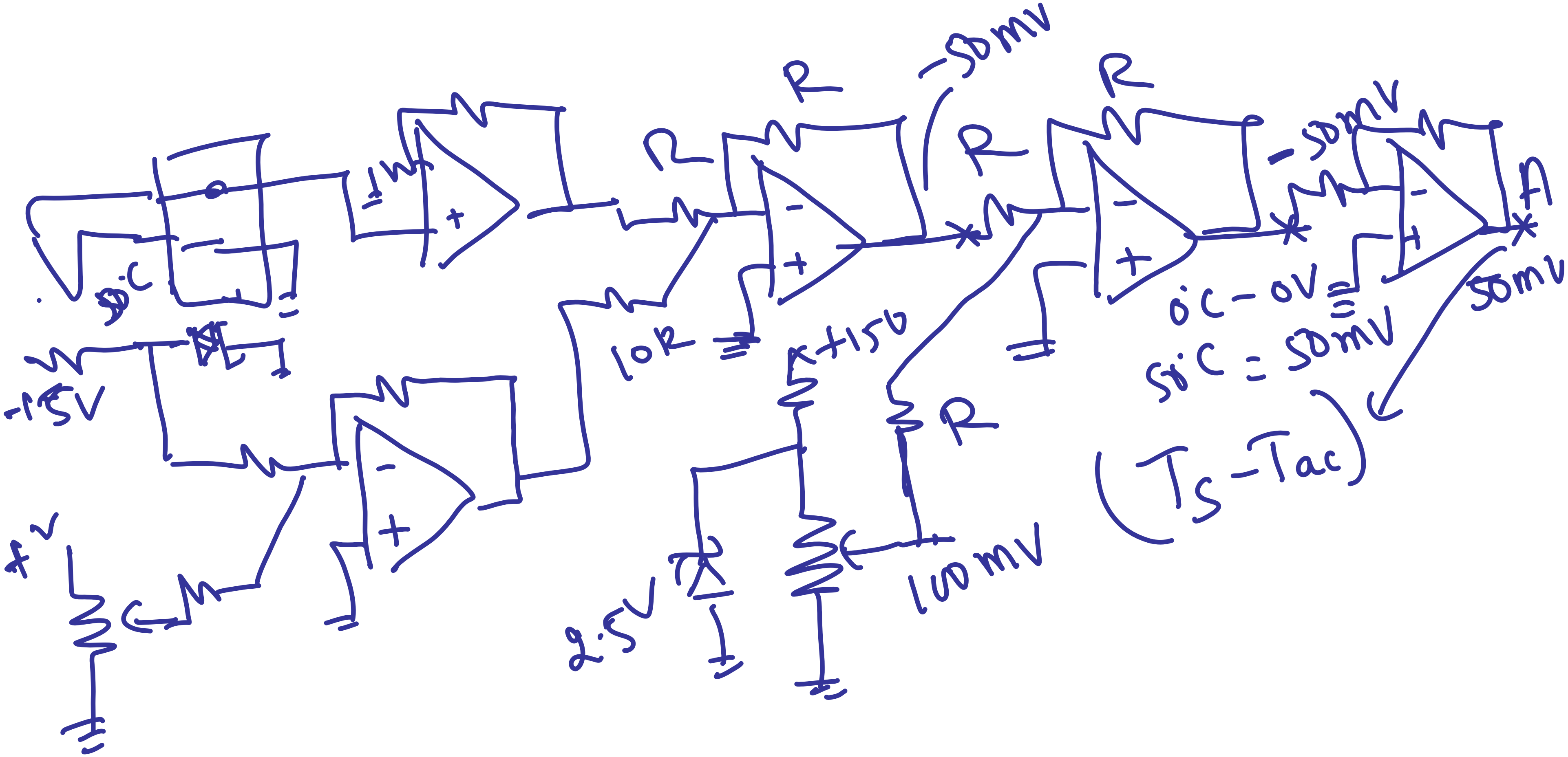
$$V_o = (T - T_{amb}) S$$

$$\frac{V_o}{S} = T - T_{amb}$$

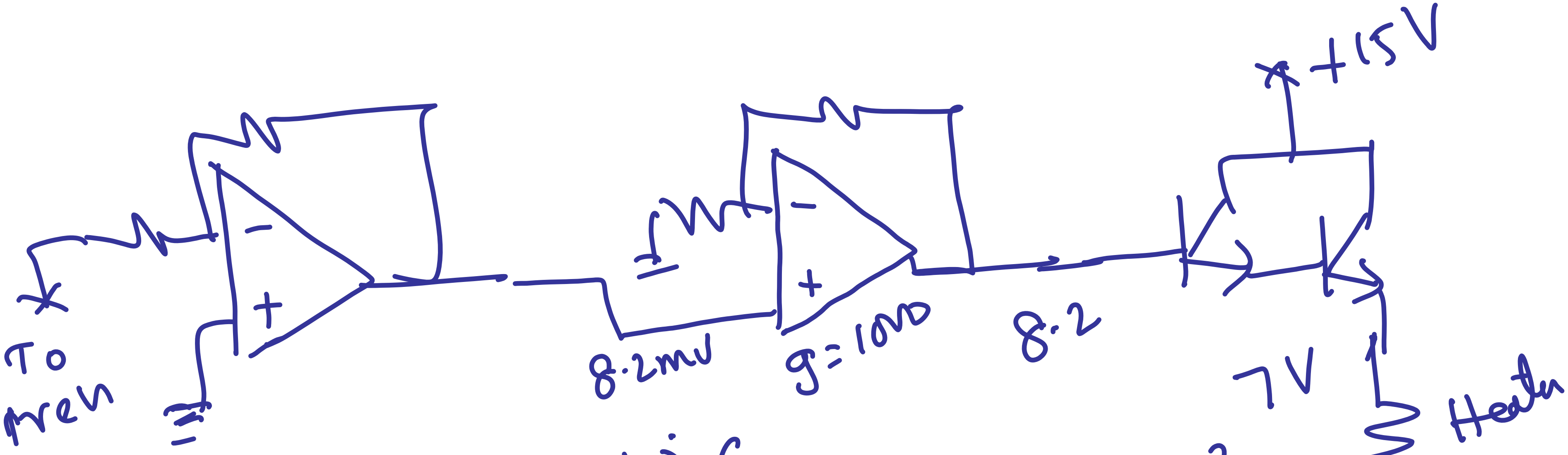
$$\frac{V_o}{S} + \text{ambient} = T$$

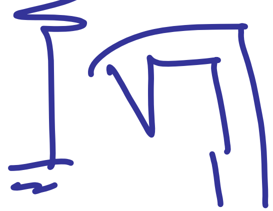


0°C → 0V
25°C = 0.25V
100°C = 1.000V



-50mV
R
R
R
R
1nF
0°C
-15V
+V
10R
+15V
R
2.5V
100mV
0°C - 0V
50°C = 50mV
-50mV
50mV
A
$(T_S - T_{ac})$





Set 100 C

Actual temp = 100 − 8.2

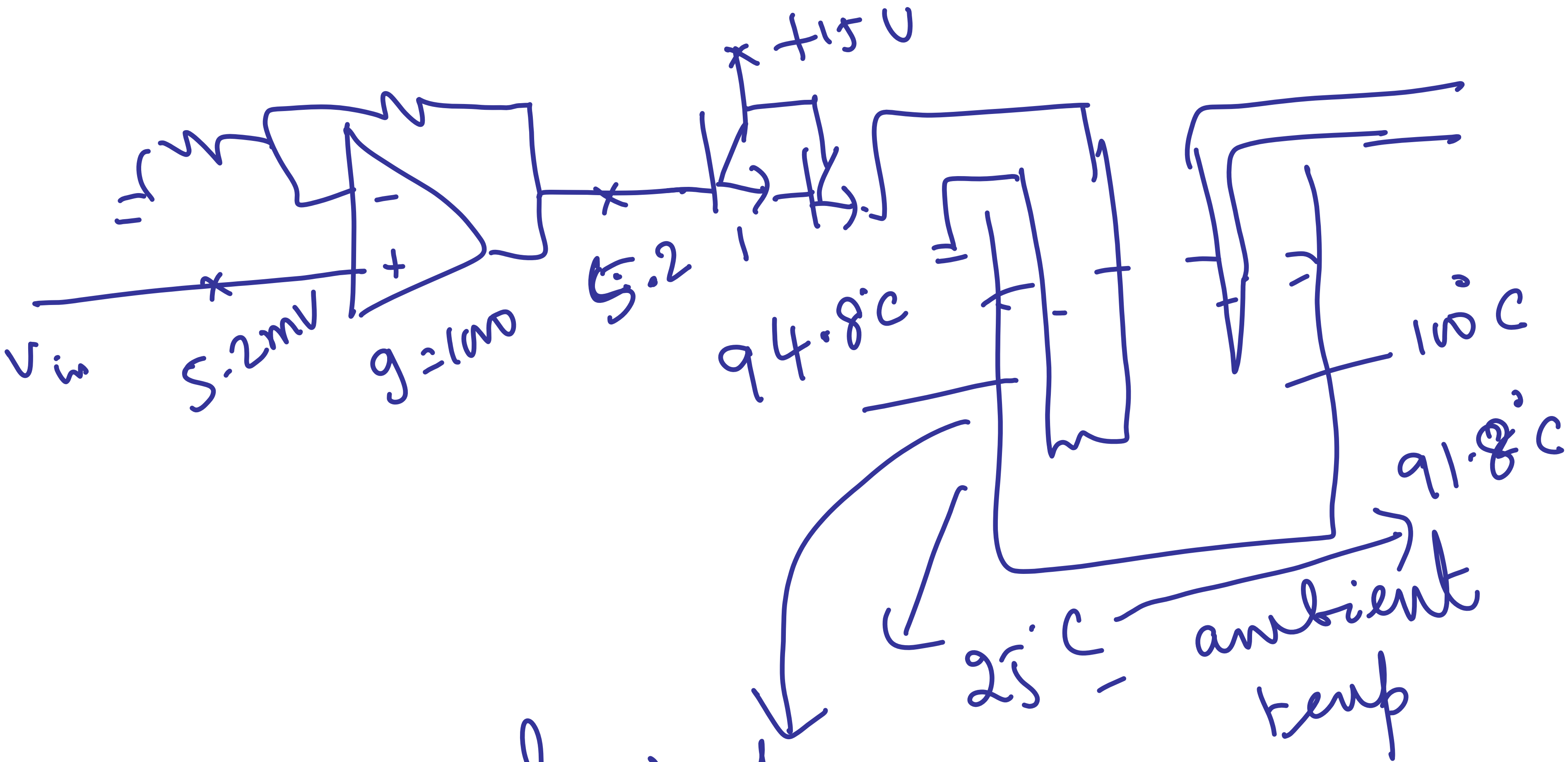
= 91.8°C

In proportional controller, Set temp is not equal to actual stabilized temp.

Set temp is different from actual stabilized temp.

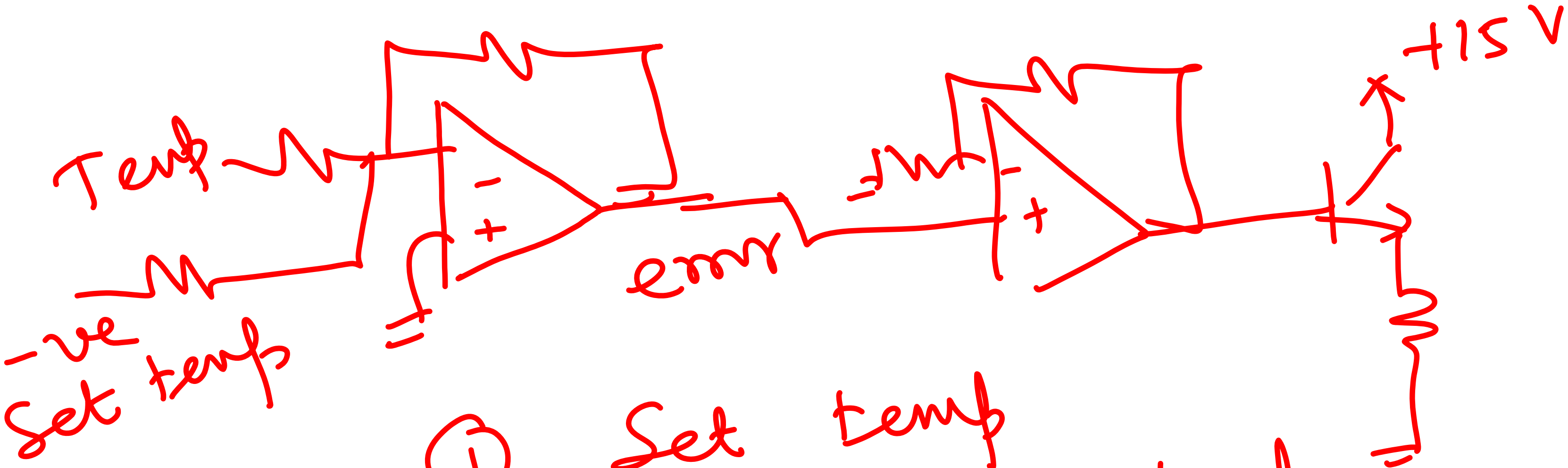
This difference only driving the Heater.

This difference produces secondary effect



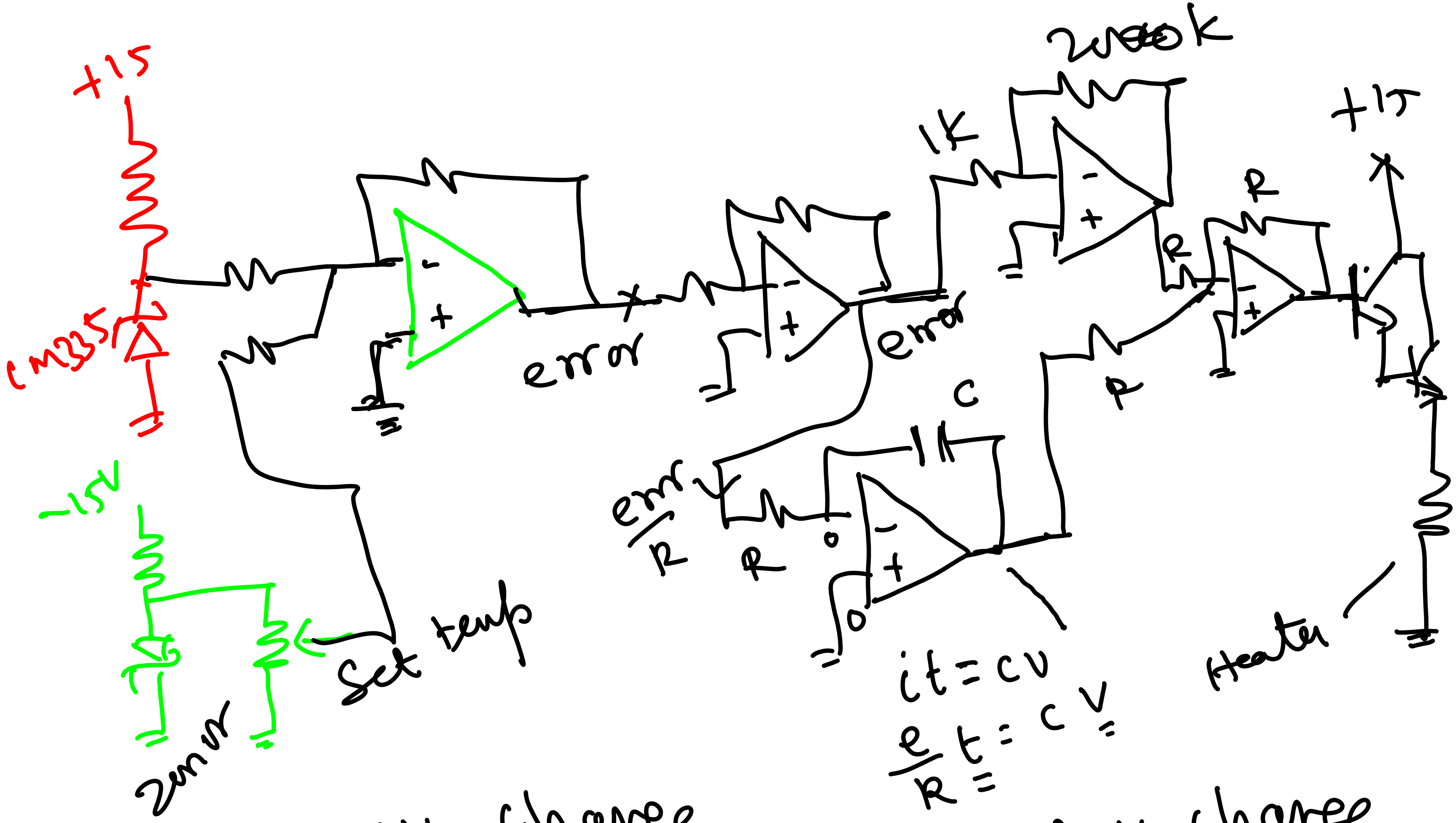
ambient temp is 50°C

PI Temperature Controller



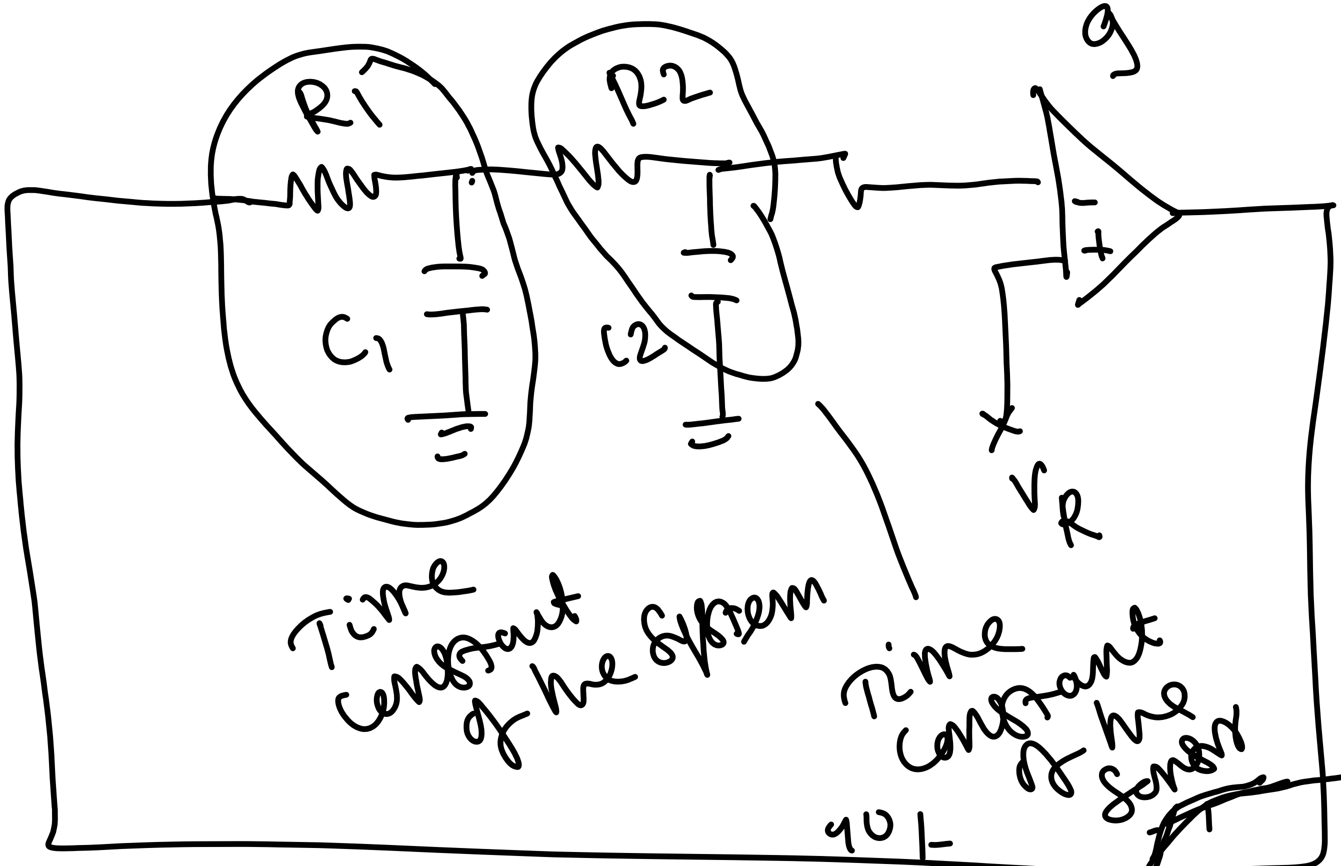
① Set temp = actual temp

error × gain = Heater vol



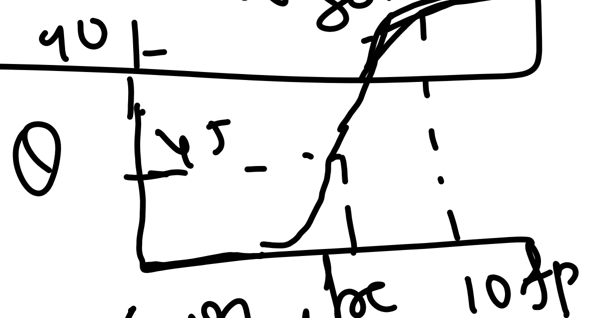
10mV change should produce 20V change at the heater

$$\text{electronic gain} = \frac{20V}{10mV} = 2000$$



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

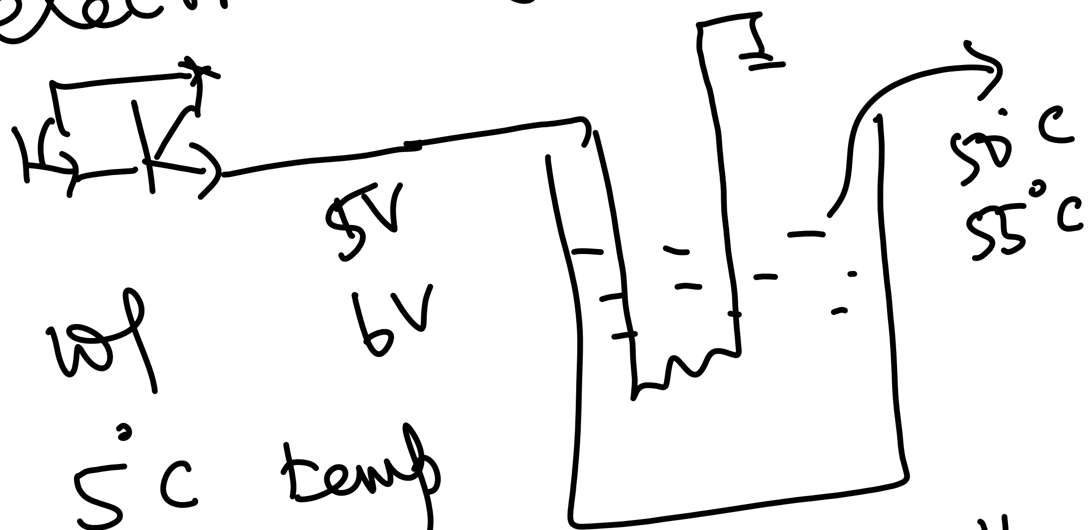
1 kHz



$A = \left(\frac{f}{f_p}\right)$

$\frac{100}{1} = \frac{\text{System time constant}}{\text{Sensor time cons}} > \text{gain}$

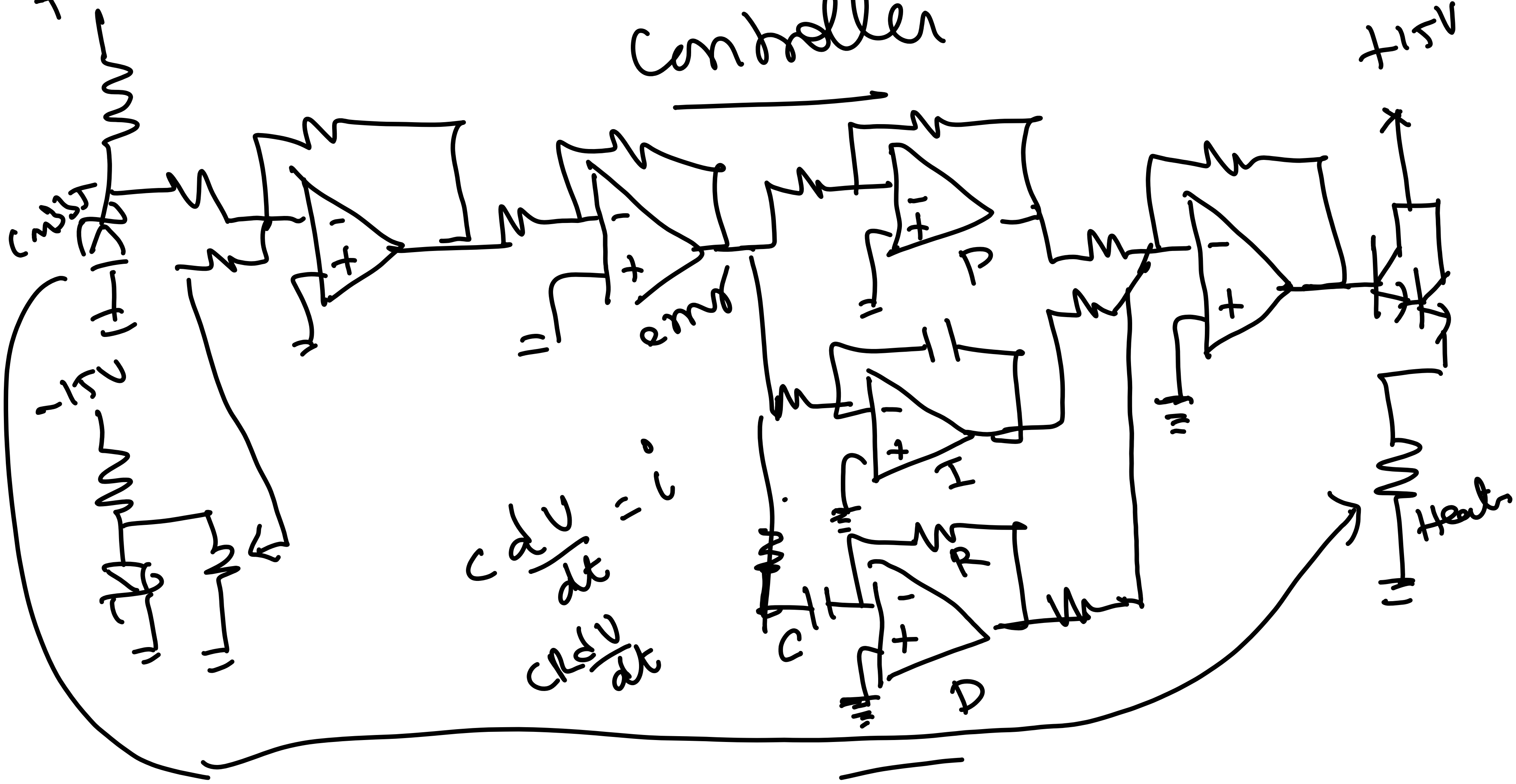
How to convert temperature gain into electronic gain?

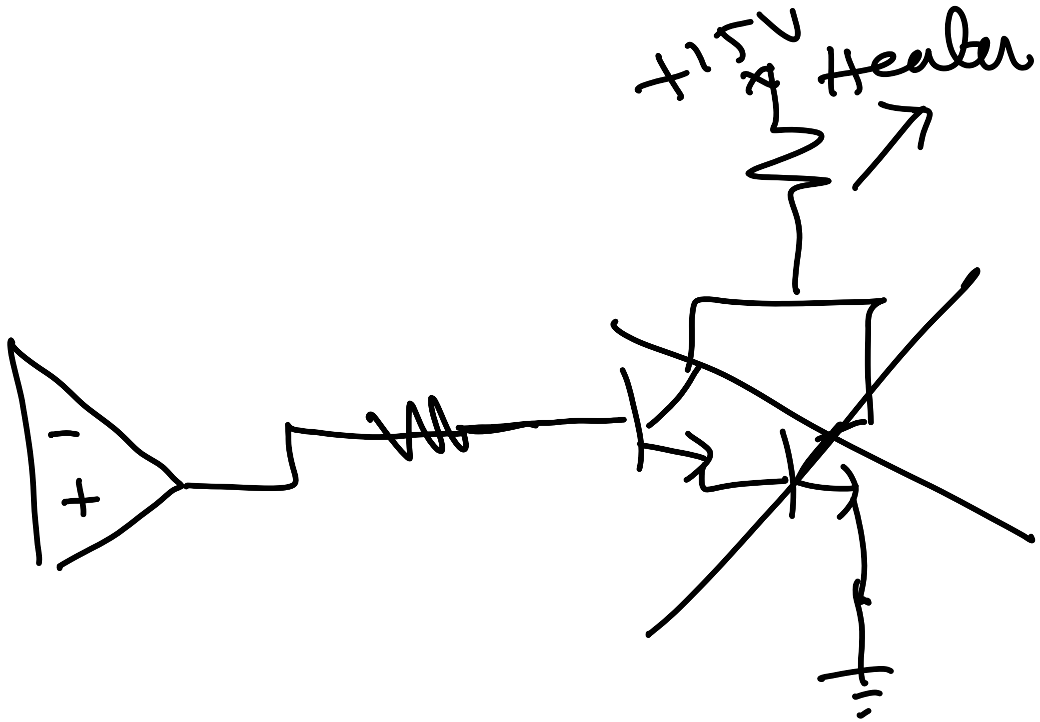


1V heater vol = 5°C temp

1°C temp rise gives us 10mV output from the sensor

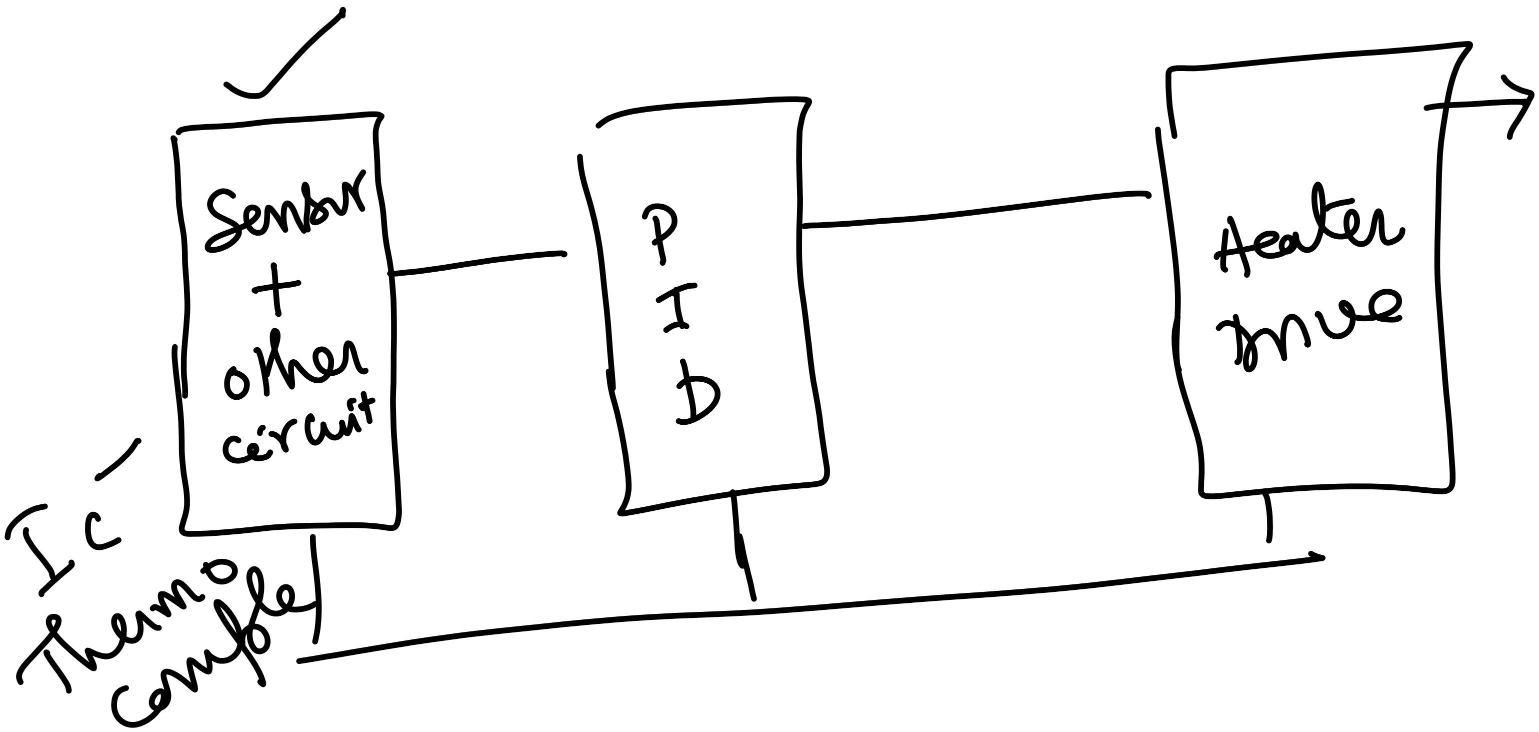
PID temperature Controller



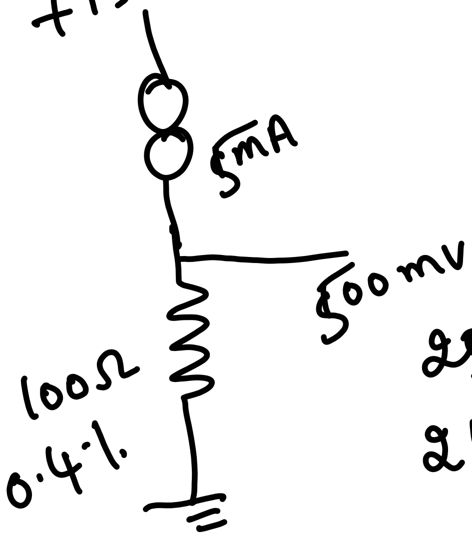


+15V
Heater
−
+

PID Controller



Platinum resistance thermometer

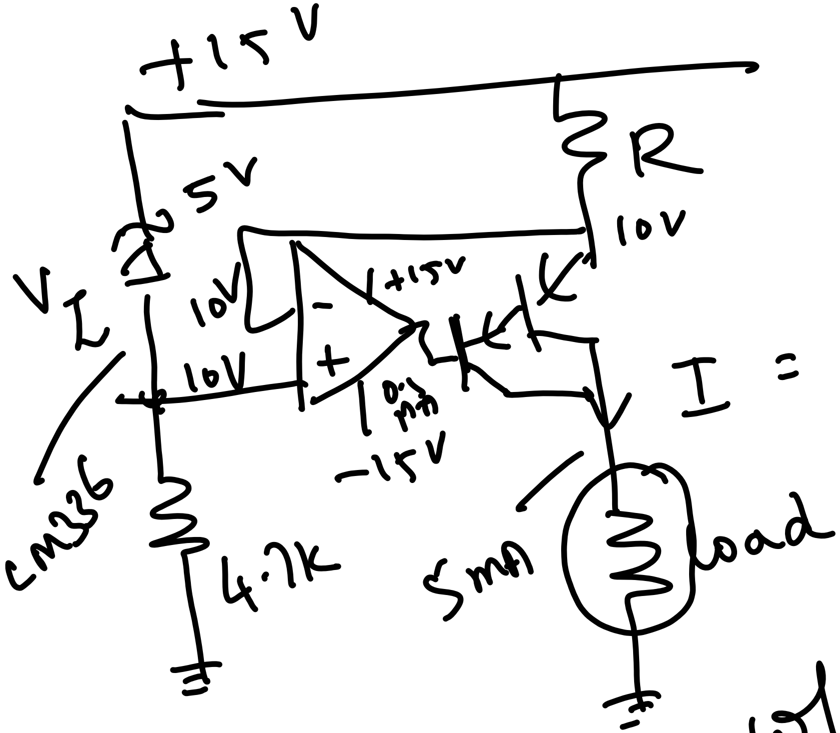


25°C → 500 mV

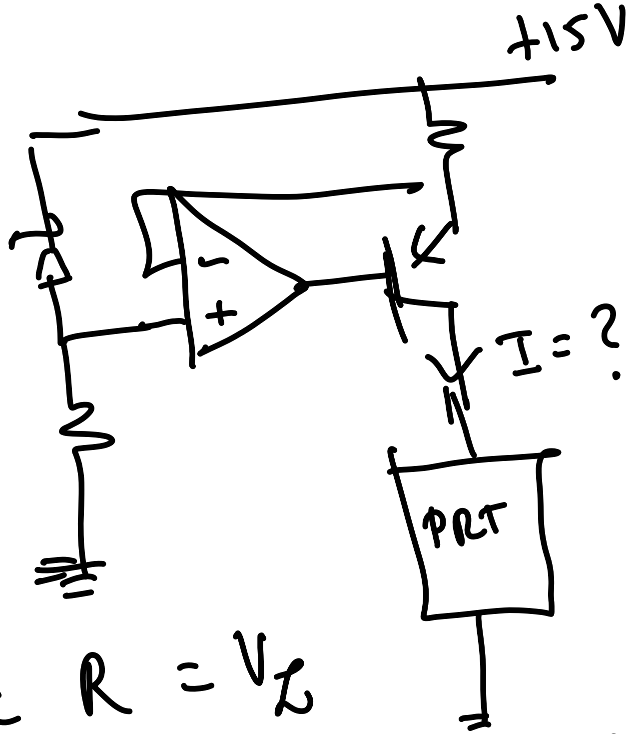
26°C → 100.4 Ω → 100.4 × 5 = 502.0 mV

Sensitivity = ~~0.4 mV~~/C

= 2 mV/C

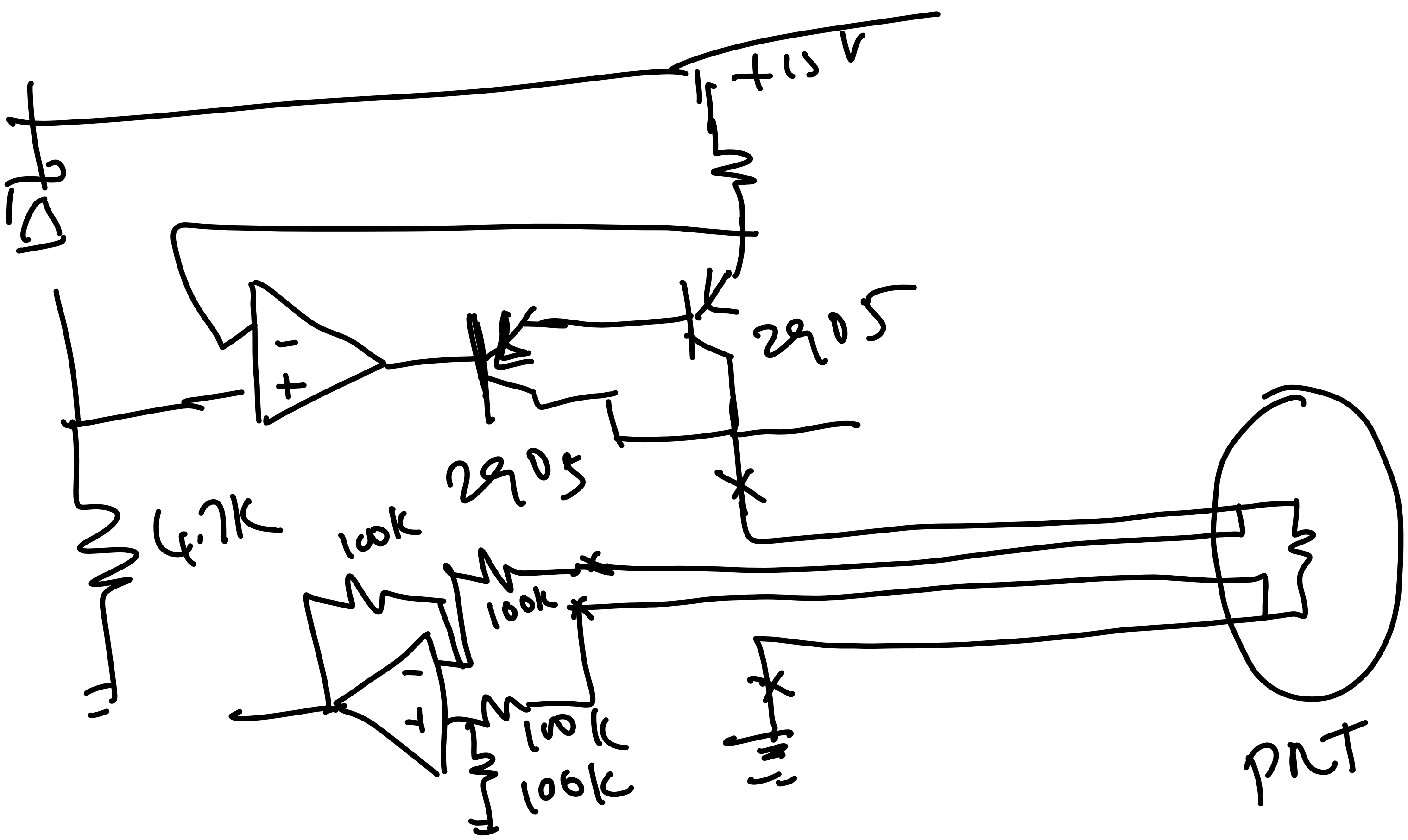


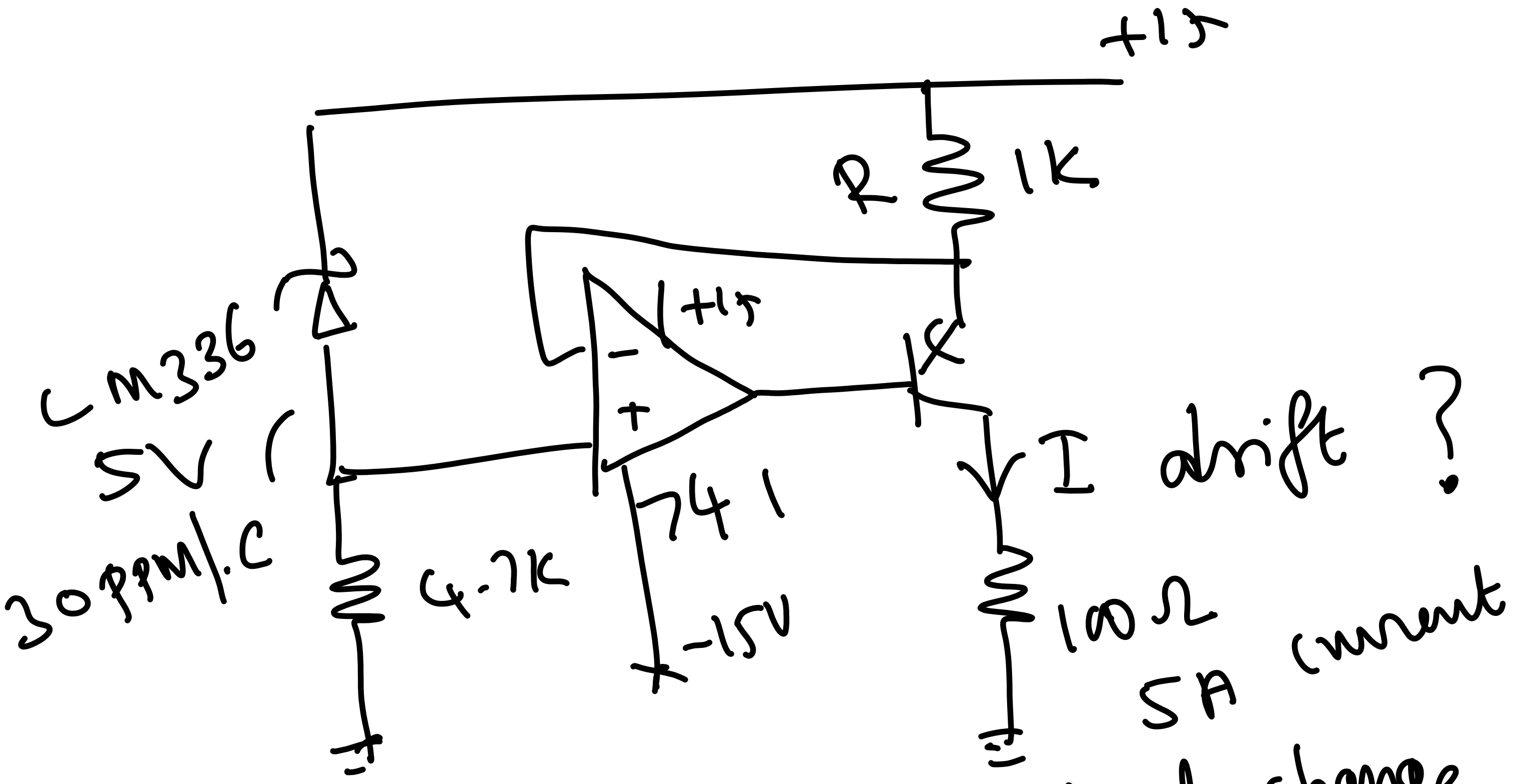
$$I = \frac{V_Z}{R}$$



w/ acc $R = V_Z$

$$I_R = \frac{V_Z}{R} = \frac{5V}{1K} = 5mA$$


+15 V
2905
2905
4.7K
100k
100k
100K
100K
PMT



For 100°C ambient temp change

$$V_Z \text{ drift} = \frac{5V \times 30 \times 100}{10^6}$$

$$= \frac{15}{10^3} = 15\,mV$$

current drift due to V_Z drift

$$= \frac{15\,mV}{1k} = 15\,\mu A$$

Current drift due to V_2 drift $= 15\,\mu A$

w/ change acc PRT due to $15\,\mu A$

current drift $= 15 \times 10^{-6} \times 100$

$= 1500\,\mu V = 1.5\,mV$

Sensitivity of PRT $= 2\,mV/°C$

So error in temp due to

$1.5\,mV$ drift $= \frac{1.5}{2} = 0.75°C$

<u>Temp error due to R change</u>

$$R = 1000\ \Omega$$

$$\text{Temp co-eff} = \pm 300\ \text{ppm}/^\circ C$$

$$\Delta T = 100^\circ C$$

$$\Delta R = 1000 \times \frac{300}{10^6} \times 100$$

$$\Delta R = 3 \times \frac{10^7}{10^6} = 30\ \Omega$$

$$\Delta I = \frac{V_z}{R} = \frac{5}{1000+30}$$

$$= \frac{5}{1030} = \frac{5000}{1030} = 4.86\ \text{mA}$$

$$\Delta I = 0.14\ \text{mA}$$

5000
4120
8800
8240
6600

voltage error for 0.14 mA

Current change in PRT $= 0.14 \times 100$
$= 14\ mV$

Sensitivity $= 2\ mV/C$

Total temp error $= \frac{14}{2} = 7^\circ C$

For ± 25 ppm drift resistance

Error due to offset voltage drift of the op amp

V_{off} drift of 741 $= \pm 15\ \mu V/^\circ C$

$\Delta T = 100^\circ C$

Total V_{off} drift $= 15 \times 100$
$= 1.5\ mV$

current error due to

V_{off} drift of 741

$$= \frac{1.5 \times 10^{-3}}{1k} = 1.5\ \mu A$$

Change in o/p

acc $100\,\Omega$ PRT $= 1.5\ \mu A \times 100$

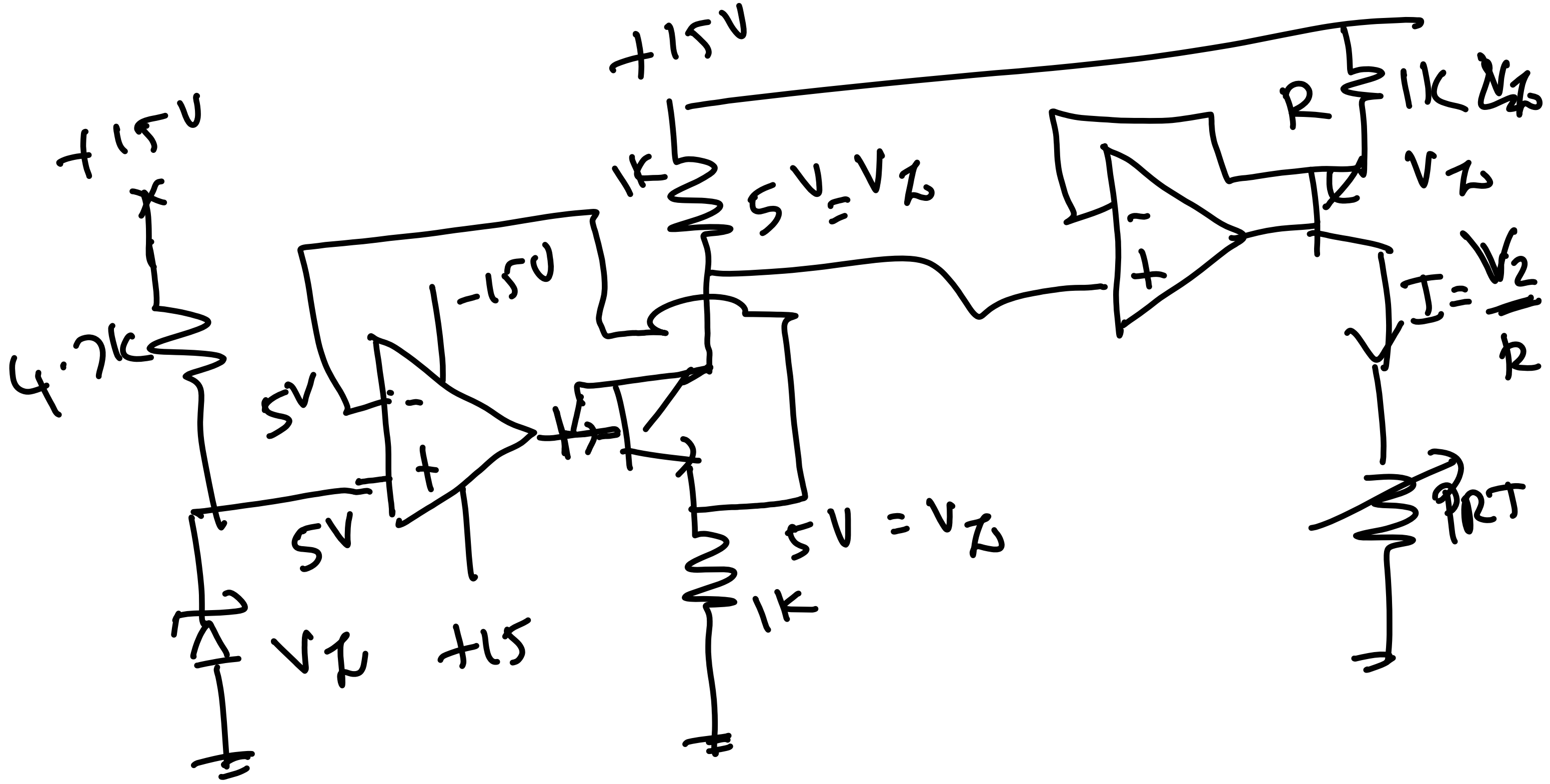
$= 150\ \mu V$

Sensitivity $= 2\ mV/°C$

$$150\ \mu V = \frac{0.15}{2} = \frac{15}{200} = \frac{7.5}{100} = 0.075$$

Temp error due to

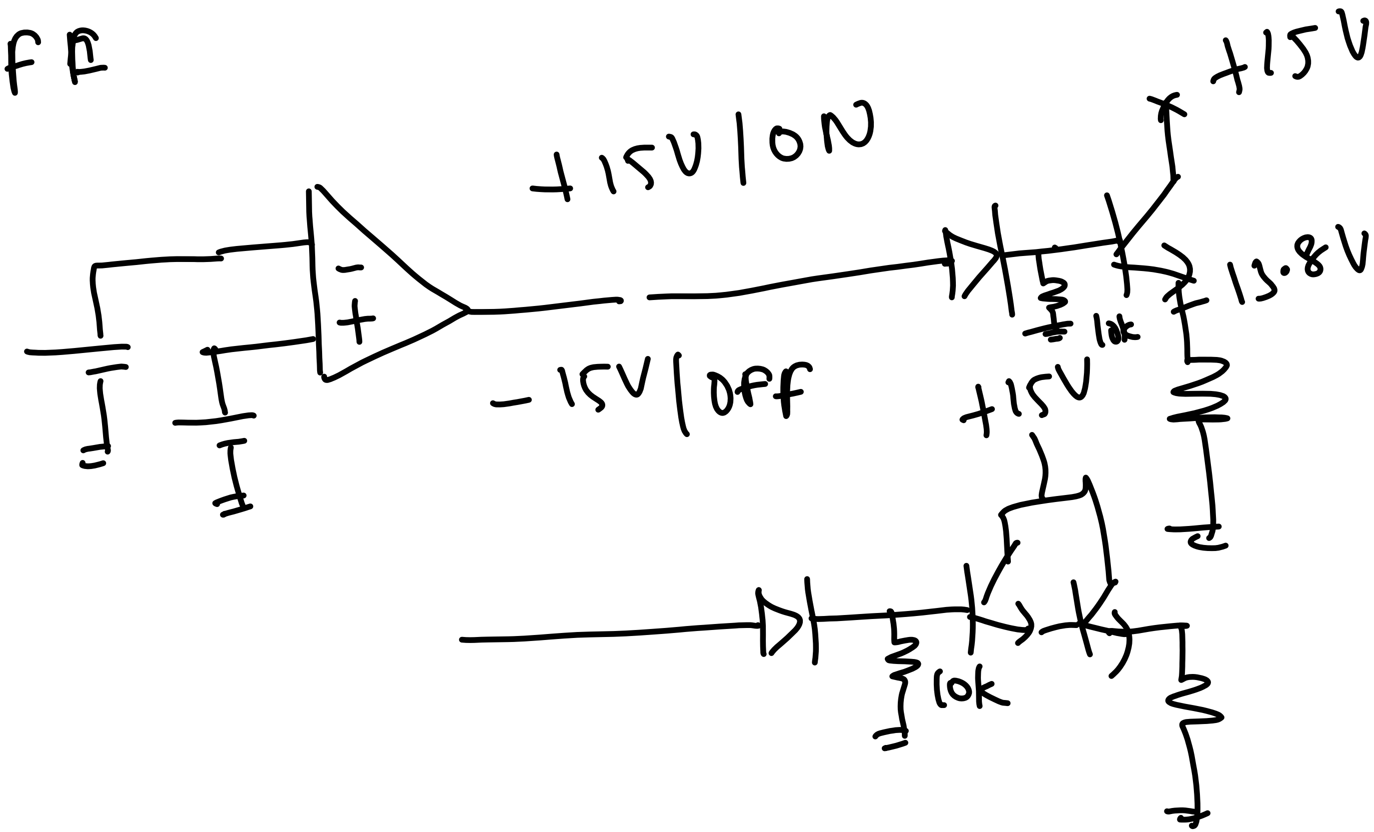
offset o/p drift of op amp $= 0.075°C$

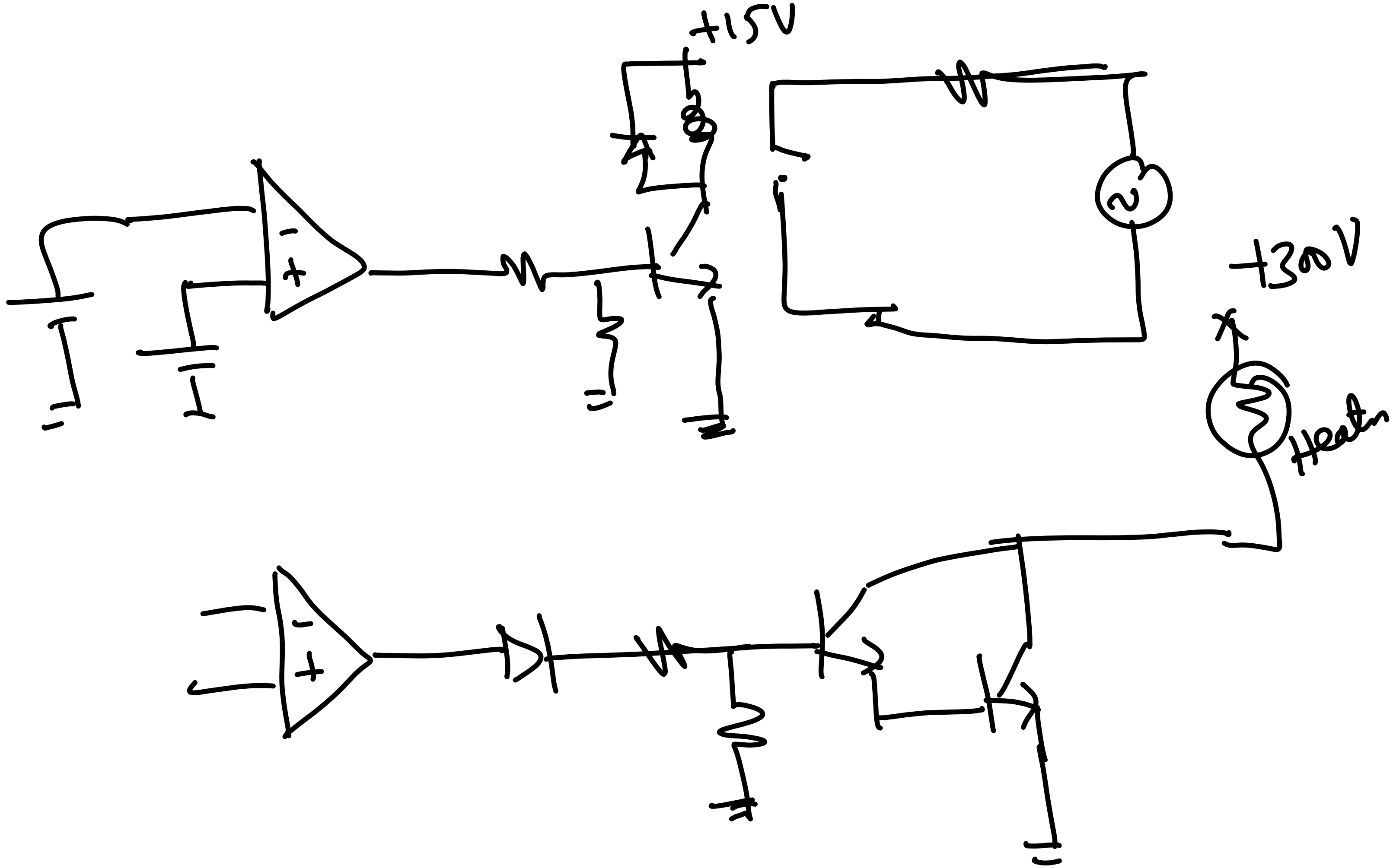


+15V
4.7k
5V
5V
Vz
-15V
+15
+15V
1k
5V = Vz
5V = Vz
1k
R
1k
Vz
Vz
I = V2 / R
PRT

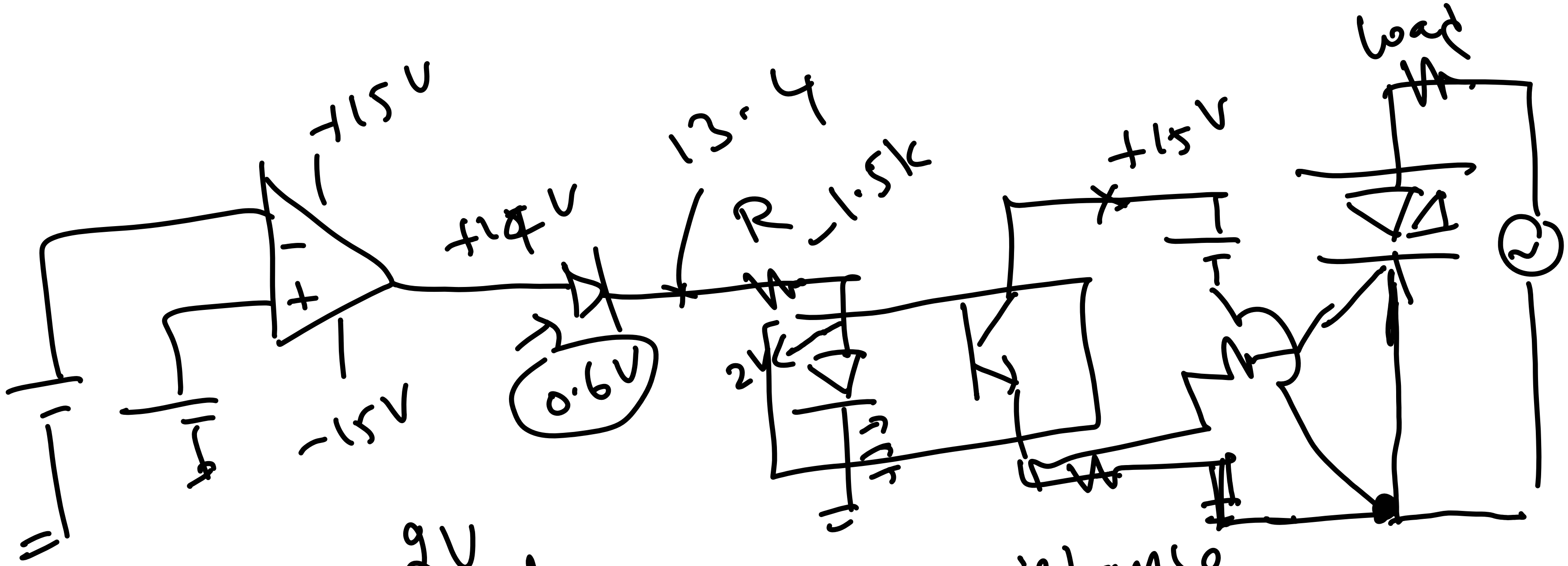
Heater Drive

ON/OFF





+15V
+300V
Heater



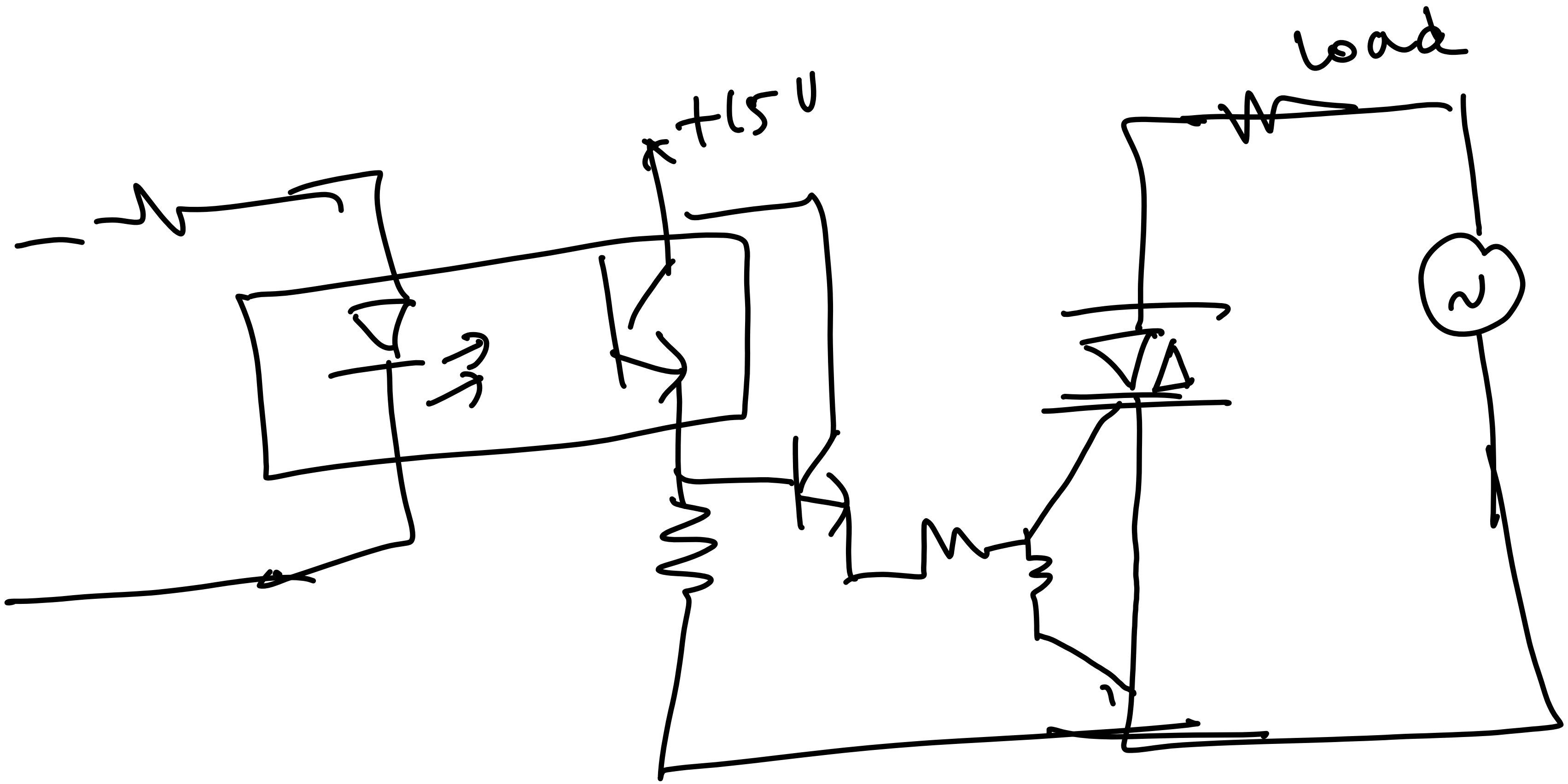
13.4 − 2V

= vol acc the resistance

11.4 = vol acc the ress

For 8mA current

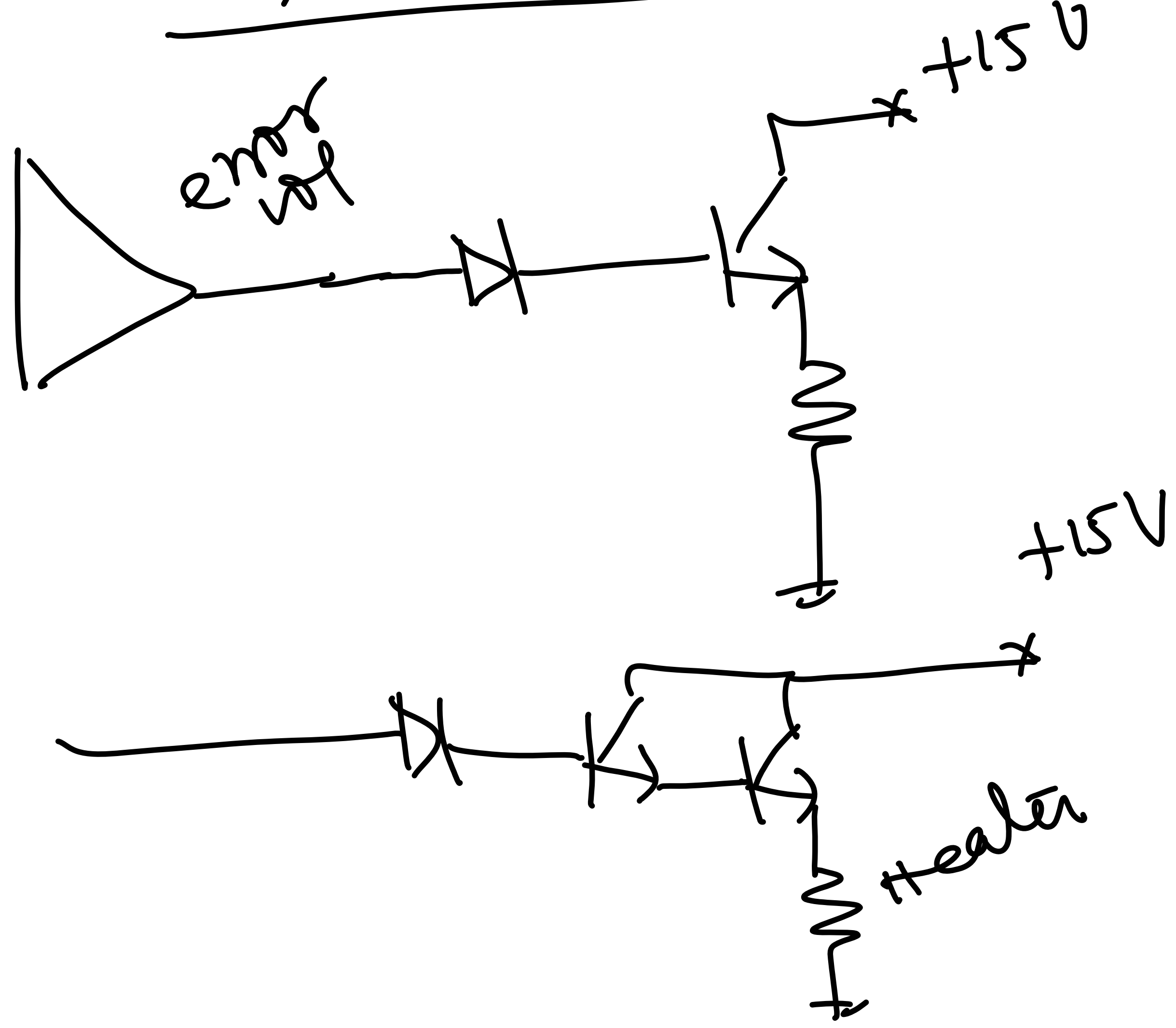
$$R = \frac{11.4\times10^{3}}{8} = 1.5K$$

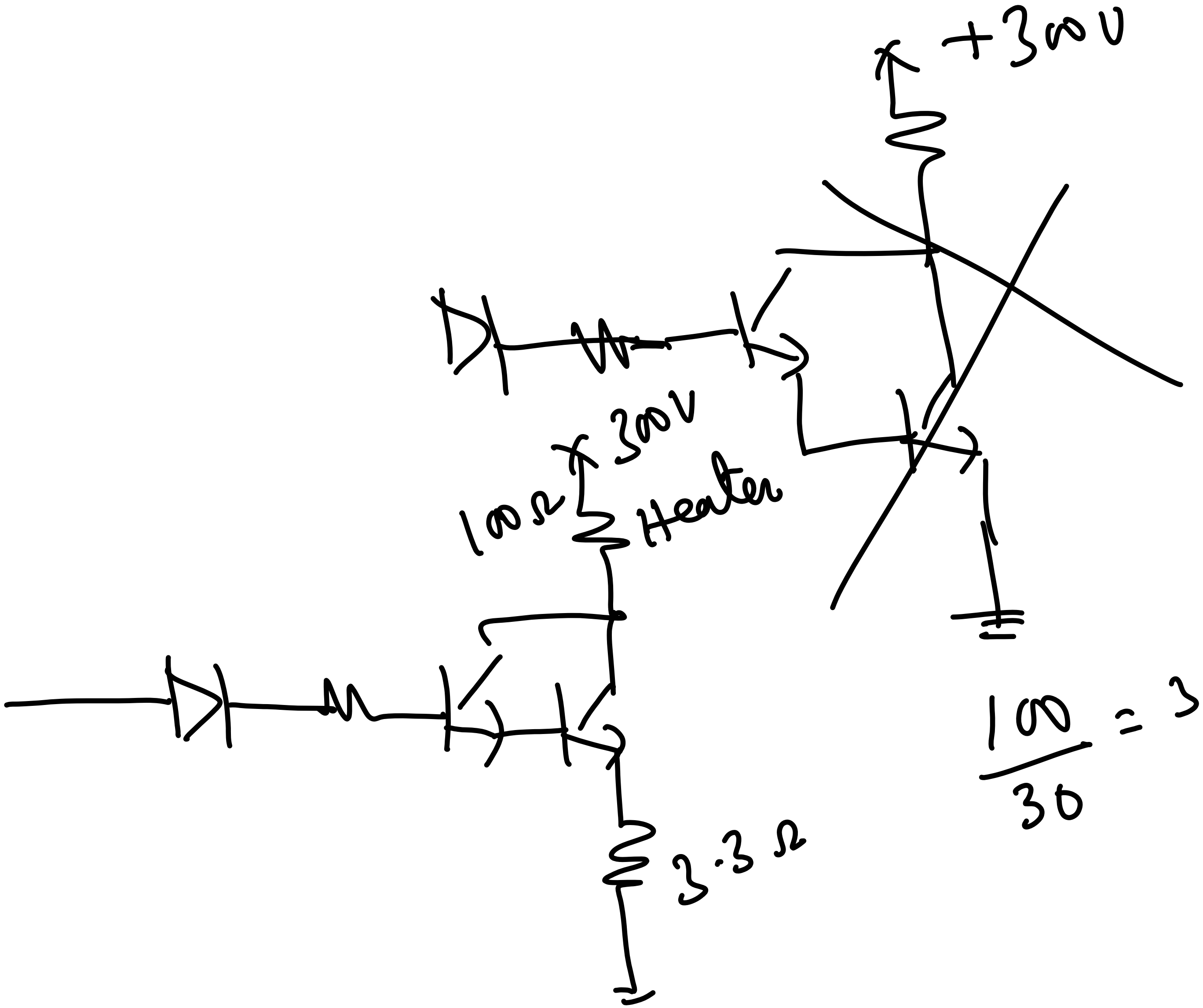


+15 V
load

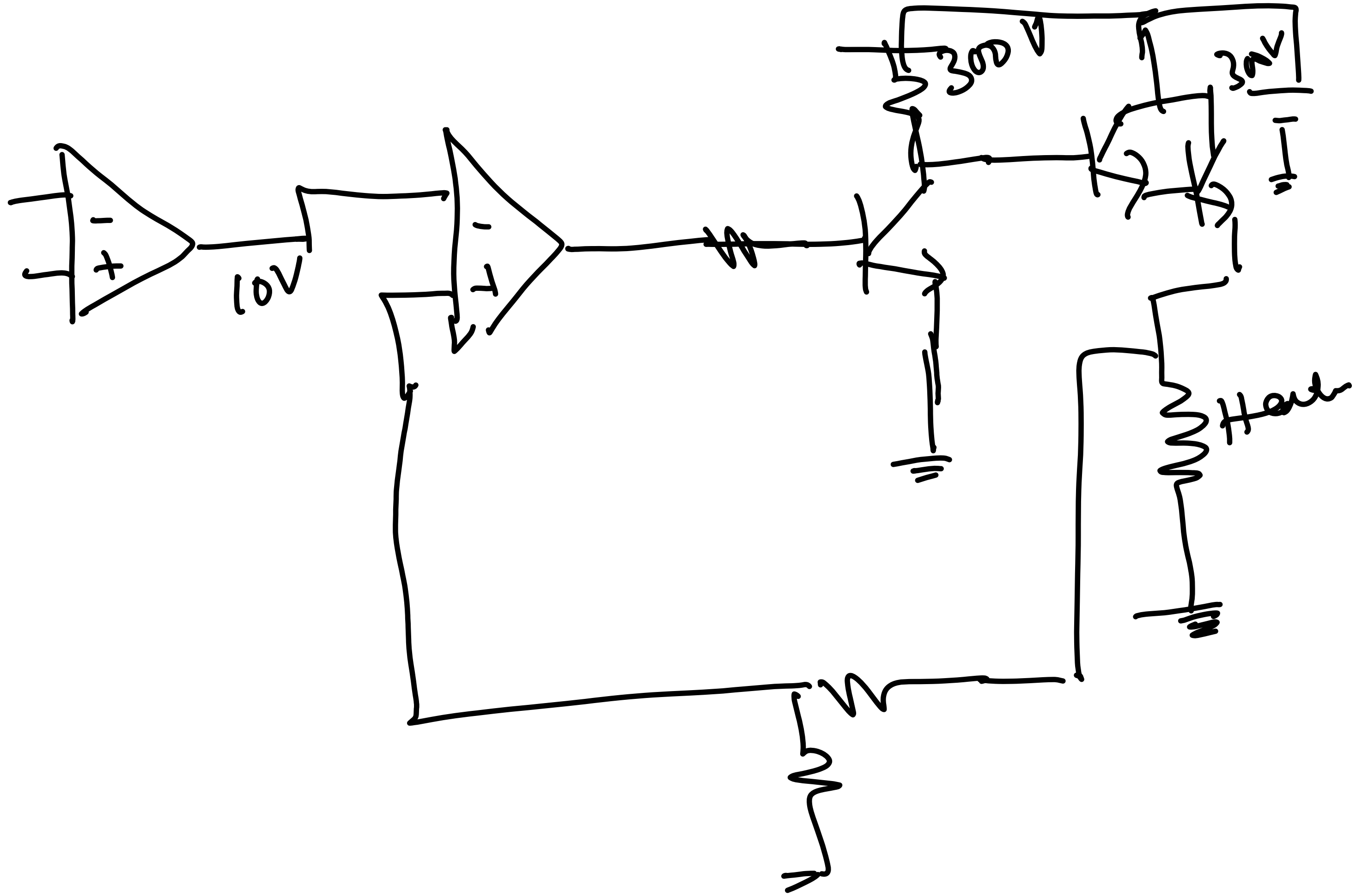
Heater drive for

P, PI, PD, PID



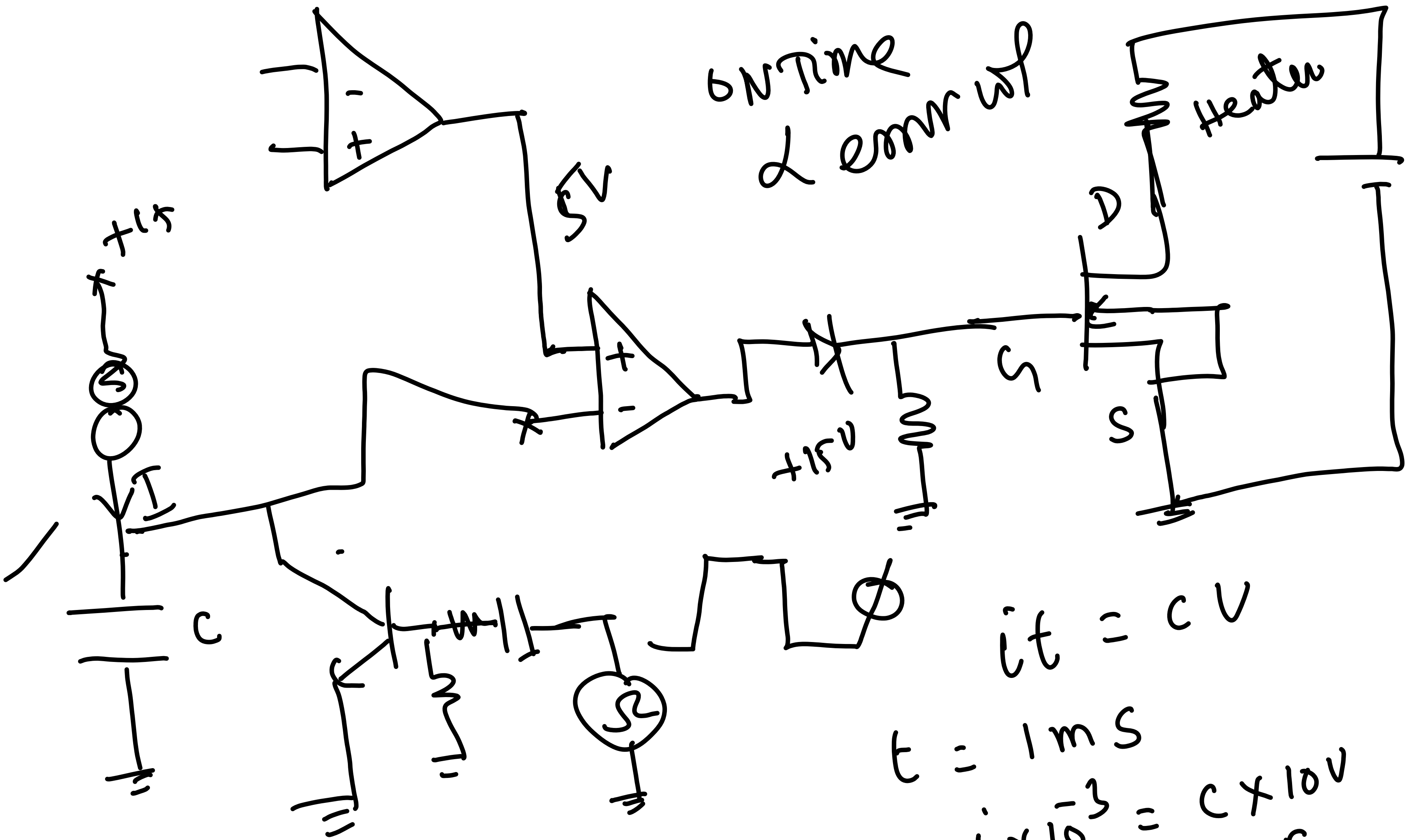


$$\frac{100}{30} = 3.3\Omega$$



10V
300V
30V
Heat

PWM Technique



$it = CV$

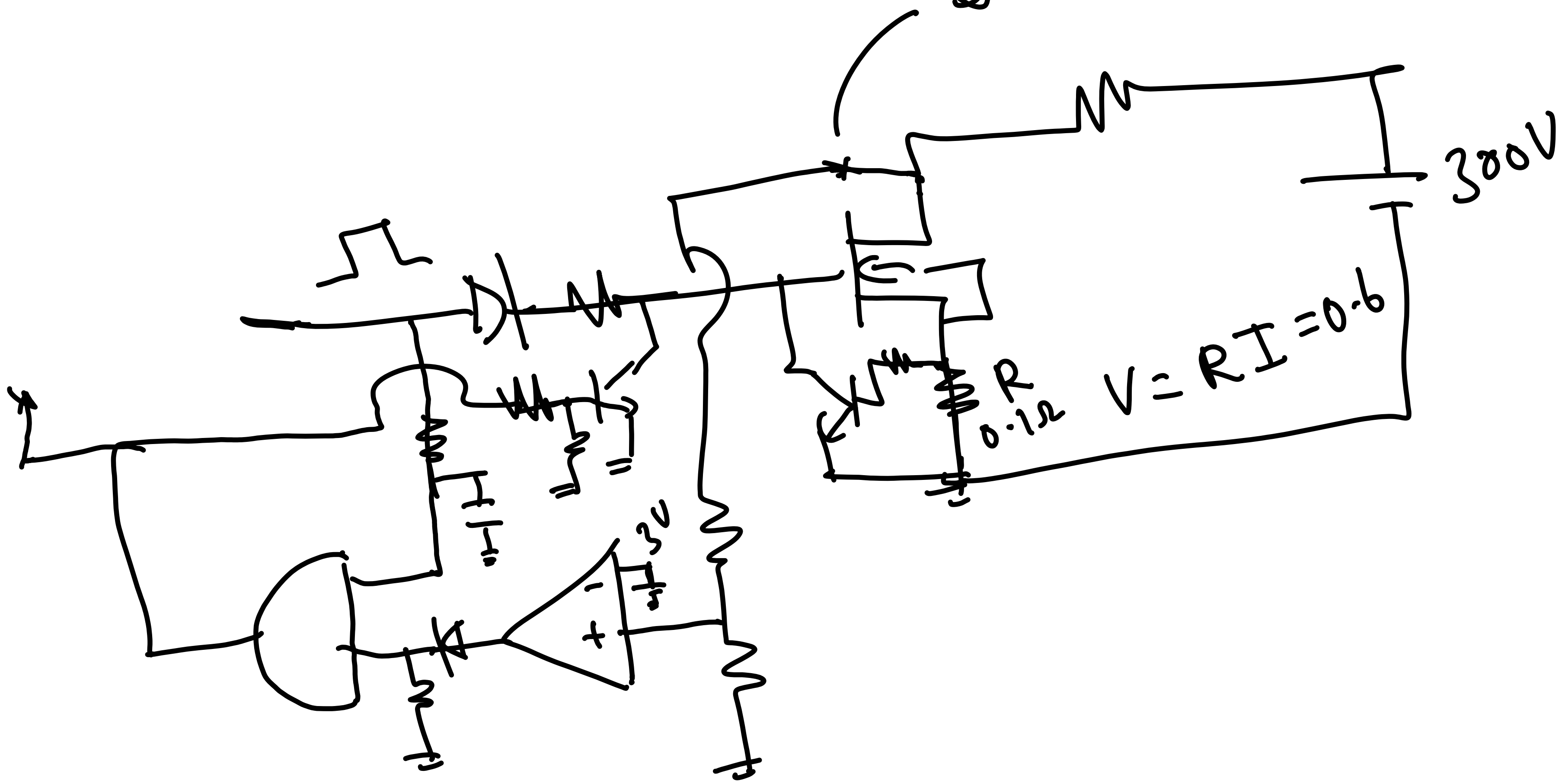
$t = 1\,ms$

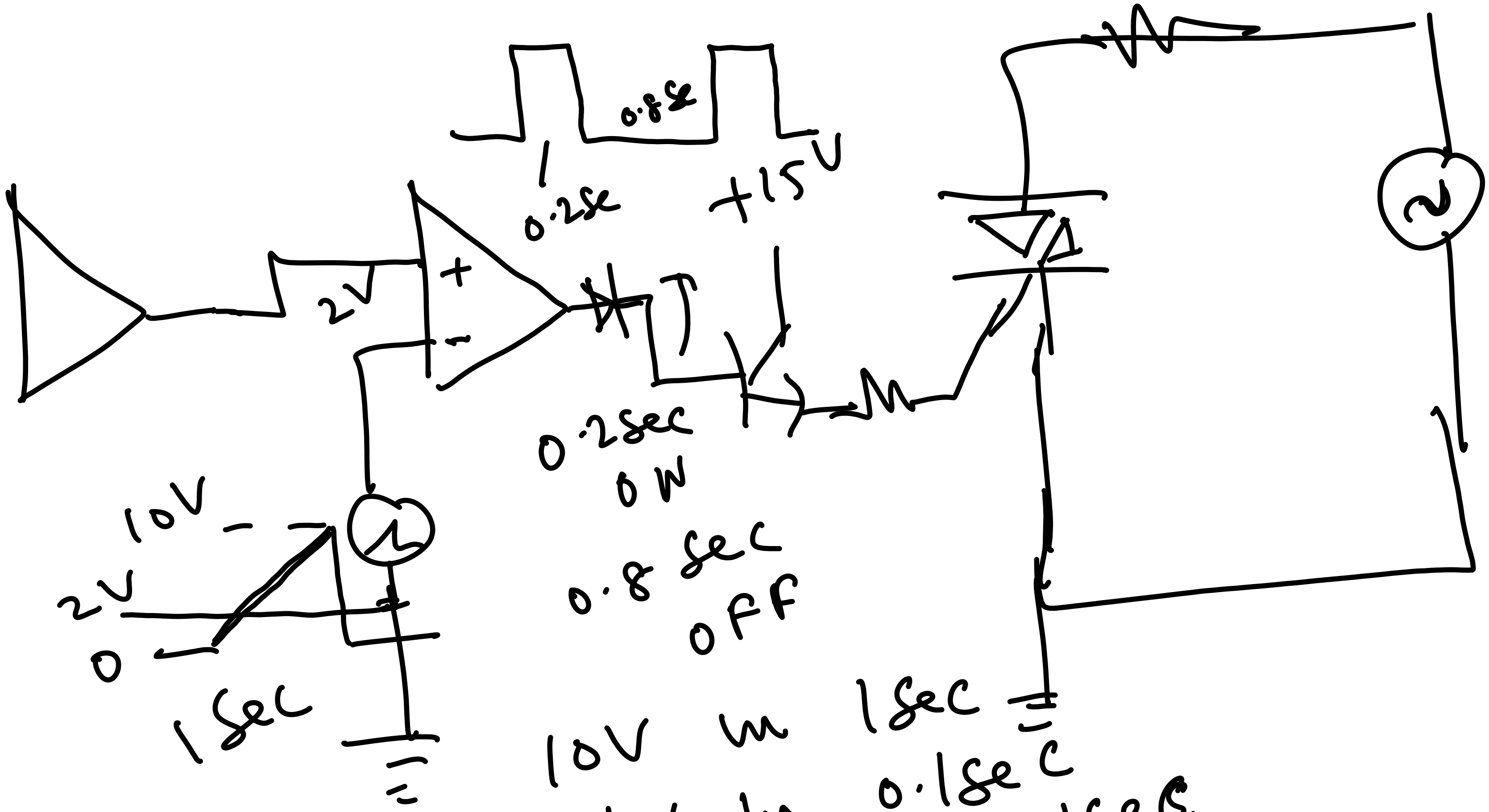
$i \times 10^{-3} = C \times 10V$

$C = 1\mu F$

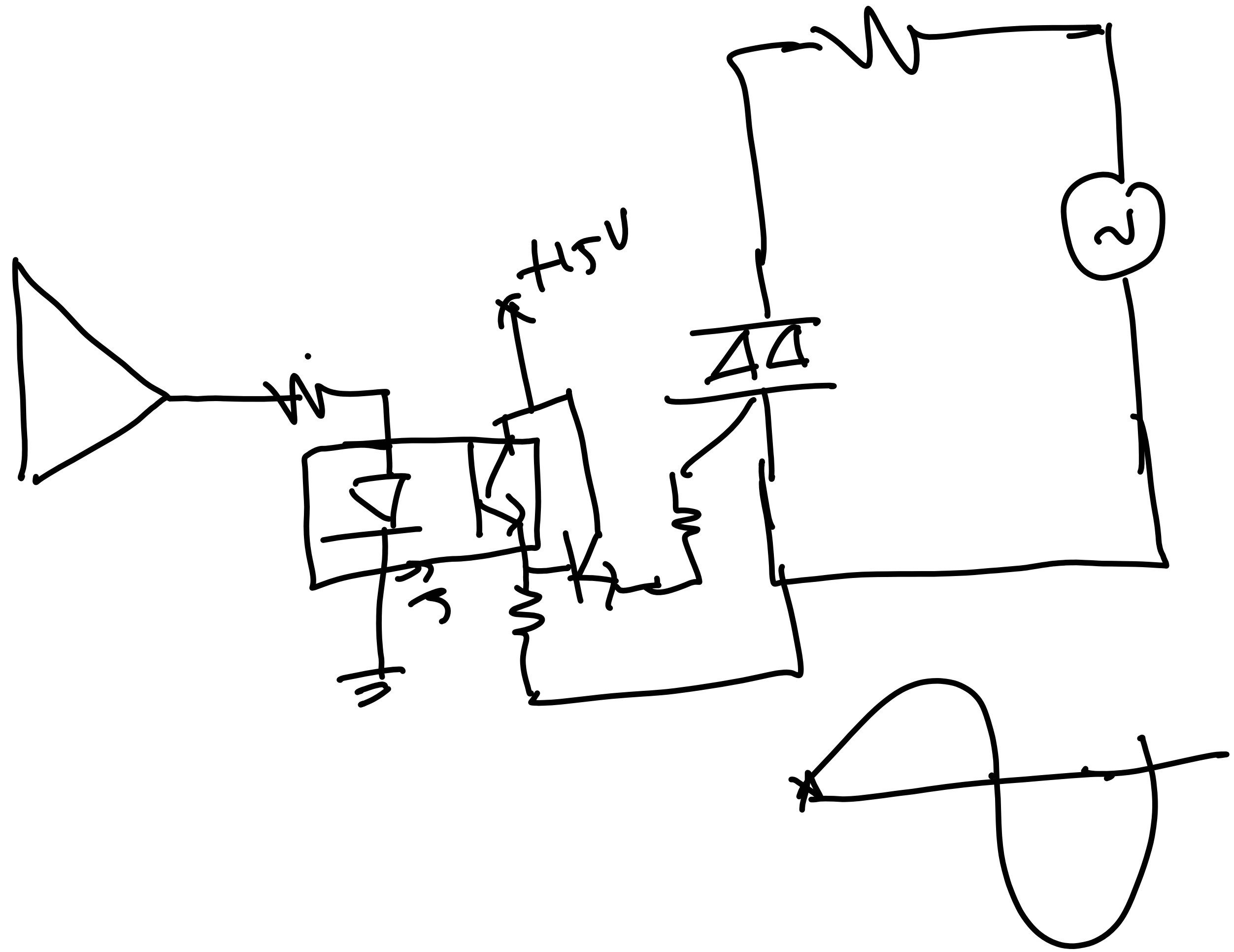
Short circuit protection for MOSFET

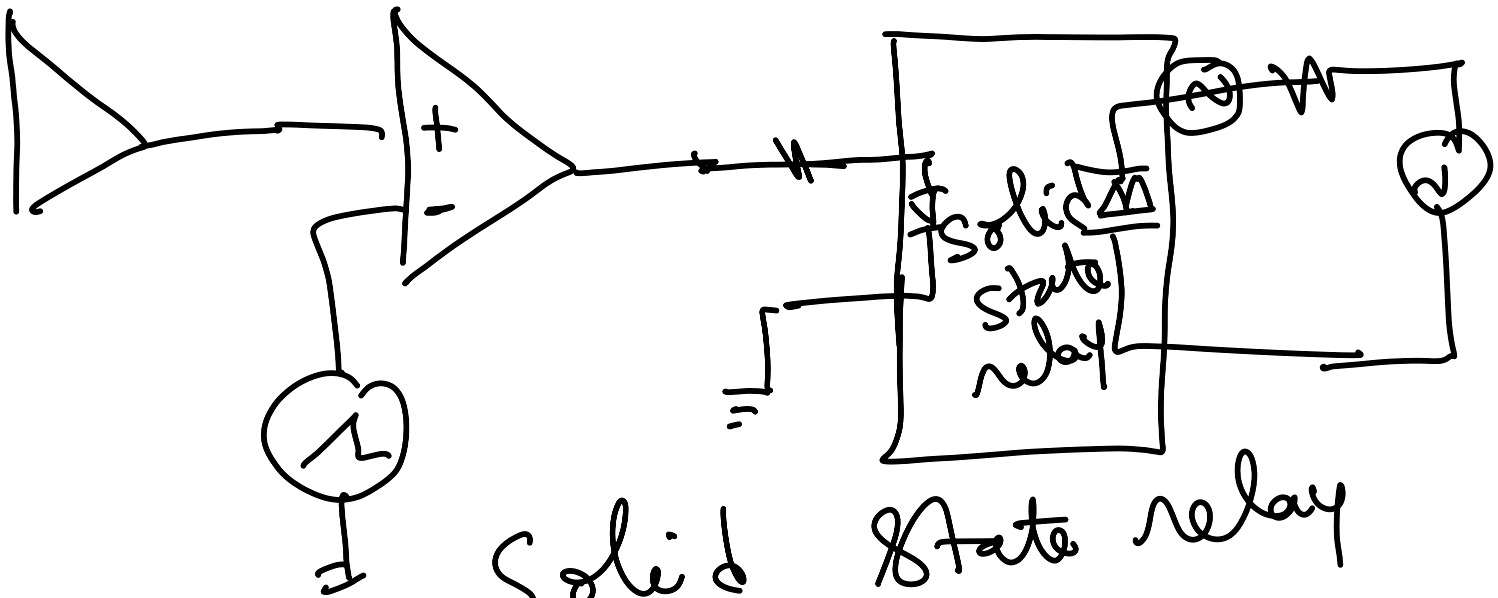
6A x 300V = 1800W





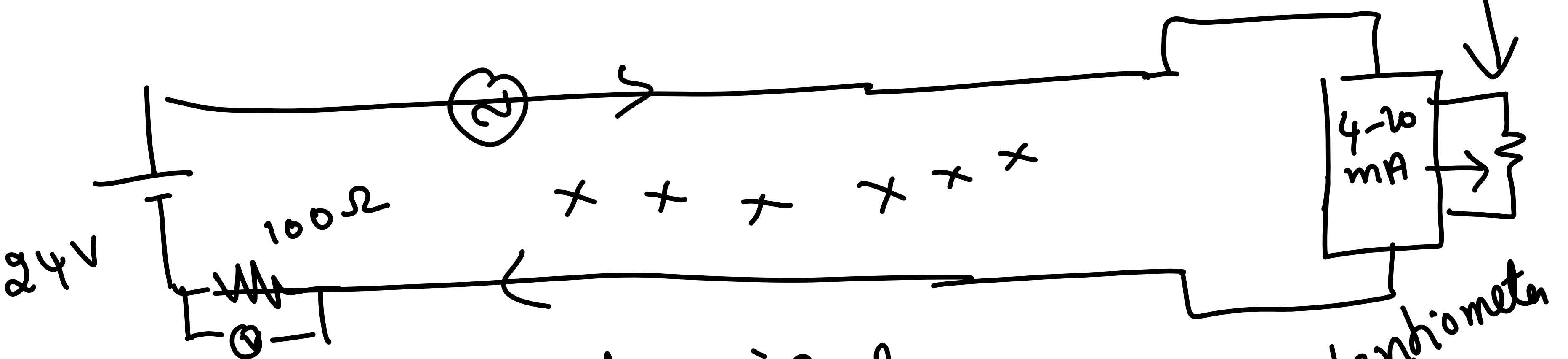
10V in 1sec
1V in 0.1sec
1V/0.1sec


+5V



Solid State relay
with zero crossing

4 - 20 mA current transmitters



When mechanical system at one end = 4 mA

At the other end I should be 20 mA

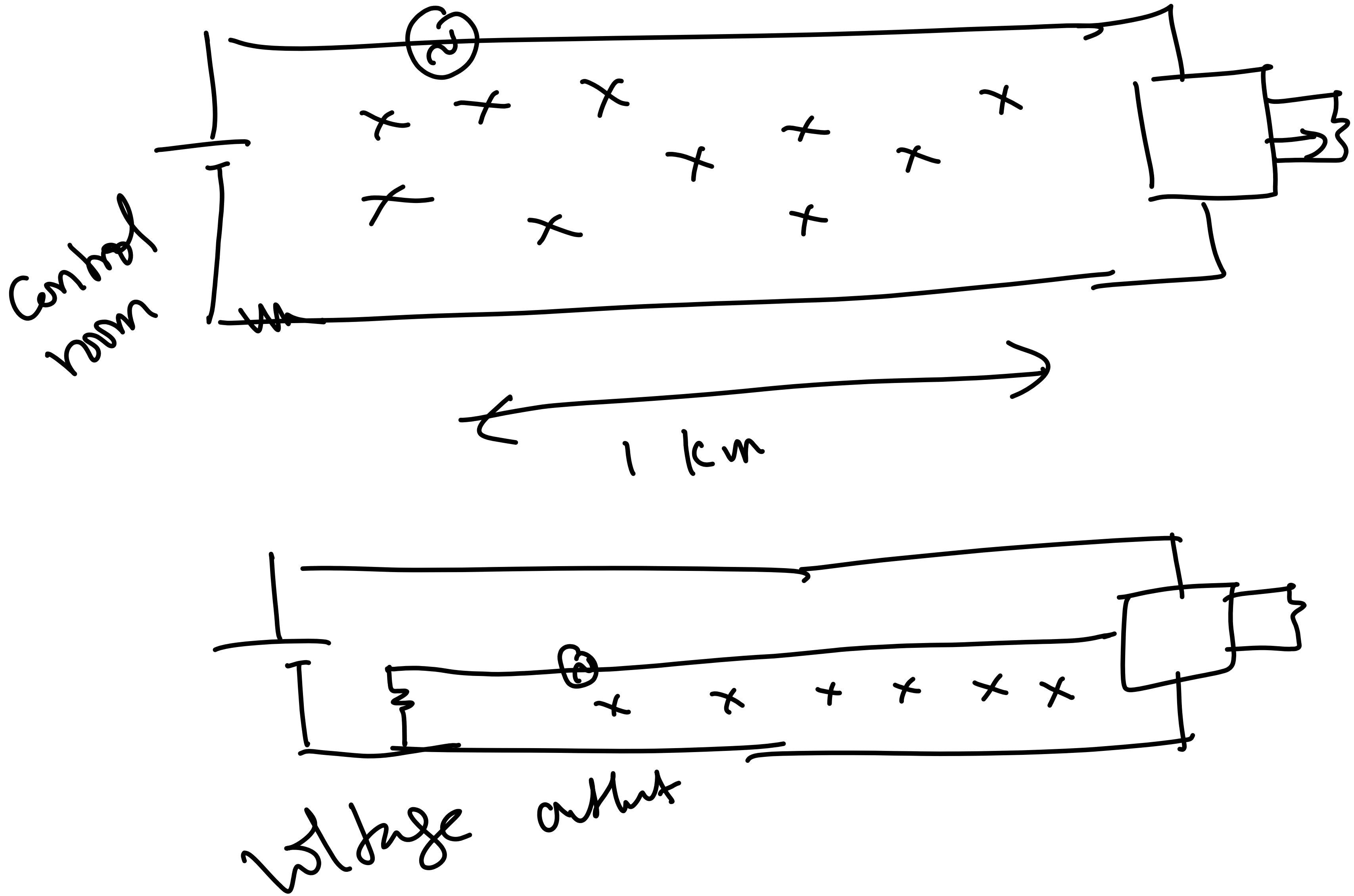
0% → 4 mA

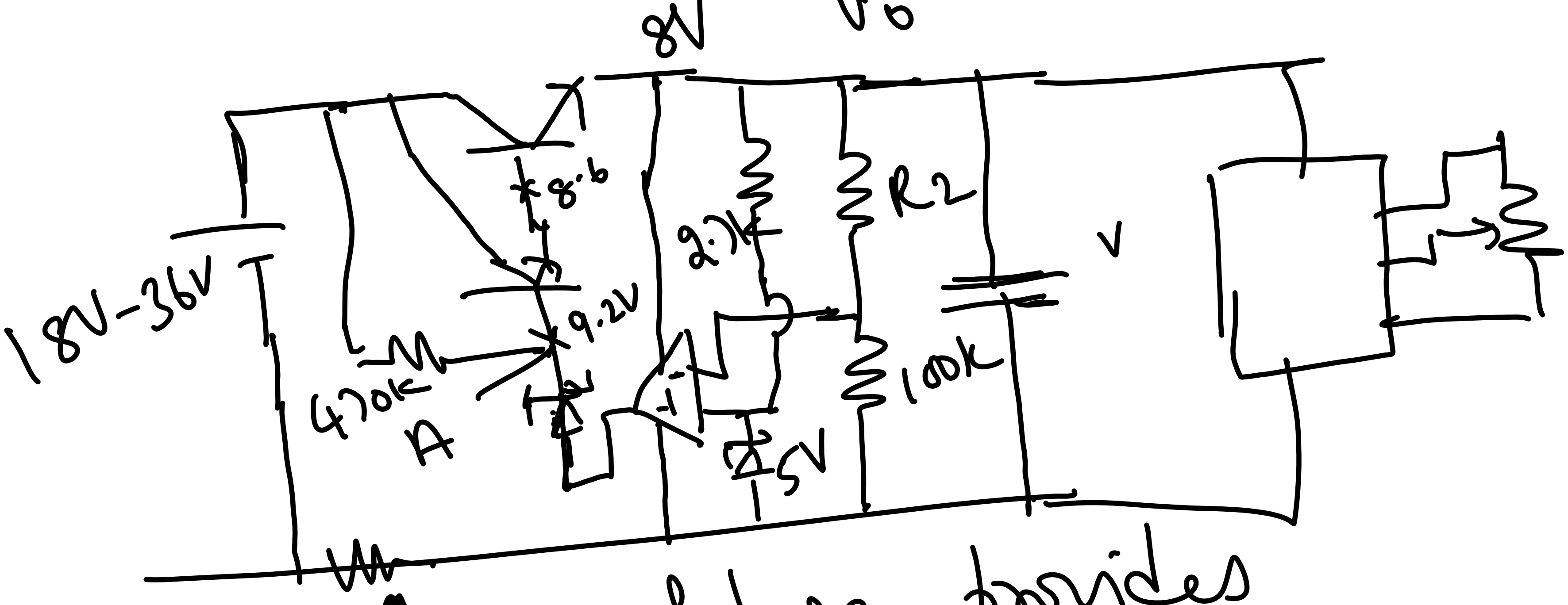
25% → 8 mA

50% = 12 mA

75% 16 mA

100% 20 mA


Control room
1 km
Voltage output



zenar voltage provides
extra voltage ie
vol at A = OP amp output
+ vol acc the zenr

vol acc the zenr ≠ breakdown
vol of the
zenr

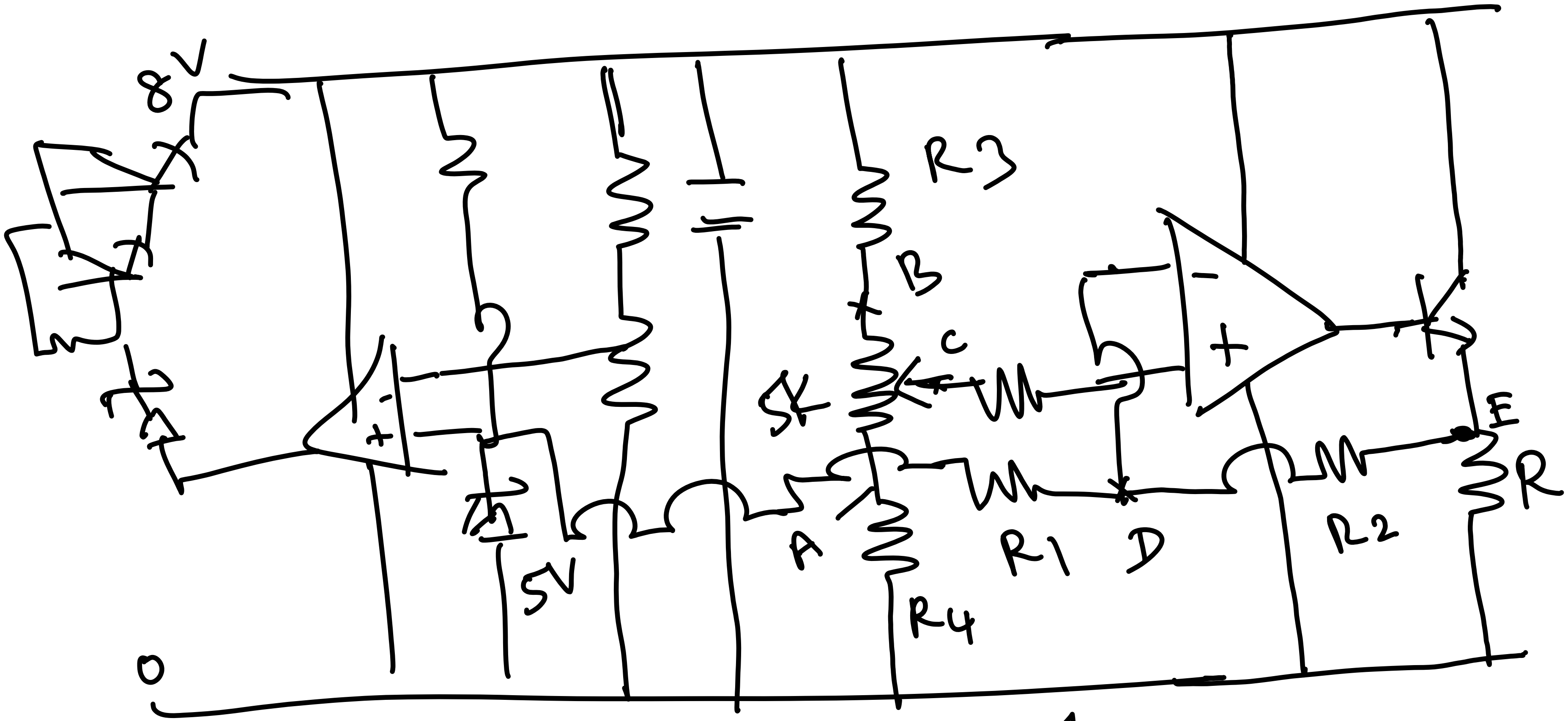
Vo acc 100k = 5V

So output Vo = $\frac{5V \times (100k + x)}{100k}$

= 8

x can be calculated

Design of a current transmitter



Select R_4, R_3, pot

such that vol at A = 2V

vol at B = 4V

vol acc the potentiometer

becomes 2V

The expected change at D = 2V

= change at C

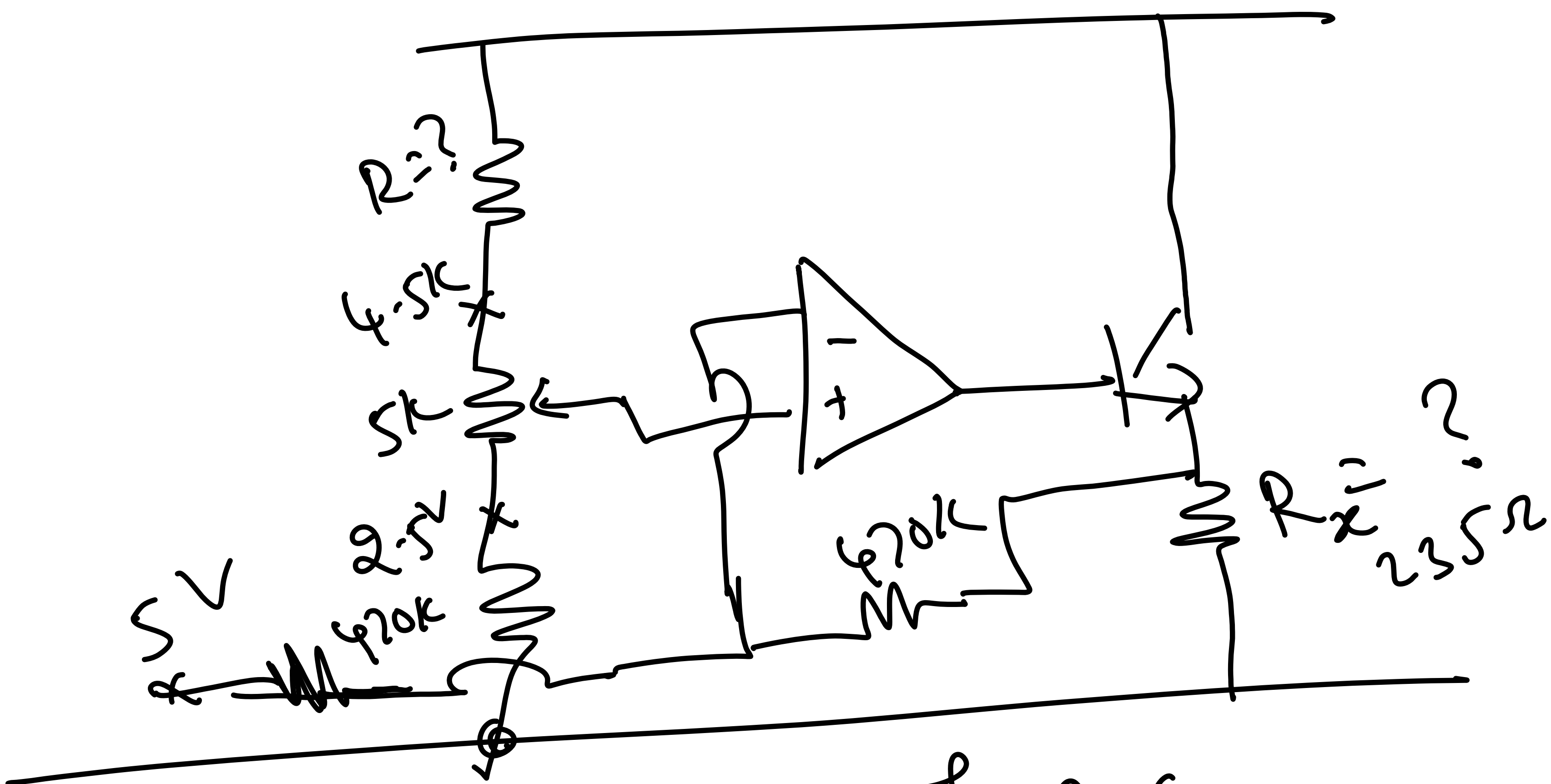
If $R_1 \simeq R_2$

Keep A at 2.5V

Then $R_1 = R_2 = 470K$

B at 4.5 V

Potentiometer V_f = 2V



Required w/ acc

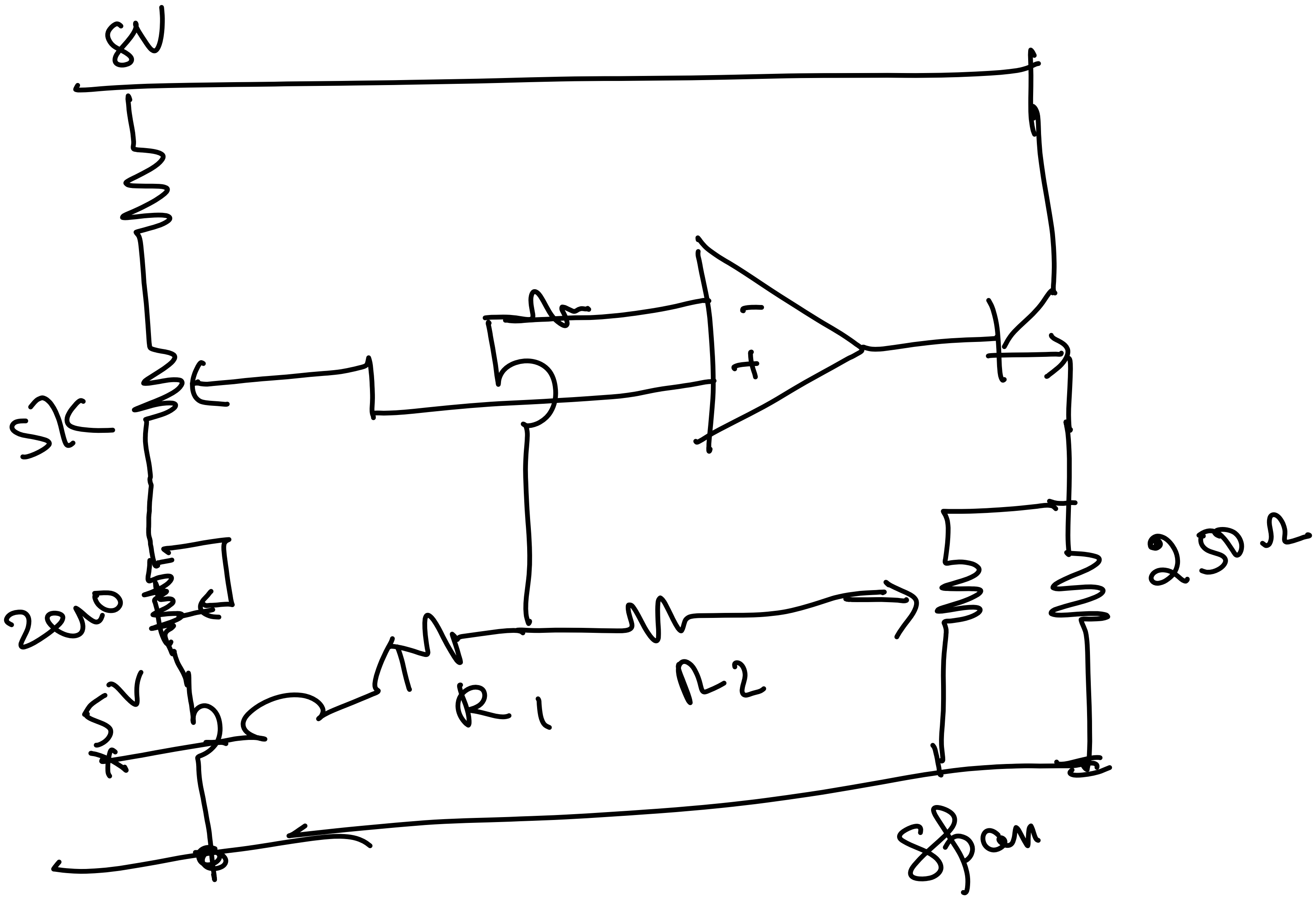
Rx = 4V at 17 mA

Required w/ acc Rx ≃ 0 V at 1 mA or less

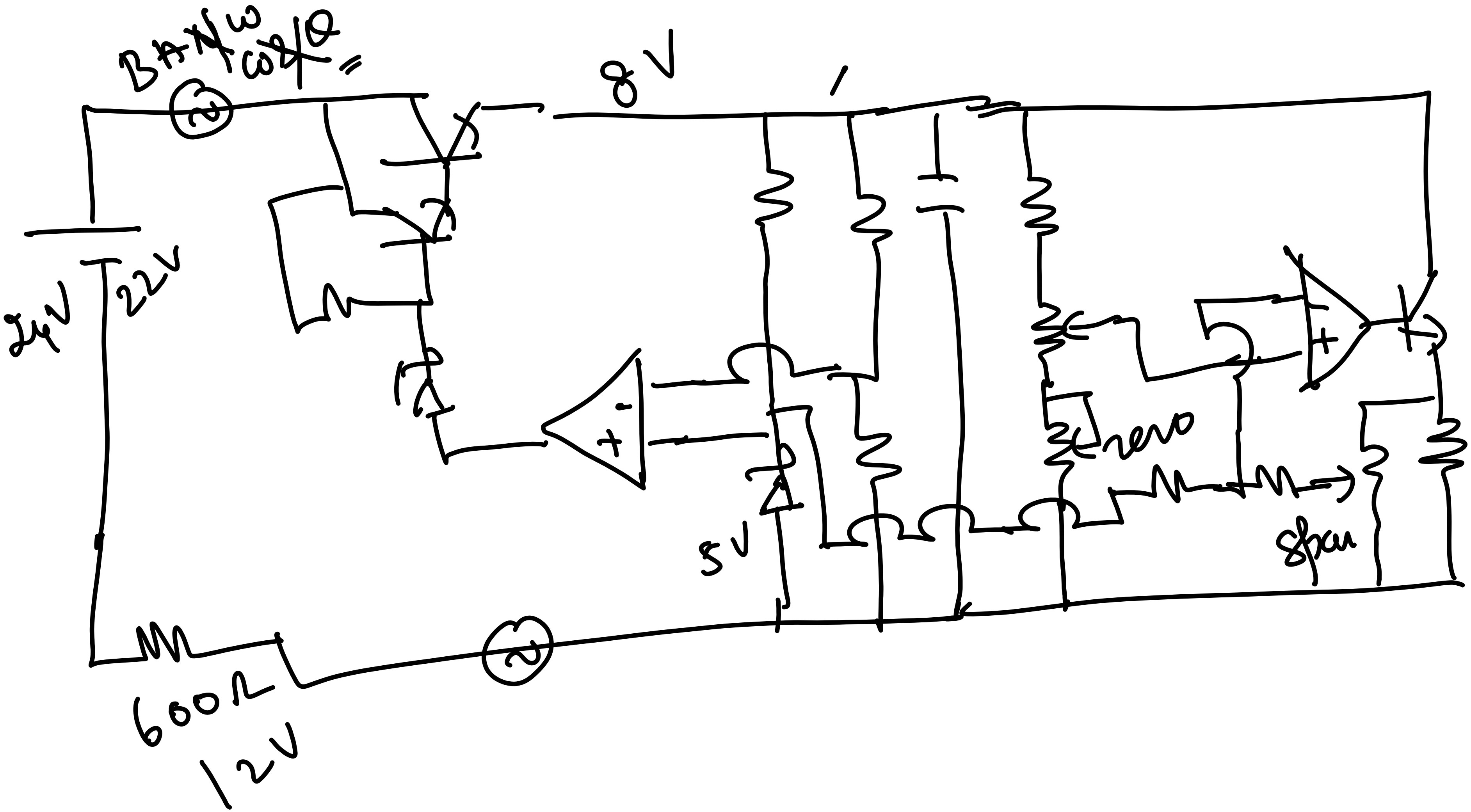
$R_x \times 17\,mA = 4V$

$$R_x = \frac{4V}{17} = \frac{4000}{17} = \underline{\underline{235}}\,\Omega$$

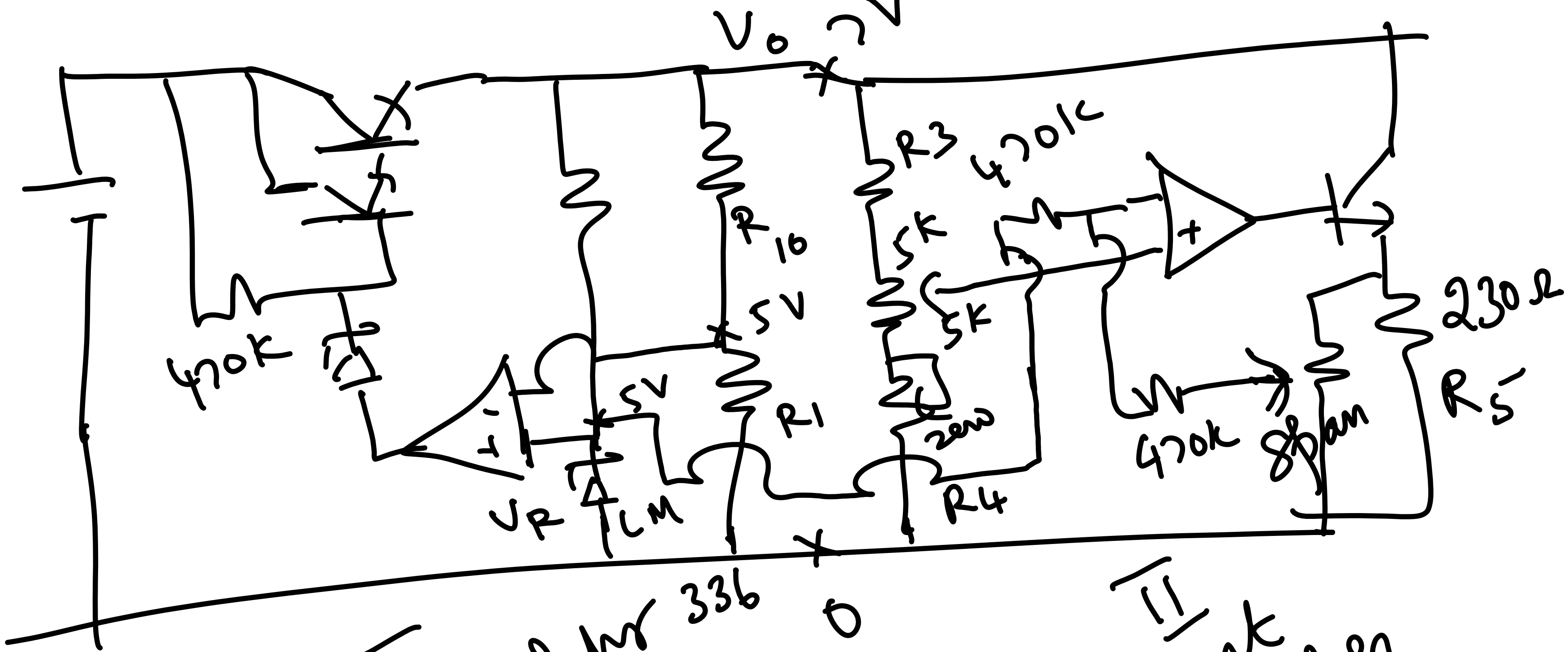
60



8V
5K
zero
5V
R1
R2
250 Ω
Span


8 V
24V
22V
5V
zero
span
600Ω
12V

Error budgeting for 4-20mA current Transmitter



I volt regulator

II current transmitter

LM336 Zener 5V 30PPM|C

Drift = 30 ppm|·C

Assuming ambient temp change = 100°C

wt of the zener initially be = 5V

ΔV of the zener

$= \frac{5 \times 30 \times 100}{10^6}$

$V \times PPM \times \Delta C$

$\Delta V = \frac{15 \times 10^3}{10^6} = 15 \times 10^{-3}$ V

$= \pm 15$ mV

ΔV of the zener $= \pm 15\,mV$

So regulated output voltage

$$\text{change} = \frac{15\,mV \times (R_1 + R_{10})}{R_1}$$

$$7V = 5 \times \frac{(R_1 + R_{10})}{R_1}$$

$$\frac{7V}{5V} = \frac{(R_1 + R_{10})}{R_1}$$

$$\Delta V_0 = 15 \times 10^{-3} \times \frac{7}{5} = 21\,mV$$

Vol acc $R_3 = R_4 = 2.5V$

For 5k the vol

drop = 2 V

So $R_3 = \frac{5k \times 2.5}{2} = \frac{12.5}{2} = 6.25k$

$R_3 = r_4 = 6.25k$

vol at the potentiometer

top end = 2.5 + 2.0 = 4.5 V

(This corresponds to current of

20 mA)

For ΔV_o of 21 mV the corresponding

voltage change at the top

end of the potentiometer

$$= \frac{21 \times 4.5}{7.0} \text{ mV}$$

$$= 3 \times 4.5 \text{ mV} = \underline{13.5 \text{ mV}}$$

Assuming the span pot at the top end the o/p change at, at inverting input = 13.5 mV = o/p change at the non-inverting input

i.e. at -ve input $\Delta V = 13.5$ mV
So o/p change acc R5
$= 2 \times 13.5 = 27$ mV

wl charge acc R_5

$= 27 mV$

So current charge $= \dfrac{27 \times mV}{230}$

$= \dfrac{27}{230} = \pm 0.08\ mA$

II <u>Calculate the current drift due to R_1, R_{10} charge</u>

$R_1 = 100K$

$R_{10} = \dfrac{5 \times (R_1 + R_{10})}{R_1} = 7V$

$= \dfrac{5 (10^5 + R_{10})}{10^5} = 7V$

$$7 = \frac{5 \times (10^5 + R_{10})}{10^5}$$

$$7 \times 10^5 = 5 \times 10^5 + 5 \times R_{10}$$

$$\frac{2 \times 10^5}{5} = R_{10}$$

$$R_{10} = \frac{200 \times 10^3}{5} = 40\text{K}$$

Assume R_1, R_{10} drifts by 50 ppm/°C

Assume R_1 is decreasing and

R_{10} is increasing with temp

$$\Delta R_1 \text{ change} = -\frac{100 \times 10^{3} \times 50 \times 100}{10^{6}}$$

$$= \frac{5 \times 10^{8}}{10^{6}} = -500\,\Omega$$

$$\Delta R_{10} = \frac{40 \times 10^{3} \times 50 \times 100}{10^{6}}$$

$$= 20 \times \frac{10^{7}}{10^{6}} = 200\,\Omega$$

R_1 decreases by $500\,\Omega$

R_{10} increases by $200\,\Omega$

So change in regulated voltage $= \dfrac{5V\,(R_1 + R_{10})}{R_1}$

$$\text{New } V_0 = \frac{5\,(100\times10^3 - 500 + 40200)}{99500}$$

$$= \frac{5\,(99500 + 40200)}{99500}$$

$$\begin{array}{r} 99500 \\ 40200 \\ \hline 139700 \end{array}$$

$$= 5\times\frac{1397\cancel{00}}{995\cancel{00}} = \frac{1397\times\cancel{5}}{\underset{199}{\cancel{995}}} =$$

New $V_0 = \dfrac{1397}{199} = \dfrac{1397}{200} = \underline{\underline{6.985\ V}}$

Change in $V_0 = 7 - 6.985$

$= 15\ mV$

So the expected current change in $20\ mA = 0.08\ mA$

III <u>Error due to V_{offset} of</u>

<u>w/ regulator stage</u>

Let $V_{offset} = 15\,\mu V/^\circ C$

For $100^\circ C$ total offset voltage

change $= 15 \times 100 = 1.5\,mV$

For $1.5\,mV$ at the i/p of

the regulator op amp the

expected change at the output

of the regulator $= 1.5 \times \frac{7}{5}\,mV$

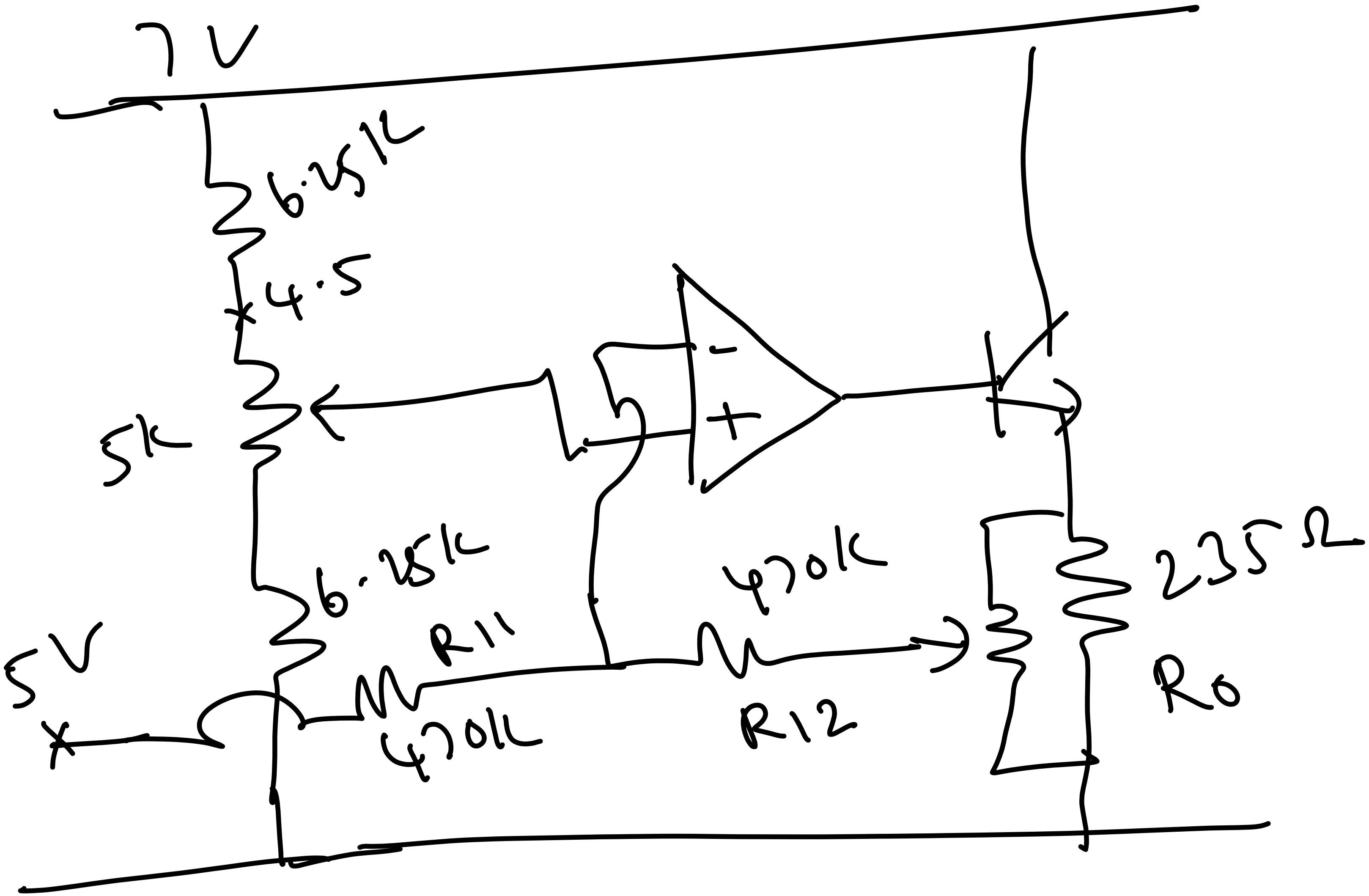
$= 2.1\,mV$

So current charge in the 20mA output

$$= \frac{0.08 \times 2.1}{15}$$

$$= \frac{0.16}{15} = 0.01 \text{ mA}$$

Error in the current converter stage


7V
6.25k
4.5
5k
6.25k
R11
470k
5V
470k
R12
235 Ω
Ro

(1) Error due to offset voltage drift of current converter op amp

$V_{offset} = 15 \mu V / °C$

Total offset voltage

Charge $= \pm 1.5 \, mV$

$(15 \mu V \times 100)$

So the expected charge in the non-inverting input $= 1.5 \, mV$

= Expected change

in the inverting input = 1.5mV

So the expected change

accross R_0 = $\frac{1.5 \times 470 + 470k}{470}$

= 3 mV

So the expected current change

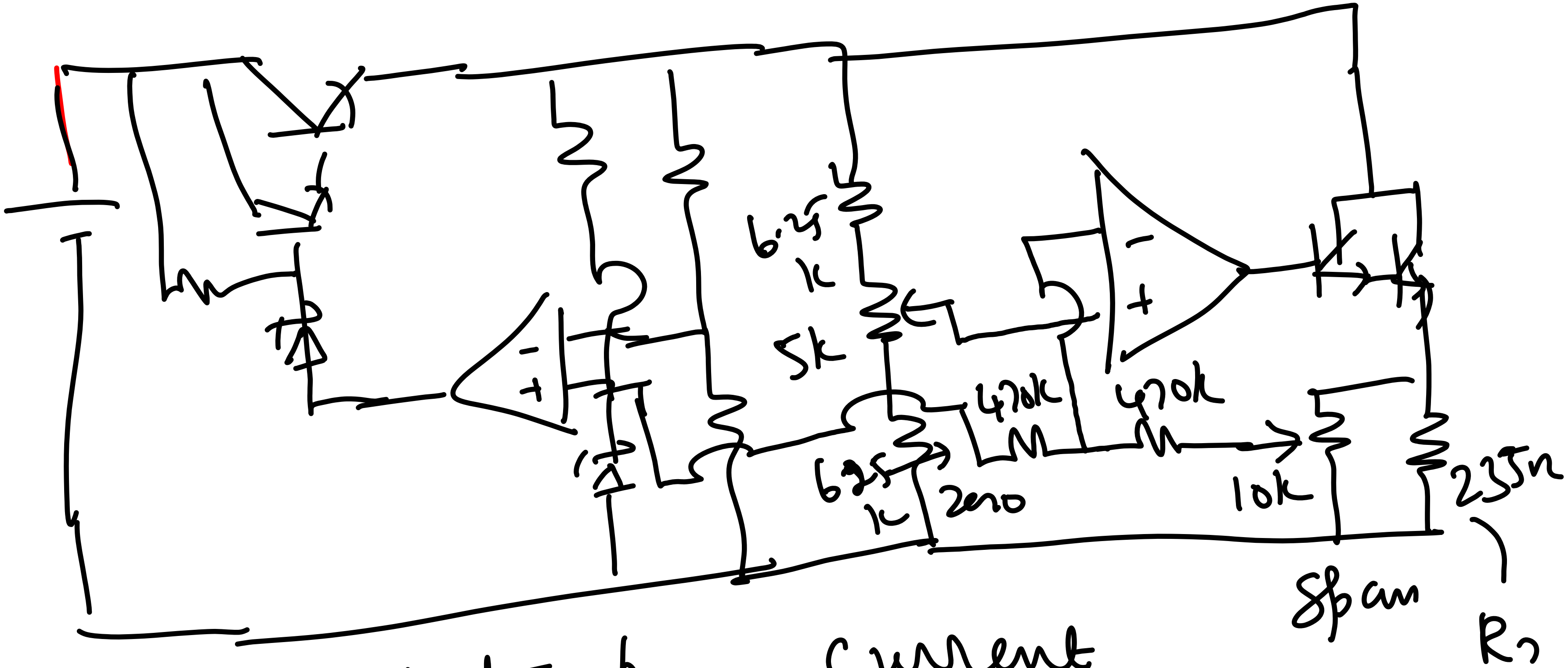
in 20 mA = $\frac{3 \times 10^{-3}}{230}$ = $\frac{3000}{230}$ µA

= 12 µA

(2) Change in current due to zener vol change

(3) Change in current due to R_{11}, R_{12} change

(4) Change in current due to 235 Ω resistance change



Calculate the current drift due to drift in R_7

Let the TC of

$R_7 = 50\,ppm/°C$

For $\Delta T = 100°C$

$$\Delta R = \frac{235 \times 50 \times 100}{10^6}$$

$$= 235 \times 5 \times 10^{-3}$$

$$= 1.175\ \Omega$$

Assuming 16 mA current in R_7

Then for $1.175\ \Omega$ change the

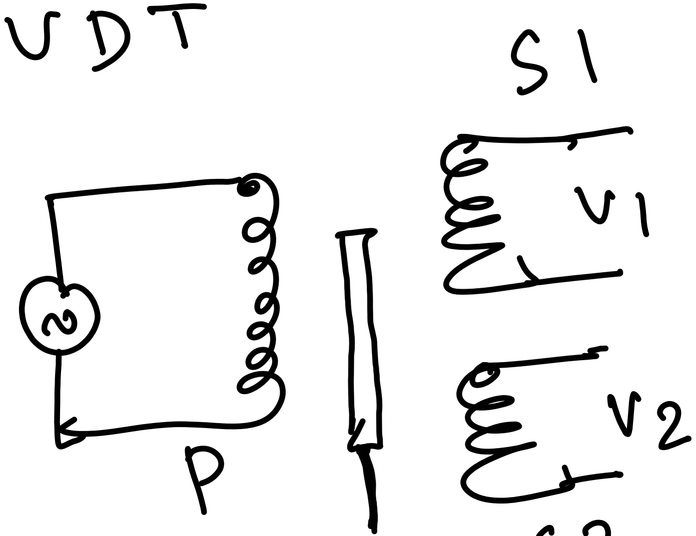
expected current change

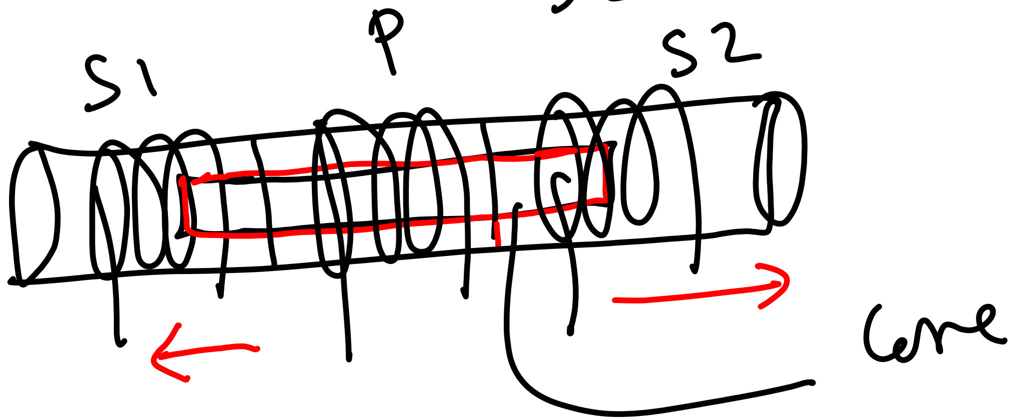
$$= \frac{16 \times 1.1}{235} = \frac{17.6}{235} \text{ mA}$$

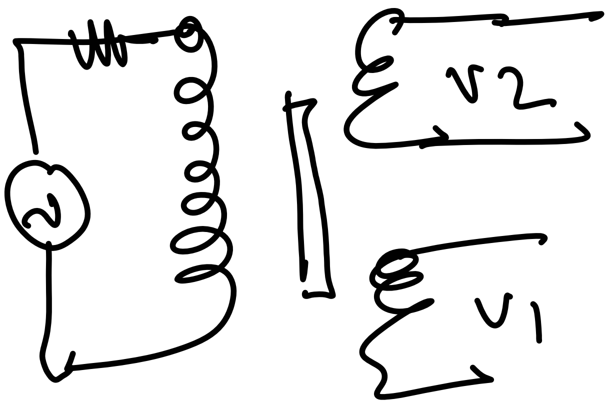
$$= \frac{1}{15} \text{ mA} = 0.06 \text{ mA}$$

Design of LVDT based position transmitter

① LVDT







$|V_2| - |V_1|$

= displace

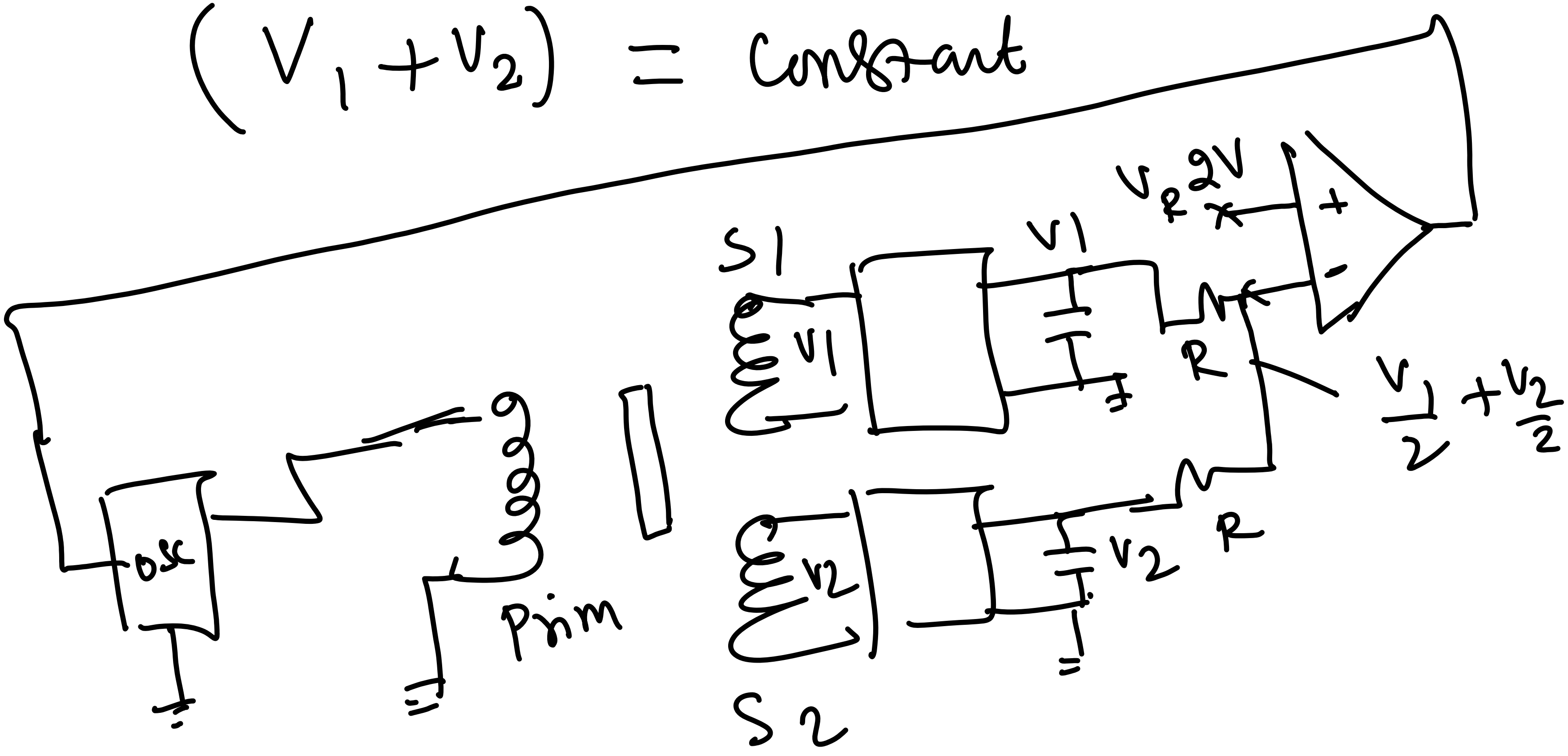
(1) Convert V_1 and V_2 into DC

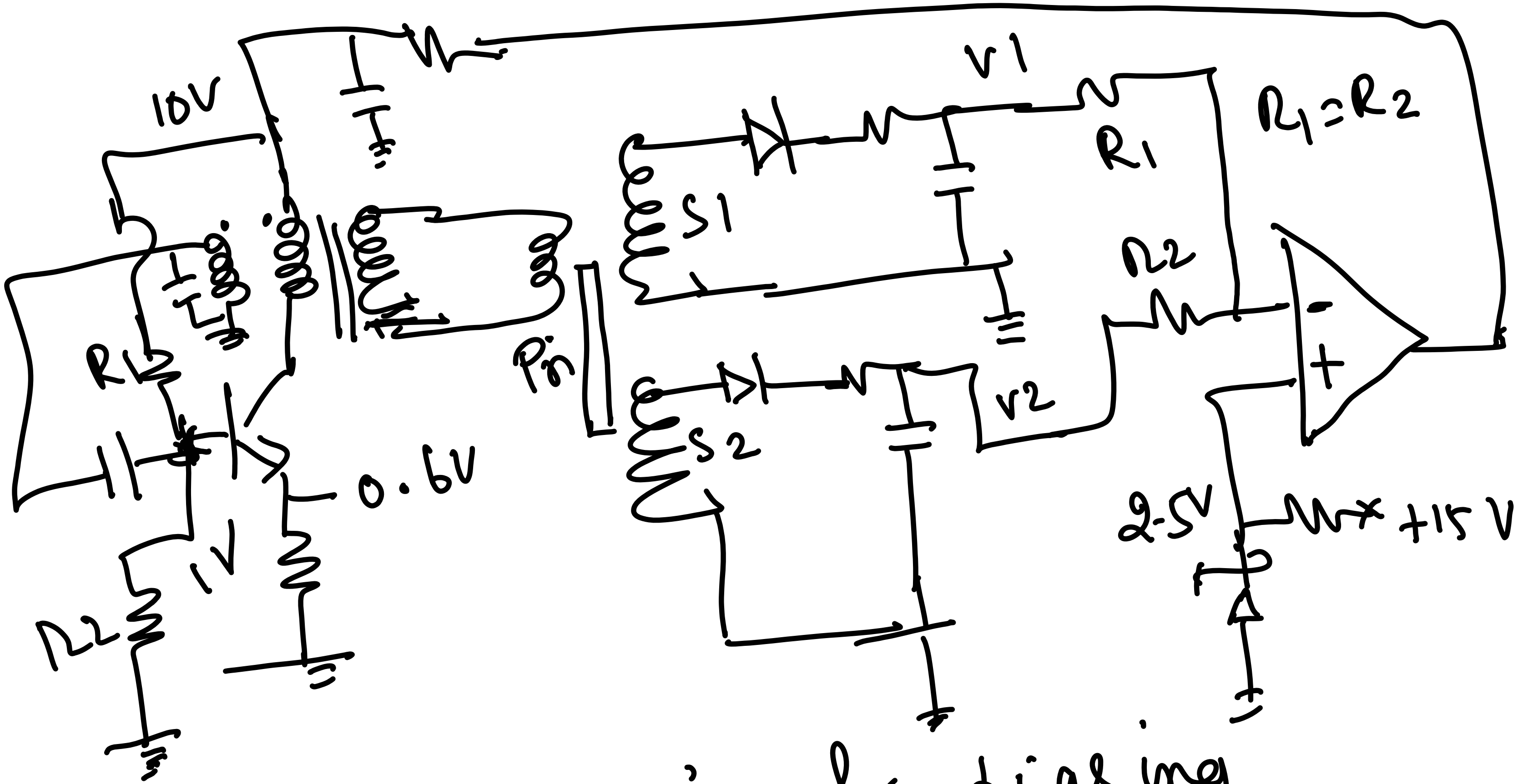
This dc w/ difference alone give you the displacement

(2) $|V_2 - V_1|$ changes with temp

How to make $(V_2 - V_1)$ independent of temp?

$$(V_1 + V_2) = \text{constant}$$





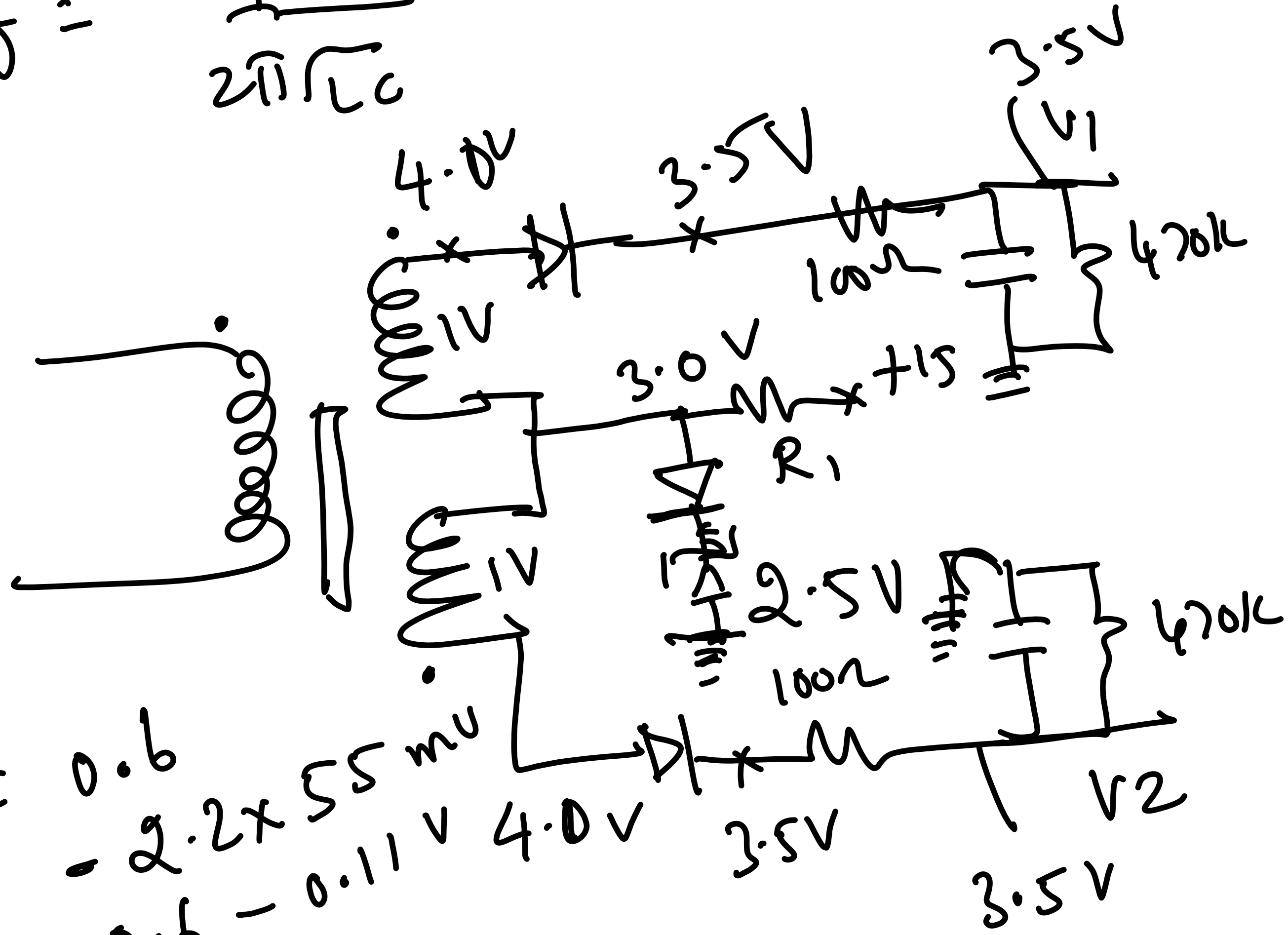
R_1, R_2 is for biasing

① The emitter current increases

② This produces increased mag field in the coil

(3) This increases base vol of the
Transistor
(4) ~~This~~ Increase in base
vol again increases the
emitter current.
(5) Again mag field increases
in the primary

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



wl acc
he diode $= 0.6$

$- 2.2 \times 55$ mV

$= 0.6 - 0.11$ V

$= 0.49$V

=

At room temp

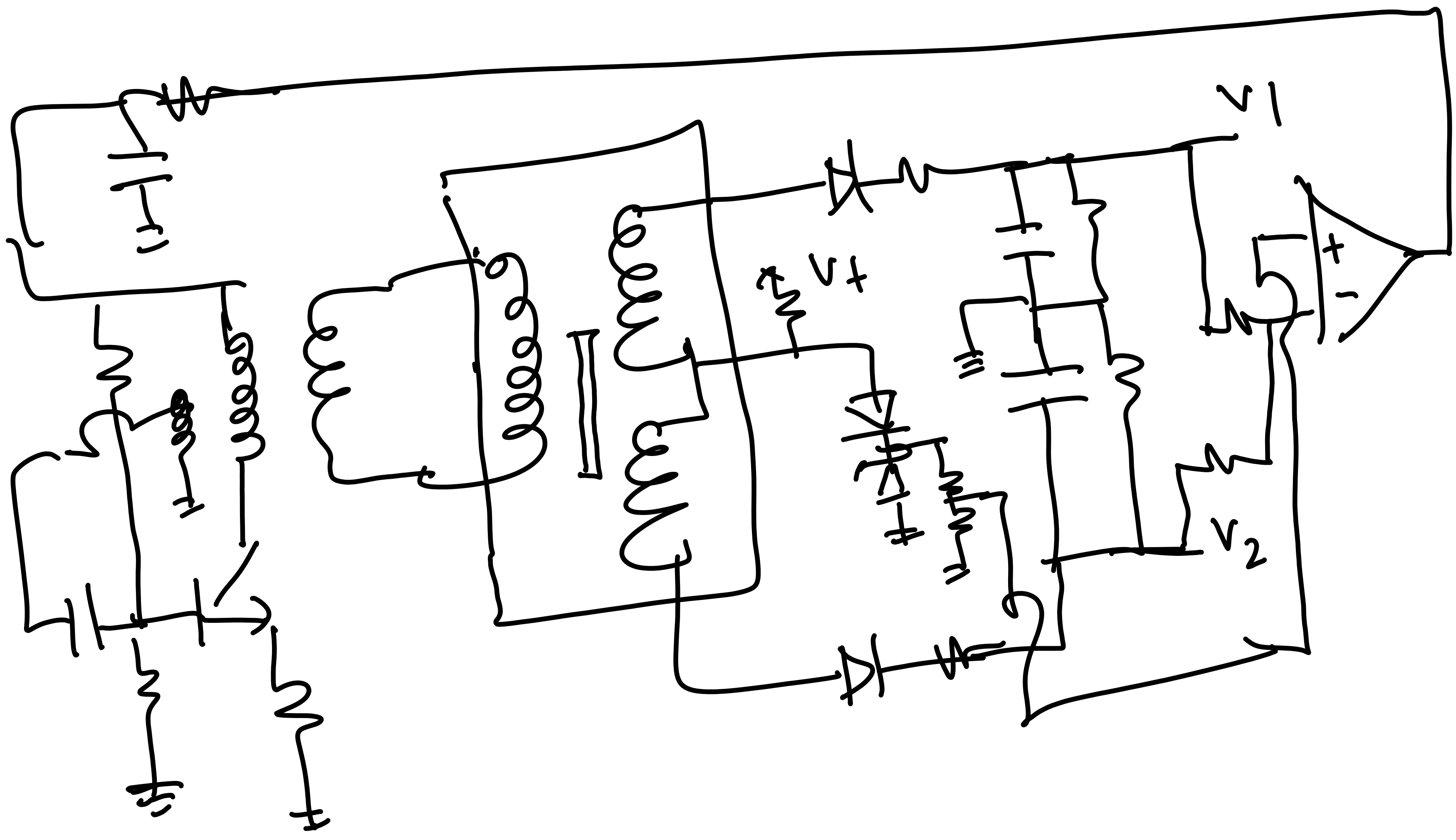
$V_1 = 3.5V$

$V_2 = 3.5V$

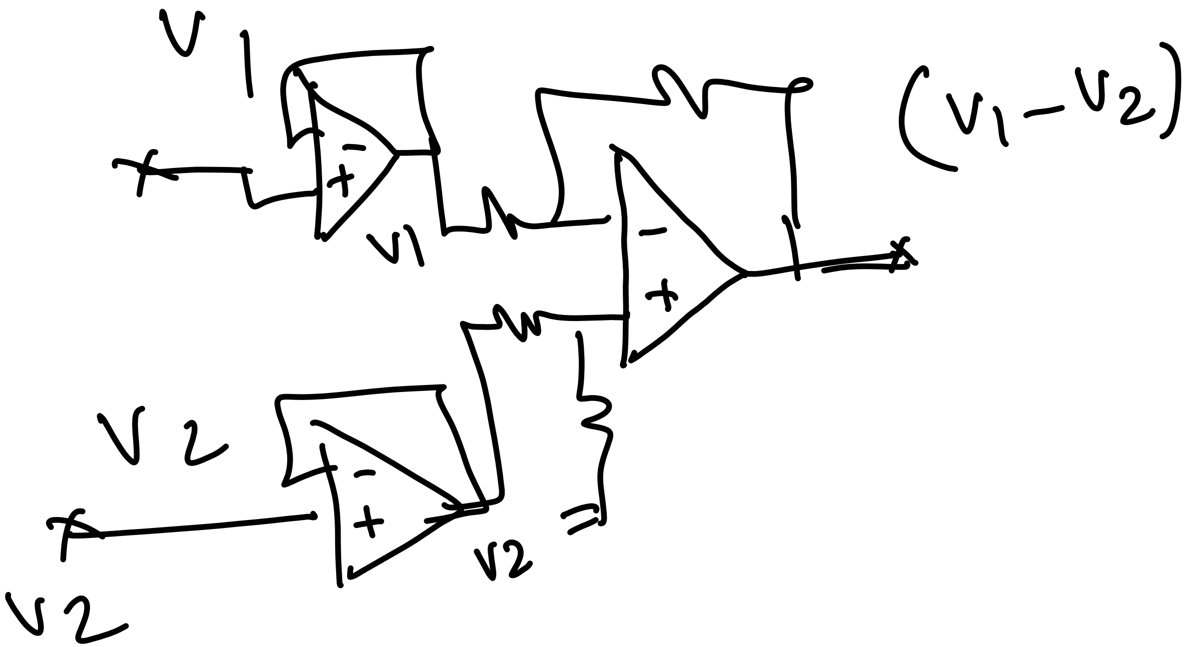
At 80°C ambient temp

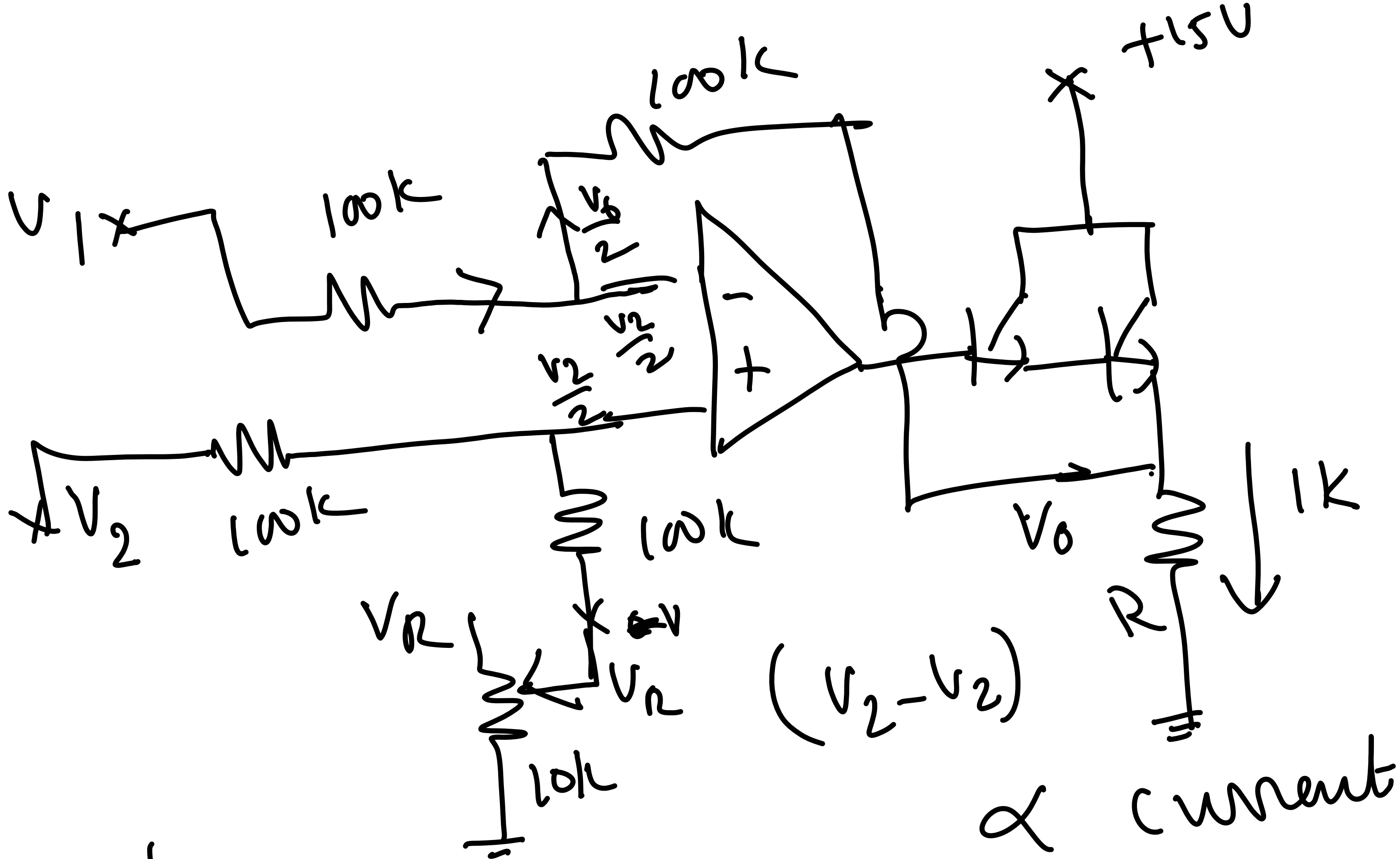
$V_1 = ? = 3.5 V$

$V_2 = ? = 3.5$



V1
V+
V2


V 1
v1
V 2
v2 =
v2
(v1 - v2)

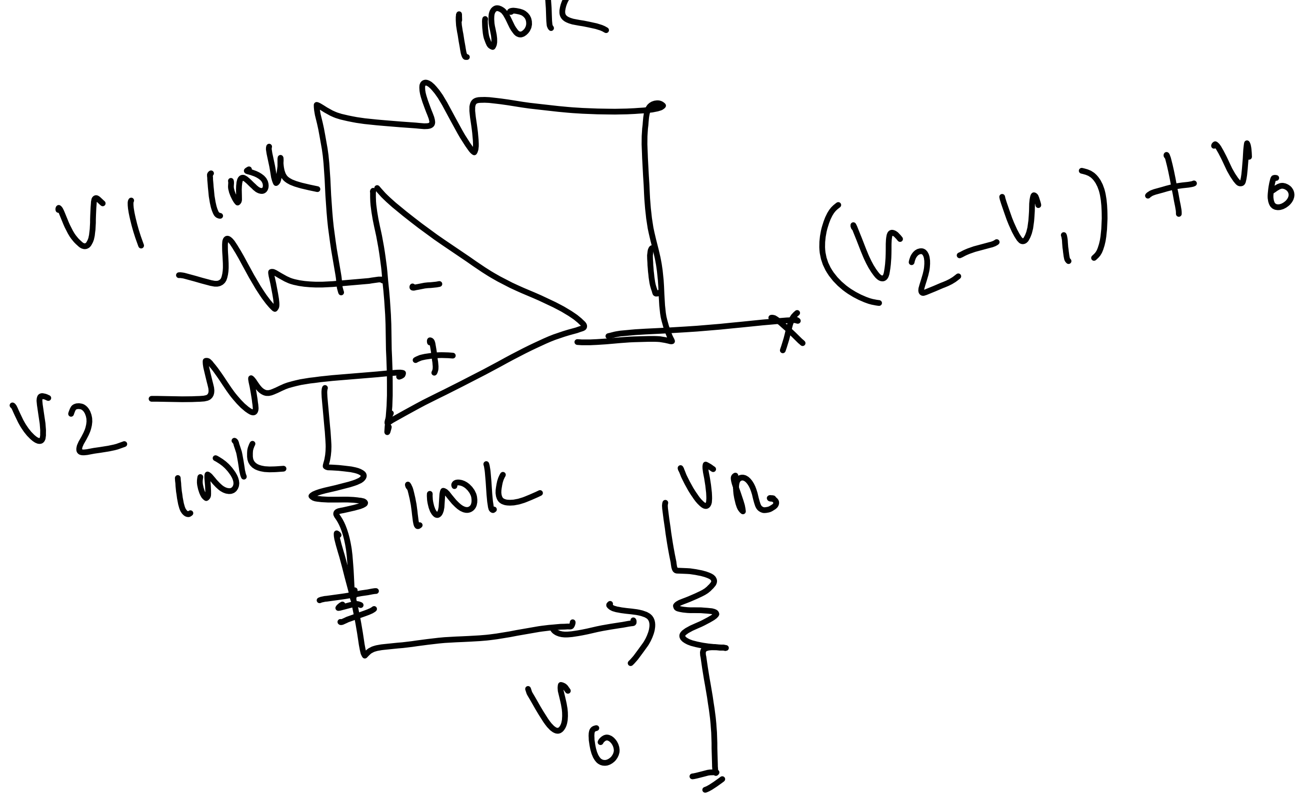


At inverting terminal $\frac{V_1}{2} + \frac{V_0}{2}$

At + terminal $= \frac{V_2}{2} + \frac{V_R}{2}$

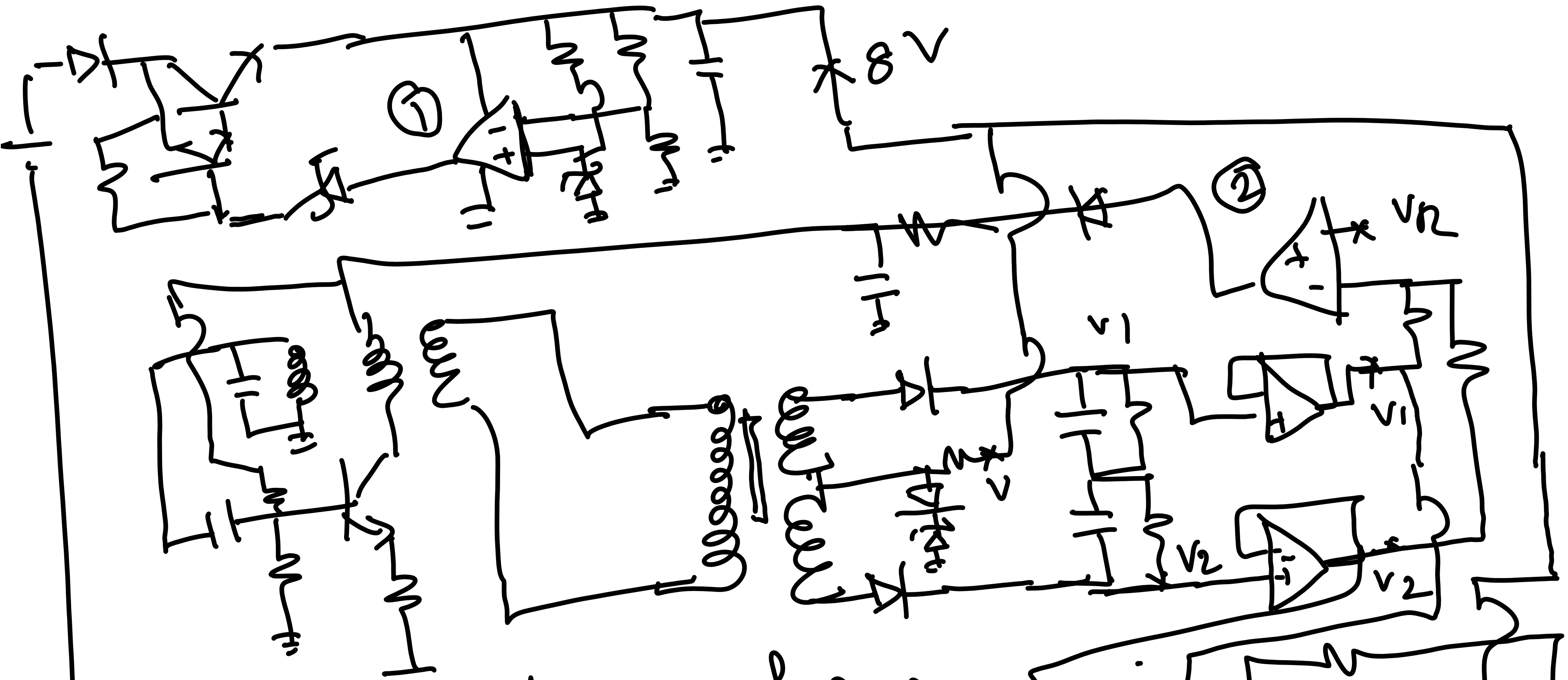
$V_1 + V_0 = V_2 + V_R$

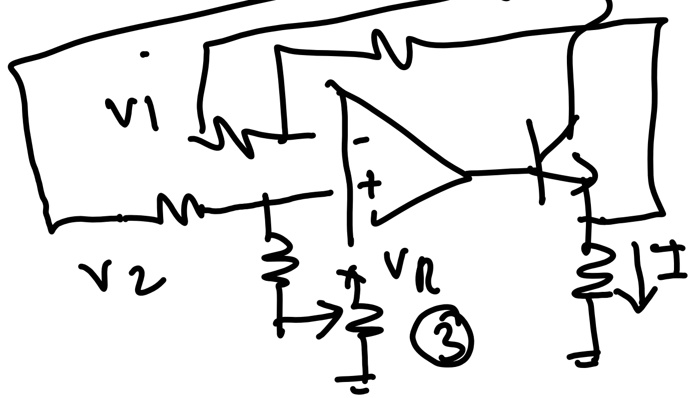
$V_2 - V_1 = V_0 - V_R$



100k
V1
100k
V2
100k
100k
Vn
V0
$(V_2 - V_1) + V_0$

1. LVDT excitation using sine wave
2. Temperature compensation
3. $(V_1 - V_2) \propto$ displacement
4. $(V_1 - V_2)$ is converted into proportional current



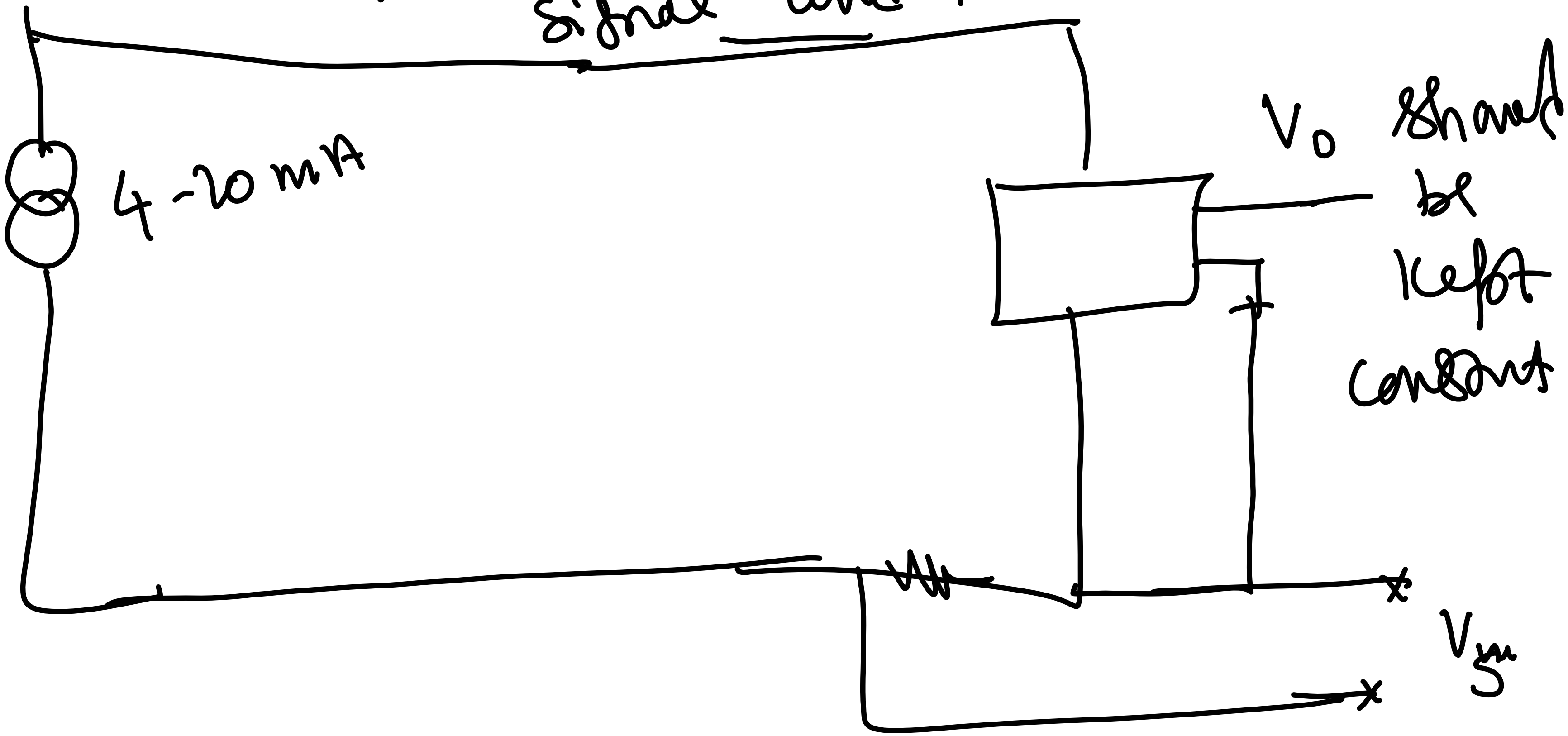


1 op-amp 1 → w/ reg

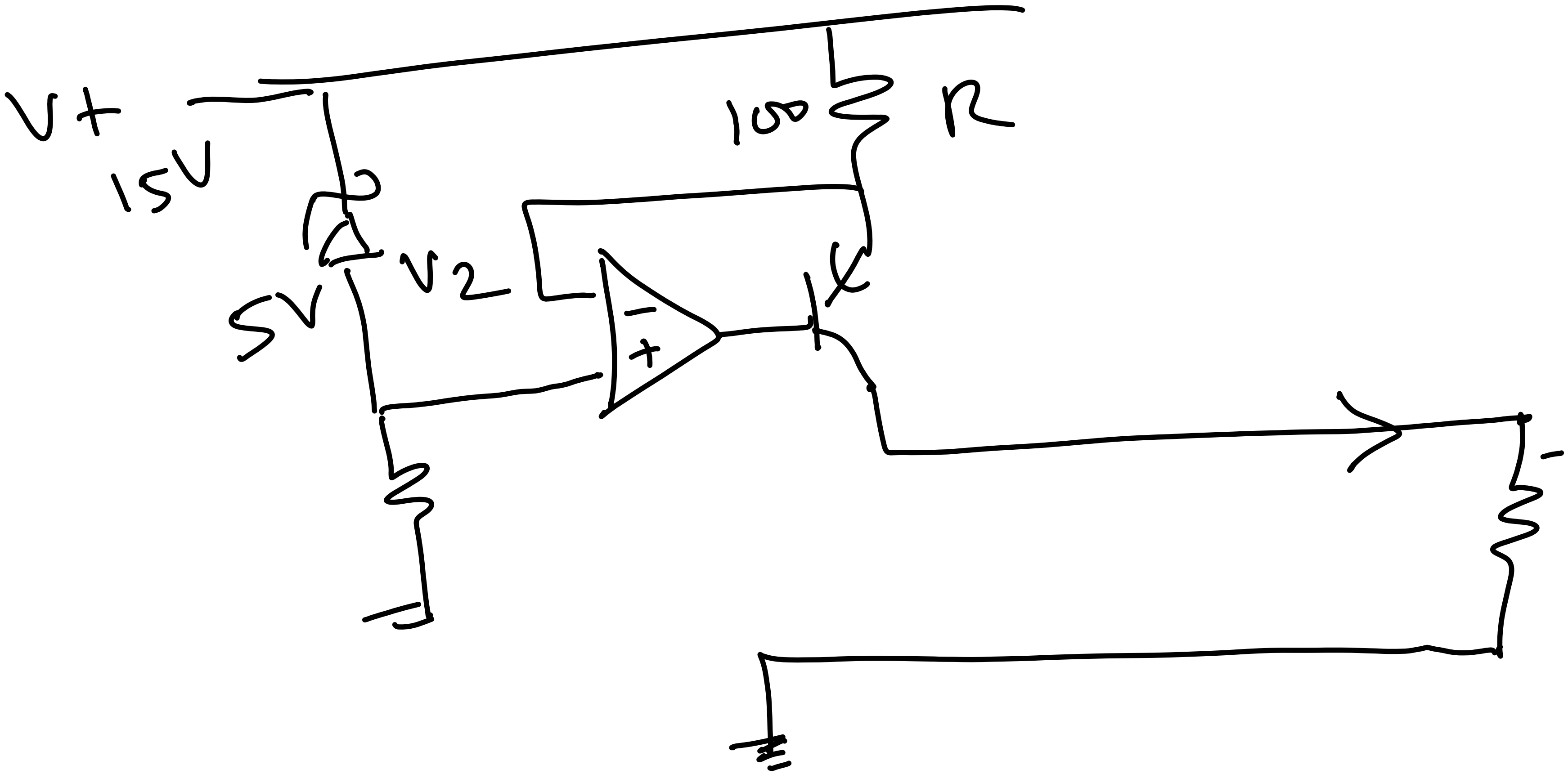
2 op-amp 2 → (V_1+V_2) constant

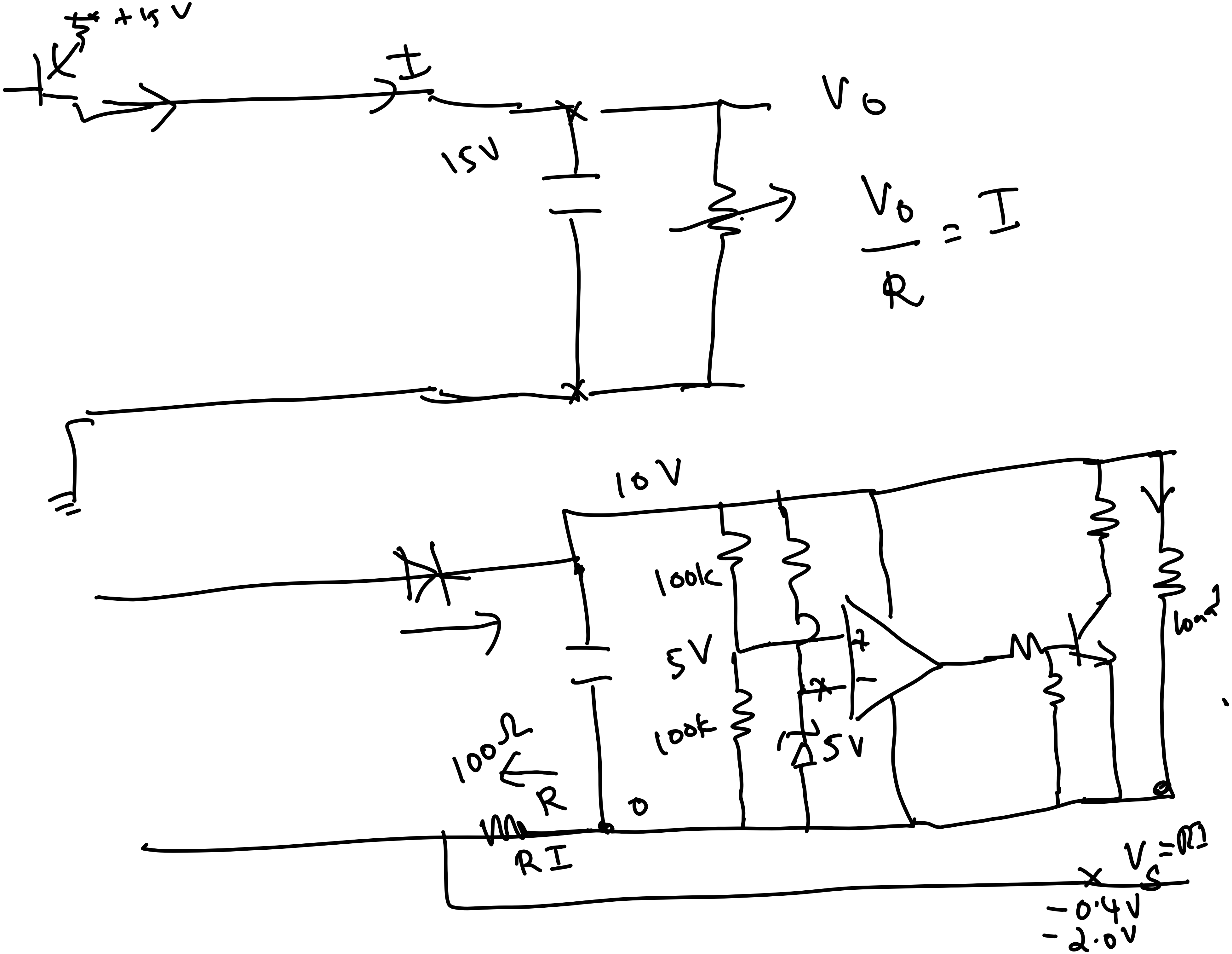
3 op-amp 3 → $(V_1-V_2) \propto I$

Spliting the current into signal and power

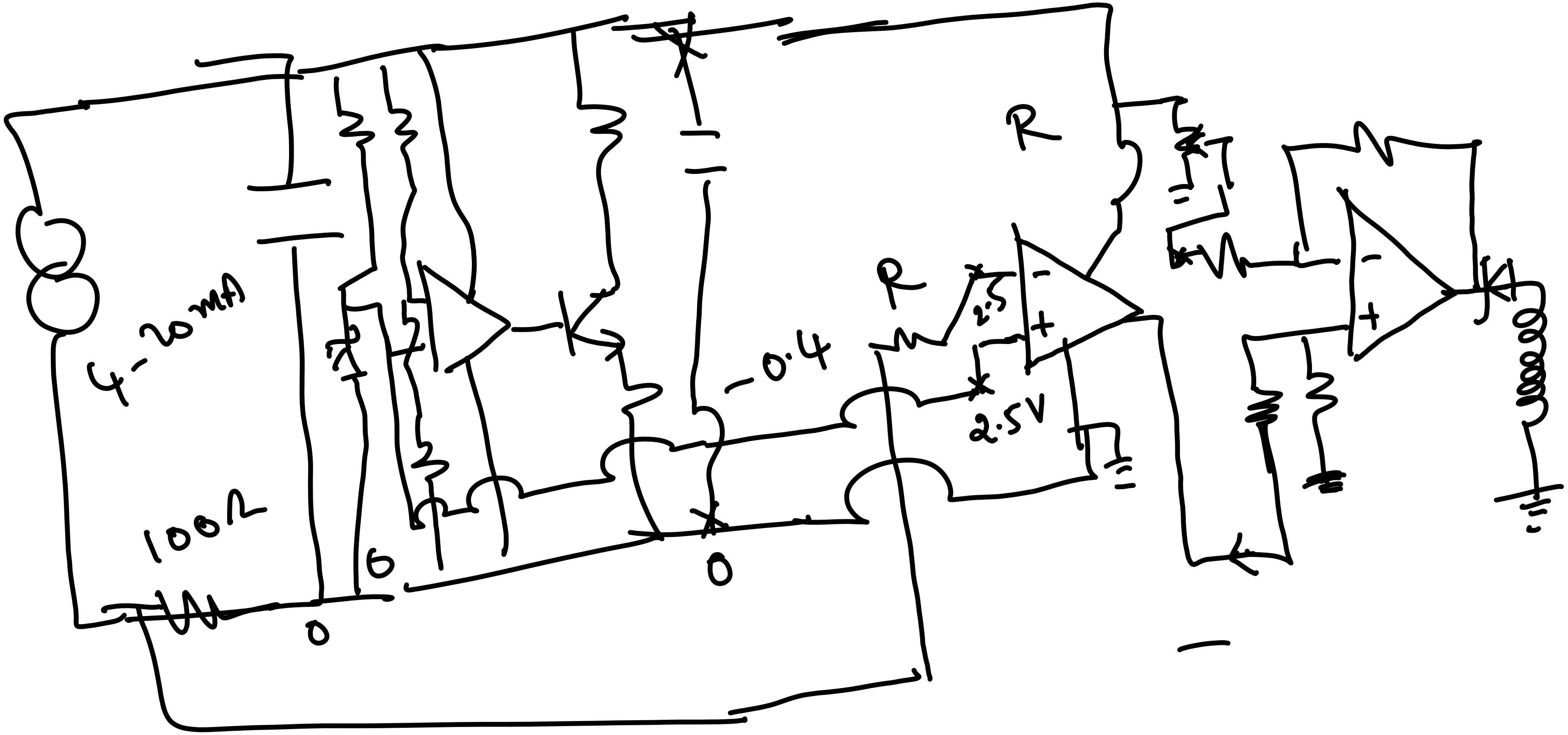


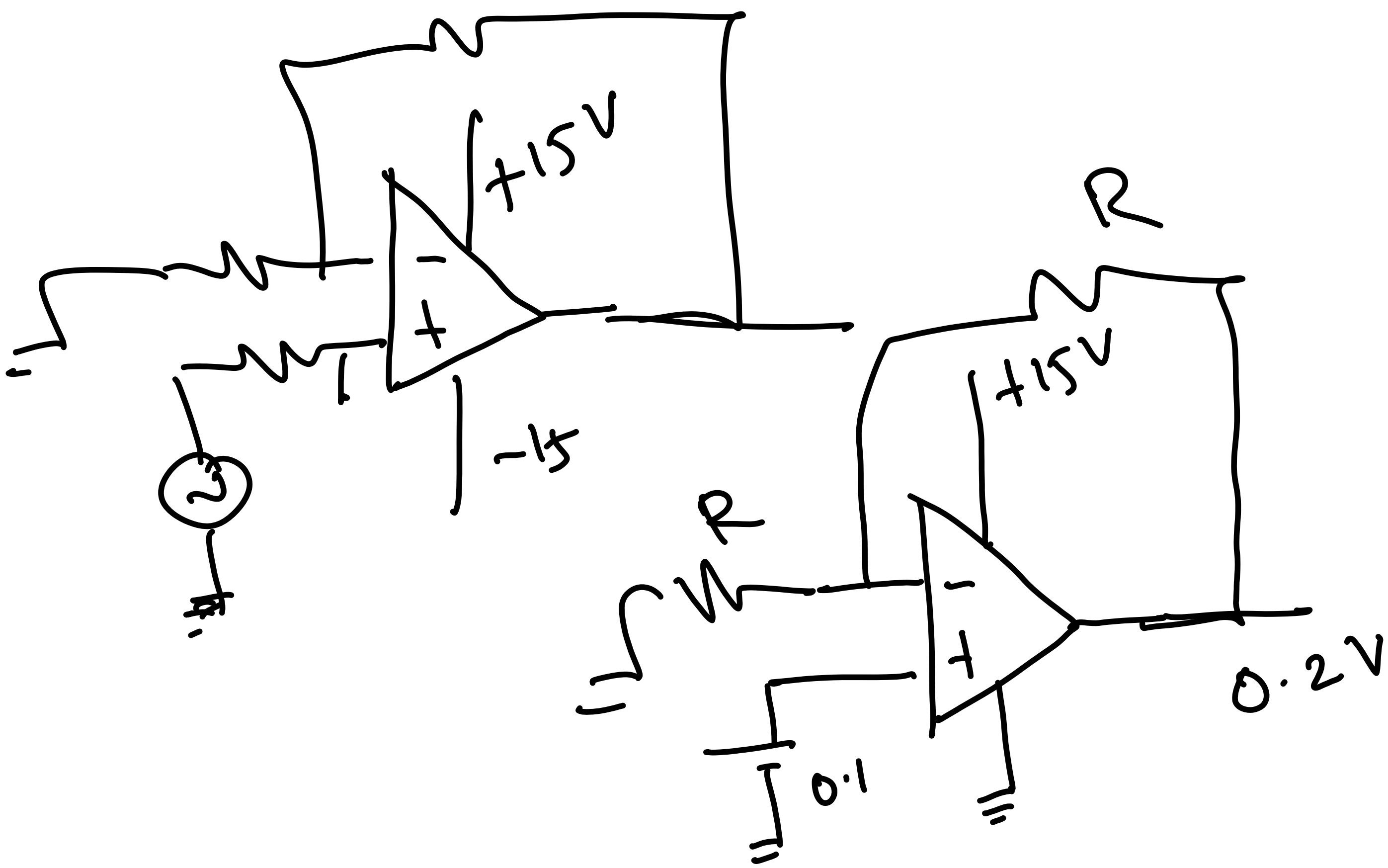
Constant current source

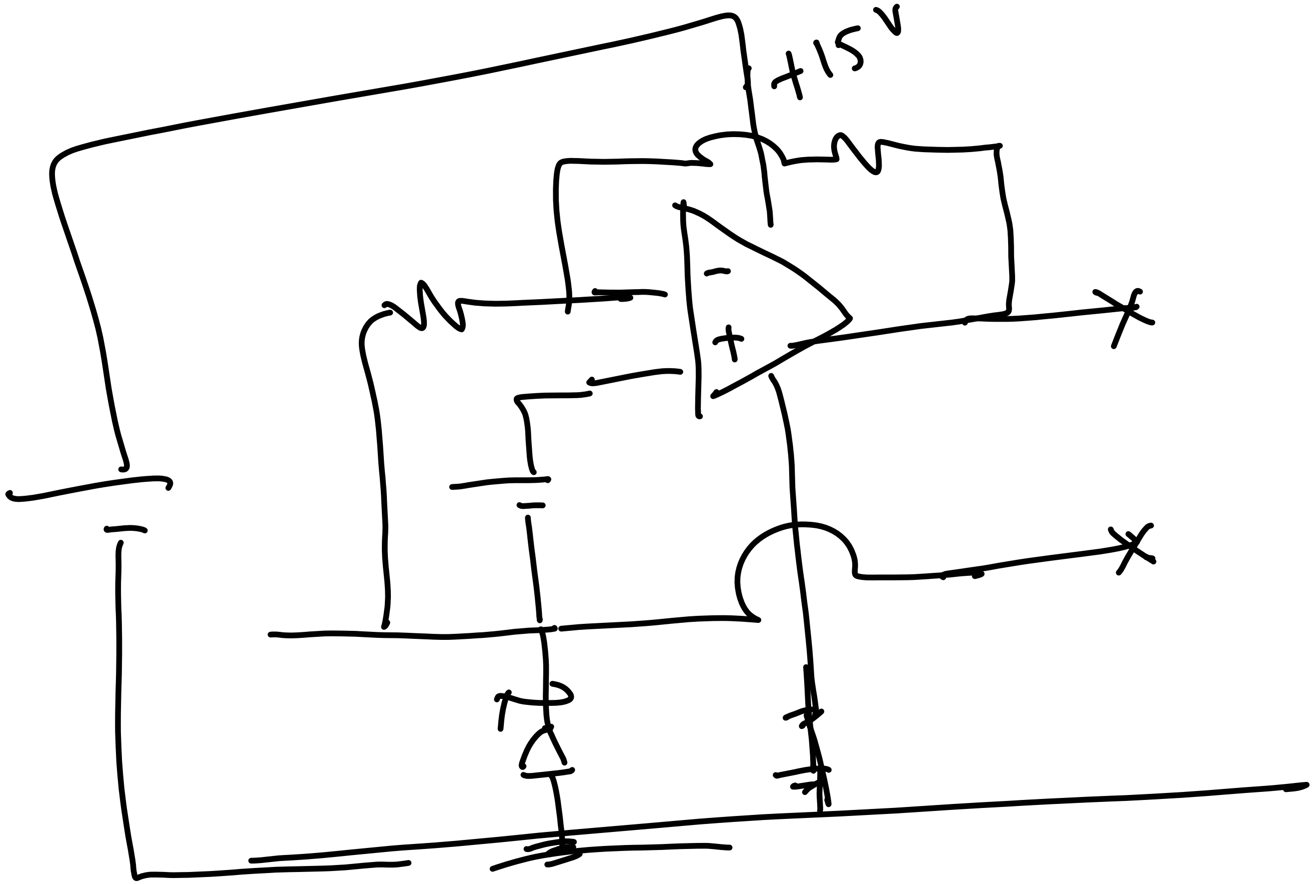




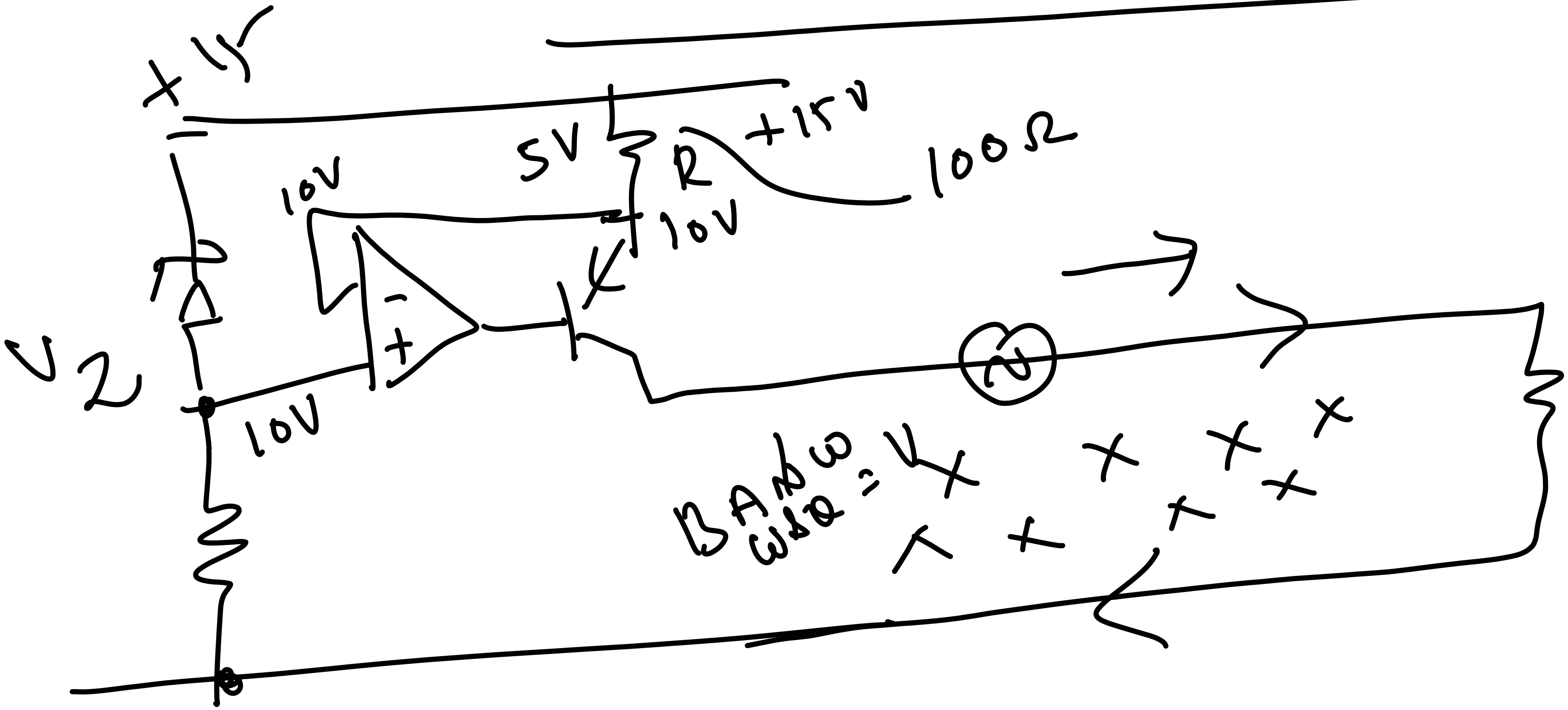
+15 V
I
15V
V0
V0/R = I
10V
100k
5V
100k
5V
100Ω
R
RI
0
load
Vs = RI
-0.4V
-2.0V

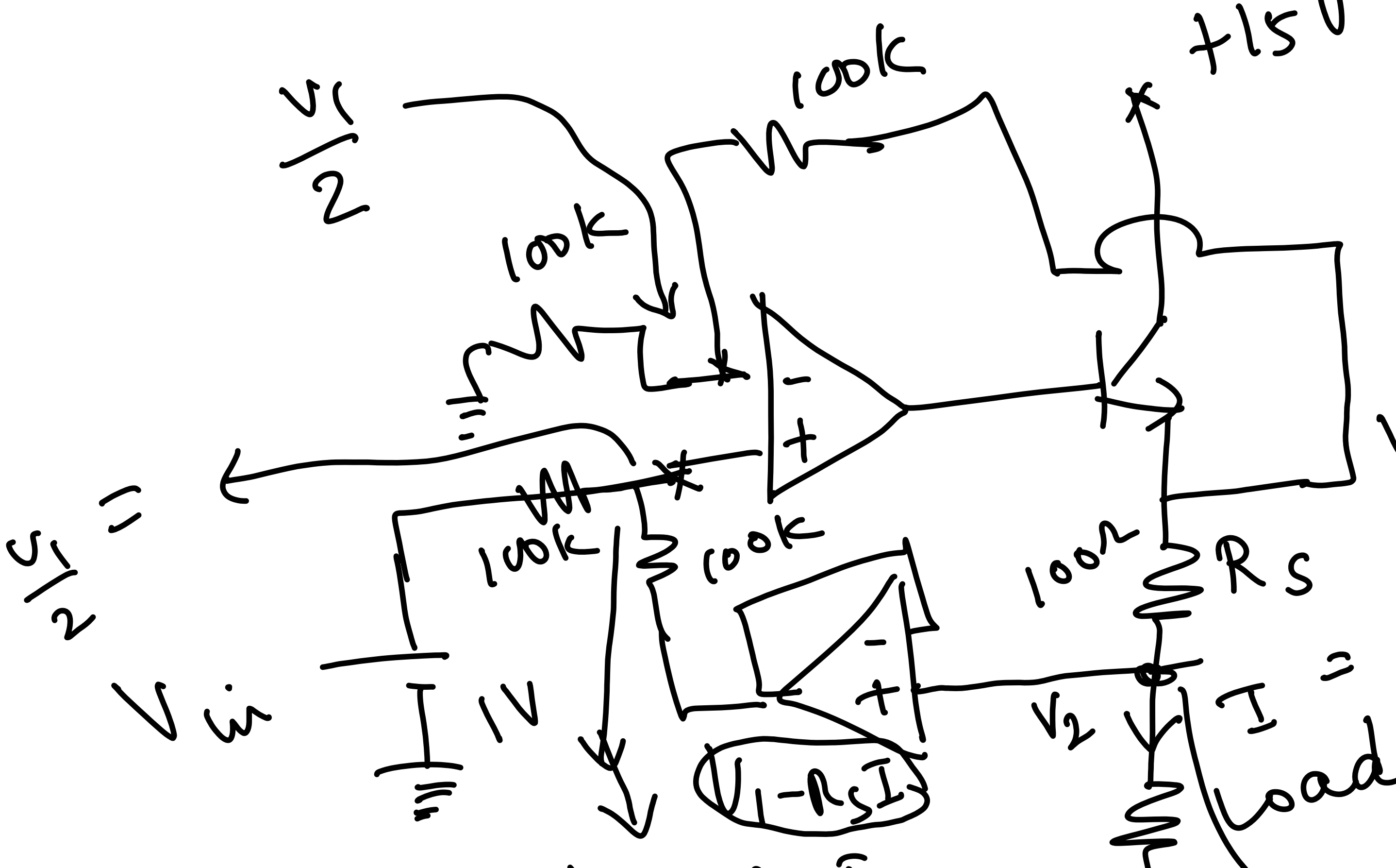

4-20mA
100Ω
0
6
0
-0.4
R
R
2.5
2.5V
-
+
-
+


+15V
-15
R
+15V
R
0.1
0.2V


+15V

Constant current source





$$I = \frac{V_{in}}{R_S}$$

$$V_2 = V_1 - R_S I$$

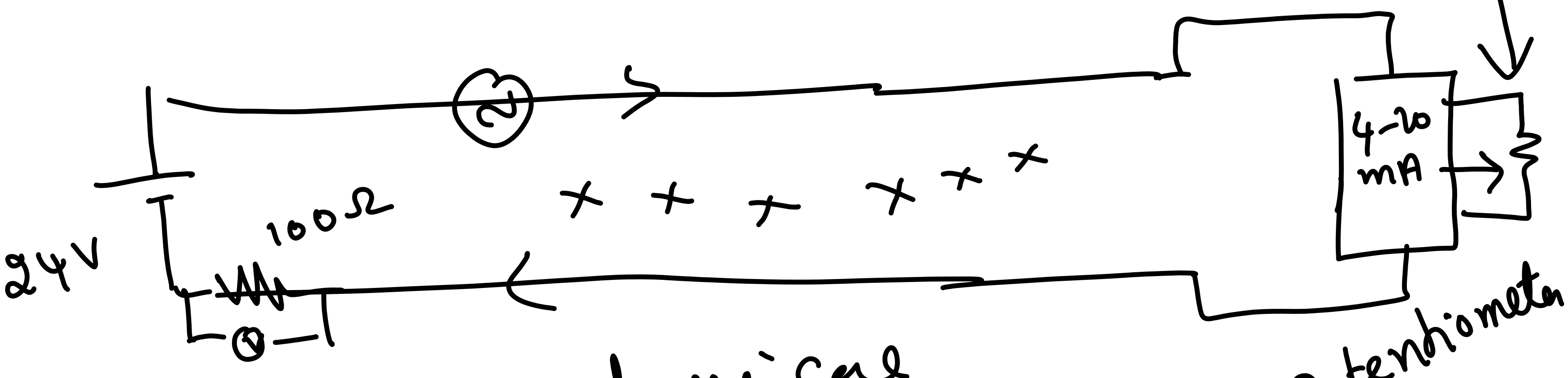
$$\frac{V_{in}}{2} + \frac{V_1}{2} - \frac{R_S I}{2}$$

$$\frac{\cancel{V_1}}{2} = \frac{V_{in}}{2} + \frac{\cancel{V_1}}{2} - \frac{R_S I}{2}$$

$$R_s I = V_{in}$$

$$I = \frac{V_{in}}{R_s}$$

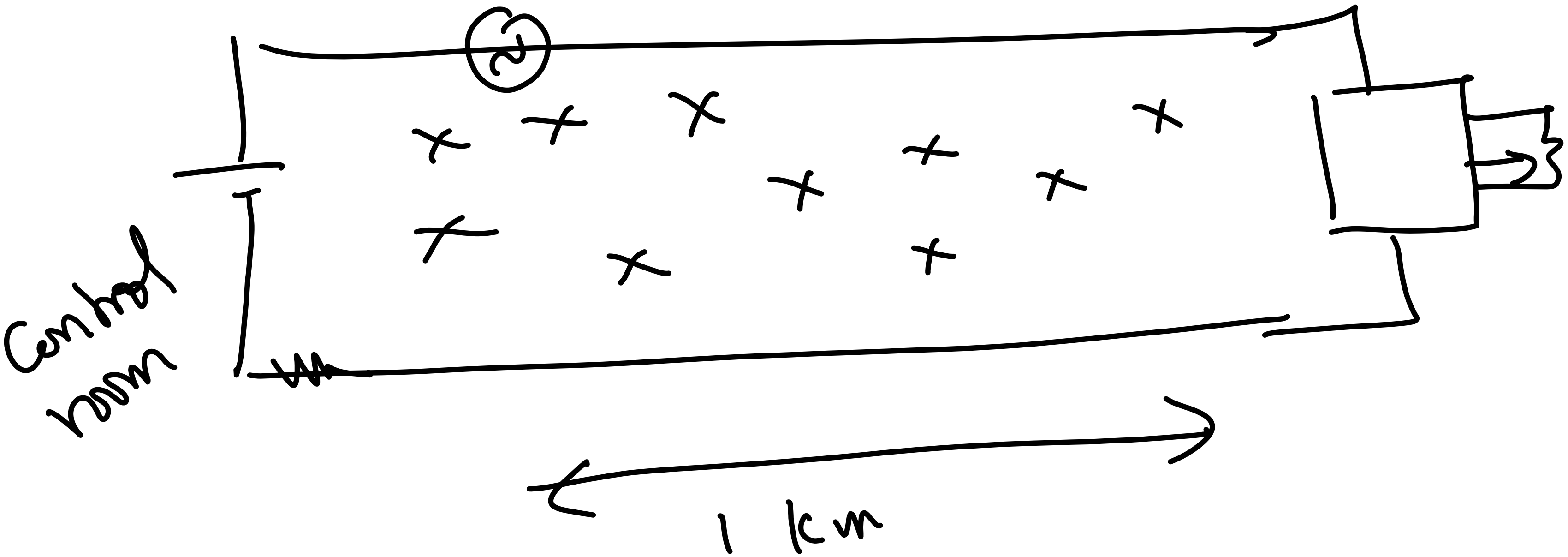
4 - 20 mA current transmitters

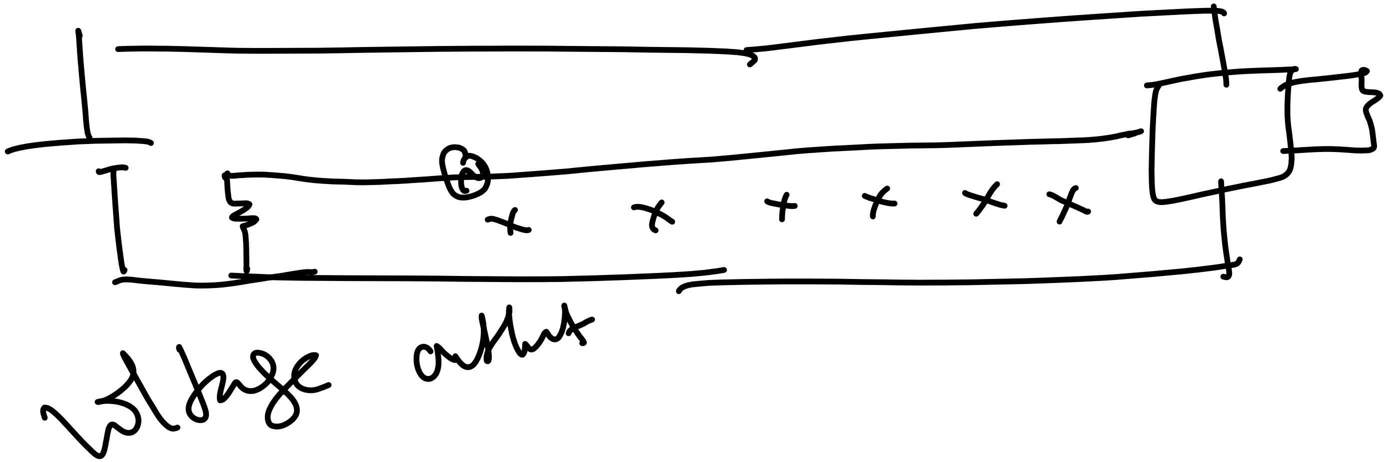


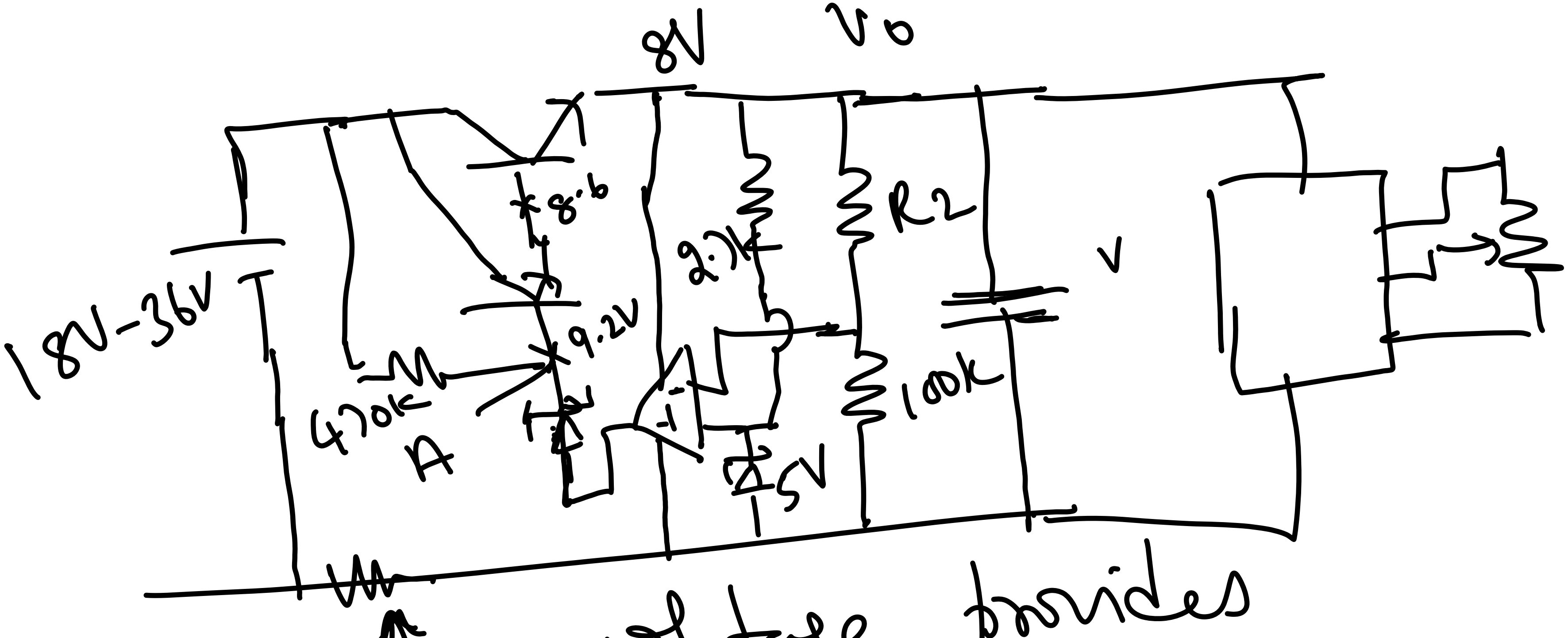
When mechanical system at ~~one~~ end = 4 mA

At the other end I should be 20 mA

0% → 4 mA
25% → 8 mA
50% = 12 mA
75% 16 mA
100% 20 mA


Control room
1 km


Voltage output



Zener voltage provides extra voltage i.e.

Vol at A = OP amp output + vol acc the zener

vol acc the zener ≠ breakdown vol of the zener

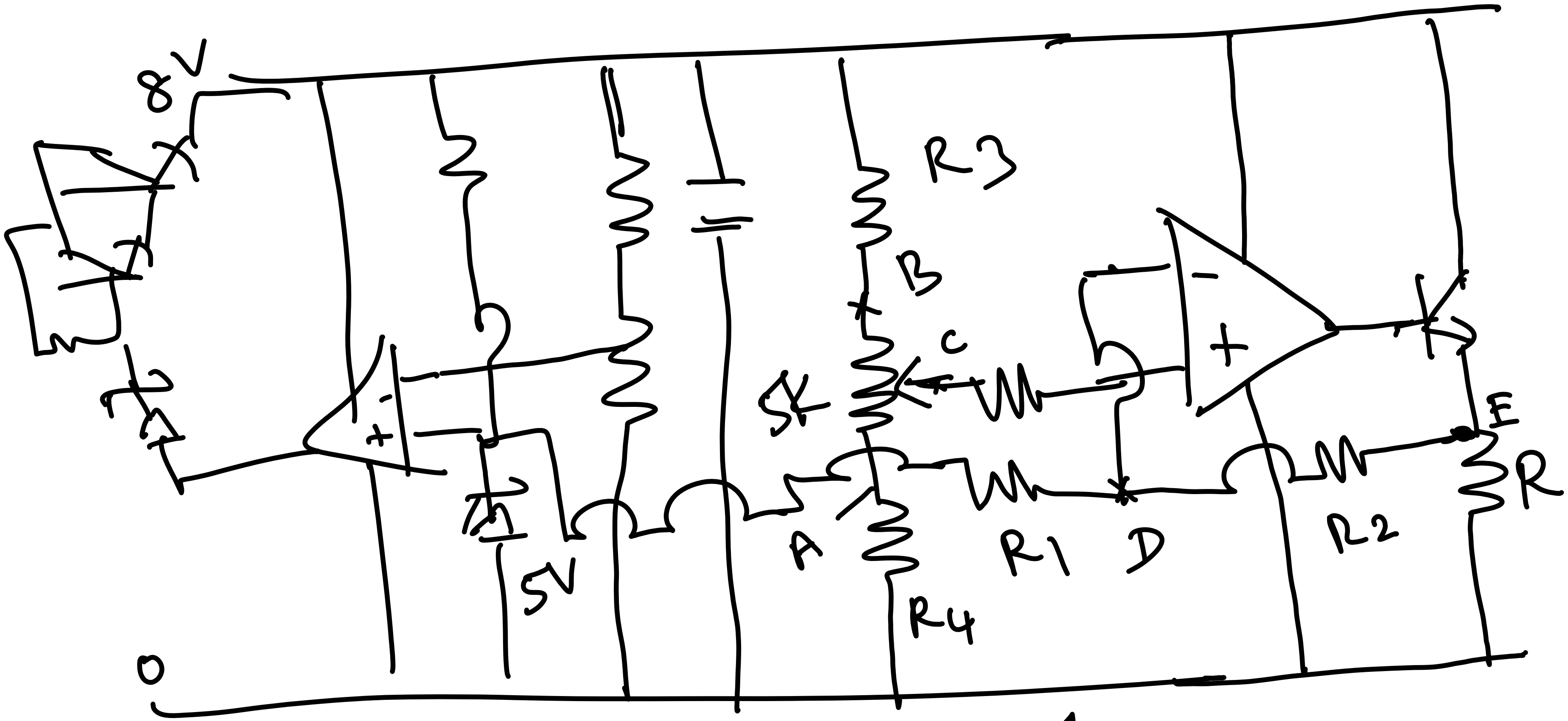
Vol acc $100k = 5V$

So output Vol $= \frac{5V \times (100k + x)}{100k}$ ✓

$= 8$

x can be calculated

Design of a current transmitter



Select R4, R3, pot
such that vol at A = 2V
vol at B = 4V
vol across the potentiometer
becomes 2V

The expected change at D = 2V

= change at C

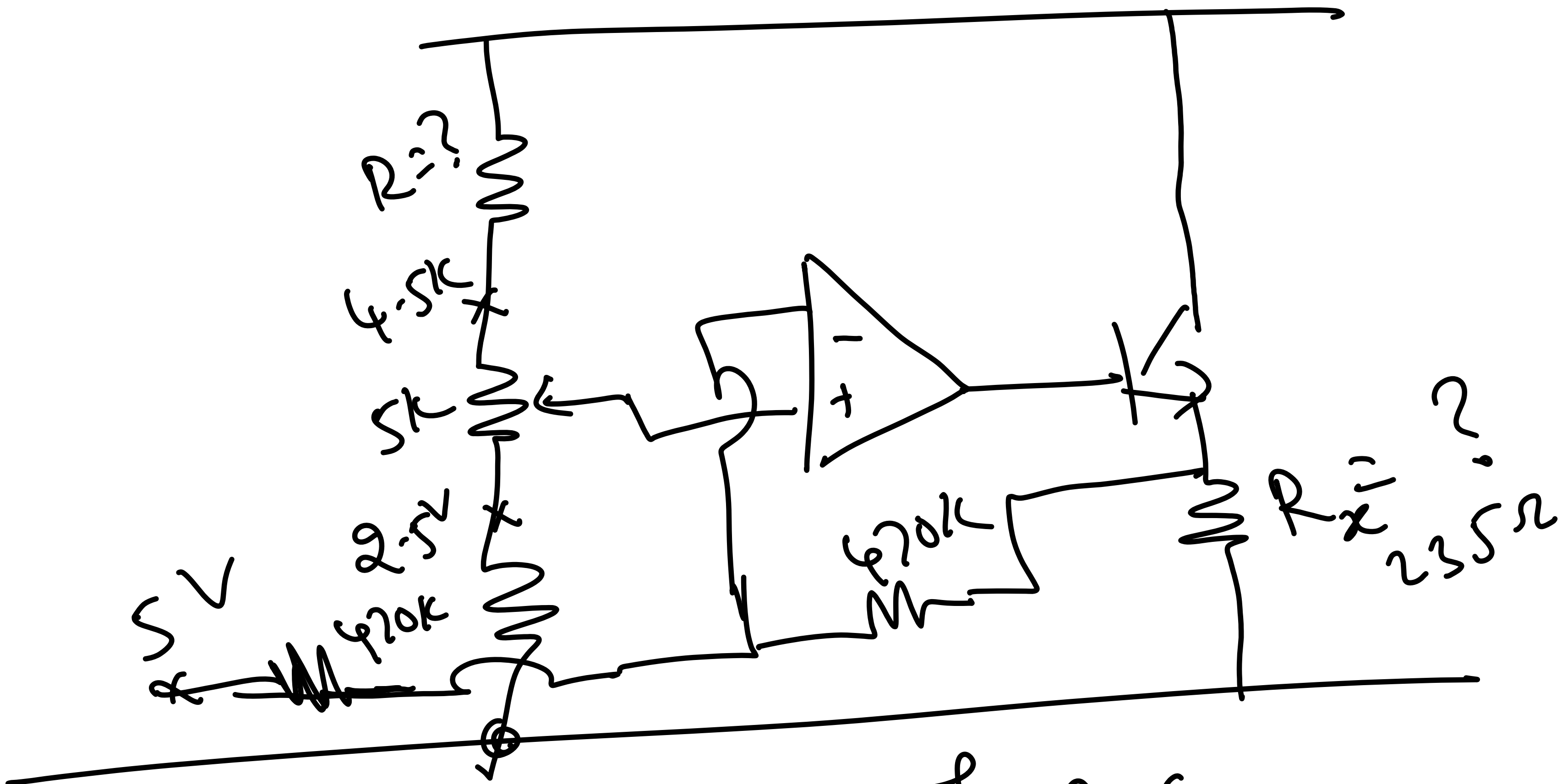
If $R_1 \simeq R_2$

Keep A at 2.5V

Then $R_1 = R_2 = 470K$

B at 4.5 V

Potentiometer W = 2V



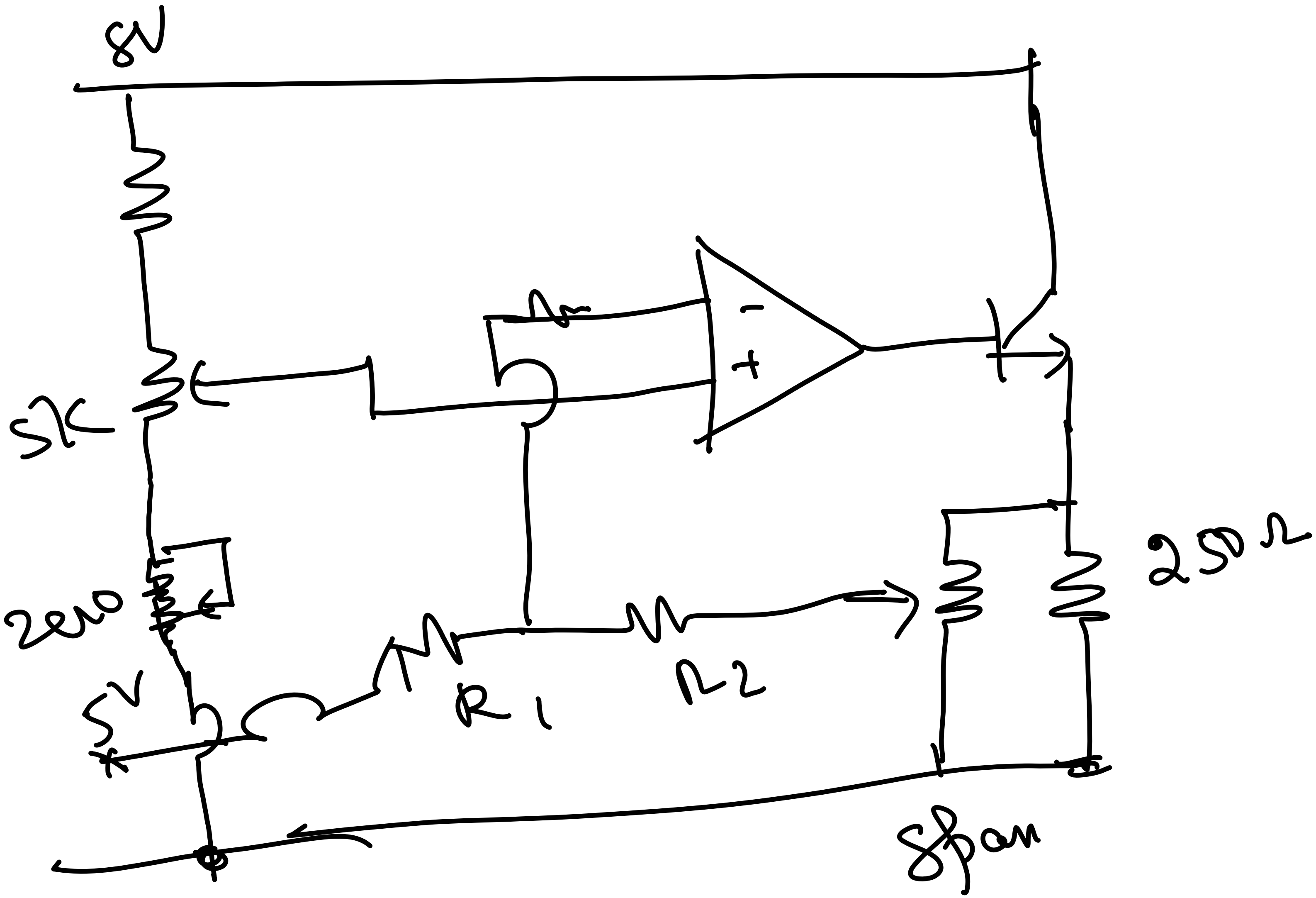
Required w/ acc
Rx = 4V at 17 mA

Required w/ acc Rx ≈ 0 V at 1 mA or less

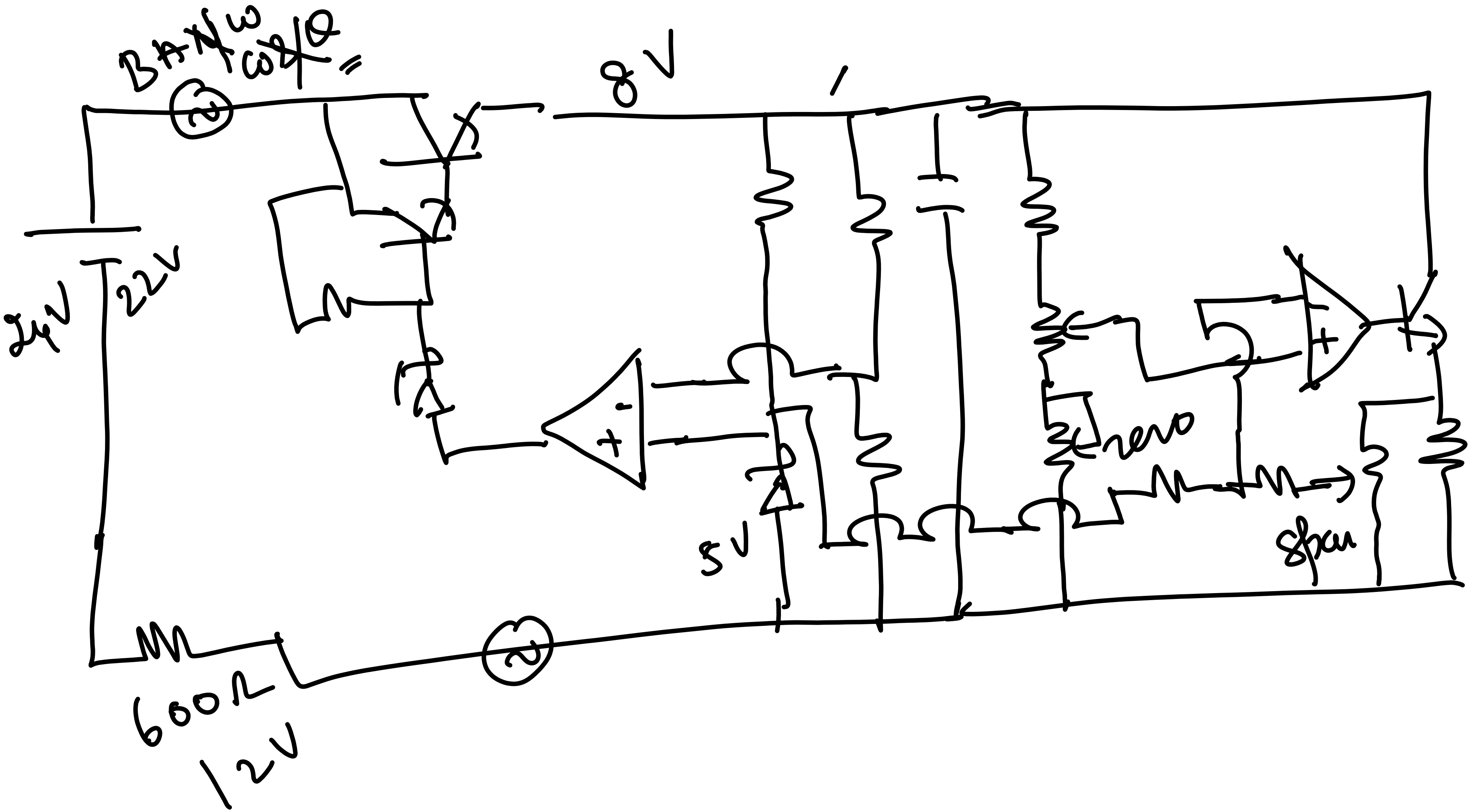
$R_x \times 17 \text{ mA} = 4V$

$$R_x = \frac{4V}{17} = \frac{4000}{17} = \underline{\underline{235}}\,\Omega$$

60



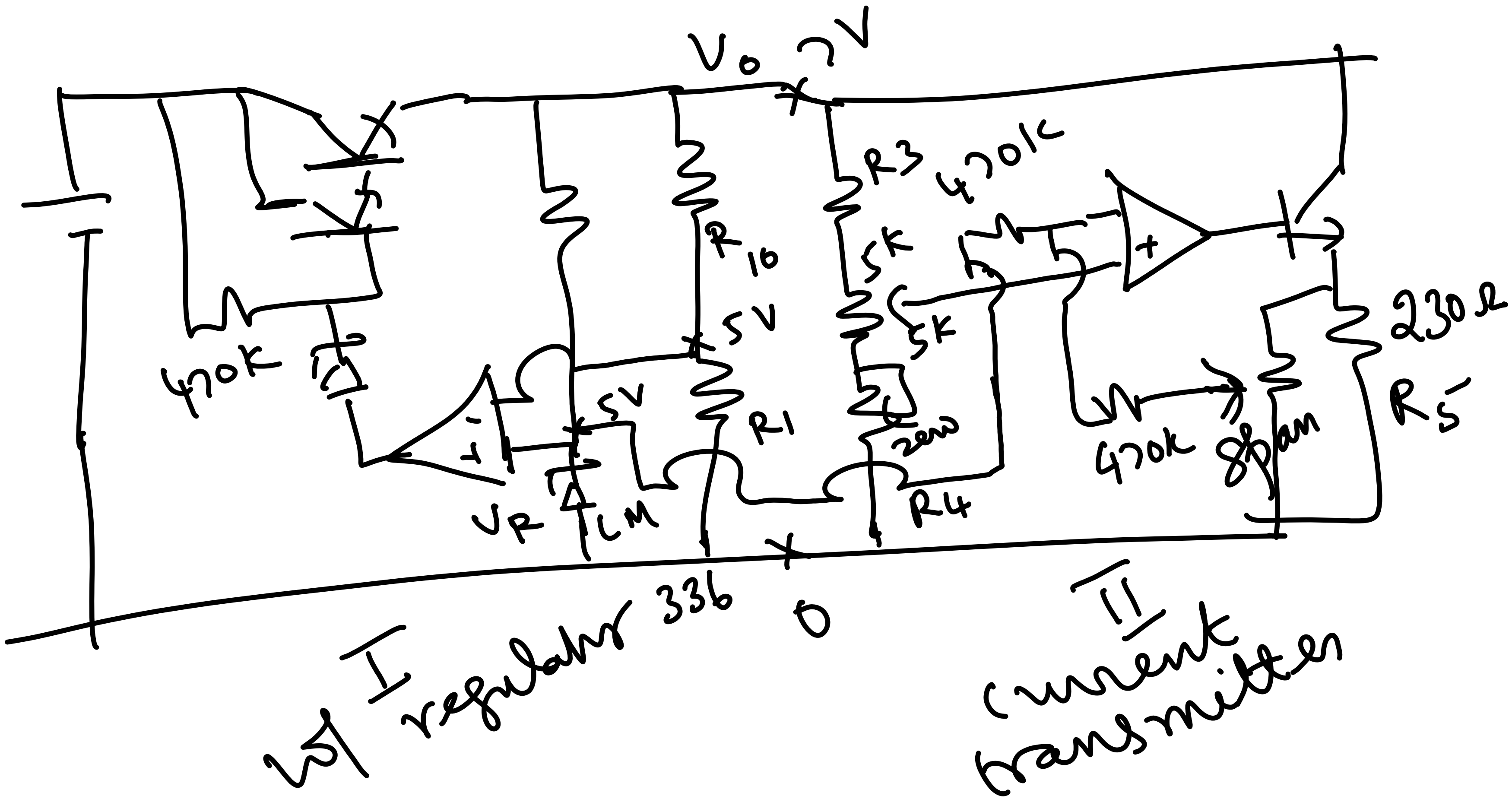
8V
5K
zero
5V
R1
R2
250Ω
span
-
+


8V
24V
22V
5V
zero
span
600Ω
12V

Error budgeting for

4-20mA current

Transmitter



LM336 Zener 5V 30PPM/°C

Drift = 30 ppm/°C

Assuming ambient temp change = 100°C

Vt of the Zener initially be = 5V

ΔV of the Zener $= \frac{5 \times 30 \times 100}{10^6}$

$V \times PPM \times \Delta C$

$\Delta V = \frac{15 \times 10^3}{10^6} = 15 \times 10^{-3}\,V$

$= \pm 15\,mV$

ΔV of the zener $= \pm 15\,mV$

So regulated output voltage

$$\text{change} = \frac{15\,mV \times (R_1 + R_{10})}{R_1}$$

$$7V = 5 \times \frac{(R_1 + R_{10})}{R_1}$$

$$\frac{7V}{5V} = \frac{(R_1 + R_{10})}{R_1}$$

$$\Delta V_o = 15 \times 10^{-3} \times \frac{7}{5} = 21\,mV$$

Vol acc $R_3 = R_4 = 2.5V$

For 5k the vol
drop = 2V

$$\text{So } R_3 = \frac{5k \times 2.5}{2} = \frac{12.5}{2} = 6.25k$$

$R_3 = r_4 = 6.25k$

vol at the potentiometer
top end = $2.5 + 2.0 = 4.5V$
(This corresponds to current of
20 mA)

For ΔV_o of 21 mV the corresponding
voltage change at the top

end of the potentiometer

$$= \frac{21 \times 4.5}{7.0} \text{ mV}$$

$$= 3 \times 4.5 \text{ mV} = \underline{13.5 \text{ mV}}$$

Assuming the span pot at the top end the o/p change at, at inverting input $= 13.5$ mV $=$ o/p change at the non-inverting input

ie at $-$ve input $\Delta V = 13.5$ mV

So o/p change acc R5

$= 2 \times 13.5 = 27$ mV

wl change acc R_5

$= 27 mv$

So current change $= \dfrac{27 \times mv}{230}$

$= \dfrac{27}{230} = \pm 0.08\ mA$

II <u>calculate the current drift due to R_1 ; R_{10} change</u>

$R_1 = 100K$

$R_{10} = \dfrac{5 \times (R_1 + R_{10})}{R_1} = 7V$

$= \dfrac{5 \ (10^5 + R_{10})}{10^5} = 7V$

$$7 = \frac{5 \times (10^5 + R_{10})}{10^5}$$

$$7 \times 10^5 = 5 \times 10^5 + 5 \times R_{10}$$

$$\frac{2 \times 10^5}{5} = R_{10}$$

$$R_{10} = \frac{200 \times 10^3}{5} = 40K$$

Assume R_1, R_{10} drifts by 50 ppm/°C

Assume R_1 is decreasing and R_{10} is increasing with temp

ΔR_1 change $= -\dfrac{100 \times 10^3 \times 50 \times 100}{10^6}$

$= \dfrac{5 \times 10^8}{10^6} = -500\ \Omega$

$\Delta R_{10} = \dfrac{40 \times 10^3 \times 50 \times 100}{10^6}$

$= 20 \times \dfrac{10^7}{10^6} = 200\ \Omega$

R_1 decreases by $500\ \Omega$

R_{10} increases by $200\ \Omega$

So change in regulated voltage $= \dfrac{5V\,(R_1 + R_{10})}{R_1}$

New $V_0 = \dfrac{5(100\times10^3 - 500 + 40200)}{99500}$

$= \dfrac{5\,(99500 + 40200)}{99500}$

$$\begin{array}{r} 99500 \\ 40200 \\ \hline 139700 \end{array}$$

$= 5\times\dfrac{1397\cancel{00}}{995\cancel{00}} = \dfrac{1397}{\cancel{995}_{199}}\times\cancel{5} =$

17

New $V_o = \dfrac{1397}{199} = \dfrac{1397}{200} = \underline{\underline{6.985\ V}}$

Change in $V_o = 7 - 6.985$

$= 15\ mV$

So the expected current change in $20\ mA = 0.08\ mA$

III **Error due to V_{offset} of o/p regulator stage**

Let $V_{offset} = 15 \mu V / {}^\circ C$

For $100^\circ C$ total offset voltage

change $= 15 \times 100 = 1.5 mV$

For 1.5 mV at the input of

the regulator op amp the

expected change at the output

of the regulator $= 1.5 \times \frac{7}{5}$ mV

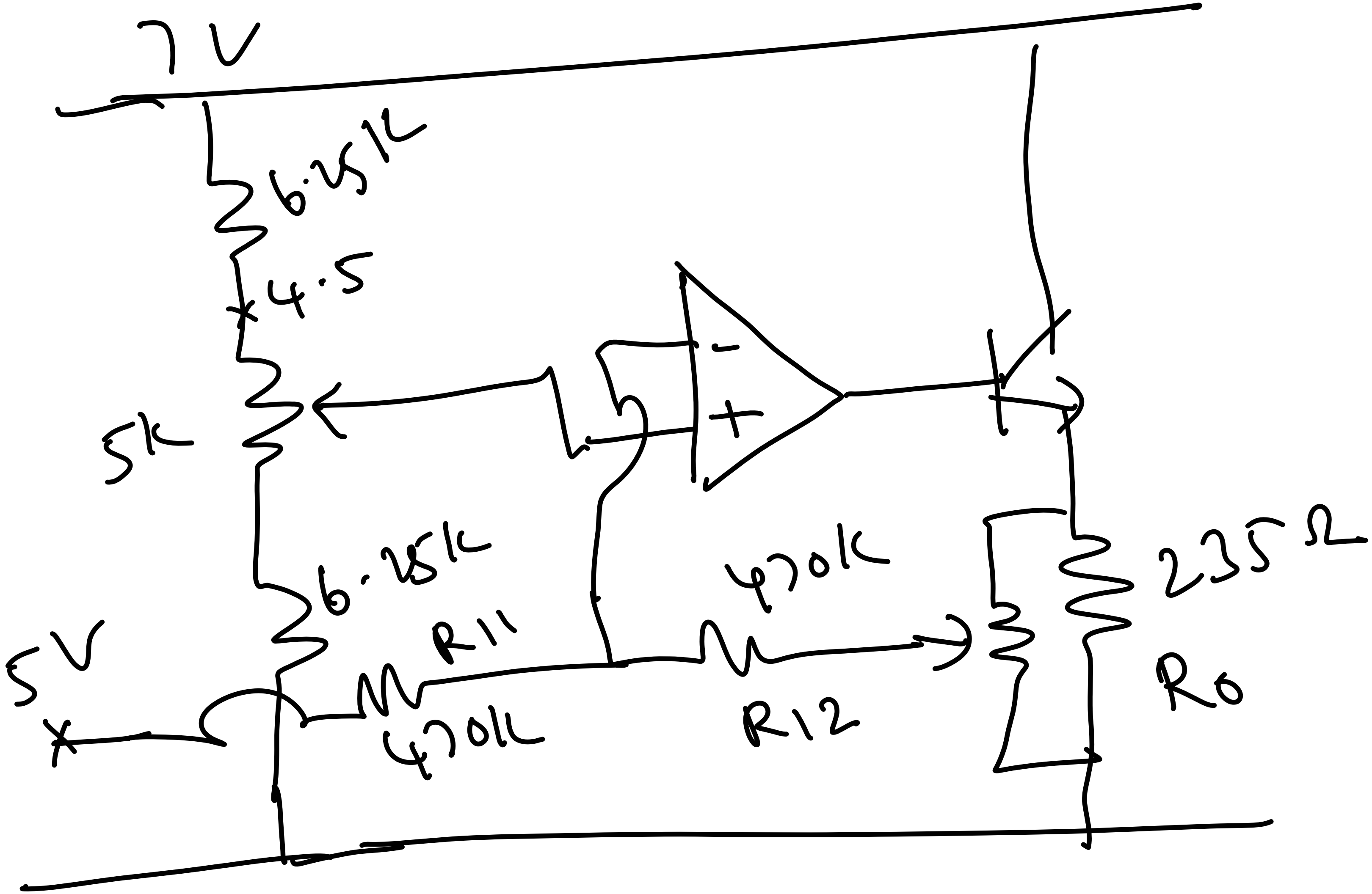
$= 2.1$ mV

So current change in the 20mA output

$$= \frac{0.08 \times 2.1}{15}$$

$$= \frac{0.16}{15} = 0.01 \text{ mA}$$

Error in the current converter stage



7V
6.25k
4.5
5k
6.25k
5V
R11
470k
470k
R12
235 Ω
Ro

(1) Error due to offset voltage drift of current converter op amp

$V_{offset} = 15 \mu V / {}^\circ C$

Total offset voltage change $= \pm 1.5\ mV$

$(15 \mu V \times 100)$

So the expected change in the non-inverting input $= 1.5\ mV$

$=$ Expected change

in the inverting input = 1.5mV

So the expected change

accross $R_0 = \frac{1.5 \times 470 + 470k}{470}$

$= 3\ mV$

So the expected current change

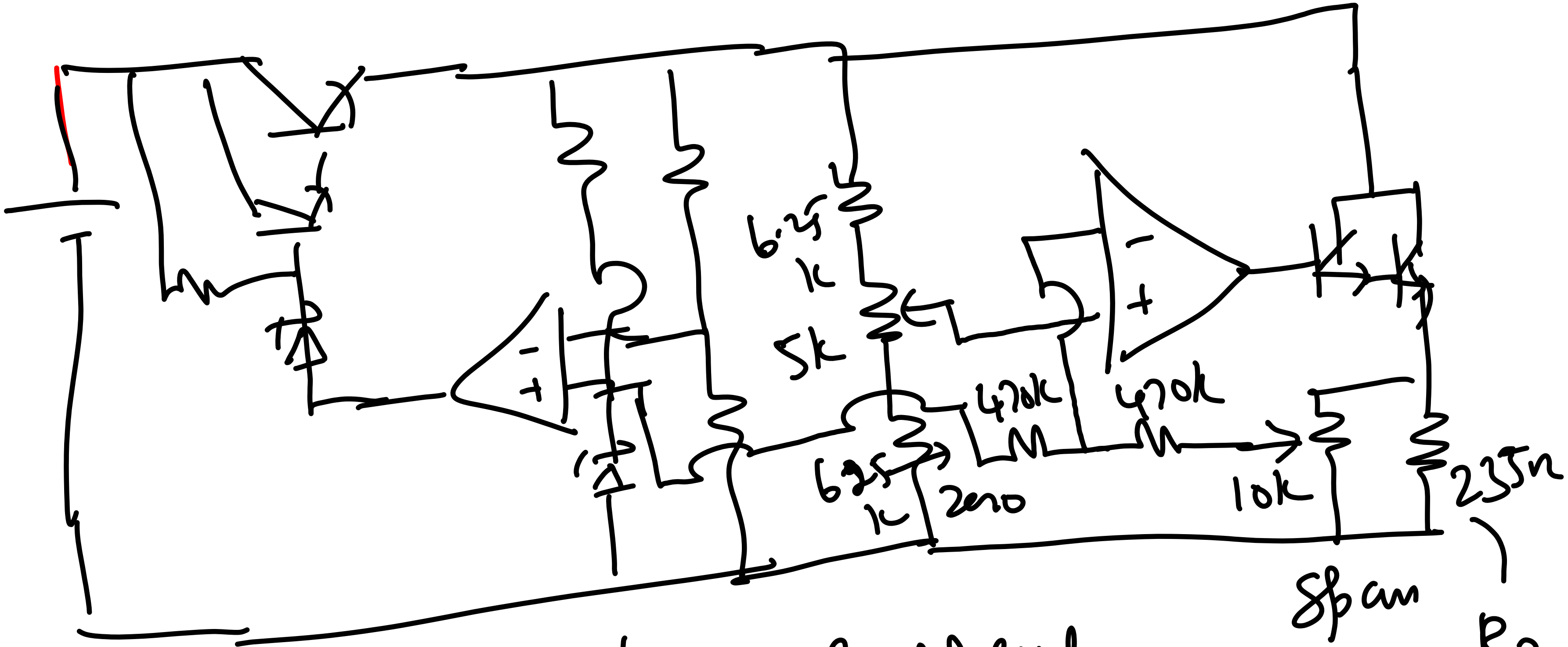
in 20 mA $= \frac{3 \times 10^{-3}}{230} = \frac{3000}{230}\ \mu A$

$= 12\ \mu A$

(2) change in current due to zener vol change

(3) Change in current due to R_{11}, R_{12} change

(4) Change in current due to $235\,\Omega$ resistance change



Calculate the current drift due to drift in R7

Let the TC of

$R_7 = 50 \text{ ppm}/°C$

For $\Delta T = 100°C$

$$\Delta R = \frac{235 \times 50 \times 100}{10^6}$$

$$= 235 \times 5 \times 10^{-3}$$

$$= 1.175\ \Omega$$

Assuming 16 mA current in R_7

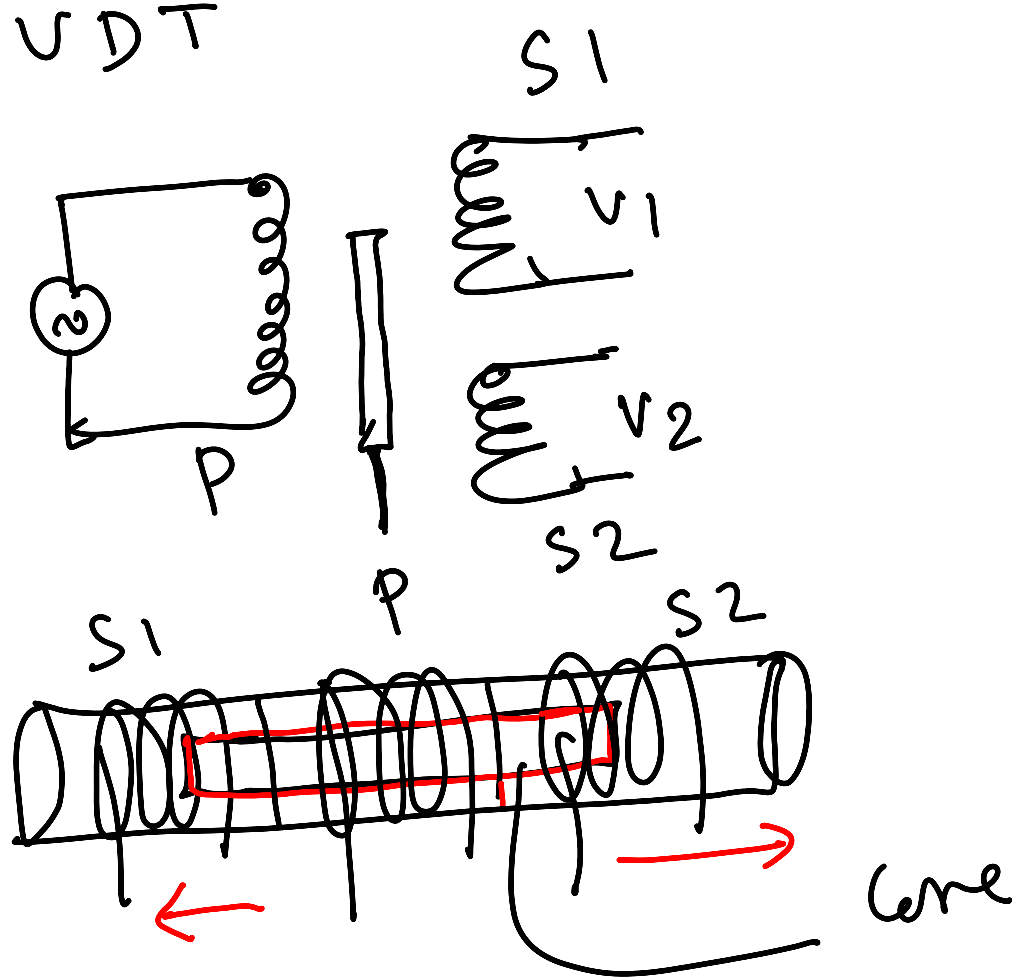
Then Δr $1.175\ \Omega$ change the expected current change

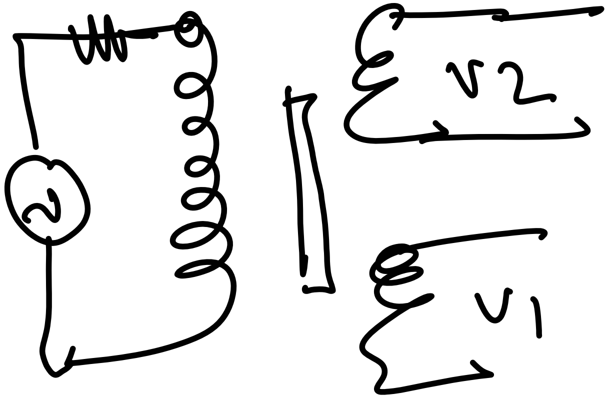
$$= \frac{16 \times 1.1}{235} = \frac{17.6}{235} \text{ mA}$$

$$= \frac{1}{15} \text{ mA} = 0.06 \text{ mA}$$

Design of LVDT based position transmitter

① LVDT





$|V_2| - |V_1|$

= displace

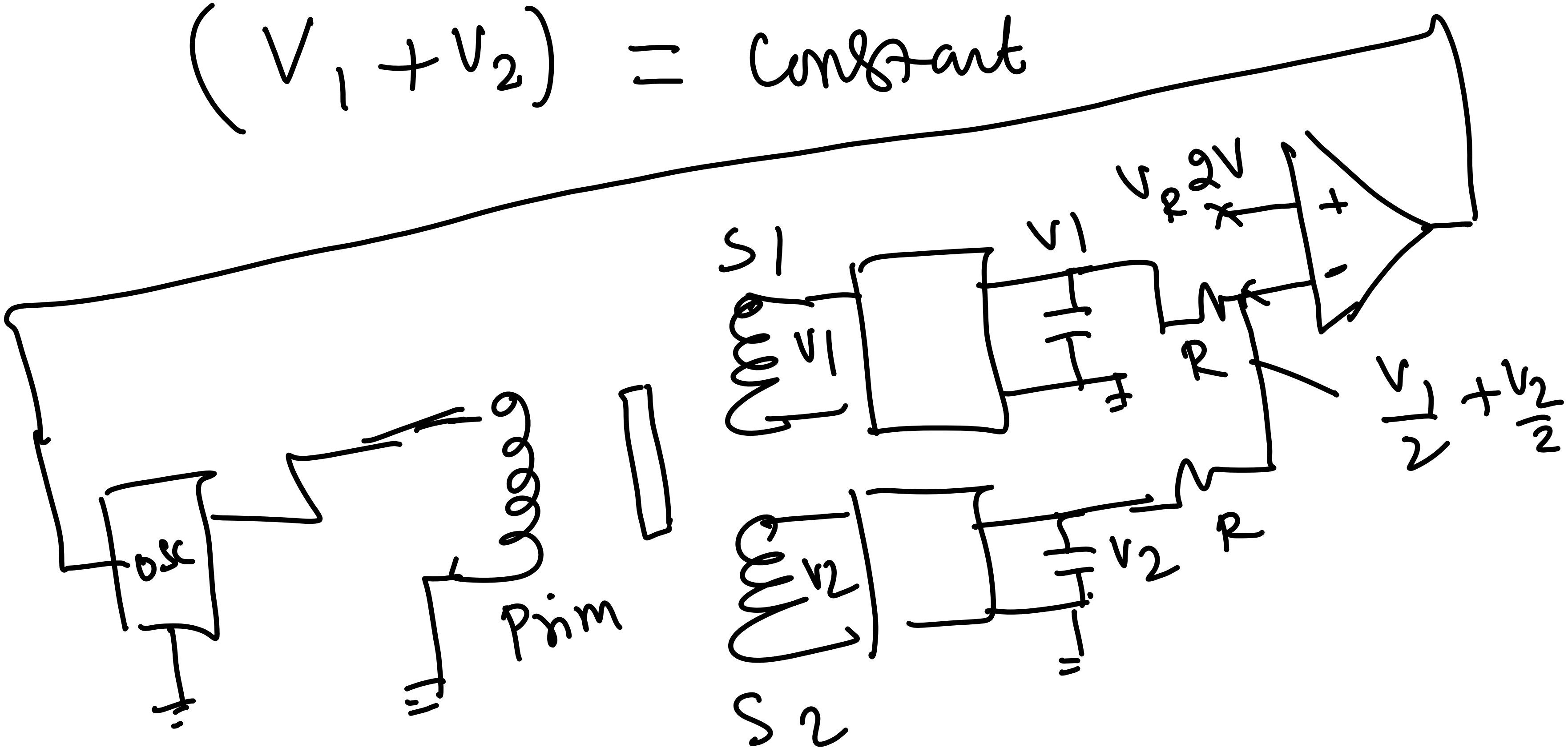
① Convert V_1 and V_2 into DC

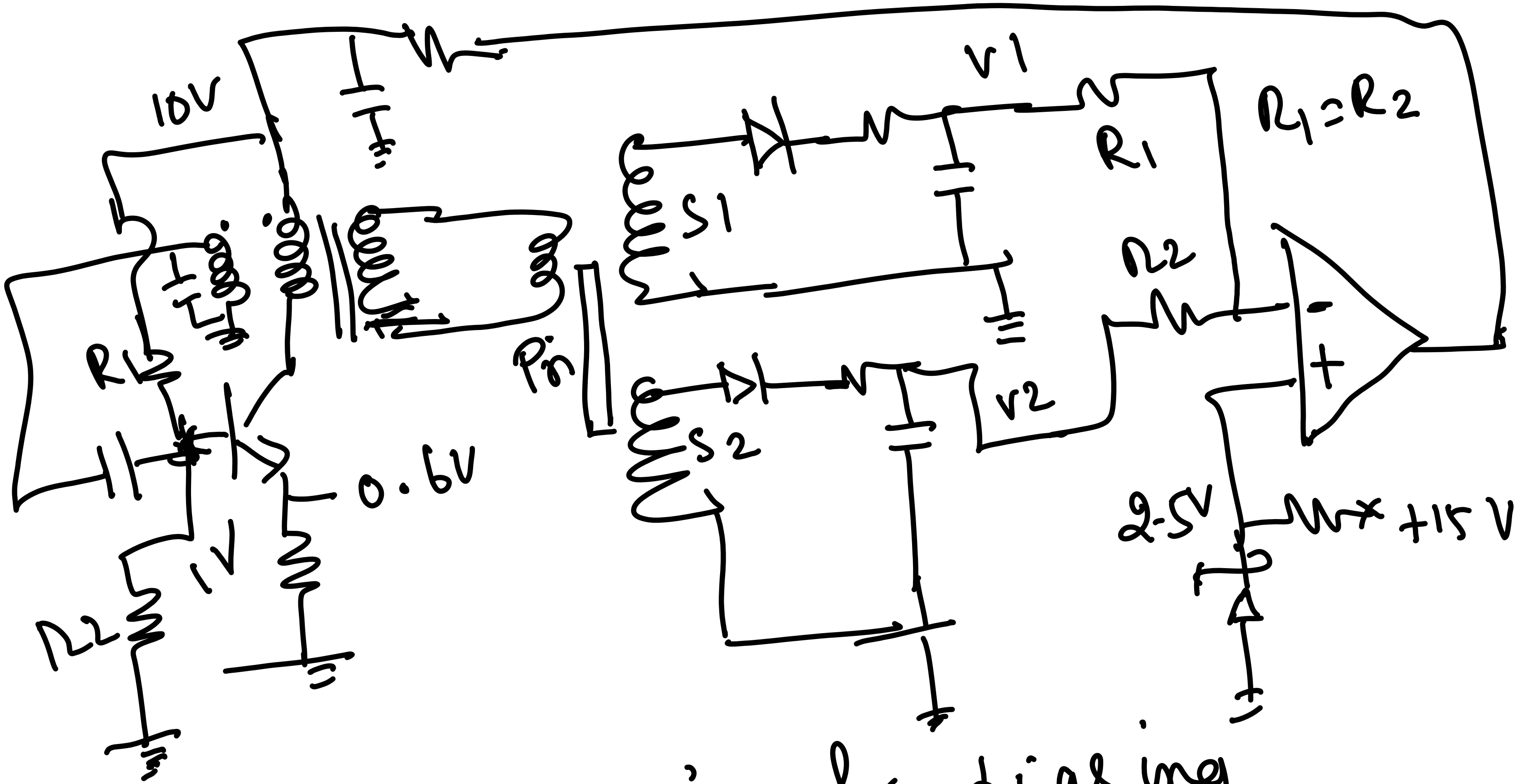
This dc w/ difference alone give you the displacement

② $|V_2 - V_1|$ changes with temp

How to make $(V_2 - V_1)$ independent of temp?

$(V_1 + V_2) = \text{constant}$





R_1, R_2 is for biasing

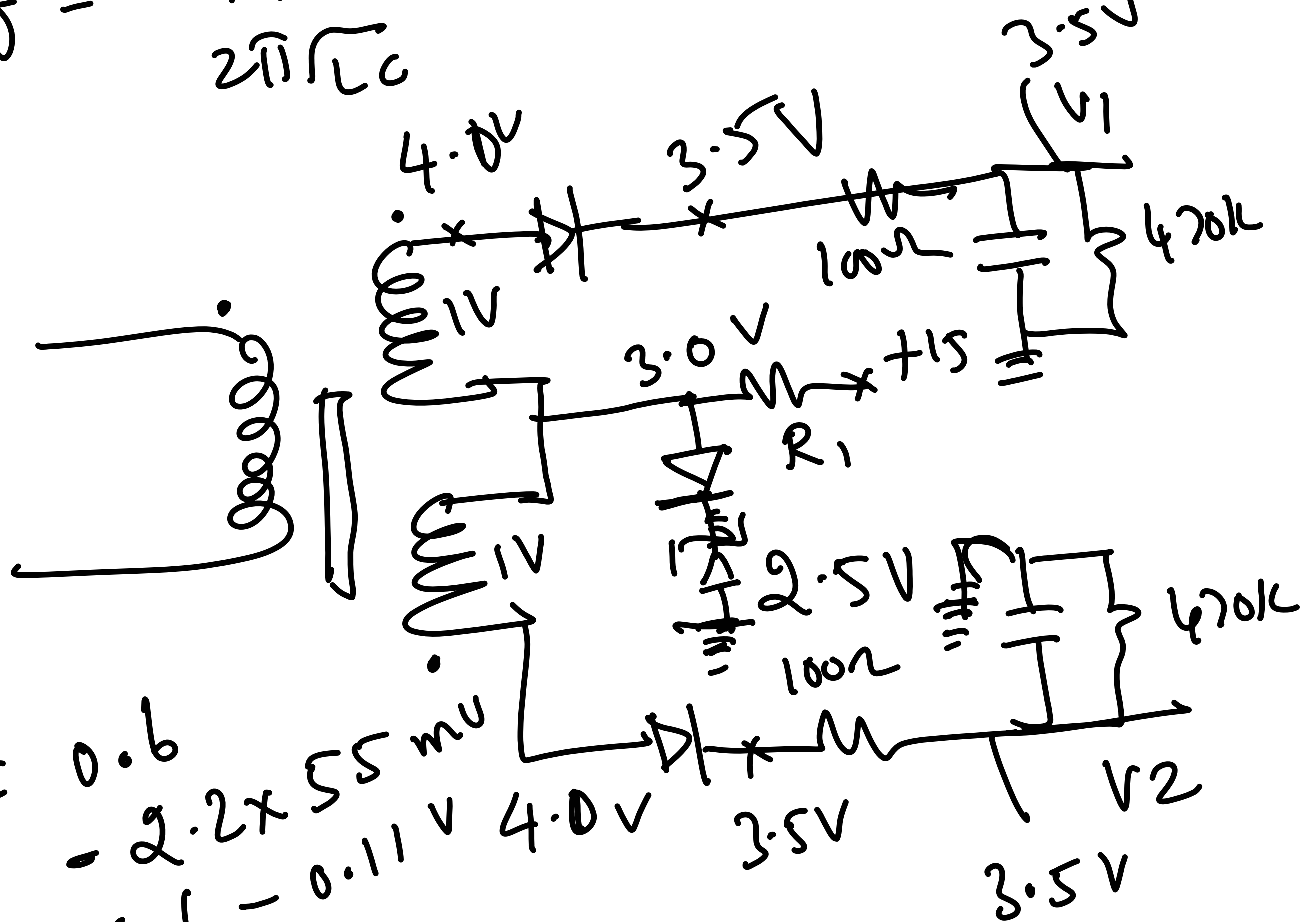
1. The emitter current increases
2. This produces increased mag field in the coil

(3) This increases base vol of the Transistor

(4) ~~This~~ Increase in base vol again increases the emitter current.

(5) ~~Again~~ mag field increases in the primary

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



wl acc
he diode = 0.6

$- 2.2 \times 55\ mV$

$= 0.6 - 0.11\ V$

$= 0.49V$

=

At room temp

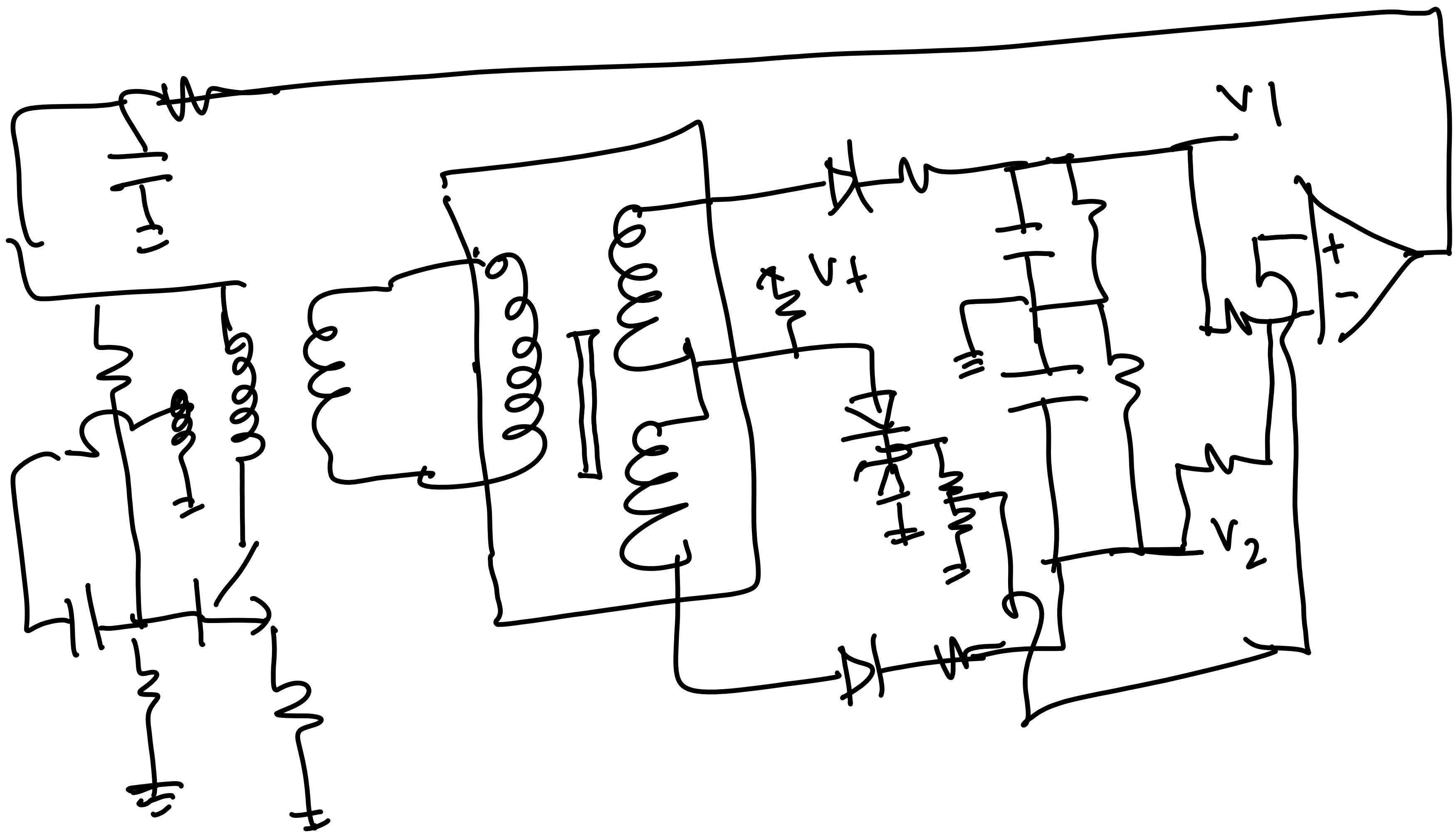
$V_1 = 3.5\,V$

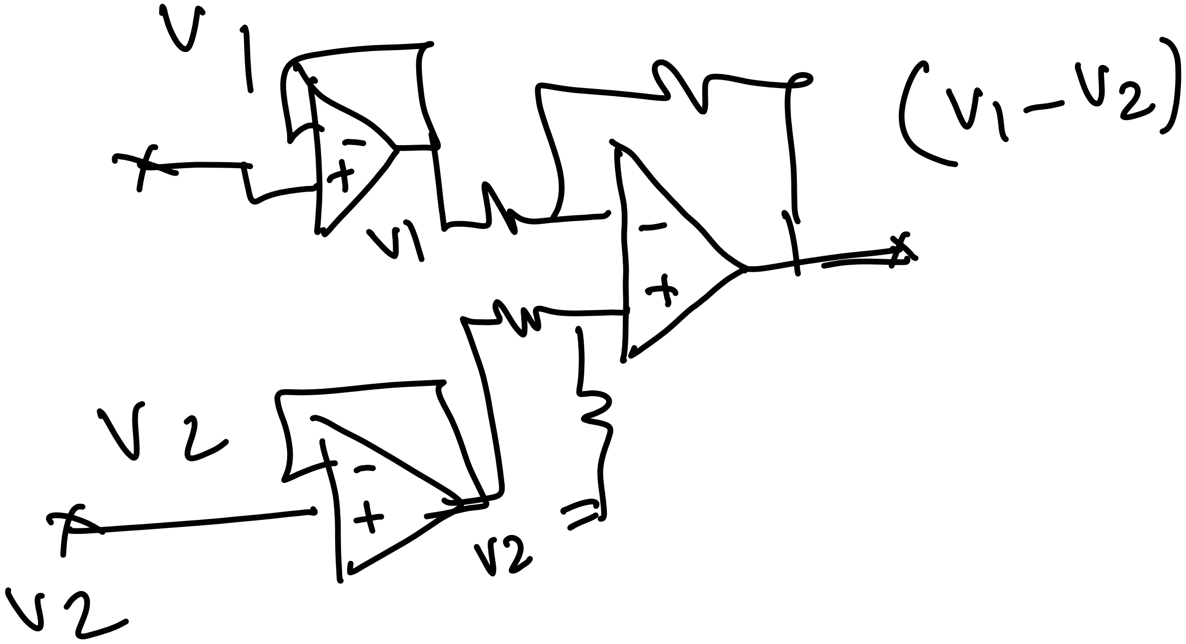
$V_2 = 3.5\,V$

At 80°C ambient temp

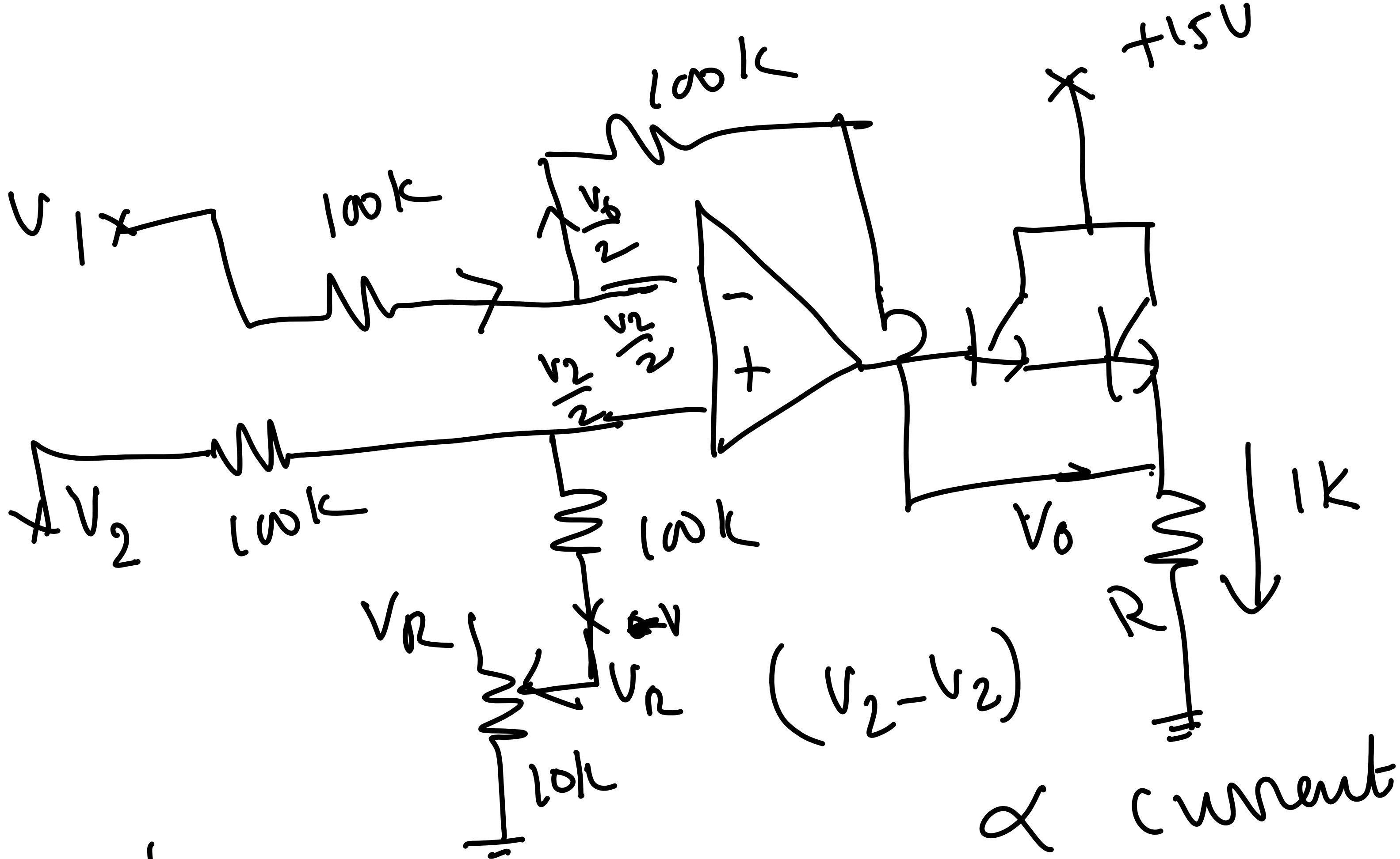
$V_1 = ? = 3.5\ V$

$V_2 = ? = 3.5$


V1
V+
+
-
V2



V 1
v1
(V1 - V2)
V 2
v2 =
V2

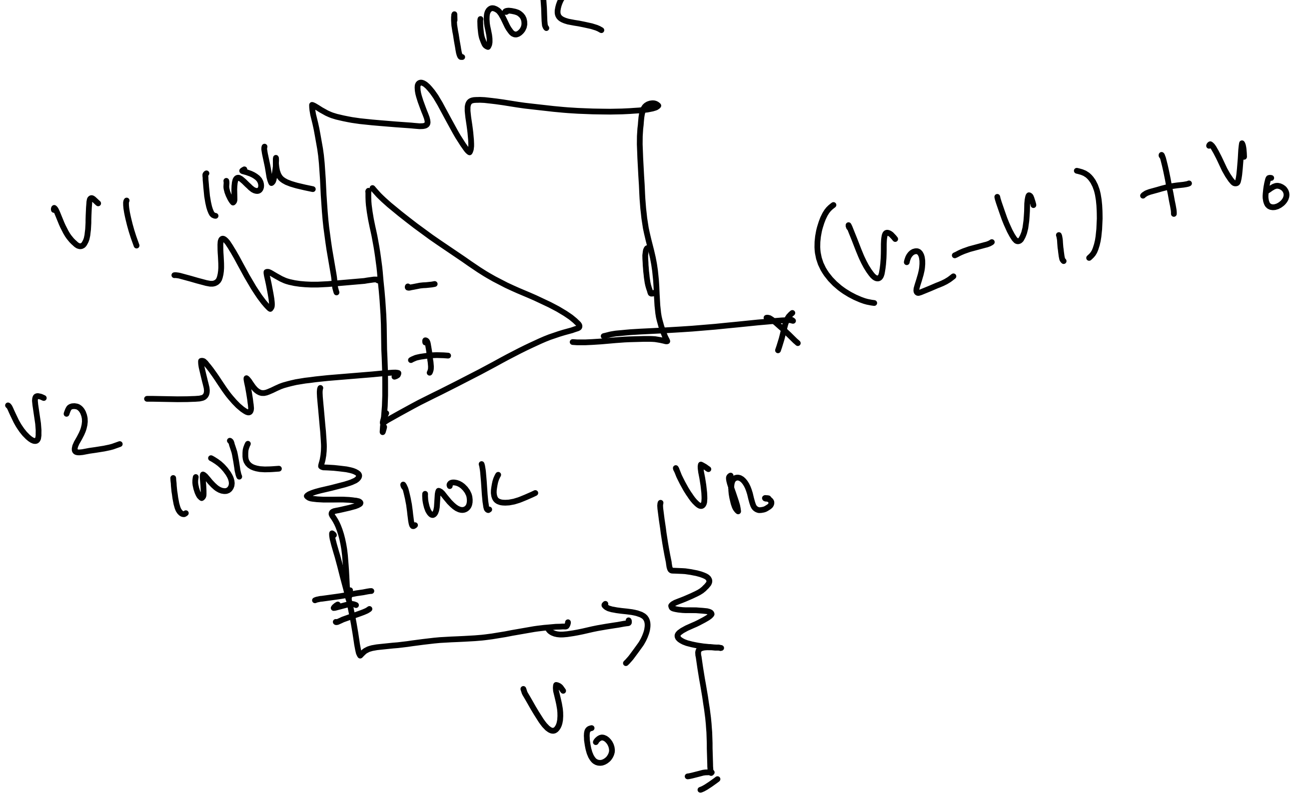


At inverting terminal $\frac{V_1}{2} + \frac{V_0}{2}$

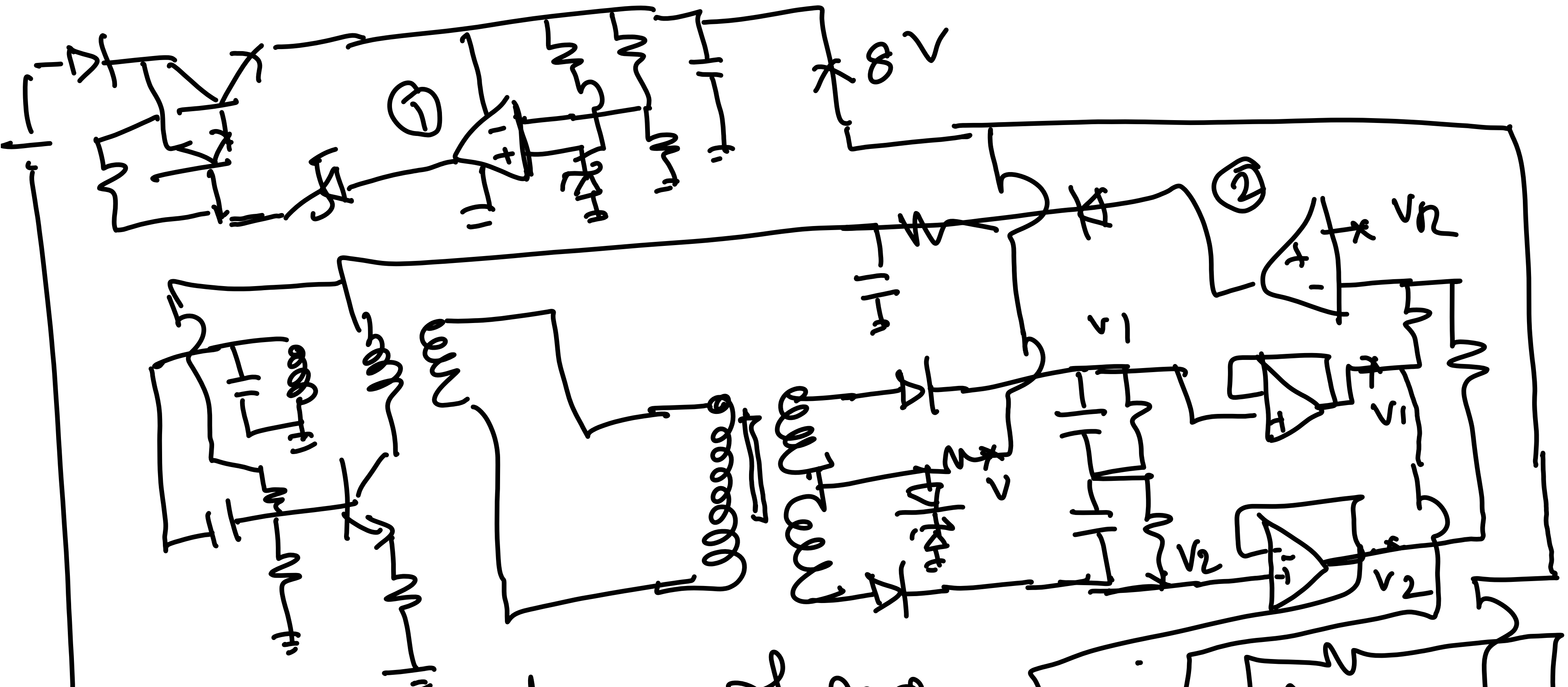
At + terminal $= \frac{V_2}{2} + \frac{V_R}{2}$

$$V_1 + V_0 = V_2 + V_R$$

$$V_2 - V_1 = V_0 - V_R$$


100k
V1
100k
V2
100k
100k
$(V_2 - V_1) + V_0$
V_n
V_0

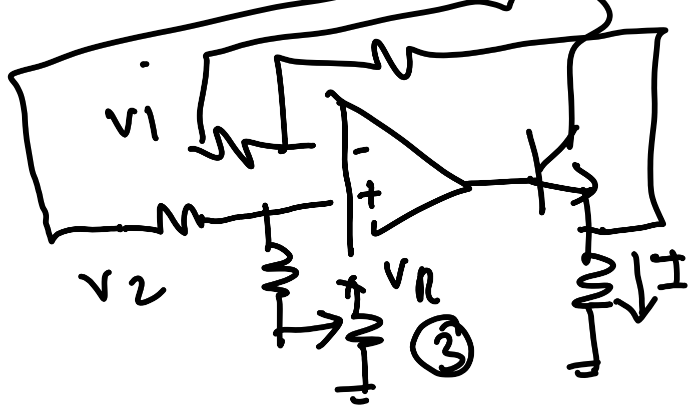
1. LVDT excitation using sine wave
2. Temperature compensation
3. $(V_1 - V_2) \propto$ displacement
4. $(V_1 - V_2)$ is converted into proportional current



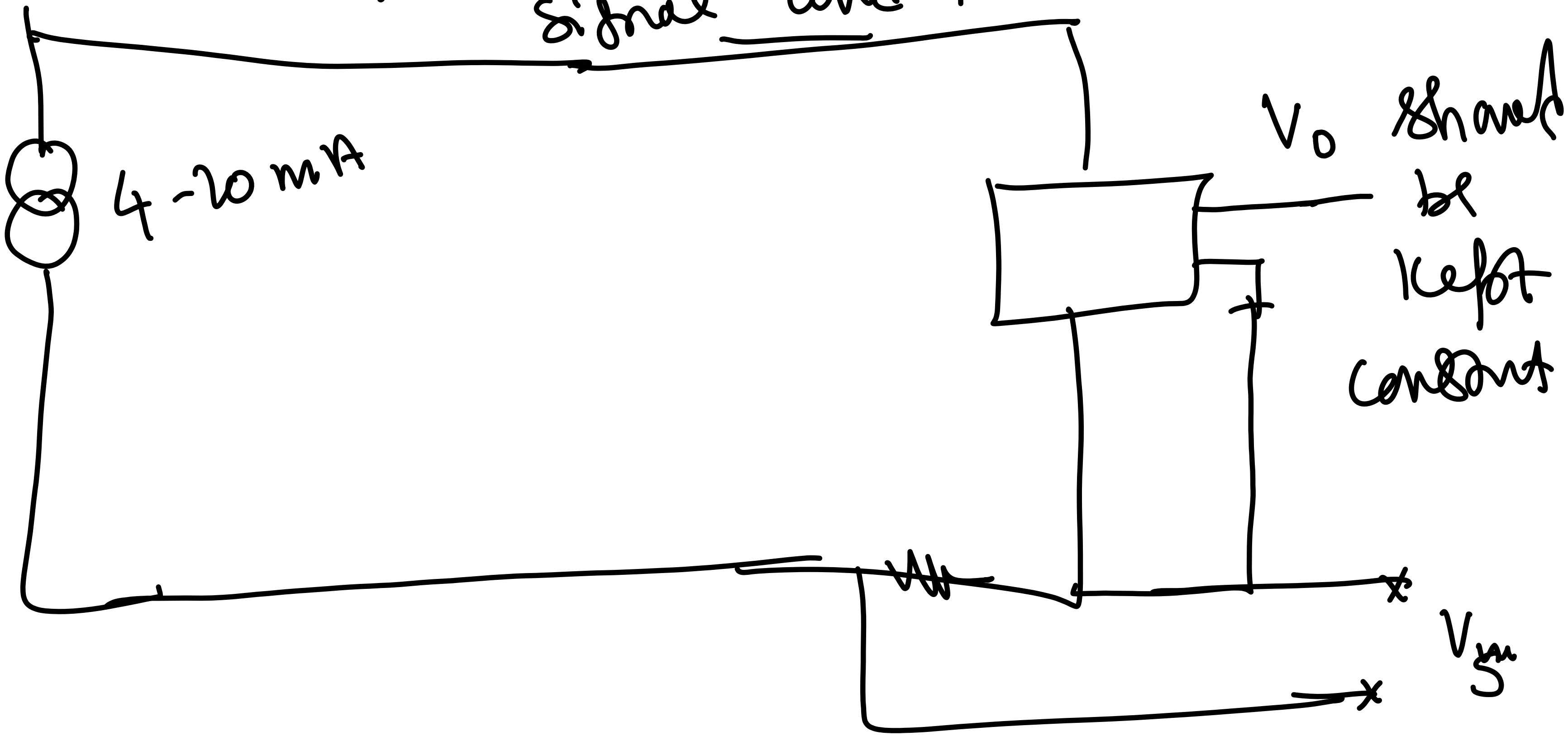
1 op-amp 1 → w/ reg

2 op-amp 2 → $(V_1 + V_2)$ constant

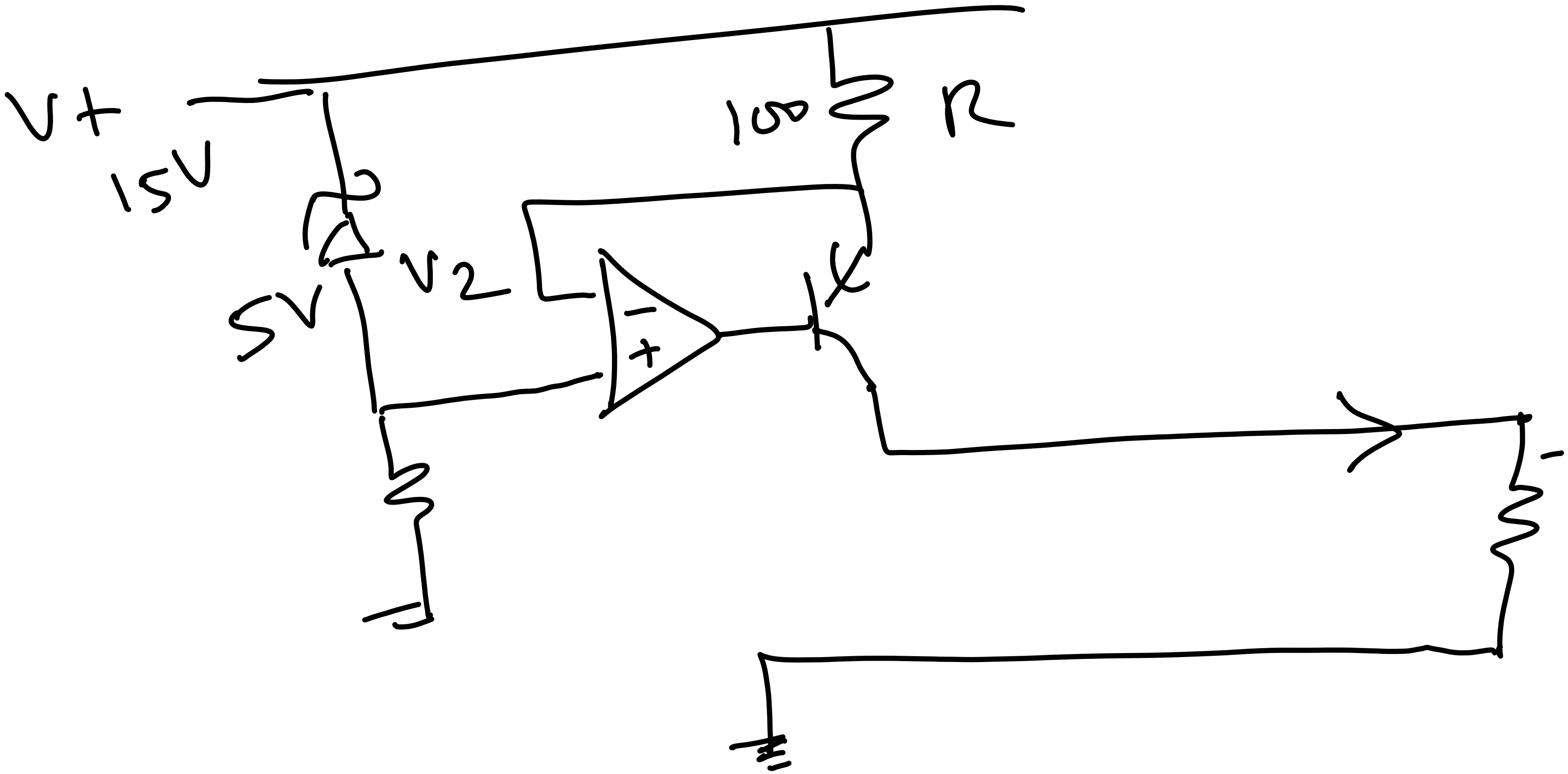
3 op-amp 3 → $(V_1 - V_2) \propto I$

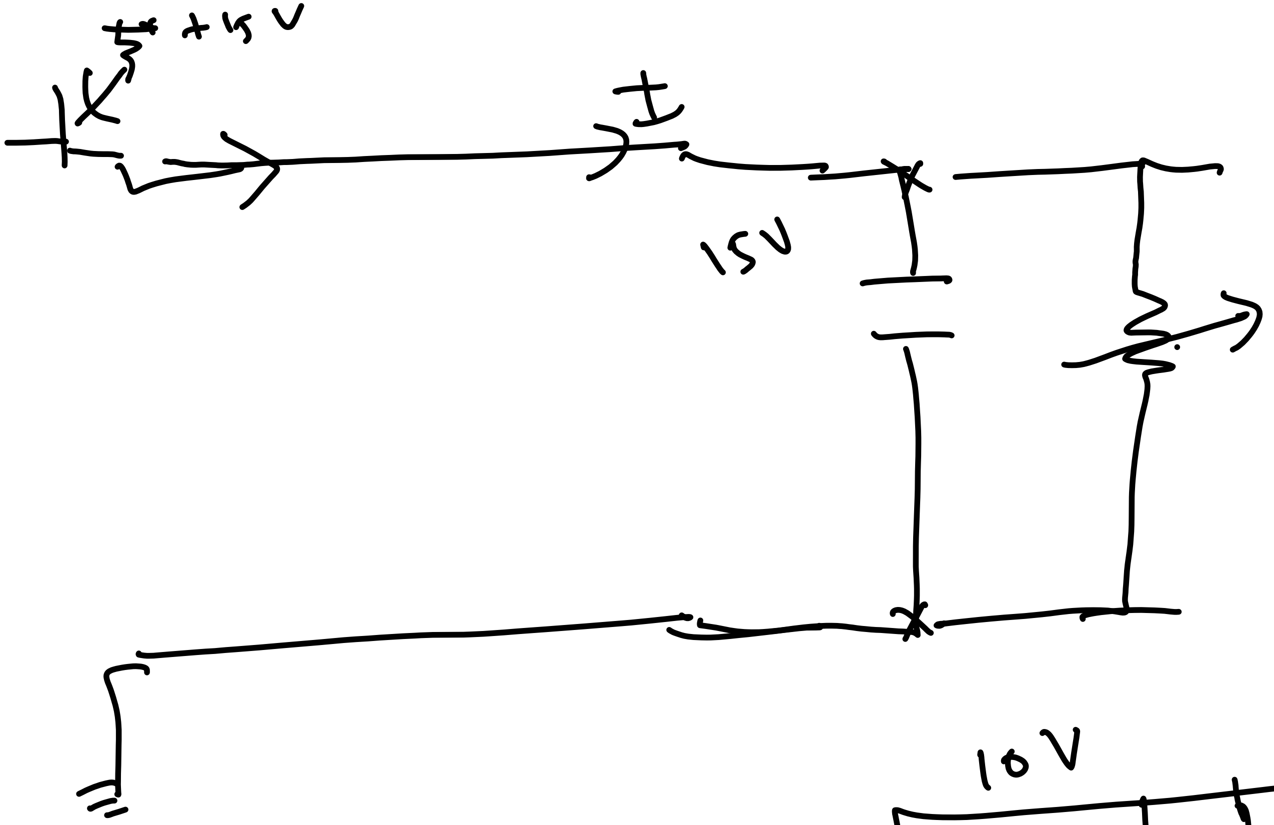


Spliting the current into signal and power

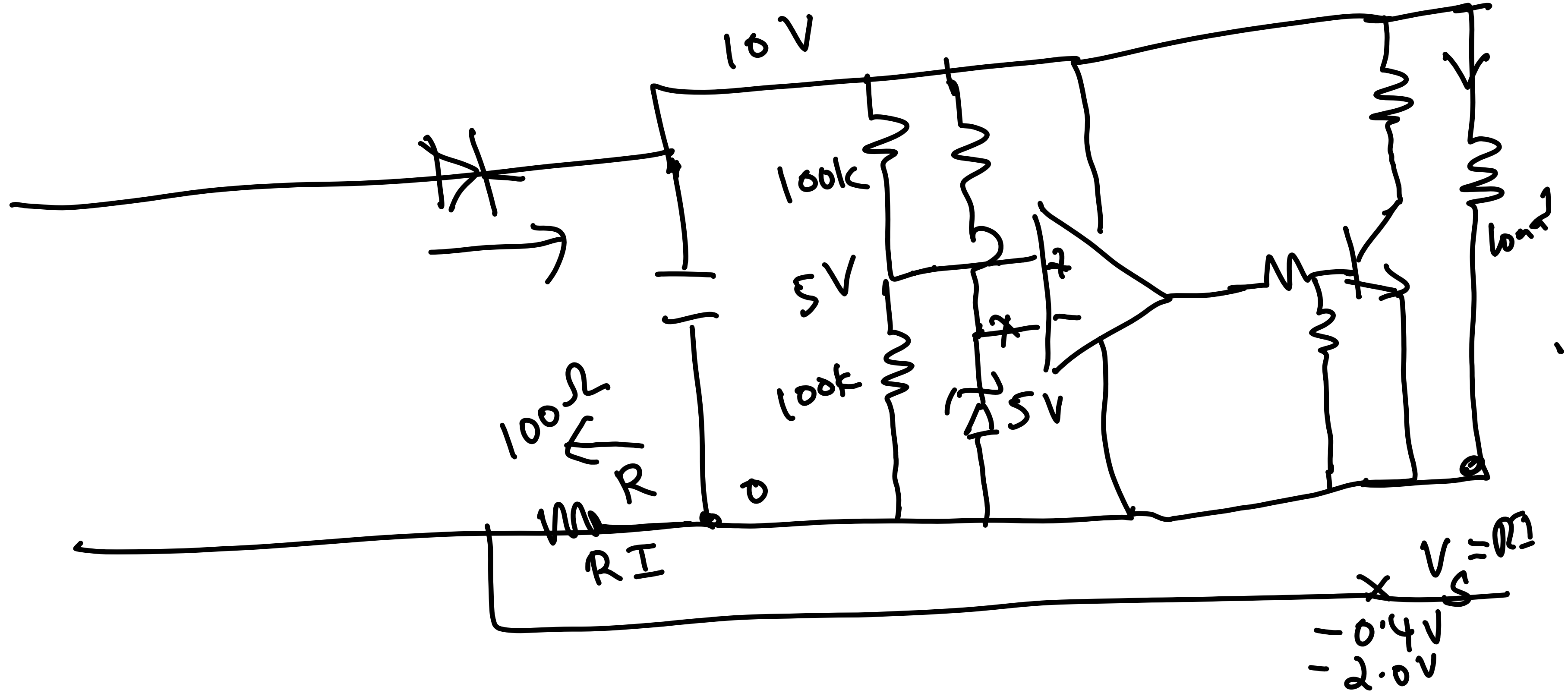


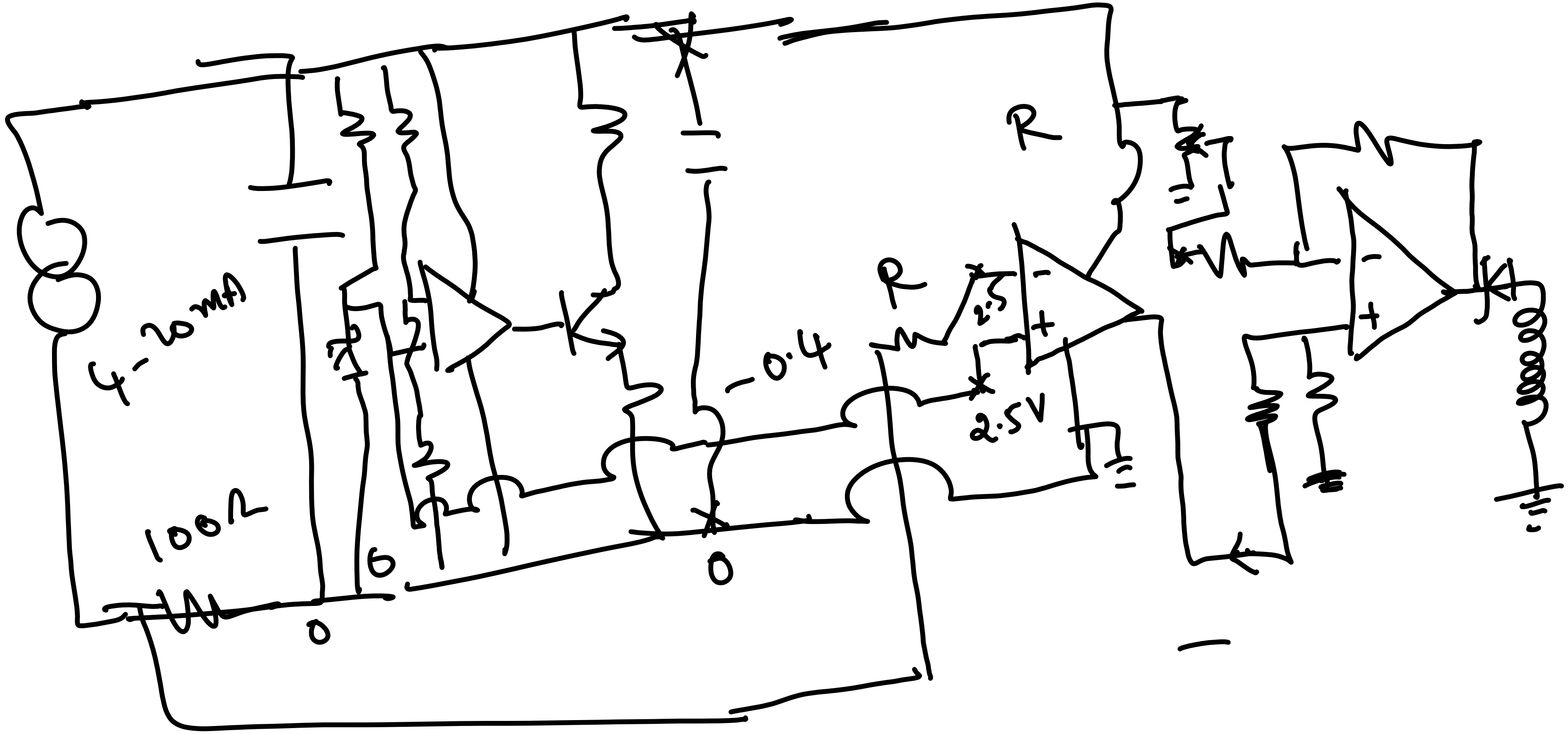
Constant current source



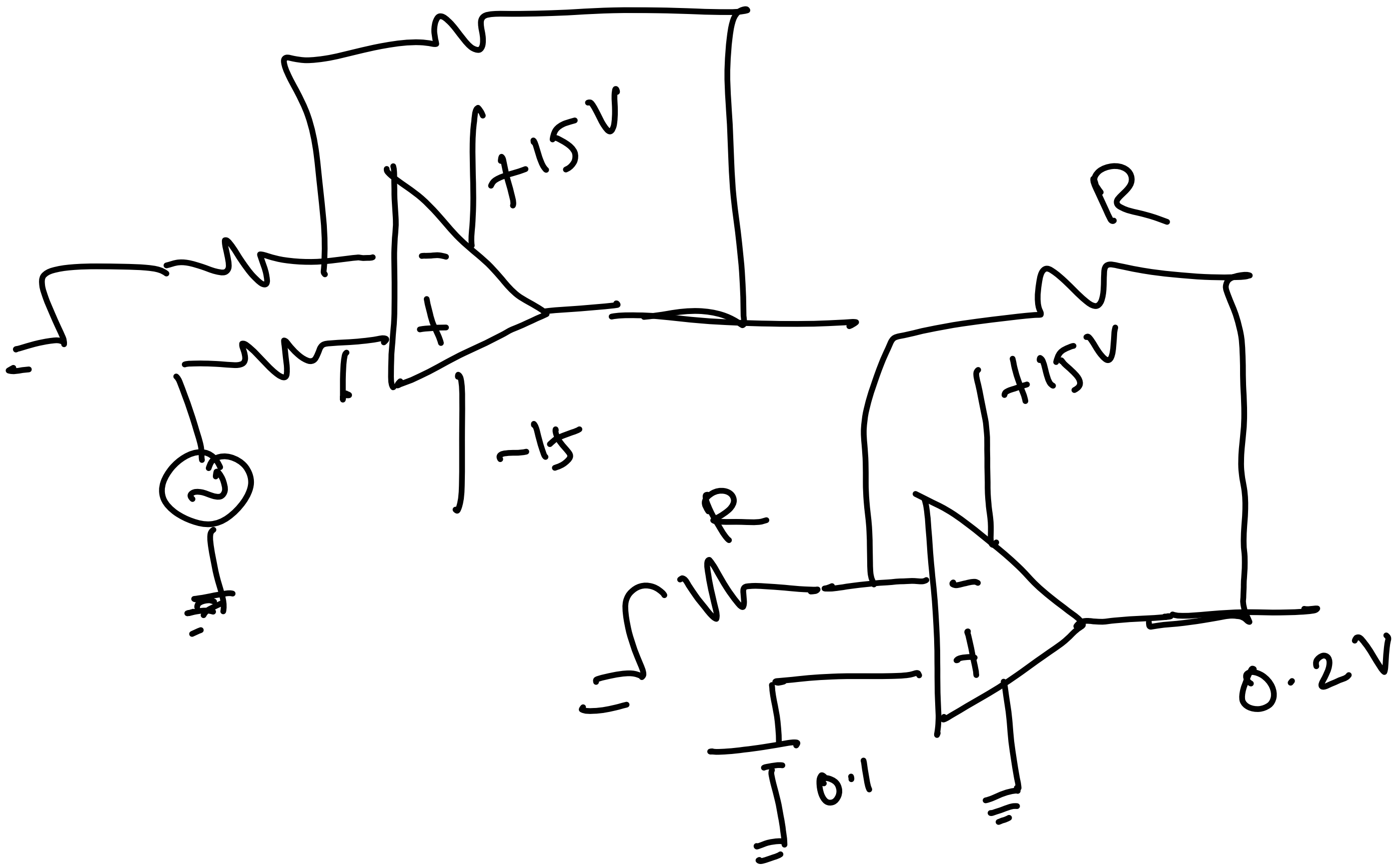


$$\frac{V_0}{R} = I$$

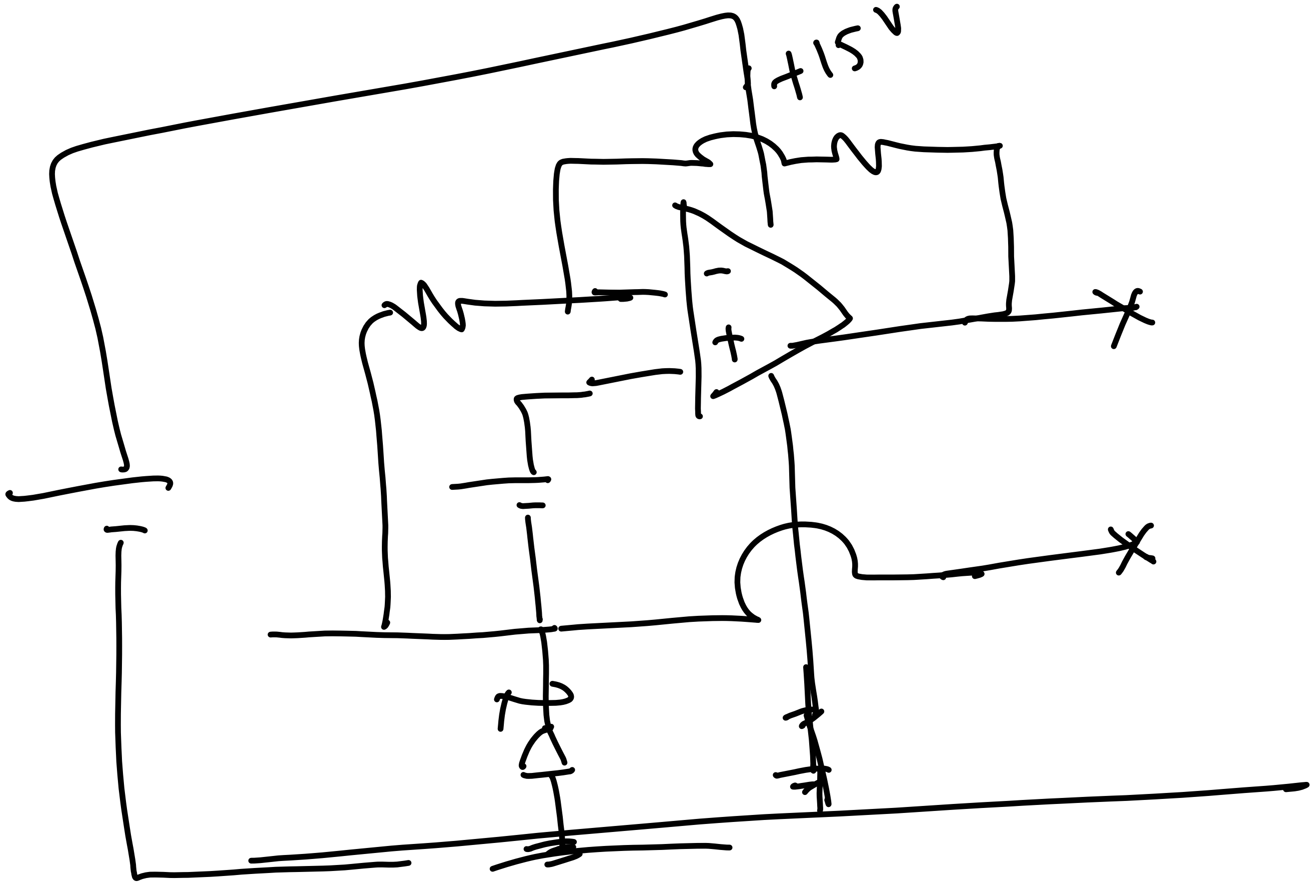




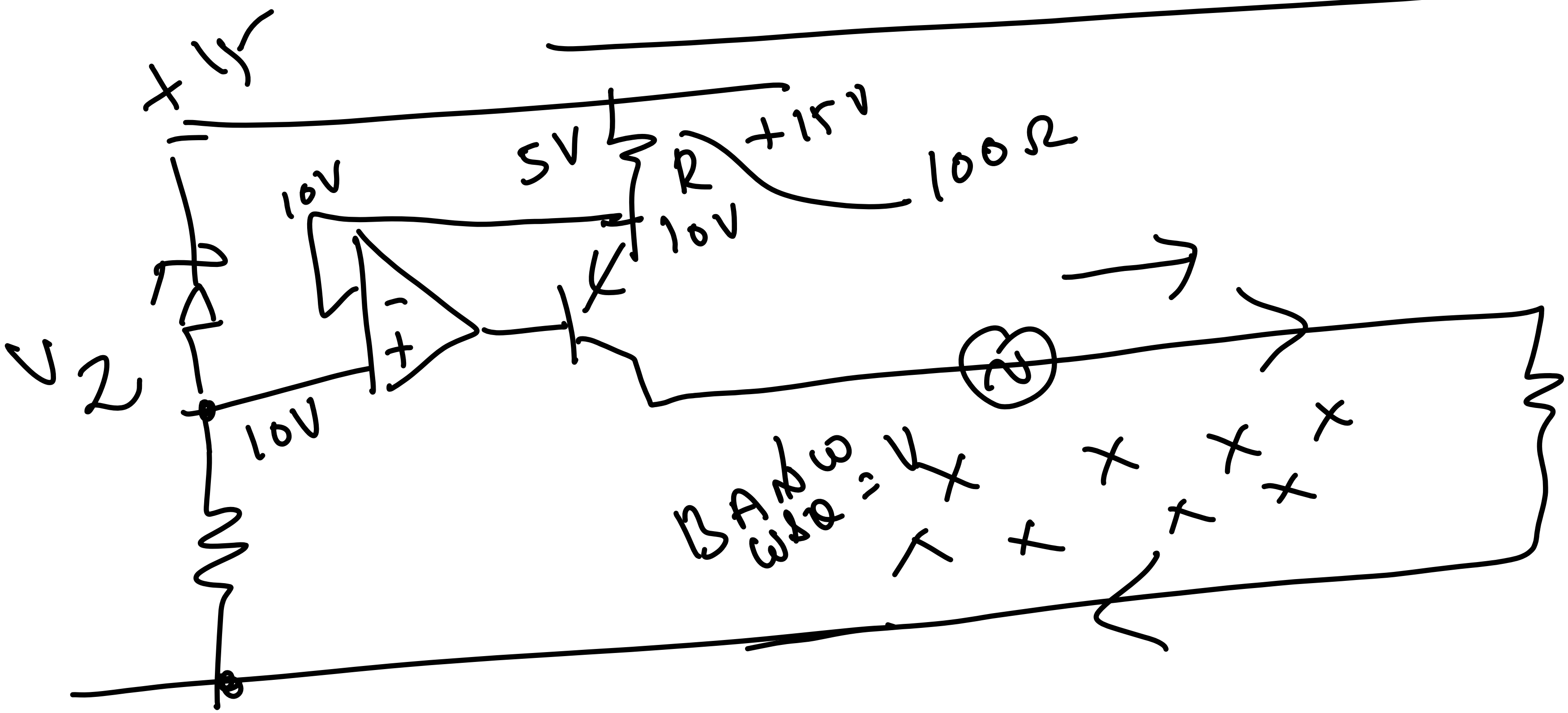
4-20mA
100Ω
0
6
0
-0.4
R
2.5
2.5V
R

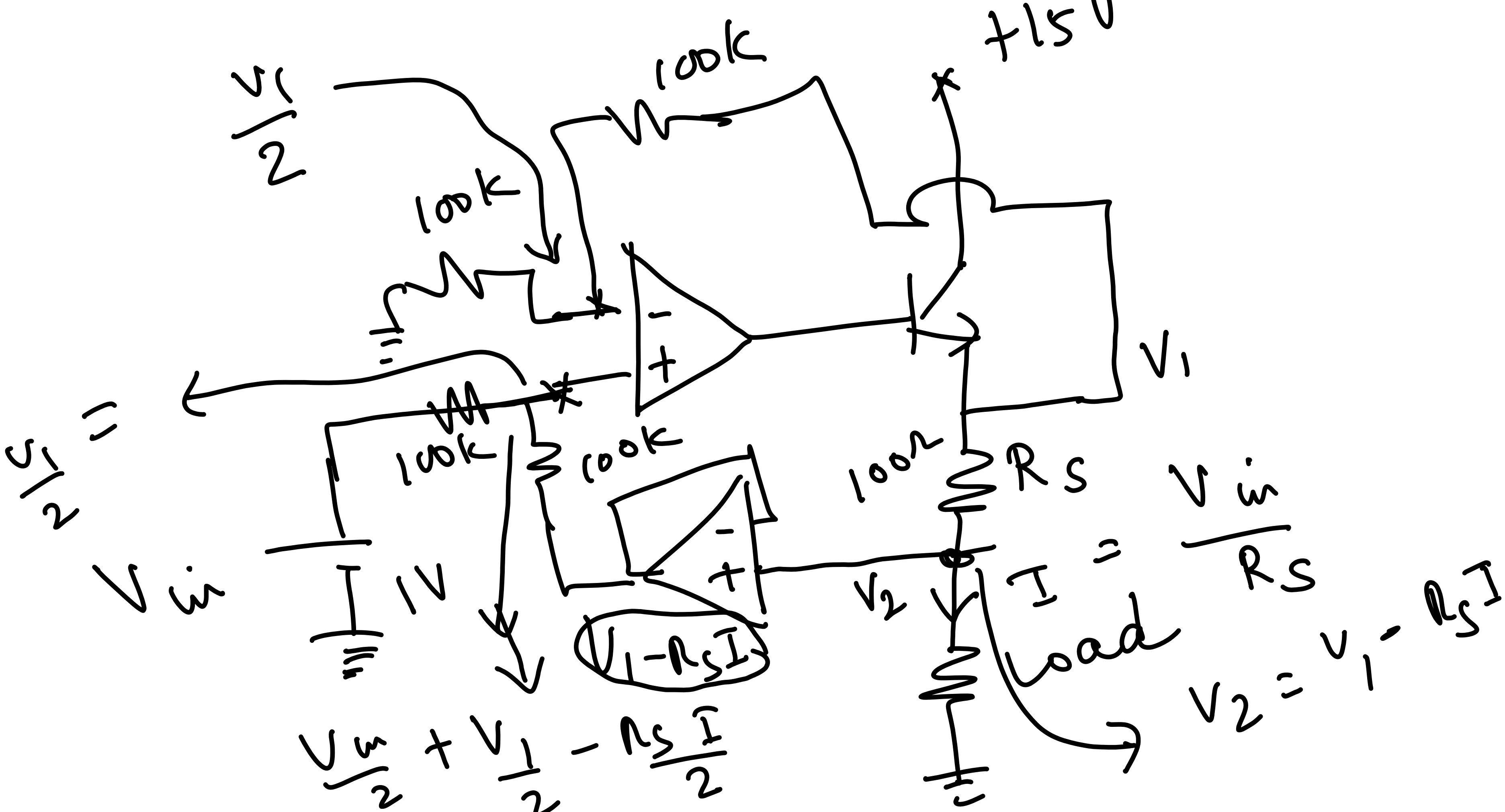


+15V
-15
R
+15V
R
0.1
0.2V


+15V

Constant current source





$$\cancel{\frac{V_1}{2}} = \frac{V_{in}}{2} + \cancel{\frac{V_1}{2}} - \frac{R_S I}{2}$$

$$R_s I = V_{in}$$

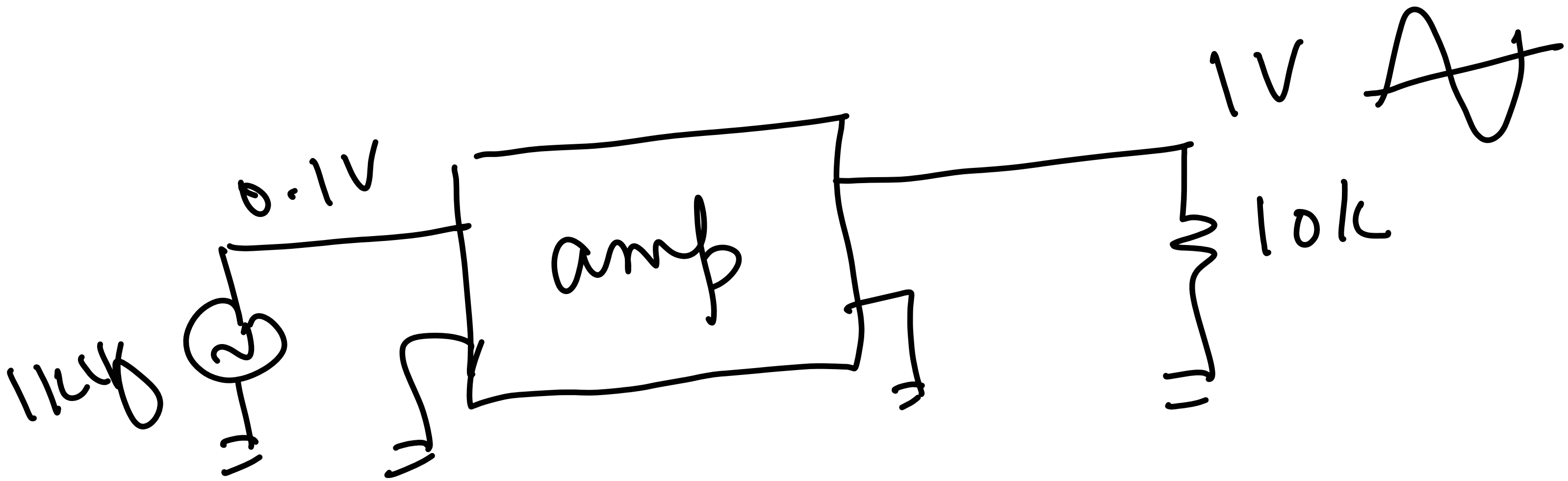
$$I = \frac{V_{in}}{R_s}$$

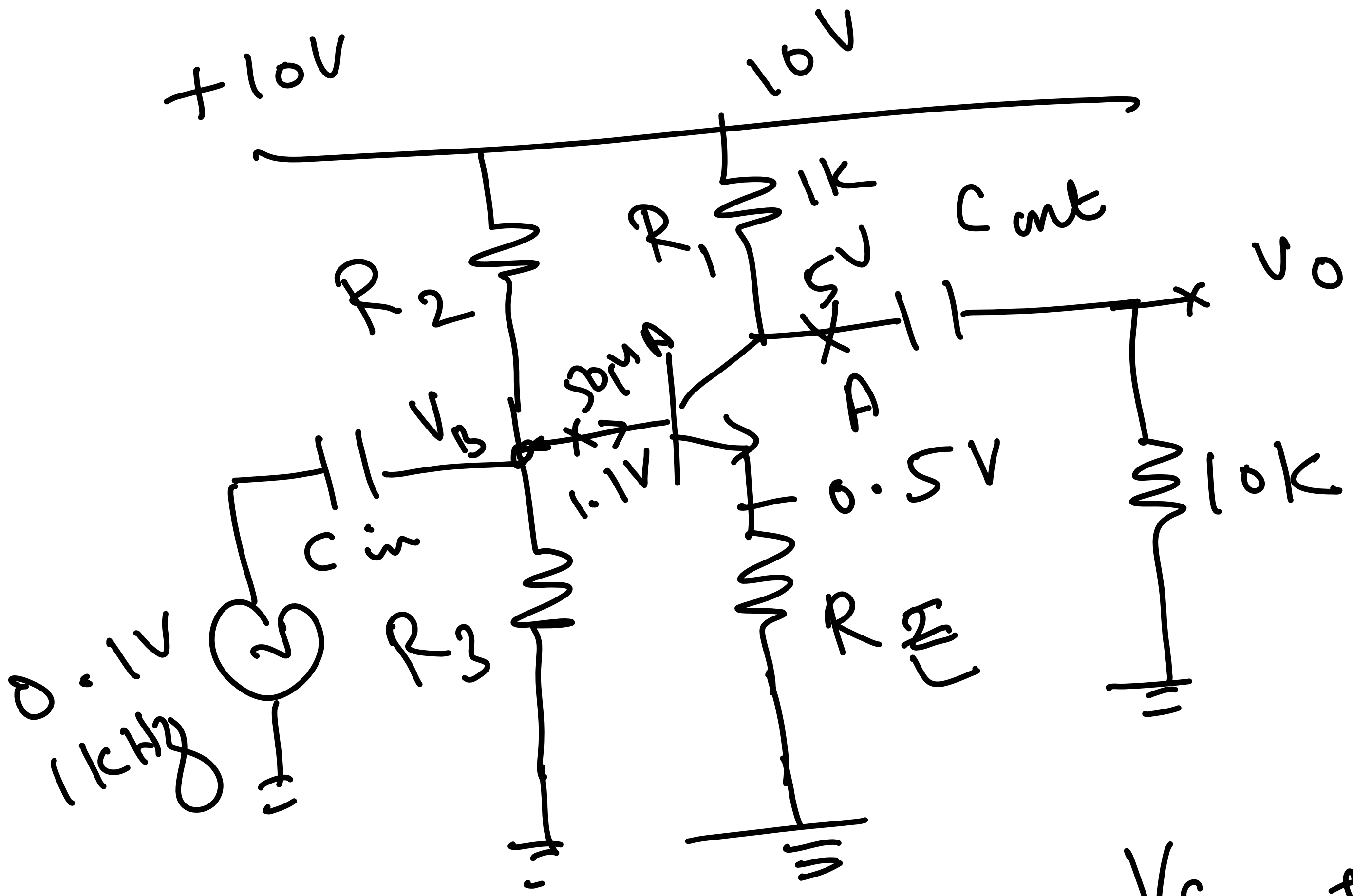
Application of transistors

(1) Transistor as an amplifier.

Spec : Input signal 0.1V
Amplification = 10
Fre = 1 kHz
Supply voltage = 10 V

Load resistance = 10k





Maintain A at $\frac{V_S}{2} = 5V$

when $V_{in} = 0$

fix R_1 — 10 times lower than load which is 10k

ie $R_1 = 1k$

$$\text{gain} = \frac{R_1}{R_E}$$

$R_E = 100\,\Omega$

What is the current in R_E?

current in R_1 = current in R_E

$$\text{current in } R_1 = \frac{(10-5)}{1k} = 5\text{ mA}$$

current through $R_E = 5mA$

voltage acc $R_E = 100 \times 5 \times 10^{-3}$

$= 0.5V$

So voltage at the base

$= 0.5V + 0.6V = 1.1V$

$$\frac{\text{base current}}{}$$

$$\frac{\text{Emitter current}}{\beta} = \text{base current}$$

base current $= \dfrac{5\times10^{-3}}{\beta} = \dfrac{5\,mA}{100}$

$= 50\,\mu A$

Current through R_2 and R_3

This is to be kept 10 times more than the base current

$= 500\,\mu A = 0.5\,mA$

voltage acc $R_3 = 1.1\ V$

current through $R_3 = 0.5\ mA$

So $R_3 = \dfrac{1.1}{0.5\ mA} = 2.2k$

vol acc $R_2 = 10 - 1.1 = 8.9\ V$

current through $R_2 = 0.5\ mA$

So $R_2 = \dfrac{8.9}{0.5 \times 10^{-3}} = 17.8k$

$\approx 18k$

$R_1 = 1k$

$R_E = 100\Omega$

$R_2 = 18k$

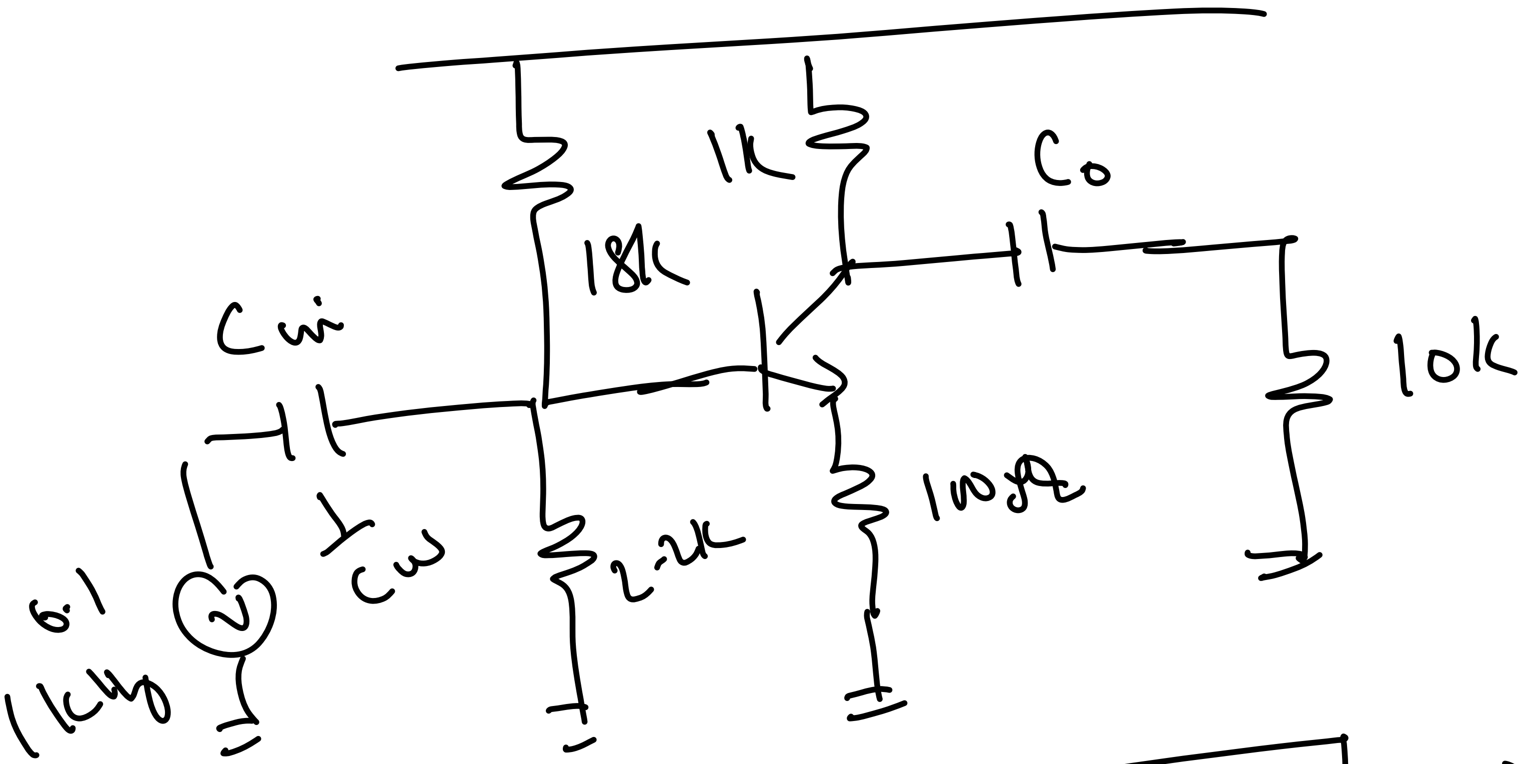
$R_3 = 2.2k$

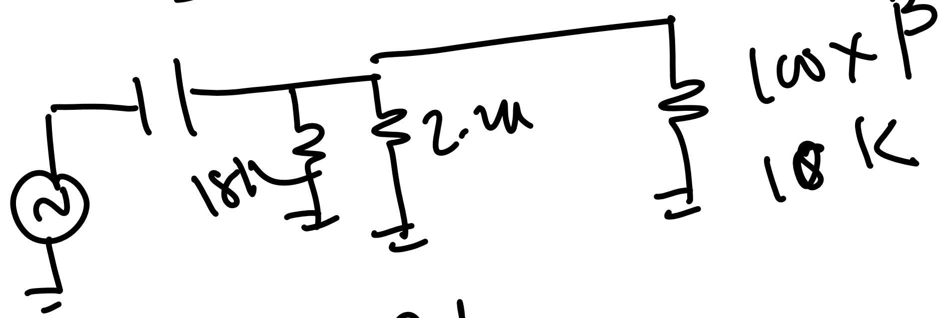
$C_{in} = ?$

$C_{out} = ?$

INPUT impedance of the amplifier

$= R_2 || R_3 || 100\Omega \times \beta$





= 2k

Input impedance is 2k

$$\frac{1}{C\omega} = 200$$

$$\omega = 2\pi \times 10^{3}$$

$$\frac{1}{200 \times \omega} = C_{in}$$

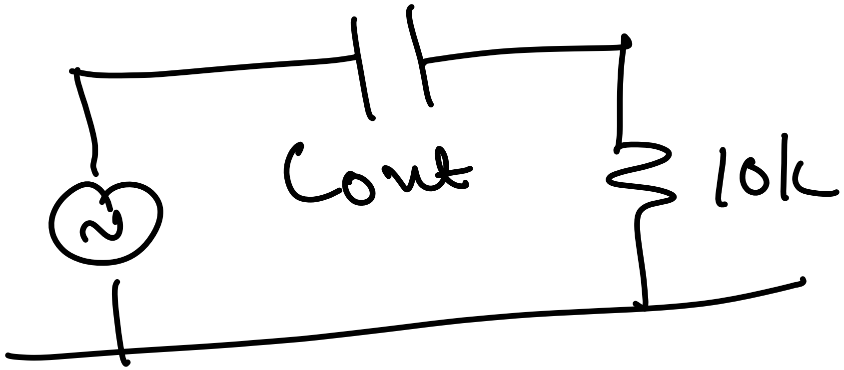
$$\frac{1}{200 \times 2\pi \times 10^{3}} = C_{in}$$

$$\frac{10^{-5}}{2 \times 2\pi} = \frac{10^{-5}}{4\pi} = 10^{-6}$$

$$= 1\,\mu F$$

$C_{in} \cong 1\mu F$

Calculation for Cout



Impedance of Cout = 1k

$$\frac{1}{C\omega} = 10^3$$

$$\omega = 2\pi \times 10^3$$

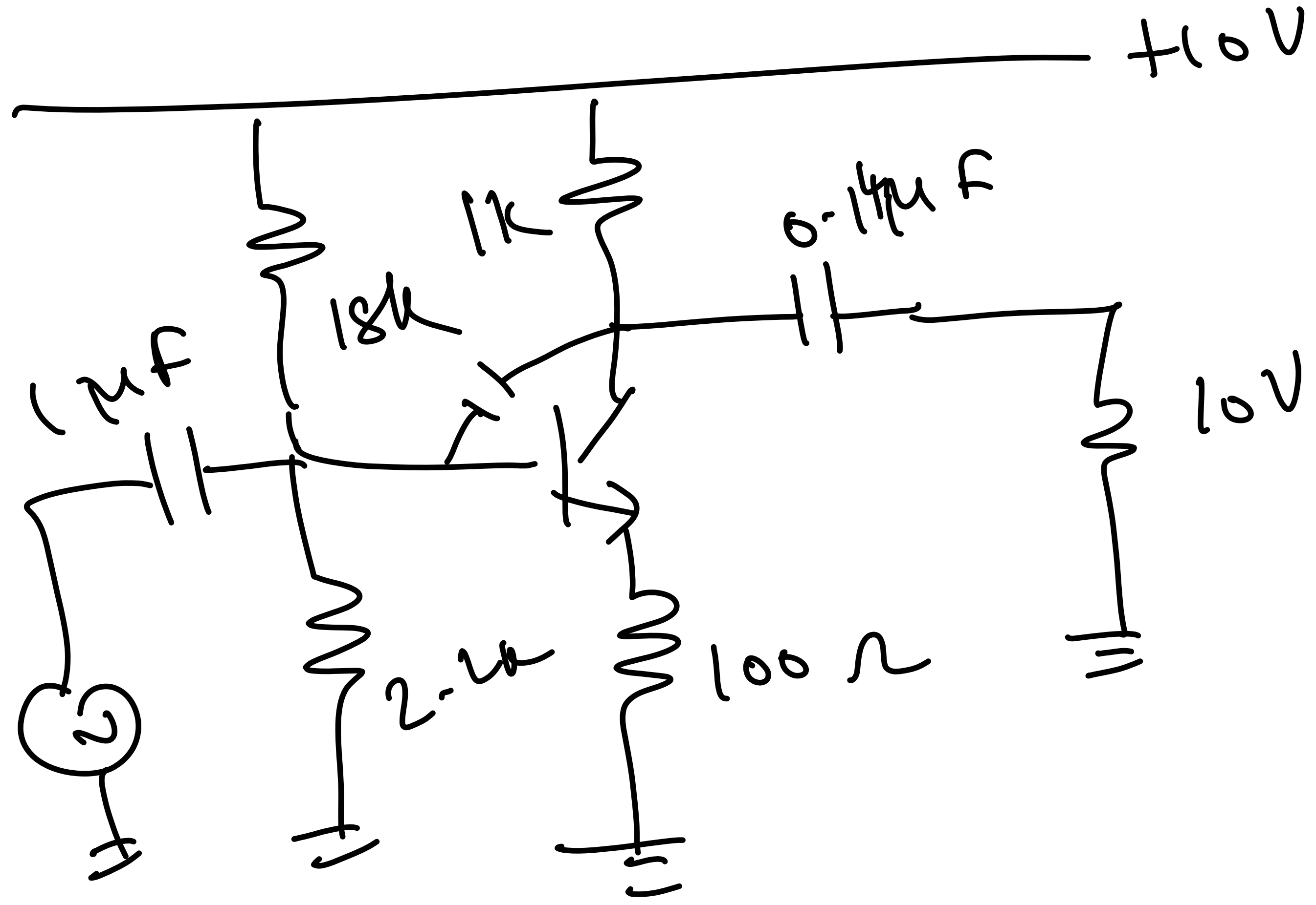
$$\frac{1}{C \times 2\pi \times 10^3} = 10^3$$

$$\frac{1}{2\pi \times 10^6} = C_{out}$$

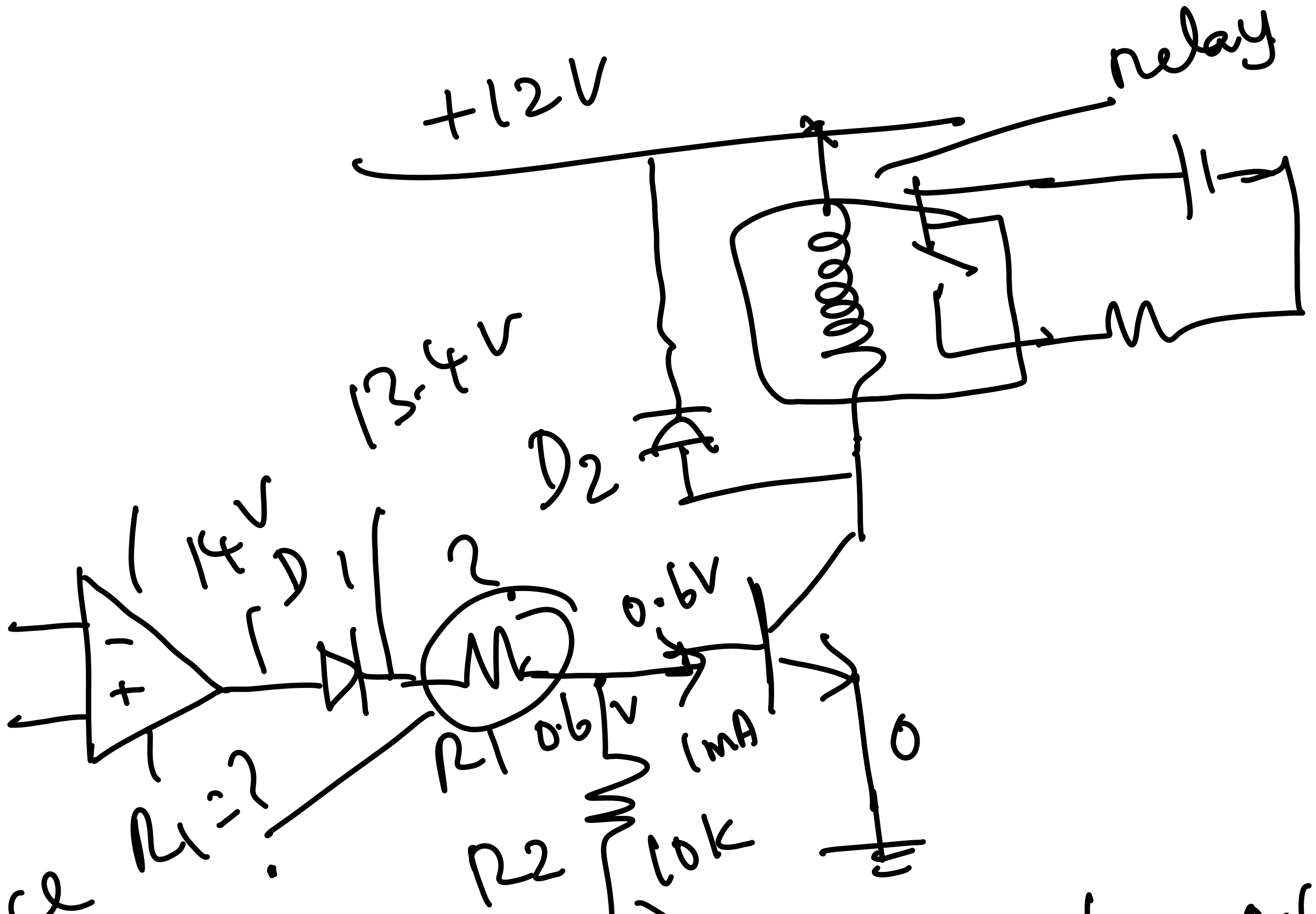
$$\frac{10^{-6}}{2 \times 3} = 10^{-7} \times \frac{10}{6}$$

$$= 1.4 \times 10^{-7}$$

$$C_{out} = 0.14\ \mu F$$


+10V
18k
1k
0.1μF
1μF
10V
2.2k
100 Ω

How to design a Relay driver using transistor



wl ace $R_1 = ?$

$$\Rightarrow I = \frac{0.6}{10^4} = 0.6 \times 10^{-4} = 6 \times 10^{-5}\ A$$

$$= 60\ \mu A = 0.06\ mA$$

V_0 of op amp is +15

Relay should be ON

V_0 of op amp is −15

Relay should be OFF

Collector current of the transistor

Collector current = Relay current

Let relay current

be = 100 mA

$$I_c = 100\ mA$$

$$I_b = \frac{I_c}{\beta} = \frac{100\ mA}{100}$$

$$= 1\ mA$$

base current = 1 mA

For base emitter cap of

100 pF

For $R_2 = 10k$

Then off time

$= 10^{-10} \times 10^{4} = 10^{-6}$ sec

$R_2 = 10k$

Current through R_2

$= \frac{0.6}{10k} = 0.06\,mA$

Total current through R_1

$= 1\,mA + 0.06\,mA$

$= 1.06\,mA$

Vol acc $R_1 = (15 - 1.2V - 0.6 - 0.6)$

$= 15 - 2.4$

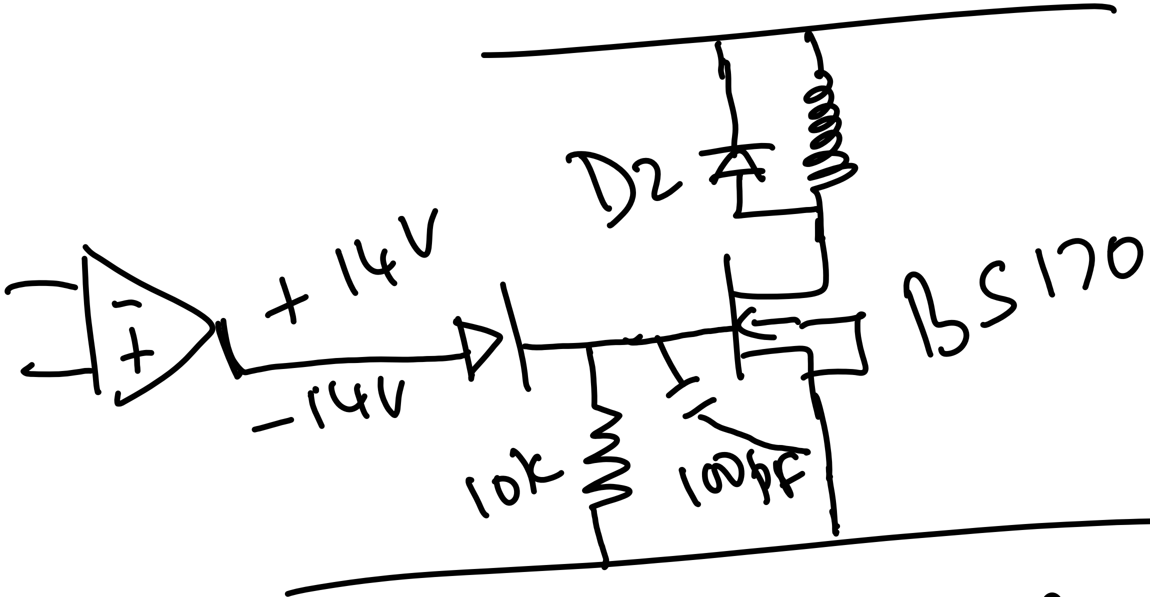
$= 12.6V$

Current through $R_1 = 1.06\ mA$

$$R_1 = \frac{12.4}{1.06\ mA} \simeq 12K$$

$$R_1 = 10K$$

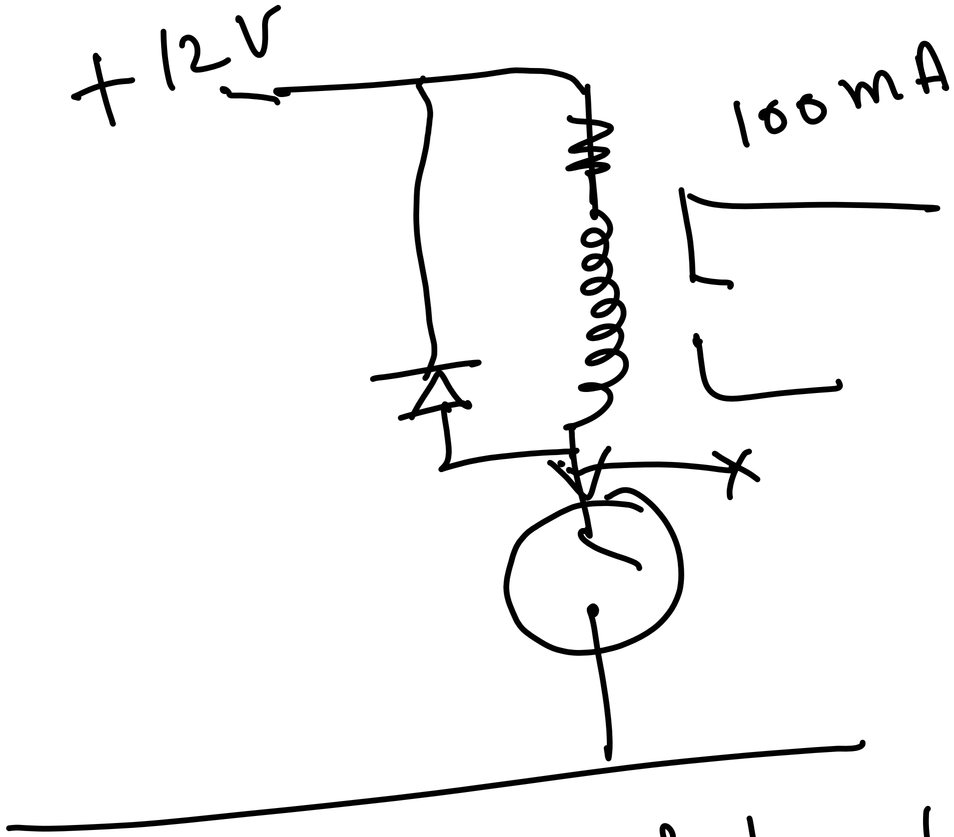
Relay drive using a MOSFET



$RC = 10^{4} \times 10^{-10}$

$= 10^{-6}$

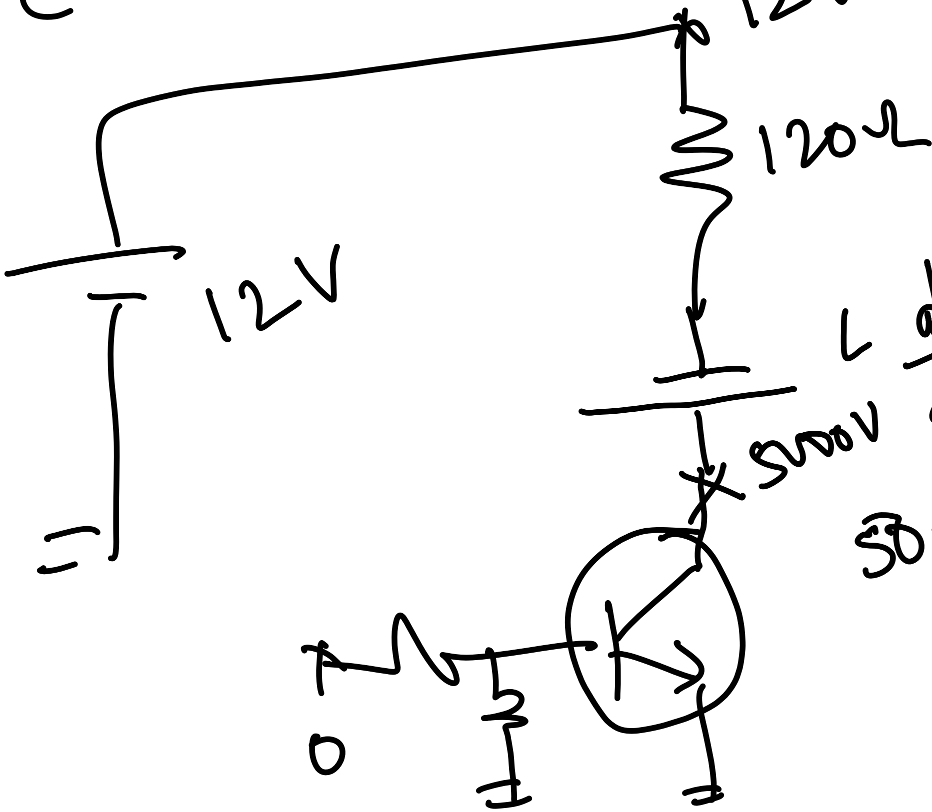
What is the use of D_2 ?



The resistance of the coil

$$= \frac{12}{100\text{mA}} = 120\,\Omega$$

$R_C = 120\,\Omega$



$L\dfrac{di}{dt}$

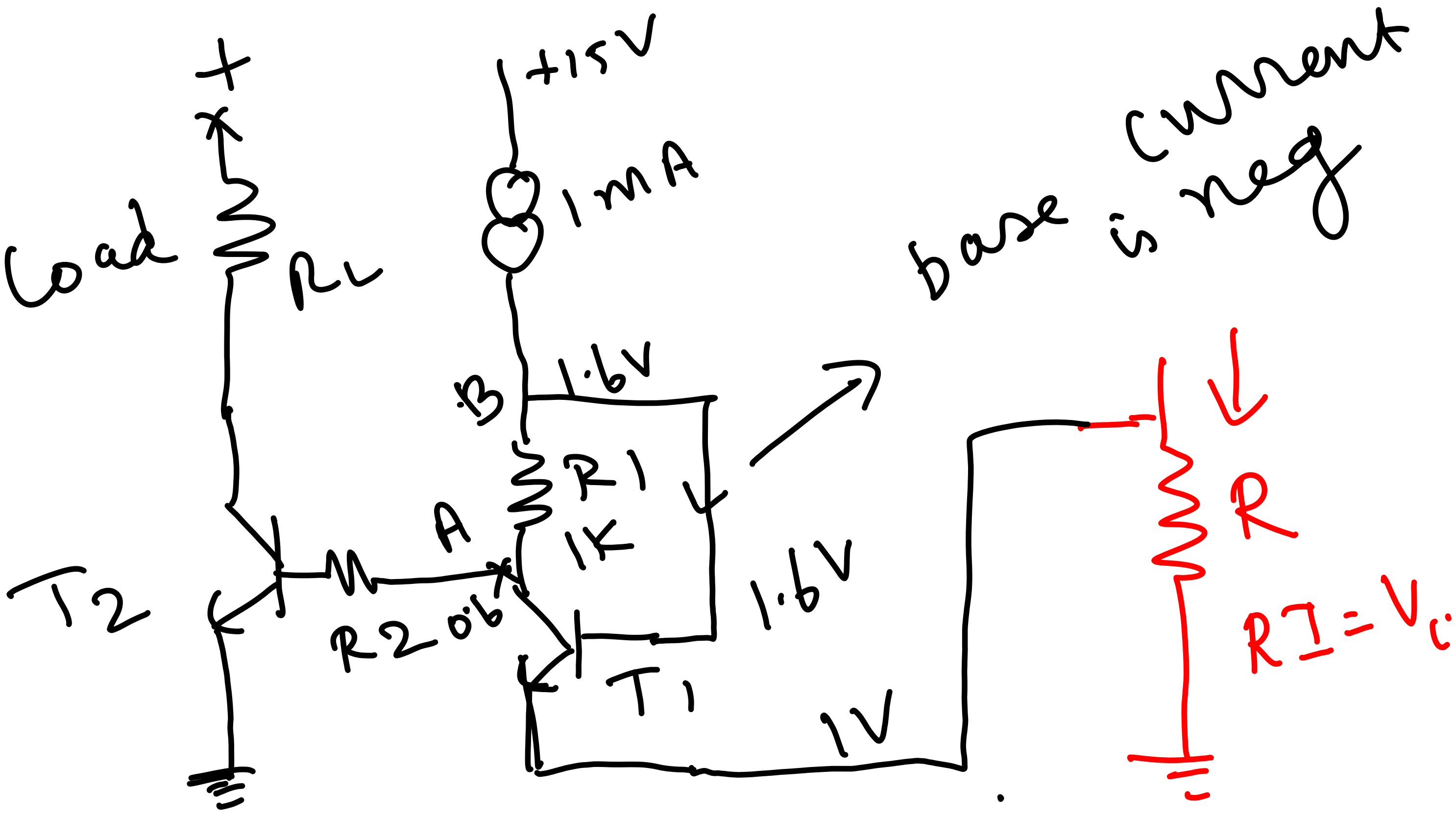
$50\times10^{-3}\times\dfrac{di}{dt}$

$dt = 1\,\mu s$

$di = 10^{-1}\,A = 100\,mA$

$$L\frac{di}{dt} = \frac{50\times10^{-3}\times10^{-1}}{10^{-6}}$$

$$= 50\times10^{2} = 5000\,V$$



① If the current in R is more than set value the load to be R_L switched off

Sol acc $R_1 = 1 \times 10^{-3} \times R1$

$R_1 = 1K$

Vol acc $R_1 = 10^{-3} \times 10^{3}$

$= 1V$

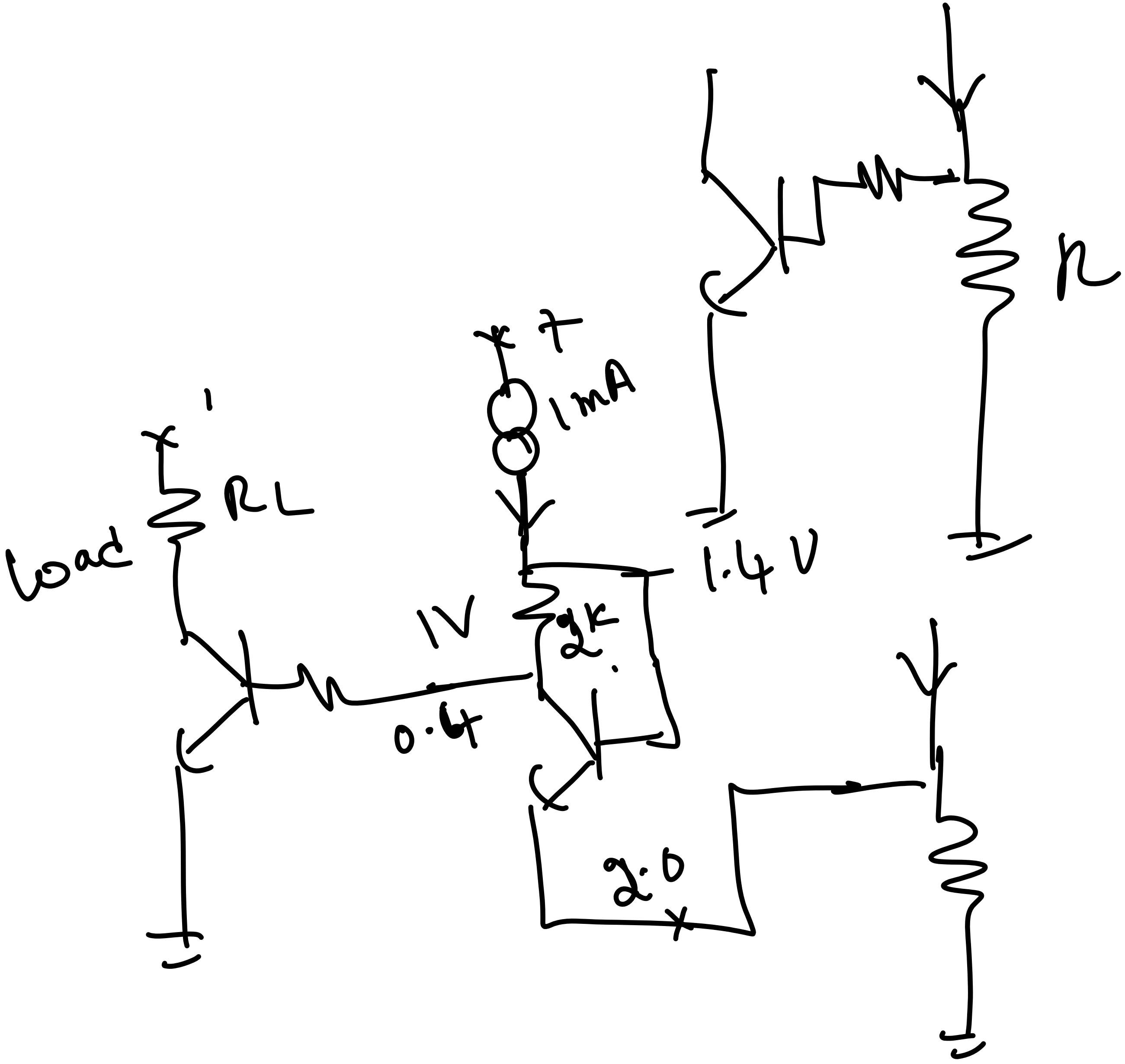
voltage at A to switch on Transistor $T_2 = 0.6V$

voltage at B

$= 1k \times 10^{-3} + 0.6 = 1.6V$

Then ~~emitter~~ emitter of $T_1 = 1.0V$

ie only when the voltage across R is $> 1V$ the transistor T_2 will be ON

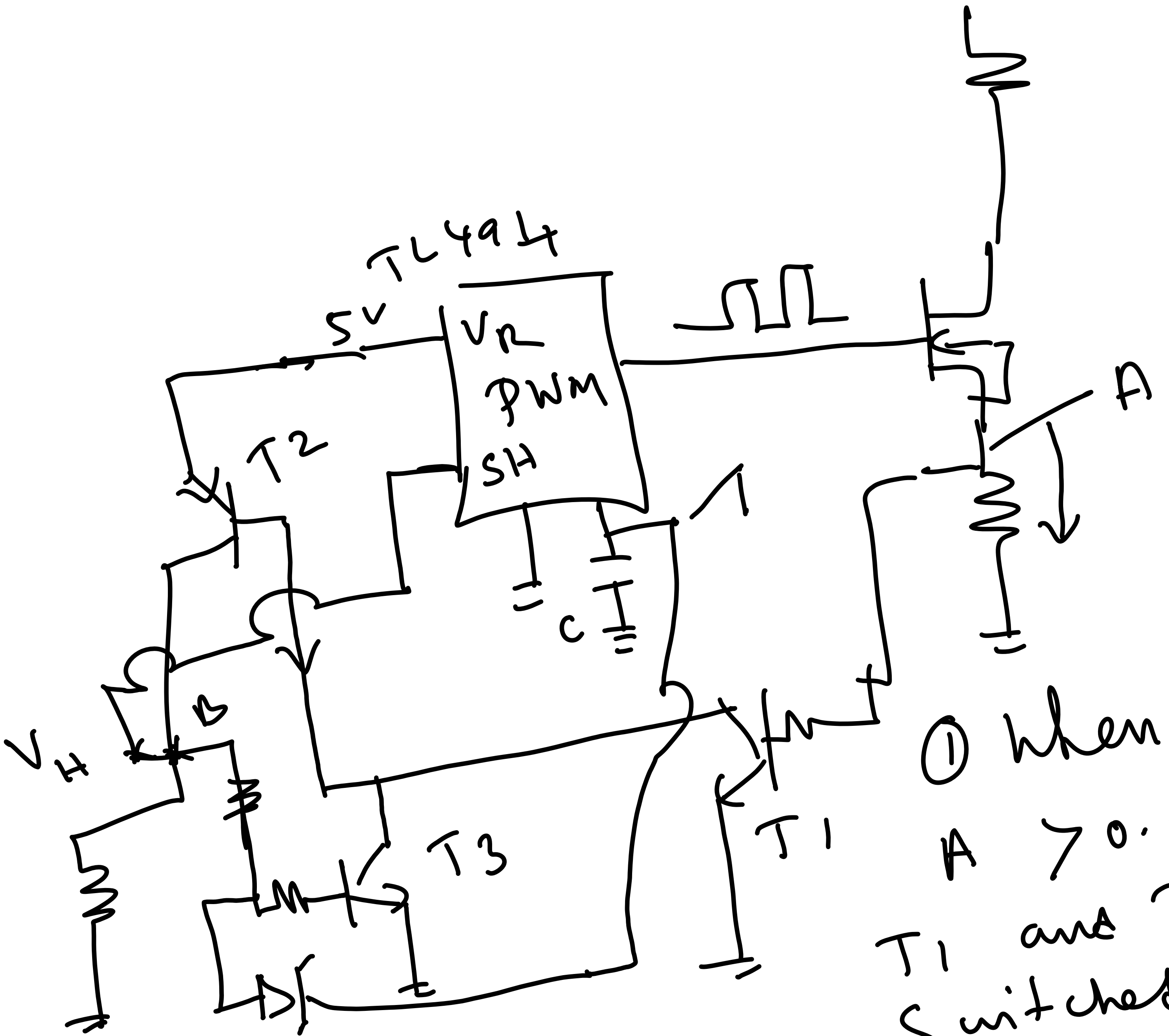


load
RL
1mA
1V
2k
0.6
1.4V
2.0
R

So with temp the
switching vol will not
change.
V_{BE} change in T_2 is
compensated by V_{BE} change in T_1

Switching voltage can be changed by changing the Resistance R_1

Transistor as a Latch



① When Vol at A > 0.6V then T1 and T2 are switched ON

So Vol appears at B

③ Then T_3 is Switched ON

When vol at A goes more than 0.6

T_1, T_2 and T_3 are ON

So PWM IC is switched OFF

④ So vol at A = 0

But T_3 will be continuously ON.

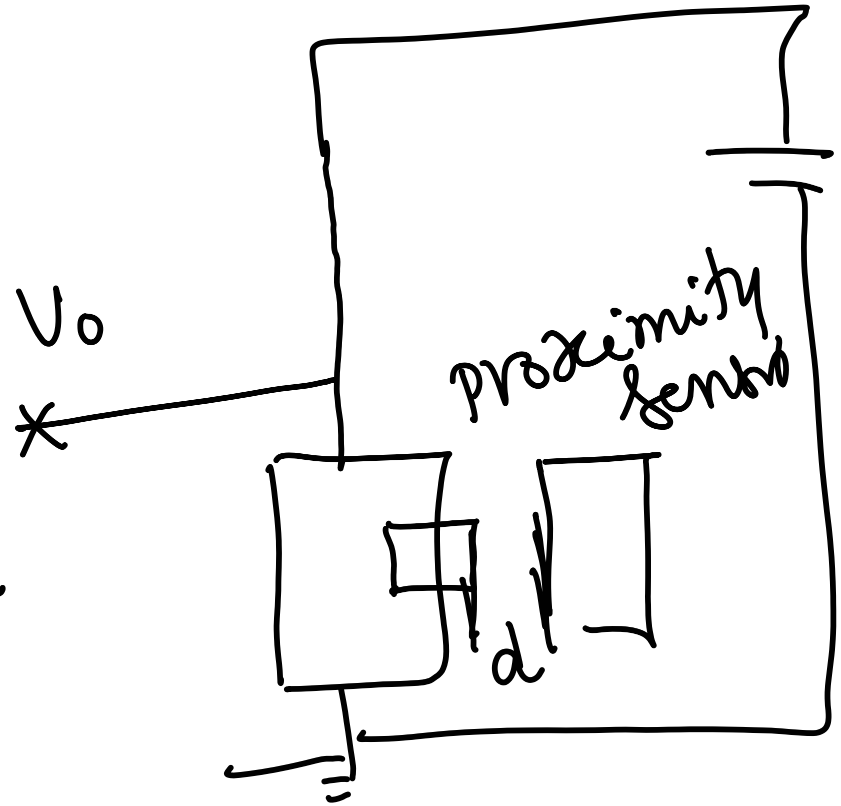
So vol at B will be present.

⑤ When the vol accross the

cap C drops to zero, T_3

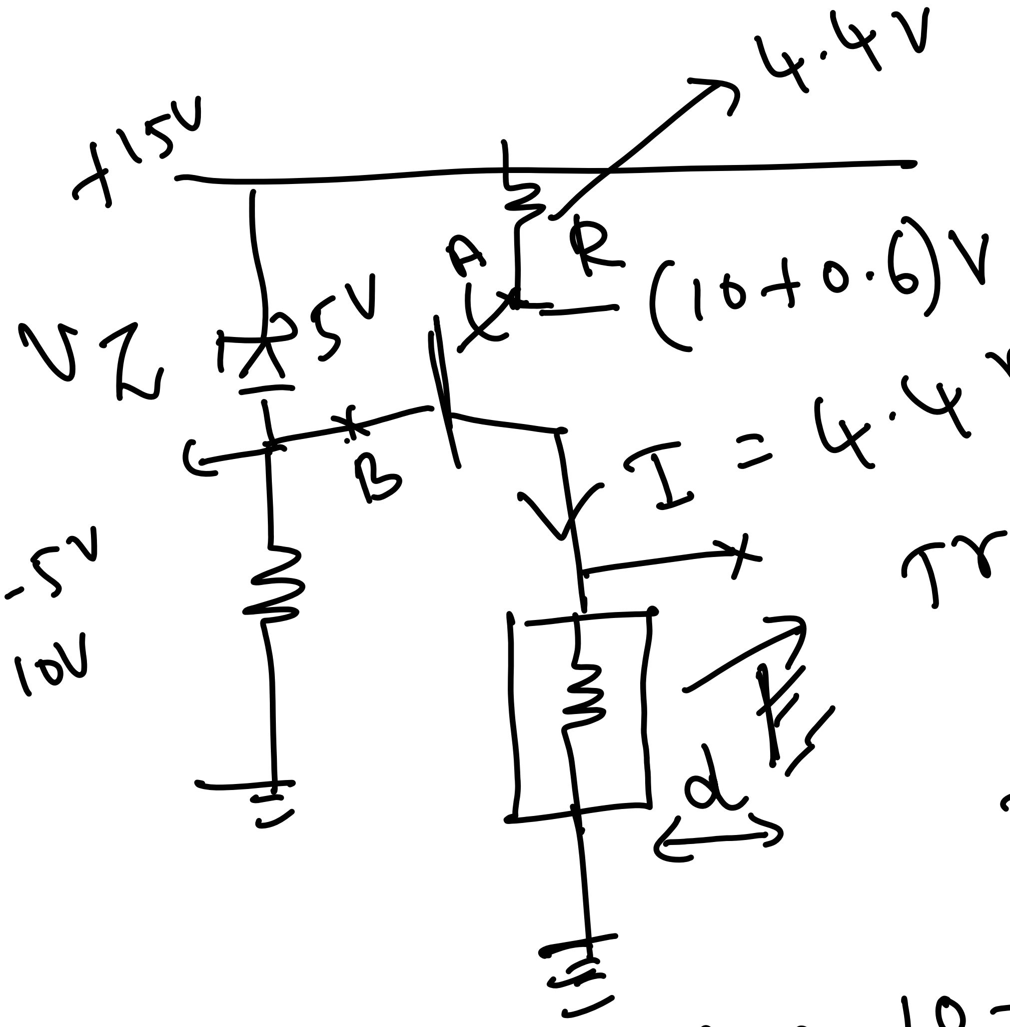
goes off. And vol at B = 0

How to use the transistor for temperature compensation in transducers



w/ V_o represents displacement d

However V_o changes with ambient temp



$(15 - V_Z)$
$= 15 - 5V$
$= 10V$

$I = 4.4\ mA$

Transducer

$I_{25} = 4.4\ mA$

$I_{80} = 4.52\ mA$

Vol at point A = 10+0.6V
= 10.6V

Because vol at B = 10V
(15-5V) = 10V

voltage across R

$= 15 - 10.6 = 4.4V$

If $R = 1K$

So current through

$R = \frac{4.4}{1K} = 4.4\ mA$

So the current $I = 4.4\ mA$

at room temperature

I will increase with temp
(ambient)

At $25^\circ C$ ambient temperature

$I = 4.4\ mA$

What is I at $80^\circ C$ ambient ?

wf at B = 10 V

V_{BE} at $100^\circ C = 0.6 - 2.2mV \times (80 - 25)$

$= 0.6V - 2.2 \times 10^{-3} (80 - 25^\circ C)$

$= 0.6 - 2.2\ mV (55^\circ C)$

$= 0.6 - 2.2 \times 55 \times 10^{-3}$

$= 0.6 - 121\ mV$

$= 0.6 - .12$

$=$

$$\begin{array}{r} 55 \times 22 \\ \hline 110 \\ 110 \\ \hline 1210 \end{array}$$

V_{BE} at $100^{\circ}C = 0.48V$

Voltage at A $= 10 + 0.48V$

$= 10.48V$

Voltage across R $= 15 - 10.48$

$= 4.52\ V$

$$I = \frac{4.52}{1k} = 4.52\,mA$$

The current I increases with temp

So if the temp co-eff of the transducer is negative (with increase in ambient temp wf across the transducer decreases) then this arrangement will compensate.

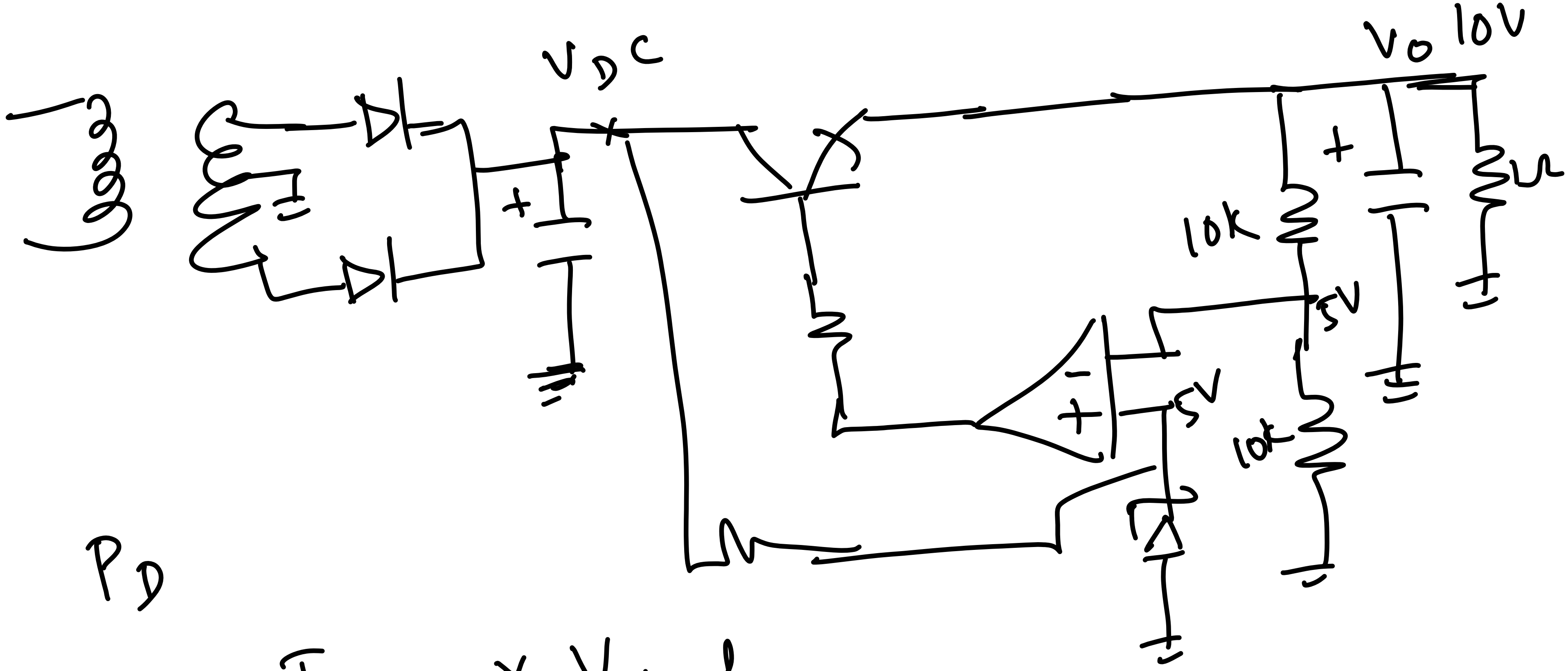
① V_{BE} of the transistor changes with temp.

② $\Delta V_{BE} = -2.2\,mV/°C$

(3) ΔV_{DE} is highly reproducable

So it is possible to get repeatable ~~&~~ result in the mass production

Transistors in parallel



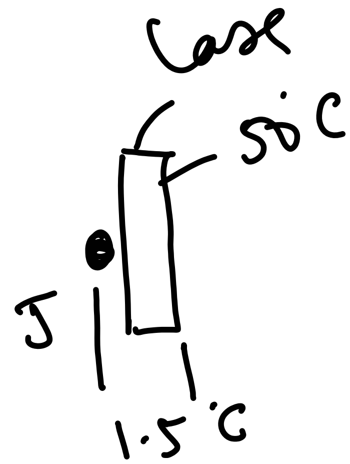
P_D

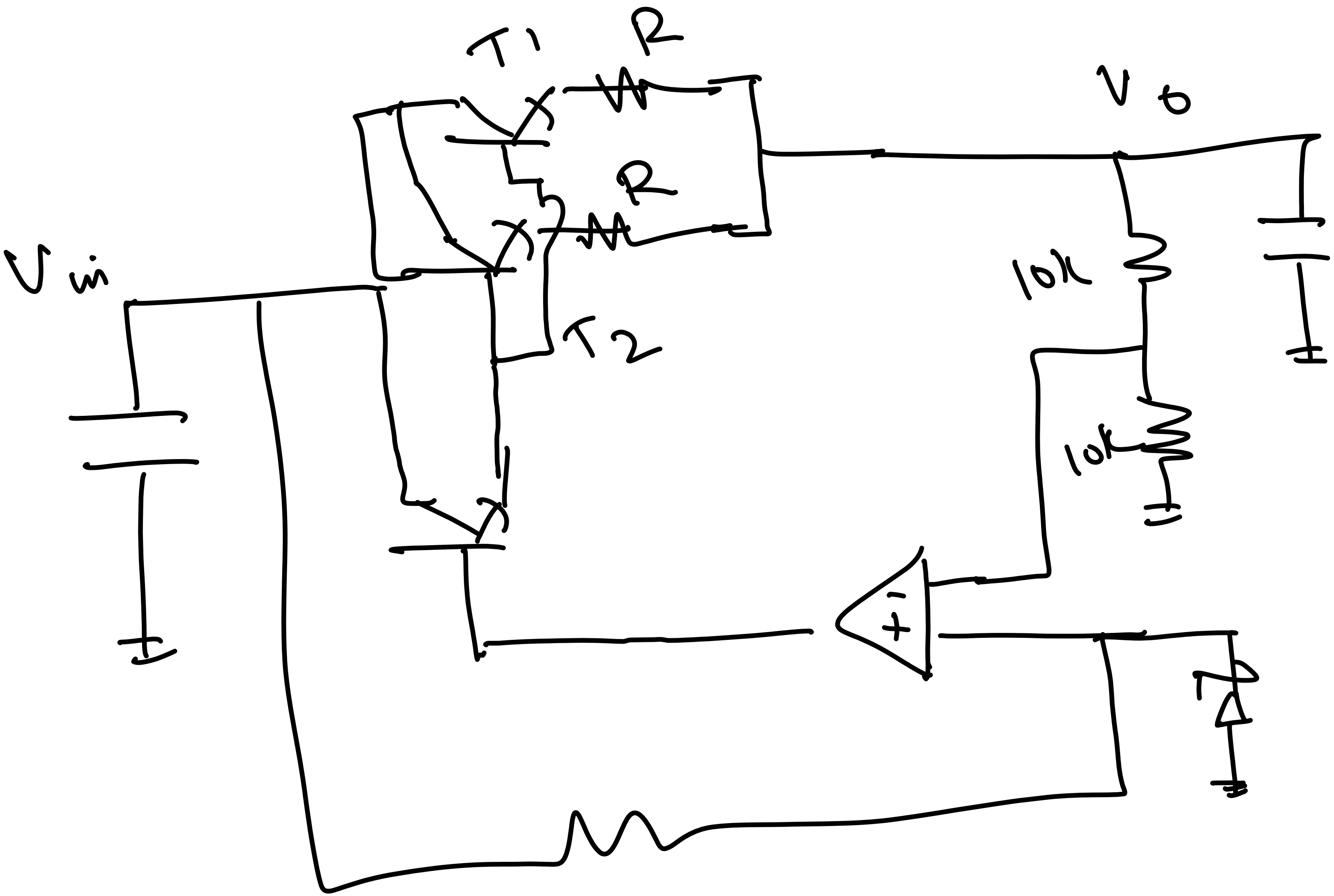
$= I_{out} \times V_{drop}$

$= 10 \times 5 = 50W$

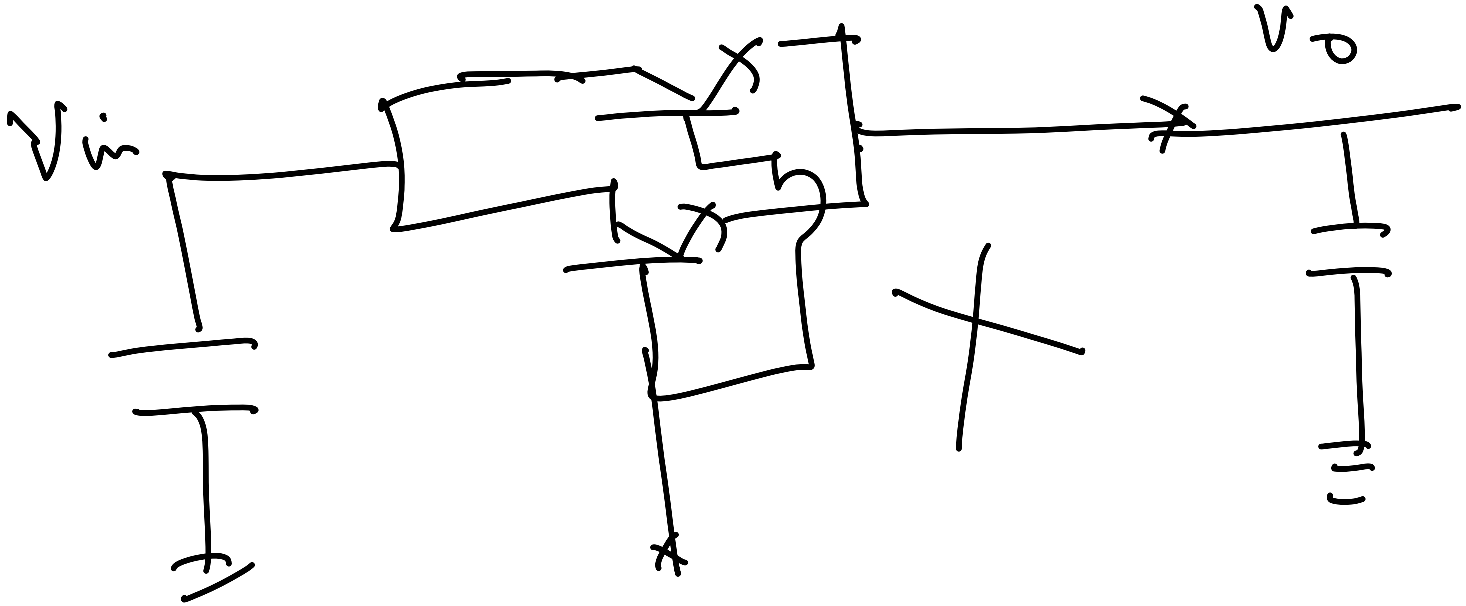
For 3055 Transistor

$\theta_{JC} = 1.5^{\circ}C/W$





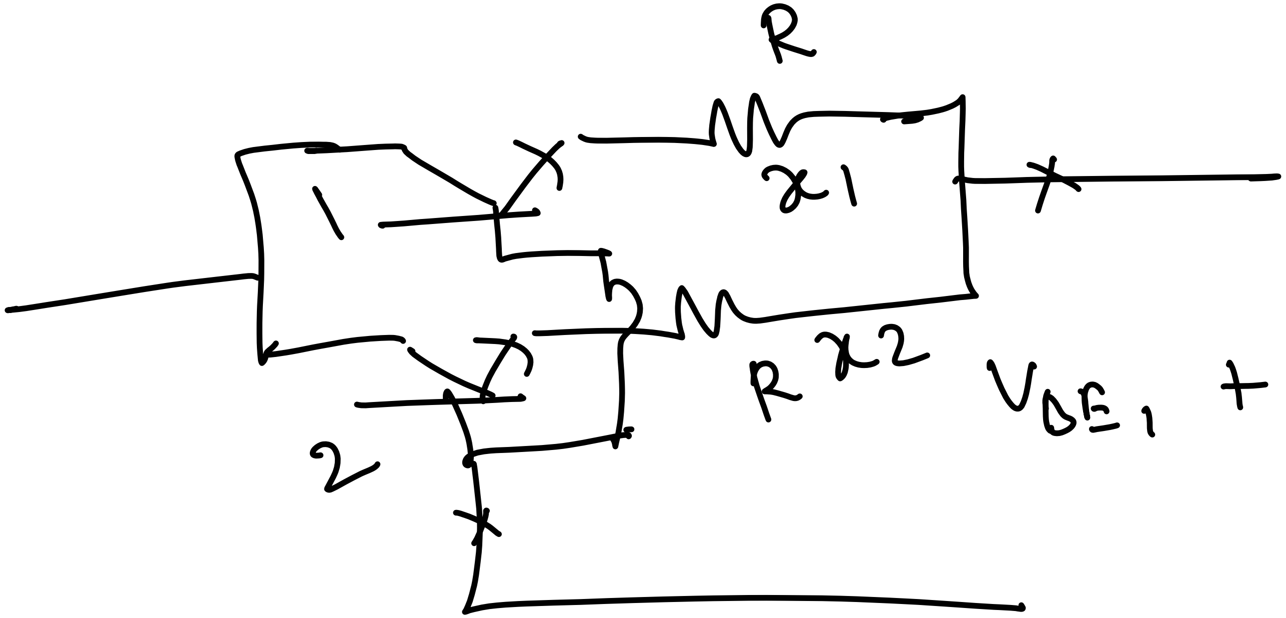
T1
R
R
Vo
Vin
T2
10k
10k
-
+


Vin
Vo

For the Same I_c

V_{BE} requirement is different for different transistors.

This difference is $\simeq 50mV$ max



$$V_{BE_1} + x_1 = V_{BE_2} + x_2$$

If V_{BE1} and V_{BE2} are
differ by 50mV

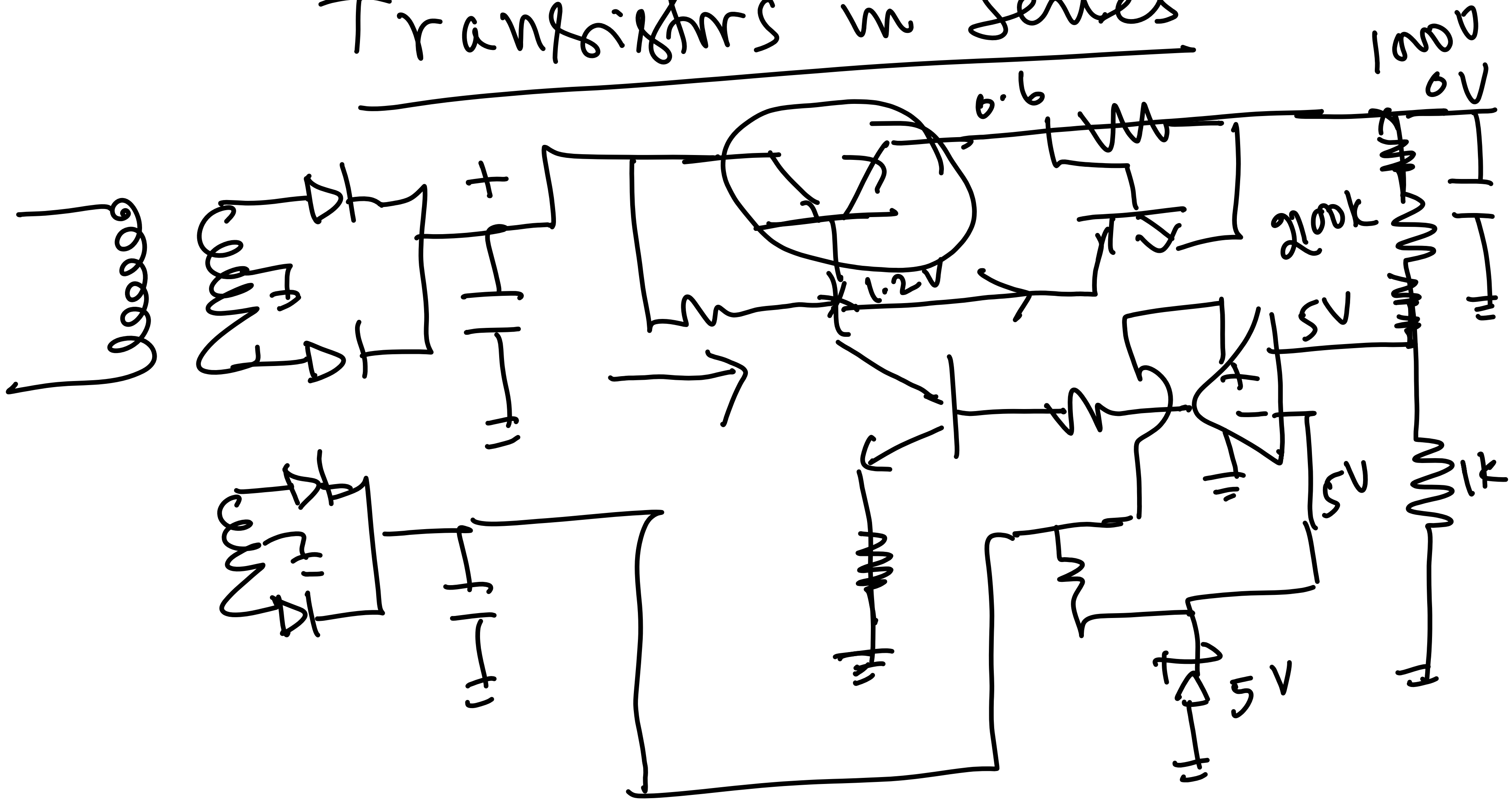
then $x_1 \sim x_2 = 50mV$

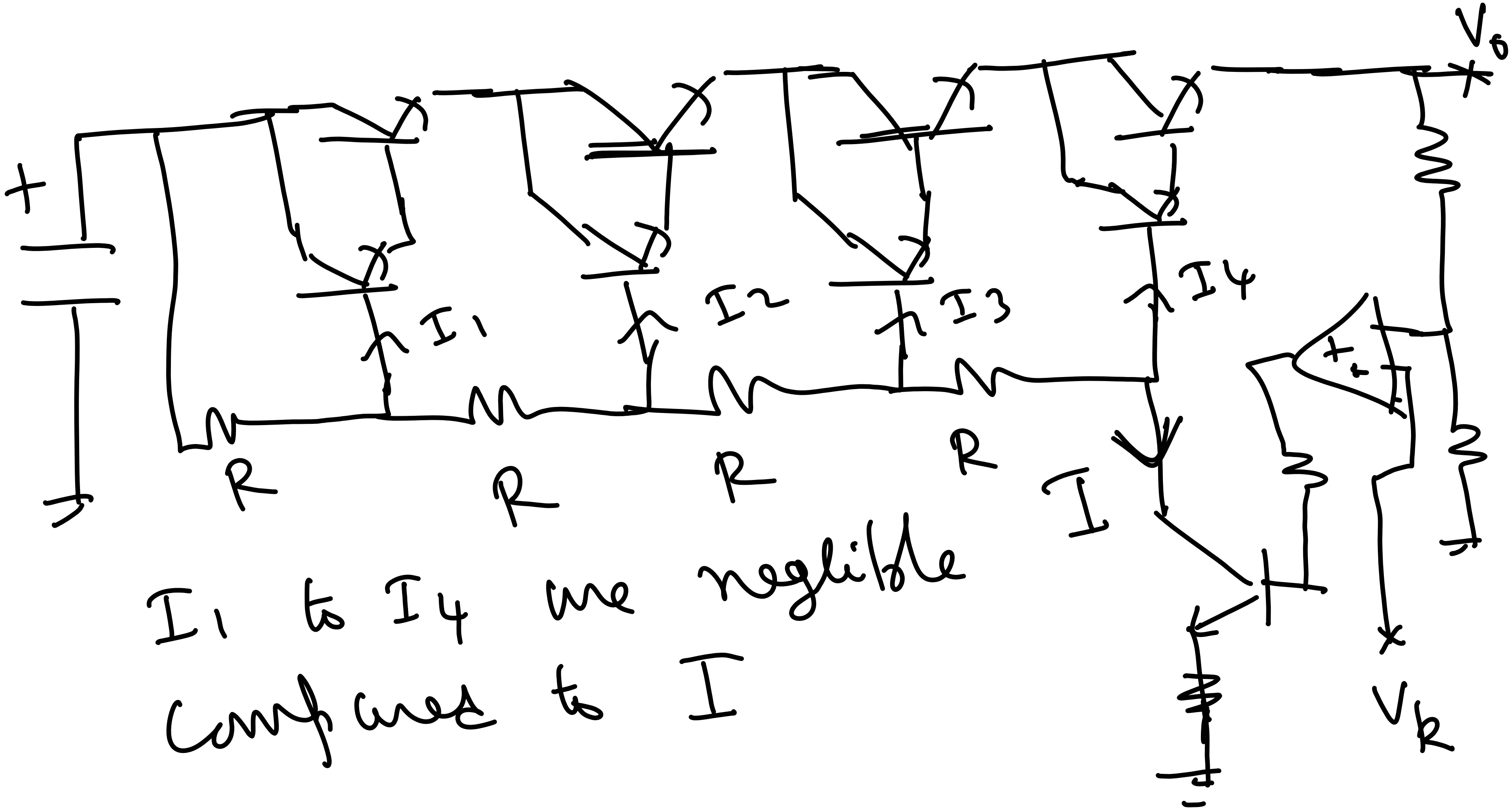
If $R = 1\Omega$

$RI_1 \sim RI_2 = 50mV$

$I_1 \sim I_2 = 50mA$

Transistors in Series



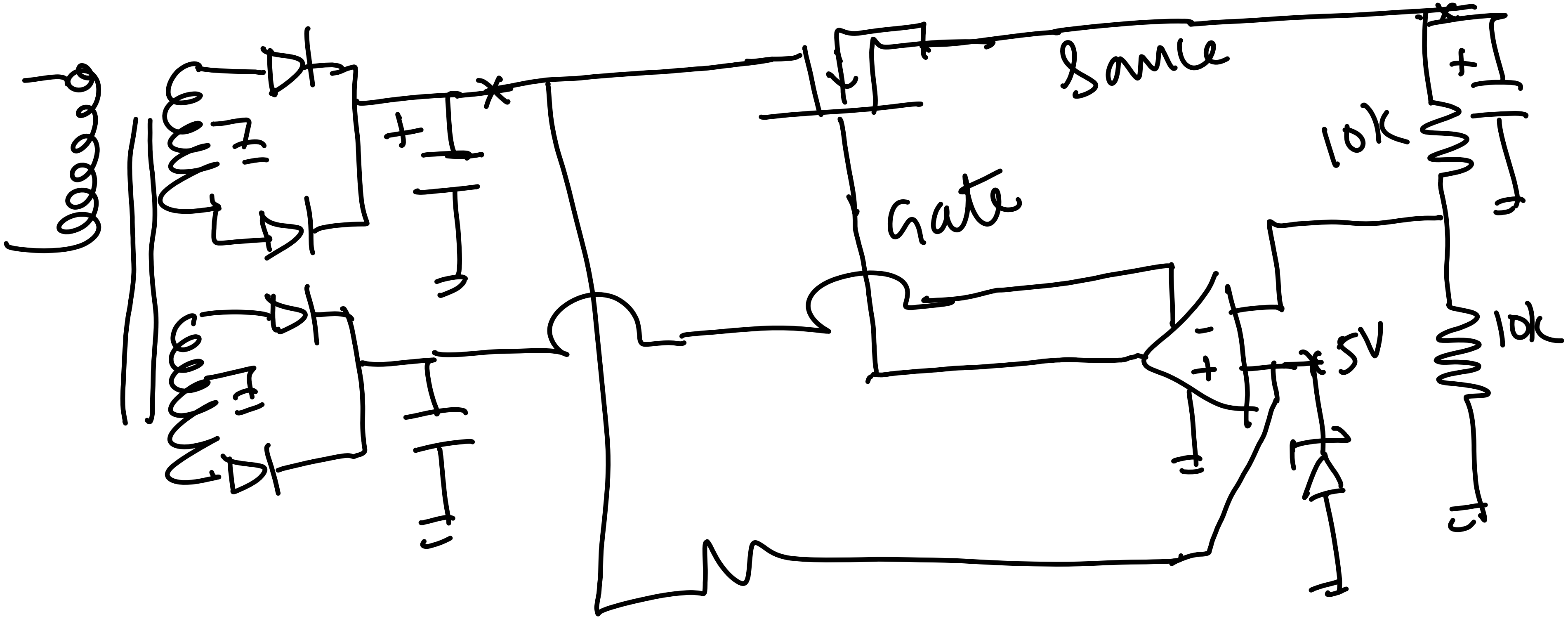


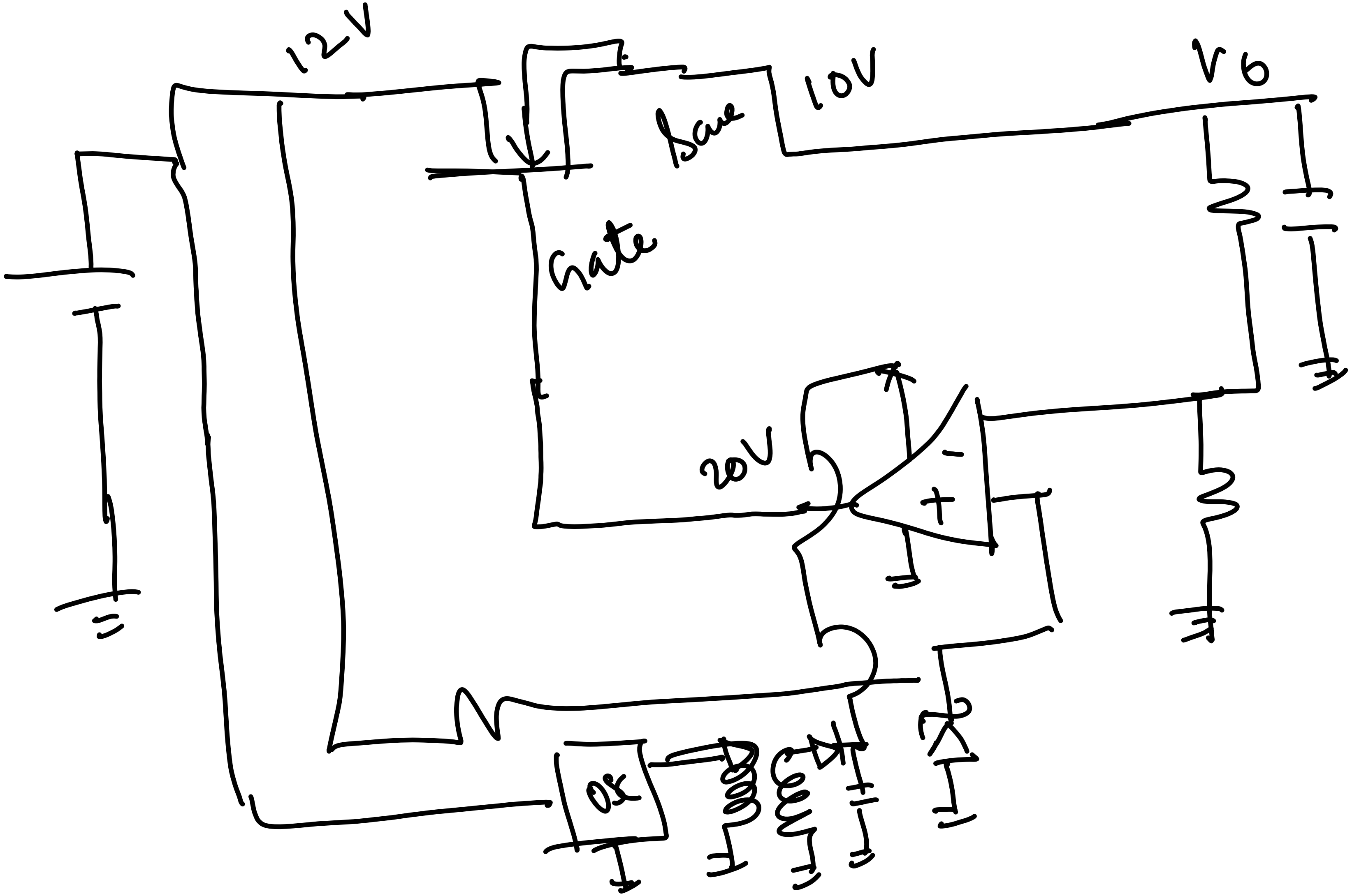
I_1 to I_4 are neglible compared to I

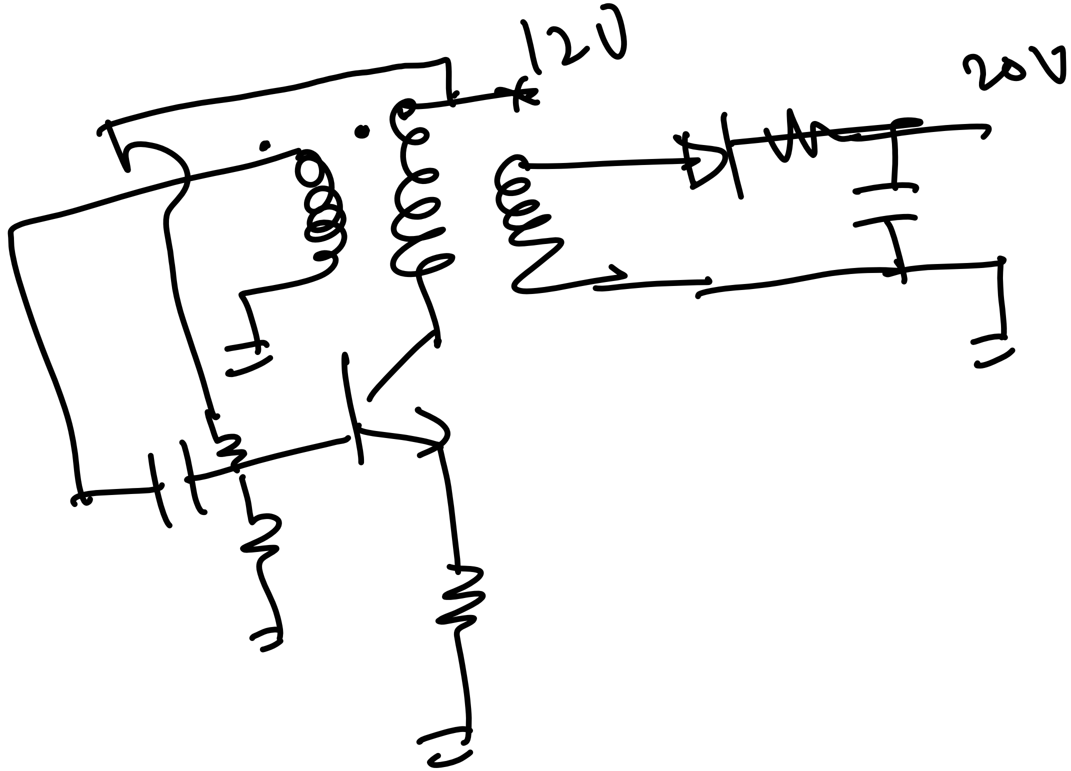
1. Connect the transistons in series to boost the vol handing capacity of the circuit.
2. Connect the transitors in parallel to boost the current handling capacity of the transitars

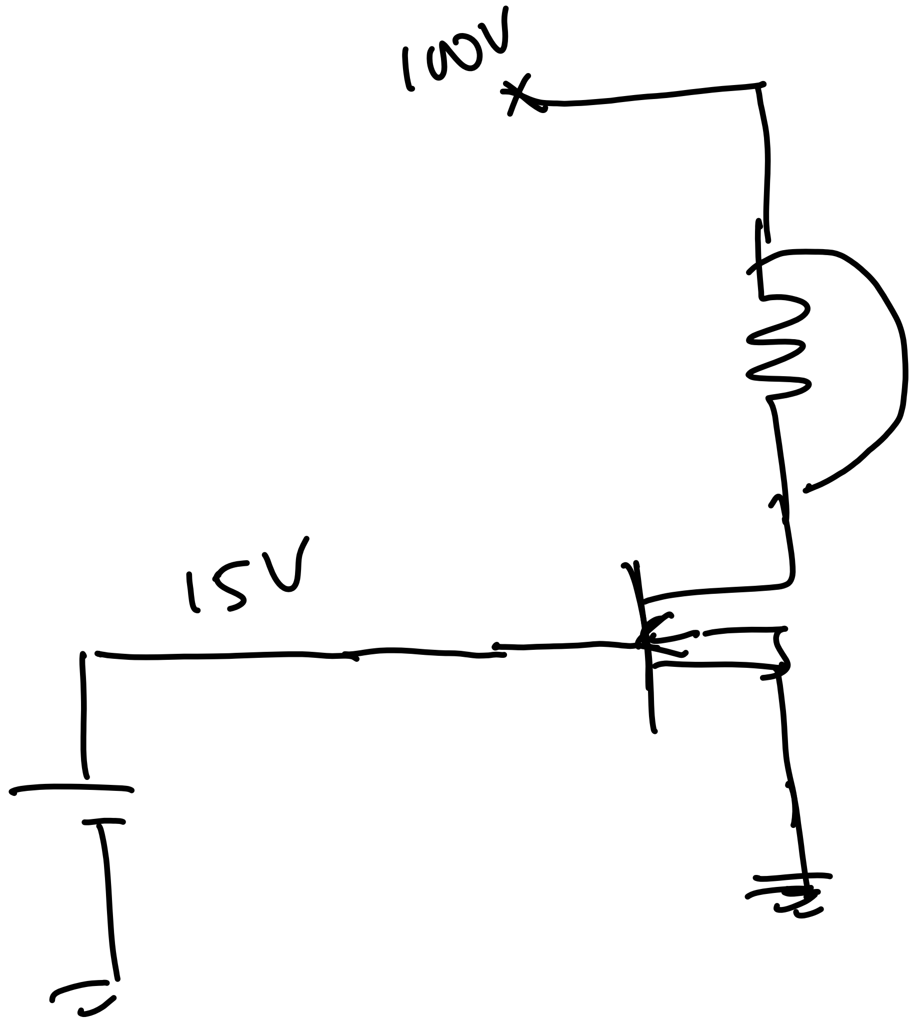
(3) Connect the transistors in parallel to reduce the heat dissipation in one one single transistor.

Low Drop out voltage regulators

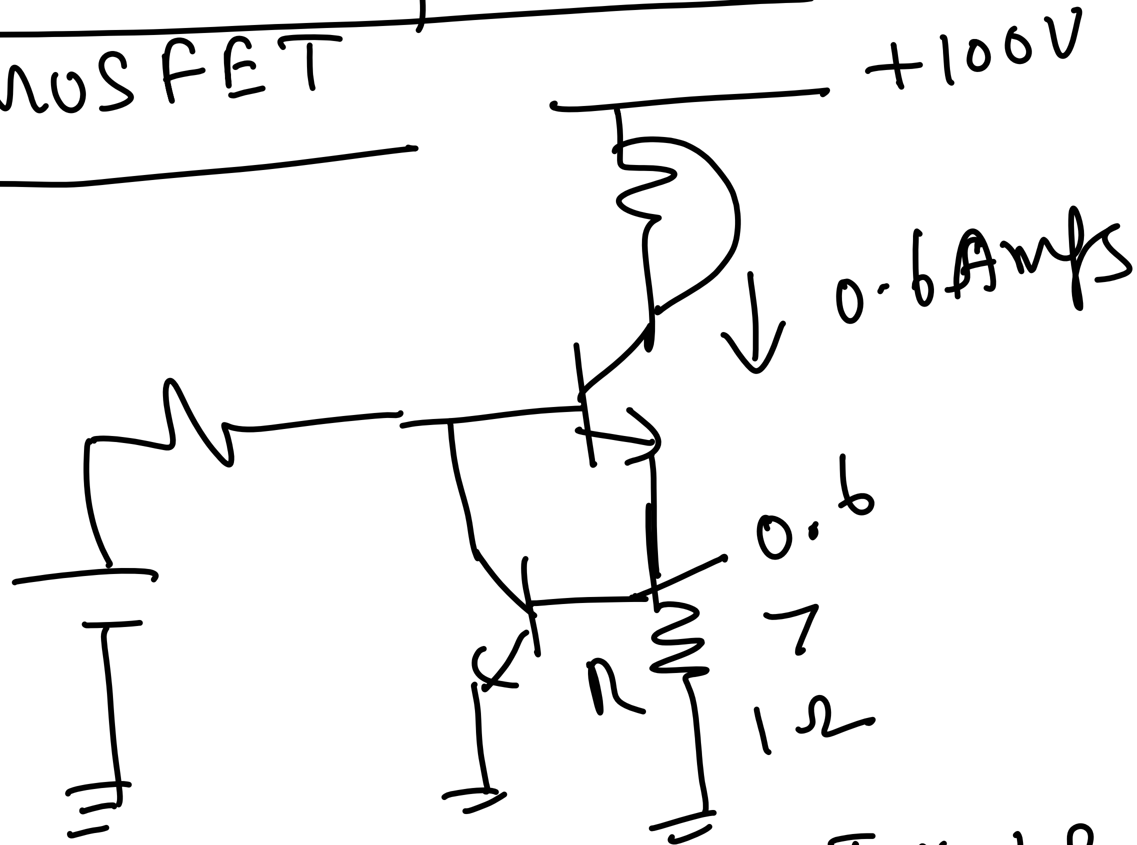



12V
Gate
Sour
10V
Vo
20V
-
+
Osc


12V
20V


100V
15V

Short circuit protection to MOSFET

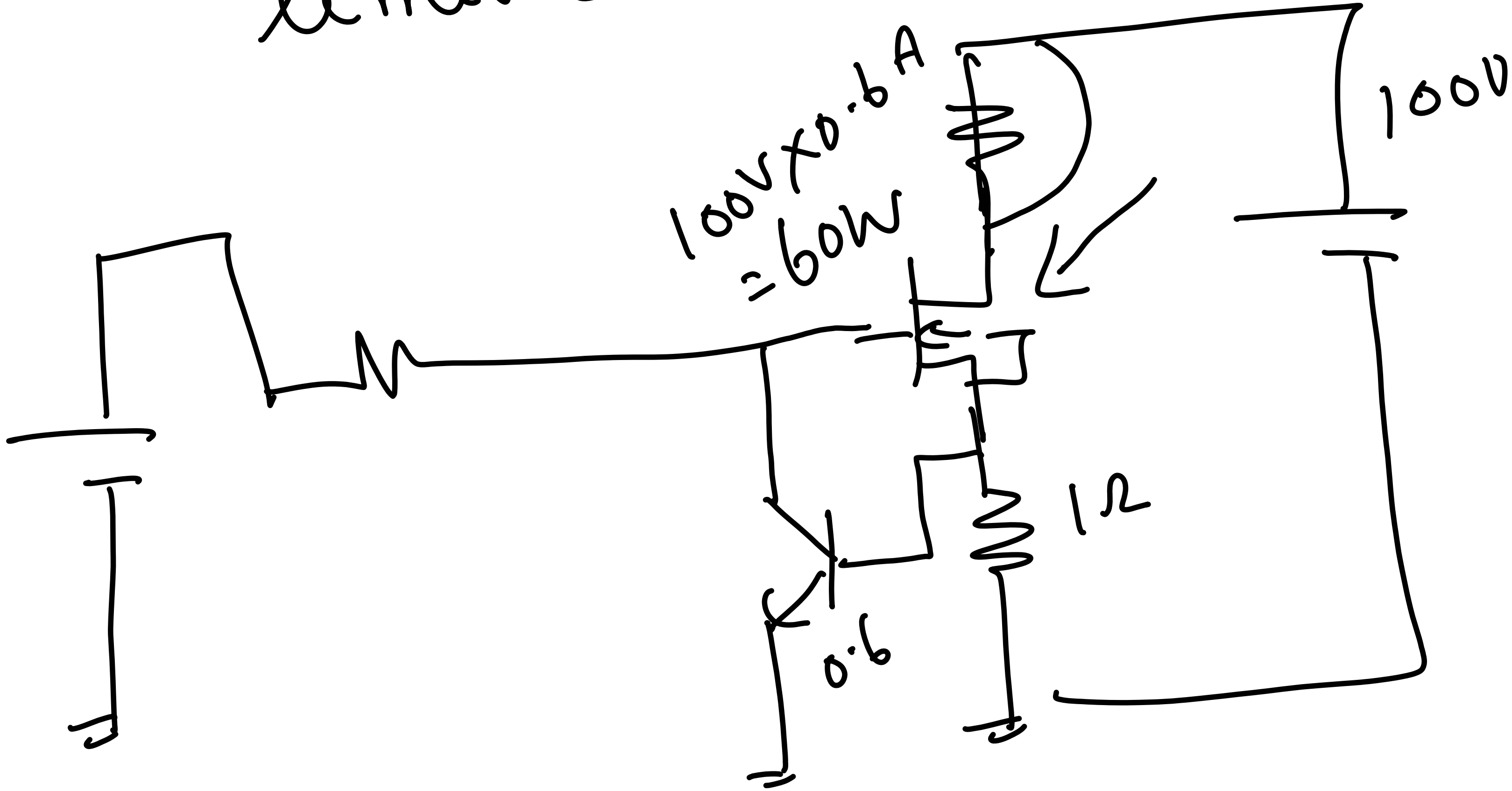


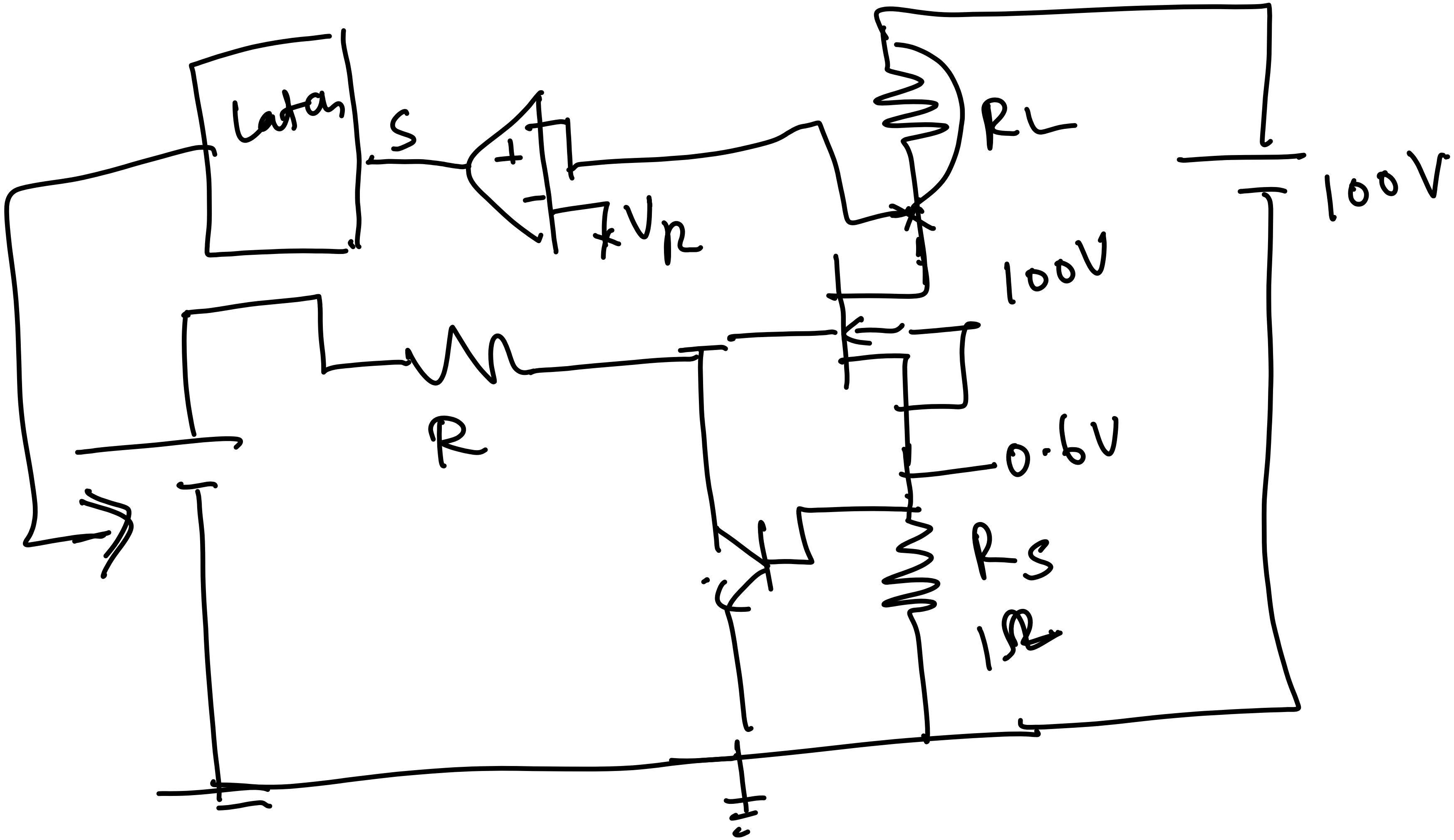
$$V_{BE} = 0.6\,V = IR = I \times 1\Omega$$

$$0.6 = 1 \times I$$

$I = 0.6A$

Short circuit current is limited to 0.6 Amps





Latch
S
+
-
Vr
RL
100V
100V
R
0.6V
Rs
1Ω

Resistance measurement

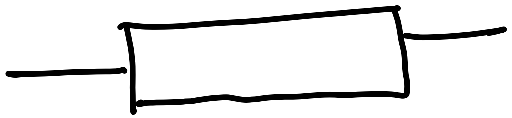
1. Temperature sensor
   - PRT
   - Thermistors
2. position measurement
3. weight measurement
   strain ~~ga~~ gauges

① Very low resistance measurement

few $m\Omega$ to $\mu\Omega$

↓

Sample resistance measurement



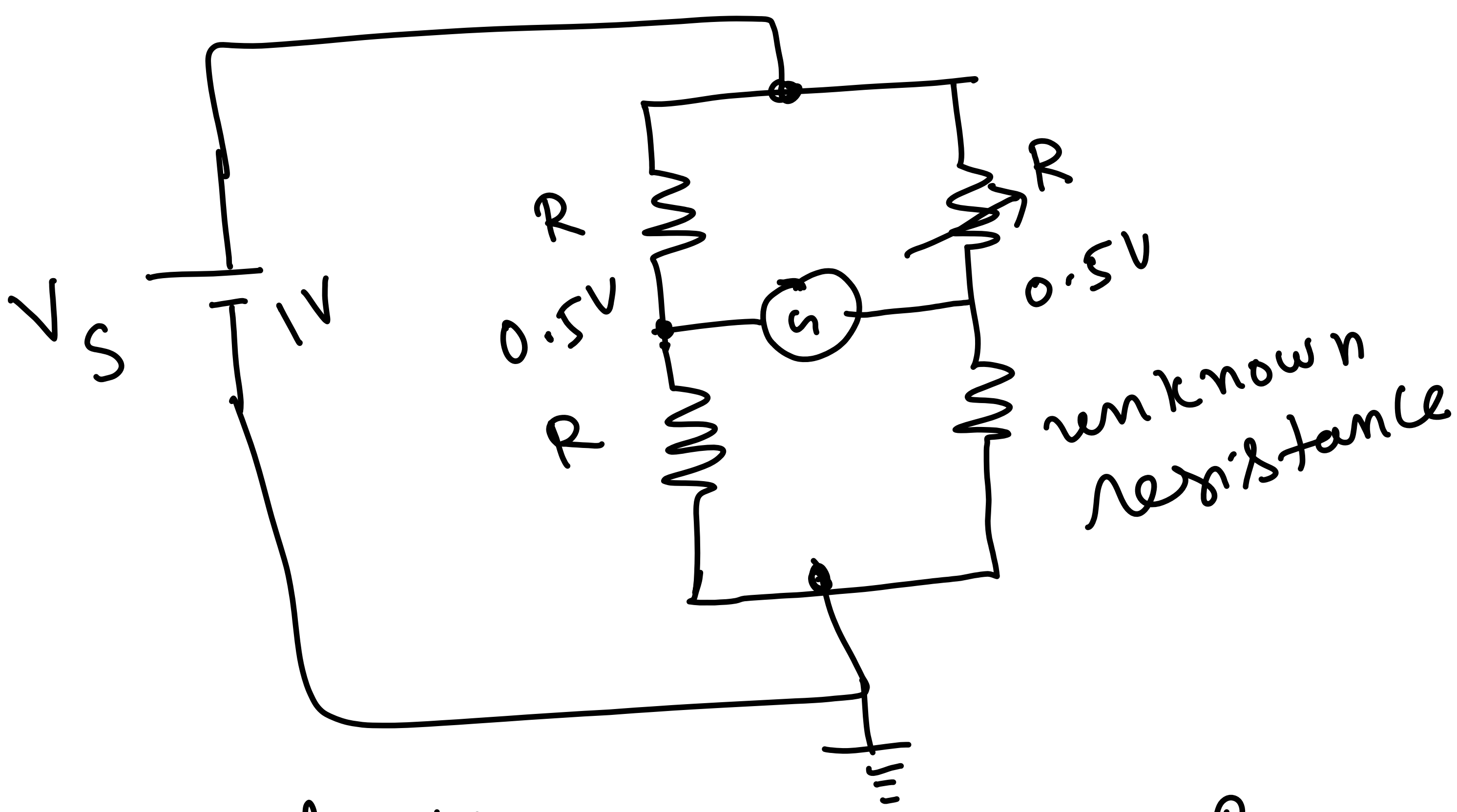
② Medium range resistance measurement

1Ω - 100k

(3) Very high resistance measurement

$100\,k\Omega - 100\,M\Omega$

Bridge based resistance measurement



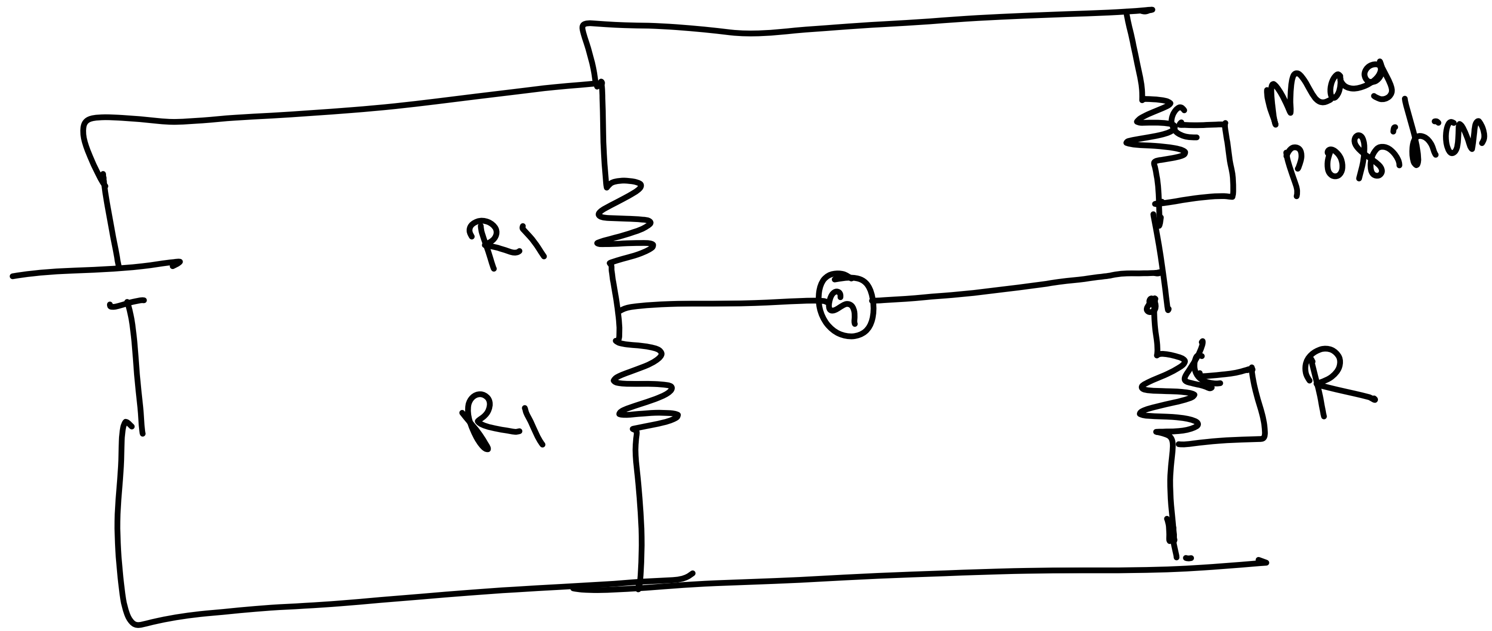
At balance
unknown resistance = R
Temp co-eff of the fixed

Resistors = 10 ppm/°C

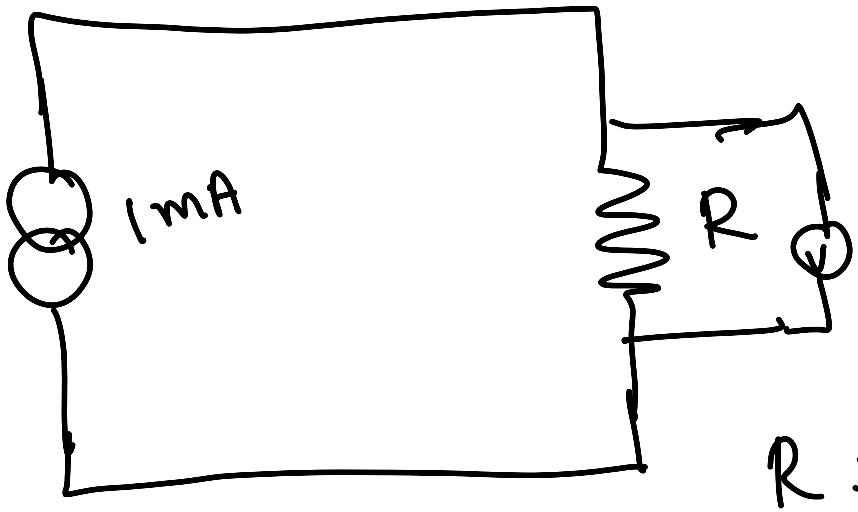
Bridge method gives high accuracy.

1. All resistors are having same temp co-eff
2. Bridge excitation voltage will not affect the balancing point

For example for position measurement if potentiometer is used then



Constant current source based resistance measurement



$$V = RI$$

$$= R \times 1\,mA$$

$$R = 100\,\Omega$$

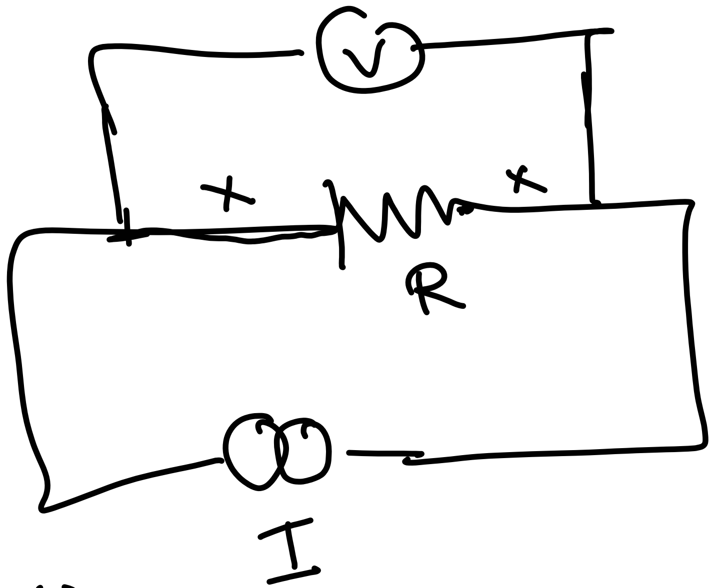
$$V = 100\,m\Omega$$

(1) The error in current or drift in constant current source due to ambient temperature variation makes the measurement less accurate

(2) Error in voltage measurement also responsible for in accuracy

Low resistance measurement

Range : few mΩ to few μΩ



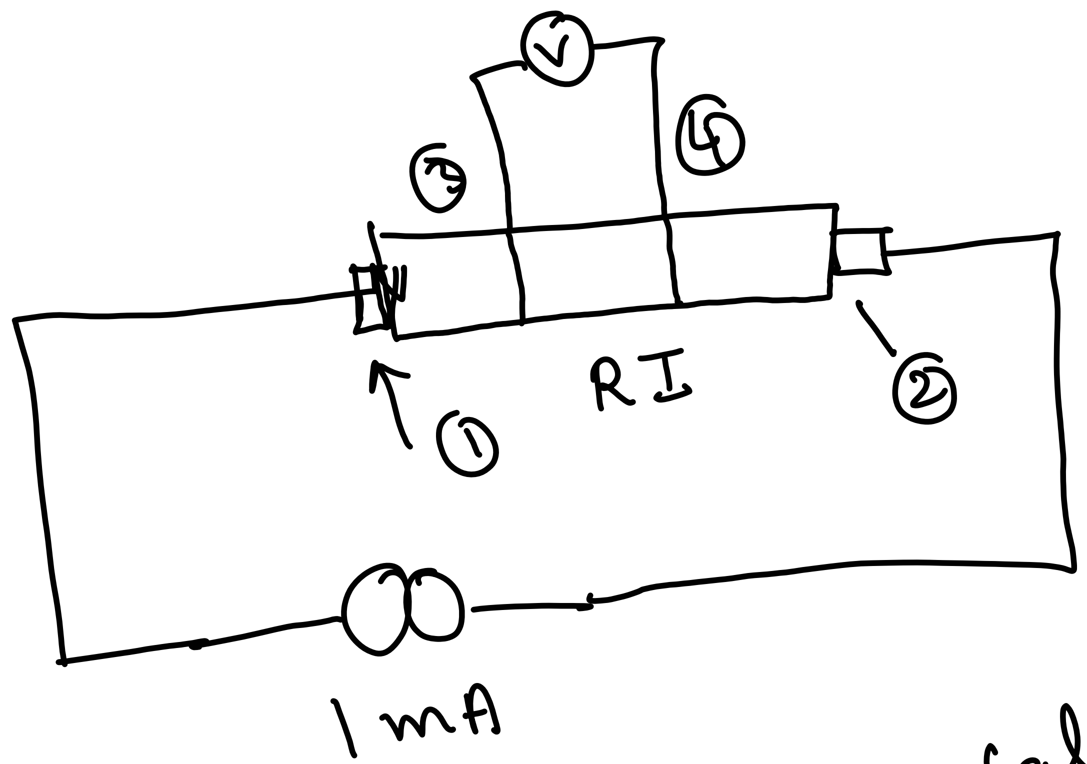
In very low resistance measurement

1. Contact resistance is added to the real value

② For very low R the contact resistance can give large error

4 - Probe resistance measurement

① This avoids contact resistance problem. So it is highly suited for very low resistance measurement



Probes 1 and 2 are called current points

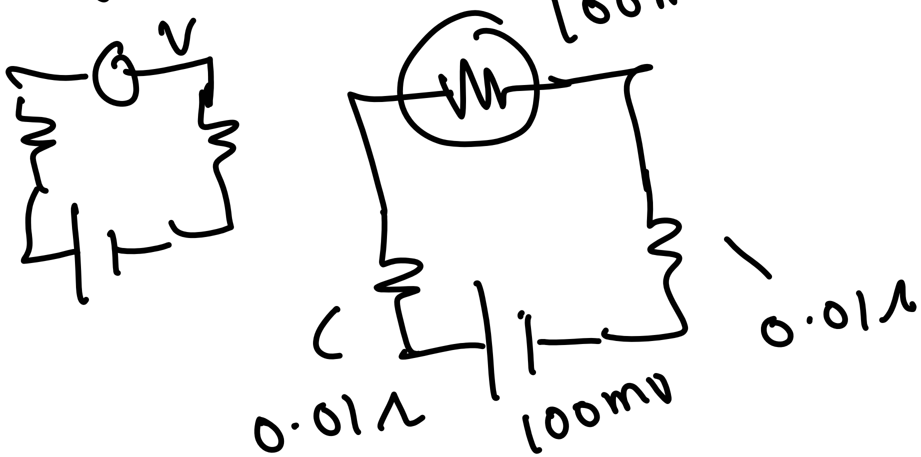
Probes 3 and 4 are called voltage points

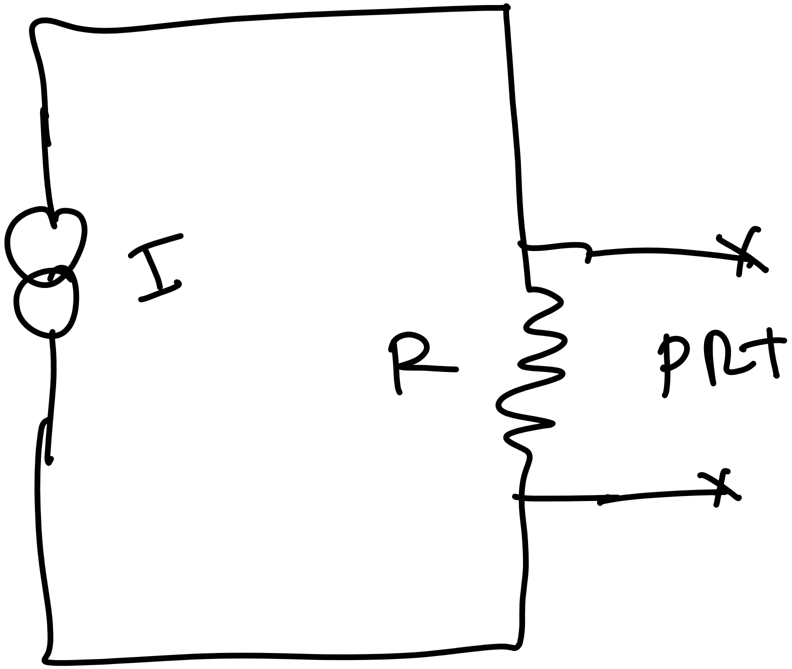
Voltmeter reading = V = RI

I is known

R can be computed

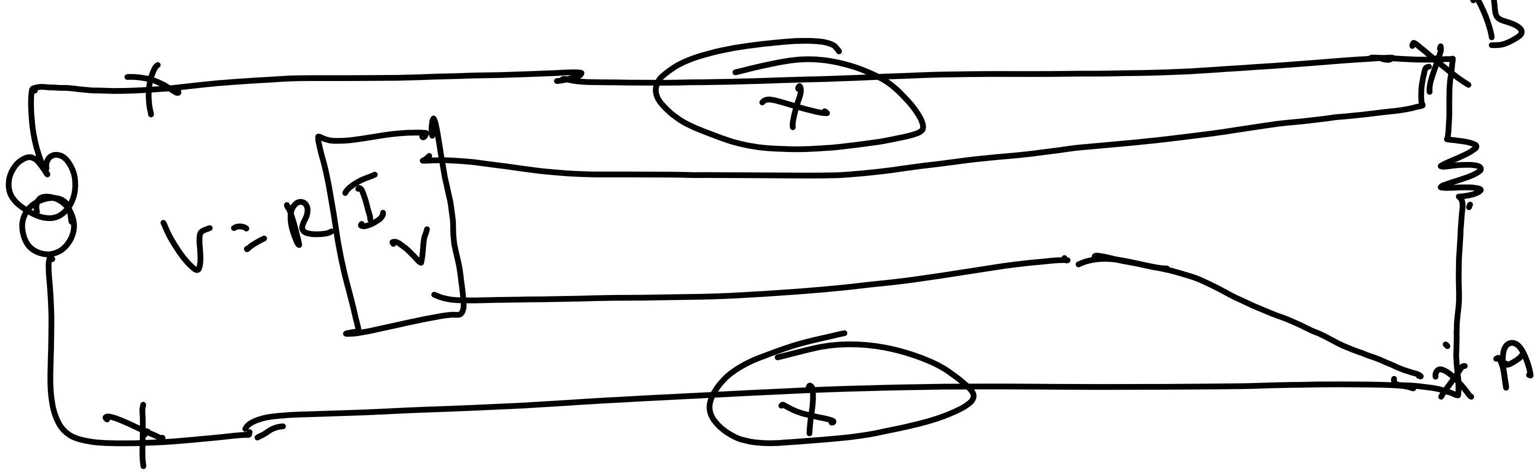
The contact resistance at the voltage points ③ and ④ will not affect the accuracy





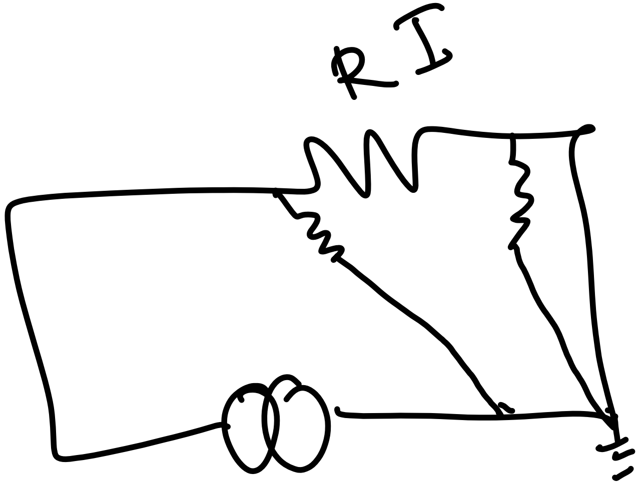
$RI = V_0$

I

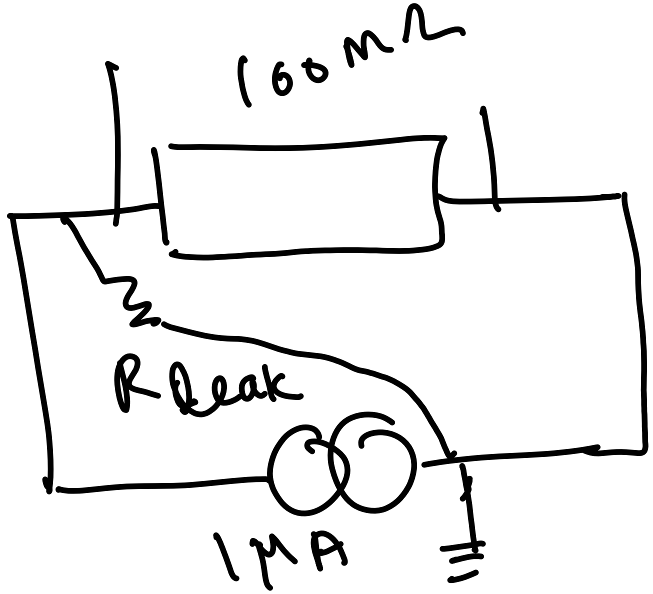


High resistance measurement

Constant current source method is not suitable



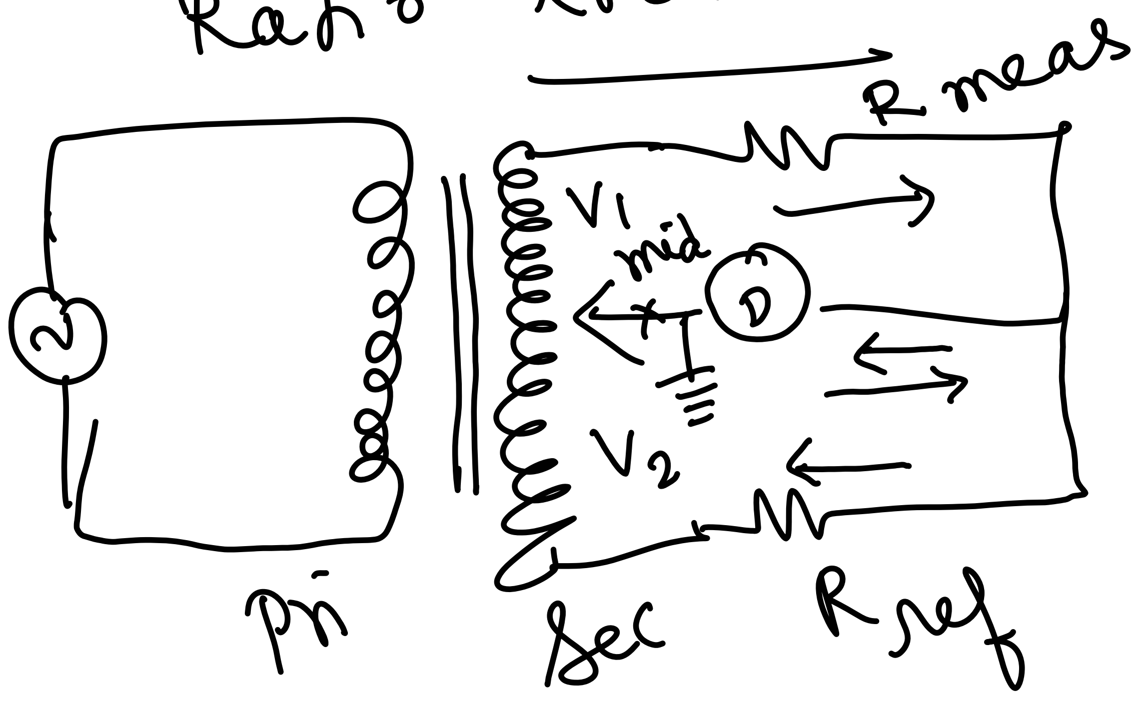
$10^6 \times 10^{-6} = 1V$



In high resistance the leakage resistance R_{leak} diverts the current that is coming from the constant current source

This makes system less accurate

For high resistance measurement use Ratio transformer bridge

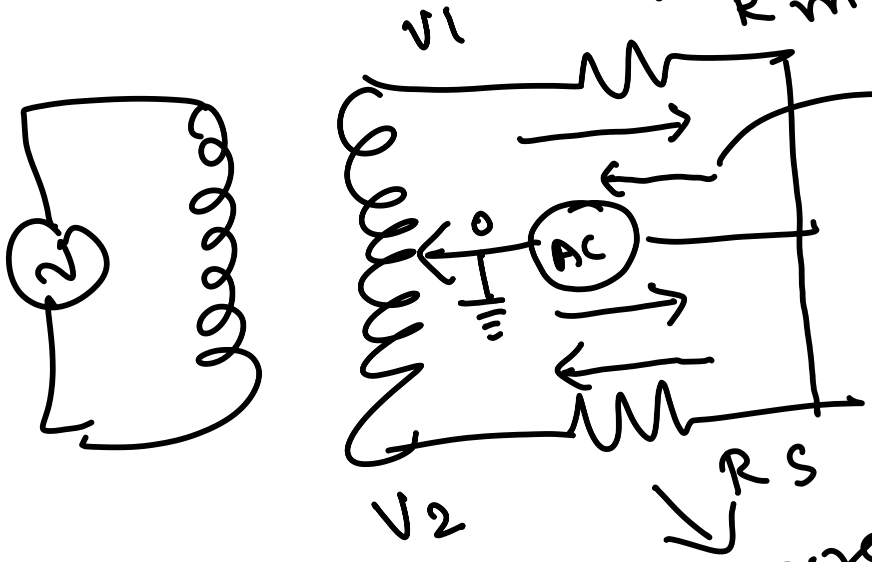


If the secondary centre tap is at mid point

Then $V_1 = V_2$

Ratio transformer bridge for high resistance measurement

High resistance R_m



$\frac{V_1}{R_m}$

Fixed standard resistance

$\frac{V_2}{R_S}$

V_1 and V_2 are 180° out of phase

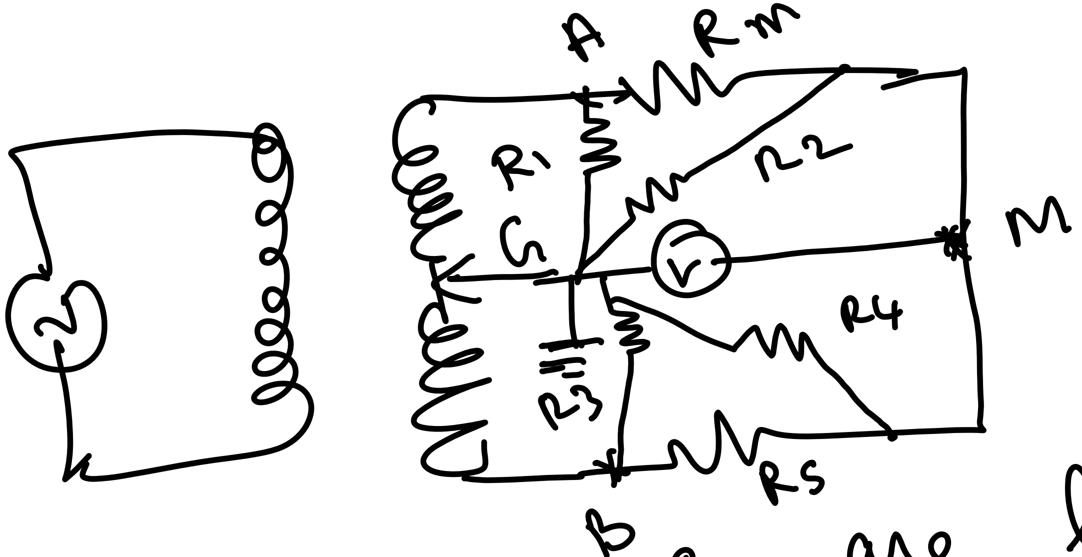
$$\frac{V_1}{R_m} = \frac{V_2}{R_S}$$

$$\frac{V_1}{V_2} = \frac{R_m}{R_S}$$

R_S is known

If V_1 and V_2 ratio is

$\frac{V_1}{V_2}$ is know then unknown resistance can be calculated



R_1, R_2, R_3, R_4 are leakage resistances

These resistances are not

contributing for the balance

why?

WT acc $R_2, R_4 = 0$
at balance

So no current is flowing
through R_2, R_4 at balance
so it is not affecting the
bridge balance

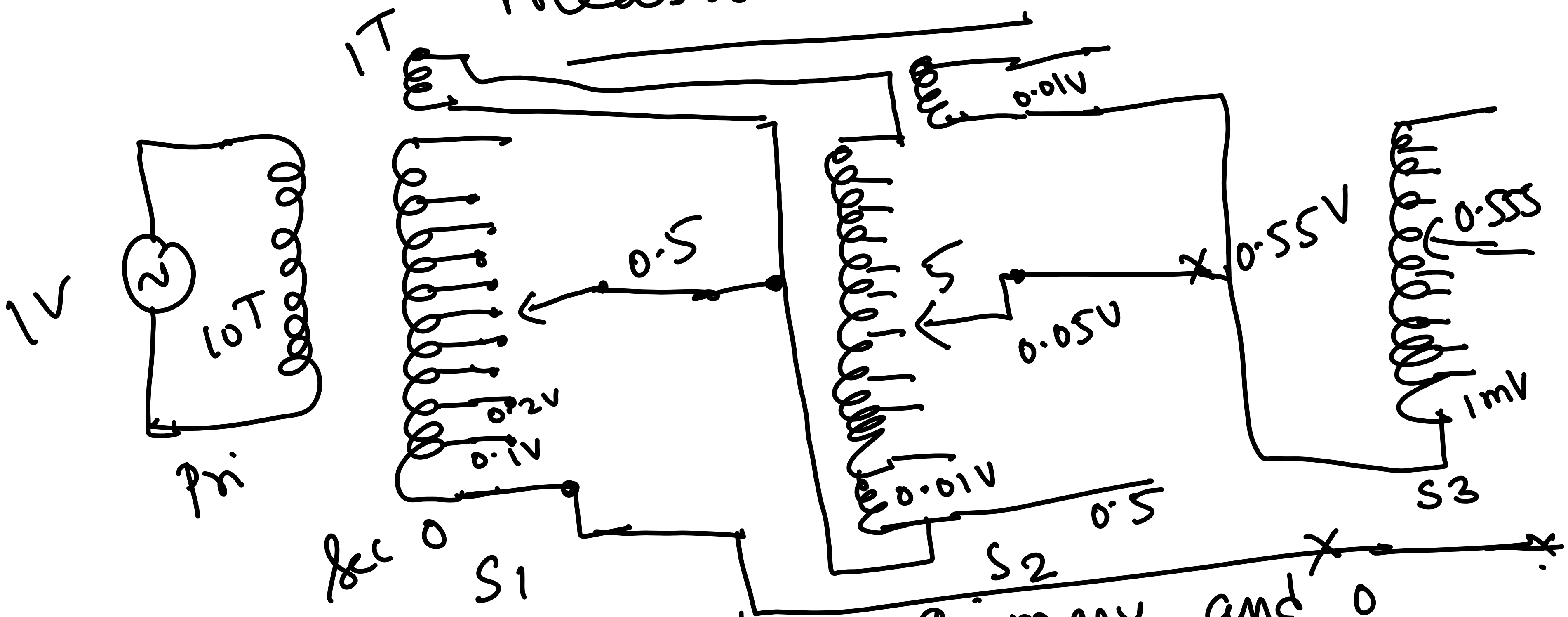
R_1, R_3 are just loading
the secondary of the transformer
but output impedance of the
transformer is very low

So the loading of R_1, R_3 will not change V_1, V_2.
So no effect on the balance of the bridge.

$$\frac{V_1}{V_2} = \frac{R_m}{R_s}$$

If V_1 and V_2 are measured using a meter, then the error involved in this measurement also to be considered.

How to avoid V_1, V_2 measurement?

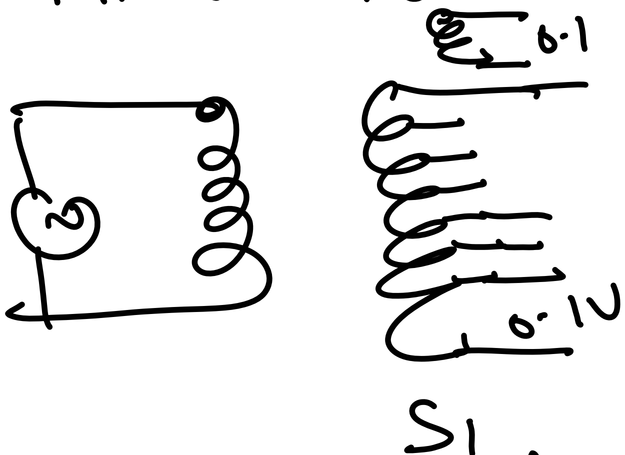


There are 10T in the Primary and 0 secondary.

So For 1V primary the v/ acc each turn of secondary = 0.1v

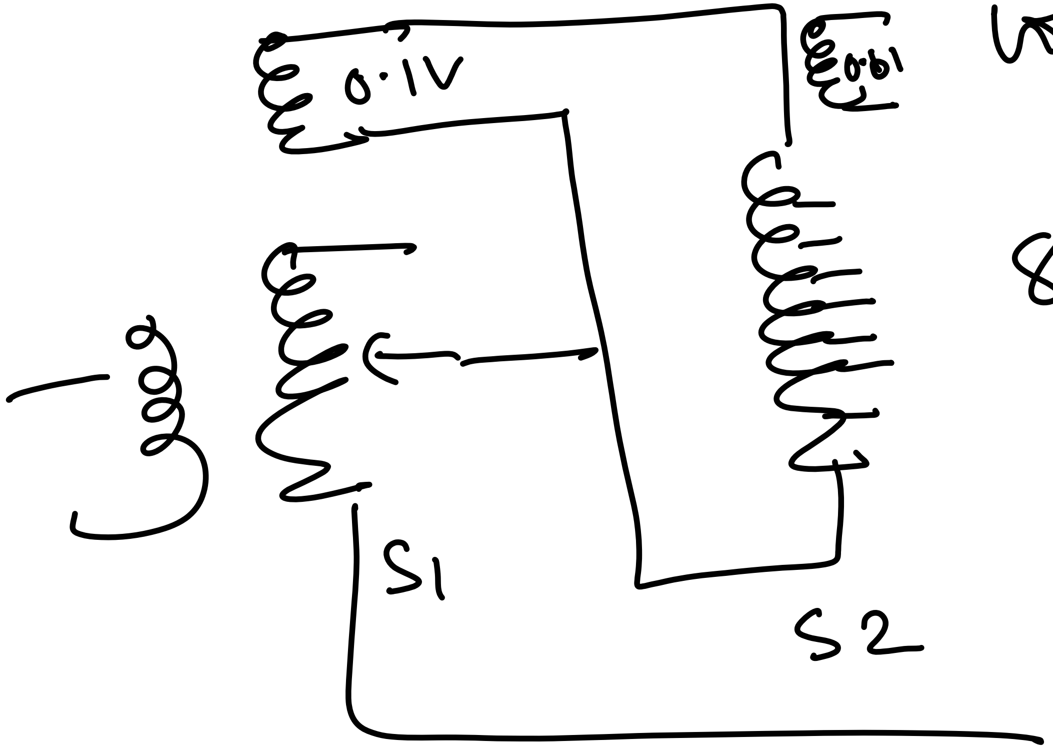
(1) For 3 digit accuracy use 3 secondaries in the ratio transformer

First ~~s~~econdary wdl is 1V



The step size is 0.1V

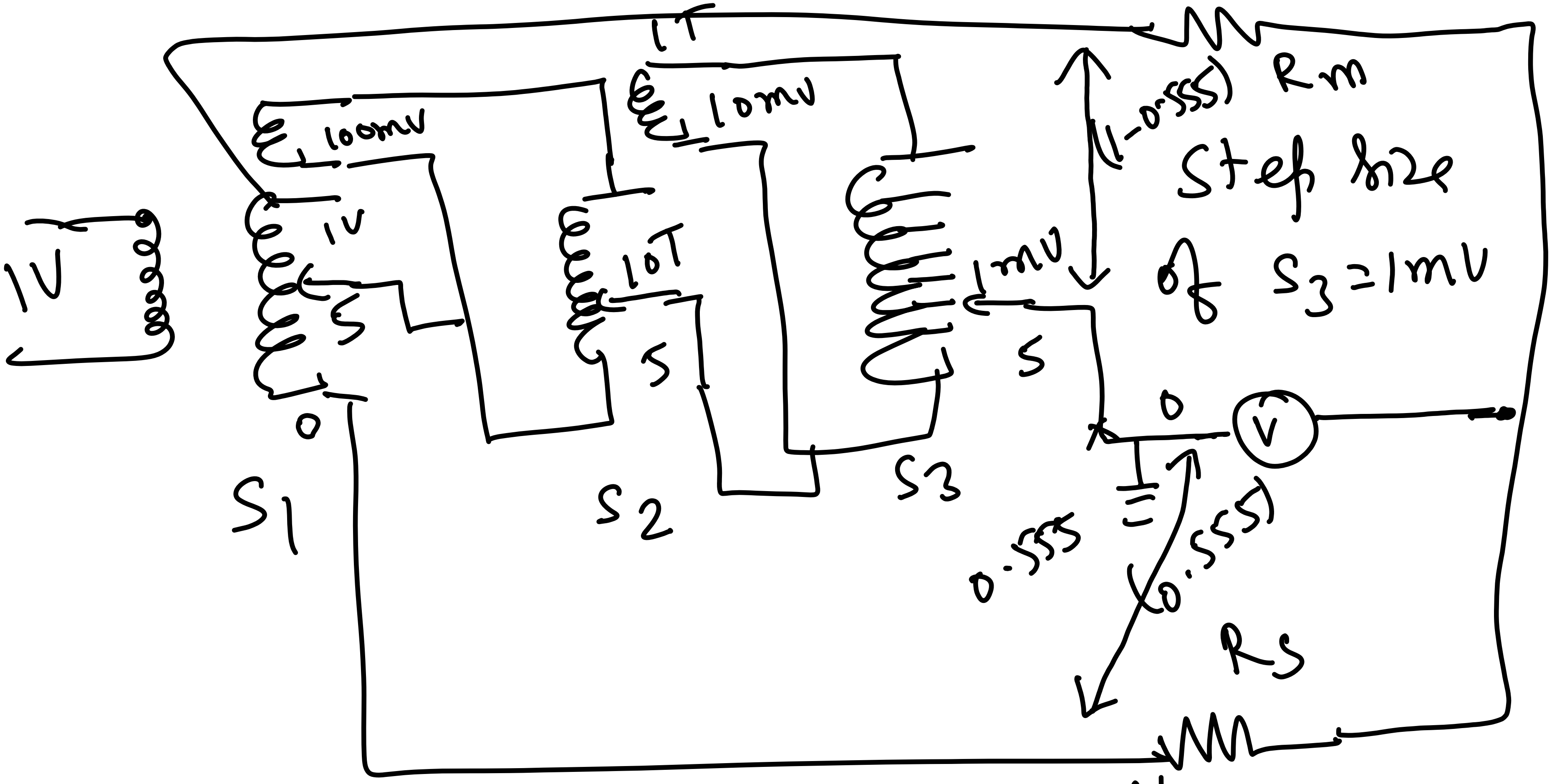
The auxiliary winding is having 0.1V



vol acc S_2
$= 0.1V$

Step size in
$S_2 = 0.01V$

The auxillary winding of S_2
is having vol of 10 mV
This winding of 10 mV is connected
to the 3rd winding S_3



1. The excitation voltage is not applied to the primary changes then also the measurement is accurate

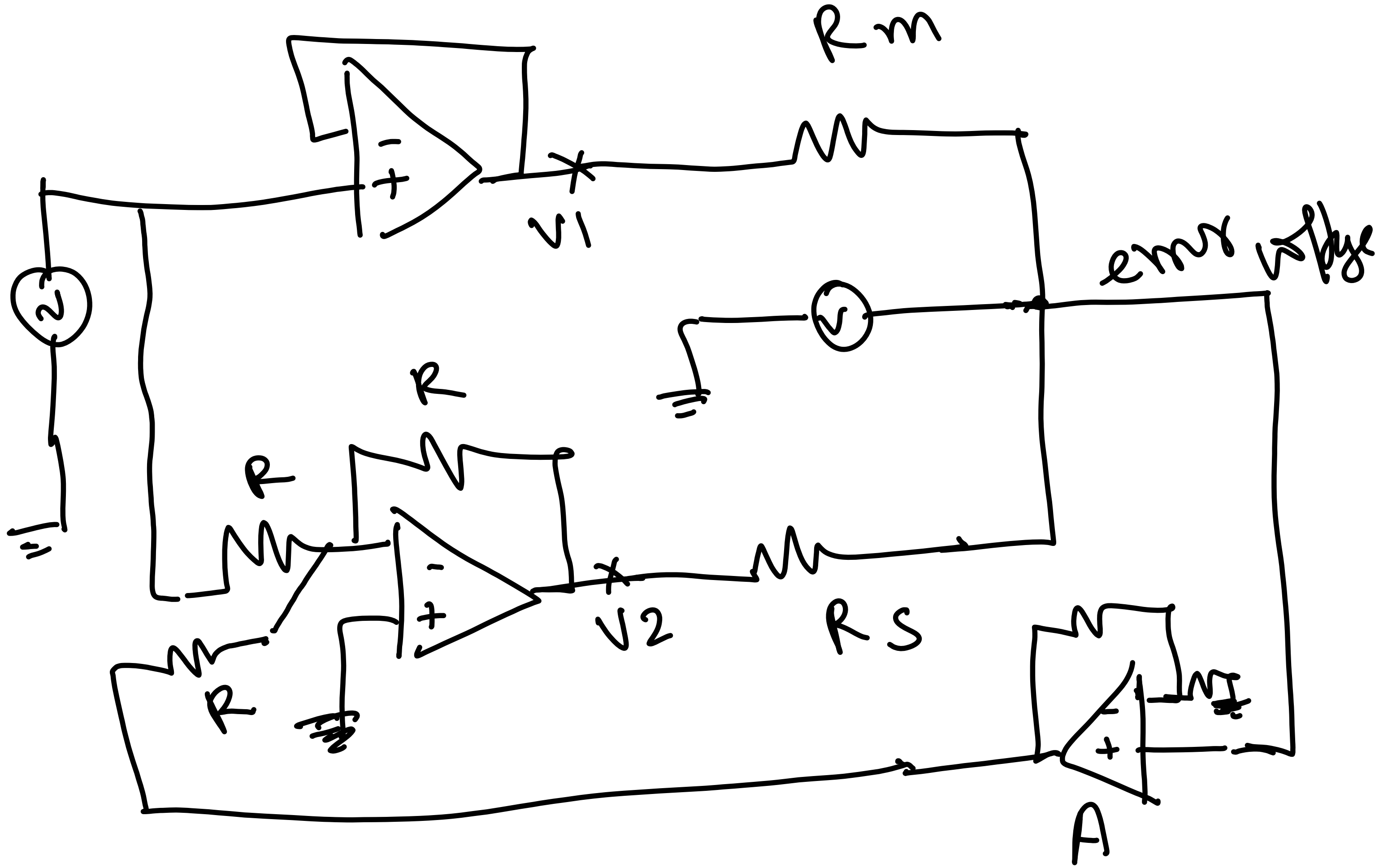
Because the vol change
in the primary affects
proportionally all the secondaries
So the ratio $\frac{V_1}{V_2}$ is not
changing.

② The voltmeter used for
balancing will not affect
the measurement, even if it is
not calibrated.
This is because it is a null
method.

(1) This method is good for high resistance measurement > 10 k up to several 100 MΩ

(2) This is not good for low resistance measurement, because the current drawn from the transformer secondary will produce lot errors.

<u>Dis-advantage</u> : Manual balancing is required



If the gain of the amplifier A is high then bridge will get balanced automatically.

This is because
V_2 is changing on a
opposite way to the
error voltage
so V_1 and V_2 ratio is
adjusted automatically to
balance the bridge

$$\frac{V_1}{V_2} = \frac{r_m}{r_s}$$

But V_1 and V_2 must be
measured.
The error in the measurement
of V_1 and V_2 affects the

Resistance measurement.

So it is not as accurate as ratio transformer bridge

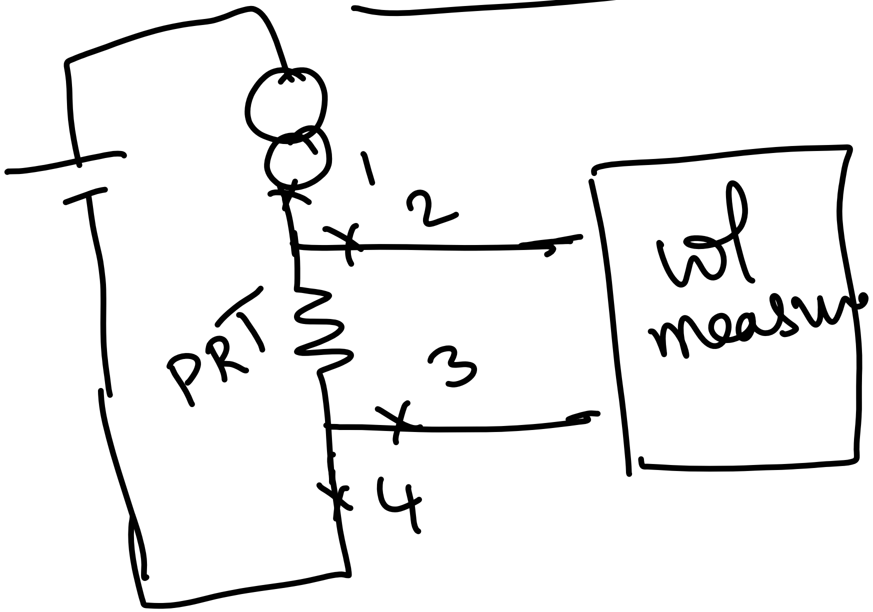
Drawback : If the op amp introduce phase shift error then measurement will not be accurate

The phase shift introduced by the op amps must be very small

So high band width of amp stage is required. or use several op amps with low gain from each stage

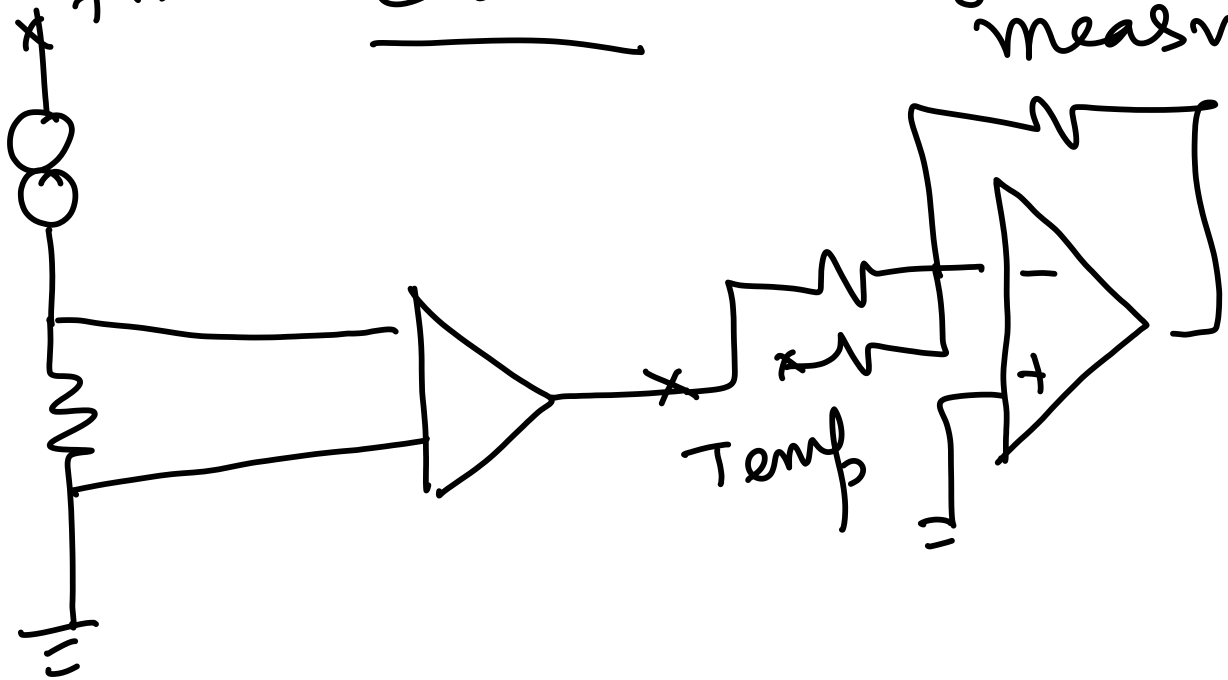
For low frequency measurement say few Hz then phase shift is not a problem.

Platinum resistance Thermometer



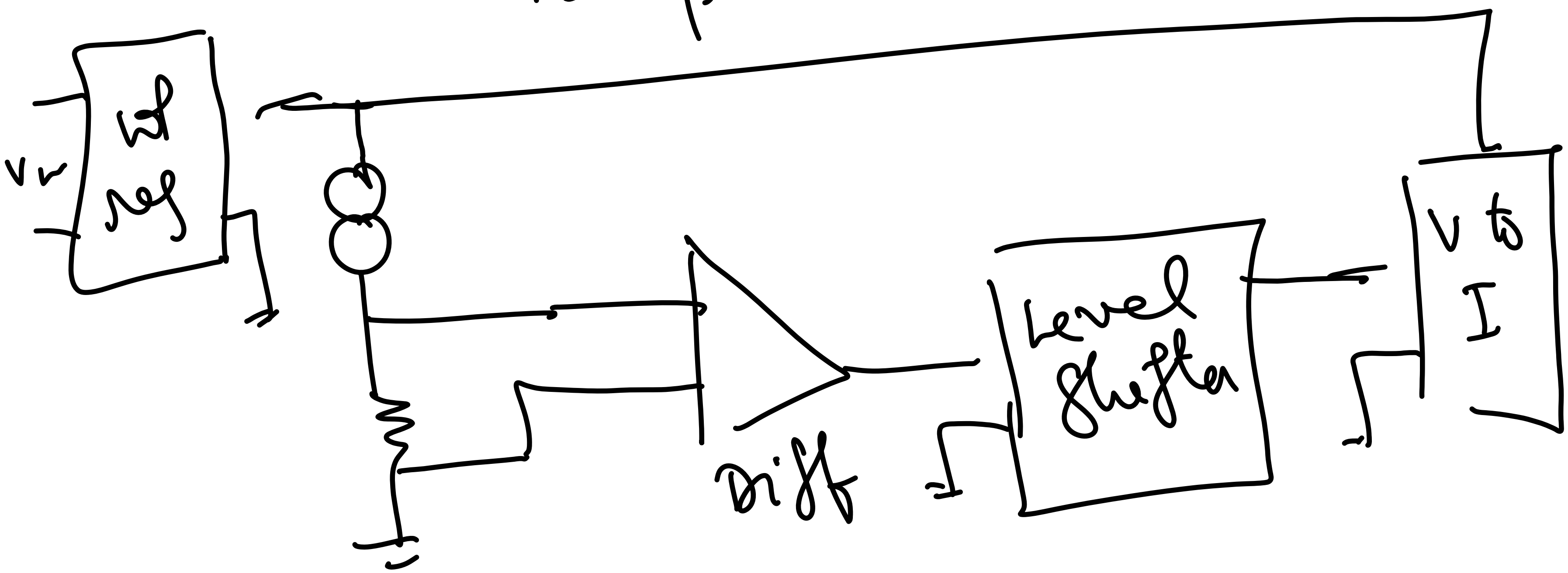
Case-1

voltage based measurement



Case - 2

4-20mA current based
Temperature measurement



4-20mA Temperature Transmitter using PRT

INPUT : ±24V DC

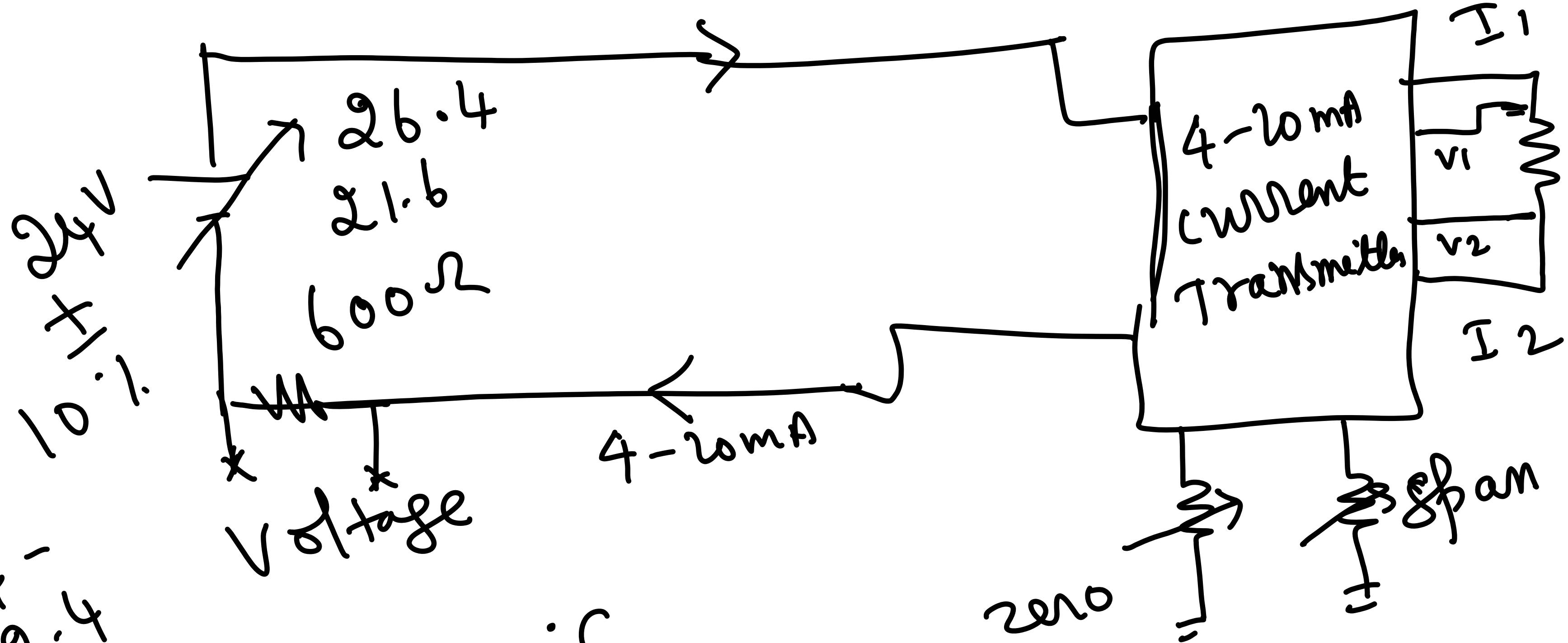
OUTPUT : 4-20mA current

Transducer = PRT

Temp range : 0 to 400°C

Ambient temp range : -20°C to 80°C

Accuracy : ± 1%



$$\begin{array}{r} 24 - \\ 2.4 \\ \hline 21.6 \end{array}$$

At 0°C → I → 4 mA

At 40°C → I → 20 mA

At 0°C → 600 × 4 = 2.4V

At 40°C → 600 × 20 = 12V

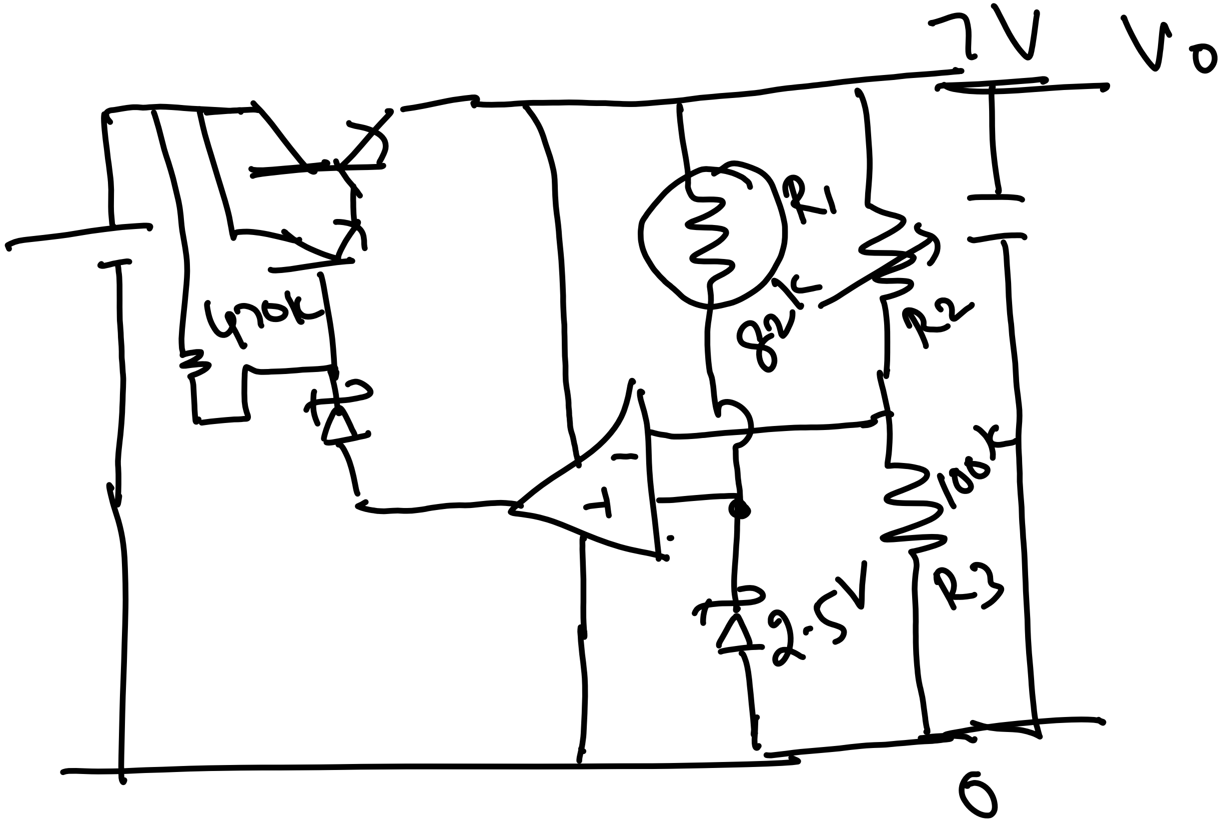
① Supply voltage variation

21.6 to 26.4 V

② Load variation

0 to 600 Ω

So the variation of the load and the variation in the 24V supply varies the available voltage to the current transmitter



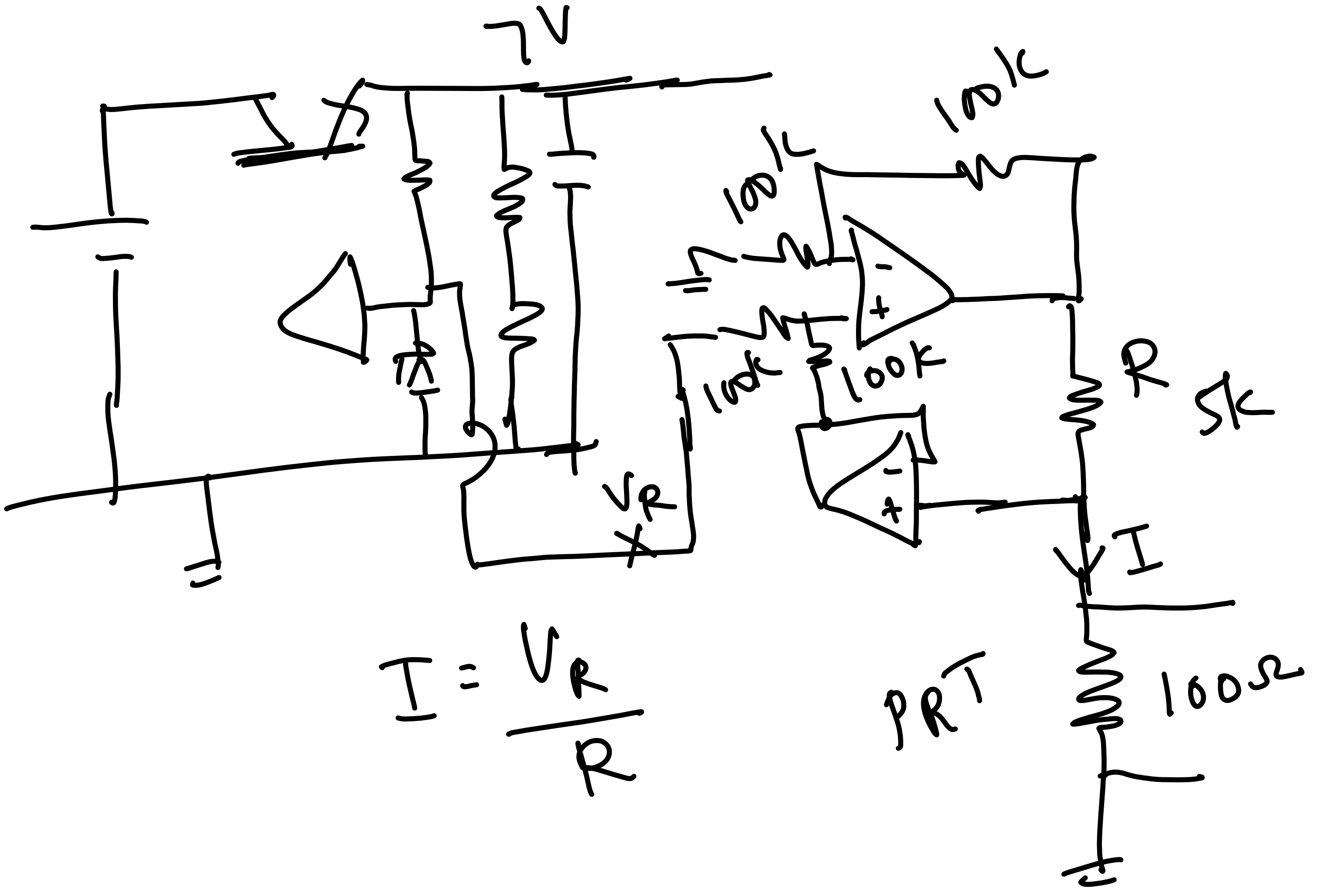
$$R_1 = \frac{(7-2.5)}{50\times10^{-6}} = \frac{4.5\times10^{6}}{50} \simeq 82K$$

Vol acc R_3 = 2.5V

$$I = \frac{2.5}{R3}$$

$$V_o = \frac{2.5}{100k}(100k + R_2)$$

R_2 can be calculated

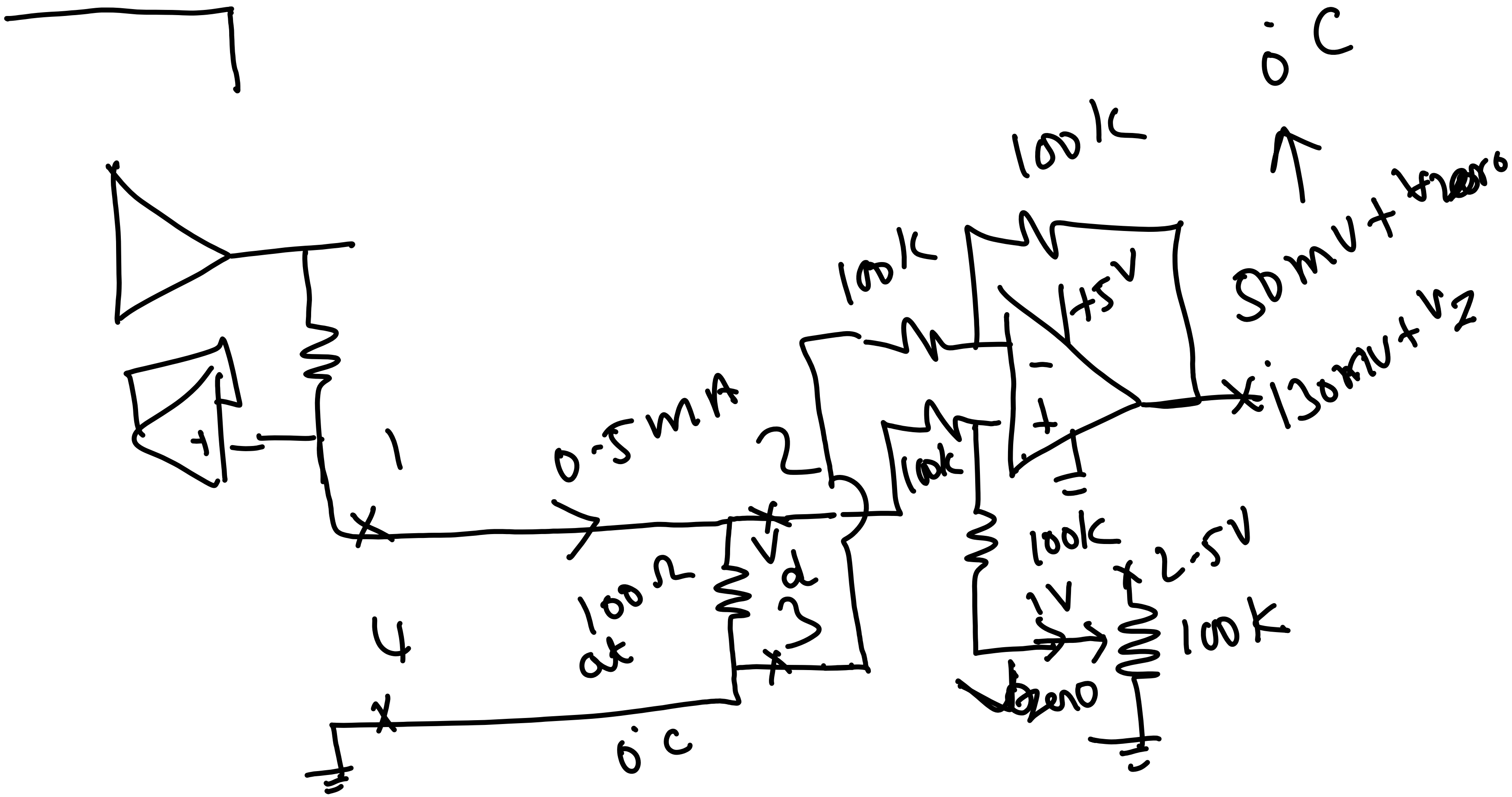


$$I = \frac{V_R}{R}$$

$I = 0.5\,mA$

$I = \frac{2.5V}{R} = 0.5 \times 10^{-3}$

$R = \frac{2.5 \times 10^{3}}{0.5} = 5K$



$$V_o = V_d + V_{zero}$$

$$= 50\,mV + 1\,V$$

$$= 1.05\,V$$

So by adjusting the zero pot we can vary the output voltage

PRT resistance value at $0^\circ C = 100\,\Omega$

PRT resistance at $400^\circ C = 100\,\Omega + \frac{0.4 \times 400}{100} \times 100$

$$= 260\,\Omega$$

At $0^\circ C$ PRT Res

$= 100\,\Omega$

Vol acc PRT $= 100 \times 0.5\,mA$

$= 50\,mV$

At $400^\circ C$ PRT resistance

$= 260\,\Omega$

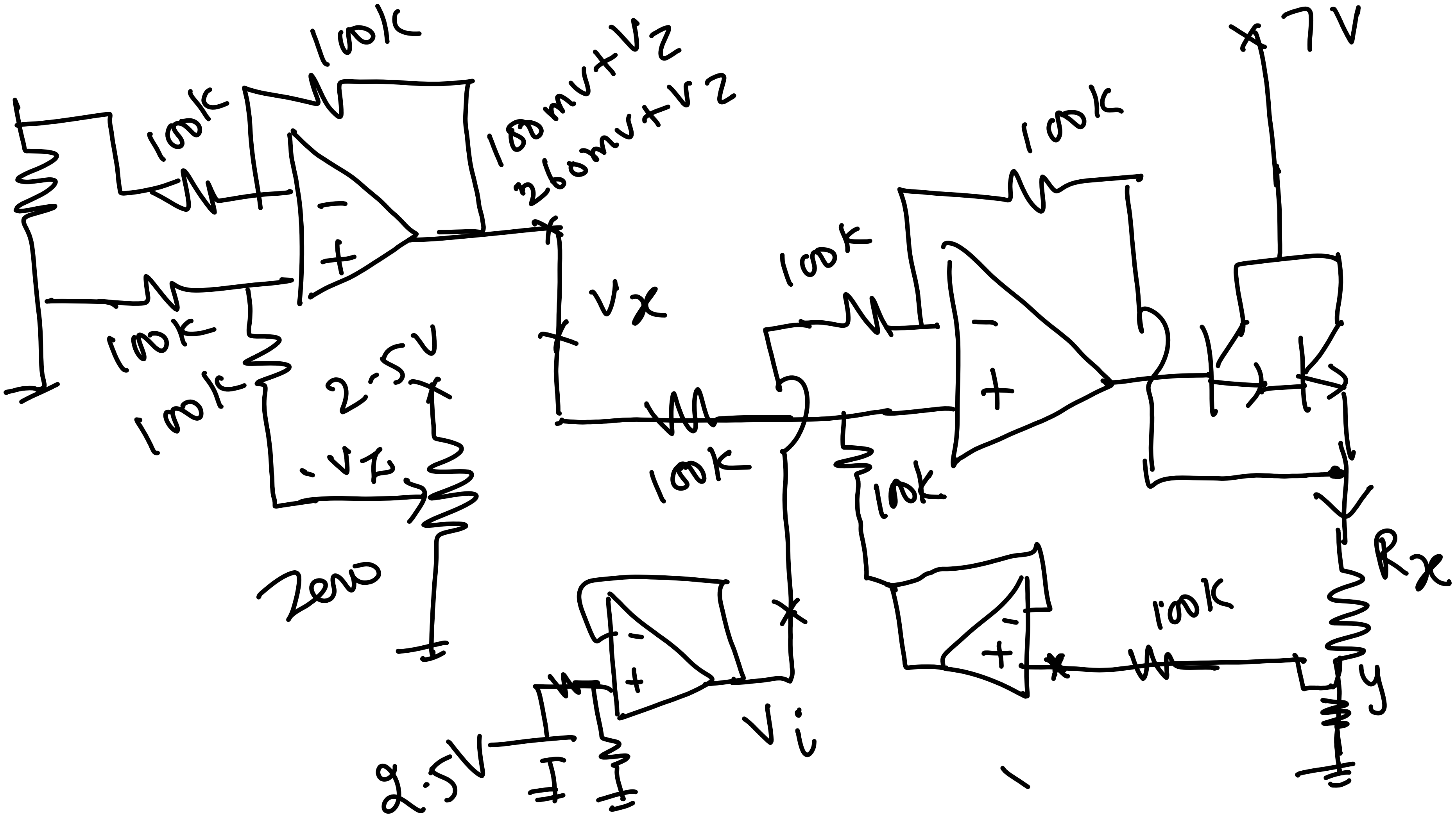
Voltage acc PRT $= 260 \times 0.5 \times 10^{-3}$

$= 130\,mV$

For PRT excitation of 1 mA

Then $V_o = 100\,mV + V_{zero}$ (at $0^\circ C$)

$V_o = 260\,mV + V_{zero}$ (at $400^\circ C$)


100k
100k
100k
100k
100k
100mV+Vz
200mV+Vz
Vx
2.5V
Vz
Zero
100k
100k
100k
100k
+7V
Rx
y
100k
Vi
2.5V

Assuming 3 mA current is drawn by the rest of the circuit we have to fix

$I = 1\,mA$ at $100\,mV + V_Z$

$I = 17\,mA$ at $260\,mV + V_Z$

$$\frac{(V_x - V_i)}{R} = I$$

$V_x = 100\,mV + V_Z$ at $0°C$

$260\,mV + V_Z$ at $40°C$

$v_x = 200\,mv$ at $0^\circ C$

$v_x = 360\,mv$ at $400^\circ C$

at $v_x = 200\,mv$ $\quad \dfrac{200\,mv - v_i}{100} = 1\,mA \Rightarrow$

For $v_i = 100\,mv$ then

& $i = 1\,mA$

For $v_x = 360\,mv$

For $v_i = 100\,mv$

$$I = \frac{260}{100} = 2.6\,mA$$

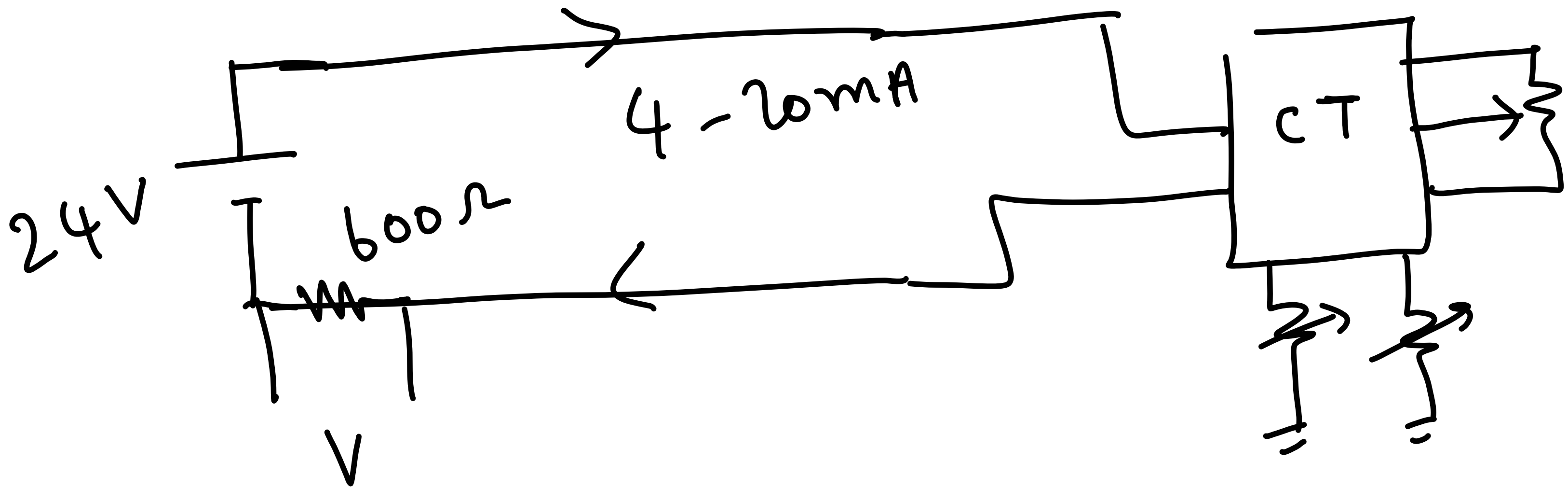
$- 1\,mA$

Adj V_i, R_x, V_Z one can

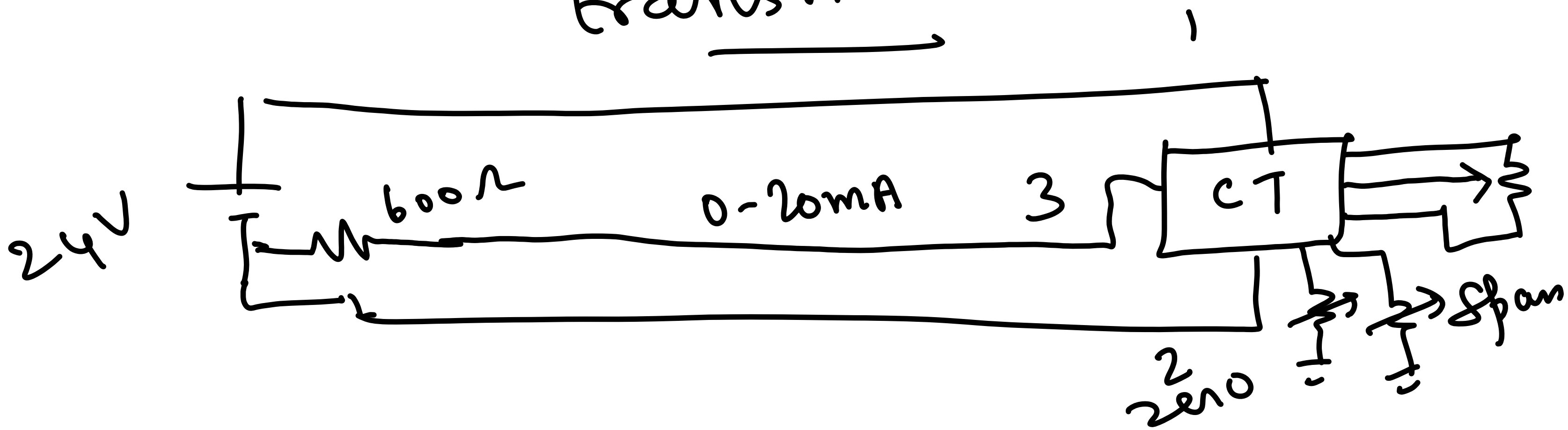
Get 1 mA at $0^\circ C$

and 17 mA at $40^\circ C$

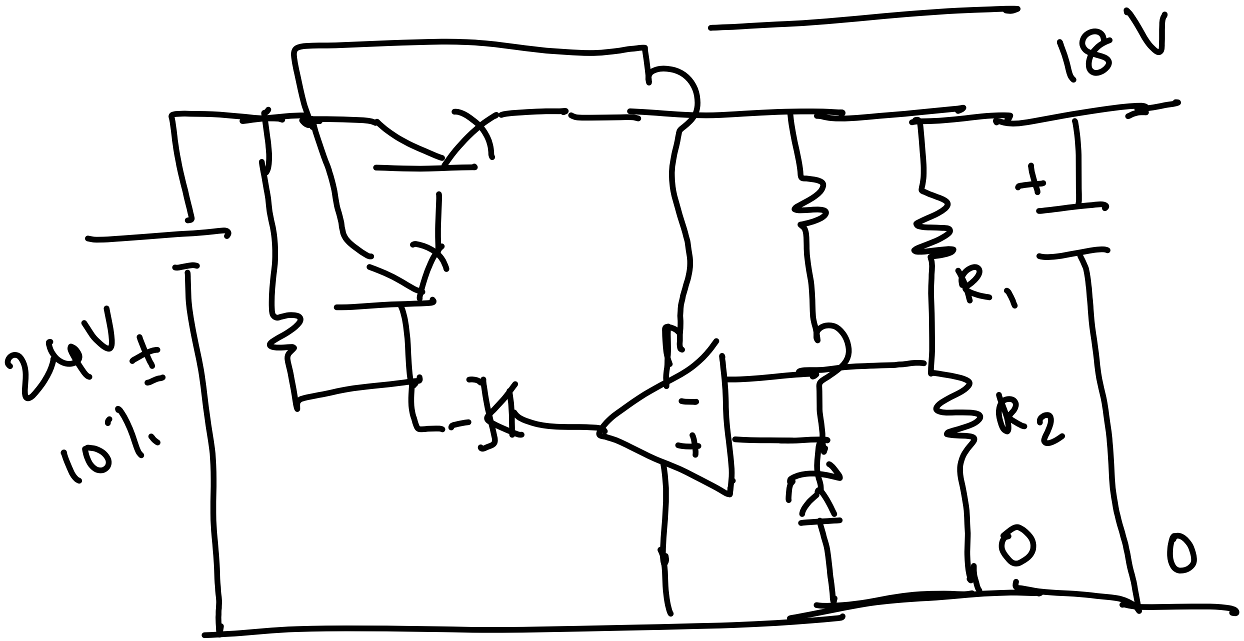
3-wire current transmitters



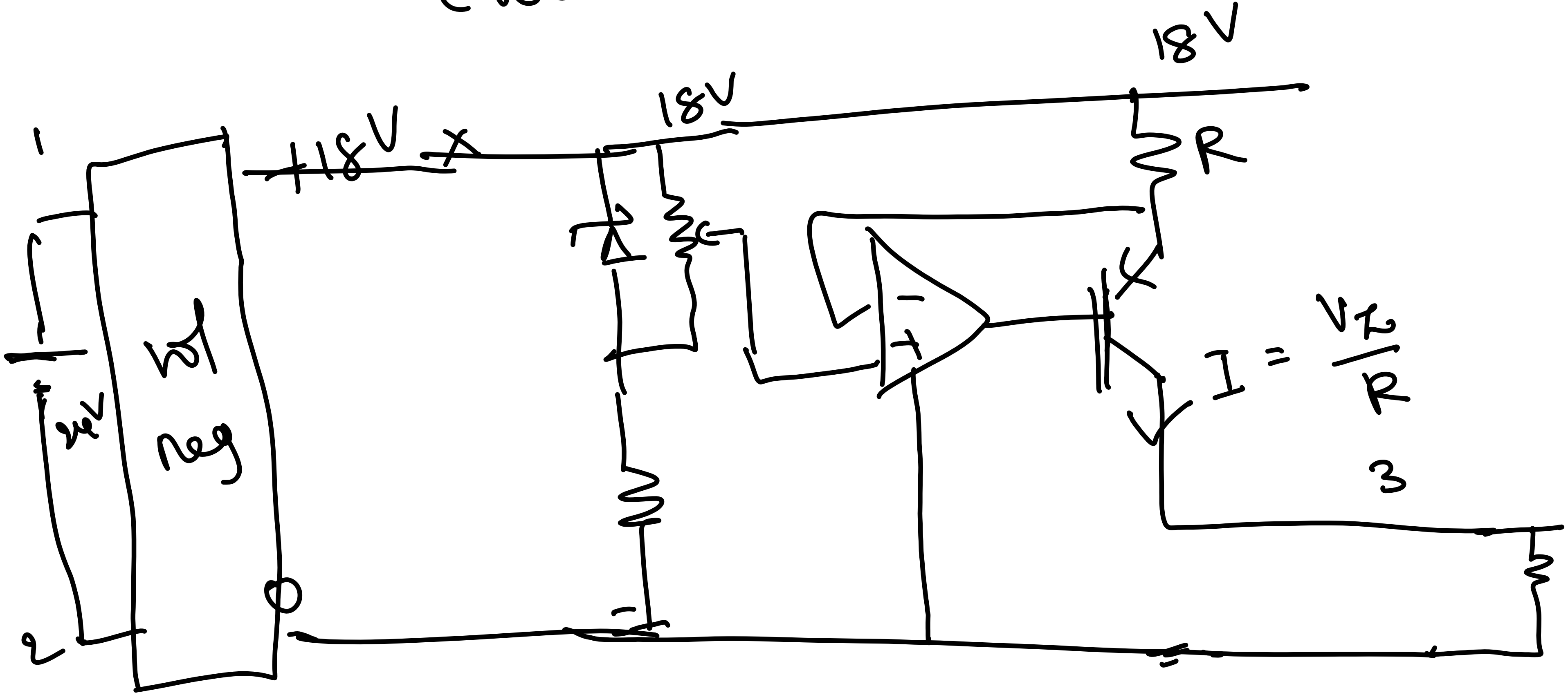
3-wire current transmitters

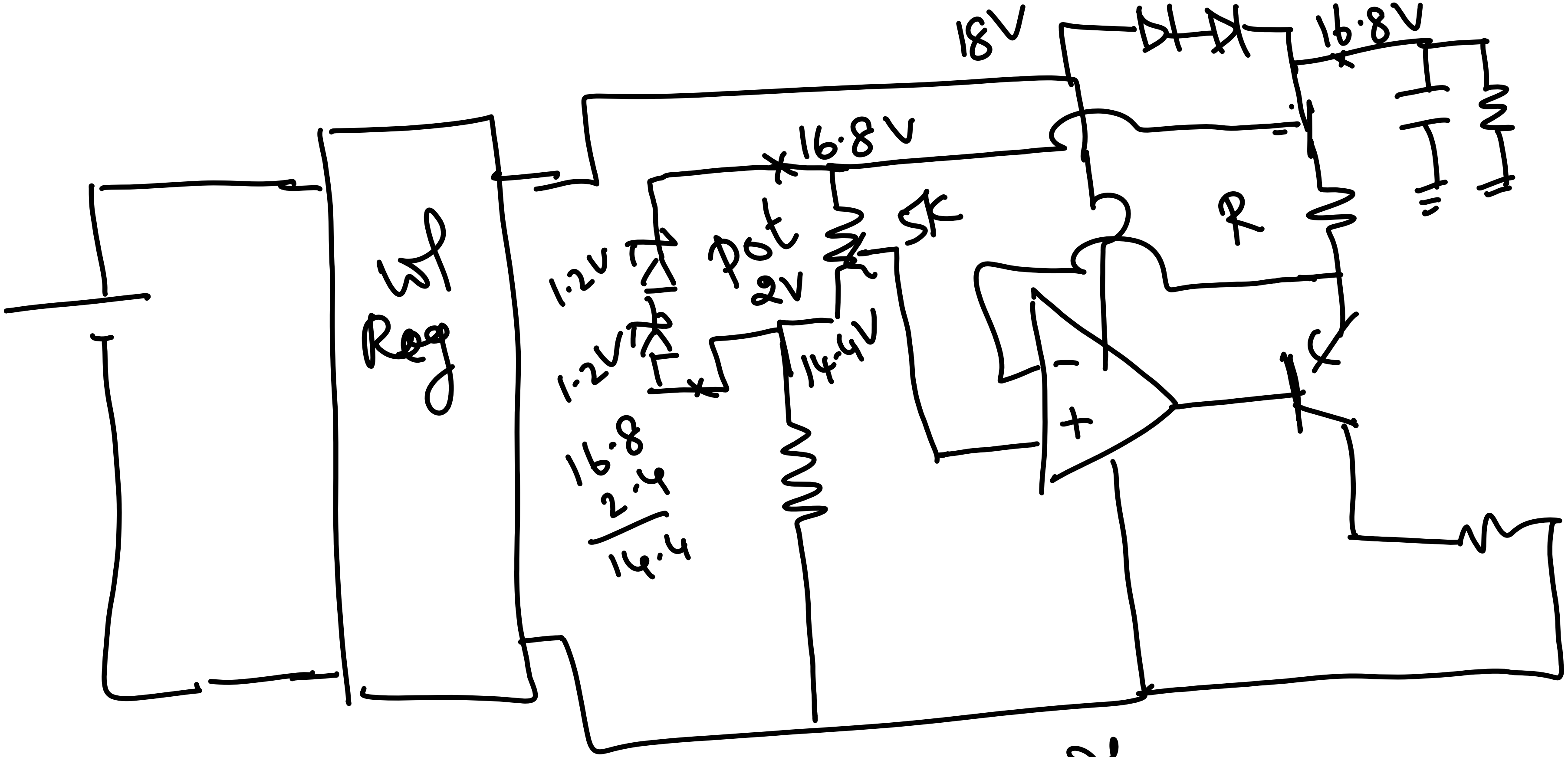


0 - 20 mA current transmitters



Current transmitter





Total expected wl charge = 2.4V

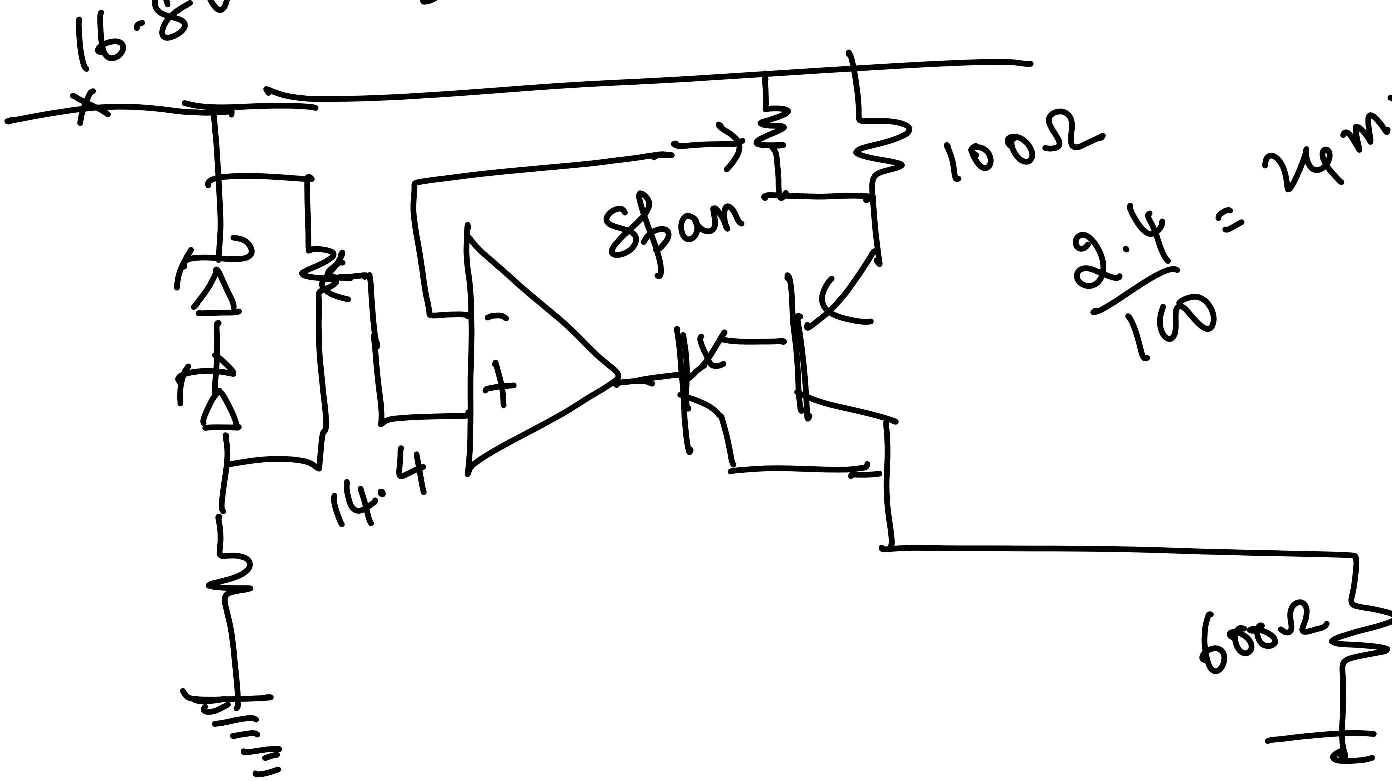
Real expected wl charge = 1.6V

For 20mA expected current charge

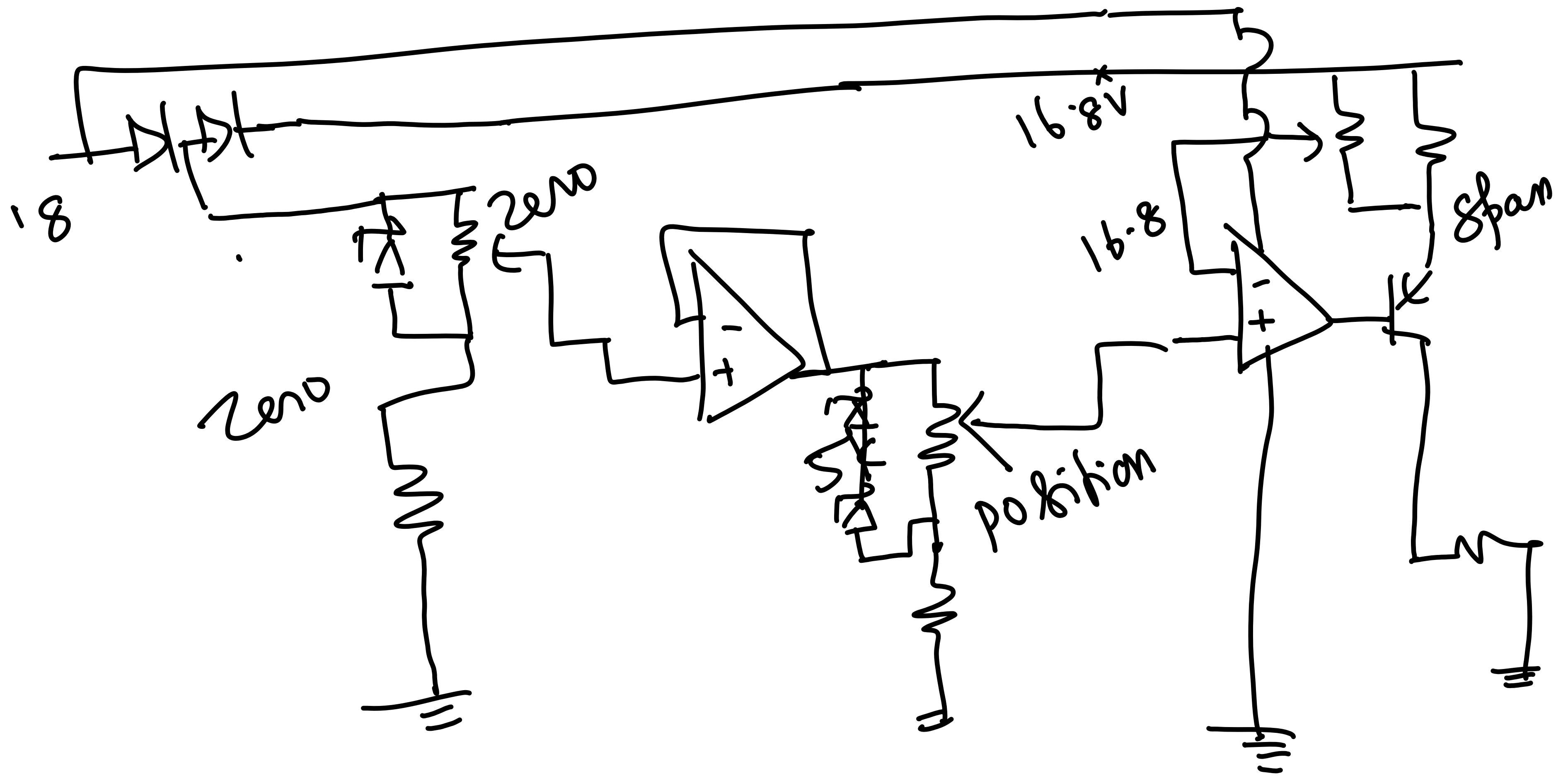
required resistance

$$= \frac{1.6}{R} = 20 \times 10^{-3}$$

$$R = \frac{1.6 \times 10^{3}}{20} = \frac{160\cancel{0}}{2\cancel{0}} = 80\Omega$$

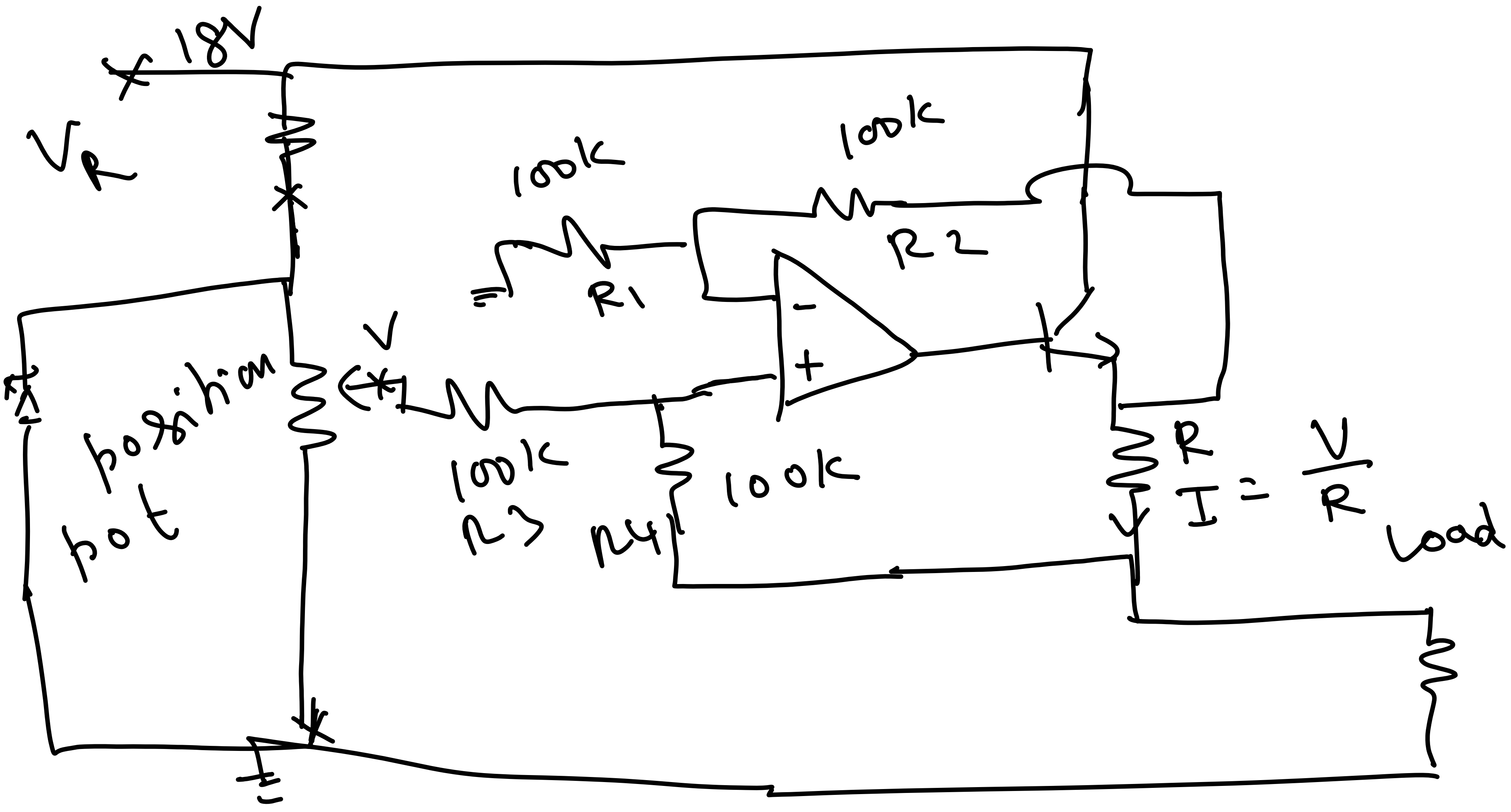


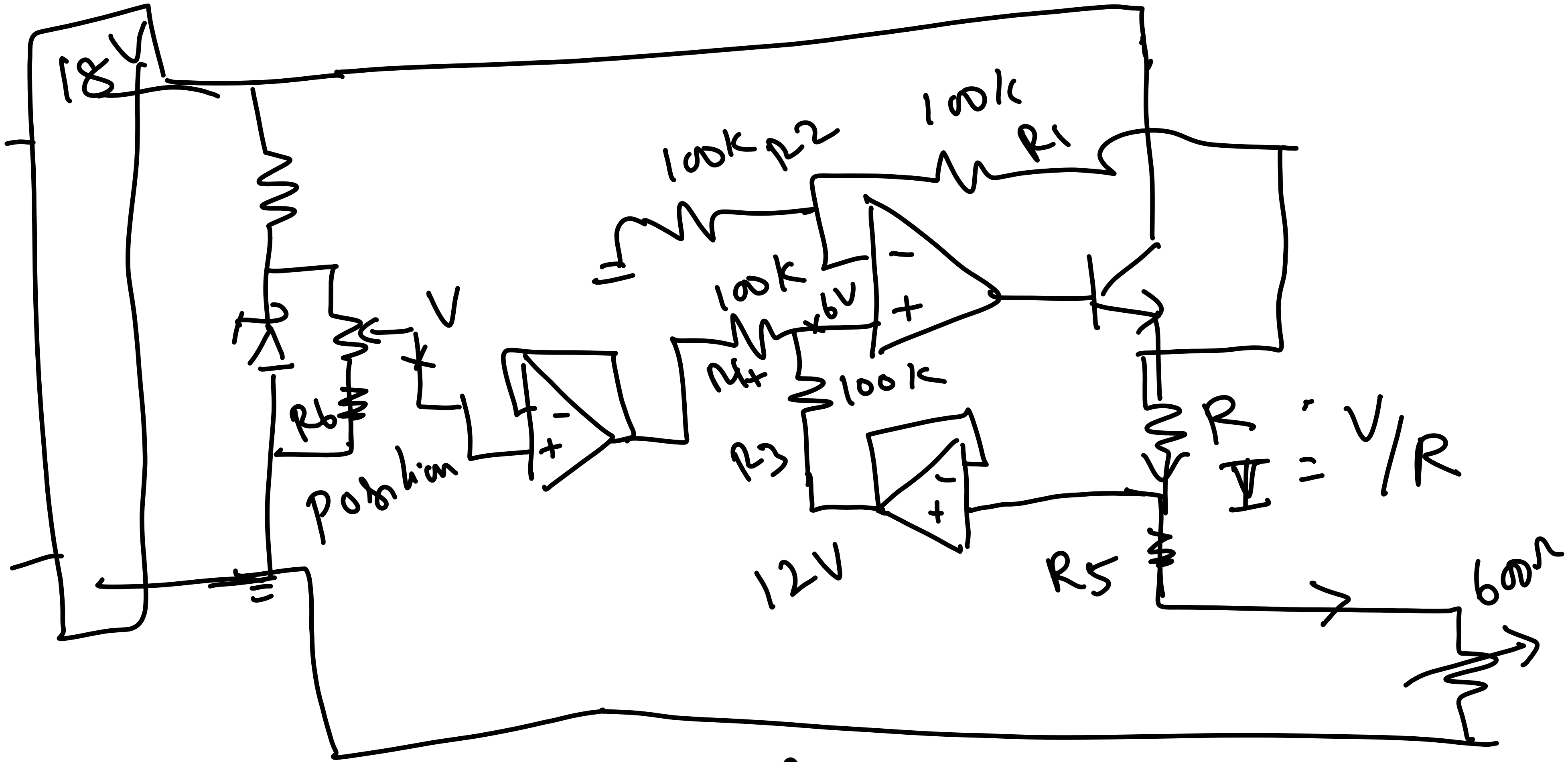
$$\frac{2.4}{100} = 24\,mA$$



18
zero
zero
16.8V
16.8
span
position

Current transmitter
with GND side reference





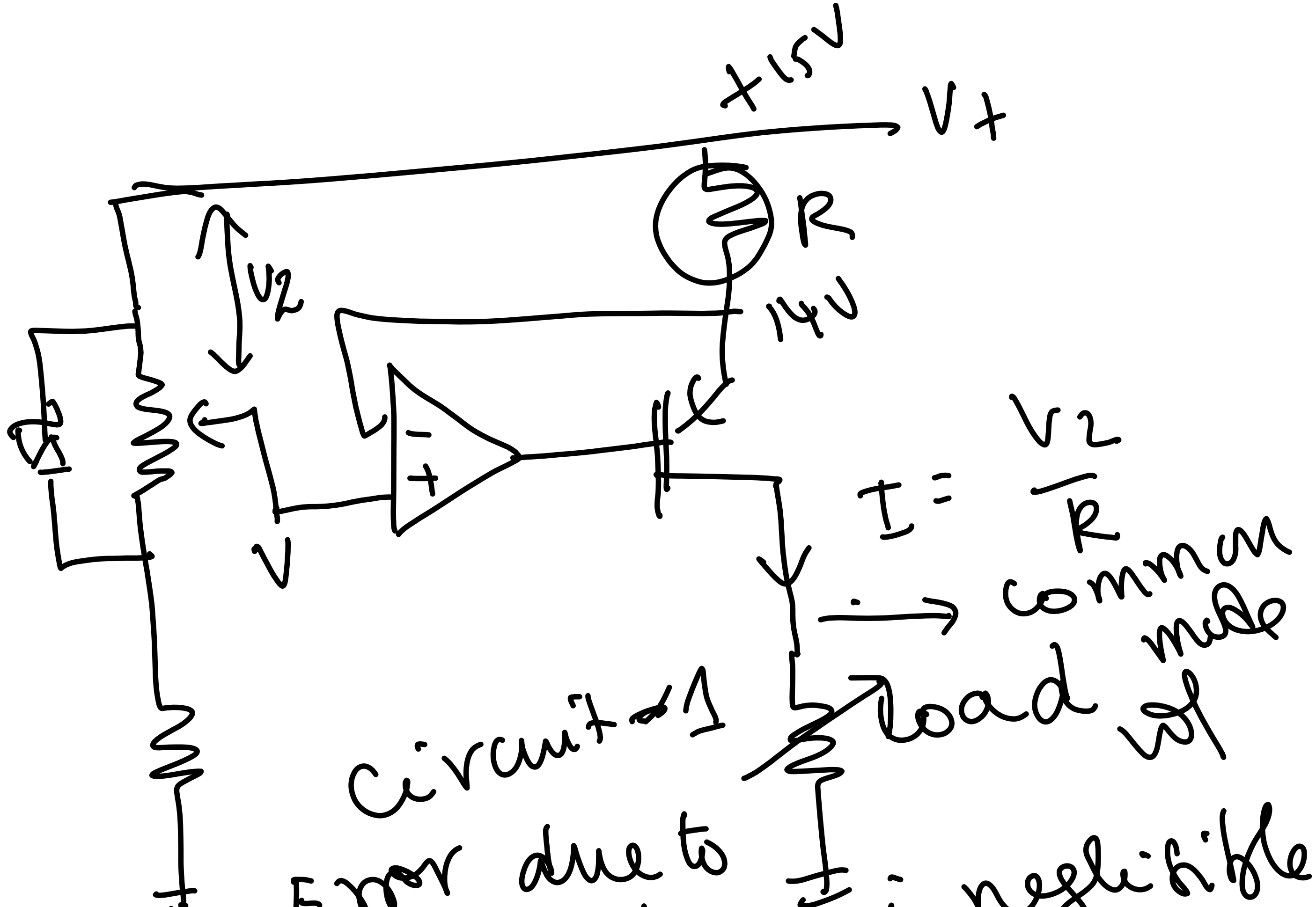
$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

(1) The input voltage to the op amp is not at the supply potential. At the

maximum of amp input terminals see only half the supply voltage

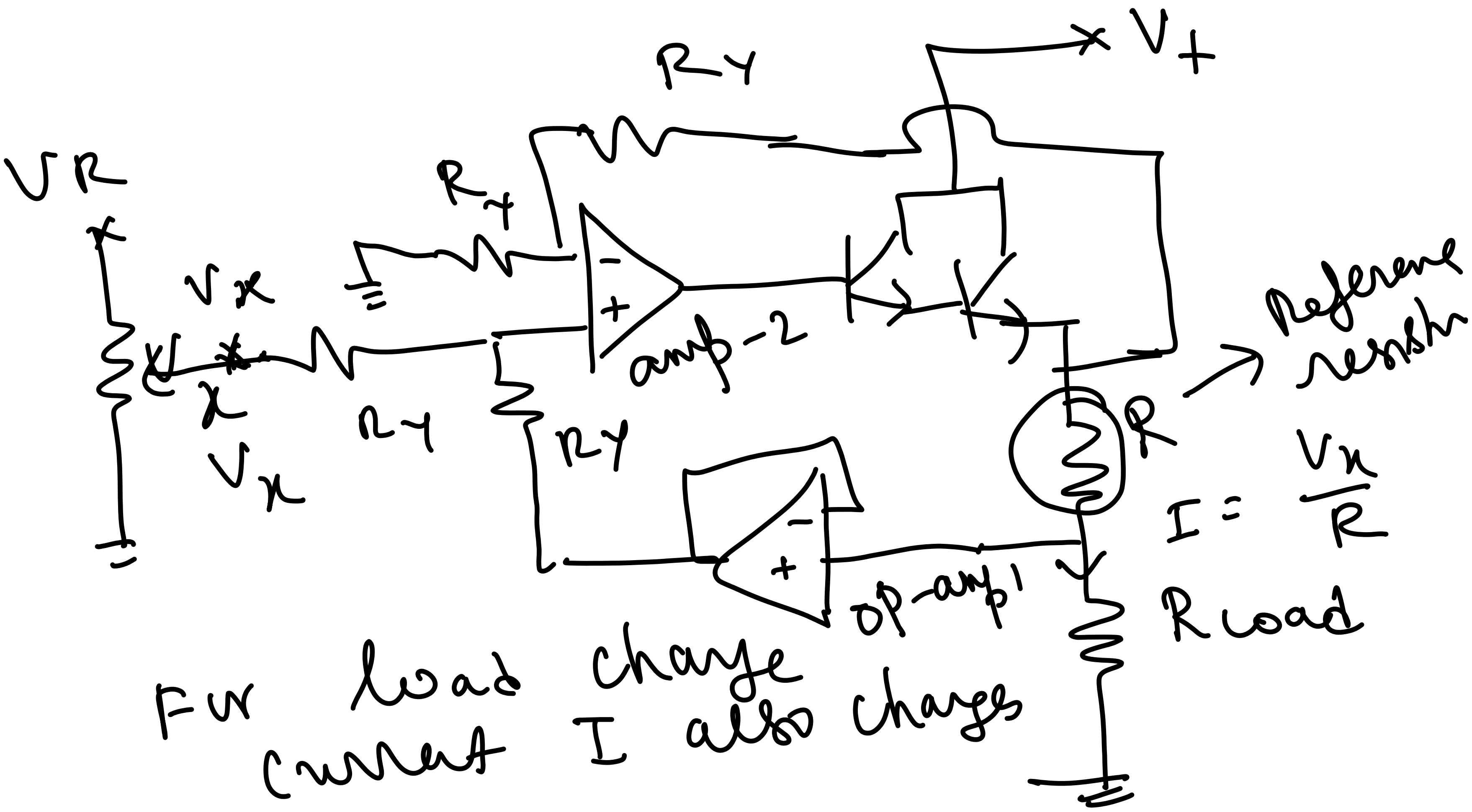
problem

At gnd end the op amps may see zero potential at zero current.
by adding resistors Rf, Rb the problem is not solved.



Circuit 1

Error due to load change is negligible



For load change current I also changes

by a small amount

① Due to finite CMRR OF THE OP AMP-1

② The four resistors (R4) may not equal

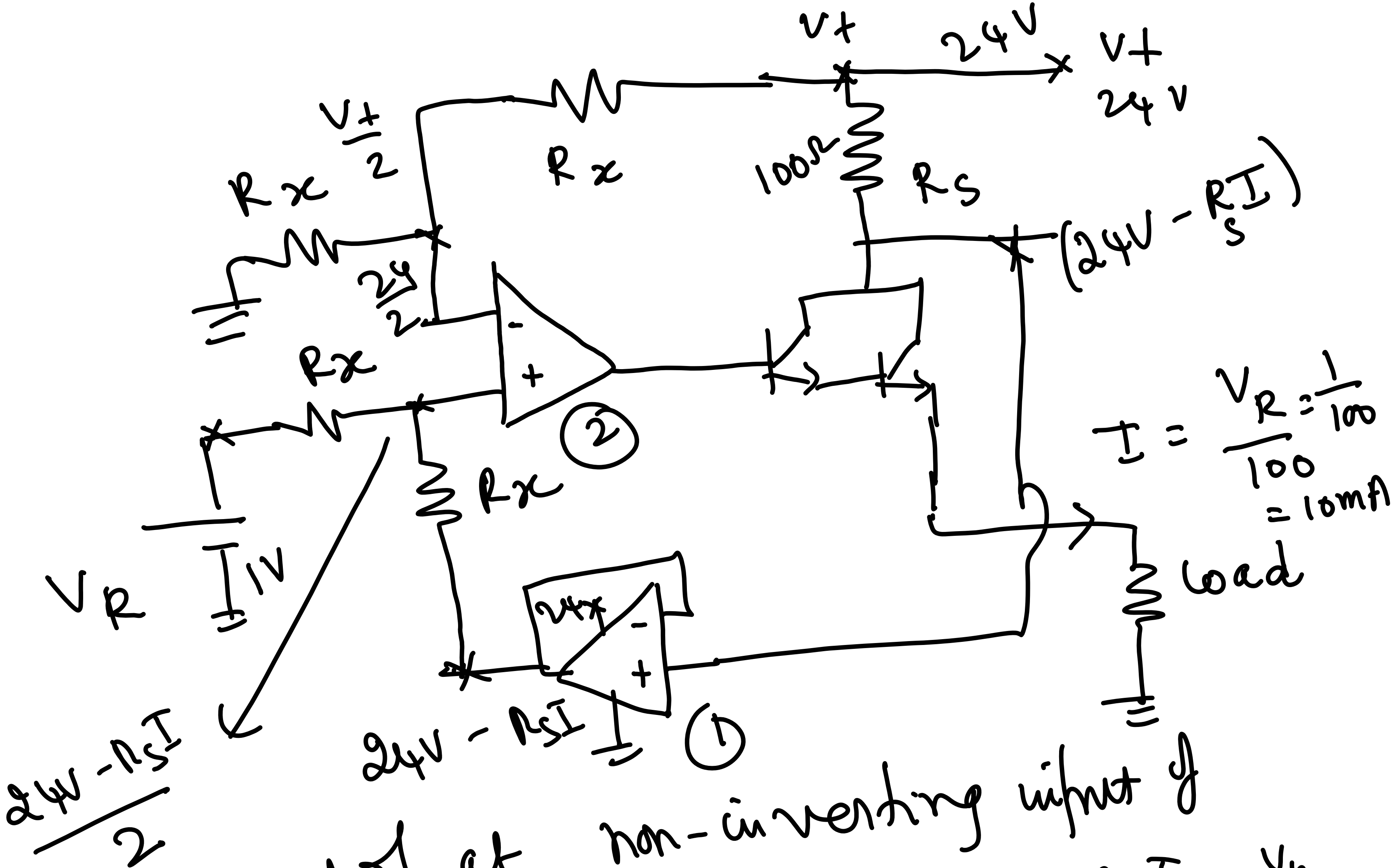
For 1% tolerence resistance then about 2% error in current is expected due to CMV

So current I is expected to change with load

For 12V change across load resistance then the voltage error at op amp 2 $= \frac{?}{?} = \frac{12 \times 2}{100}$

$= \frac{24}{100} = 240\,mV$

$\Delta I = \frac{240\,mV}{R}$



Vol at non-inverting input of op amp-2 $= \frac{24}{2} - \frac{R_S I}{2} + \frac{V_R}{2}$

$V_+ = V_-$

$$R_s I = V_R$$

$$I = \frac{V_R}{R_s}$$

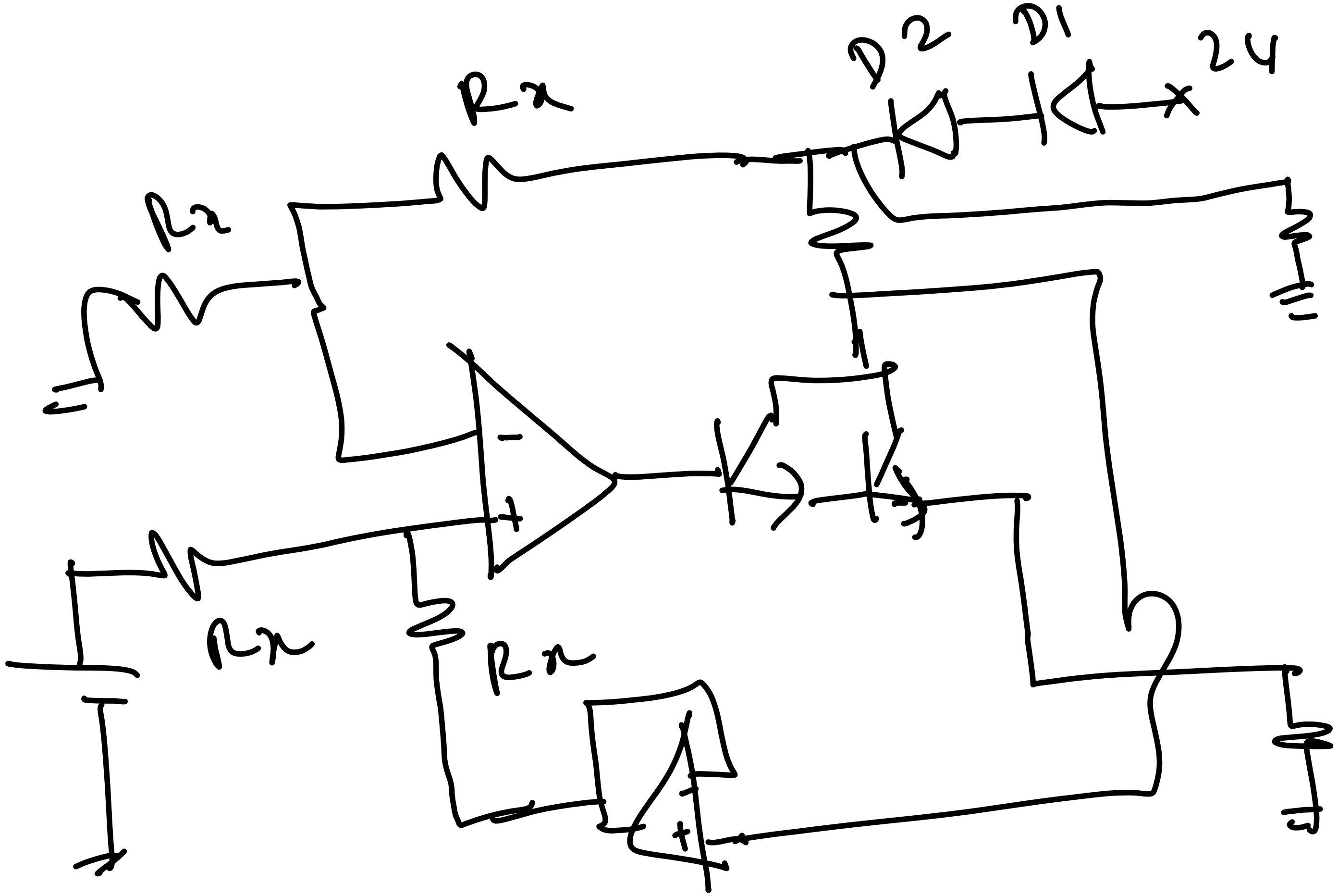
Advantage

When load resistance changes, the voltage seen by the op-amp-1 is not changing. That is no common mode voltage problem.

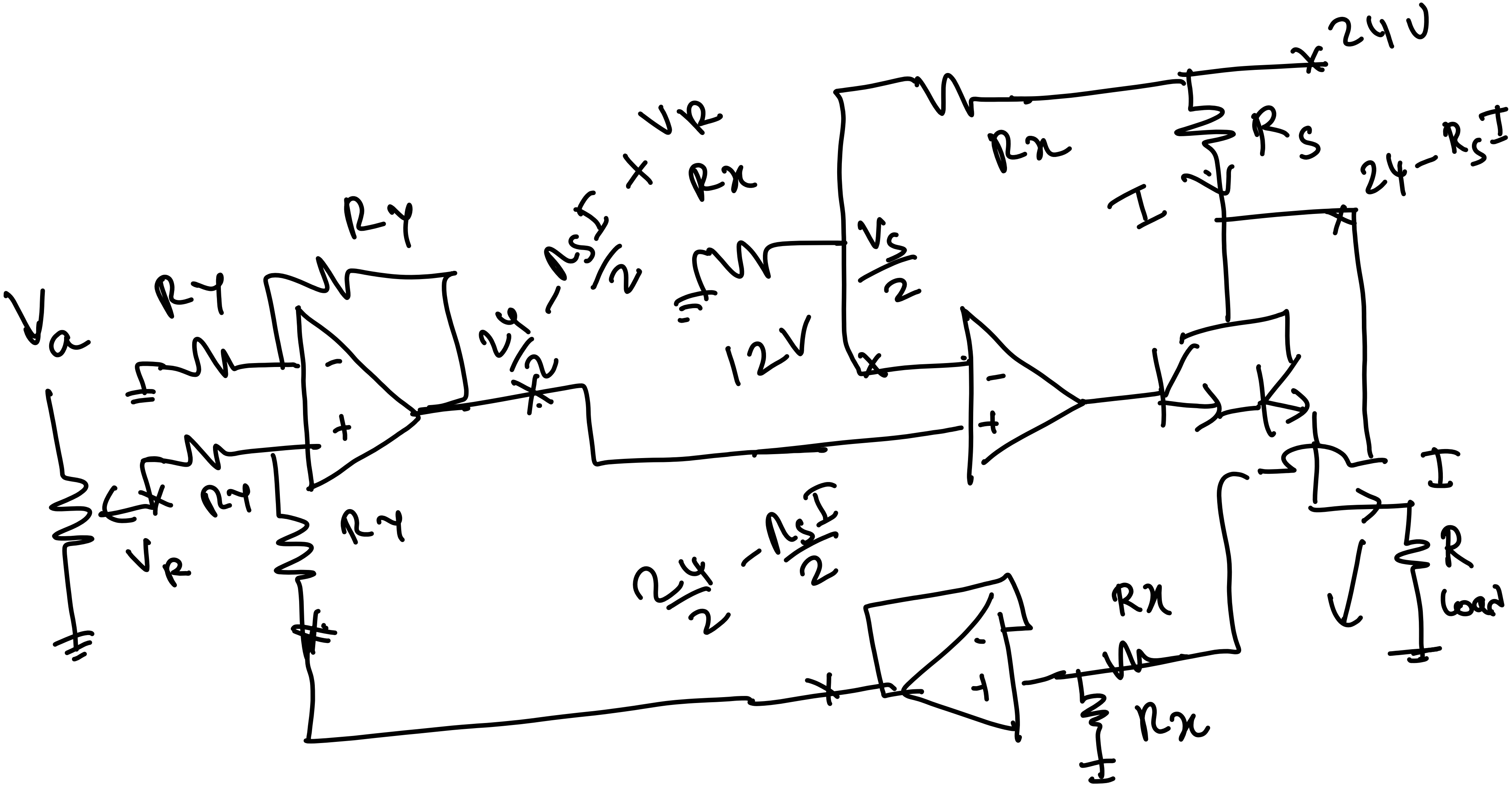
The resistors R_x need not be matched. The mismatch gives error only for a small amount if input supply voltage changes.

Dis advantage

OP - amp-1 sees input voltage very close to its supply voltage. This calls for slightly higher supply voltage to the OP-amps



D2
D1
24
Rx
Rx
-
+
Rx
Rx
-
+



$$I = \frac{2V_R}{R_s}$$

$$V_+ = \frac{24}{2} - \frac{R_s I}{2} + V_R = V_-$$

$$V_- = \frac{24}{2}$$

$$\frac{24}{2} = \frac{24}{2} - \frac{R_s I}{2} + V_R$$

$$\frac{R_s I}{2} = V_R$$

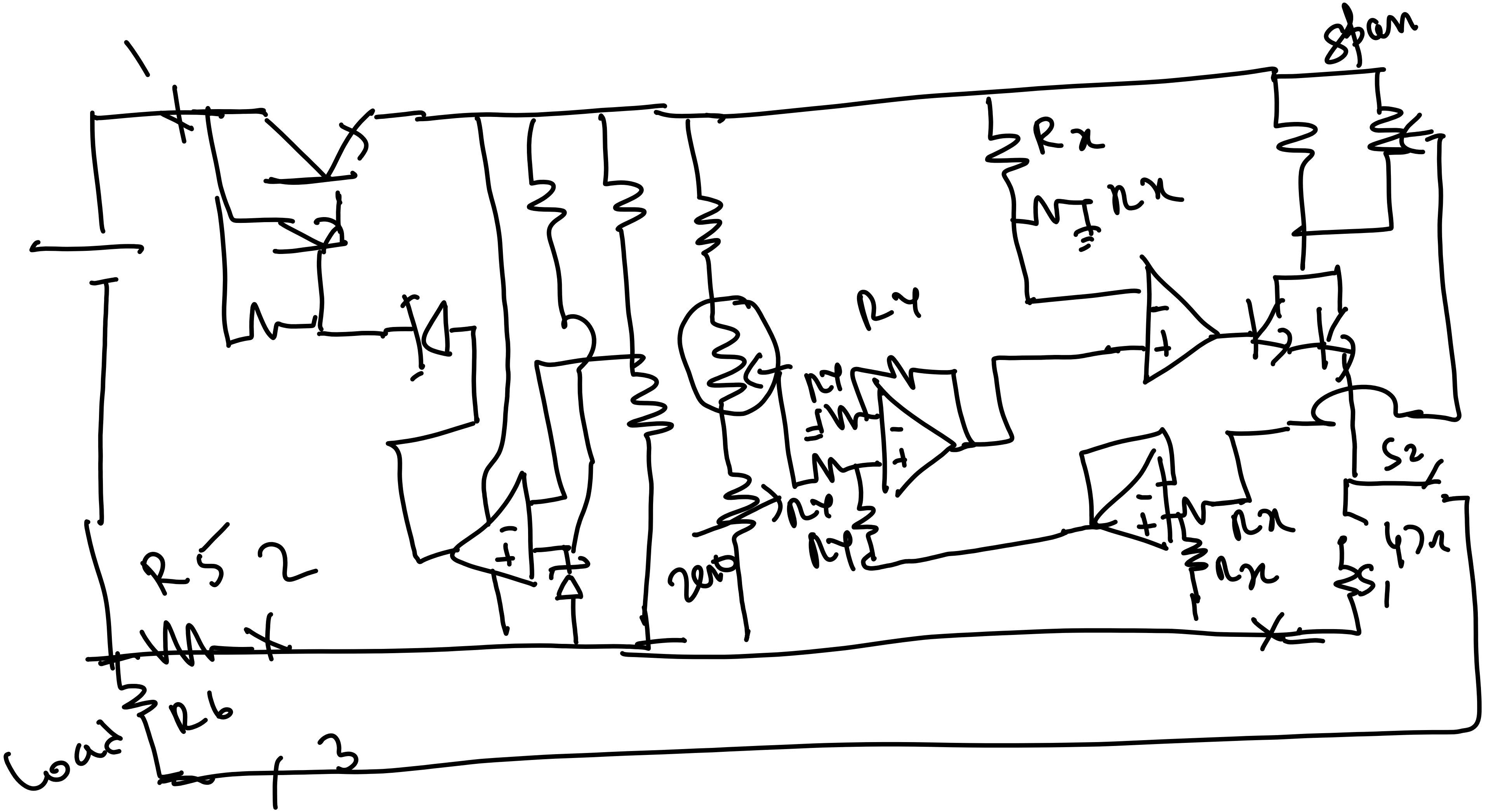
$$I = 2 \frac{V_R}{R_s}$$

Errors

1. R_s drift will introduce a current error
2. All the resistors R_x, R_y drift will introduce error

(3) offset voltage drift and offset current drift of all three of amps introduce current drift error

(4) Reference voltage drift will also introduce error.



1
span
Rx
rx
Ry
ry
zero
S2
S1
rx
R5
2
R6
Load
3

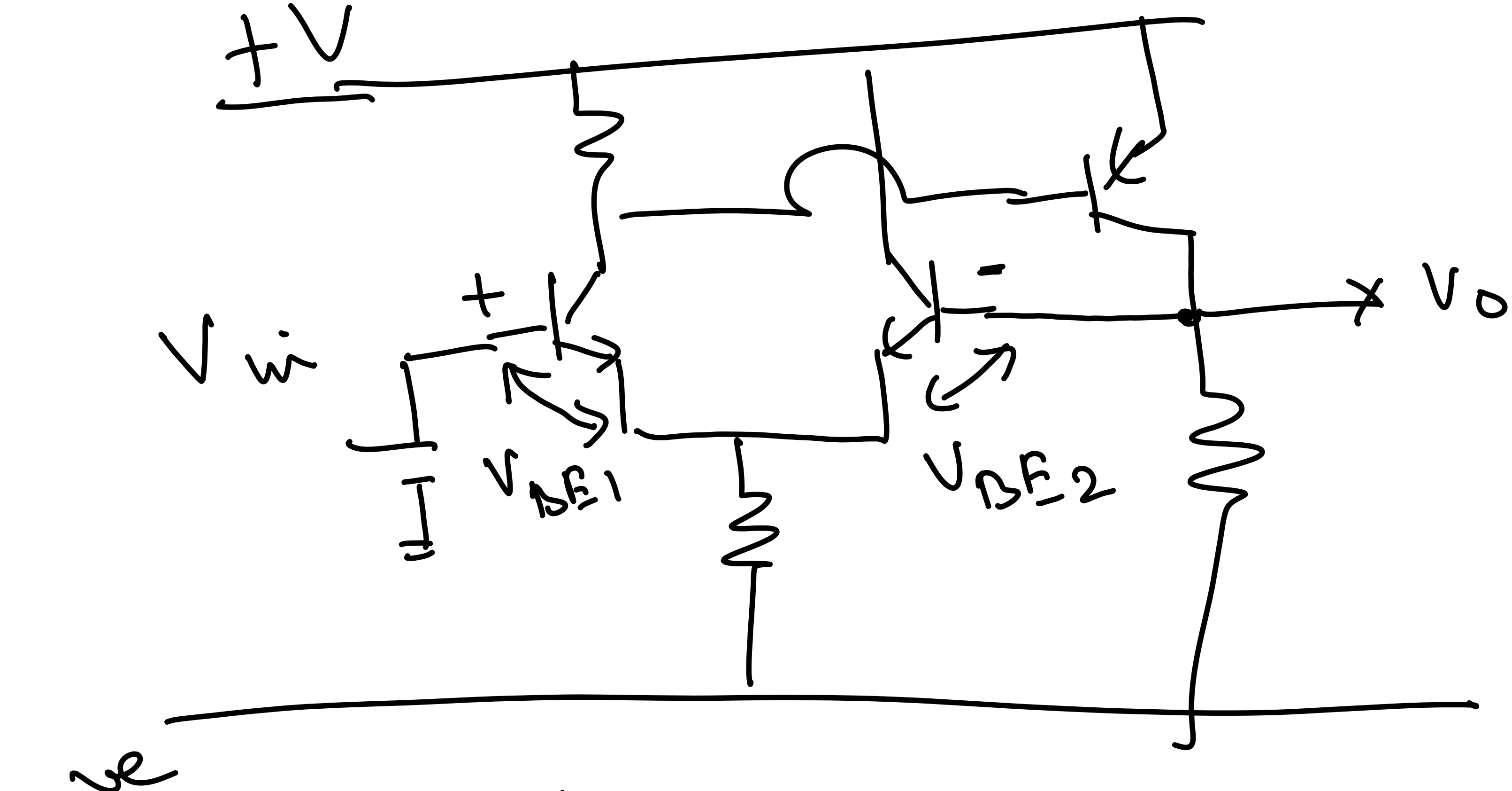
For 2 wire System
S_1 is ON S_2 is off
R_6 is open and R_5 is
connected

For 3 wire System
S_1 is OFF S_2 is ON
R_6 is connected and R_5 is
shorted.

This resistance based / potentiometer based current transmitters are used for displacement measurement in the process control industry.

op amp errors and error budgeting

1. OFF set voltage drift
2. Bias current drift - offset current drift
3. Gain drift error
4. Input resistance
5. Output resistance of the op amp
6. CMRR introduces error

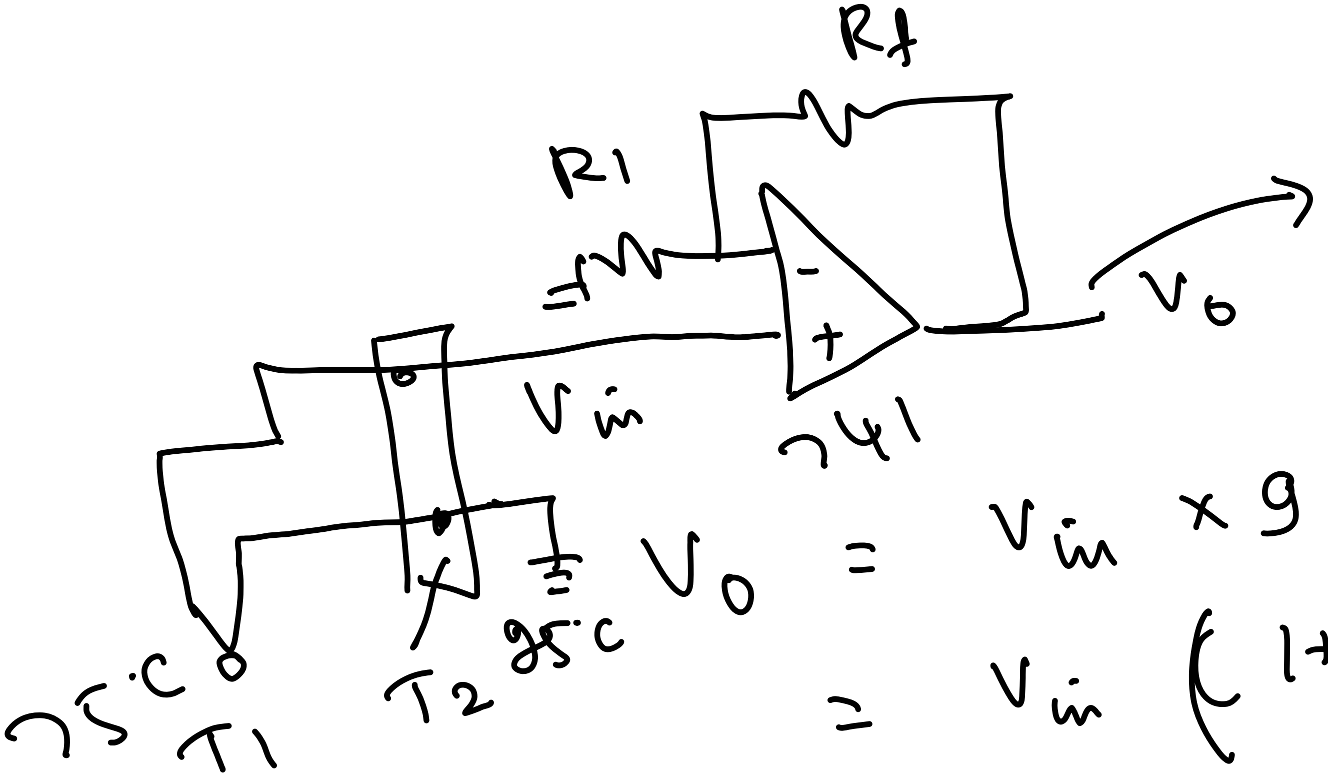


$V_{in} = V_o$

Base - emitter voltage of the input transistors are ~~iden~~ same over a working temp range

ΔV_{BE} = offset voltage

ΔV_{BE} change with temp is called offset voltage drift



$$V_o = V_{in} \times g$$

$$= V_{in}\left(1 + \frac{R_f}{R_1}\right)$$

Sensitivity of the thermocouple

$= 40\,\mu V/°C$

Thermocouple output voltage

$(T_1 - T_2)$ Sensitivity $= V_o$

$T_1 - T_2 = 50^\circ C$

Sensitivity $= 40\,\mu V/^\circ C$

$V_0 = 50 \times 40\,\mu V = 2\,mV$

gain $= 25 = \left(1 + \frac{R_f}{R_1}\right)$

$R_f = 24k$

$R_1 = 1k$

Output of the op amp

$= 25 \times 2\,mV = 50\,mV$

For 741 op amp

Offset vol drift = $\pm 15 \mu V/^\circ C$ max

For ambient temp variation of $-20^\circ C$ to $+80^\circ C$

Total offset vol drift

$= 15 \times 100 = \pm 1.5\,mV$

Total output vol

$= (2\,mV \pm 1.5\,mV)\,25$

error due to offset vol drift $= \pm 25 \times 1.5\,mV = \pm 37.5\,mV$

The expected error
at the output $= \pm 37.5^\circ C$
This error is large for this
$40 \mu V/^\circ C$ signal provided
by the thermocouple

What is to be done

(1) Go for low drift
of amps $\rightarrow$ LM714
LM07

offset vol drift $\rightarrow$ $0.5 \mu V/C$

For $\pm 0.5\,\mu V/^\circ C$ of amp

the thermocouple amp error

$= 100 \times 0.5\,\mu V$

$= 50\,\mu V$

Total error $= 50\,\mu V$ at the input

At the input signal sensitivity

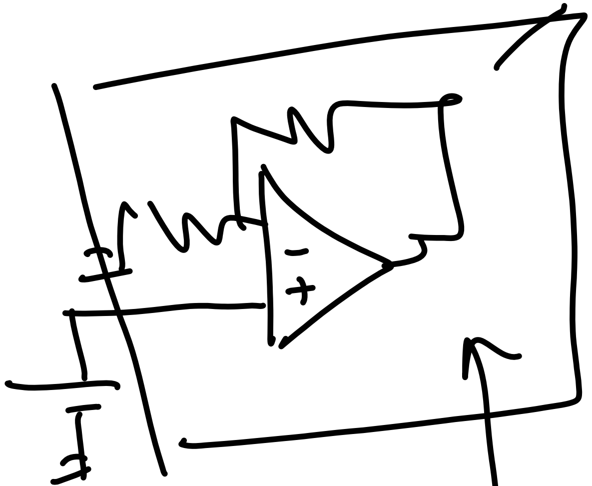
$= 40\,\mu V/^\circ C$

Total error in $^\circ C$ $= \frac{50}{40} = 1.25^\circ C$

Very low DC signals

① Use oven

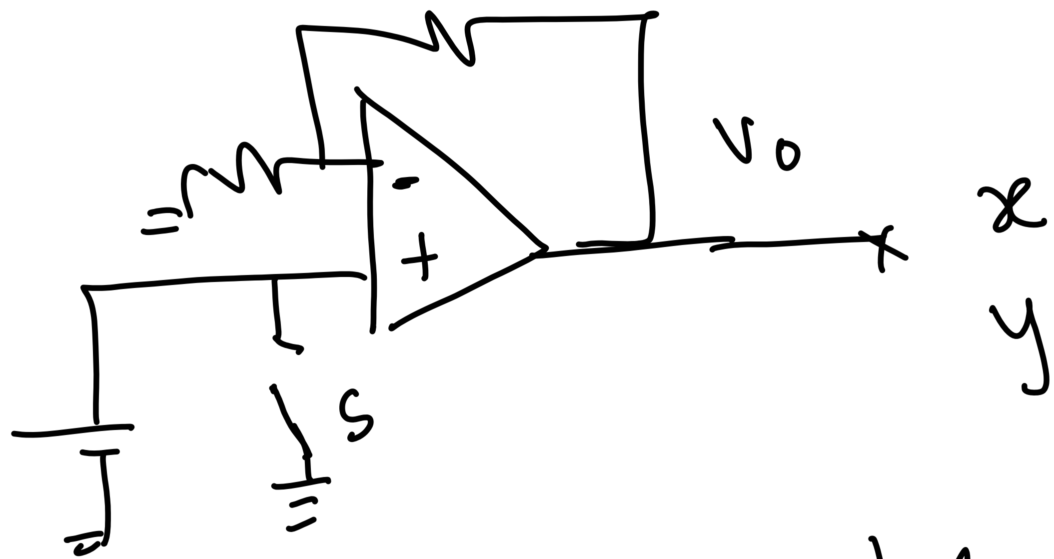
90°C



② Chopper stabilized op amp

0.01 μV/°C

③ Go for Auto zero method



When the input is grounded

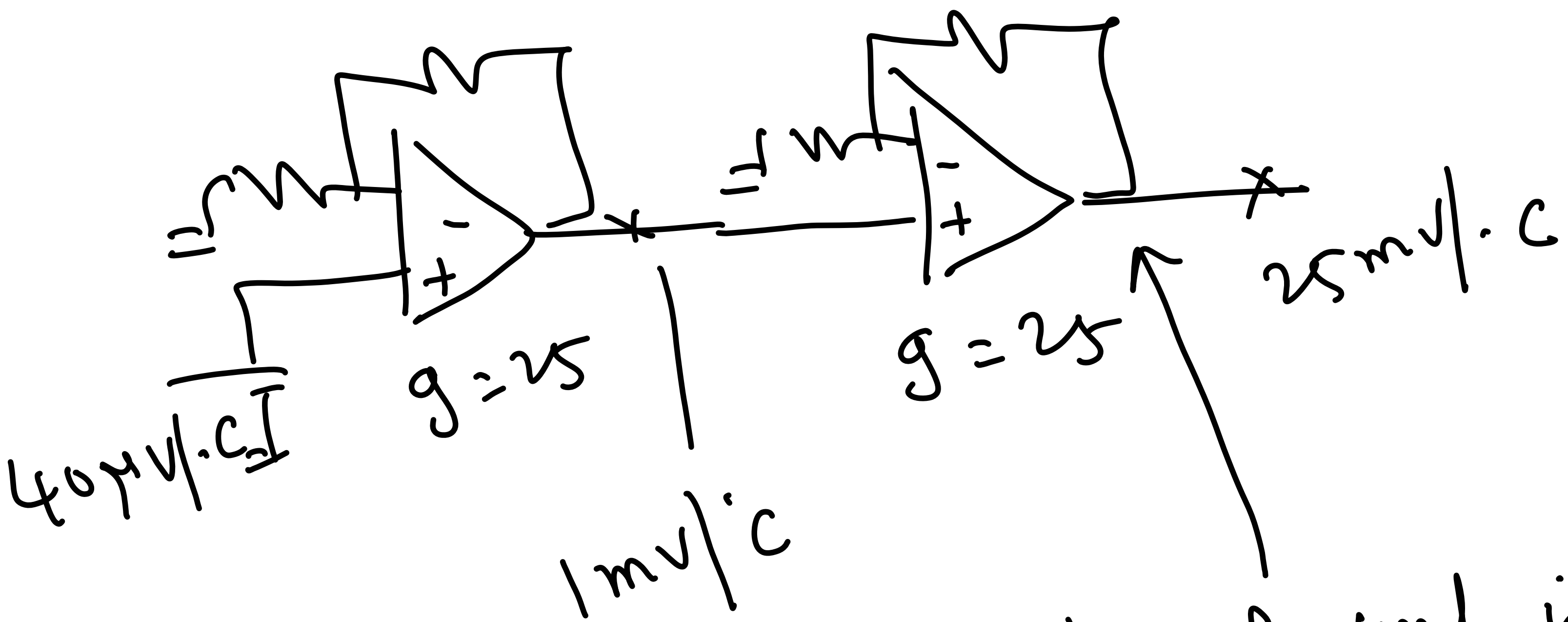
$$V_o = V_{OFF} \times g$$

V_{OFF} changes with ambient temp.

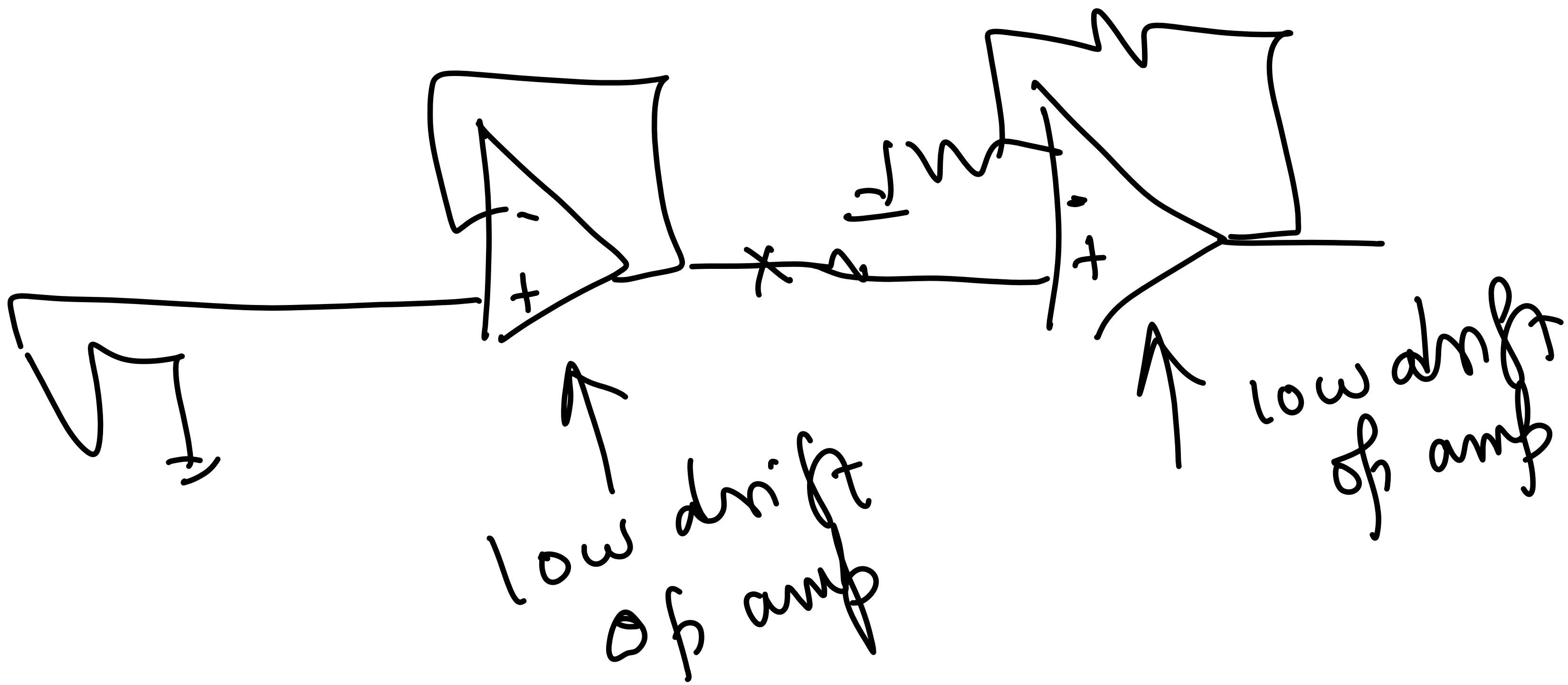
① Measure the output by grounding the input
Let it be x

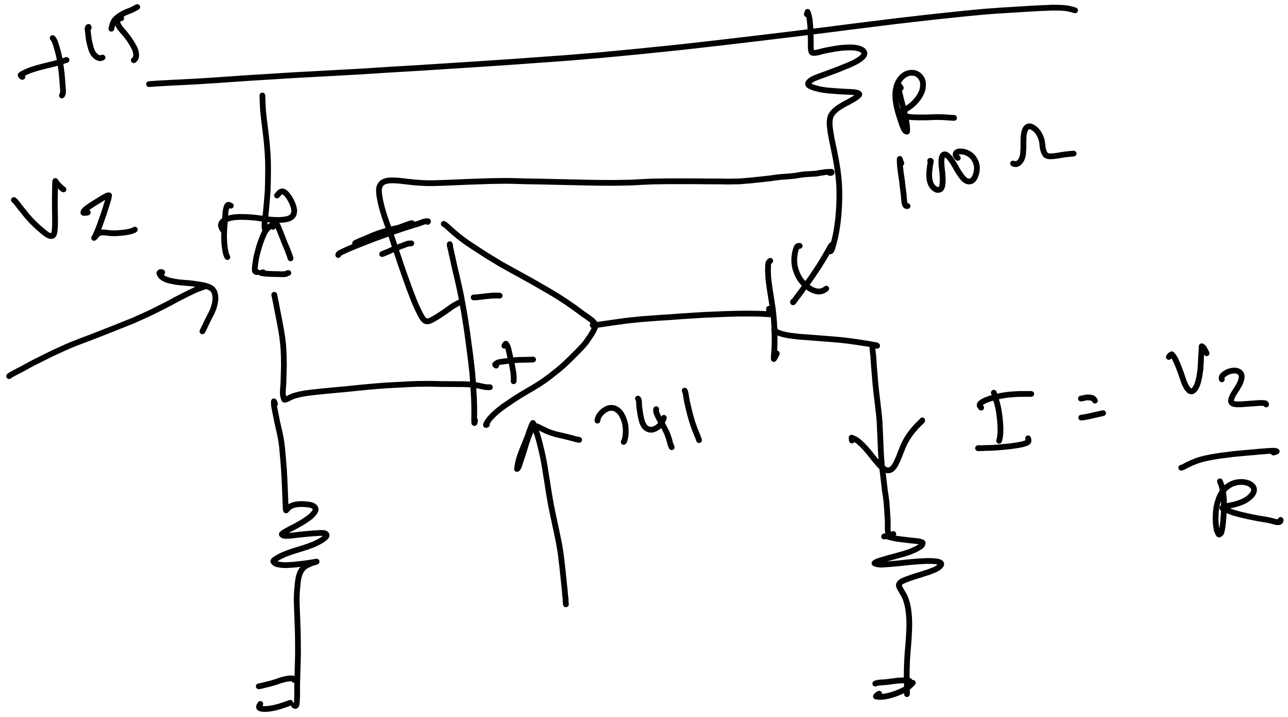
② Measure the output by applying the input ω
Let it be y

$x - y$ = Signal



Drift of the Ist op amp is very important and not the second op amp drift


low drift
op amp
low drift
op amp

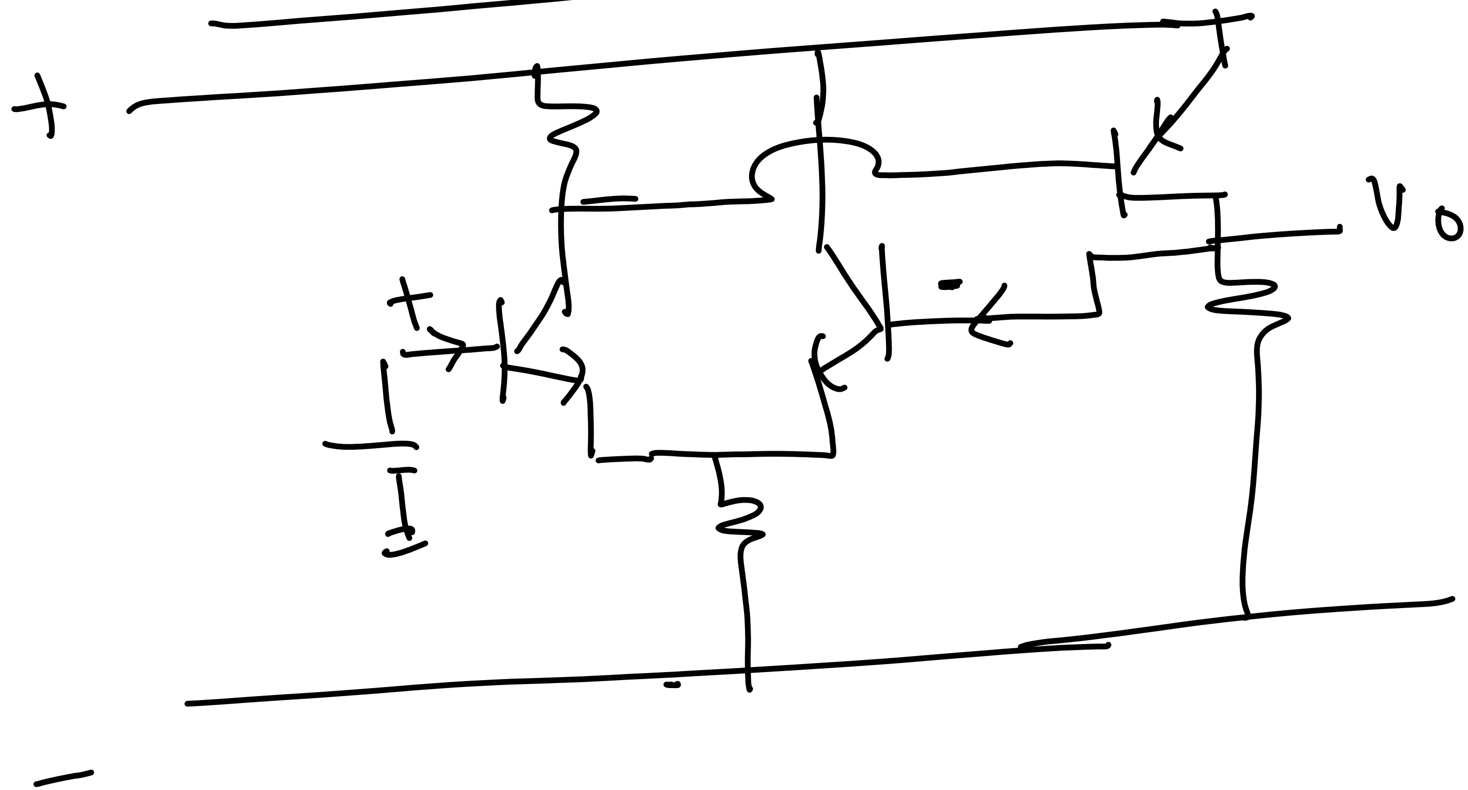


worst case current

$$\text{drift} = \frac{1.5\,mV}{R} = \frac{1.5}{100}$$

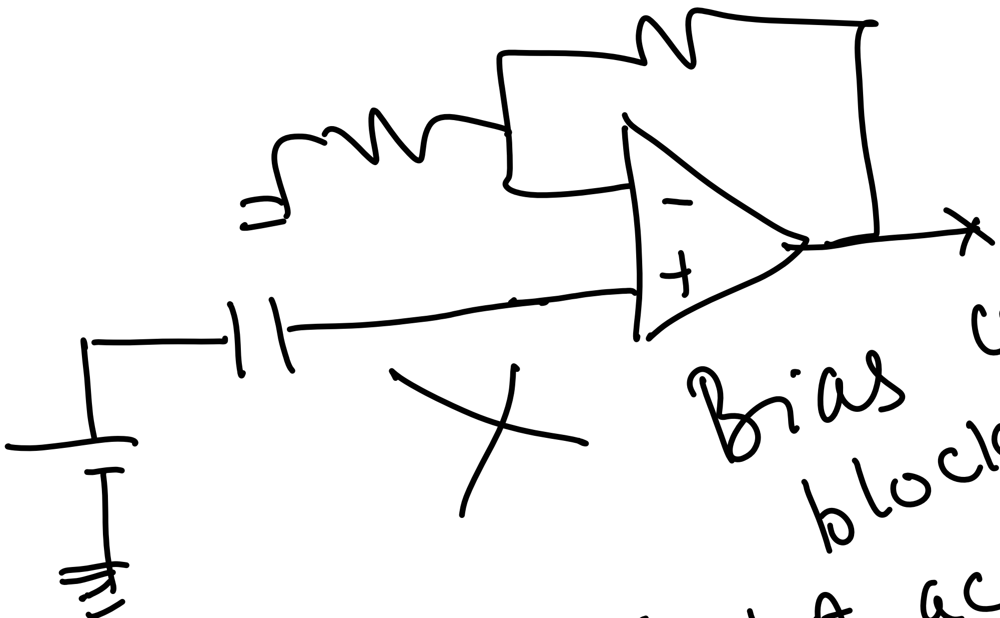
$$= \frac{1500}{100} = 15\,\mu A$$

Bias current error



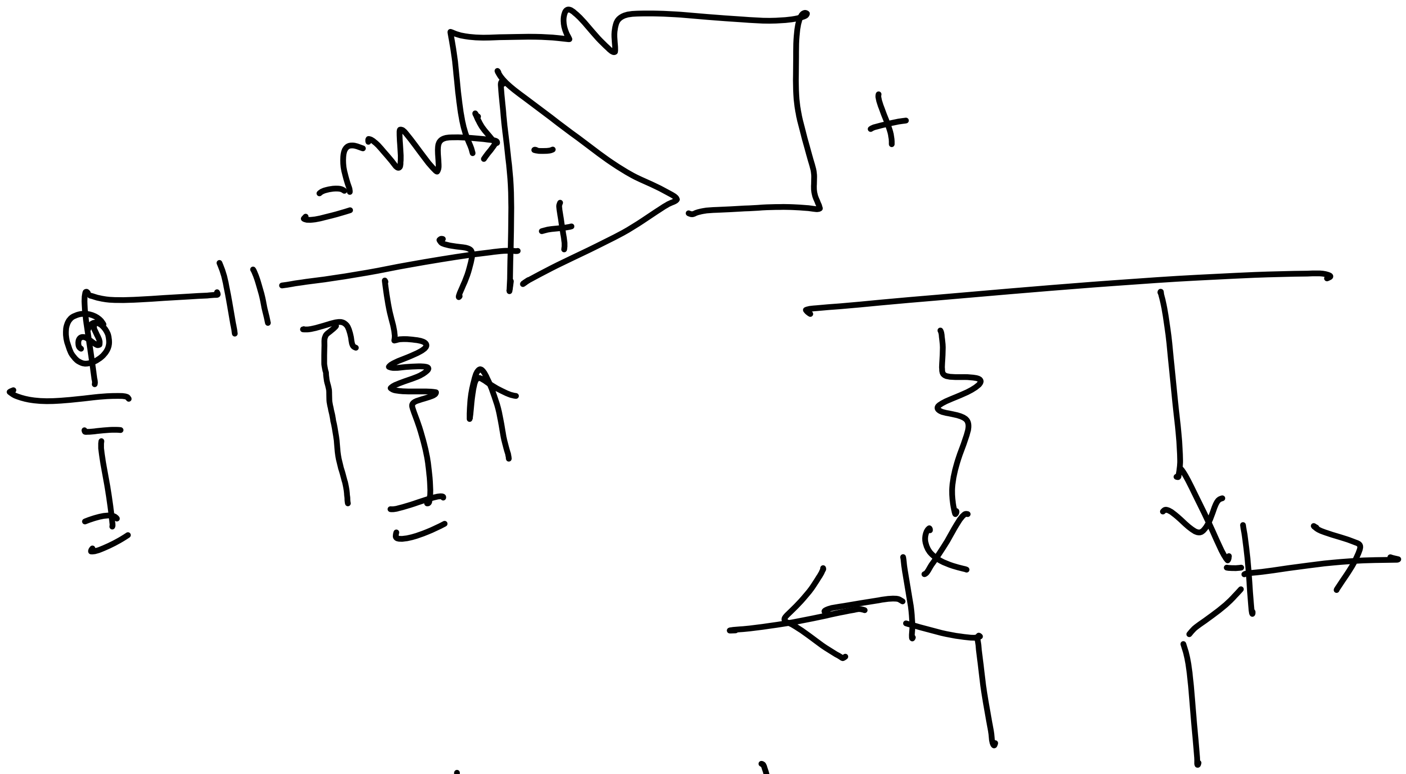
Base current of the input transistors are called bias current

bias current = 50 nA



Bias current is blocked !!

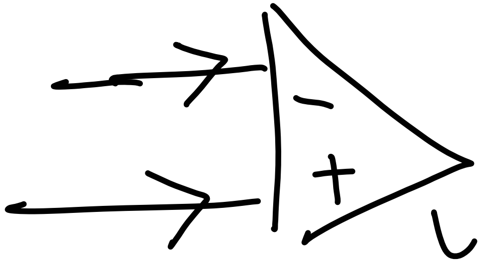
It is nA acceptable



In NPN transistor bias current input of amps flows into the input (ex 741)

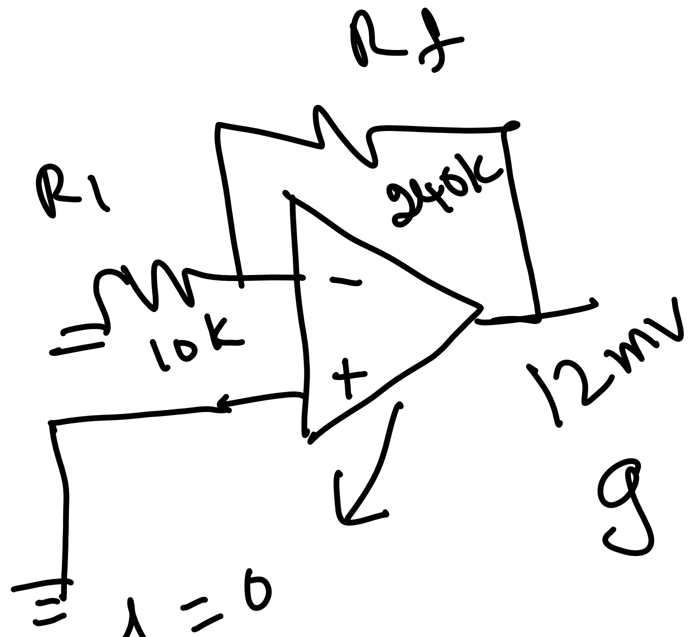
In PNP transistor input of amps

bias current flows out
from the op amp
(LM324)



LM741

LM324



Offset vol = 0

$V_{in} = 0$

$V_0 = 0$

$g = 25$

$R_f = 24k$

$R_1 = 1k$

The expected $V_0 = 0$

but in real life $V_0 \neq 0$

For $I_{B-} = 50\ nA$

Then current through $R_f = 50\ nA$

So vol acc $R_f = 50 \times 10^{-9} \times 240 \times 10^{3}$

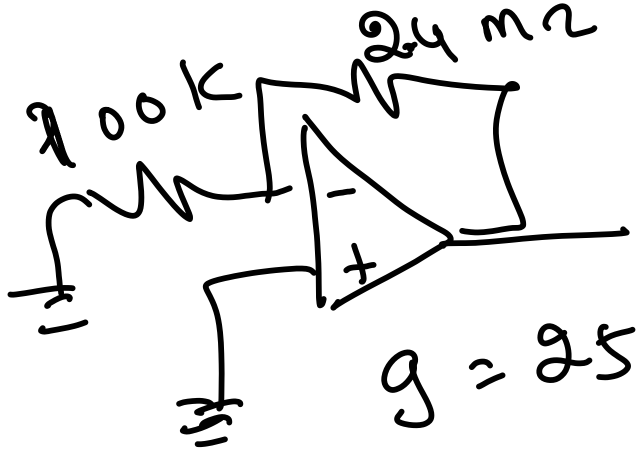
$= 50 \times 240 \times 10^{-6}\ V$

$= 5 \times 24 \times 10^{-4}\ V$

$= 120 \times 10^{-4}\ V$

$= 12\ mV$

The error due to bias current is $12\ mV$



$g = 25$

The bias current error

$= 2.4 \times 10^{6} \times 50 \times 10^{-9}$

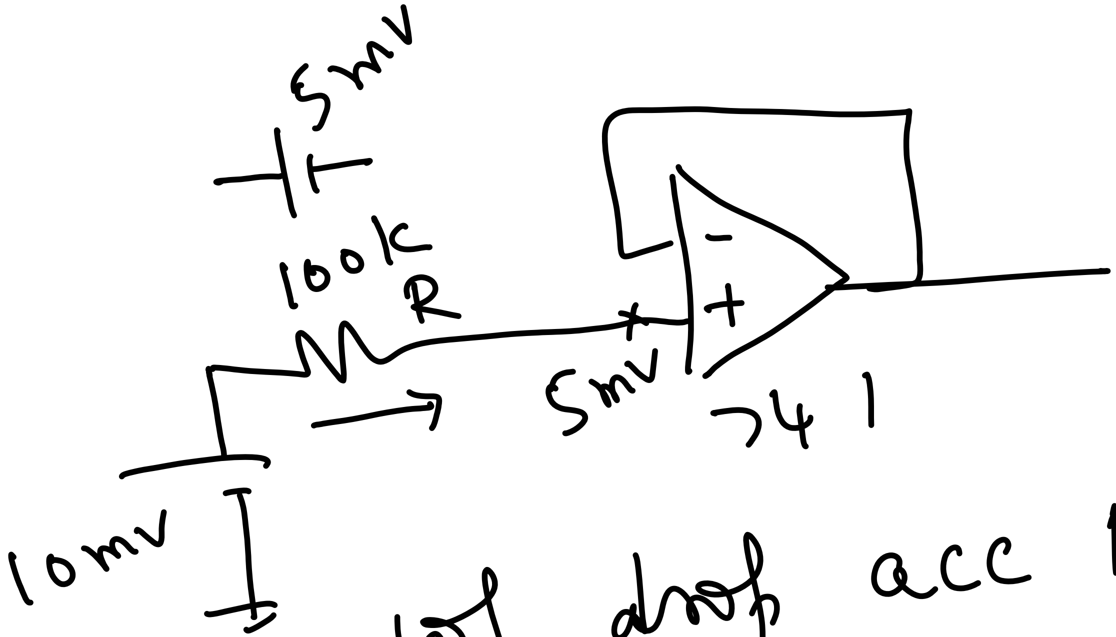
$= 24 \times 5 \times 10^{-3}$

$= 120$ mV

For $\left.\begin{array}{l} R_f = 24k \\ R_1 = 1k \end{array}\right\} g = 25$

bias current error $= 24 \times 10^{3} \times 50 \times 10^{-9}$

$= 1.2$ mV



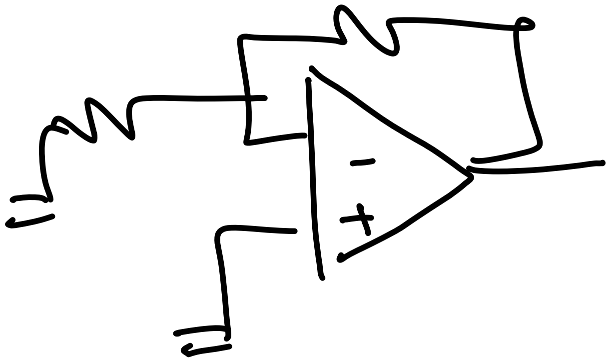
Vol drop acc R

$$= 10^{5} \times 50 \times 10^{-9}$$

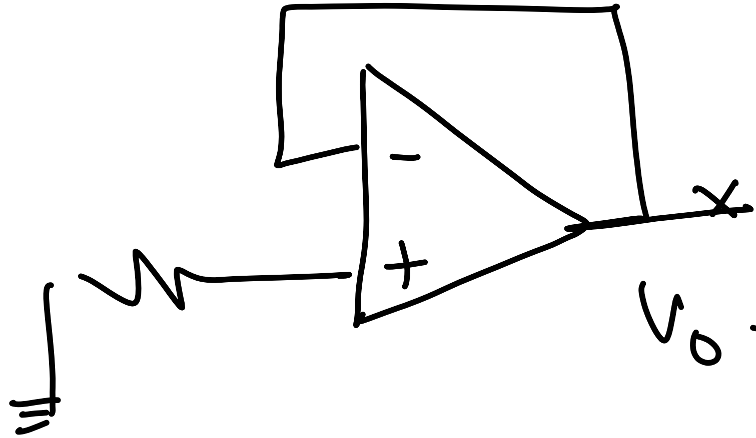
$$= 50 \times 10^{-4}$$

$$= 5\ mV$$

The op amp actually amplifies only 5 mV and not 10 mV

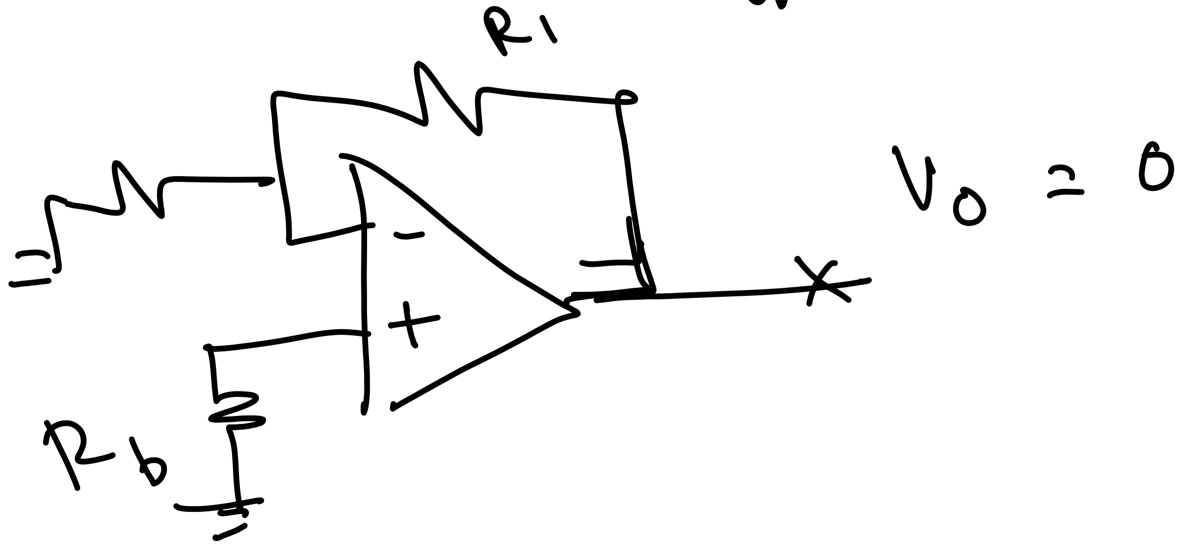


$V_o = +ve$



$V_o = -ve$

How to combine bias current error at non-inverting terminal with bias current error in inverting terminal



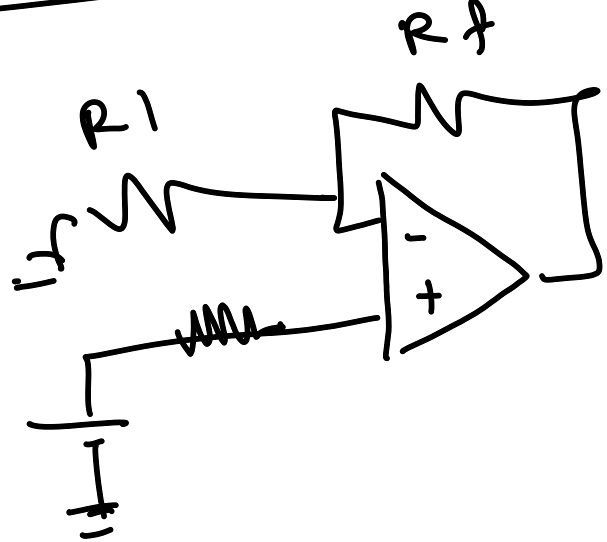
by assuming $I_{b+} = I_{b-}$

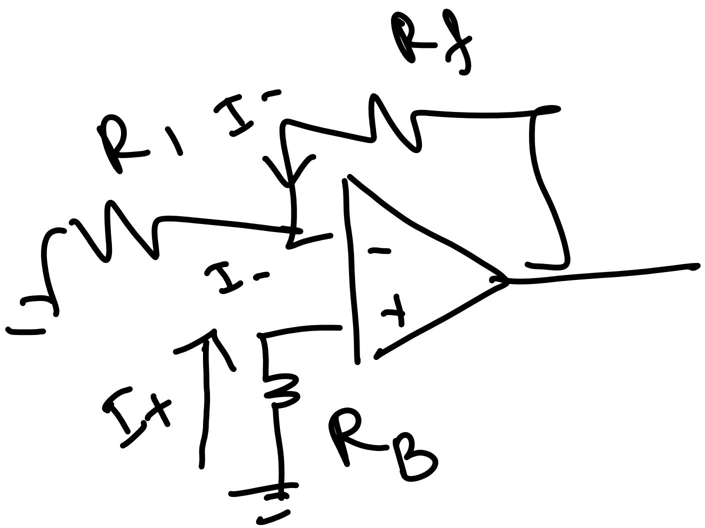
we can make error due to bias current to zero by selecting the resistor values

— x — x — x —

No: 26

Bias current error





$I_- = I_+$

By adding R_B, -ve vol is introduced at the non-inverting input

$$R_f I_- = R_B I_+ (1+g)$$

$$= R_B I_+ \left(1 + \frac{R_f}{R_1}\right)$$

$$R_f = R_B\left(1 + \frac{R_f}{R_1}\right)$$

$$R_B = \frac{R_f}{1 + \frac{R_f}{R_1}} = \frac{R_f R_1}{R_1 + R_f}$$

parallel combination of R_f and R_1

The assumption is

$$I_- = I_+$$

If I_- is not equal to I_+

In fact this is the real

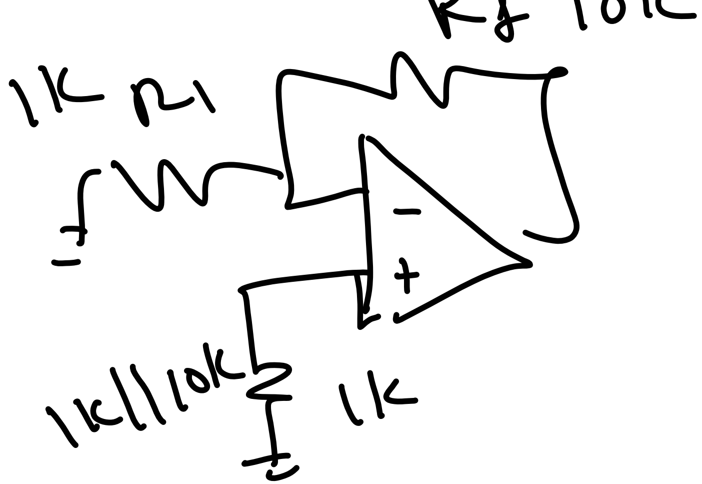
life situation.

So how to reduce the error?

Assume

$$(I_+ - I_-) = I_{offset} \text{ current}$$

If offset current is known then what is the error?



Then V_o error

$$= (I_{offset} \times R_b) g$$

The error voltage at the output

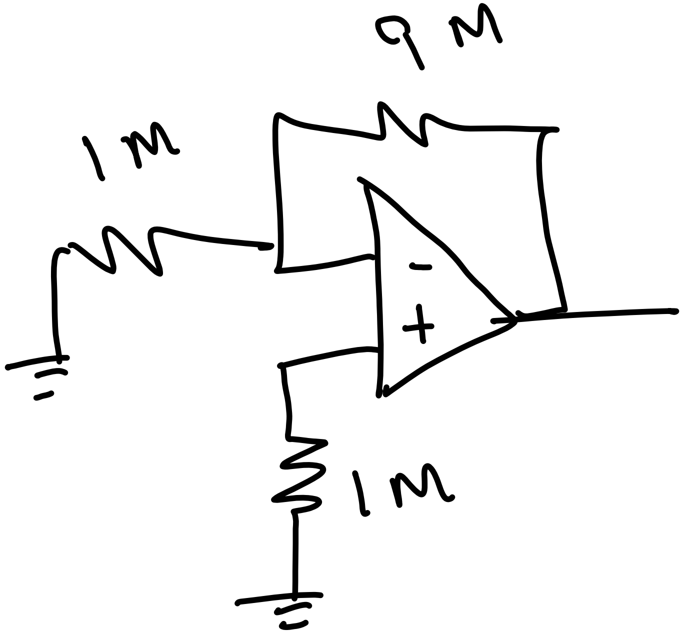
$I_{offset\ current} \times R_B \times$ gain

For 741 op amp

$I_{offset\ current} = 10\,nA$

$$V_o \text{ error} = 10 \times 10^{-9} \times 10^{3} \times 10$$

$$= 10^{-4}\ V$$



$$V_o = 1M \times 10^{-8} \times 10$$
$$= 10^{6} \times 10^{-8} \times 10$$
$$= 0.1V$$

The error voltage due to bias current increases with the resistance that is used to set the gain

The offset current is not constant. It is changing with temperature.

This is called offset current drift.

offset current drift is a major problem and not the offset current by itself

(1) offset voltage drift is a major problem and it doesn't depend up on gain setting resistors

② The offset current drift is also a serious problem and it depends on the gain setting resistors.

What is that we can do to reduce the offset current drift? (other than reducing the resistor values)

① Go for low bias current or low offset current op-amps. Like FET input op amps.

For 741

OFF set current is 10 nA

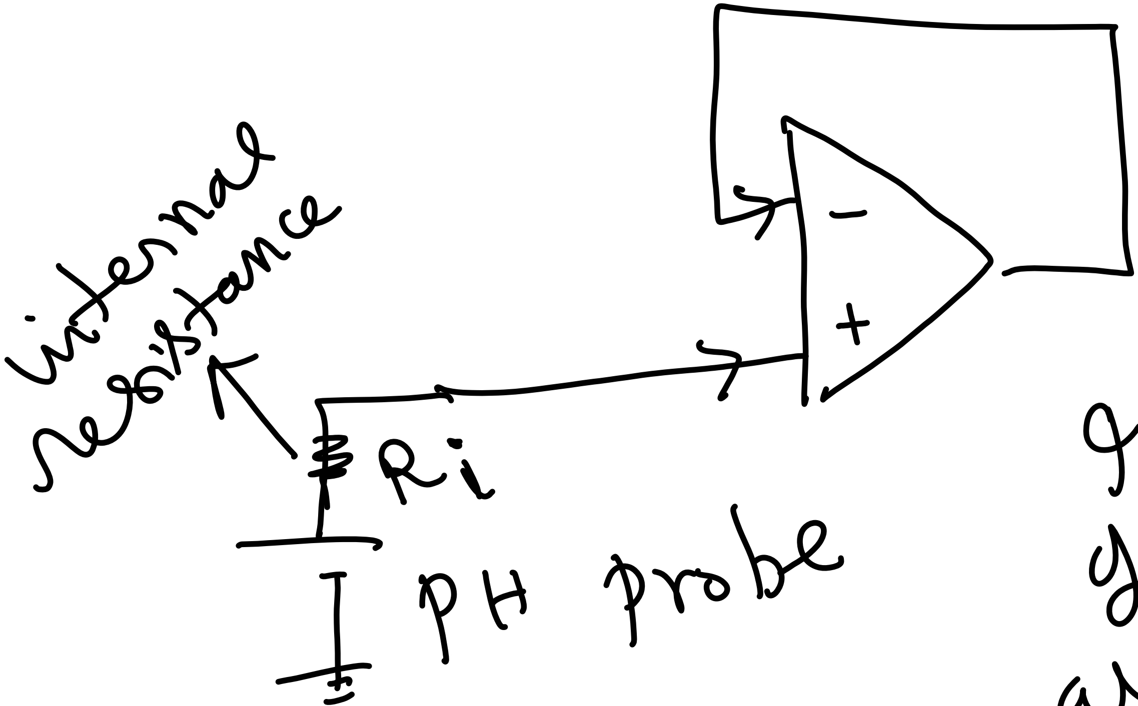
which is Fet input op amp

[3741 the offset current is 10 pA

Design of PH meter amplifier

① Select Fet input op amp

② use the op amp in voltage follower mode



Internal resistance of the PH probe is about 100 MΩ

The bias current I_+ must flow through R_i

Vol drop across R_i = $R_i I_+$

Assume that
PH probe vol = 50 mV

Then vol acc R_i

$= 100\,M\Omega \times I_+$

$= 10^8 \times 10 \times 10^{-12}$

$= 10^{-3}$ V

~~Here~~ $I_+ = 10$ PA

The actual voltage that is appearing at the input

$= 50$ mV $- 1$ mV $= 49$ mV

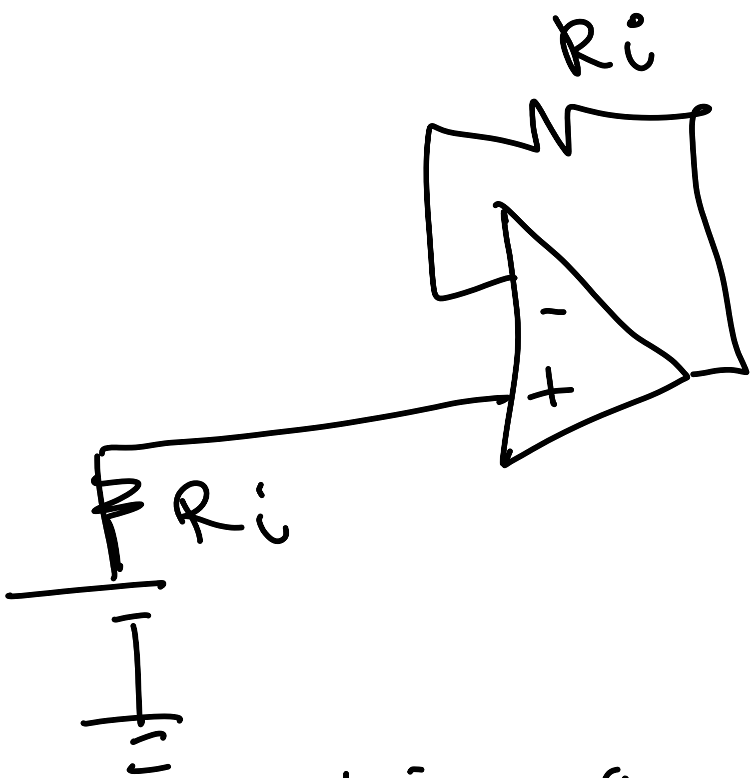
If the bias current of the op amp is 1 nA

Then the voltage drop on R_i

$= 10^8 \times 10^{-9} = 100 mV$

Then net voltage at the non-inverting terminal

$= 50 mV - 100 mV = -50 mV$

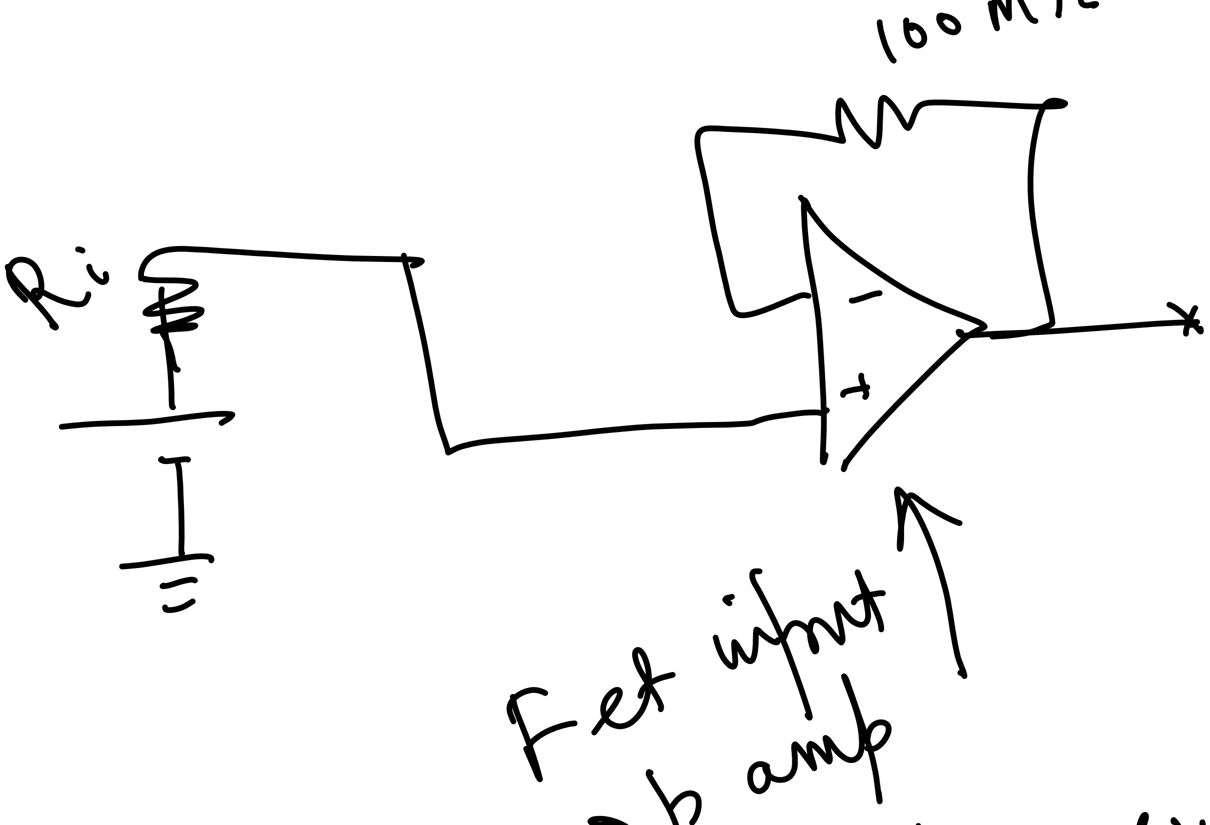


① by adding a resistance equal to R_i in the feedback path one can reduce the bias current error

(2) Even if we manage to connect R_i exactly in the feed back path due to offset current there will be an error voltage at the output

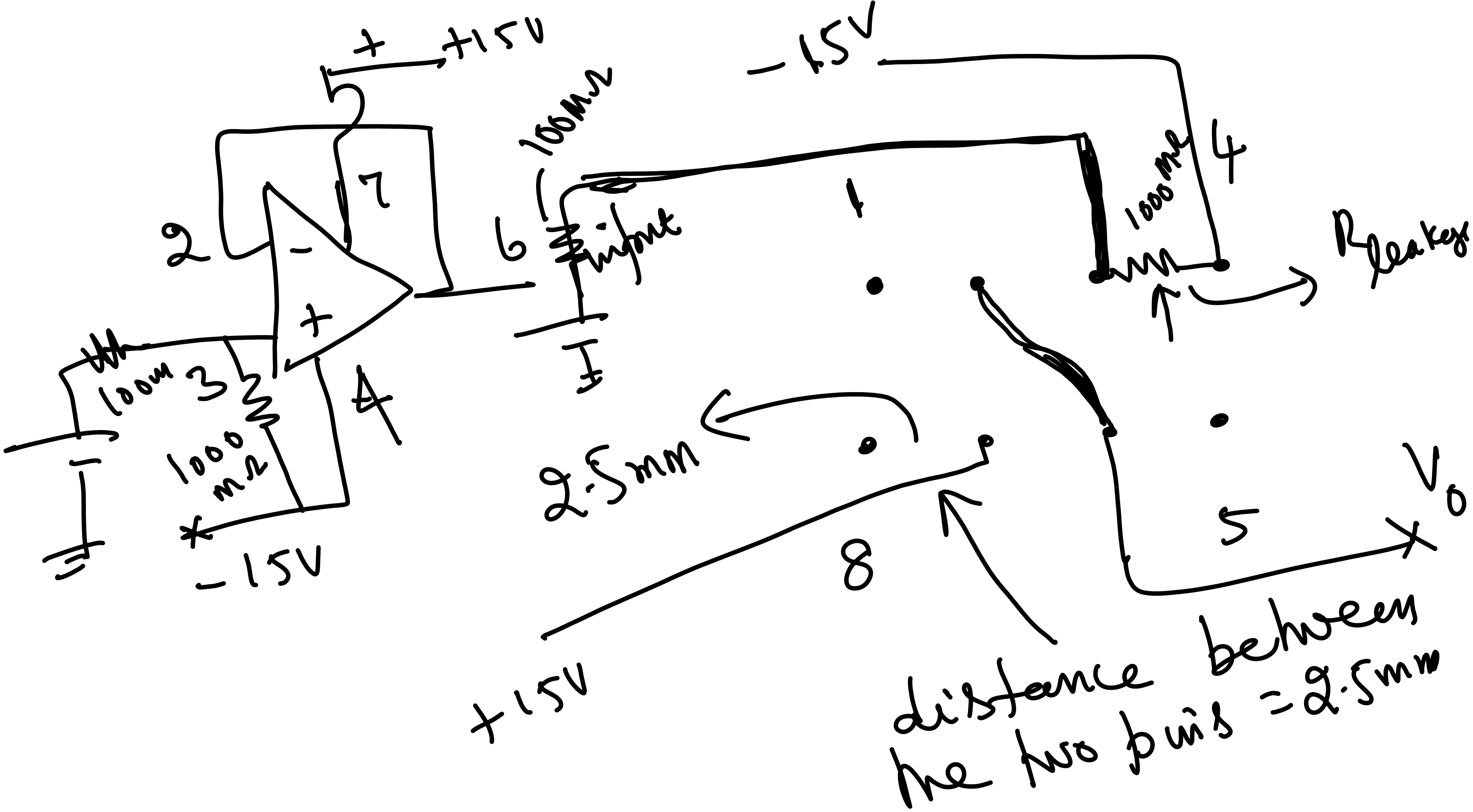
The error voltage = $R_i\, I_{\text{offset current}}$

Since offset current is normally less by a factor of 10, adding resistance in the feed back path reduces the error



Issues in the low bias current op amp

① The leakage current in the PCB will introduce large error



Assume that

$V_{in} = 0$

$R_{in} = 100\ M\Omega$

$R_{leakage} = 1000\ M\Omega$

Because of the R_{leakge}

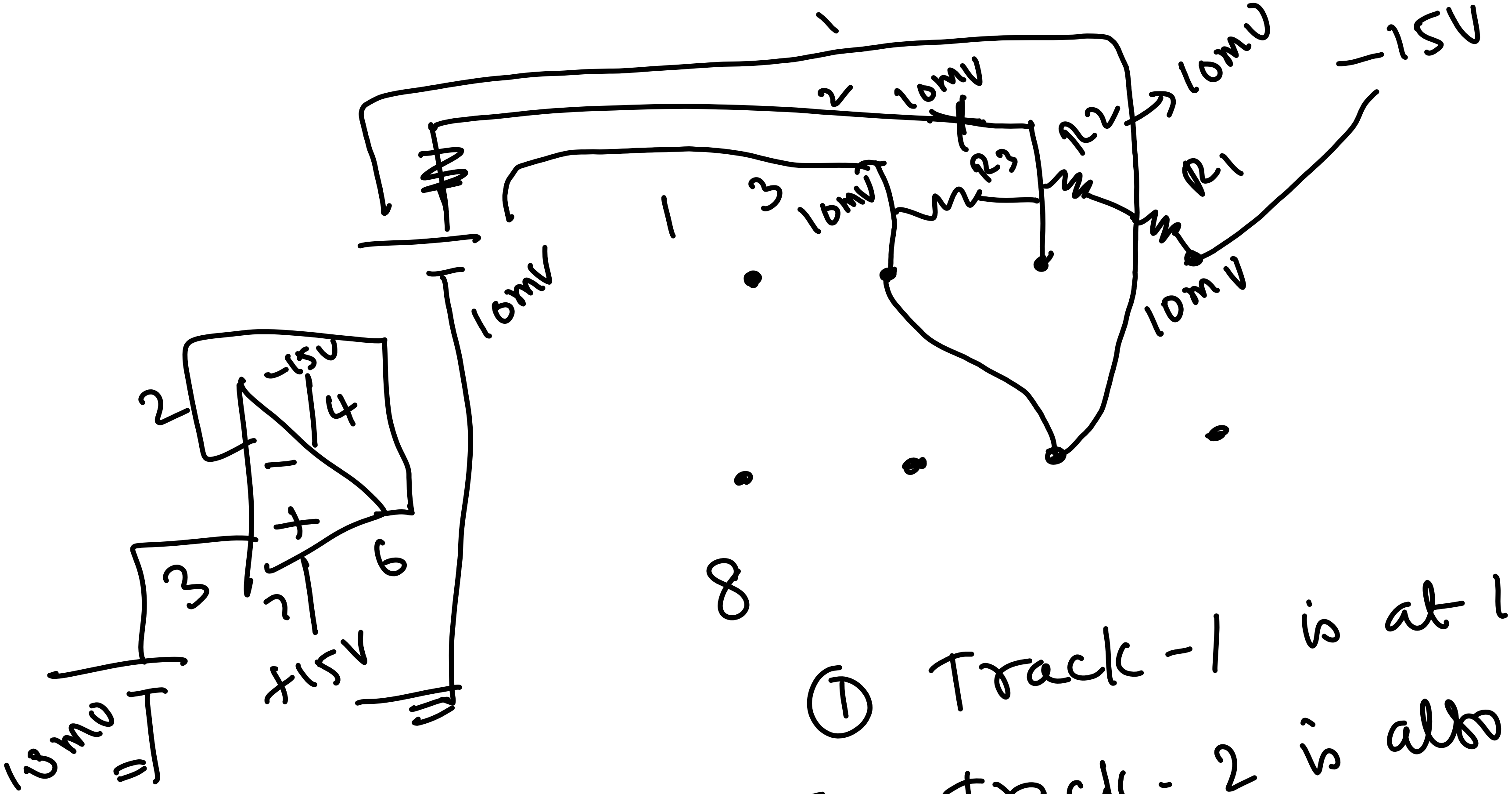
vol at the non-inverting input

$$= -15V \times \frac{100}{1100}$$

$$= \frac{15}{11} = \frac{150}{11} = \frac{1.4}{} V$$

How to reduce this error?

We use Guarding technique to reduce the error due to leakage resistance.



① Track-1 is at 10mV

② Track-2 is also 10mV

③ Track-3 is also 10mV

So current through R_2 and R_3 are zero

However the leakage resistance R_1 sees one side $-15V$ and at the other side it sees $-10\,mV$

So there will be current through R_1

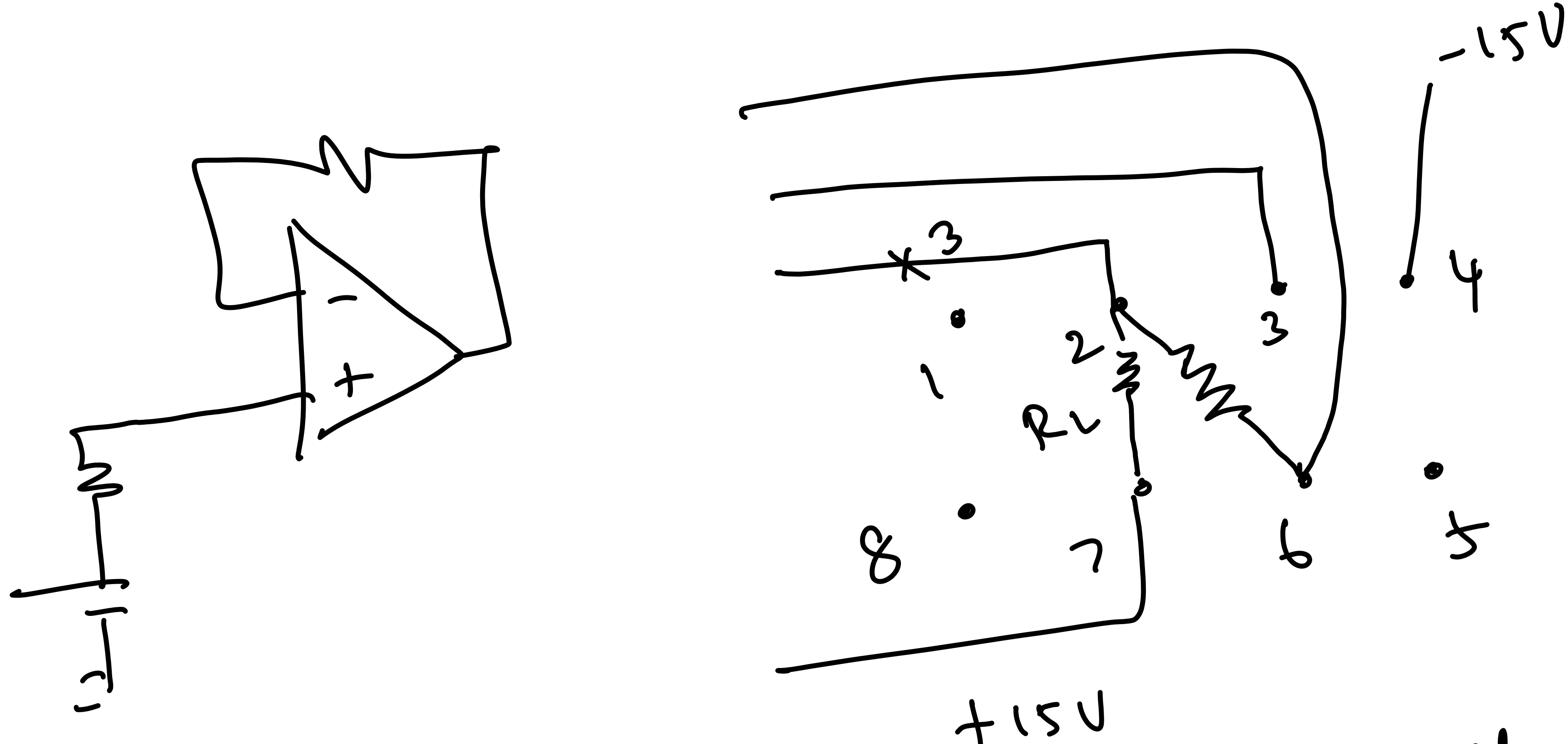
This current R_1 flows through the output terminal of the op amp. However this will not introduce any error at the output

Why?

Because the output resistance of the op amp is nearly zero.

So current flowing through the output will not introduce any error

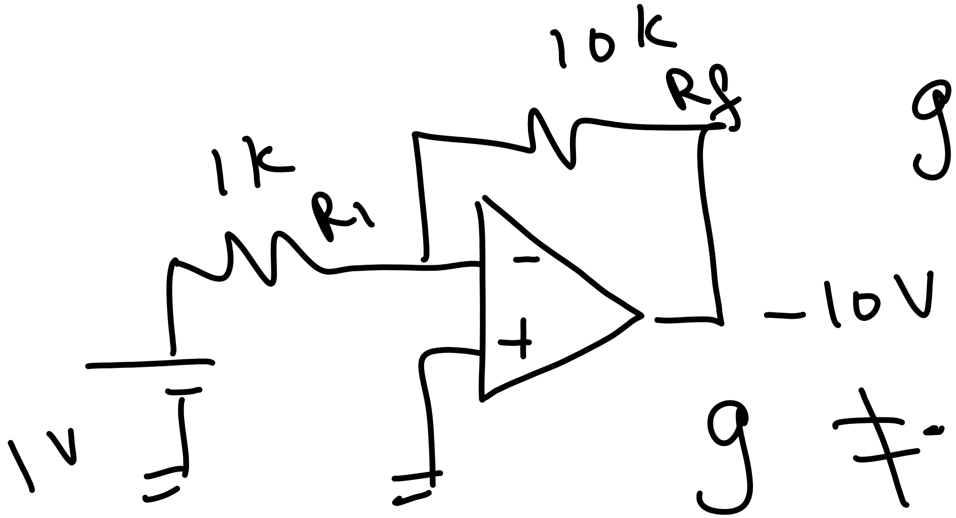
What happens if we introduce compensating resistance in the feed back path?



R_L will introduce a error at pin 2. The Guarding point 3 is of no use.

Lec No: 27

GAIN error in op amp



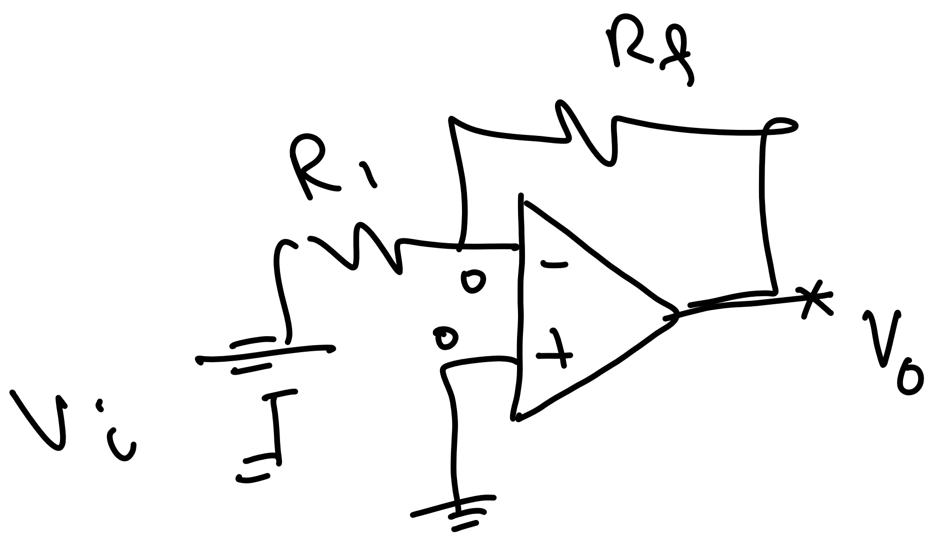
$g = -10$

$g \neq -10$

Actual gain = ?

Set gain $= -\frac{R_f}{R_1}$

Set gain ≠ actual gain



– = inverting terminal

+ = non inverting terminal

vol

In actual case

voltage difference between

non inverting and inverting terminal is amplified by open loop gain

$$V_o = (V_+ - V_-) \text{ open loop gain}$$

non zero

$$(V_+ - V_-) = \frac{V_o}{A}$$

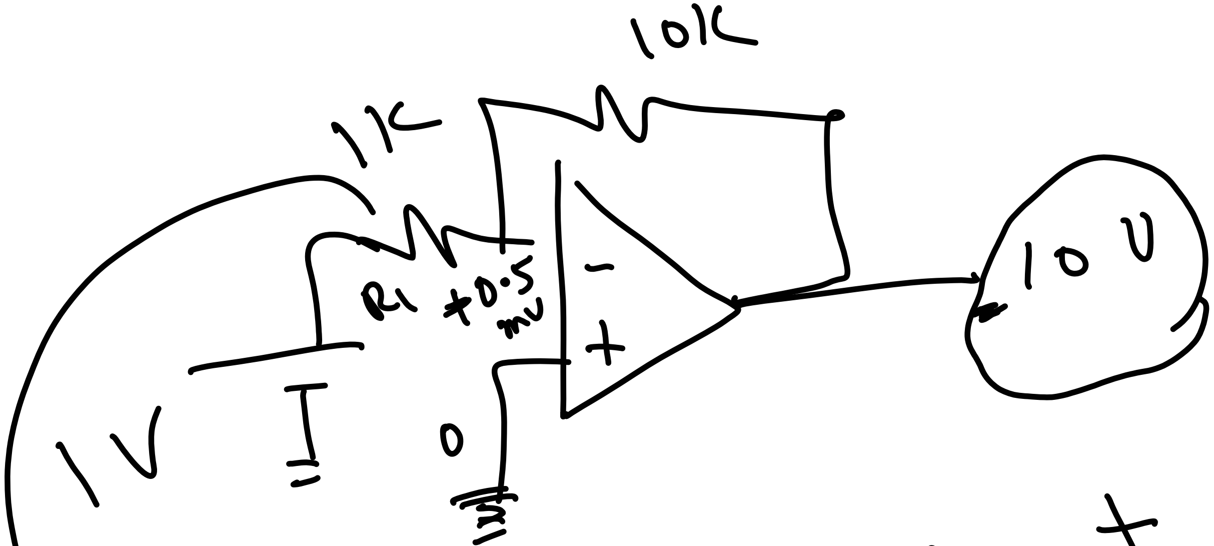
For example

For $V_o = 10V$

$A = 20000$

$$(V_+ - V_-) = \frac{10}{20000}$$

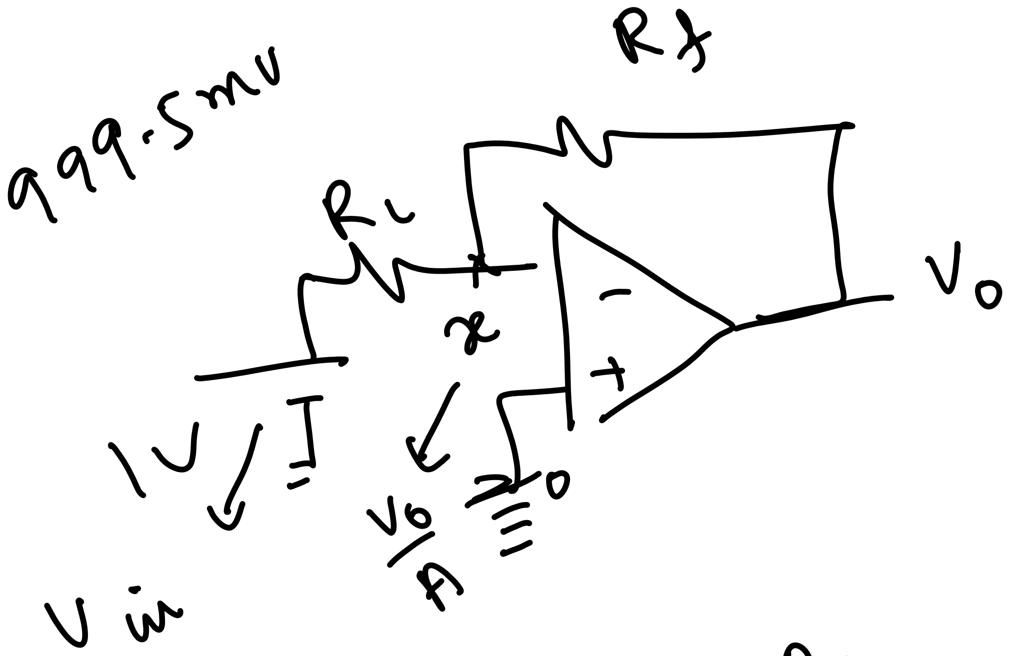
$$= \frac{1}{2000} = \cancel{10}\ 0.5\ mV$$



This will be less than 10

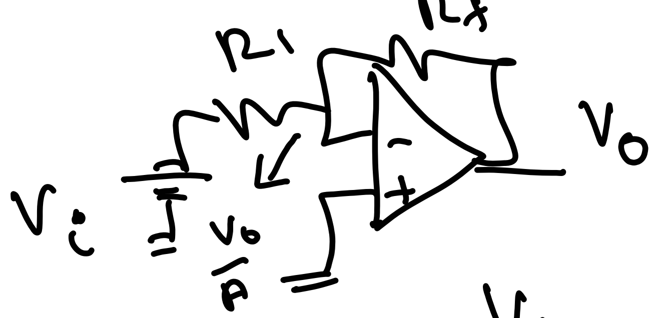
wt acc R1 $\neq V_i$

Actual gain is always less than set gain



Set gain $= \frac{R_f}{R_1}$

actual gain $= \frac{V_0}{V_{in}}$



$$\left(V_i - \frac{V_o}{A}\right)\frac{R_f}{R_1} = V_o \qquad \frac{V_o}{A} = x$$

The input diff $= \dfrac{V_o}{A}$

The effective input vol $= V_i - \dfrac{V_o}{A}$

$$\left(V_i - \frac{V_o}{A}\right) g_s = V_o$$

$$\frac{\left(V_i A - V_o\right)}{A} g_s = V_o$$

$$\frac{V_o}{V_i} = \text{actual gain} = g_a$$

$$\left(V_i - \frac{V_o}{A}\right) g_s = V_o$$

$$(A V_i - V_o)\, g_s = V_o \times A \longrightarrow \text{div by } V_i$$

$$\left(A - \frac{V_o}{V_i}\right) g_s = \frac{V_o}{V_i} \times A$$

$$(A - G_A)\, g_s = G_A \times A$$

$$G_S = \frac{G_A \times A}{A - G_A}$$

$$G_S = \frac{G_A}{1 - \frac{G_A}{A}}$$

$$G_S = \frac{G_A}{\left(1 - \frac{G_A}{A}\right)}$$

What is G_A?

$G_S = 100$

Assume $A = 10000$

$$100 = \frac{G_A}{1 - \frac{G_A}{10000}}$$

$$100\left(1 - \frac{G_A}{10000}\right) = G_A$$

$$100 - \frac{G_A}{100} = G_A$$

$$10000 - G_A = 100\, G_A$$

$$10000 = 101\, G_A$$

$$G_A = \frac{10000}{101}$$

$$\text{Actual} = 99$$

If the set gain and actual gain are to be close then open loop gain must be very large

(1) open loop gain must be large compared to set gain

② The open loop gain A changes with temp

∴ So the gain obtained from the op amp changes with the temperature

$\downarrow$

because open loop gain A changes with temperature

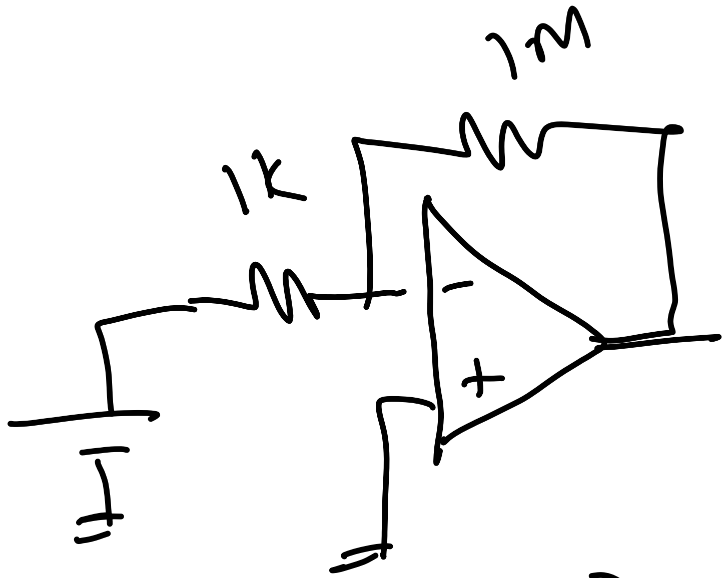
To keep the gain change to a minimum we have to make sure $\frac{G_s}{A}$ is small

Example - 1

Required gain $= 1000 = G_S$

of amp 741

$A = 20,000$



$$1000 = \left(\frac{G_A}{1 - \frac{G_A}{20000}} \right)$$

$$\boxed{G_A = \frac{G_S}{1 + \frac{G_S}{A}}}$$

$$G_A = \frac{1000}{1 + \frac{1000}{20000}}$$ → Set gain

→ open loop gain

$$G_A = \frac{1000}{1 + \frac{1}{20}} = \frac{1000}{1.05}$$

550

$$G_A = 955$$

Is it Stable against temp ??

The open loop gain of the op amp changes with temp

Assume that gain increases from 20000 → 30000

Then the new actual gain is given by

$$G_A = \frac{G_S}{1 + \frac{G_S}{A}}$$

$$= \frac{1000}{1 + \frac{1000}{30000}}$$

$$G_A = \frac{1000}{1+\frac{1}{30}} = \frac{1000}{1.03}$$

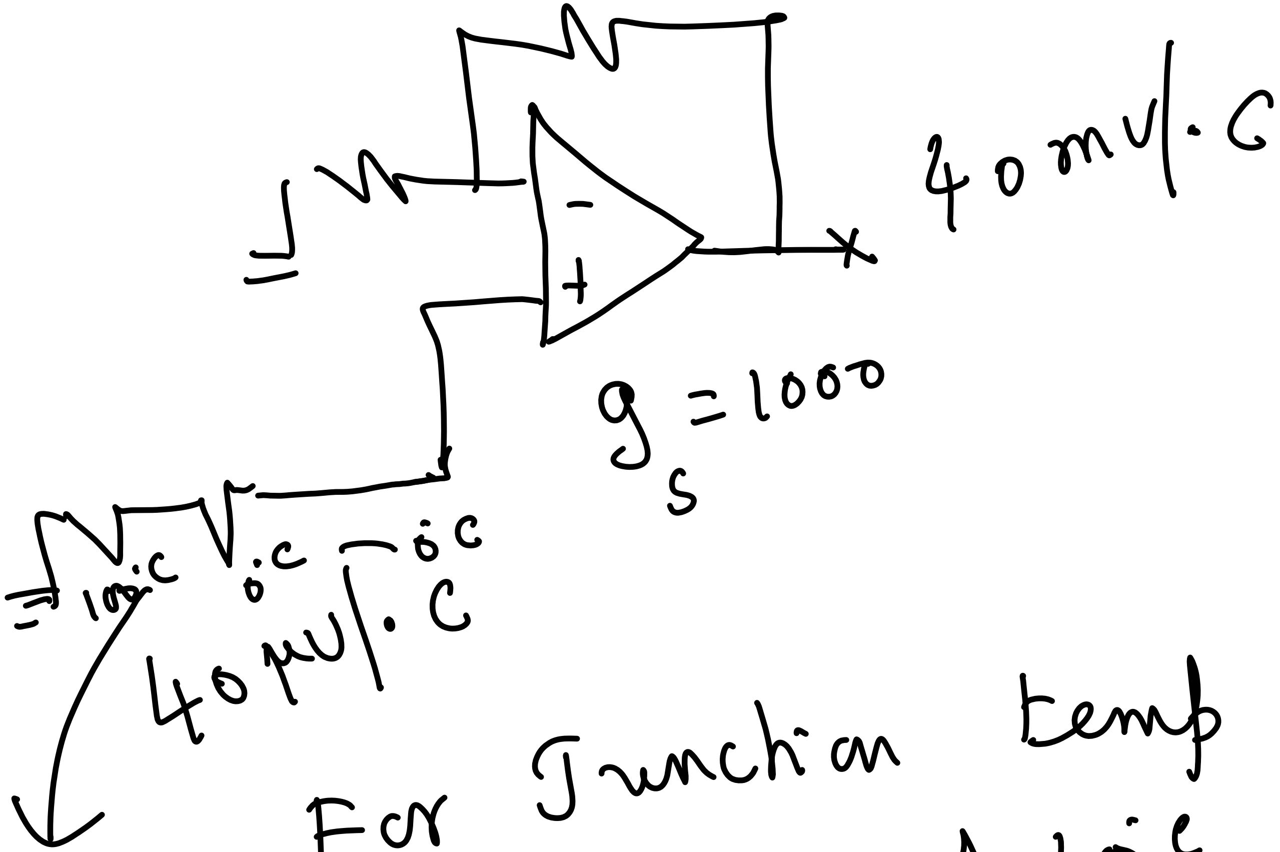
$$G_A \cong 970$$

The gain which is supposed to be 1000 is

at 95°C ambient → 955

85°C ambient → 970

Example 2 Thermo Couple amp



For Junction temp difference of 100°C the expected thermocouple vtg = 40 × 100 = 4 mV

For gain of 1000 (set gain)

the expected output vol

= 4 mV x 1000 = 4 V

In real case you will never get

4 V output even though

thermocouple voltage is correctly at

4 mV

We know at 25°C ambient

the A = 2000

At 80°C ambient

A = 30000

actual gain at
25°C ambient = 955

actual gain at 2
80°C ambient = 970

Vo at 25°C ambient = 4 × 955 mv

Vo at 80°C ambient = 4 × 970 mV

at 25°C = 3.820 V 2

at 80°C = 3.880 V

Difference in output = 60 mV

At the output of the

of amp Sensitivity $= 40 \times 1000\,\mu V$

$= 40\ mV$

The expected emr $= 1.5\,^{\circ}C$

at the output

The abesolute emr at $25\,^{\circ}C$

$= (4 - 3.820)\ V$

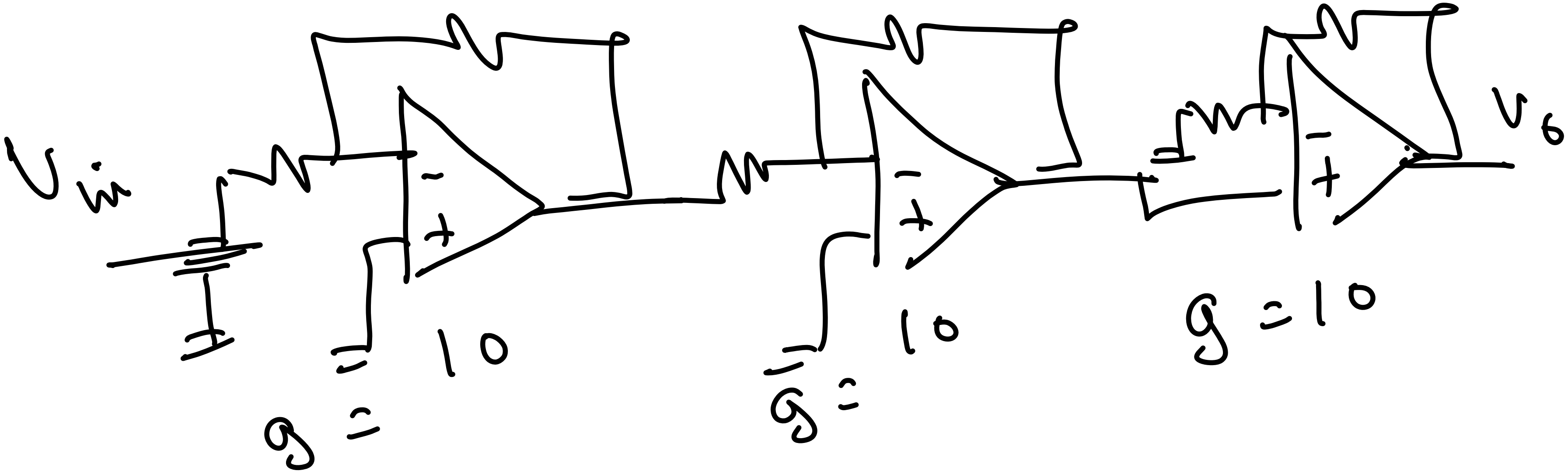
$= 180\ mV$

$= \dfrac{180}{40} = 4.5\,^{\circ}C$

At room temp 100°c temp
will be read as = 96.5°c
At 80°C room temp 100°c temp
will be read as = 98°C

What is to be done ?

Obtain the required gain of
1000 in 3 Stages.



Total gain = 1000

Single stage gain

$$G_A = \frac{10}{1+\frac{10}{20000}} = \frac{10}{1+\frac{1}{2000}}$$

$$= \frac{10}{1+0.0005} = \frac{10}{1.0005} = 9.995$$

For 3 Stages

total gain $= \frac{10}{1.0005} \times \frac{10}{1.0005} \times \frac{10}{1.0005}$

$\simeq 999$ — check this

total gain at 85°C ambient $= \frac{10}{1+\frac{10}{3000}} \times \frac{10}{1+\frac{10}{3000}} \times \frac{10}{1+\frac{10}{3000}}$

$= 999.5$ check this

To get larger gain use many stages

ie get the required gain from multiple stages. This reduces the gain error and also gain drift.

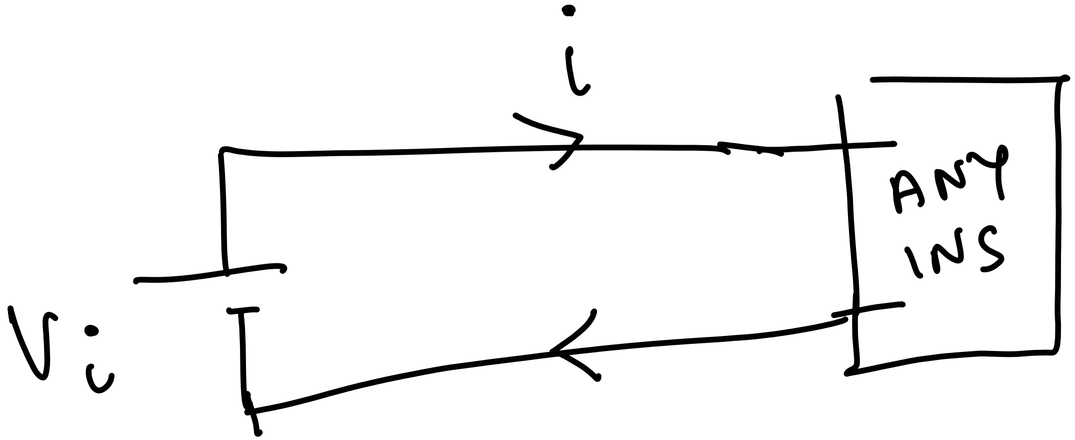
Do not set large gain for one single stage

Lec No: 28

① INPUT RESISTANCE

② OUTput resistance

INPUT RESISTANCE

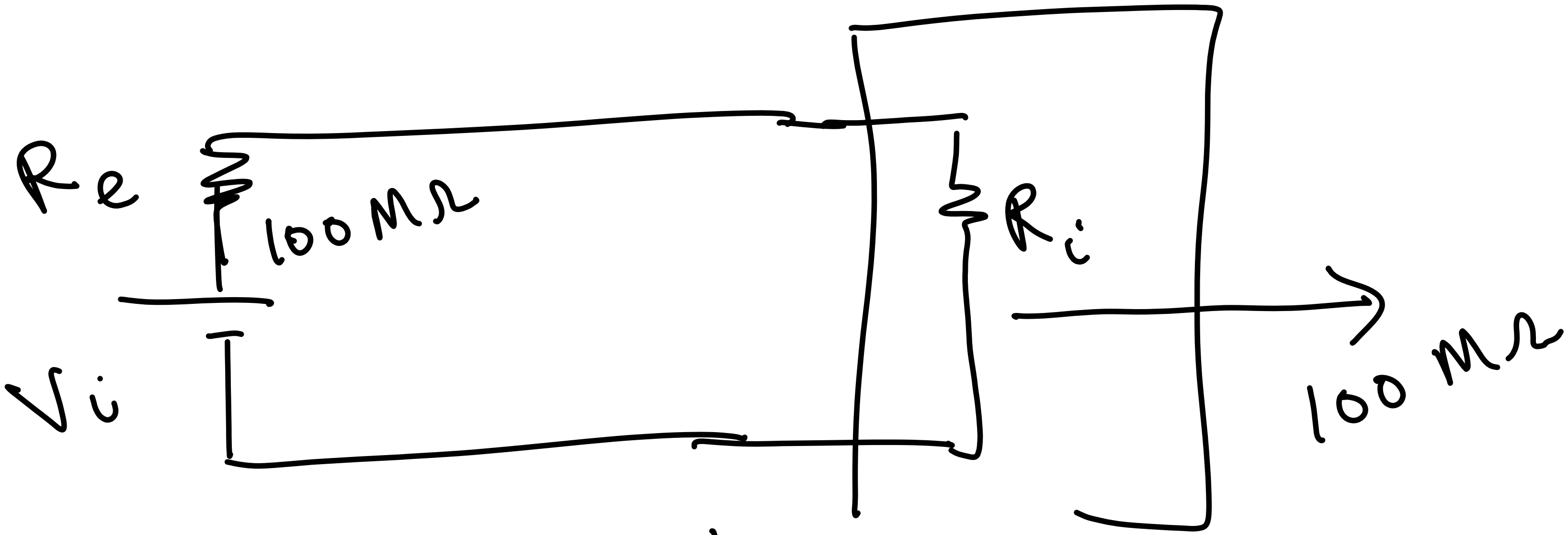


what is the input resistance offered?

The effective input resistance

$$= \frac{V_i}{i} = \frac{1V}{1\mu A} = 1M\Omega$$

As long as source impedance is zero the input resistance offered by the instrument is of no importance



For PH Probe

internal resistance = 100 MΩ

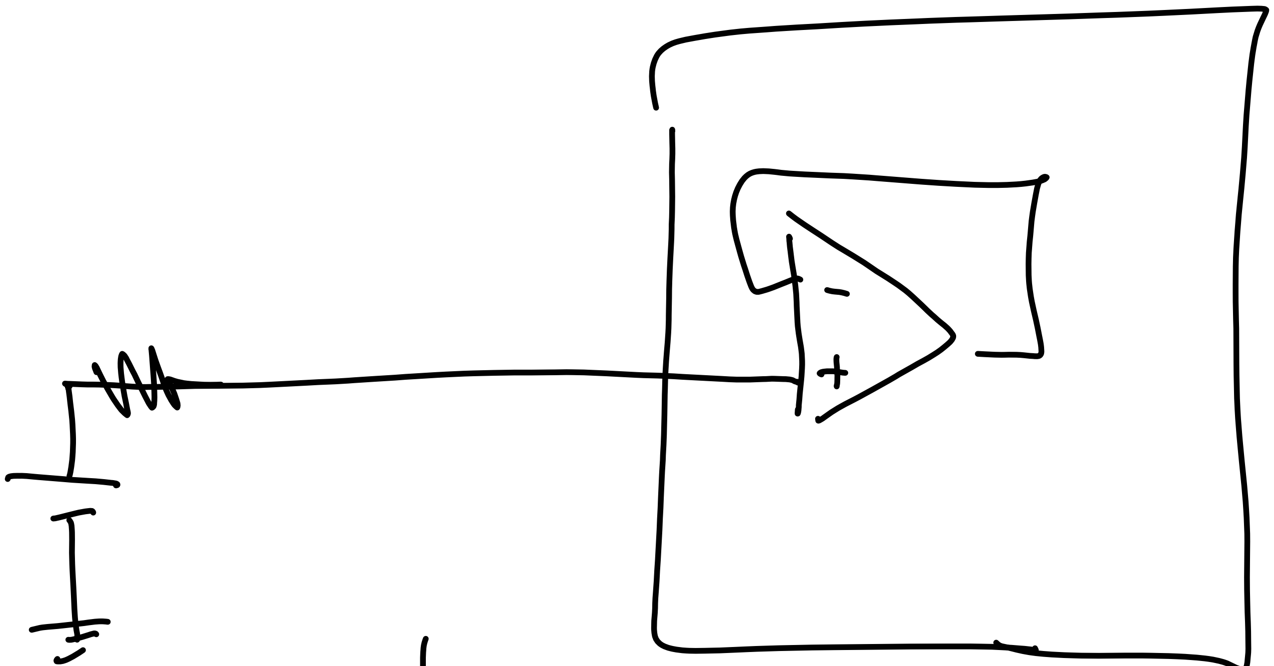
Even if R_i = 100 MΩ

Then half of V_i alone

appear as input at the

instrument

So ideally speaking
i/put resistance must be
infinite for voltage measurement



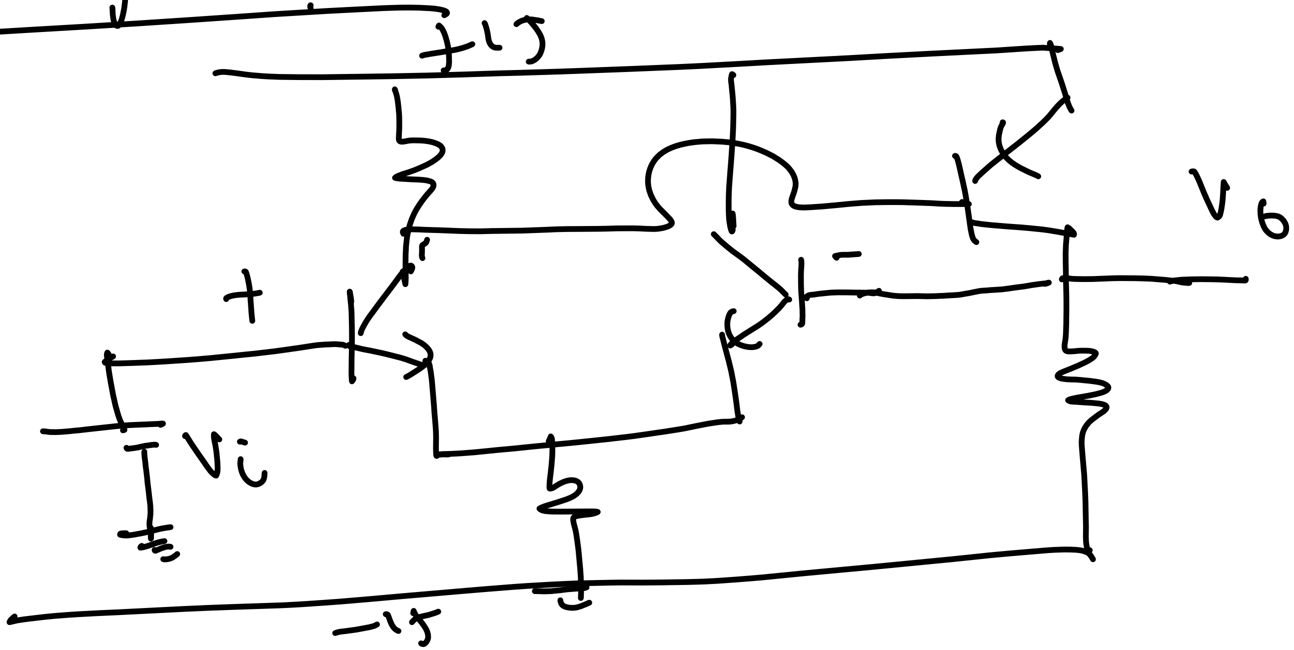
(1) What is the
i/put resistance offered by the
op amp

(2) For ideal op amp it is infinite

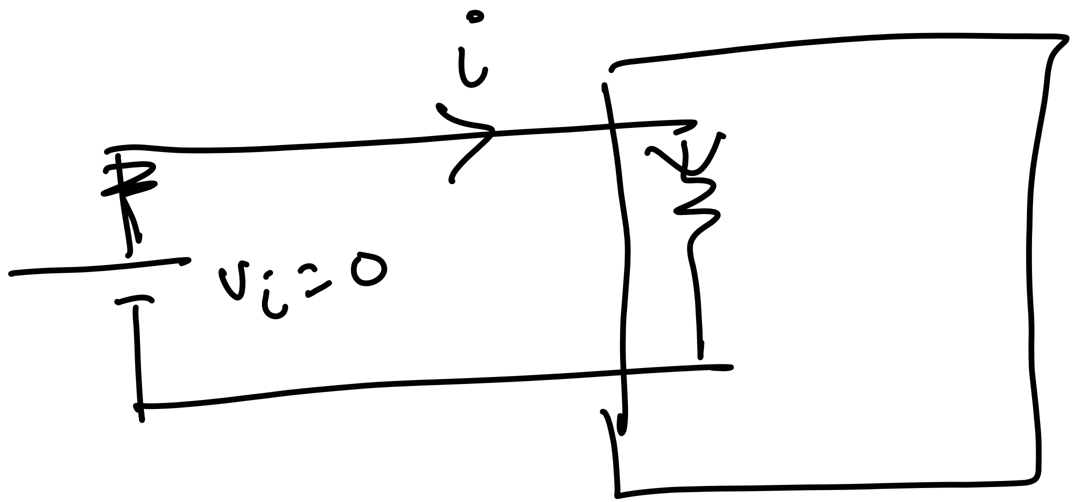
(3) But for real op amp it is not infinite.

It also varies with gain

Real op amp



In case of op amp even for $V_i = 0$ there is a current



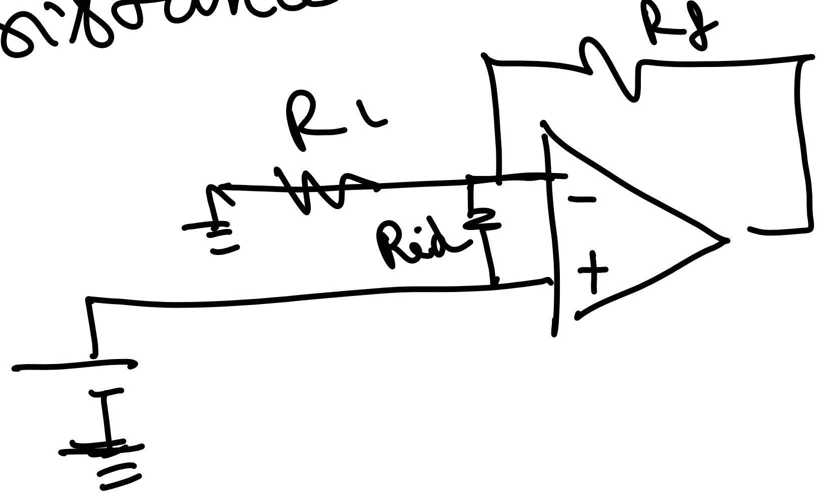
Here even for $V_i = 0$ there is a current to the input. This is taken as bias current.

Bias current is not considered directly for input resistance

Calculation.

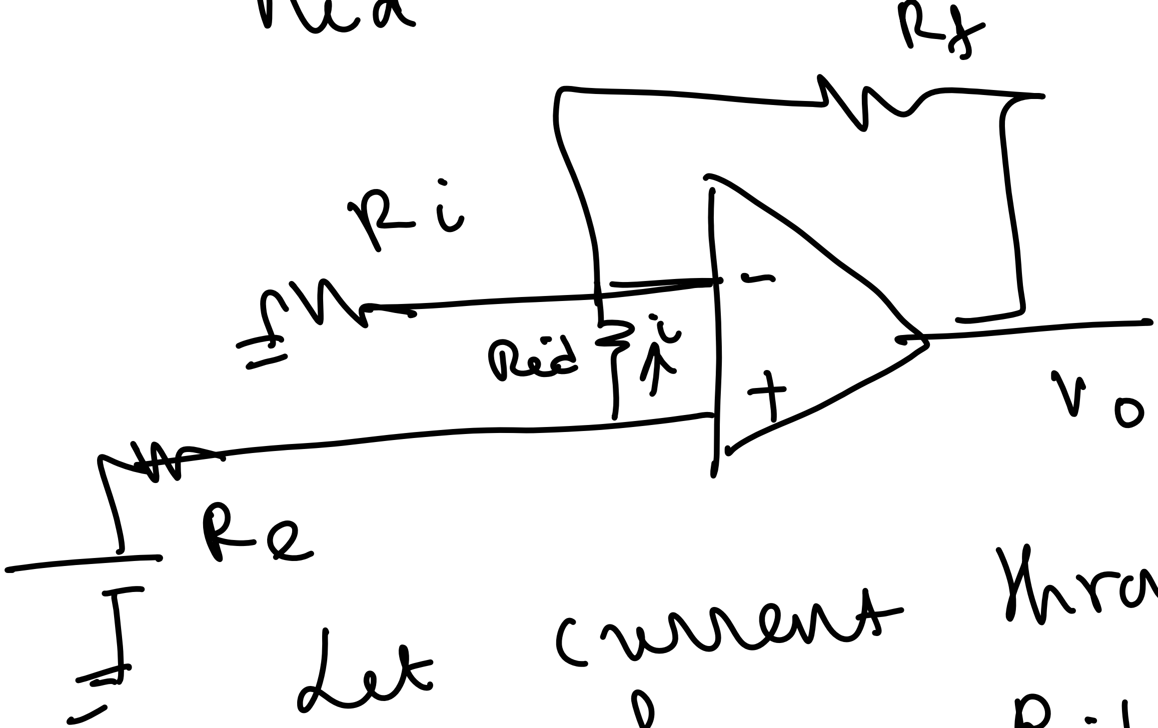
However when the input voltage goes up bias current increases. This increase in current is due to input resistance

How to represent the input resistance ~~ird~~ inthe op amp



Rid is called differential input resistance of the op amp

Rid is not input resistance



Let current through Rid = i

Then vol acc Rid = Rid × i

$(R_{id} \times i)$ = differential vol at the input

$(R_{id} \times i)\, A = V_0$ --- (1)

when A is open loop gain

V_0 = output voltage

$$i = \frac{V_i}{R}$$

where V_i is the input voltage

R is the input resistance

$$\left(R_{id} \times \frac{V_i}{R}\right) A = V_0$$

$$\frac{R_{id} \times A}{R} = \frac{V_o}{V_i}$$

but $\frac{V_o}{V_i}$ = gain

$$\frac{R_{id} \times A}{R} = g$$

g is a closed loop gain

$$g = \frac{R_{id} \times A}{R}$$

$$R = \frac{R_{id} \times A}{g}$$

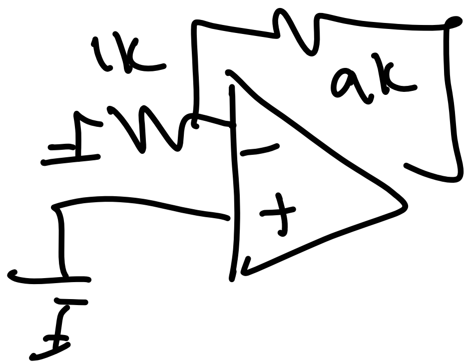
1. The input resistance offered by the op amp varies with gain g. Higher the closed loop gain lower the input resistance

<u>For example</u> For 741 op amp

$$R_{id} = 200k$$

$$A = 20\,000 \quad (\text{min})$$

For $g = 10$



What is the input resistance offered by this circuit?

$$R = \frac{R_{id} \times A}{g} = \frac{200 \times 10^{3} \times 2000\cancel{0}}{1\cancel{0}}$$

$$= 200 \times 10^{3} \times 2 \times 10^{3}$$

$$= 2000 \times 10^{6}$$

$$= 2000 \ M\Omega$$

For $g = 1000$

$$R = \frac{200 \times 10^{3} \times 20\cancel{000}}{1\cancel{000}}$$

$$= 200 \times 10^{3} \times 20$$

$= 4 \times 10^6 = 4\ M\Omega$

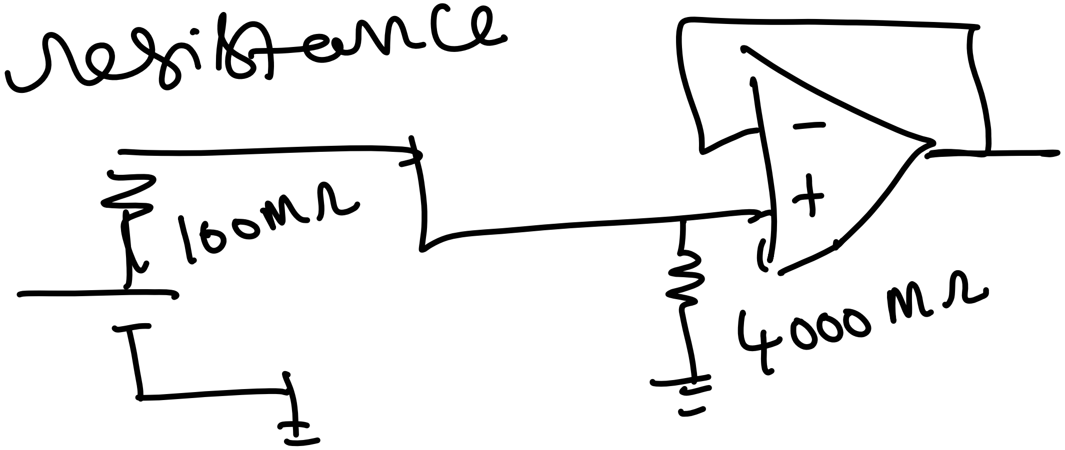
For voltage follower.

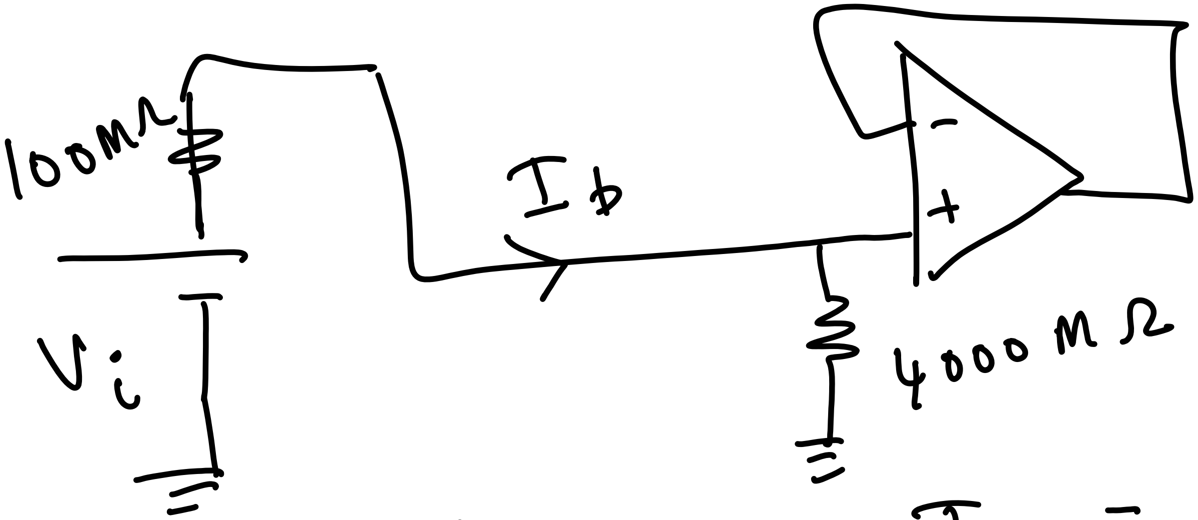
$$R = \frac{R_{id} \times A}{g} = \frac{200 \times 10^3 \times 20000}{1}$$

$= 4 \times 10^9$

$= 4000\ M\Omega$

Voltage follower gives highest input resistance





For example if $I_b = 10\,pA$

For Fet input of amp

vol drop on the internal resistance

$= 10^8 \times 10 \times 10^{-12}$

$= 10^9 \times 10^{-12} = 1\,mV$

voltage loss due to bias current

$= 1\,mV$

Voltage loss due to input resistance of the op amp

$$= V_i \times \frac{100}{4000+100}$$

$$= V_i \times \frac{100}{4100} = V_i \times \frac{1}{41}$$

If $V_i = 100mV$

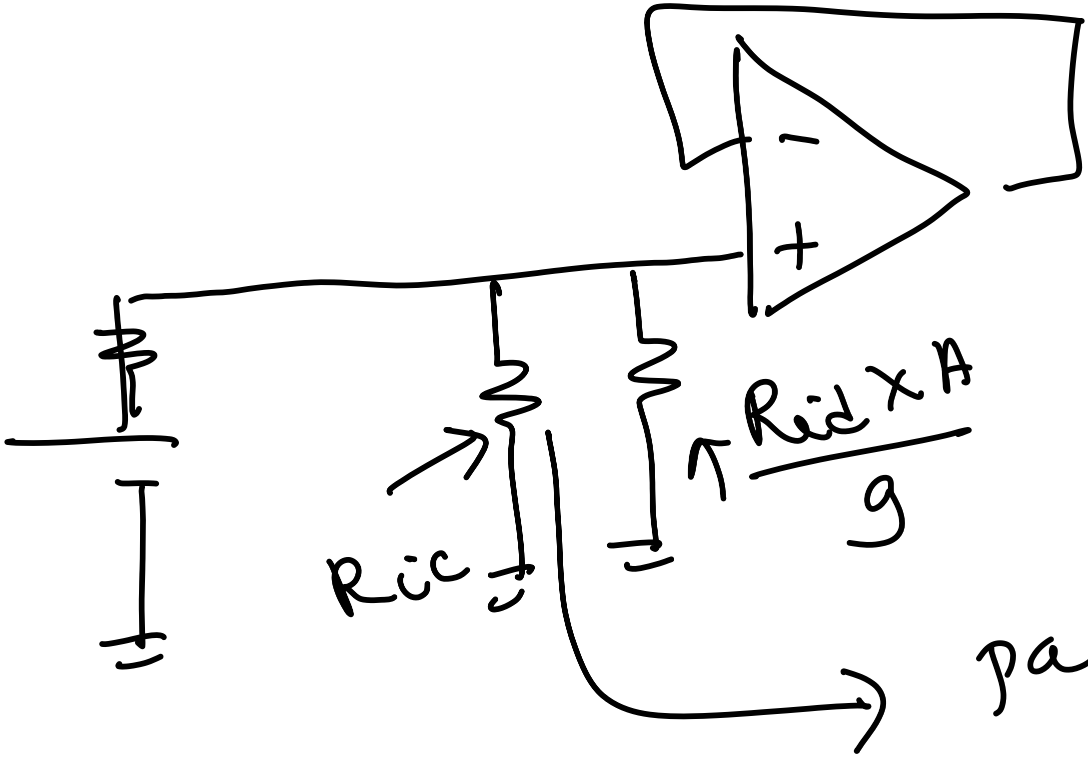
Then voltage loss $= \frac{100}{41} = 2.5mV$

So for 100mV PH probe output 1mV is lost due to bias current and 2.5mV is lost due to input resistance

Total loss in IGBT

= bias current loss + input resistance loss

Input resistance is also affected by package resistance



Package resistance is constant.
It is not changing with gain etc.

The effective input resistance of the op amp $= R_{ic} \| R$

where $R = \dfrac{R_{id} \times A}{g}$

For 741 for $g = 1$

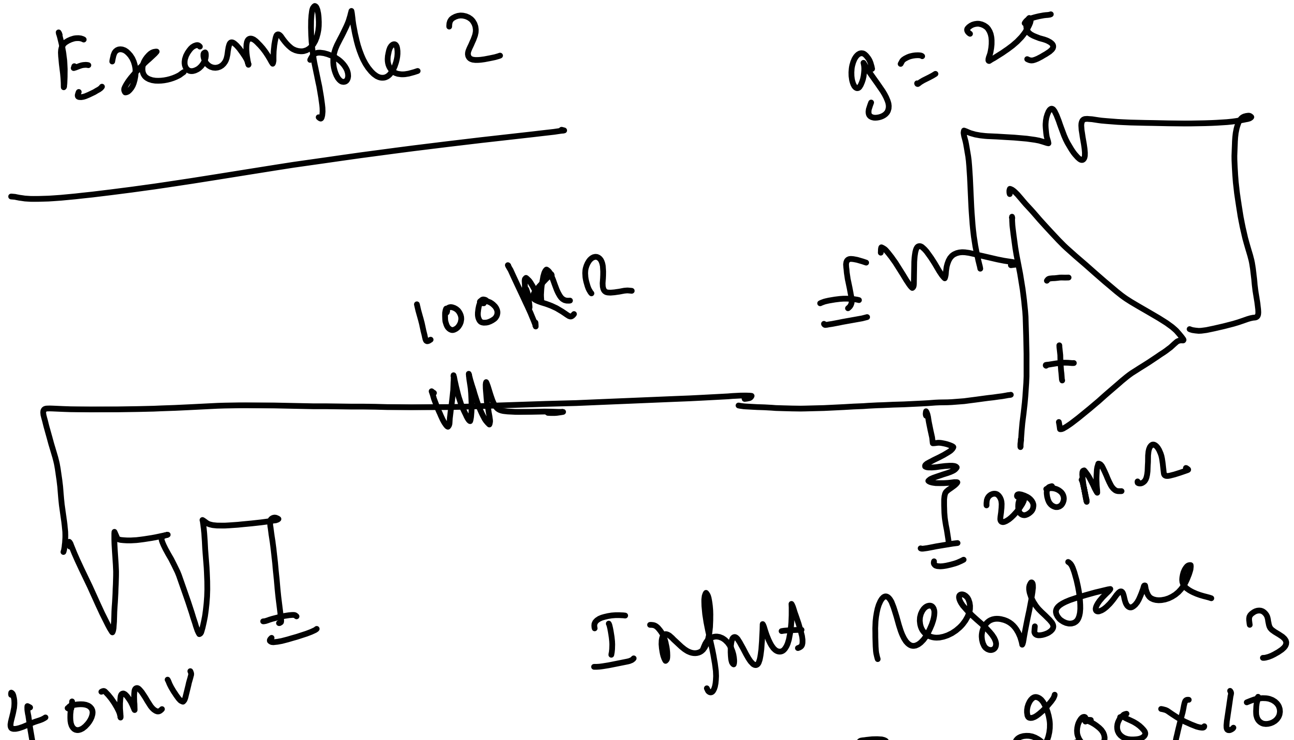
$R = 4000\ M\Omega$

$R_{ic} = 4000\ M\Omega$

So for 741 op amp the highest input resistance $= 4000 \| 4000\ M$

$= 2000\ M\Omega$

Example 2



Input resistance

$$= \frac{200 \times 10^{3} \times 25000}{25}$$

$$= 200 \text{ M}\Omega$$

The internal resistance $= 100 \text{ k}\Omega$

Voltage loss due to input resistance $= \dfrac{40 \times 10^{-3} \times 100 \times 10^{3}}{200 \times 10^{6}}$

$$= \frac{40 \times 100}{200 \times 10^{6}} = \frac{40}{2} \times 10^{-6}$$

$$= 20\,\mu V$$

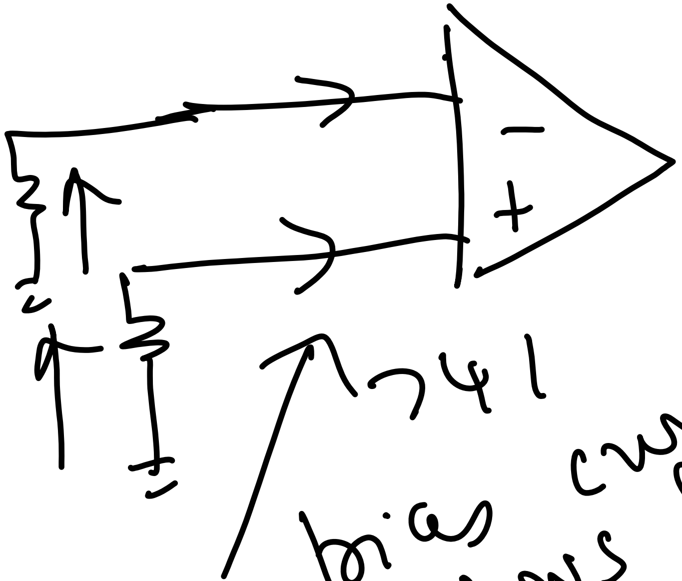
loss due to bias current $= R_{wif} \times I_b$ 10nA

$$= 100 \times 10^{3} \times 10^{-8}$$

$$= 100 \times 10^{-5} = 1\,mV$$

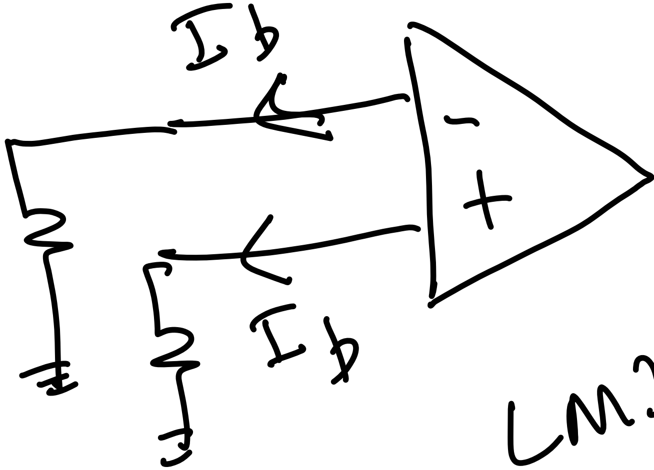
In this loss due to bias current is more than wiport resistance loss

Bias current direction depends on the op amp !!



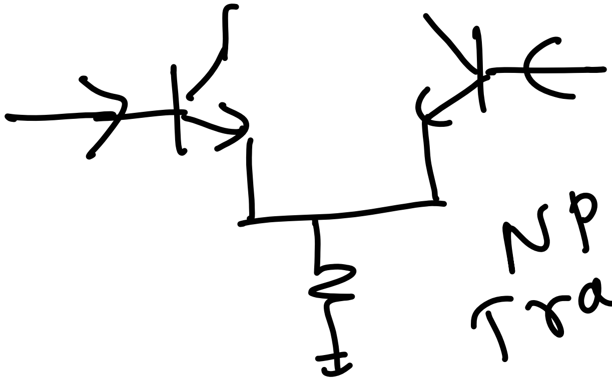
741

bias current flows from ground to input

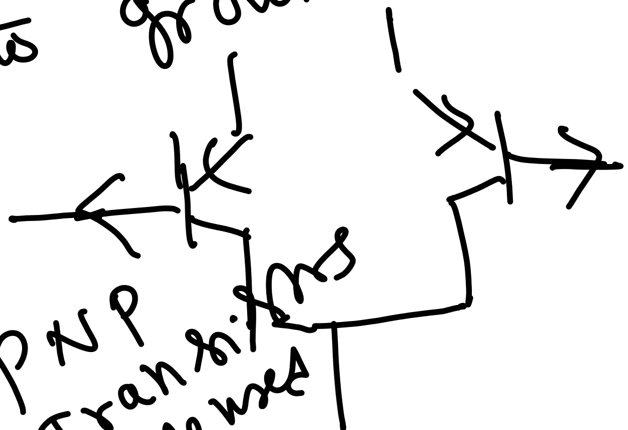


LM324

Here bias current flows from input to ground



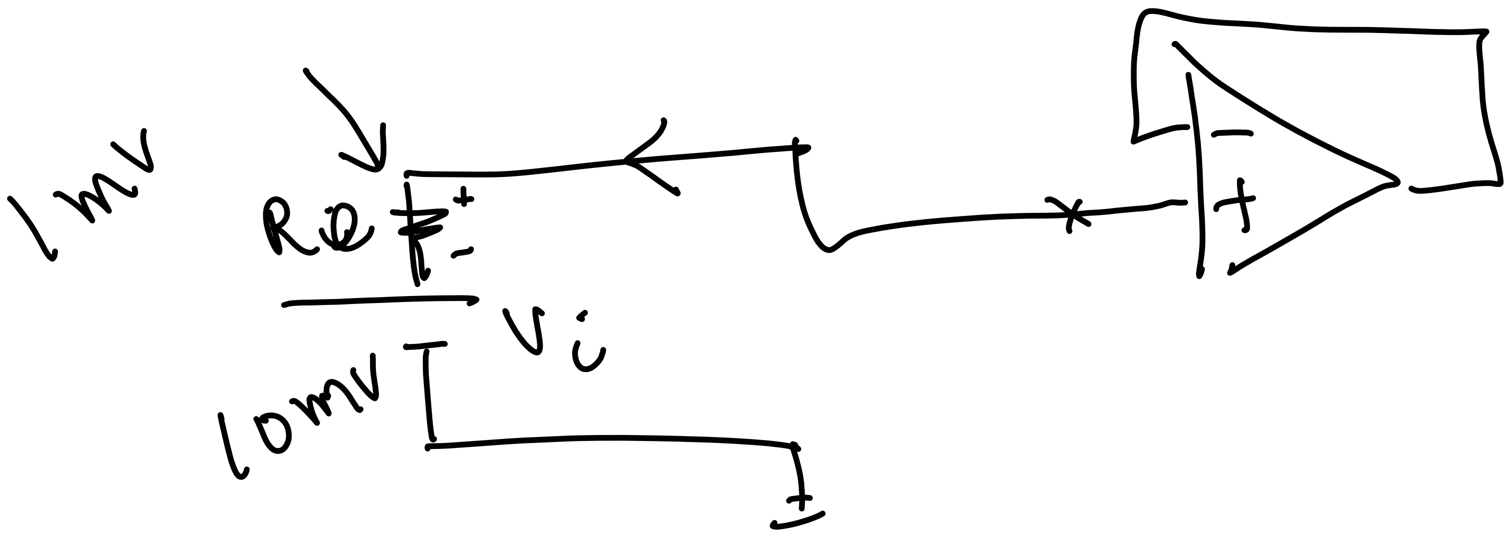
NPN Transistors



PNP Transistors are used

(1) Most of the op amps are NPN TYPE. So bias current flows into the terminal

(2) Only single supply op amps use PNP transistors. (rail to rail op amp)



Voltage drop across the internal resistance R_e

$$= R_e \times I_b$$

$$= 10^{5} \times 10^{-8} = 10^{-3}\,V = 1\,mV$$

This voltage is added to the input source.

So the net effective voltage seen by the op amp $= 10\,mV + 1\,mV$

$$= 11\,mV \text{ (For 324 op amp)}$$

In case of 741 op amp

The effective voltage at the input of the op amp

$= 10\,mV - 1\,mV = 9\,mV$

In addition to this voltage loss due to input resistance must be calculated as before.

This error is same for both the types of op amp. There is no difference.

— x — x —

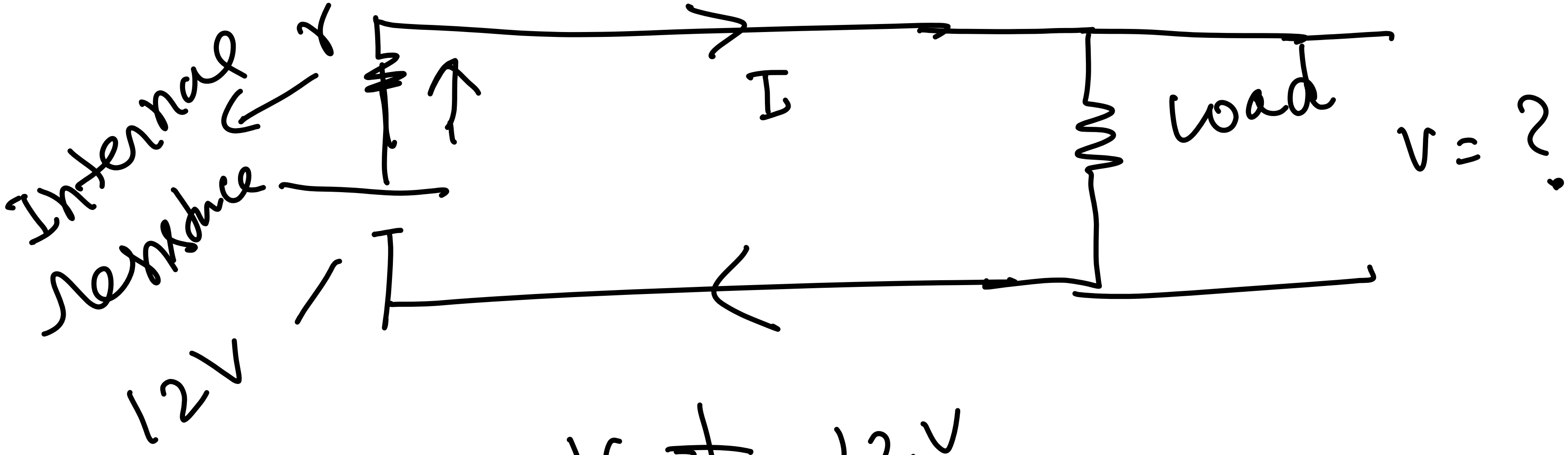
L. No: 29

Output resistance of the op amp

R_o = output resistance $\neq 0$

① Output resistance of the op amp is not zero
② It varies with the gain

Output resistance ?



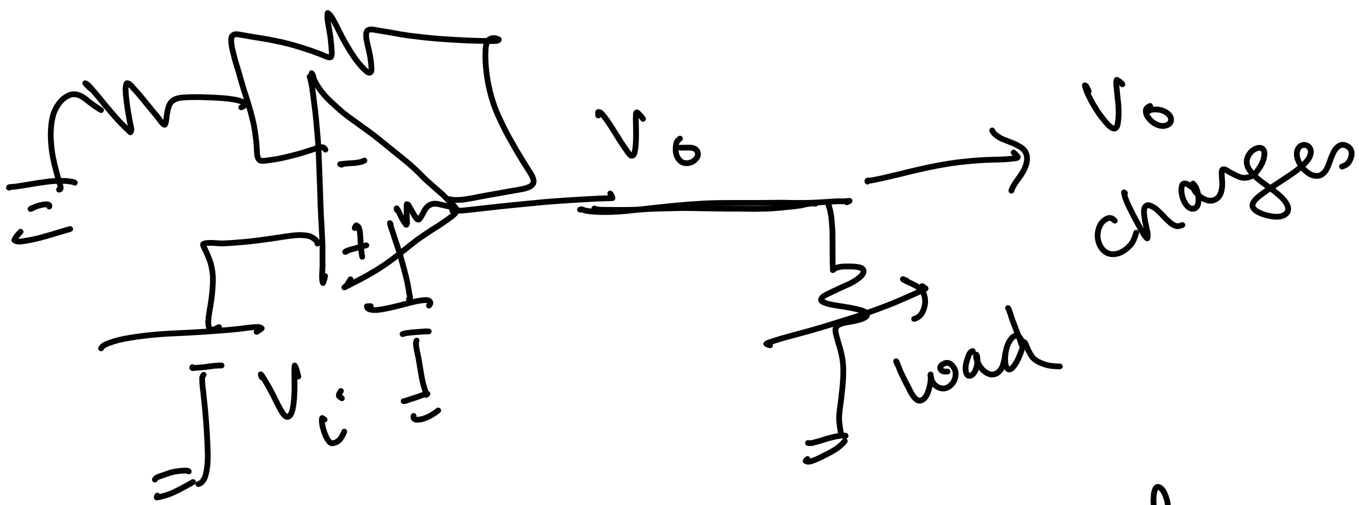
$V \neq 12V$

Voltage drop on the internal resistance $r = rI$

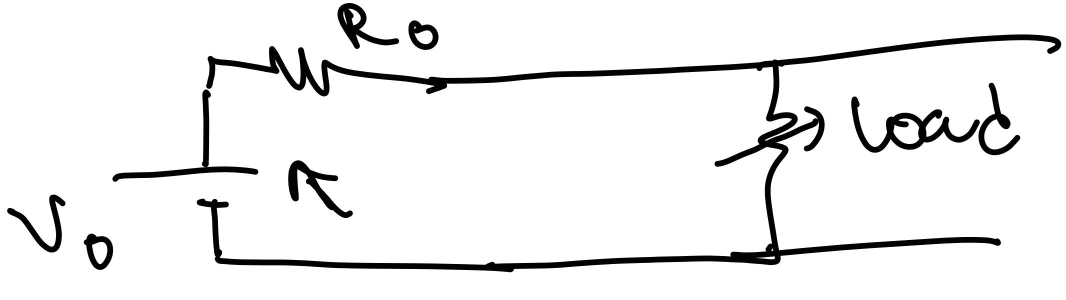
Load Vol $= 12 - rI$ then I changes

So $V_L = 12 - rI$

So with load current
load voltage changes



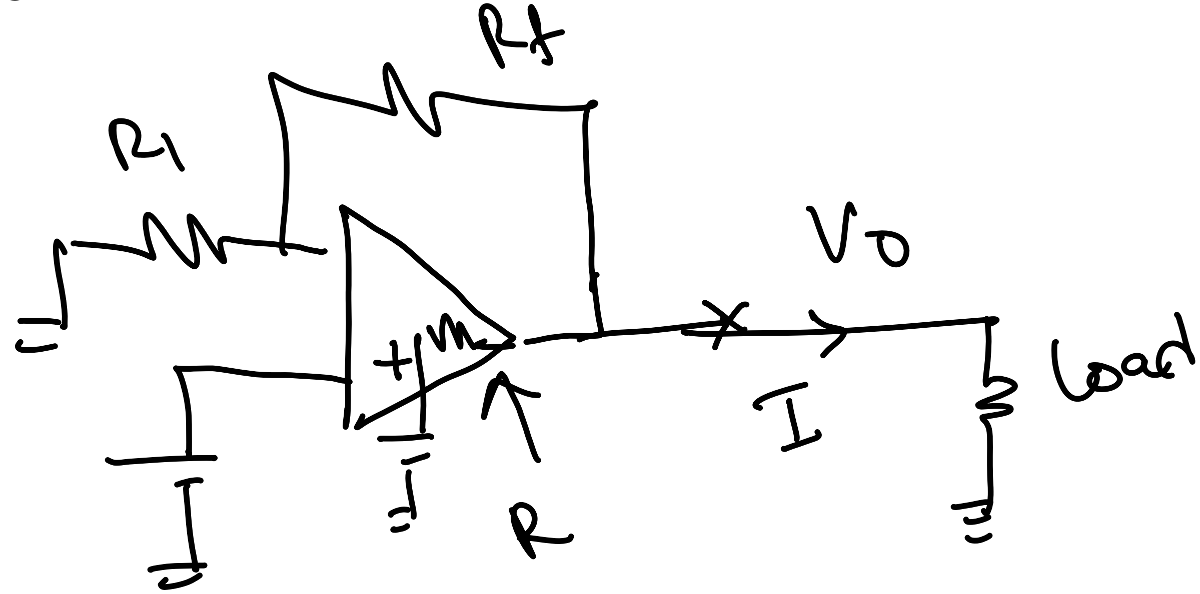
ie output voltage of the op amp
changes with the load resistance



For any voltage source there is a finite internal resistance. So output voltage will change with the load.

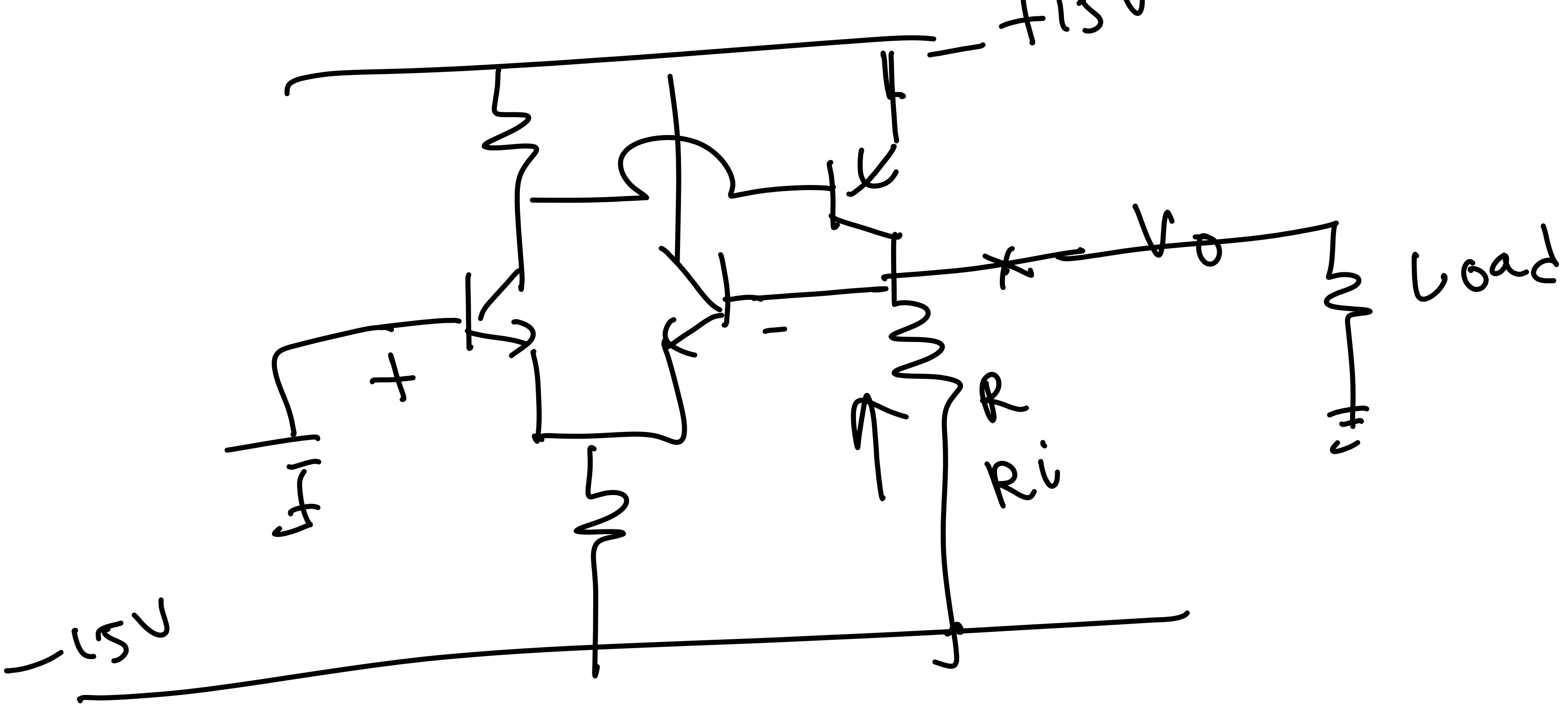
In the case of op amp

What is R_o ?



Assuming R_f and R_1 are very large

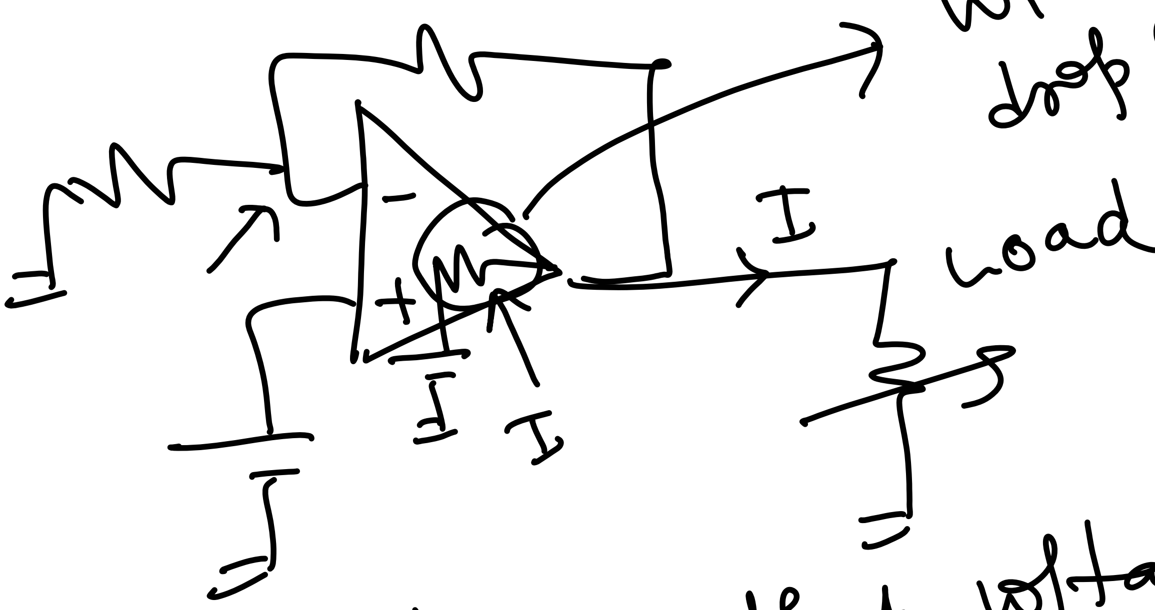
We can take i I as the full current of op amp.



In this case the resistance R is acting as a internal resistance.

But the effective output resistance is not R because of the closed loop feedback most of the voltage that is lost is compensated

So the effective output resistance is much smaller than R.



The effective output voltage change is much smaller

$= R_o I$

Where R_o is effective output resistance of the op amp

In a typical op amp circuit = $R = 100\,\Omega$

but R_o is few milli ohms

In real life

when load resistance is decreased the voltage at the inverting terminal decrease so that output vol is put back.

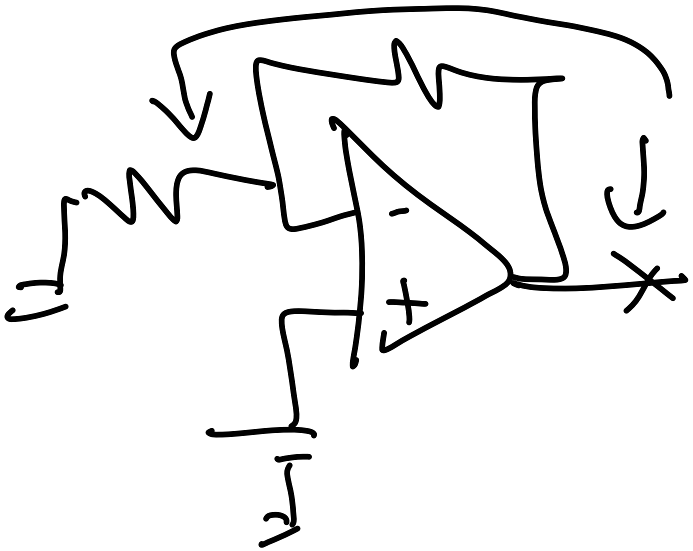
But output vol will not be equal to original.

voltage drop on the internal resistance of the op amp = RI

Assume the effective output resistance is R_o

i.e. $R_o I$ = output voltage change

So what is the voltage change at the inverting ter. of the op amp ??



The voltage change at -ve ~~input~~ is

$$= \frac{R_o I}{g}$$

where g is the closed loop gain

what is the relation between

change in vol at -ve ter.
and output terminal

$\left(\text{change at inverting terminal}\right) \times A =$ output vol change

$$RI = \left(\frac{R_0 I}{g}\right) A$$

$$R = \frac{R_0 A}{g}$$

$$R_0 = \frac{R \times g}{A}$$

Output resistance

$$R_o = R \times \frac{g}{A}$$

For example for most of the op amp $R = 100\,\Omega$

$$R_o = \underbrace{100}_{\text{internal resistance}} \times \frac{\overbrace{100}^{\text{closed loop gain}}}{\underbrace{20{,}000}_{\text{open loop gain}}}$$

$$R_o = \frac{10^4}{2 \times 10^4} = 0.5\,\Omega$$

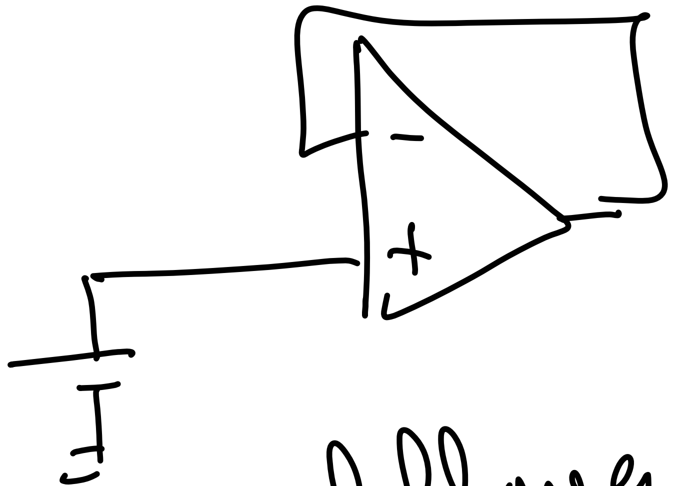
For 741 for voltage follower

$$R_o = 100 \times \frac{1}{20\,000}$$

$$= \frac{1}{200} = 5\,m\Omega$$

When the closed loop gain is less than the output resistance also low

So voltage follower gives lowest output resistance



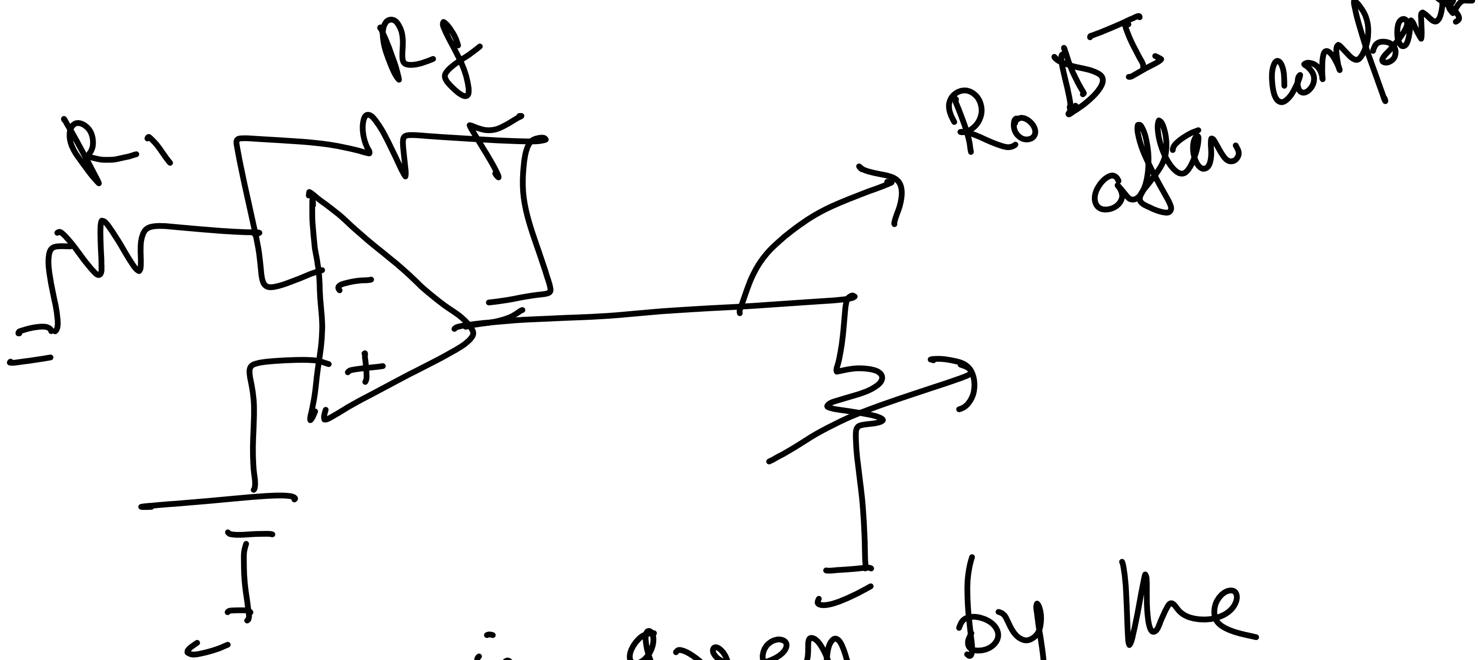
(1) voltage follower gives highest input resistance

(2) Voltage follower gives lowest output resistance

$$R_0 = 100 \times \frac{1\cancel{000}}{20\cancel{000}}$$

$$= 5\Omega$$

For $g = 100$ then output resistance goes up to 5Ω !!



R value is given by the manufacture in the data sheet

In real case

Rf also acts as a load

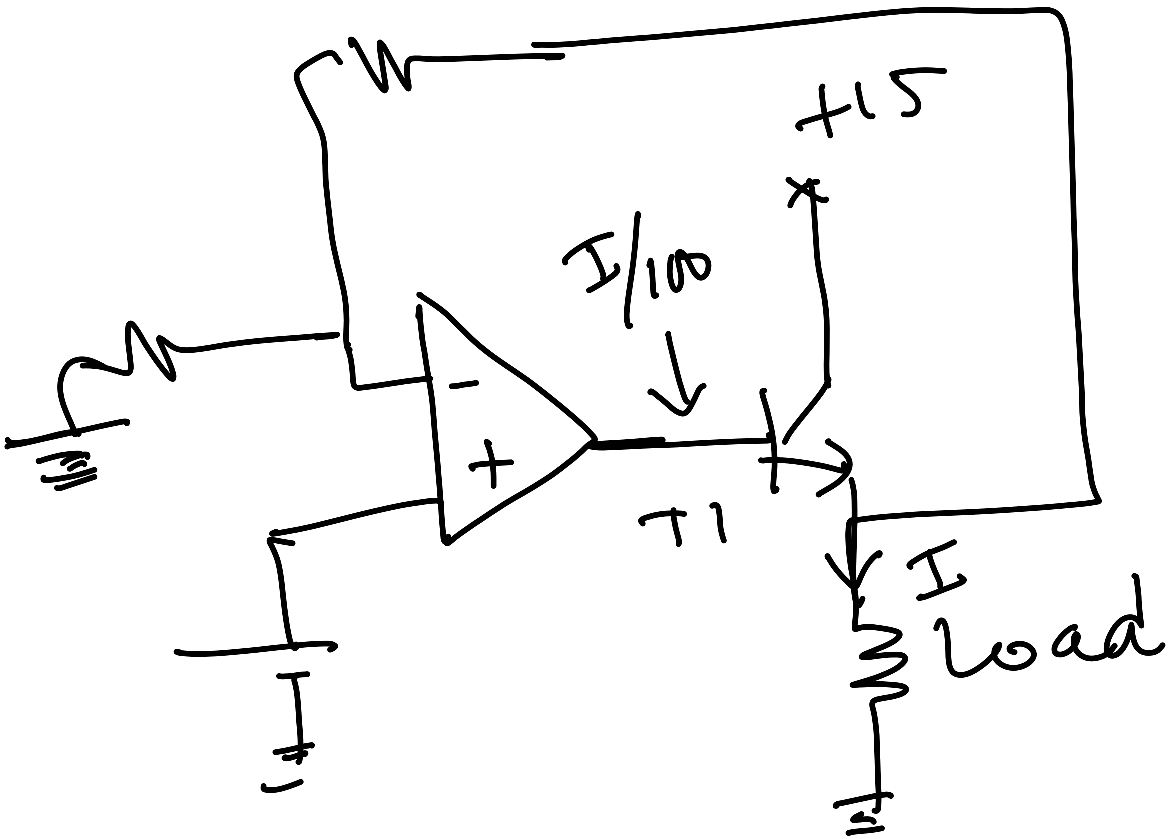
If R_F is small then output voltage will decrease

So one should not use low value of R_F.

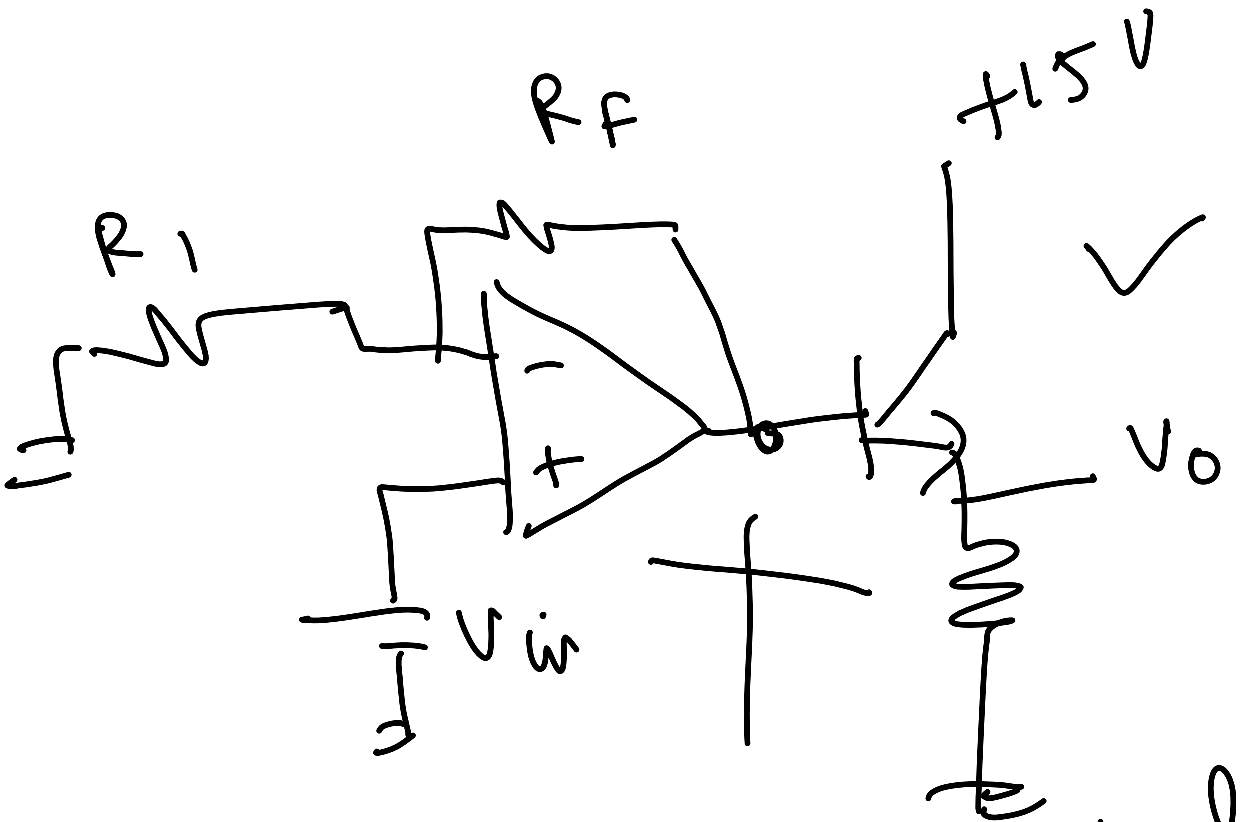
Low value of R_F reduces the bias current error. but it will increase the output resistance related error

It should be notes that normally there is a internal current limit for every op-amp.

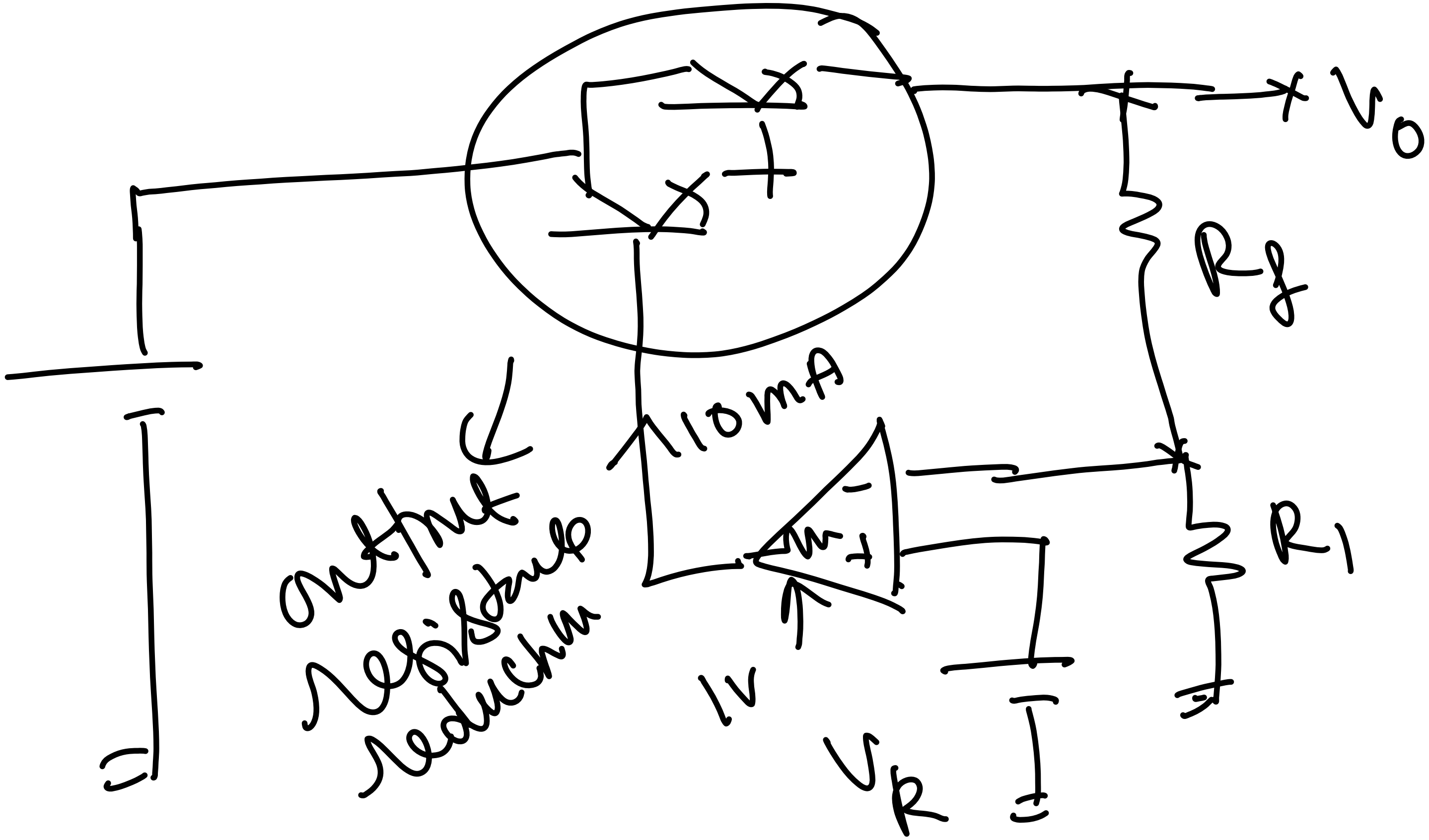
For most of the op amps
~~out~~ I is limited to 10mA
So normally bias current
related error is much larger than
output resistance related error



For $\beta = 100$



This is not acceptable
V_{BE} is a loss !!
As V_{BE} changes with temp



voltage regulator

It can also be considered as non-inverting amp.

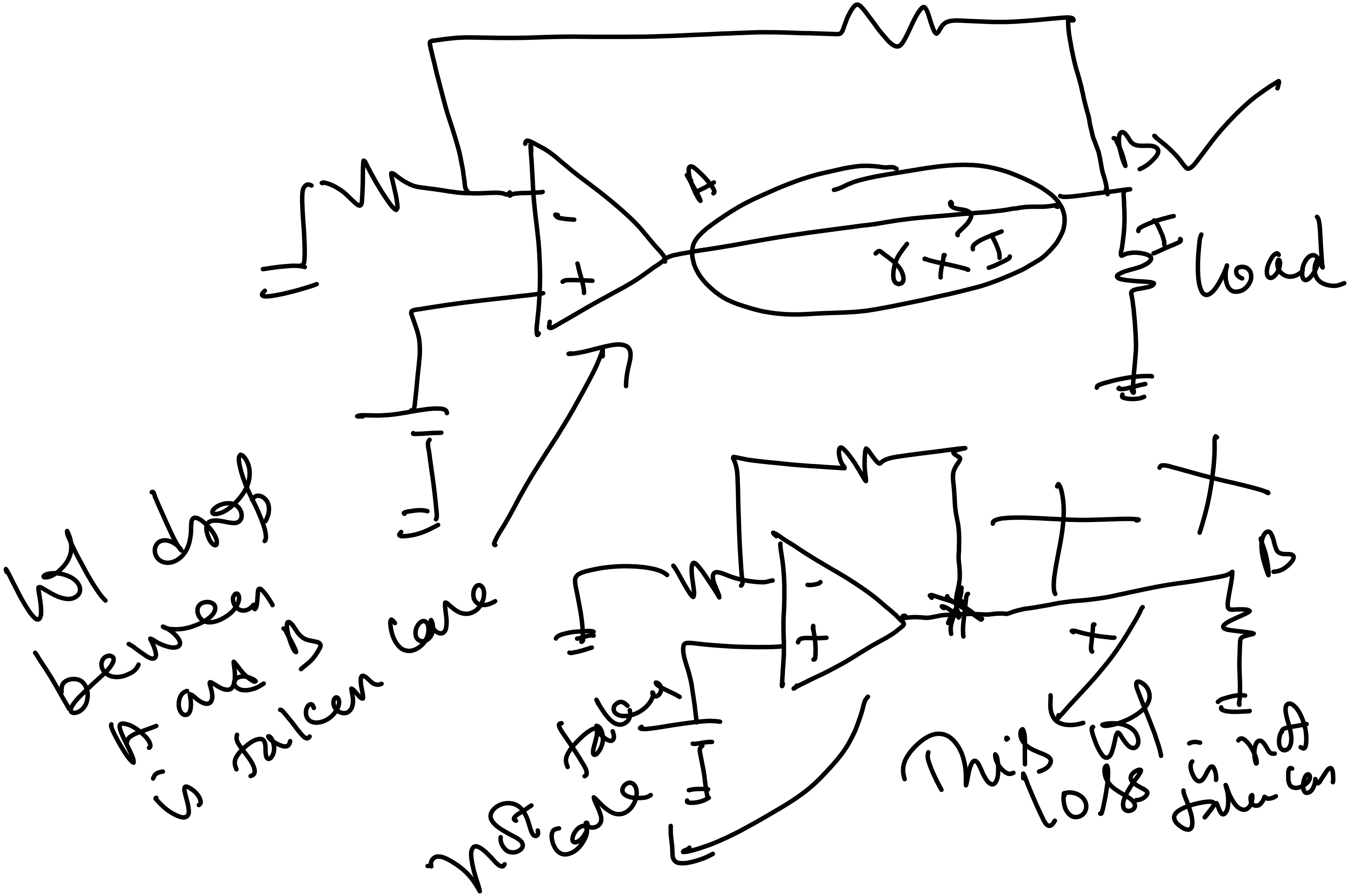
$$V_o = V_R\left(1+\frac{R_f}{R_1}\right)$$ gain

It is a non-inverting amp with low output resistance

At the output of the op amp

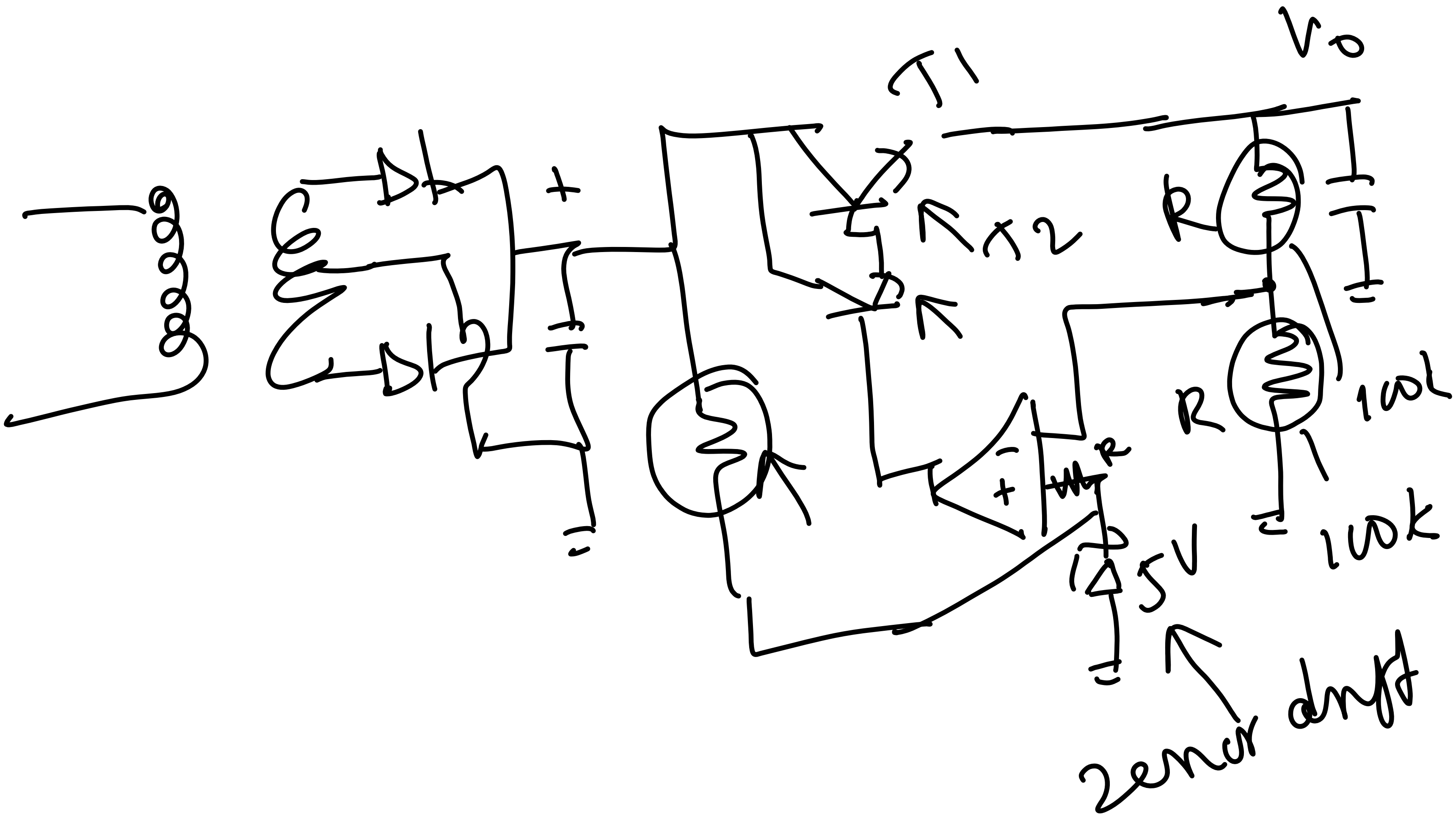
$R_o \times 10\,mA$ = is the expected change

$$R_o = 100 \times \frac{2}{20000} = \frac{200}{20000} = 10\,m\Omega$$


A
B
$\gamma \times I$
I
load
Vol drop between A and B is taken care
not taken care
B
This vol loss is not taken care

Lecture-30

Calculations of op-amp errors



$V_o = 5V \times 2 = 10V$

V_{BE} drift of T_1 and T_2 is actually compensated by the loop gain

(1) V_{OFFset} drift of op amp directly affects the output

For 741 drift $= 15\mu V/°C$

For 100°C $= 1.5mV$

So total drift is 3mV

Zener drift error

LM336 → drift

Spec = 30ppm/°C

For 100°C drift

$$= \frac{30 \times 100}{10^6} \times 5 = \frac{150 \times 10^2}{10^6}$$

$$= 150 \times 10^{-4} = 15\,mV$$

The expected output drift = 30 mV

③ Resistance drift error – R_1 and R_2

④ Bias current
drift error → R_1 and R_2
Reduce the bias current
error by reducing the R_1
and R_2

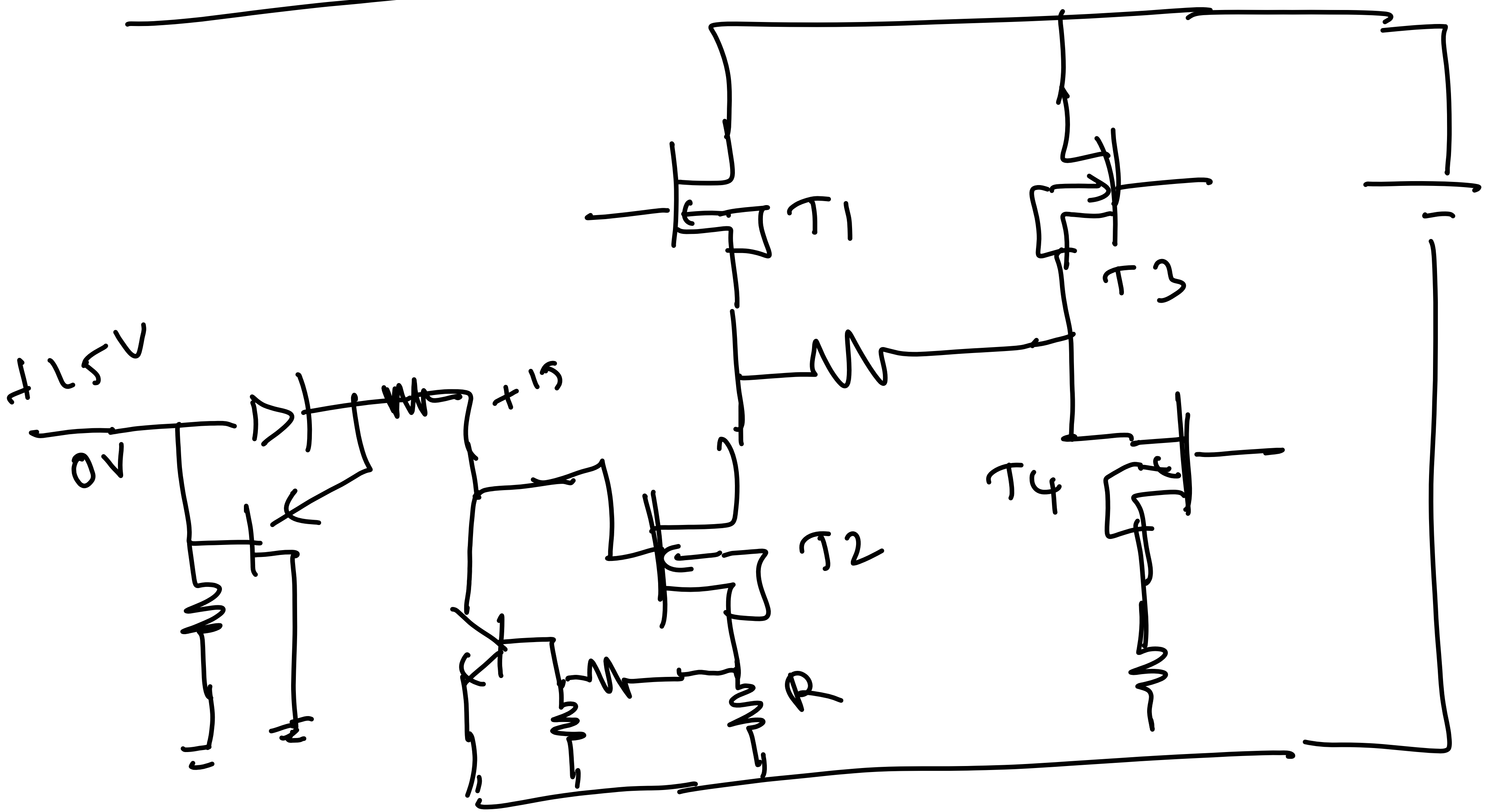
or

Add Resistance in the
non inverting input.

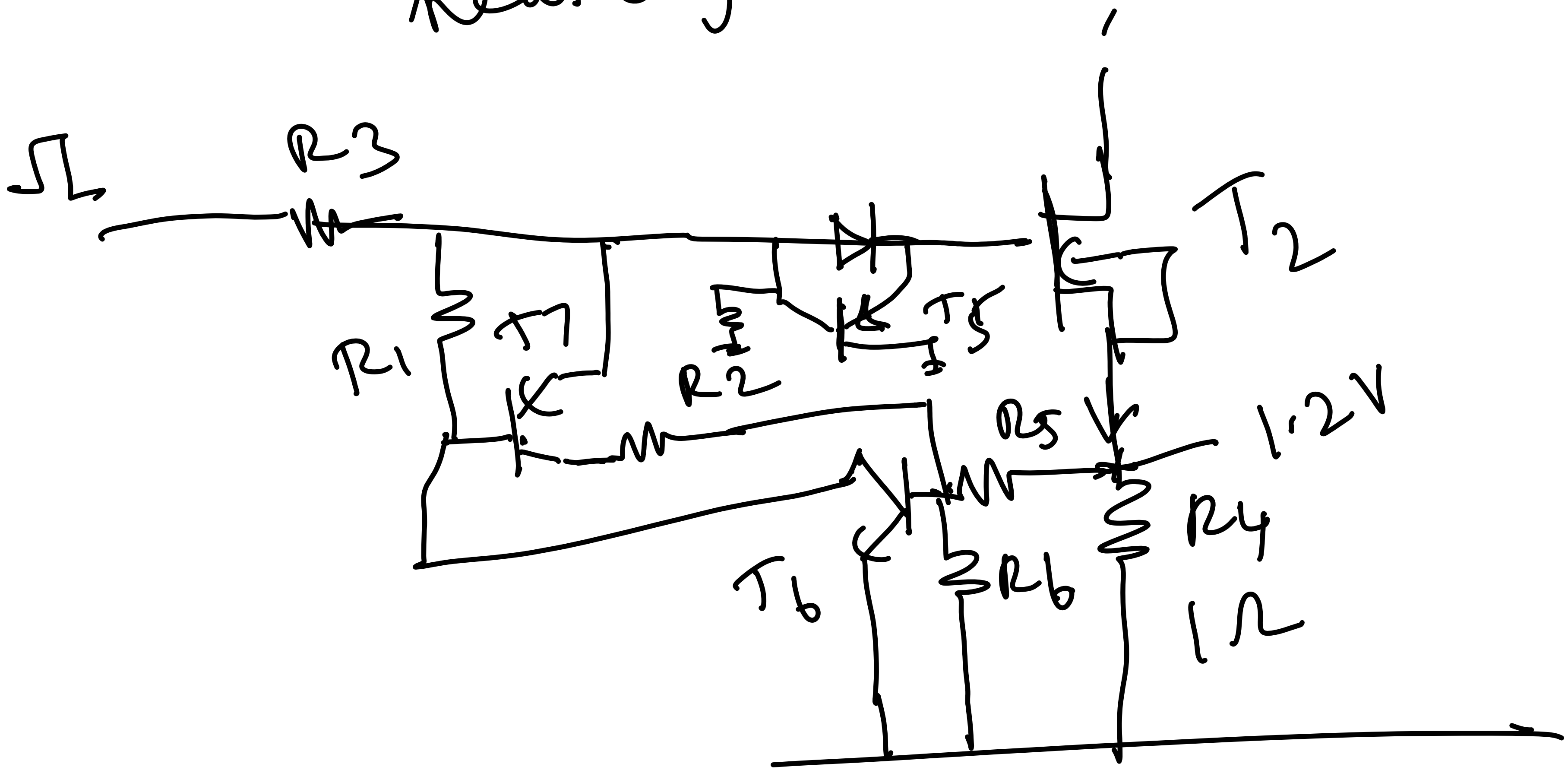
Circuit - 2

H - bridge drive

Current sensing error



This circuit will limit the current but it will damage the T_2 because of heating



wl ace R_4 is applied to T_6

When T_6 $V_{BE} > 0.6V$

T_6 is turned ON

This makes T_7 to turn ON

This makes wl at the base of T_6 to be present even with out the current

i.e T_6 is latched

This makes T_2 to go off.

So T_2 will be off for the remaining period of the pulse

The latch will be reset at the next zero pulse input

If $R_4 = 1\Omega$

Then at 1.2A current

0.6 V will appear at the

base of T_6

So current limiting is taking

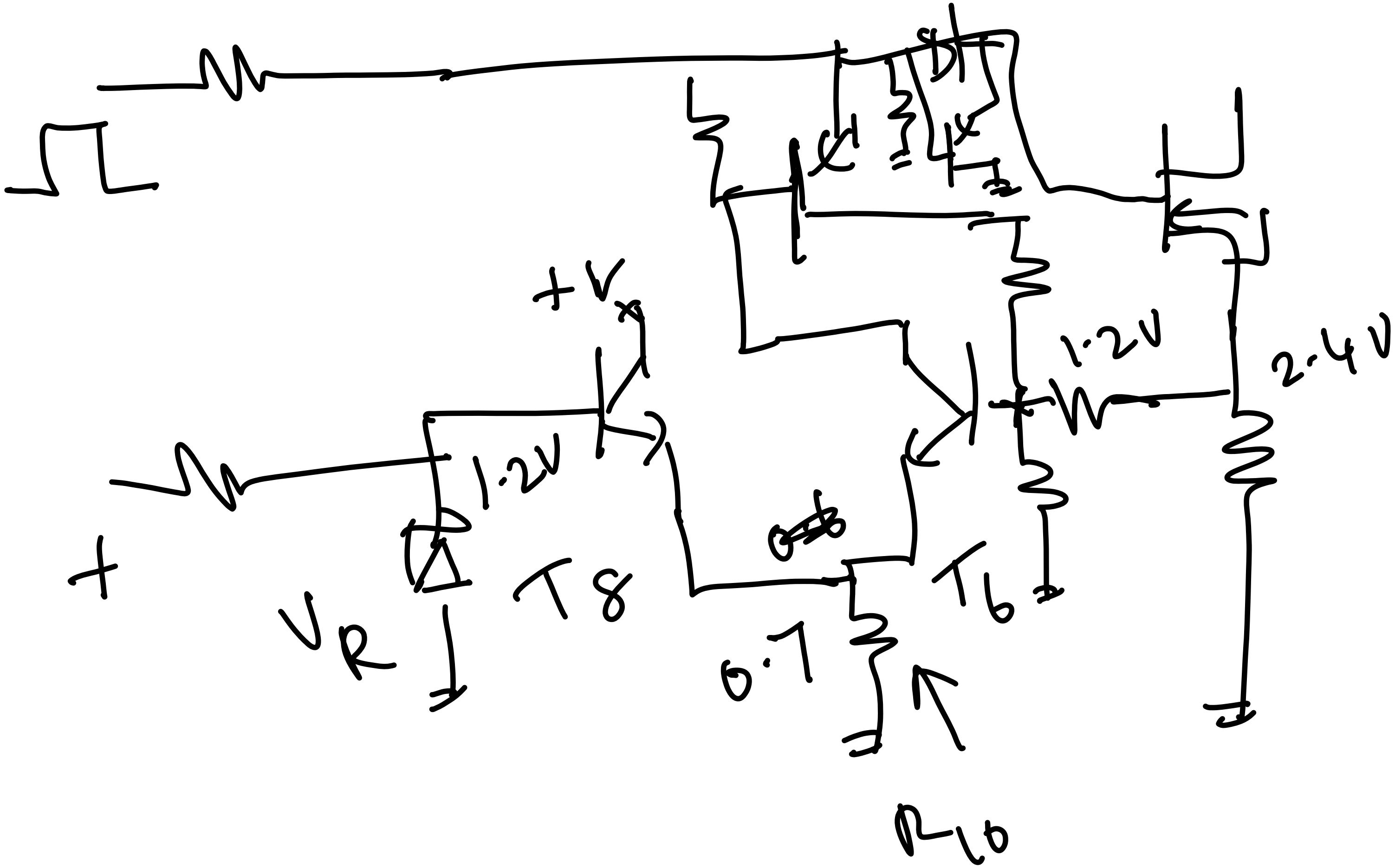
place at 1.2A → This is at 25°C ambient

At 75°C ambient

then T_6 needs only 0.5 V

So current limiting will

take place at 1A


+Vx
1·2V
+
VR
T8
0·7
R10
T6
1·2V
2·4V

Base emitter
V_{BE} drift of T_6 is
eliminated by using T_8 and
V_R
This is because when V_{BE} changes
the voltage across R_{10} also
changes

For example at 25°C
vol R_{10} $\longrightarrow$ 0.6 V
At 75°C $\longrightarrow$ 0.5 V

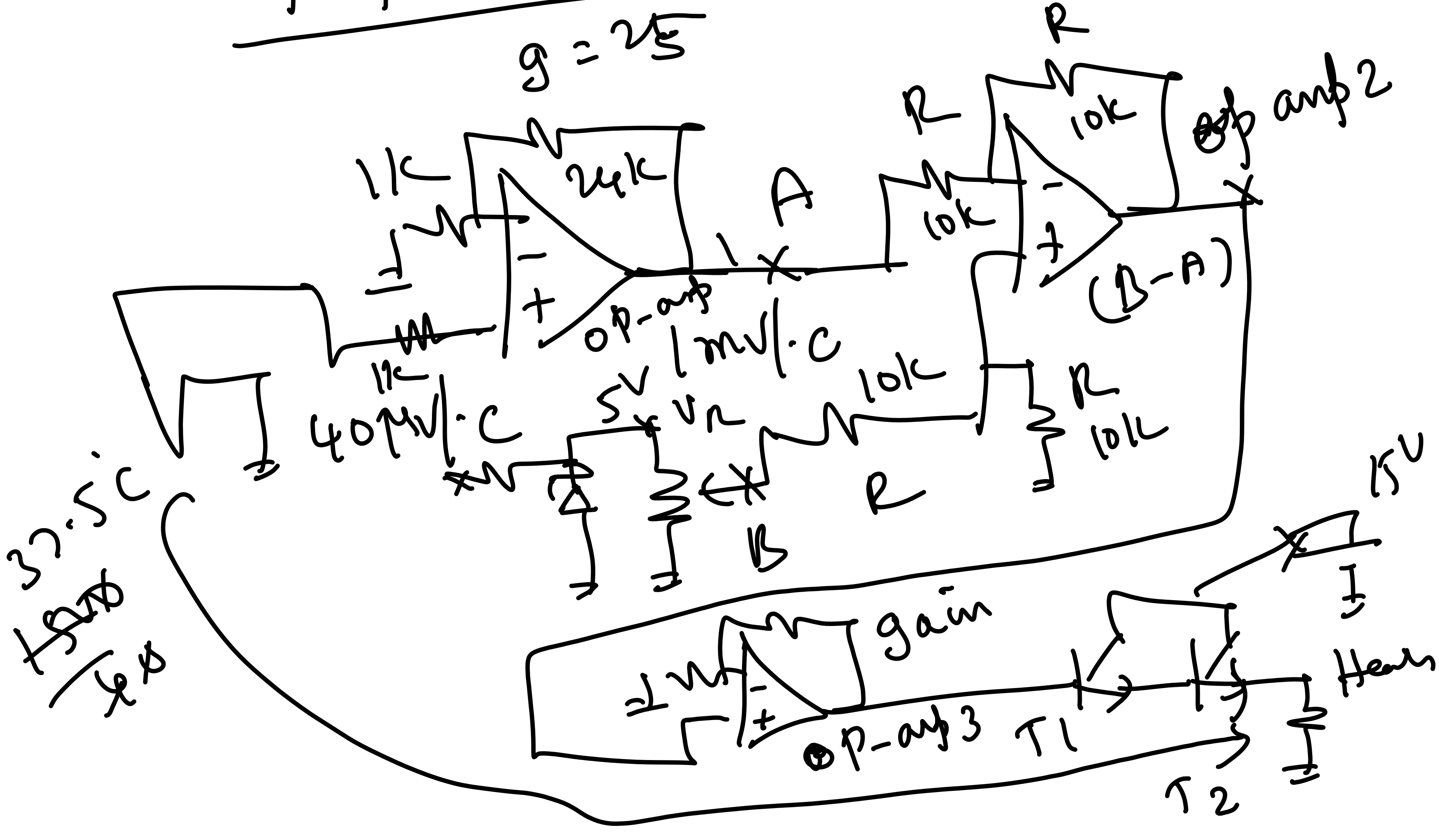
T_6 conducts at 25°C ambient
when V_{BE} is 0.6 V ie
voltage at the base is 0.6 + 0.6
= 1.2 V

T_6 conducts at 75°C ambient
when V_{BE} is 0.5 V ie
voltage at the base is 0.5 + 0.70
= 1.2 V

ie at all temperatures current
limiting is taking place when
T_6 base vol is at 1.2 V

Circuit 3 — Drift Calculation

Thermo couple based proportional temp controller



Point A gives actual temp.

point B gives set temp

(B − A) → error vol

op amp-3 amplifies the
error voltage and the
amplified vol is applied to the
heater through the Transistors

Calculation of errors

(1) op amp-1 handles the low level signal
So offset voltage drift of op amp-1 is a concern
So use low offset voltage drift op amp.
LM-07 → V_{offset} drift is 0.5 μV/°C. For 100°C ambient temp change 50 μV drift is

expected. This is equal to

$\frac{50}{40} = 1.25^\circ C$

Total error due of offset voltage drift

Op amp 1 $\longrightarrow$ $\frac{50\,\mu V}{40\,\mu V} = 1.25^\circ C$

Op amp 2 $\longrightarrow$ offset vol drift $\rightarrow 1.5\,mV$

error $\longrightarrow$ $\frac{1.5\,mV}{1\,mV} = 1.5^\circ C$

signal $\leftarrow$ 1 mV

op amp 3 $\longrightarrow \dfrac{1.5\,mV}{1.0\,mV} = 1.5\,^\circ C$

Total error $=$

$$\begin{array}{r} 1.25\ + \\ 1.50 \\ 1.50 \\ \hline 4.25\,^\circ C \\ \hline \end{array}$$

If gain in op amp 1 is kept as 50

Then the error comes down

op amp 1 — $1.25\,^\circ C$

op amp 2 error = $\frac{1.5 mV}{2 mV} = \frac{3}{4}$

$= 0.75 °C$

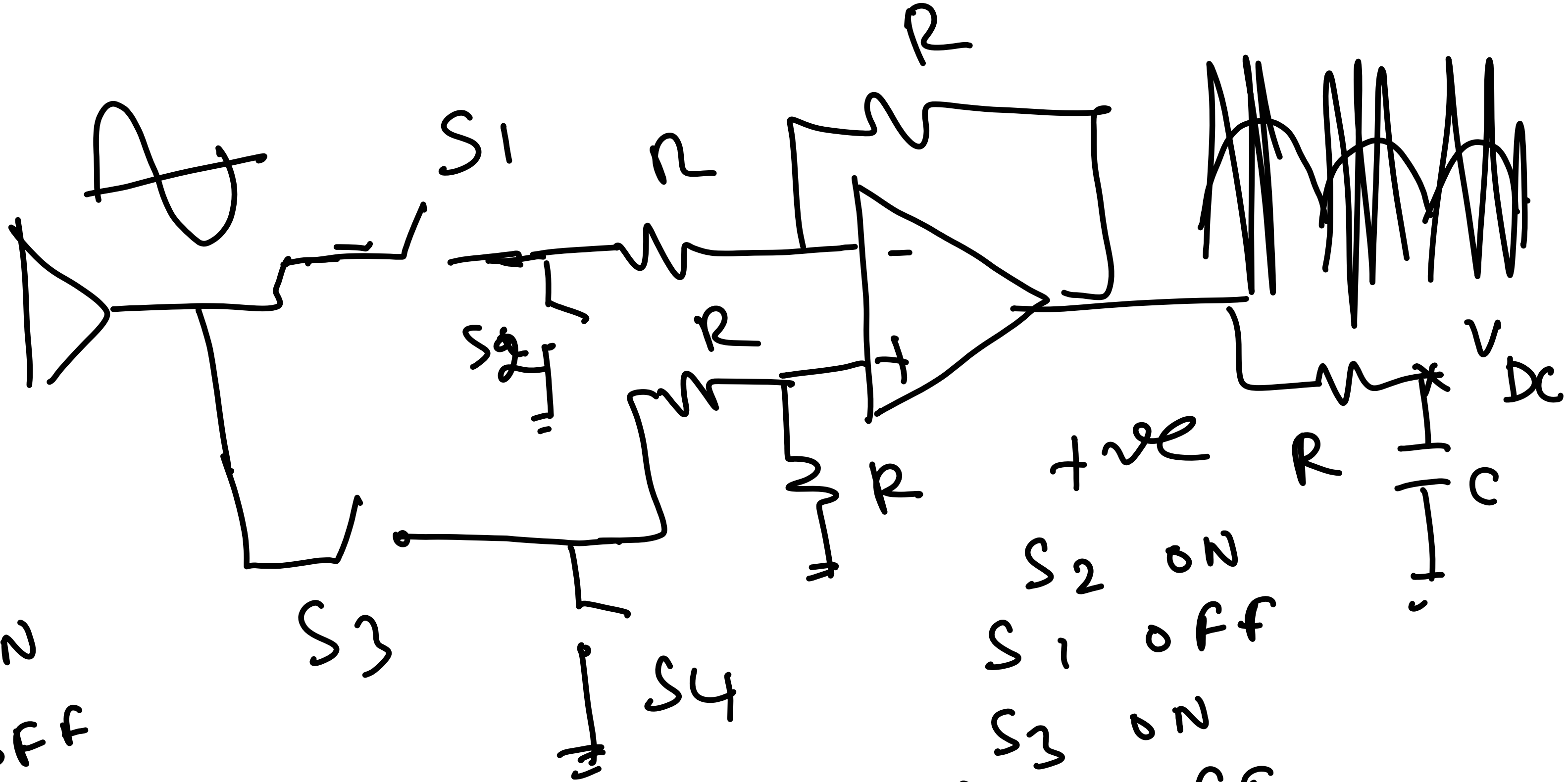
op amp 3 error = $\frac{1.5 mV}{2 mV} = 0.75 °C$

Total error = 1.25 +
0.75
0.75

$\underline{2.75 °C}$

~x ~x ~x ~

Lecture No: 32



-ve

S1 ON
S2 off
S3 off
S4 ON

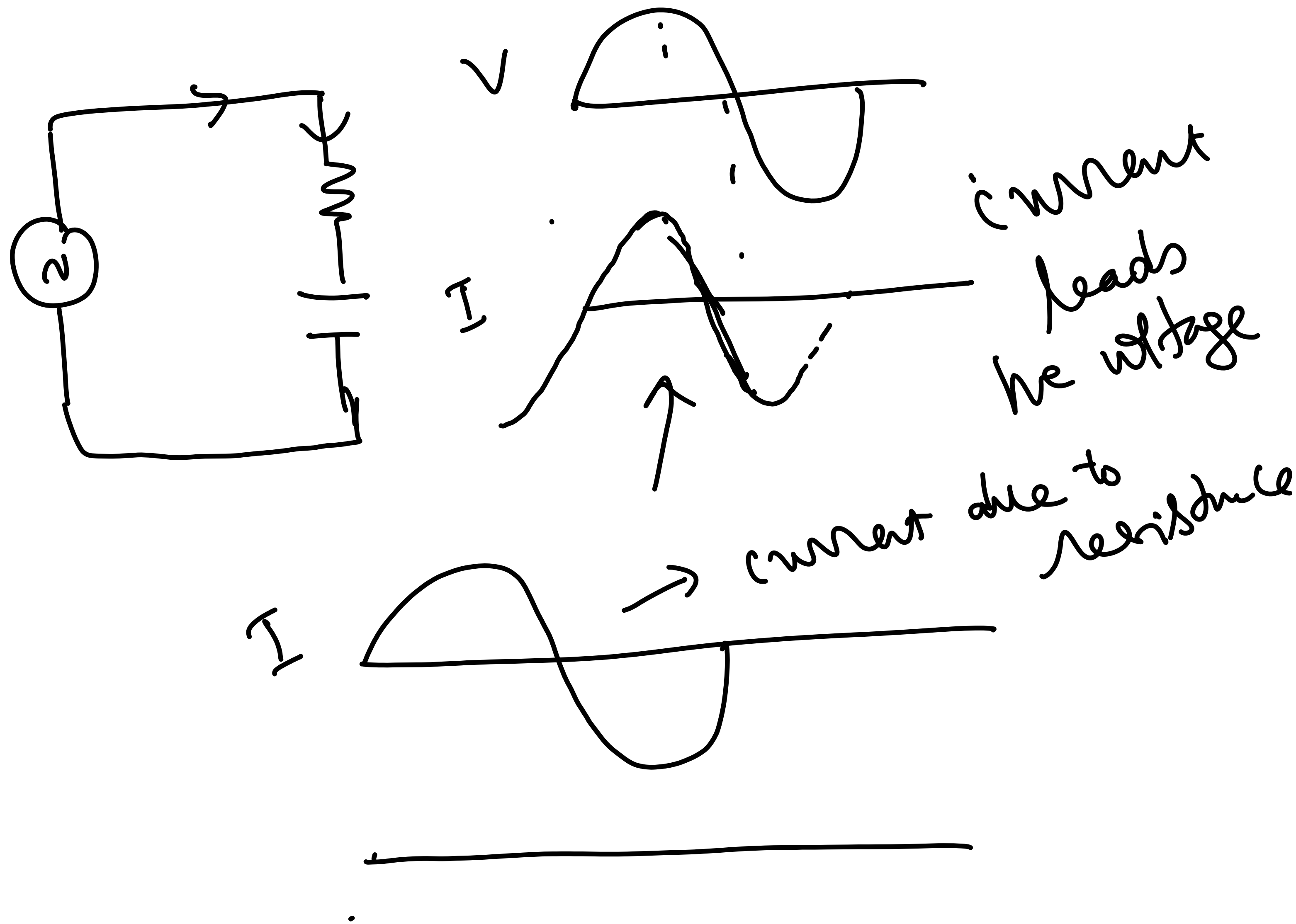
+ve

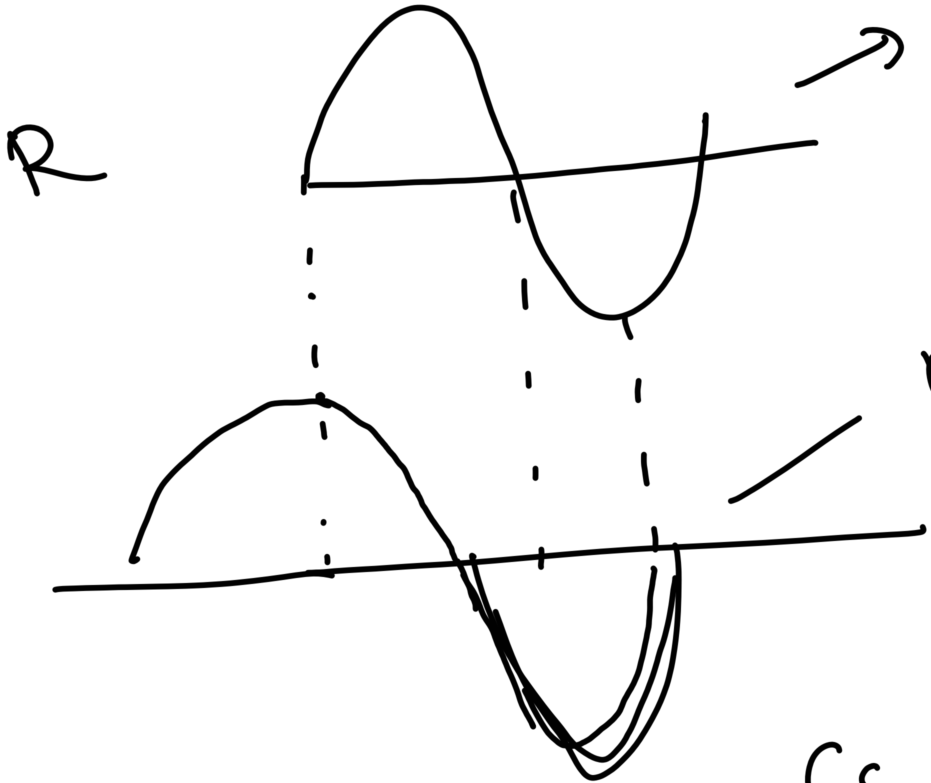
S2 ON
S1 off
S3 ON
S4 off

By means of averging
noise is removed
For this to be effective

(1) RC time constant
much larger than
lowest noise freqnency.

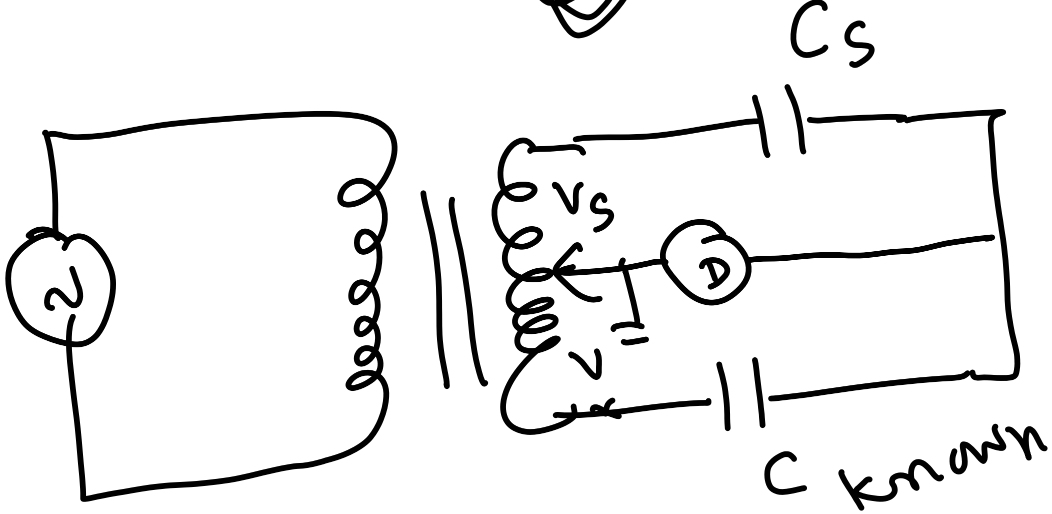
Lock in amplifier to separate out in phase and out of phase components





current due to resistance

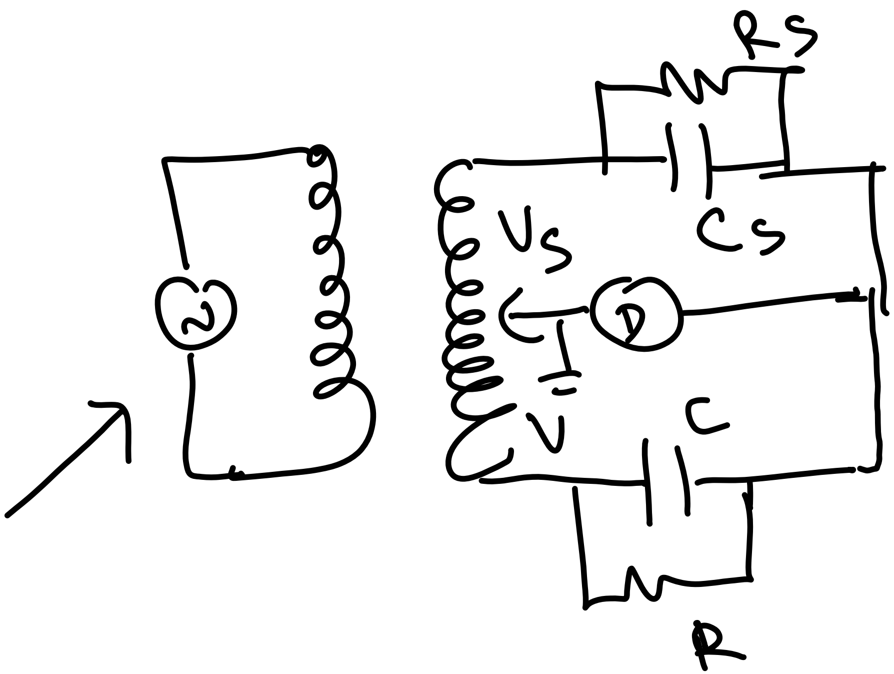
current due to cap



Radio transfr bridge for cap measurement

$$V_s C_s \omega = V C \omega$$

$$\frac{C_s}{C} = \frac{V}{V_s}$$



The Detector current is due to both R and C

Assuming R_s and R is large

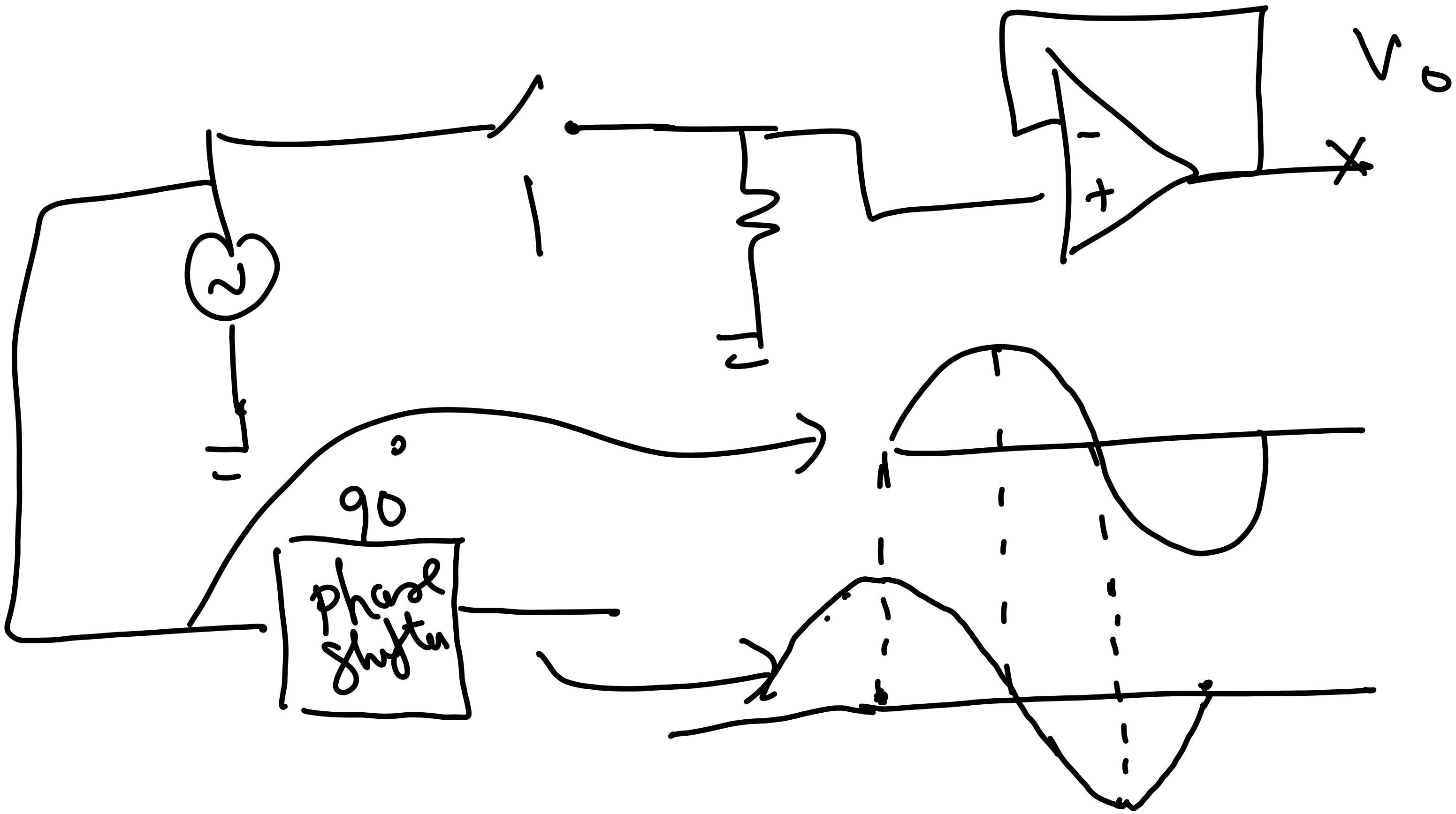
We can balance the bridge for resistance and capacitance separately.

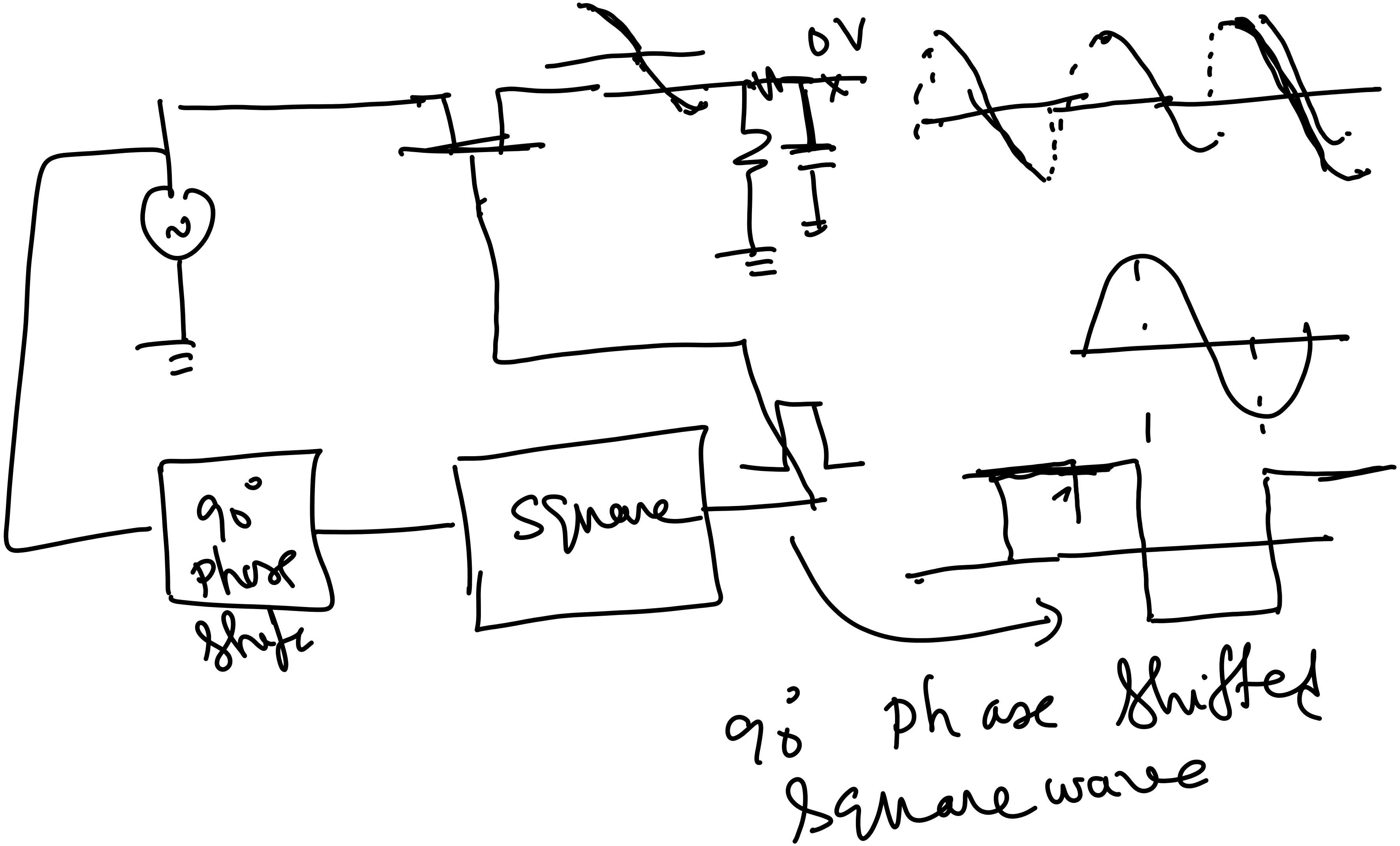
1. To balance resistance
   to make D sensitive to in phase compt. (in phase compared to the applied vol i.e. V_s or V)

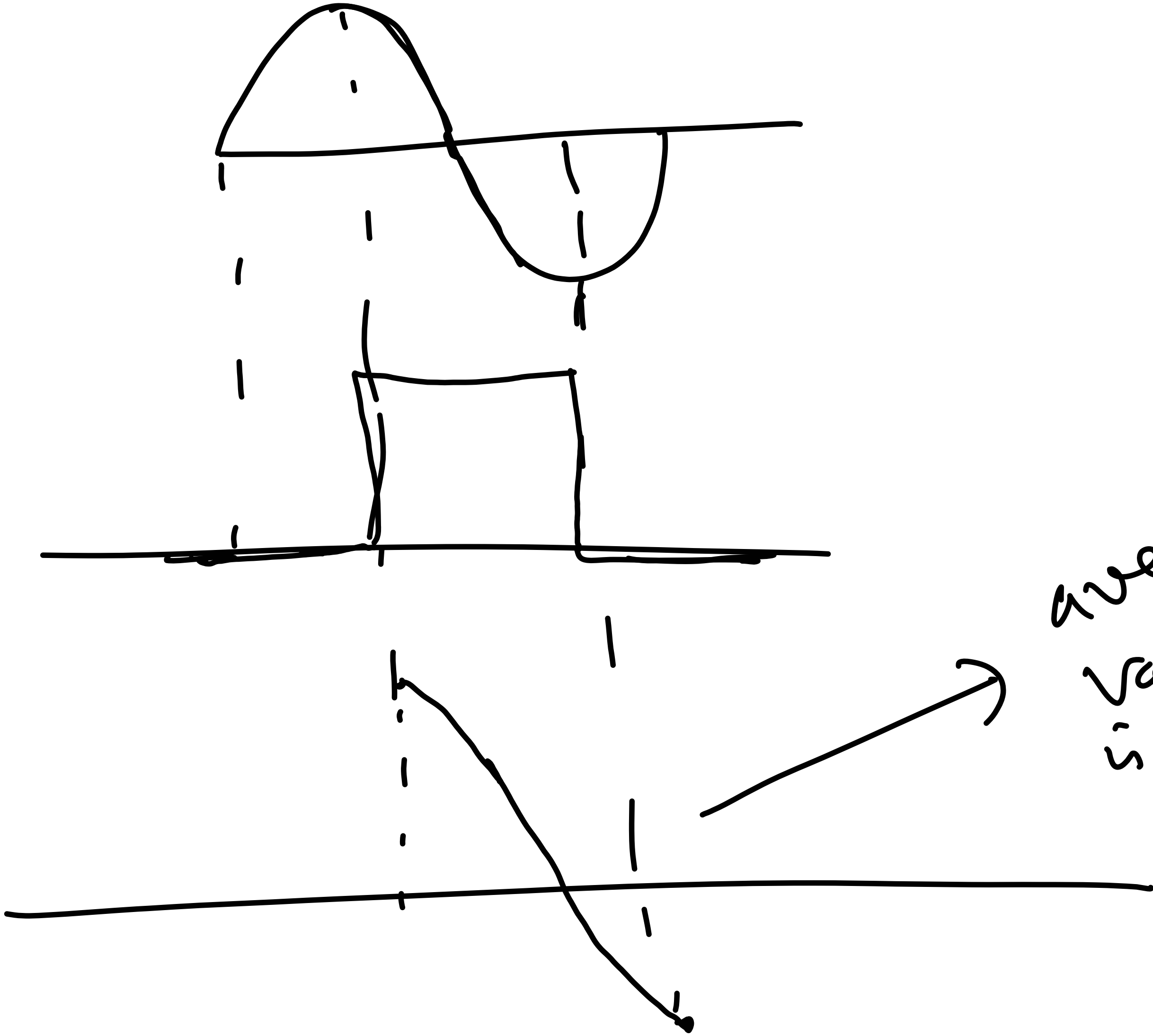
2. To balance the capacitance
   to make D sensitive to 90° out of phase compt.

The Detector D is now
lock-in amplifier

How lock in amp is
sensitive to 90° phase component

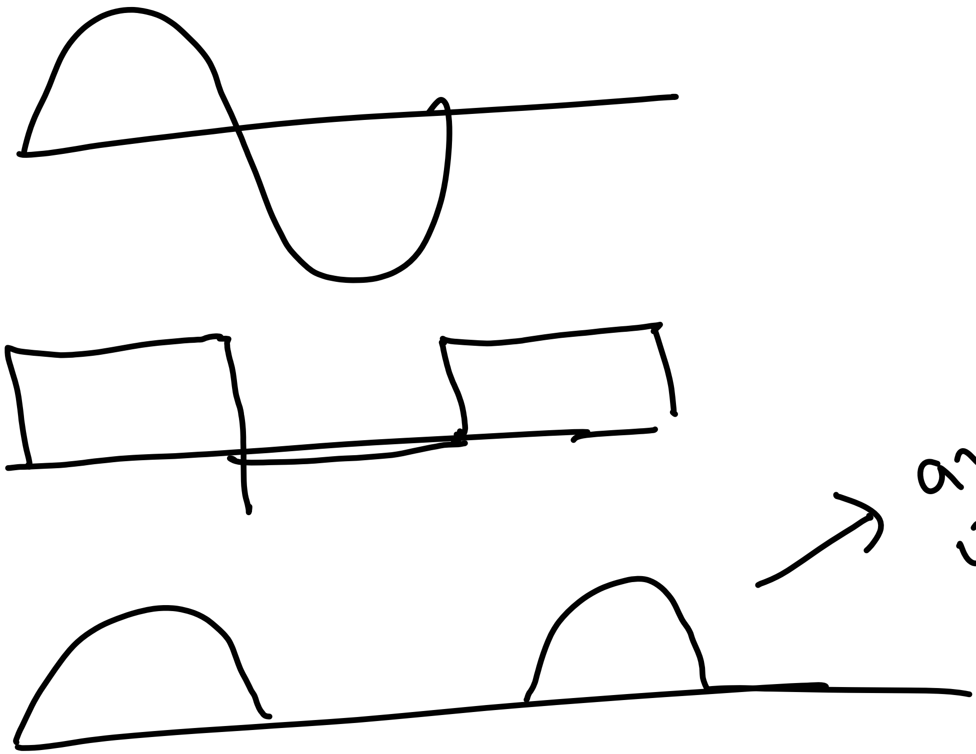



0V
90° phase shift
Square
90° phase shifted square wave

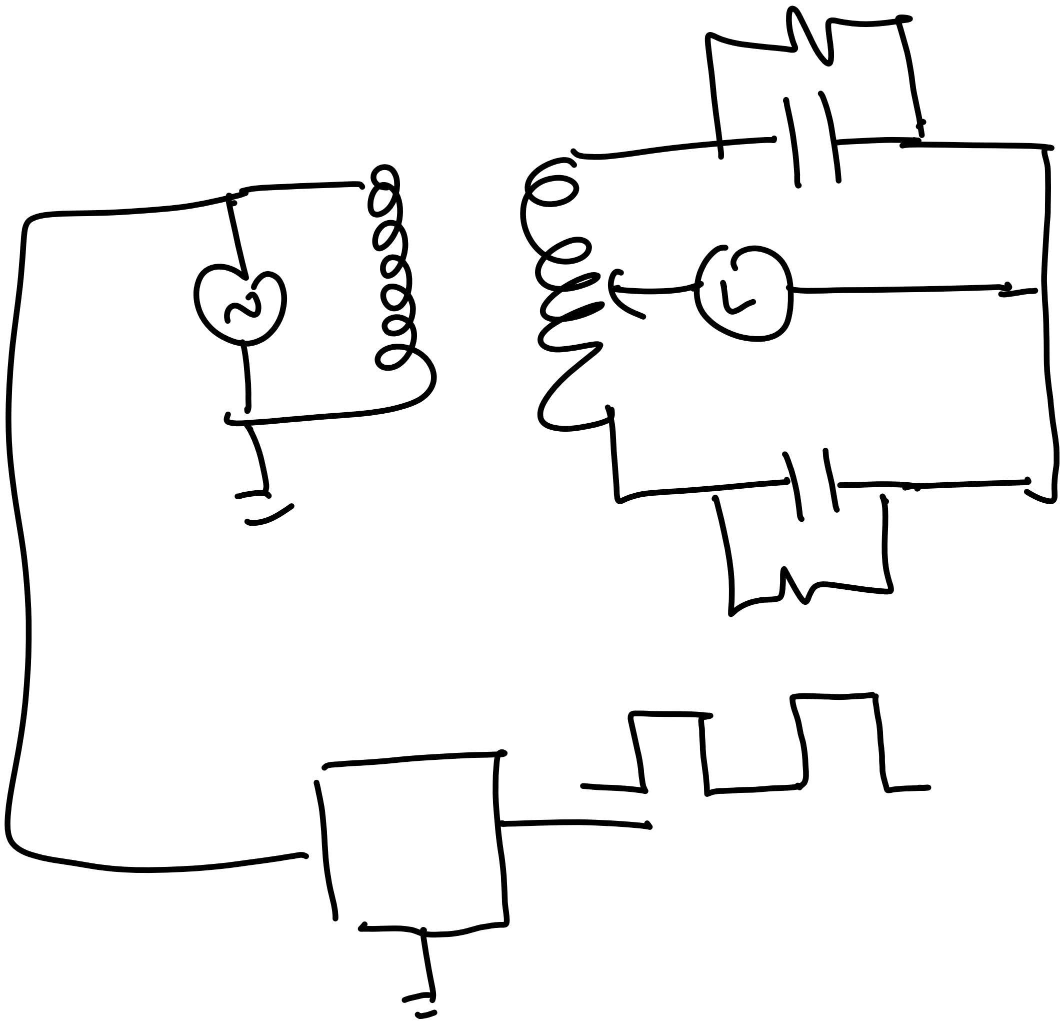


average value is zero

without phase shifter

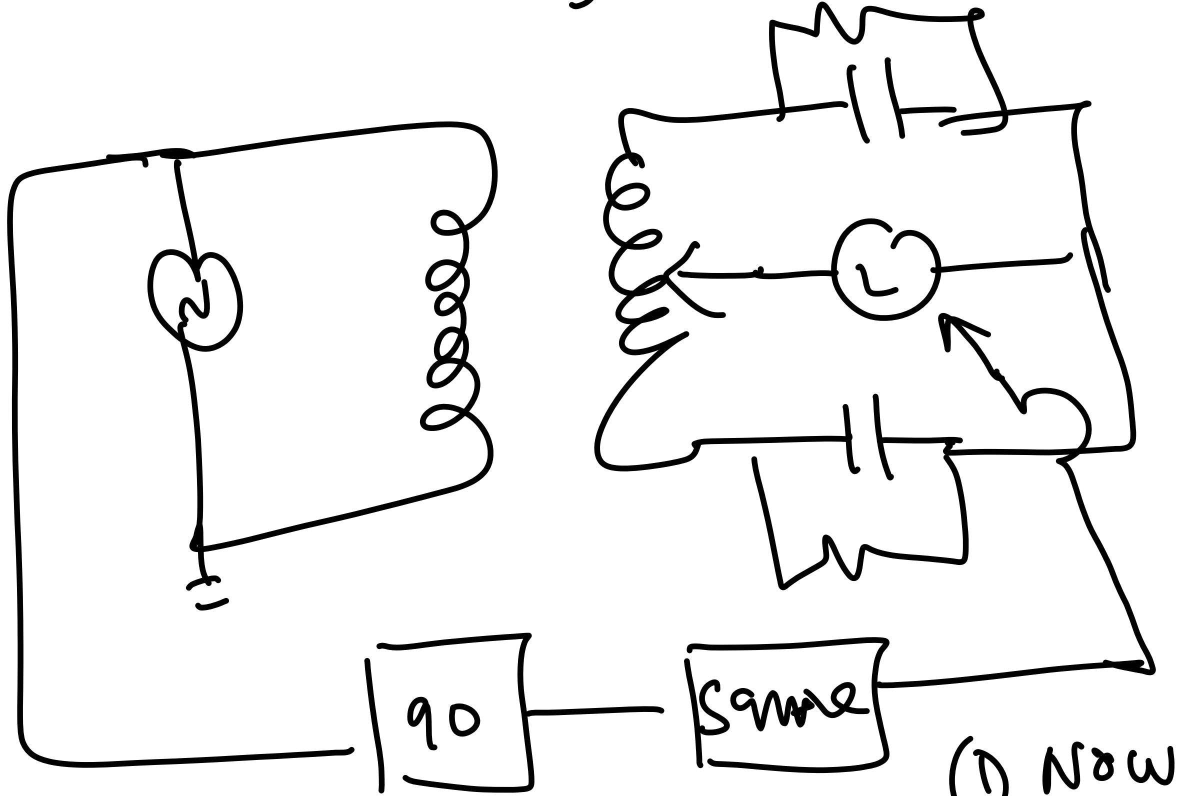


average is not zero



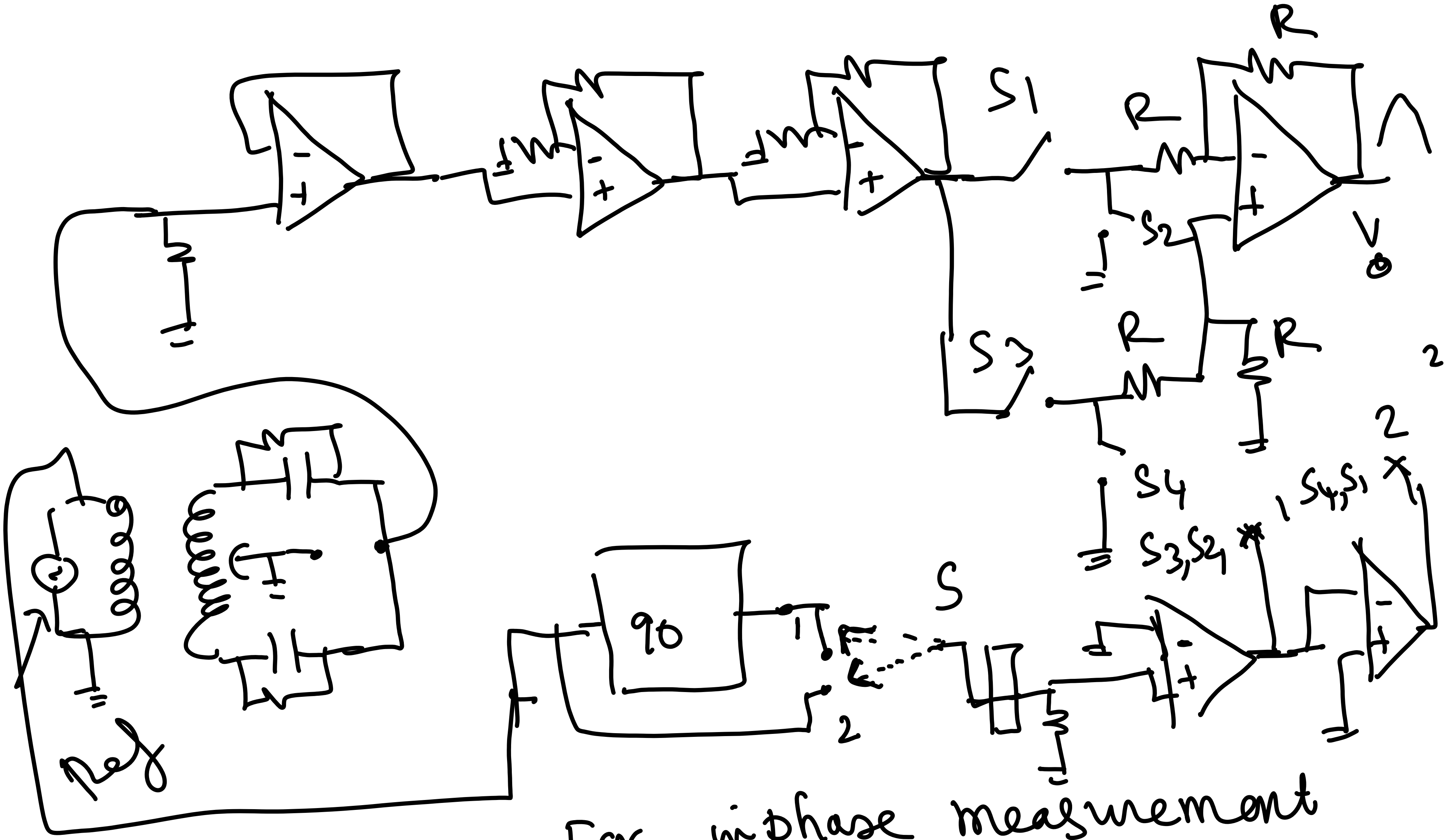
(1) If lock in
amp is
switched with
in phase square
wave then
only resistance
current will
produce output.

② The capacitance current will not produce any output.



① Now lock in sensitive only to capacitance change and not for resistance change.

Designing of lock in amplifier



For inphase measurement

Connect S to point 2

S_3 and S_2 will be ON, S_4, S_1 OFF

So output V_o will be +ve

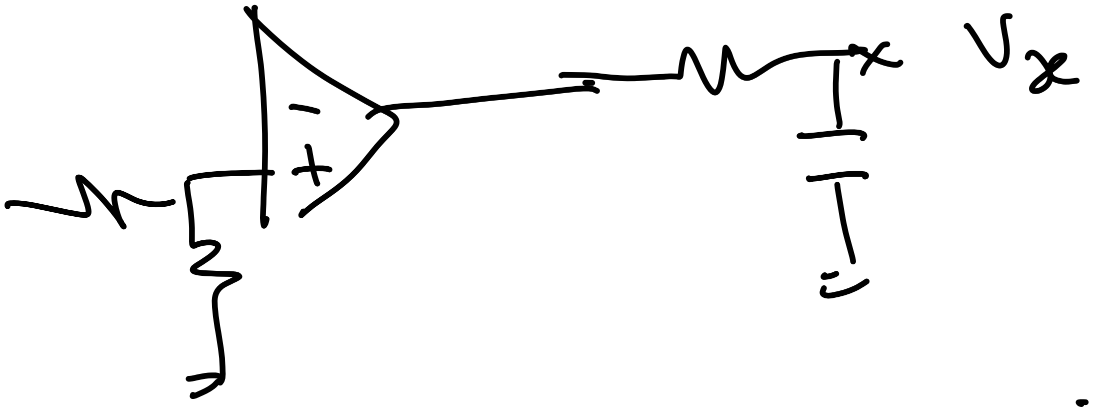
This measures the resistance

i.e. it measures the in phase compt.

To measure the capacitance

Connect Switch S_2 to position 1

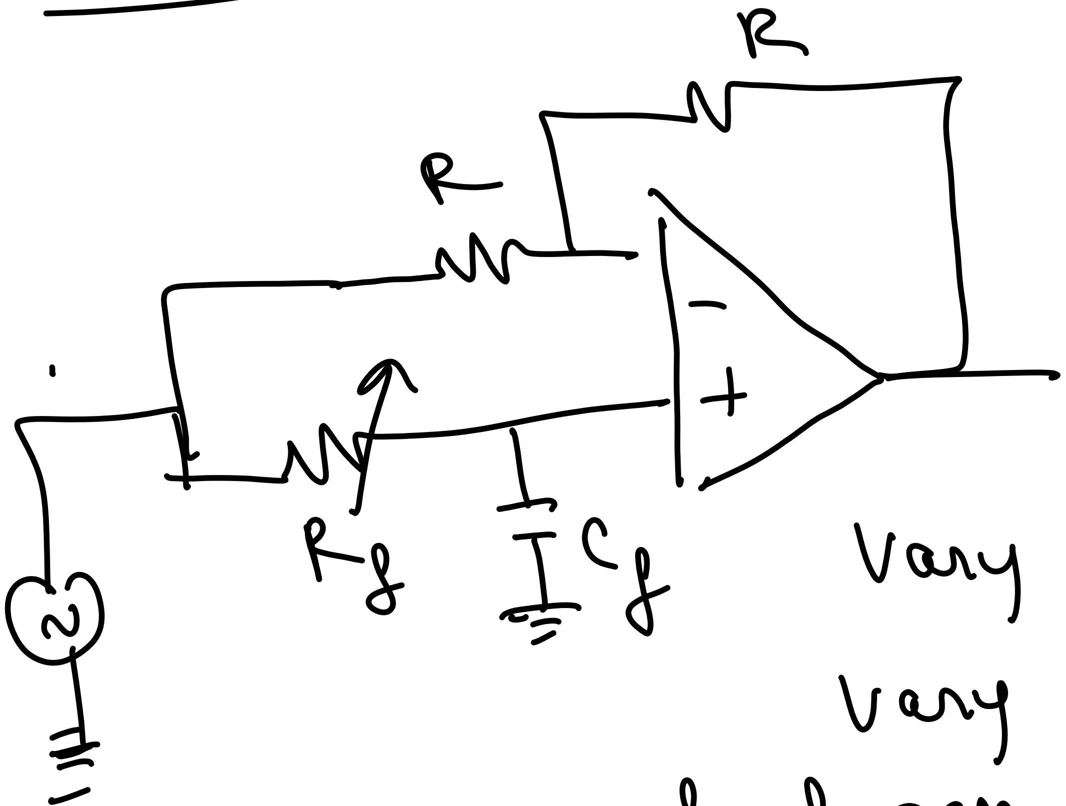
Now the bridge is sensitive for capacitance measurement



So using Lock in amp

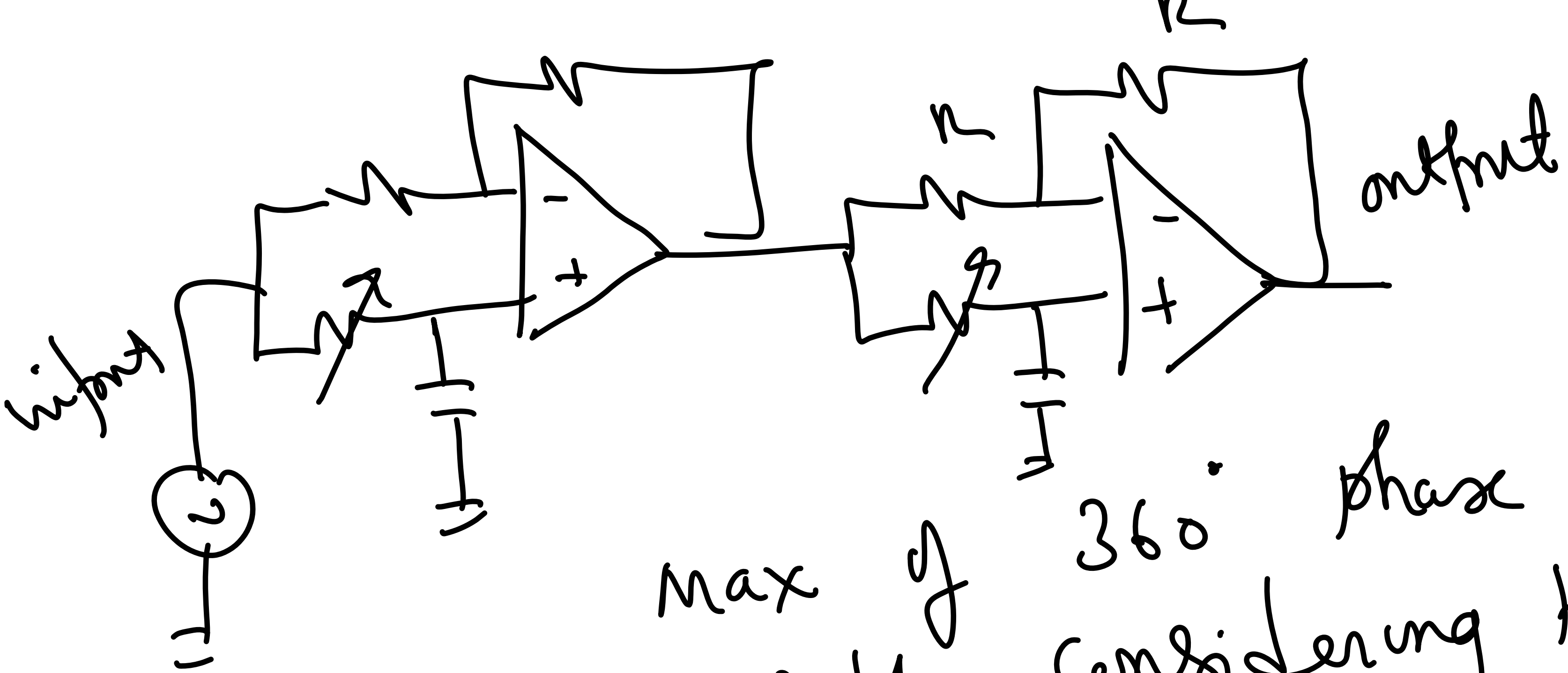
1. Separate out the in phase and out of phase compt
2. Remove the noise from the signal

Phase Shifter circuit

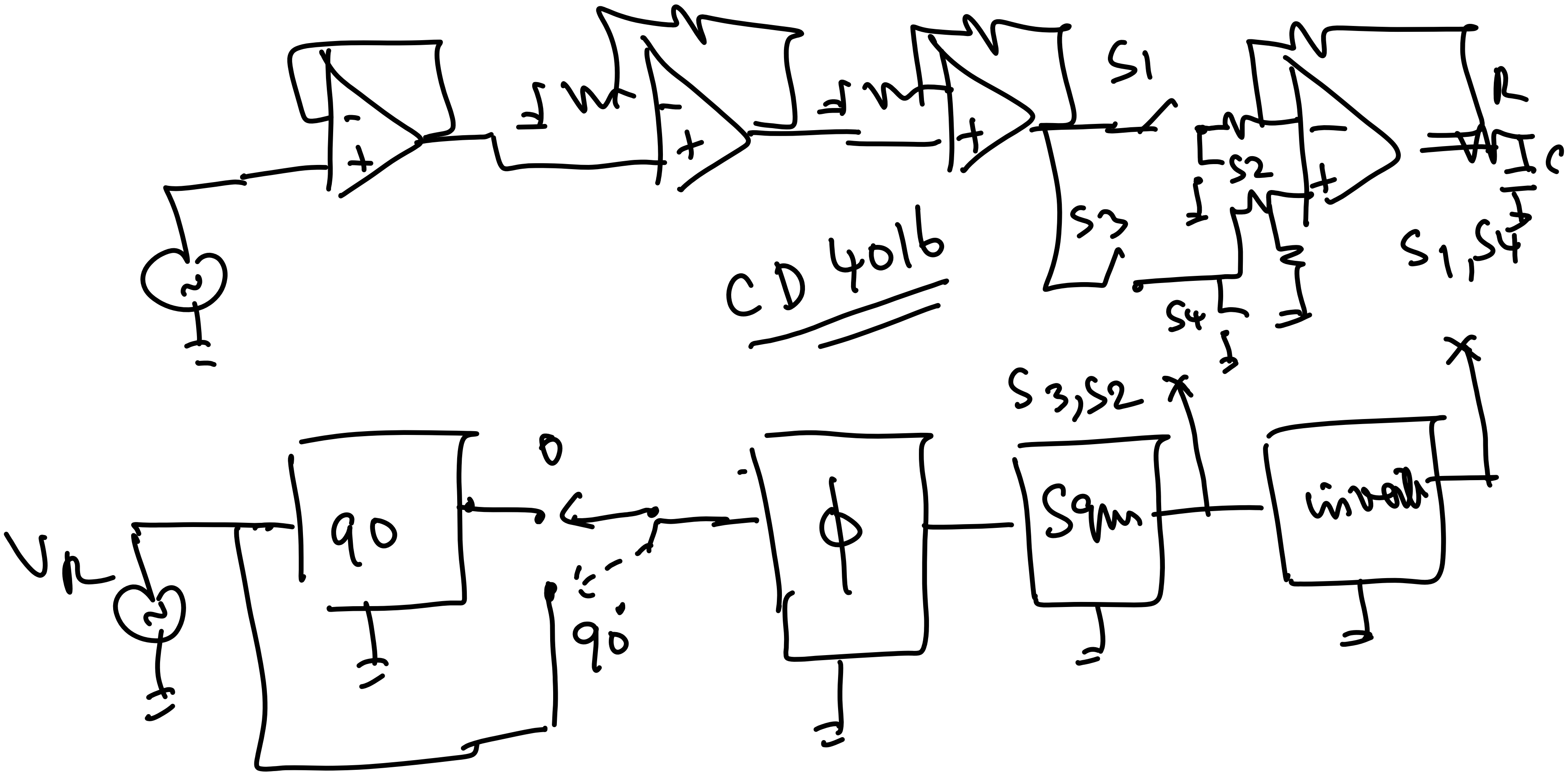


Vary the R_f to vary the phase shift between input and output.

For 90° phase shift select suitable R_f and C_f



Max of 360° phase shift possible considering both the stages


S1
R
C
S2
S3
S4
S1, S4
CD 4016
S3, S2
Vr
90
0
90°
ϕ
Sqm
invert

Lecture No: 33

Capacitive sensors

1. Small in size
2. low power consumption

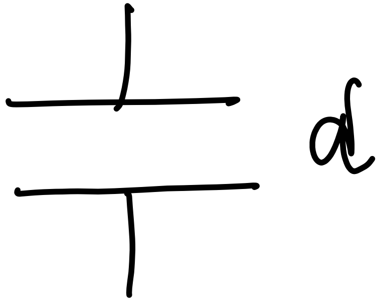
Disadvantge

Noisy

Signal conditioning is difficult

Problems

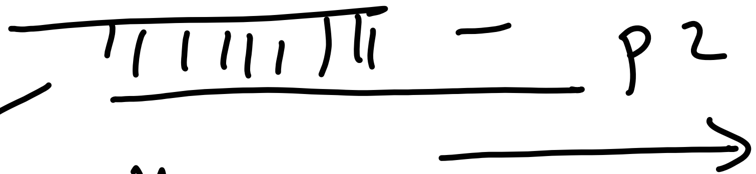
$$C = \frac{\varepsilon_0 \varepsilon_r A}{d}$$



Area

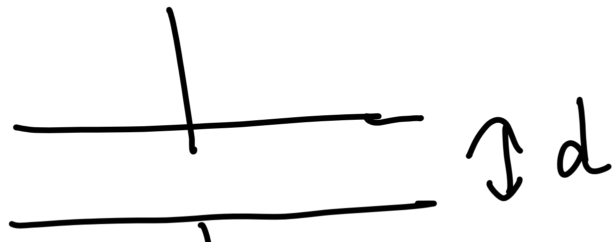
distance between the two plates

① Area variation



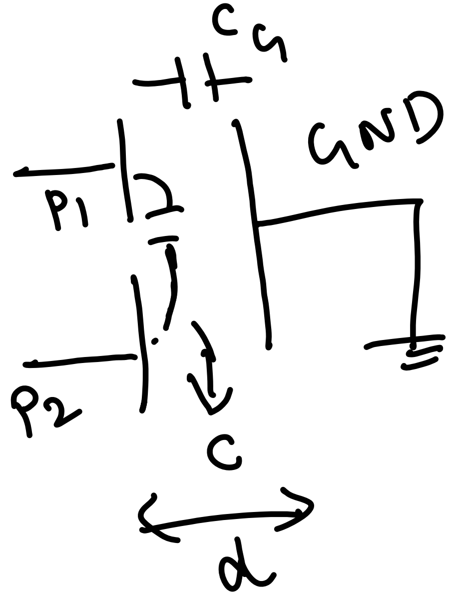
effective area changes with displacement

② Space variation



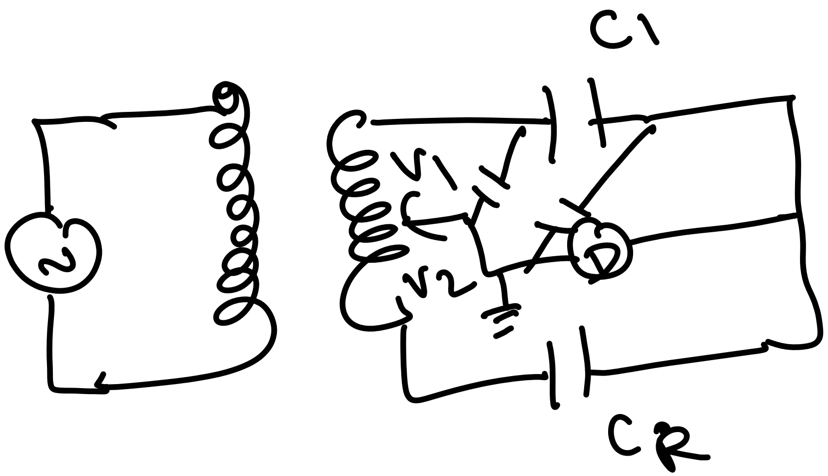
distance between the two plates varied and A is constant

③ proximity variation



Here grounded electrode is moved and capacitance between the two fixed plates are measured.

1. Using ratio transformer technique
2. Diode pump circuit
3. How to use of amps ?



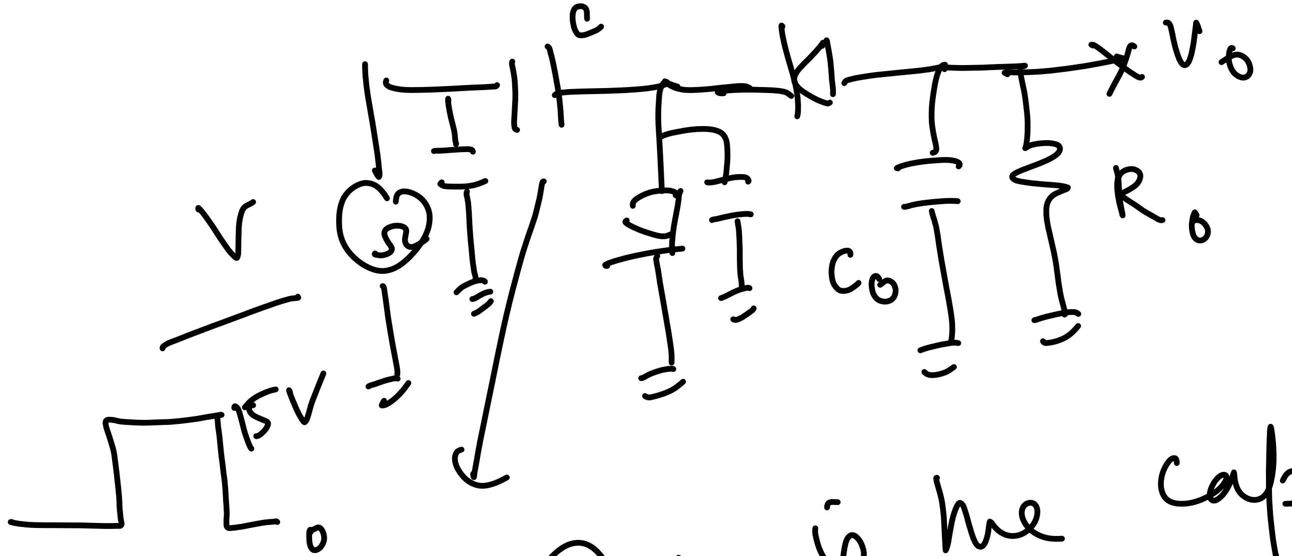
Ratio transformer bridge

$$V_1 C_1 \omega = V_2 C_2 \omega$$

$$\frac{V_1}{V_2} = \frac{C_2}{C_1}$$

Stray cap → no role to play in the balancing of the bridge

So balacing to be done manually. This is the main draw back for instrumentation work.



Diode pump circuit

This is the cap under measurement

If f is the frequency of the source of vol V.

Total charge transfered for se

$Q = CVf$

This charge is delivered
to the output capacitor
If output vol is V_o

The amount of charge
leaking through R_o $= \frac{V_o}{R_o}$

$=$ In coming charge $= CVf$

$$CVf = \frac{V_o}{R_o}$$

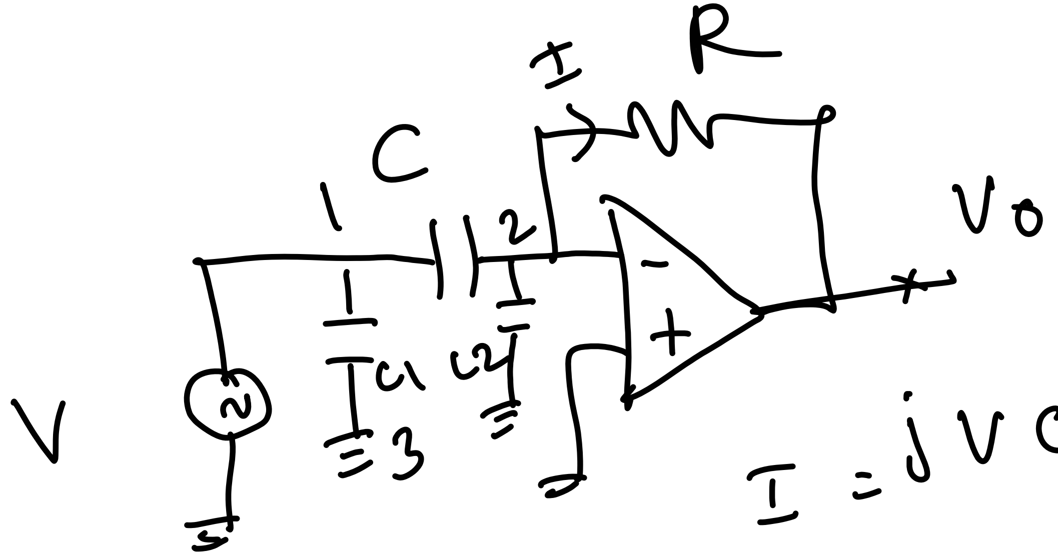
$$CVR_of = V_o \qquad C = \frac{V_o}{VR_of}$$

This is valid only if
V_o is small compared to V

So keep V_o with in few mV by selecting R_o

It is good for 1% to 2% accuracy level.

③ <u>using op amp</u>



$I = jVC\omega$

$V_o = jVC\omega R$

$V_0 = VC\omega R$

(1) Output vol is linear to C

Always consider cap as a 3 terminal device. This brings two stray caps.

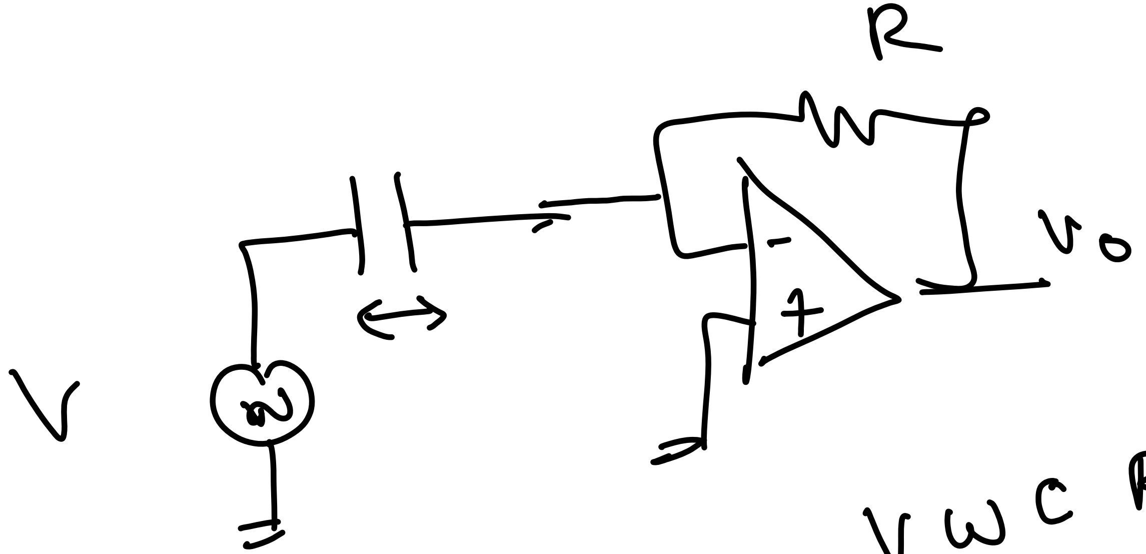
(2) C_1 and C_2 will not produce any charge in V_0

It is valid provided

(1) Source impedance zero

(2) virtual ground is at zero

Displacement measurement



$$V_o = V \omega C R$$
$$= K V R \omega \left[\frac{A}{d} \right]$$

d changes V_o varies by $\frac{1}{d}$

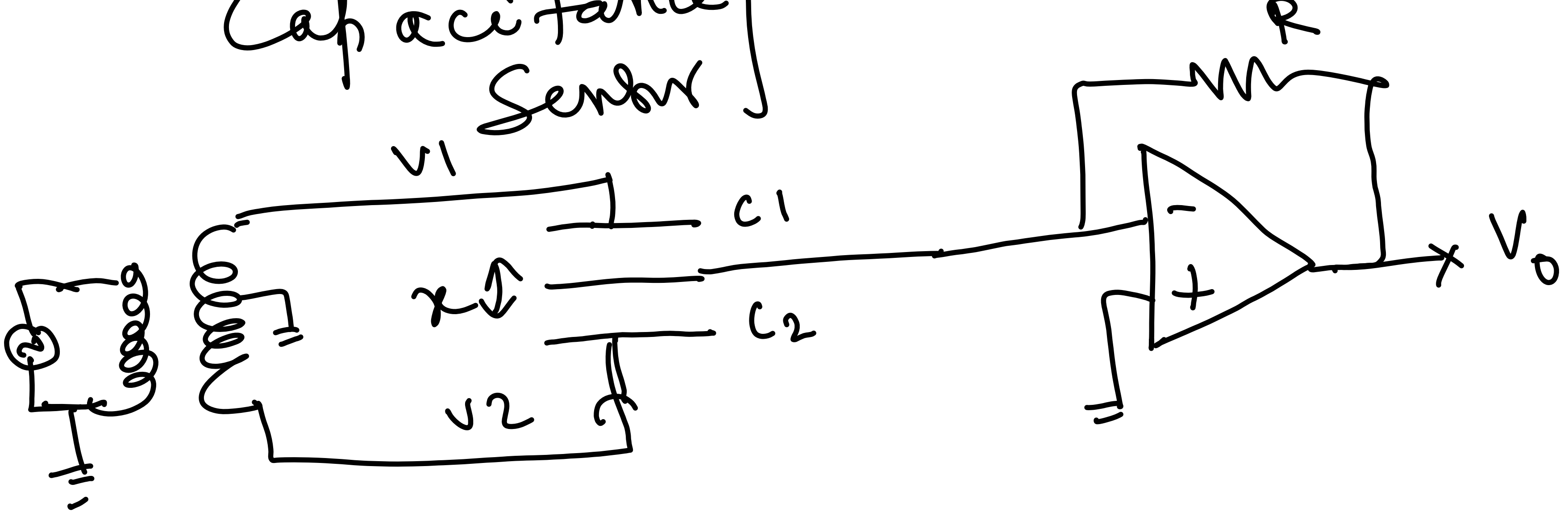
$$V_o = \frac{V \omega R \epsilon_0 \epsilon_r A}{d}$$

This is not as accurate as ratio transformer bridge

The errors are

1. V variations —
2. ϵ_r variations → moisture
3. ω variation

Capacitance Sensor] with three plates



① current through C_1 and C_2 are 180° out of phase

current through C_1 $= j V_1 C_1 \omega$

current through C_2 $= j V_2 C_2 \omega$

$$(V_1 C_1 \omega - V_2 C_2 \omega) R = V_0$$

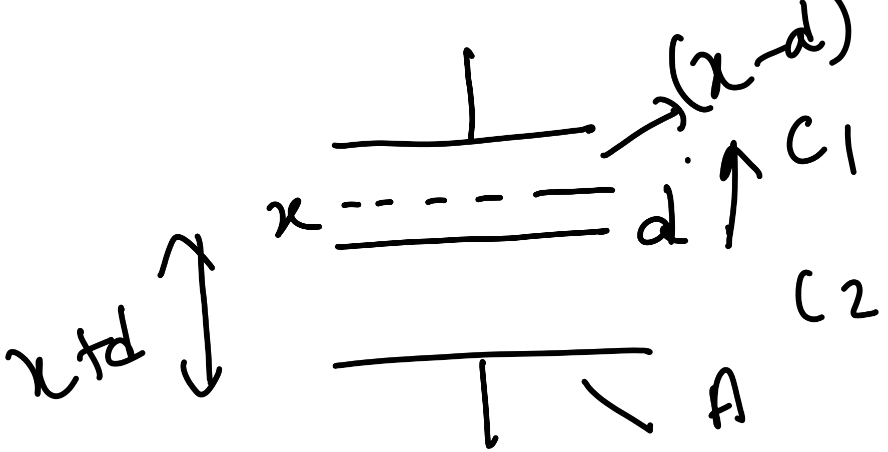
$$\omega R (V_1 C_1 - V_2 C_2) = V_0$$

$$V_1 = V_2 = V$$

$$V \omega R (C_1 - C_2) = V_0$$

① Area of the plates are A

② x is the displacement from the centre distance d



$$C_1 = \epsilon_0 \epsilon_r \frac{A}{d-x}$$

$$C_2 = \epsilon_0 \epsilon_r \frac{A}{d+x}$$

$$(C_1 - C_2) = A \epsilon_0 \epsilon_r \left(\frac{1}{d-x} - \frac{1}{d+x} \right)$$

$$= A \epsilon_0 \epsilon_r \left(\frac{d+x - (d-x)}{d^2 - x^2} \right)$$

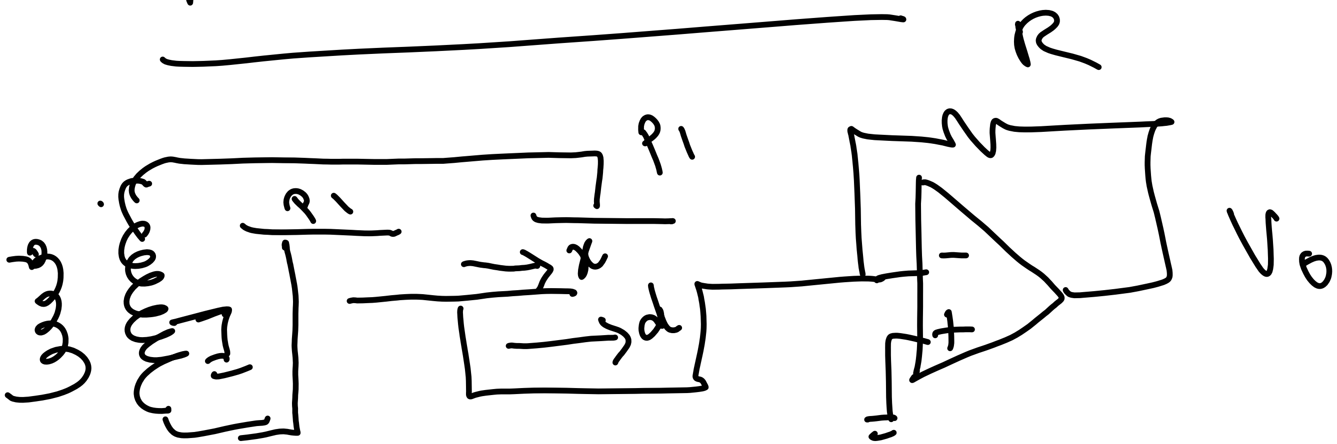
$$(C_1 - C_2) = A \epsilon_0 \epsilon_r \left(\frac{2x}{d^2 - x^2} \right)$$

$$V_0 = V \omega A \epsilon_0 \epsilon_r \left(\frac{2x}{d^2 - x^2} \right)$$

For $x << d$

$$V_0 \propto x$$

Area variation



For a displacement x
area under one plate is
decreasing and equal area
is increasing in the other plate

$$A(l+x)$$
$$A(l-x)$$

$$V_0 = VRW(C_1 - C_2)$$

$$C_1 = \frac{\epsilon_0 \epsilon_r A(l+x)}{d}$$

$$C_2 = \frac{\epsilon_0 \epsilon_r A(l-x)}{d}$$

$$(C_1 - C_2) = \frac{\epsilon_0 \epsilon_r A}{d}\left[(1+x) - (1-x)\right]$$

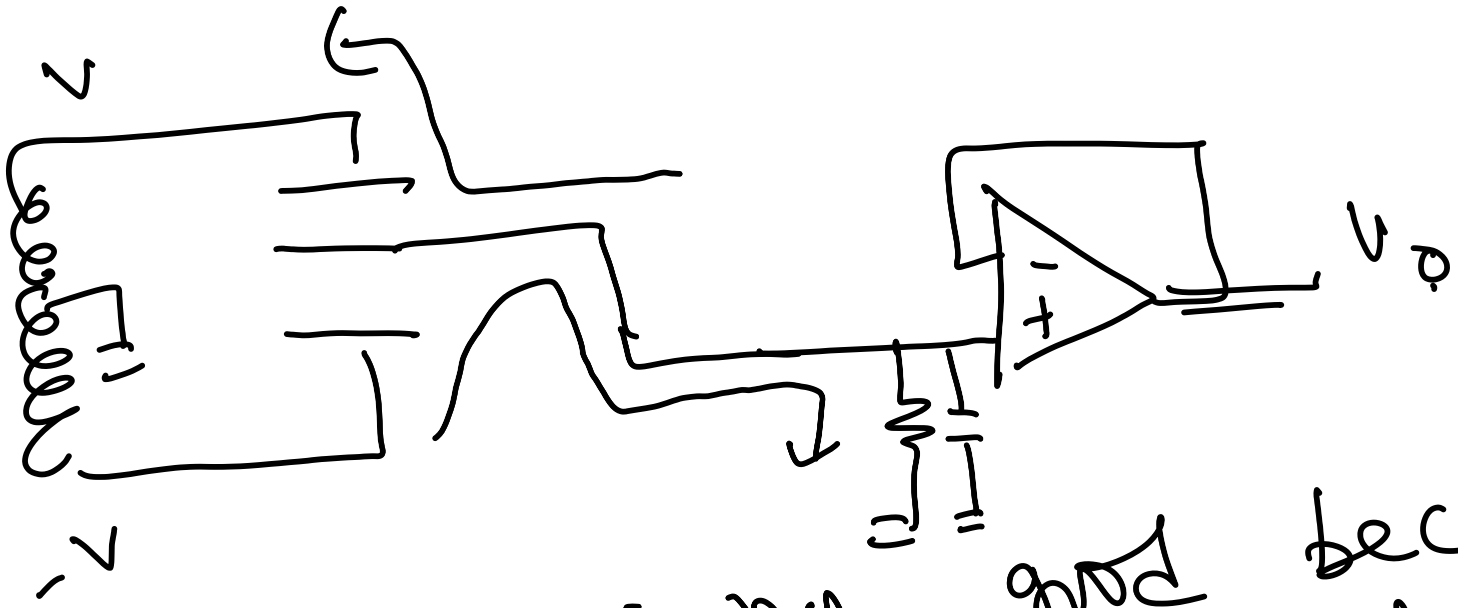
$$= \frac{\epsilon_0 \epsilon_r A}{d}\left[2x\right]$$

$$V_0 = \frac{V R \omega \, \epsilon_0 \epsilon_r A}{d}\left[2x\right]$$

Output voltage is linear for displacement x.

This is also not as accurate as ratio transformer bridge

Use of non-inverting amp

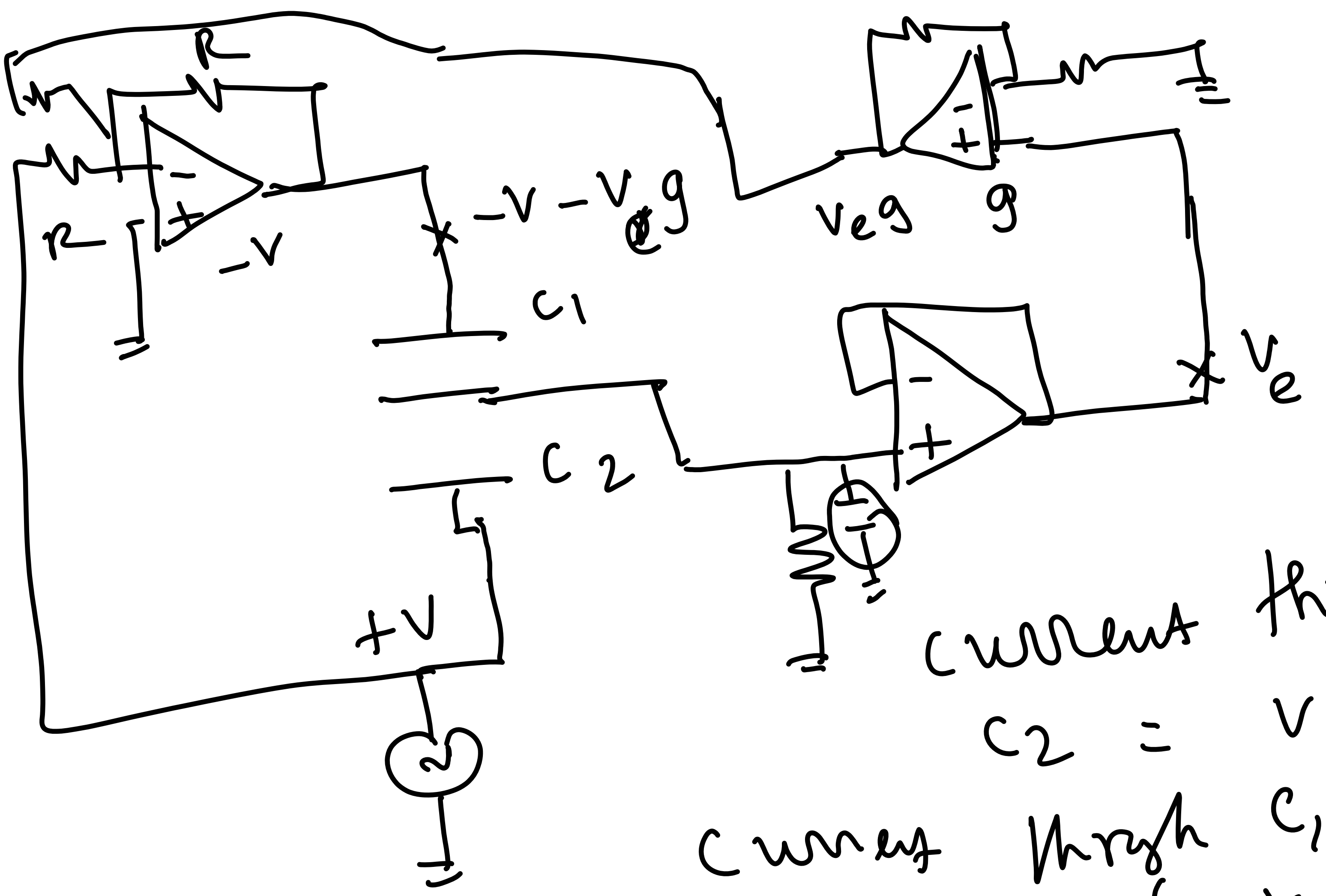


This is not good because it is sensitive to stray cap.

For this method to work

$$\boxed{V_o = 0}$$

How to make
$V_0 = 0$??



current through
$C_2 = V C_2 \omega$

current through $C_1 =$
$= (-V - V_e g) C_1 \omega$

$V C_2 \omega = (-V - V_e g) C_1 \omega$

$$V C_2 = (V + V_{e}) C_1$$

. V_e does the balancing technique

V_e is small for large g

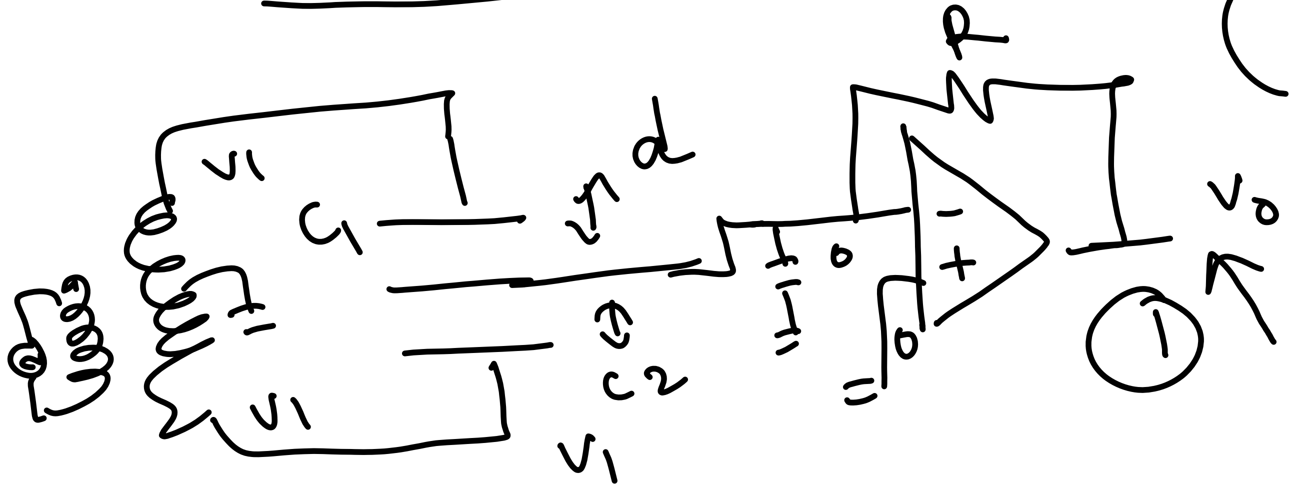
$$C_2 = \epsilon_0 \epsilon_r \frac{A}{d+x} \qquad C_1 = \epsilon_0 \epsilon_r \frac{A}{d-x}$$

Non-inverting amp also can be used

Linear or non linear for x??

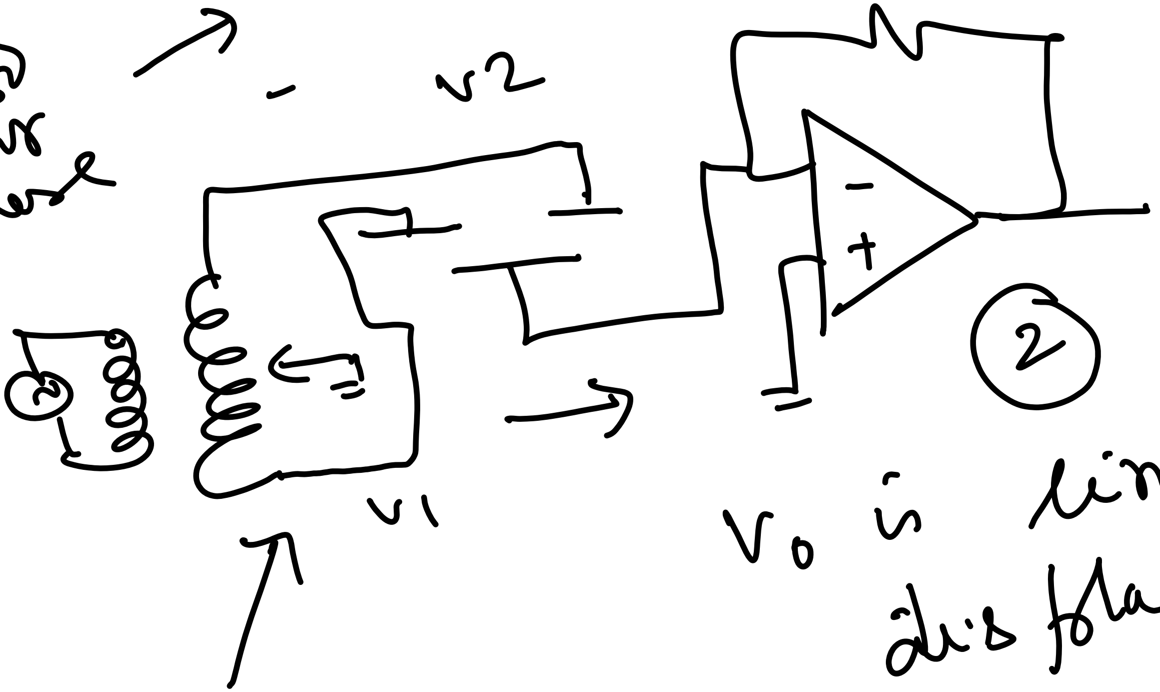
Lec No: 34

Output is not linear with 'd'



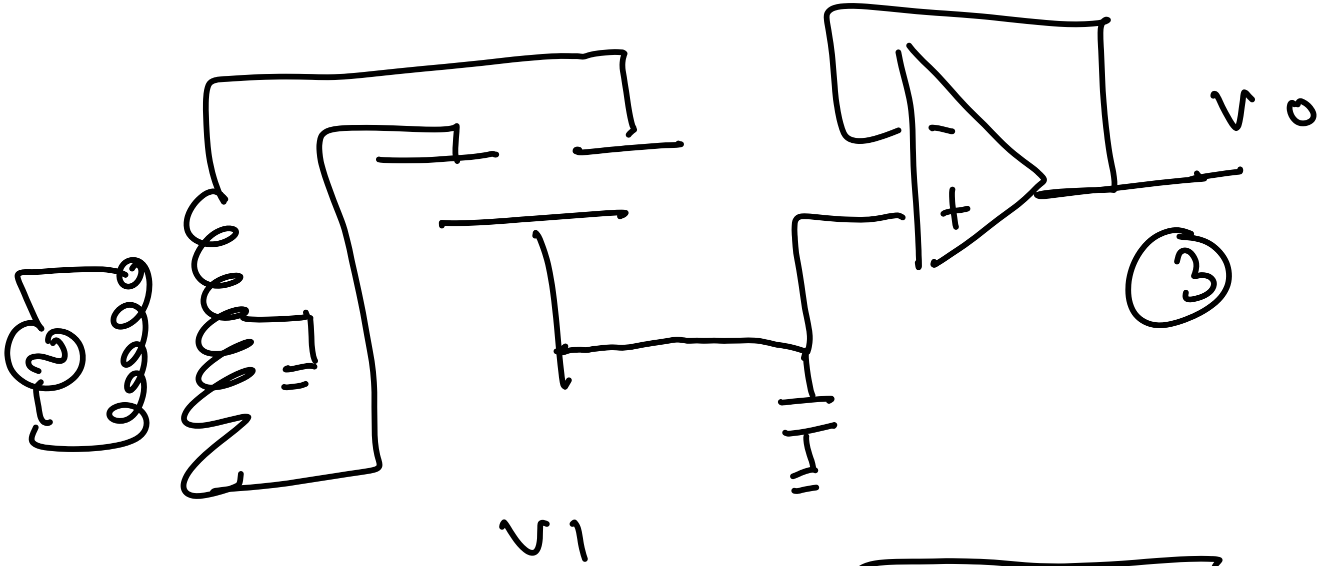
Low Z
with space
variation

180 out of phase



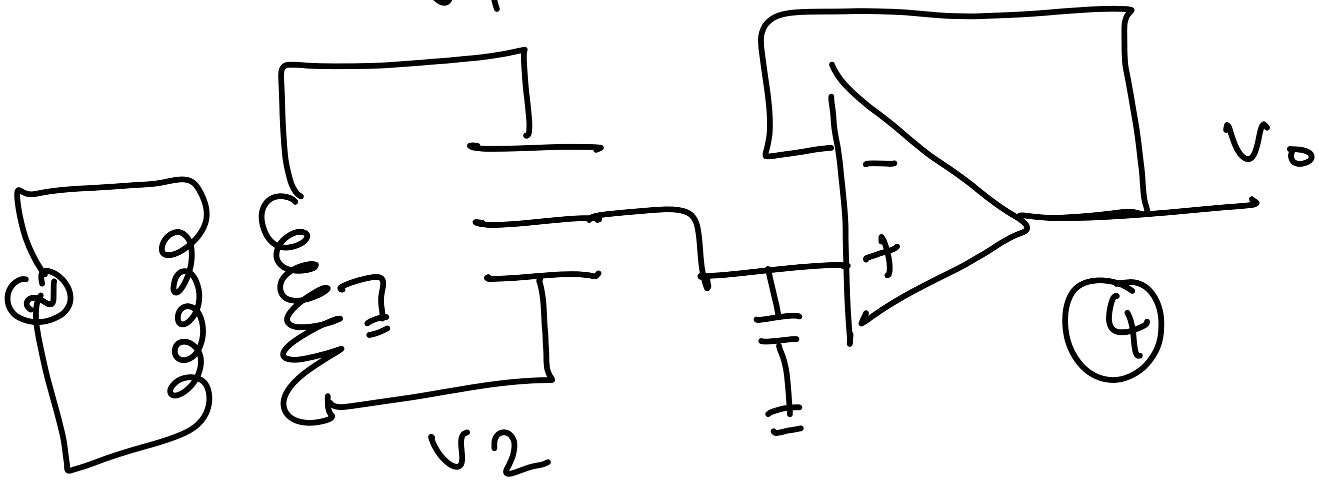
Low Z
with area
variation

V_0 is linear with
displacement



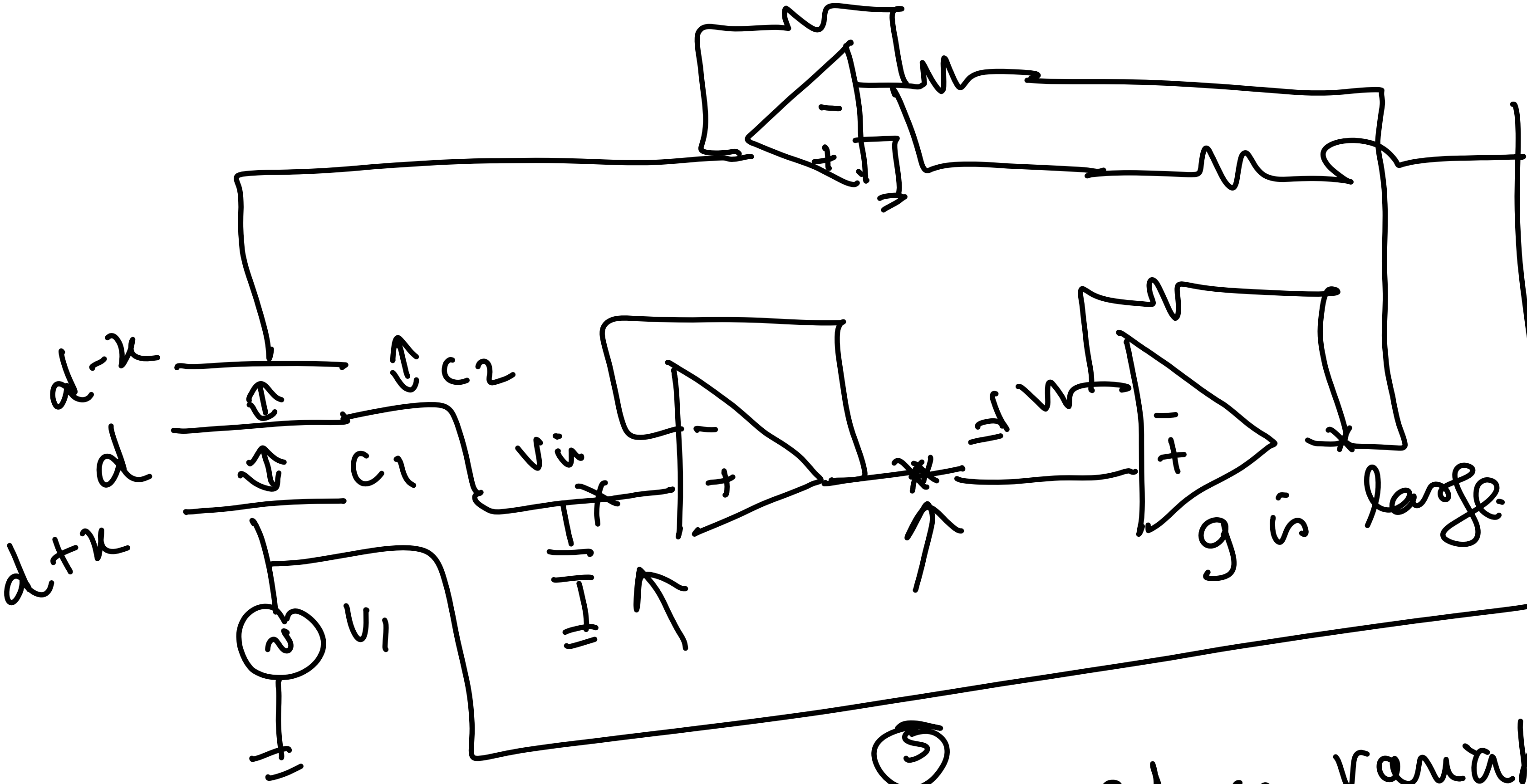
3

High Z
area variation



4

High Z
Space variation



5

High 2 amp space variation with feed back

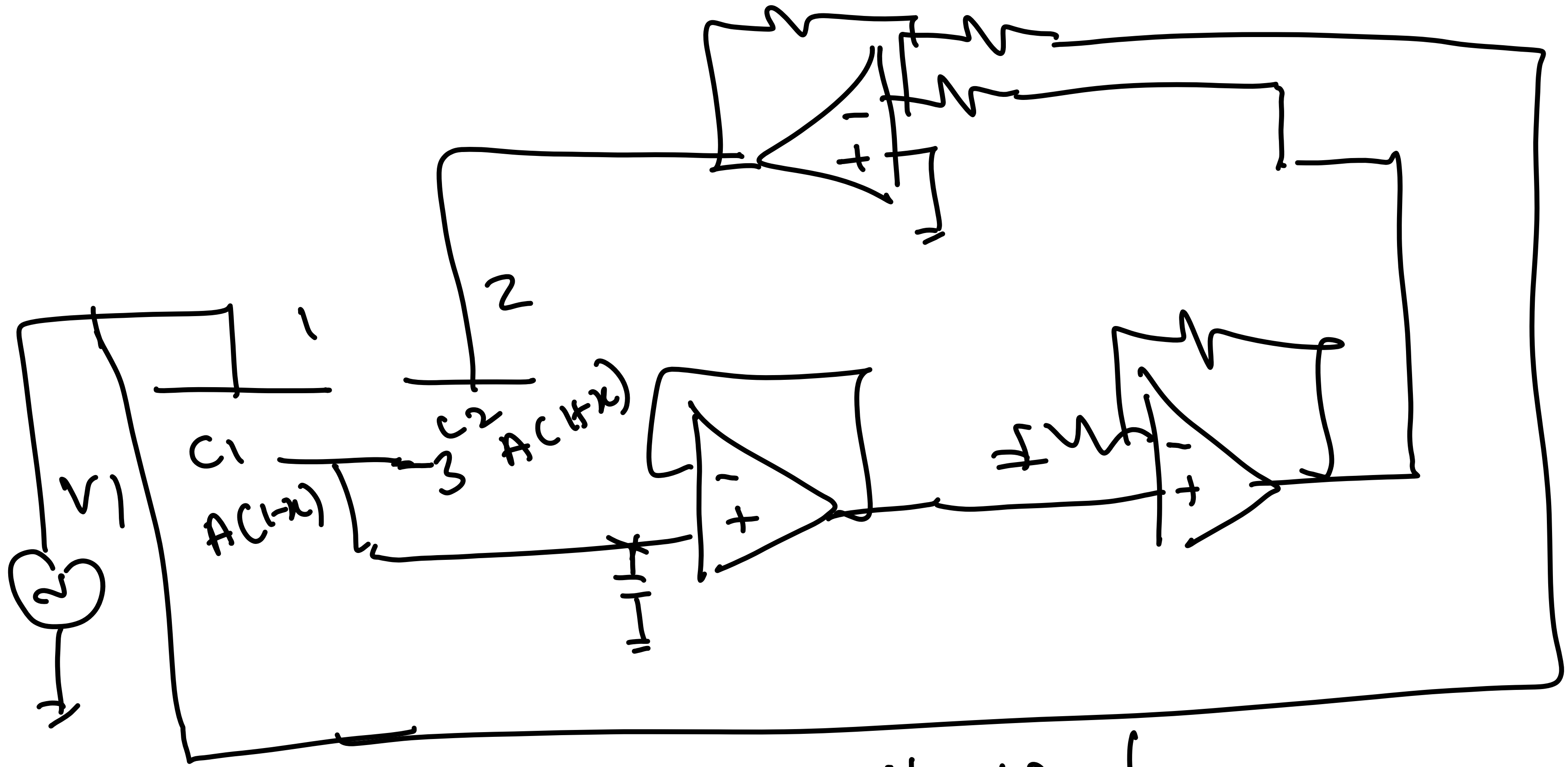
If $V_{in} \simeq 0$

$V_1 C_1 \omega = V_2 C_2 \omega$

$$\frac{V_1}{V_2} = \frac{C_2}{C_1}$$

$$= \frac{A}{d-x} \Big/ \frac{A}{d+x}$$

$$\frac{V_1}{V_2} = \frac{d+x}{d-x}$$



Circuit no. 6

High Z amplifier with area variation
with feed back $V_1 C_1 \omega = V_2 C_2 \omega$

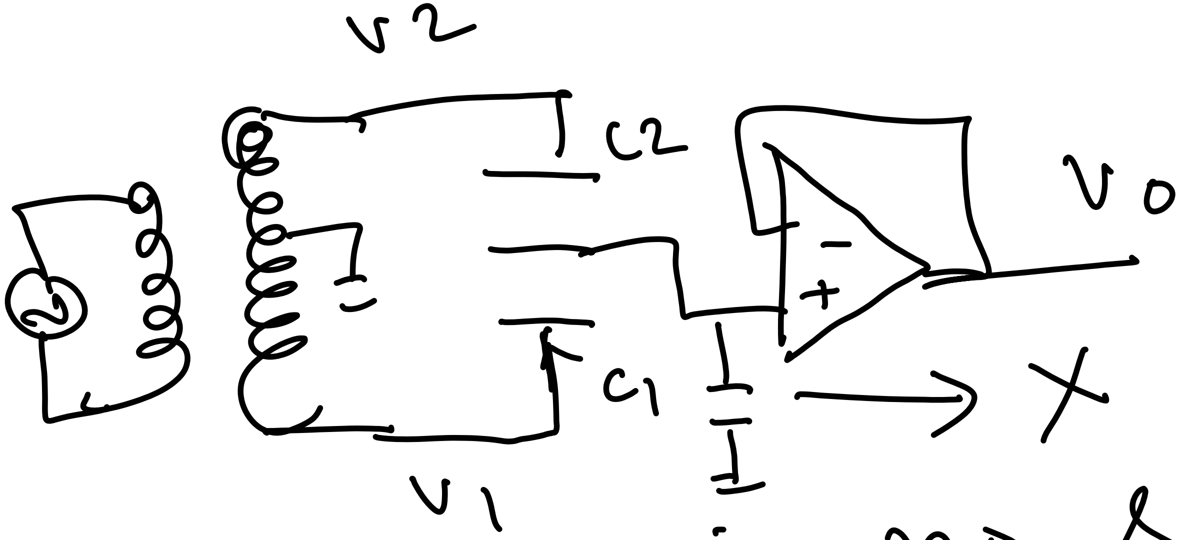
$$\frac{V_1}{V_2} = \frac{C_2}{C_1} = \frac{A(1+x)}{A(1-x)}$$

non linear with displacement

circuit 5 and circuit 6

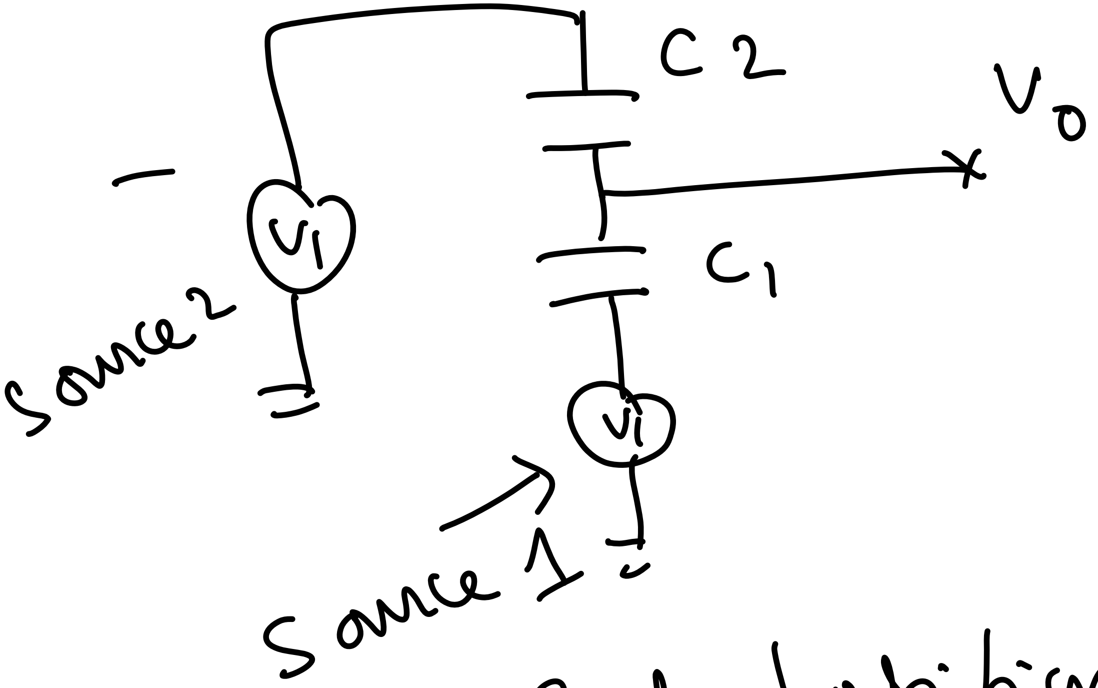
Feed back keeps the non-inverting input of the first amplifier near to ground potential

This makes the stray caps are in effective



① Assuming no stray cap at the input of the op amp

② Assuming $v_1 = V_2$ and they are 180° out of phase



Using Superposition theorem V_o can be calculated

$$\frac{V_1 \times 2C_2}{2C_1 + 2C_2} \quad \text{I}$$

→ contribution due to V_1 by connecting source 2 ~~V_2~~ to ground

ZC_2 is impedance of cap C_2

ZC_1 is the impedance of cap C_1

II Contribution due to Source 2

$$\rightarrow \frac{V_1 \times ZC_1}{ZC_1 + ZC_2} \quad - \text{II}$$

by grounding Source 1

Adding I and II

$$V_0 = \frac{V_1 \times ZC_2}{ZC_1 + ZC_2} - \frac{V_1\, ZC_1}{ZC_1 + ZC_2}$$

$$V_o = \frac{V_1}{(ZC_1 + ZC_2)} \left[ZC_2 - ZC_1 \right]$$

$$ZC_1 = \frac{j\omega A}{d+x}$$

$$ZC_2 = \frac{j\omega A}{d-x}$$

$$\frac{ZC_2 - ZC_1}{ZC_1 + ZC_2} = \frac{\frac{1}{d-x} - \frac{1}{d+x}}{\frac{1}{d-x} + \frac{1}{d+x}}$$

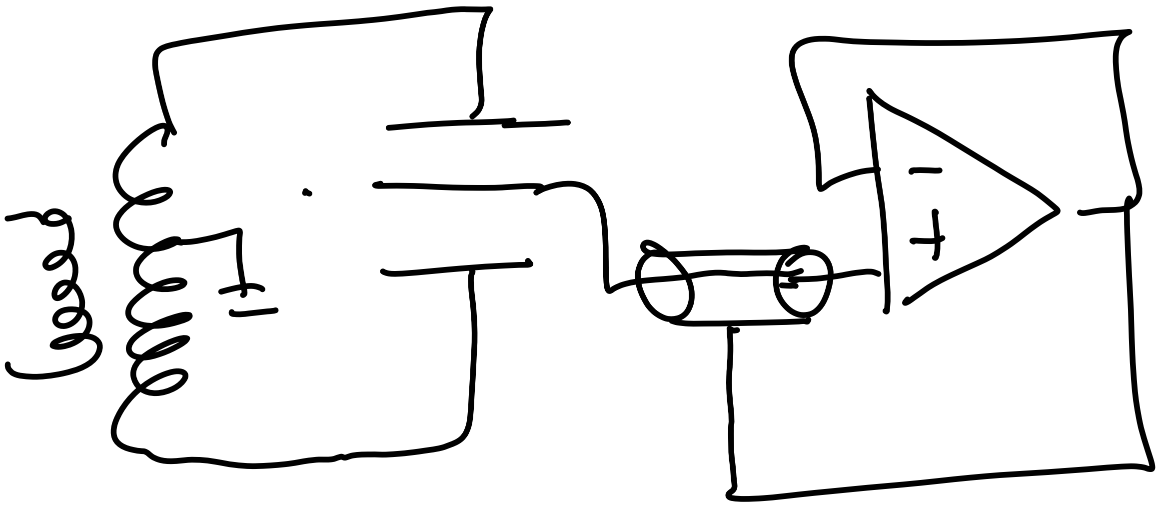
$$= \frac{d+x-(d-x)}{d+x+d-x}$$

$$\frac{2c_2 - 2c_1}{2c_1 + 2c_2} = \frac{2x}{2d} = \frac{x}{d}$$

$$V_0 = V_1 \times \frac{x}{d}$$

d is fixed

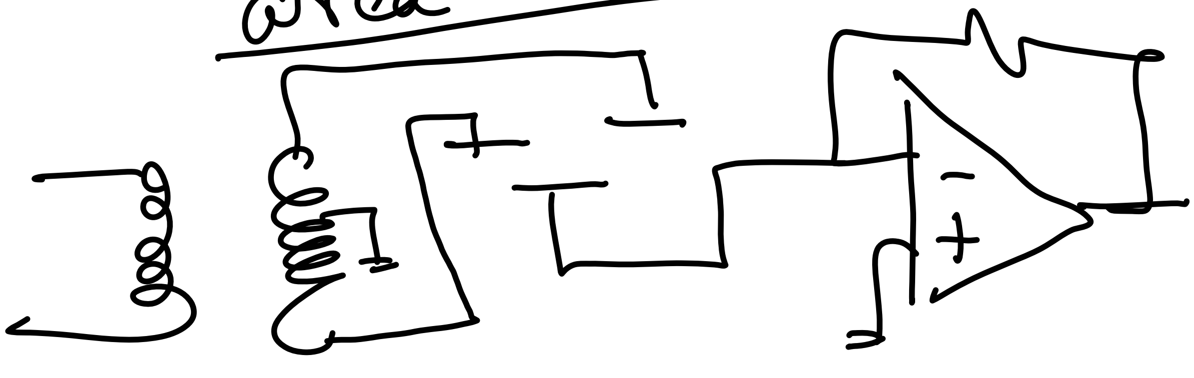
V_1 is fixed



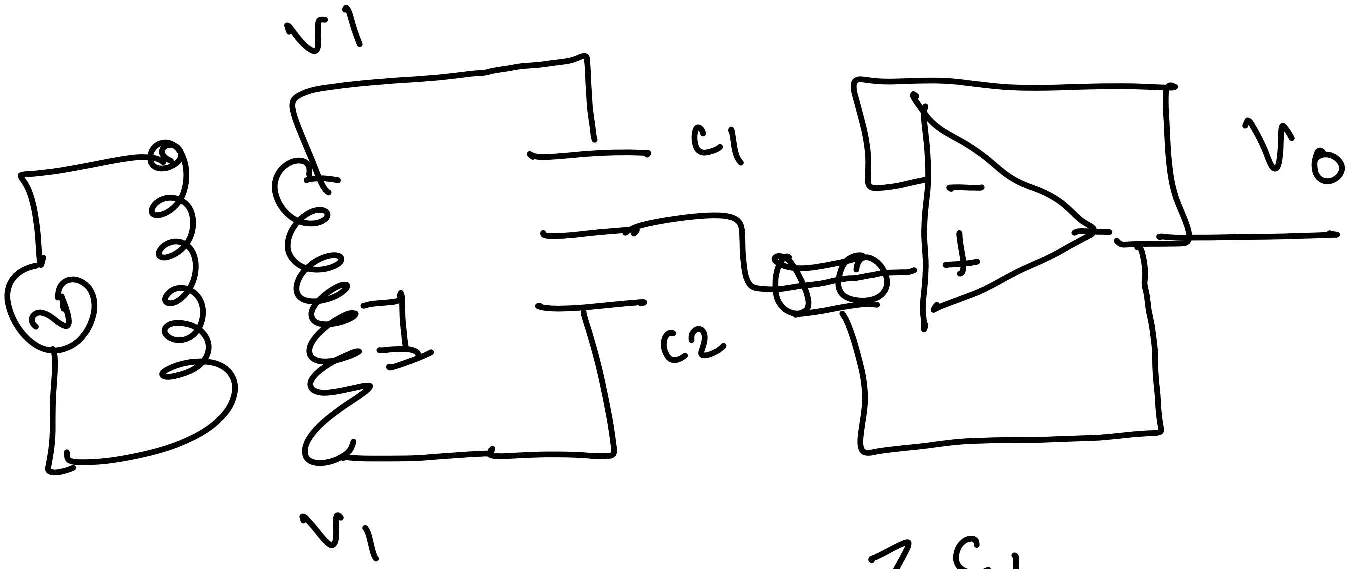
Linear for displacement d

space variation

Low Z amp is linear for area variation



High Z amp for space variation



$$V_o = V_1 \frac{Z C_1}{2C_1 + 2C_2} - \frac{V_1 Z C_2}{2C_1 + 2C_2}$$

$$Z C_1 = \frac{A}{d+x} \times K$$

$$Z C_2 = \frac{A}{d-x} \times K$$

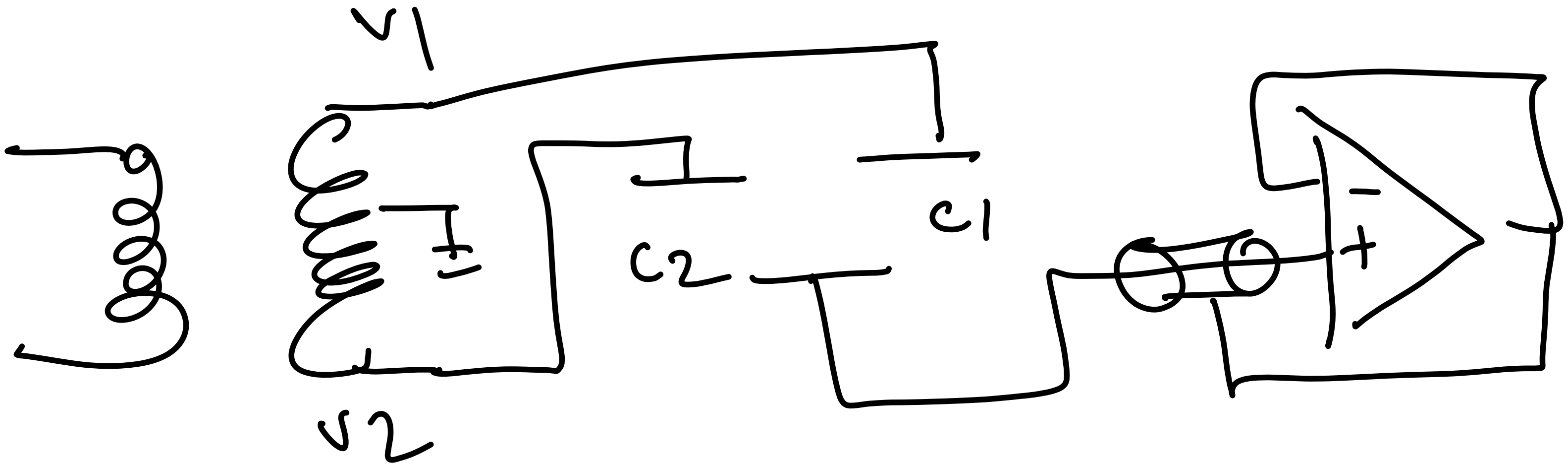
$$2c_1 + 2c_2 = \frac{A}{(d+x)} + \frac{A}{(d-x)}$$

$$2c_1 - 2c_2 = \frac{A}{(d+x)} - \frac{A}{(d-x)}$$

$$\frac{2c_1 - 2c_2}{2c_1 + 2c_2} = \frac{\frac{1}{d+x} - \frac{1}{d-x}}{\frac{1}{d+x} + \frac{1}{d-x}}$$

This $\underline{\text{is}}$ already done

High 2 for area variation



$$V_2 \frac{Z C_1}{Z C_1 + Z C_2} - V_1 \frac{Z C_2}{Z C_1 + Z C_2} = V_0$$

$$Z C_1 = \frac{A(1+x)}{d}$$

$$Z C_2 = \frac{A(1-x)}{d}$$

$$2C_1 - 2C_2 = \frac{A}{d}\left[1+x-(1-x)\right]$$

$$2C_1 + 2C_2 = \frac{A}{d}\left[(1+x)+(1-x)\right]$$

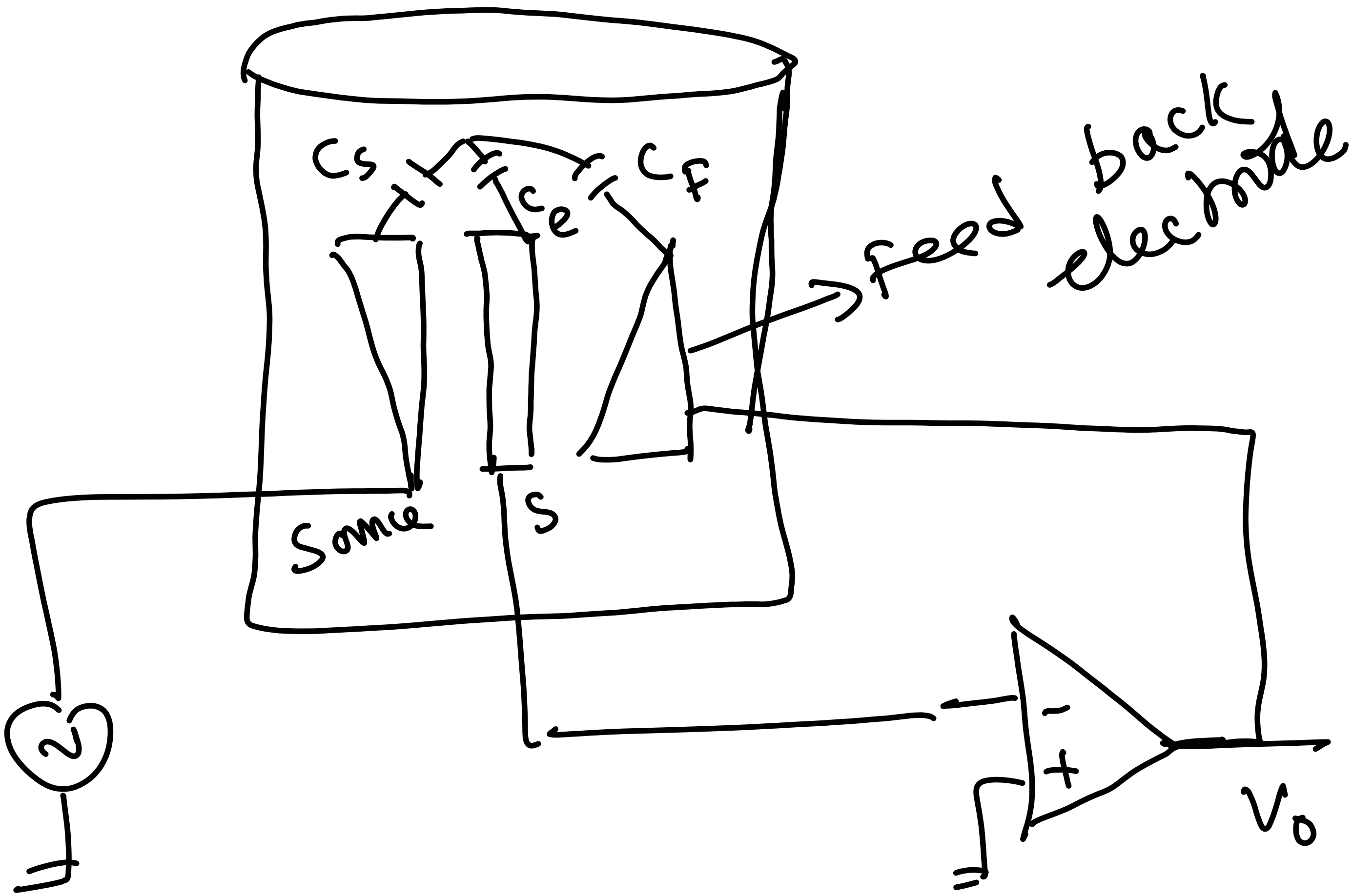
$$\frac{2C_1 - 2C_2}{2C_1 + 2C_2} = \frac{1+x-1+x}{1+x+1-x} = \frac{2x}{2}$$

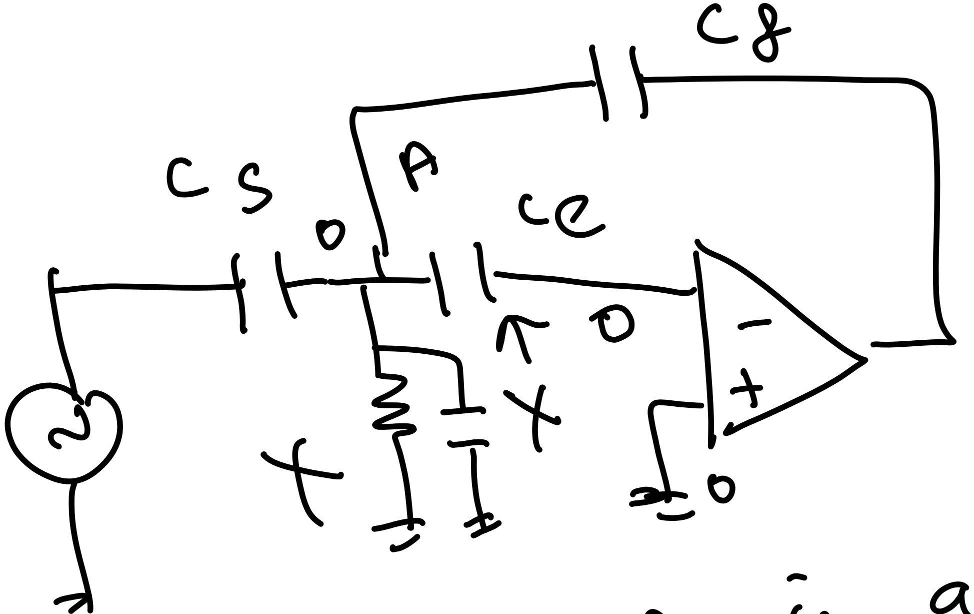
$$= x$$

$$V_1 = V_2$$

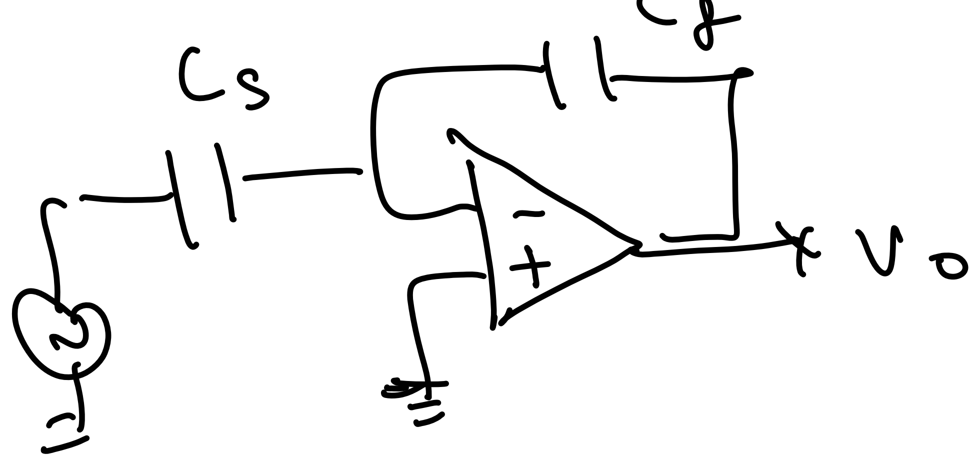
$$V_0 = Vx$$

Water level measurement using Cap. Sensor





point A is at zero potential!
This is because -ve termine is at zero potential and no current through Ce



Both c_f and c_s depend on the water level.
The plates are inverted so with water level v_o varies.

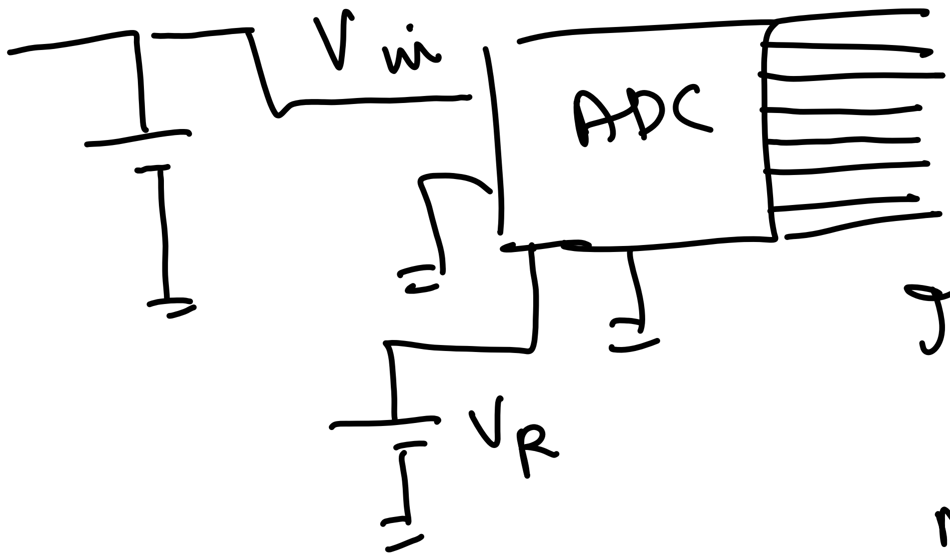
① Best circuit is Ratio transformer bridge.

② op amp with inverting amp is ~~bes~~ next best

Lecture - 35

Analog to Digital Conversion

ADC



8 bit

V_o

If $V_{in} = V_R$

1 1 1 1 1 1 1 1

MSB LSB

For $V_{in} = V_R$

$1V = 1V$

Then MSB 1 1 1 1   1 1 1 1 LSB   All bits are high

For $V = 0$ 0 0 0 0   0 0 0 0   All bits are zero

Total binary level $= 2^8$

$= 256$

1 LSB to change the reference

Vol $= \frac{1}{256} \simeq 4$ mV

Input 4 mV $\longrightarrow$ 0 0 0 0   0 0 0 1

For 1 LSB the required

$$\text{vol} = \frac{V_R}{2^n} = \frac{V_R}{256}$$

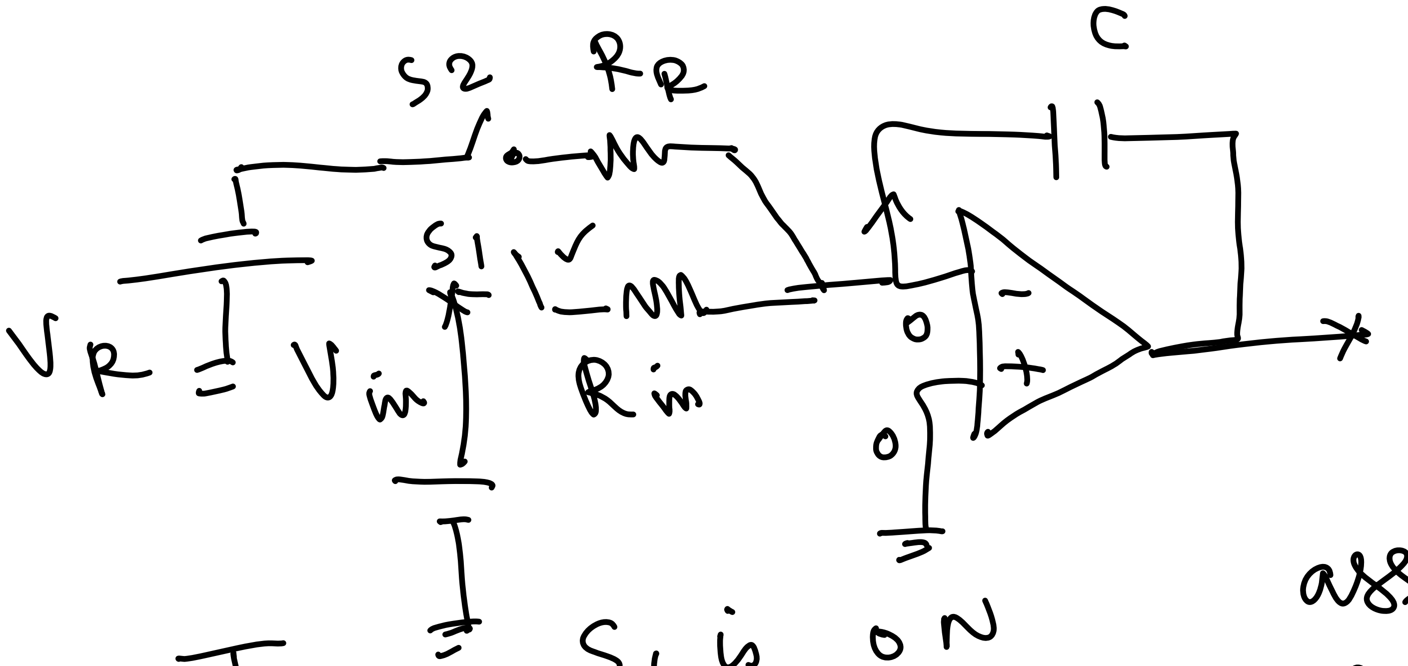
LSB gives us the minimum resolution one can get using this converter.

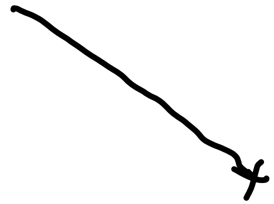
LSB value depends on the reference voltage

Types of Converters

1. Dual Slope ADC
2. Sucessive approximation ADC
3. Flash ADC
4. Σ - D ADC

Dual slope ADC





I S_1 is ON
S_2 is OFF

assuming C is empty

when S_1 is ON
I_{in} flows through the R_{in}
−ve input is at zero w/

$$I_{in} = \frac{V_{in}}{R_{in}} = \text{current through } C$$

ie the input source voltage charges the capacitor C at a constant rate

Capacitor vol will increase with time.

So output vol = Capacitor vol.

capacitor

w/f

$Q = it$

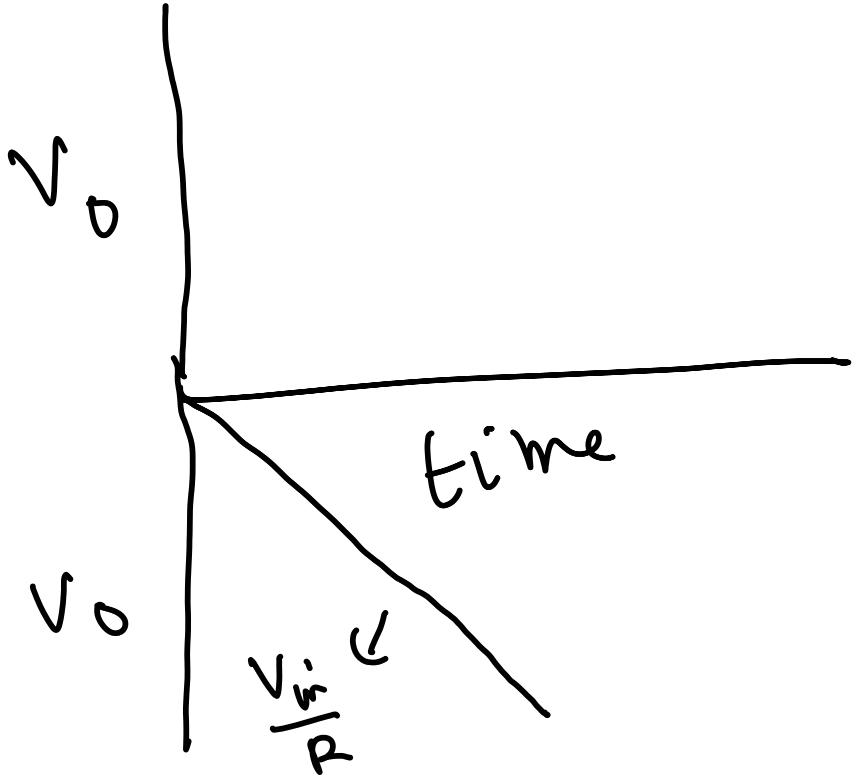
$Q = CV$

$it = CV$

$$V_0 = \frac{it}{C}$$

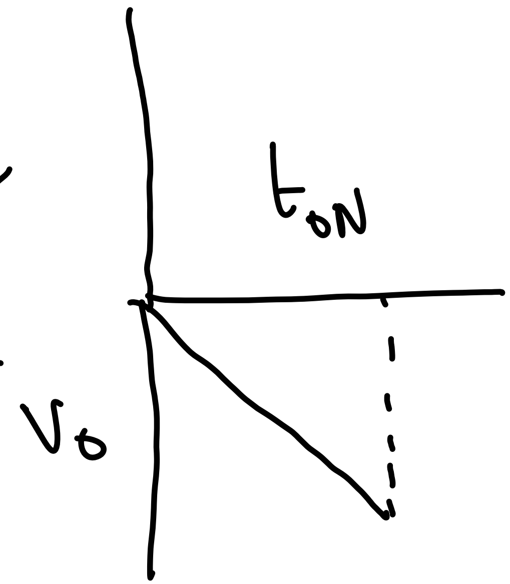
$$i = \frac{V_{in}}{R_{in}}$$

$$V_0 = \frac{V_{in}}{R_{in}} \times \frac{t}{C} \longrightarrow \text{output w/f}$$



Due to Op amp inversion

$$V_o = -\frac{V_{in} \times t}{R_{in}\, C}$$



Keep S_1 ON only for a fixed time t_{on}

After t_{on} time switch OFF S_1

Switch on S_2

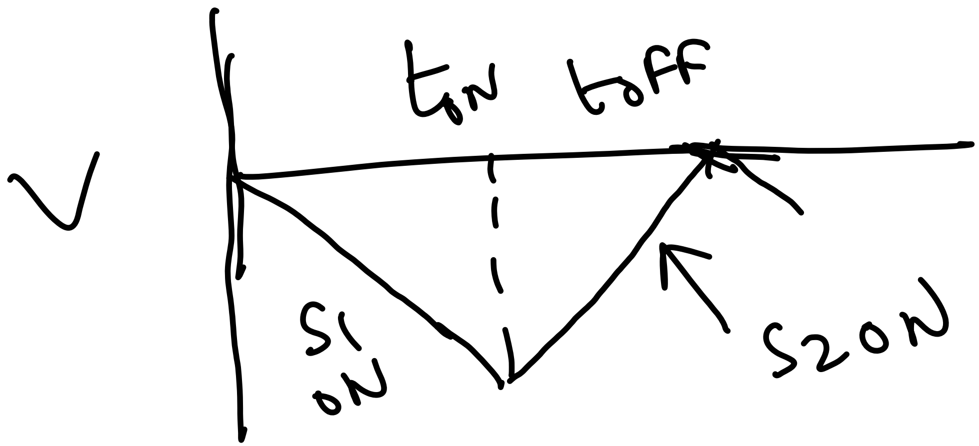
Switch S_2 is connected to V_{Ref} through the Ref resistance R_{Ref}

The current through R_R is opp to that of R_{in}

So capacitor C now discharges

So output vol V_o will start coming down.

Keep S_2 ON so that V_o after some time becomes zero



Discharge equation

$$I_D = \frac{V_R}{R_R}$$

$$it = CV$$

$$\frac{it}{C} = V_o$$

Total charge delivered to C when S_1 ON $= it = \frac{V_{in}\, t_{ON}}{R_{in}}$

Total discharge when S_2 ON $= \frac{V_R\, t_{OFF}}{R_R}$

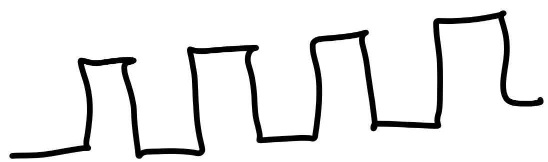
Total input charge = Total output charge

$$\frac{V_{in}}{R_{in}} t_{ON} = \frac{V_R}{R_R} t_{OFF}$$

$$R_{in} = R_R$$

V_R is fixed $\quad \dfrac{V_{in}}{V_R} = \dfrac{t_{OFF}}{t_{ON}} \quad$ fixed

$$V_{in} \propto t_{OFF}$$

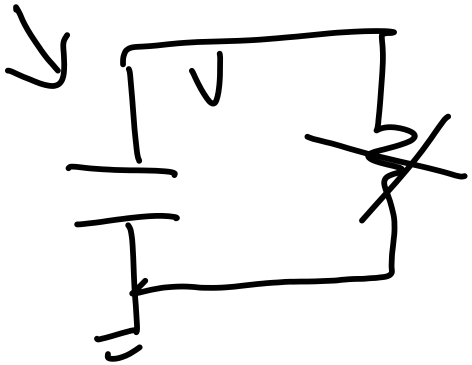


t_{OFF} period can be measured by counting the no. of pulses

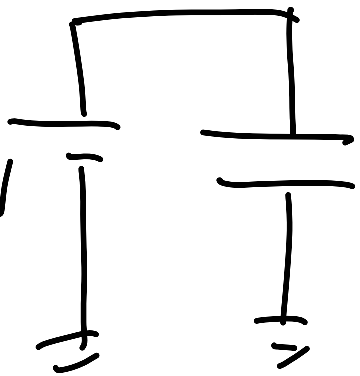
(1) The C is not appearing in the final equation So the long term drift of the capacitor is not a issue

(2) The slow drift of the C due to temp is not a issue. Because tON and tOFF are in milli seconds

③ The dielectric absorption of the C is a major issue



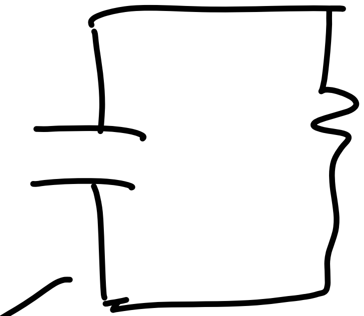
10V



I

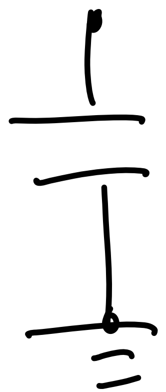
$\rightarrow$

II



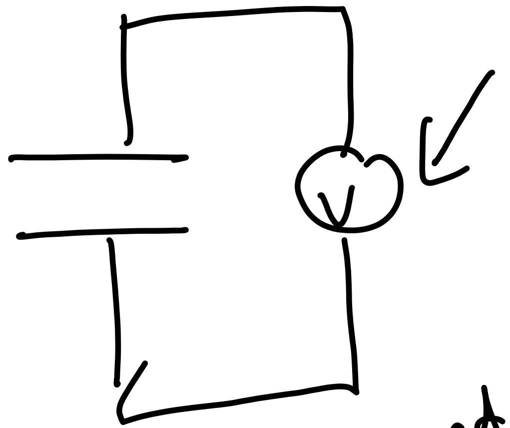
1 or 2 Ω for few secs

Allow the cap to discharge fully



remove the load

III



IV

now connect a high impedance multimeter (volt)

with time V is increasing

ie some of the charge re appears in the cap

This is called di-electric absorption

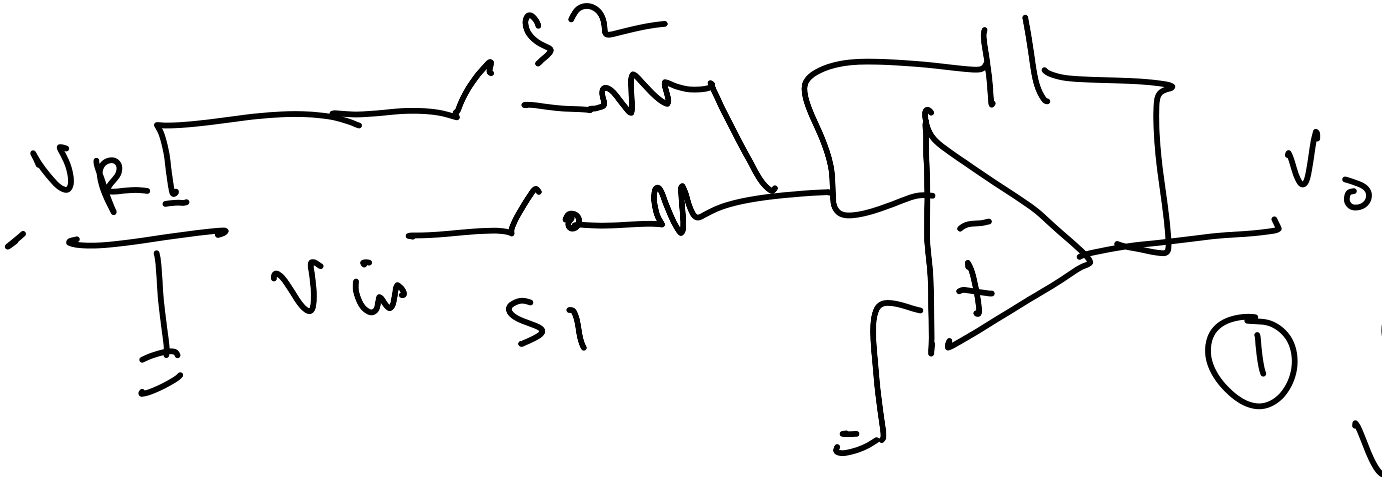
For Electrolytic cap

it is about 10%

For ceramic 1-2%

For plastic film is is less

ie 0.01% to 0.1%



(1) assuming

Vin large

Then c will charge

to high vol

Then S_2 discharges C quickly

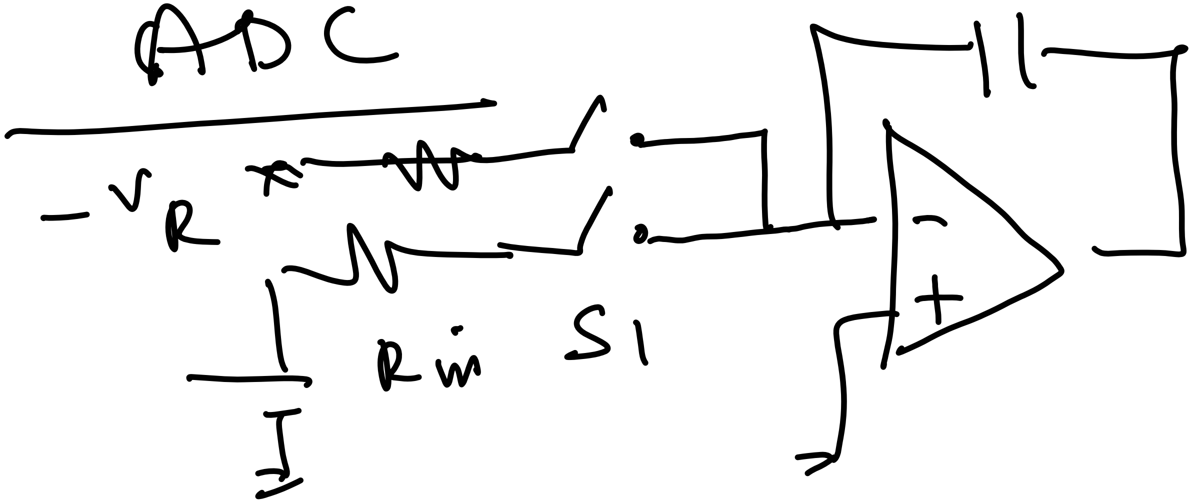
If the input vol is changed to a lower value say from 1V to 10mV or so

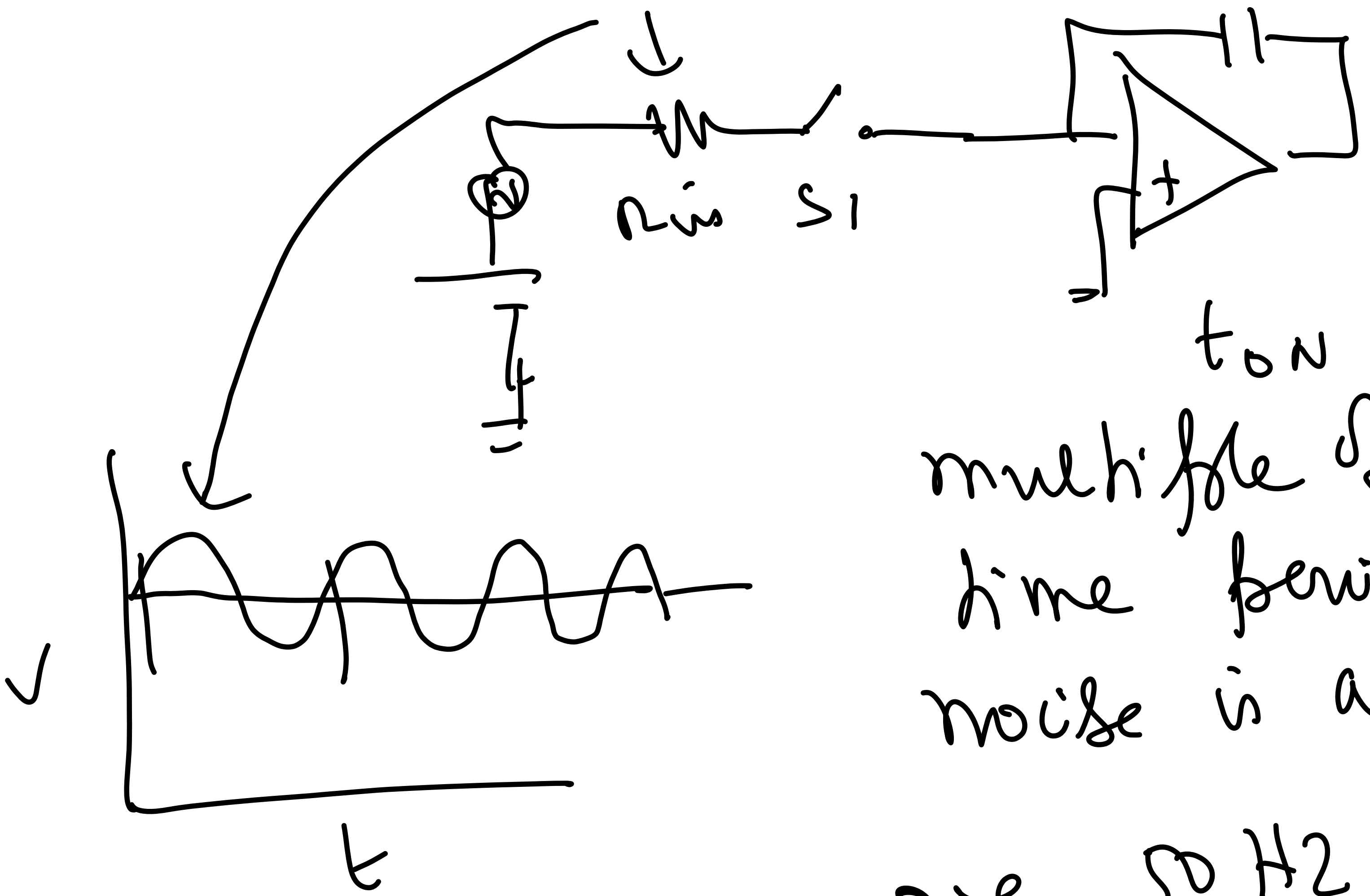
No when S_1 is on this 10mV
will deliver small charge to C
but in addition part of the earliar charge also ~~start~~ re-appear
due to di-electric absorbtion

This will make our measurement to go wrong.

So it is essential that c is having very low di-electric absorption.

§ Noise issues in dual Slope ADC





t_{ON} is integral multiple of noise time period then noise is averaged out

To remove 50 Hz noise

If t_{ON} = 60 ms (3 cycle time)

Then averaging is done for
3 ac cycles
The average charge due to
ac component is zero
The ADC is not sensitive
to input noise.
This is the merit of the
Dual slope ADC

Dis - advantage

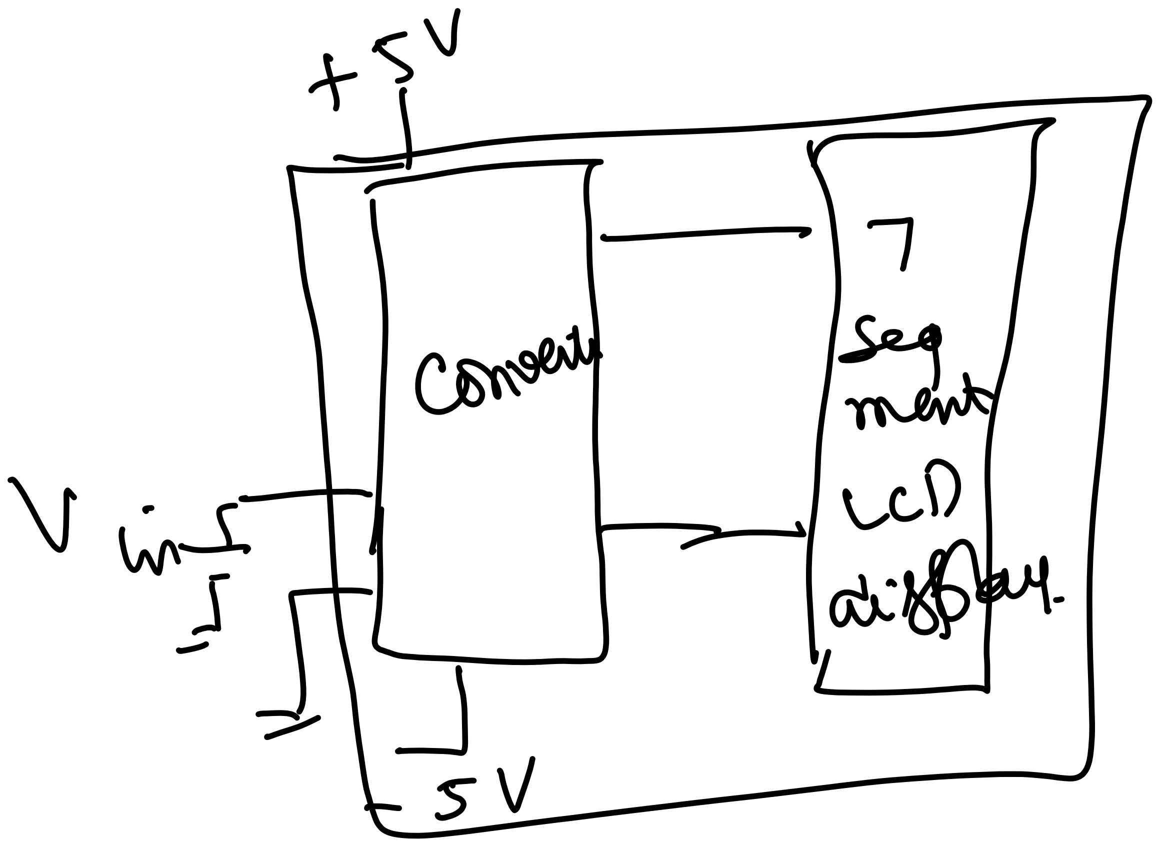
Conversion is very slow

Advantage

In sensitive to noise

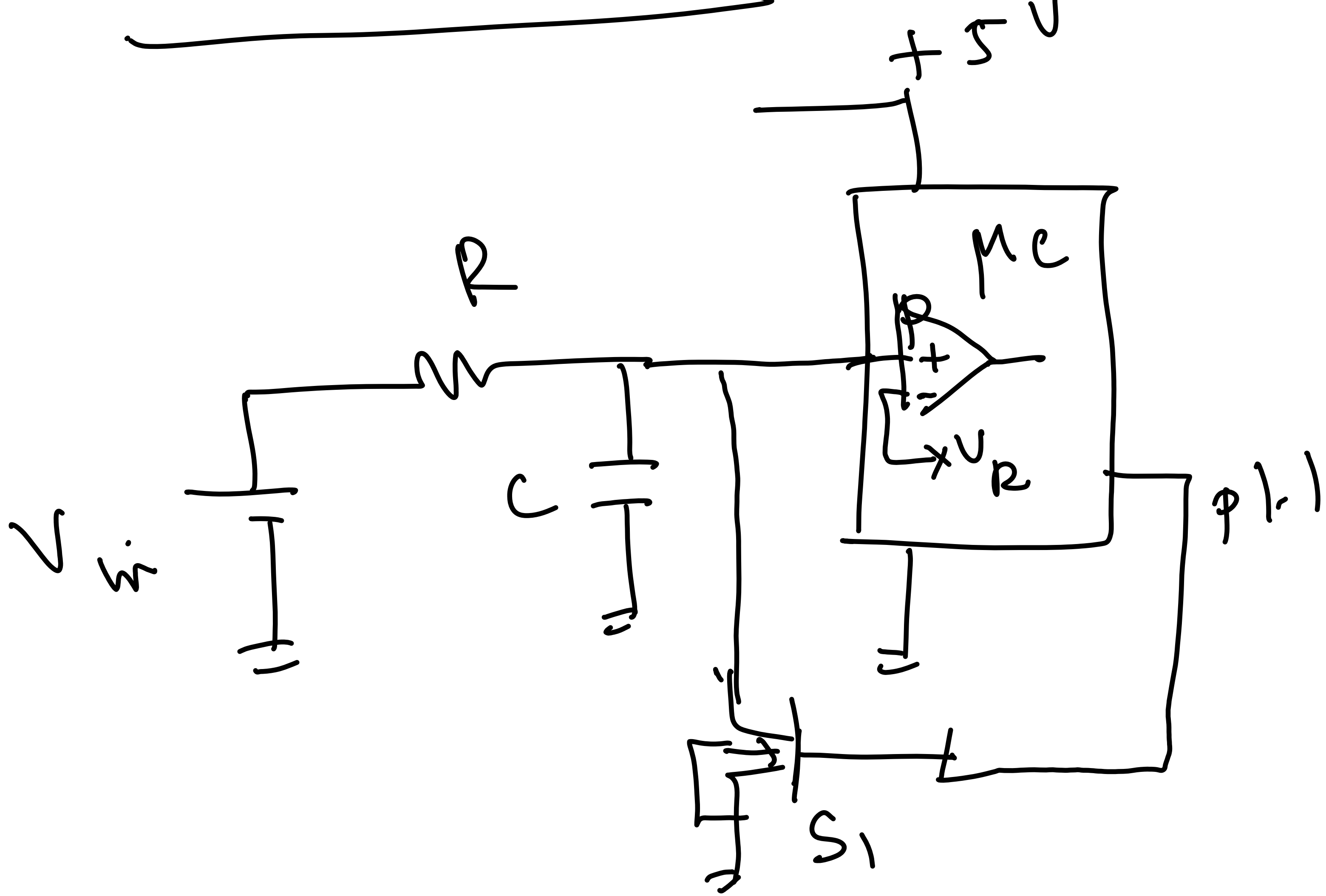
This types of converters are used for LCD display etc. At the max 8 to 10 conversion are possible

All displays use this type of converters.



micro processor / micro controller

based ADC



Discharge C by switching on S_1

Then start the timer in μC and simultaneously switch off S_1

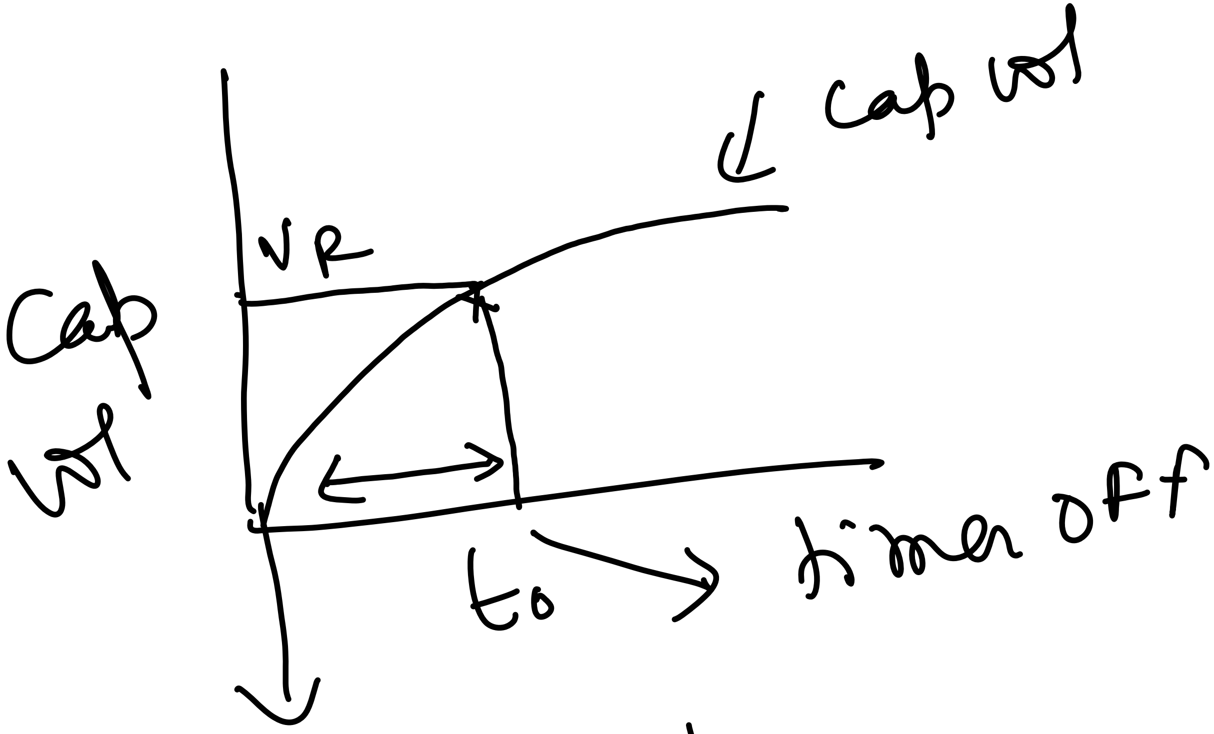
Allow this OFF condition for some time

Now the capacitor C voltage will slowly increase.

The comparator in the μC will change stage once the capacitor voltage goes above reference fixed inside the μC.

Once stage change in the comparator occurs the timer is stopped.

The time period can be used to calculate the V_{in}



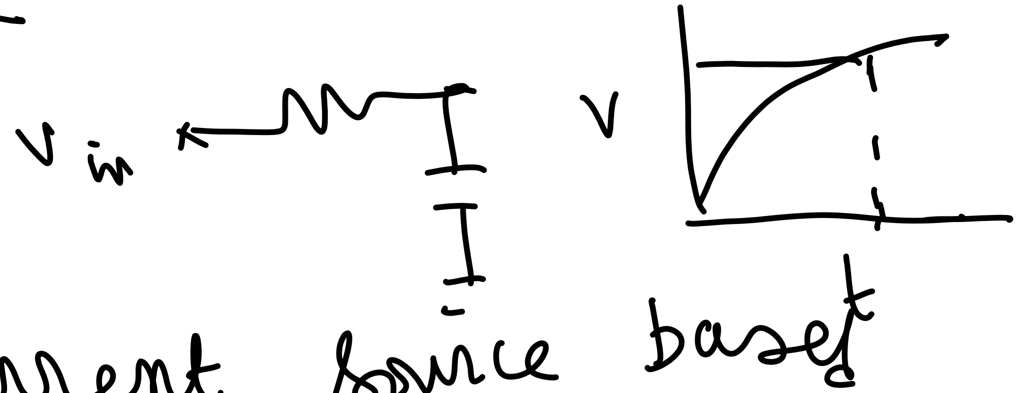
Period to timer ON
gives you the input voltage
This is not a dual slope
converter.

1. The capacitor drift
Resistance drift
all will give error

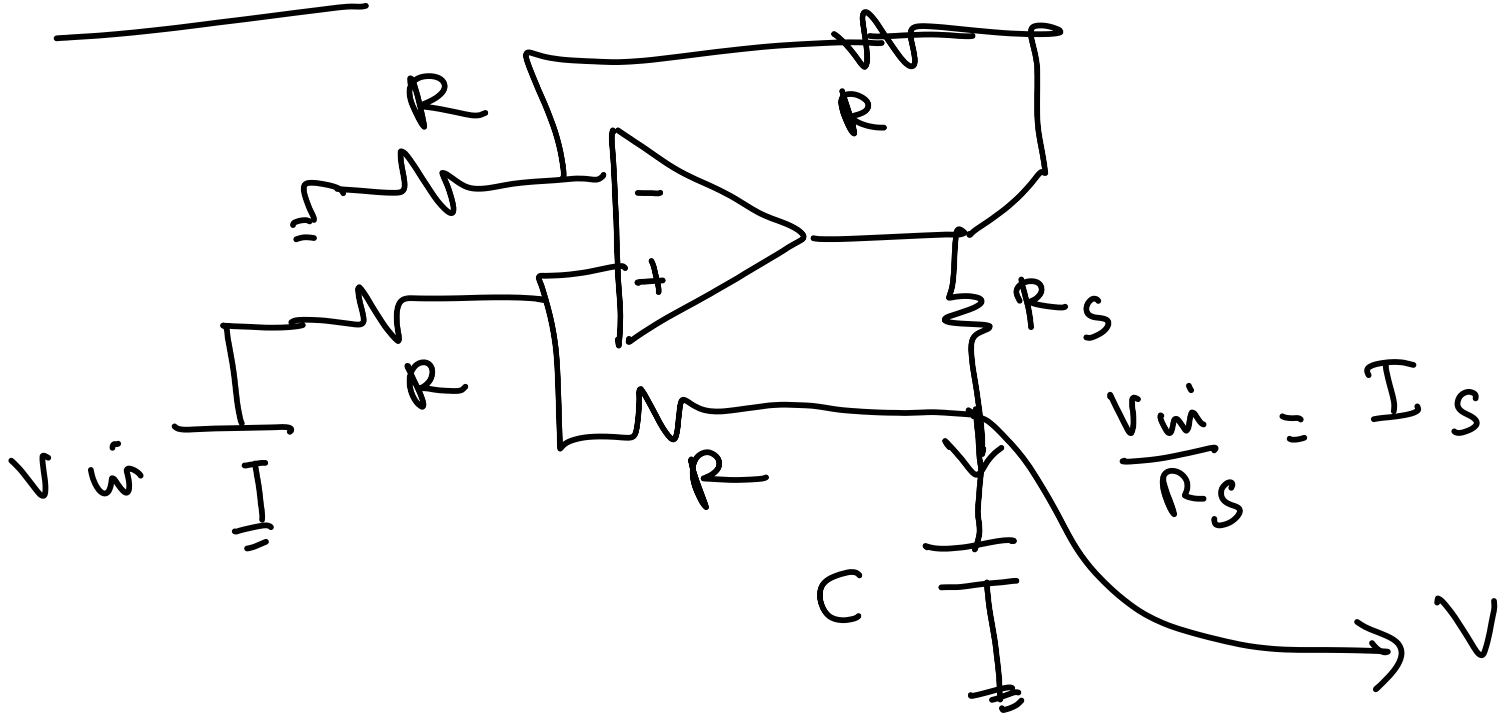
2. The noise averaging
benefit is available.
but not completely because
Ton is not fixed. It
changes with Vin

This converter can be improves
by modifying the circuit.

Lecture No: 36



Constant current source based ADC



$$\frac{V_{in}}{R_S} = I_S$$

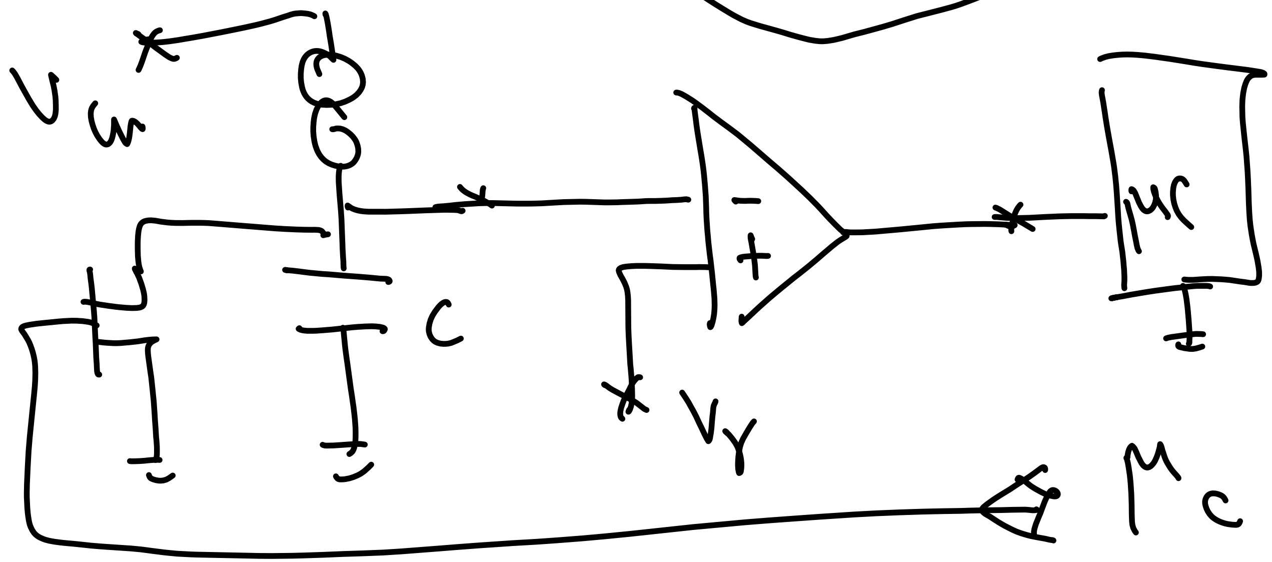
The capacitor c charges linearly with time

$$CV = it$$

$$V = \frac{it}{C} = \frac{V_{in}\, t}{R_S\, C}$$

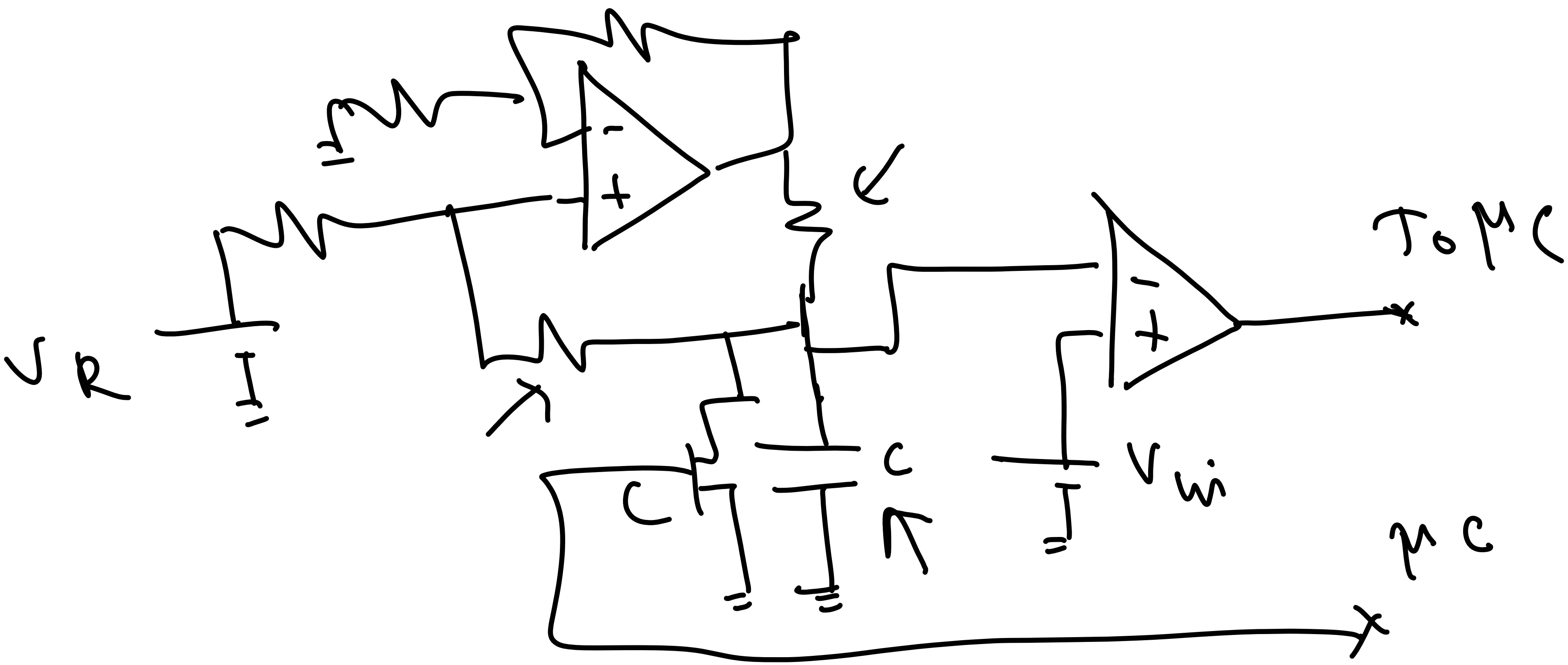
If V_{in} is the vol to be measured

Then $V_{in} = \dfrac{V R_S C}{t}$



① First discharge c
After discharge start the timer

② Wait for comparator roll over
Then stop the timer



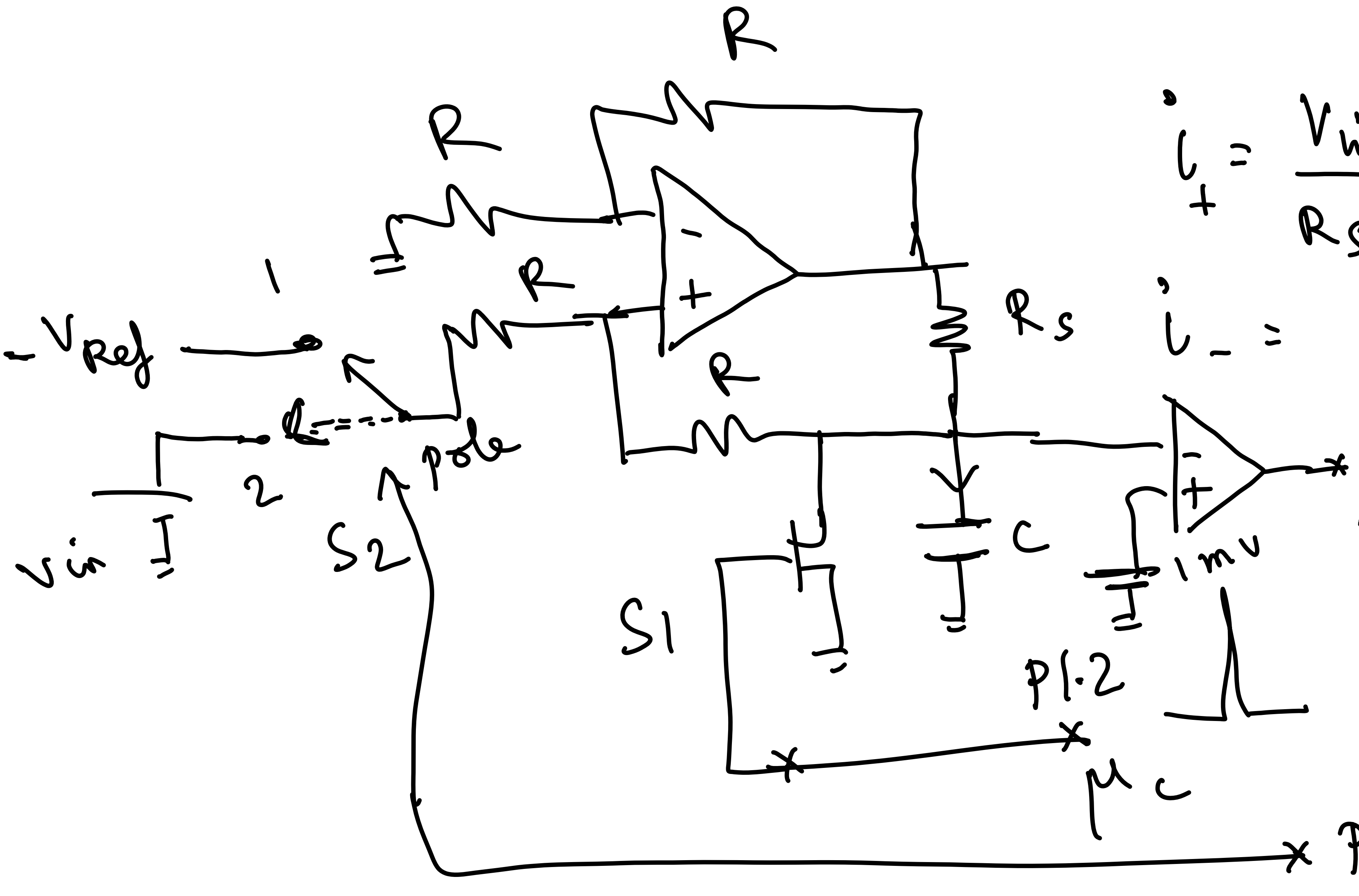
$$it = CV_{in}$$

$$V_{in} = \frac{it}{C}$$

$$V_{in} \propto t$$

1. discharge C and start the timer
2. wait for comparator roll over and stop the timer

Dual Slope Converter with μC



$$i_+ = \frac{V_{in}}{R_S}$$

$$i_- = \frac{V_{Ref}}{R_S}$$

1. Discharge c using the switch S_1
2. Allow the switch S_2 to connect to V_{in}
3. Keep the switch S_2 for a fixed time 't' $it = CV$ ✓

$$\underline{\underline{V}} = \frac{it}{C} = \frac{V_{in} \times t_{ON}}{R_S C}$$

4. Now change the switch S_2 to $-V_{ref}$

This discharges the capacitor C

Time required to discharge can be computed

$it = CV$

$$\frac{V_{Ref} \times t_{dis}}{R/S \times \cancel{C}} = V = \frac{V_{in} \times t_{ON}}{R/S \times \cancel{C}}$$

$$V_{Ref} \times t_{dis} = V_{in} \; t_{ON}$$

$$\frac{V_{Ref} \times t_{dis}}{t_{ON}} = V_{in}$$

(1) This is very accurate

(2) Noise averaging can be controlled → we are fixing the on time

Select the on time in multiples of noise frequent

To remove 50Hz noise

ton can be in multiples of 20 ms.

(3) Low dielectric absorption capacitor must be used

4. Discharge can be checked by the μC using the comparator

Advantage

1. Good noise rejection

Dis advantage

1. It is very slow
2. 5 to 6 conversions only possible per sec

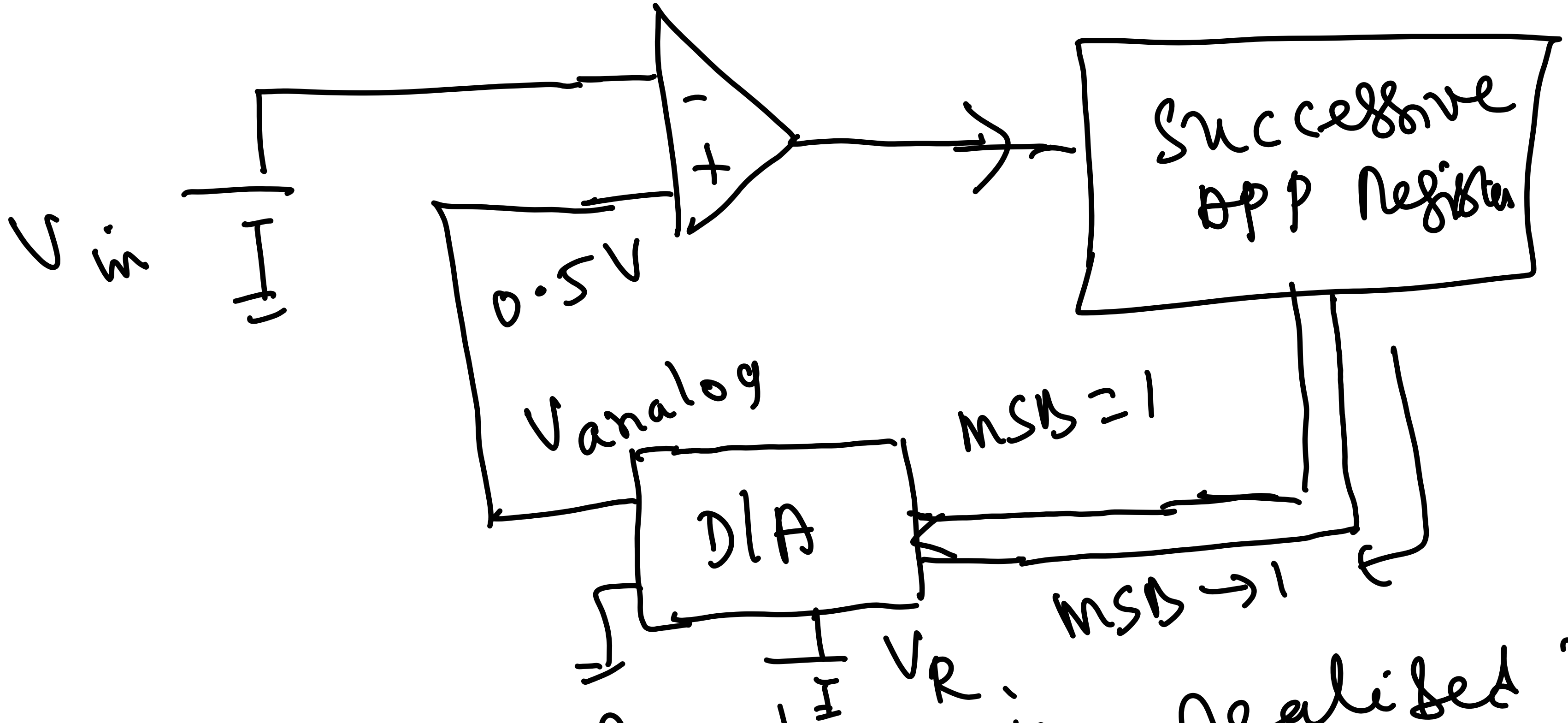
Sucessive approximation ADC

Adv

① Very Fast

dis-adv

② noise rejection is very poor



ADC function is realised using a Digital to analog converter

For 8 bit converter

MSB							LSB
1	0	0	0	0	0	0	0
7	6	5	4	3	2	1	0

1. Keep MSB ie bit no: 7 is high all the other bits 0 - 6 are low
2. The D/A converter responds for this by outputting its analog voltage

③ If the DAC $V_R = 1V$

Then $V_0 = 0.5 V$

V_0 of DAC $= \dfrac{\text{Digital value} \times V_R}{2^n}$

$= \dfrac{128}{256} \times 1 = 0.5$

④ If the input $V_{in} > 0.5V$

Then comparator output is zero

⑤ Now SAR its next trial

Since there is no change

in the comparator output

(a) The MSB ie bit 7 is kept as it is at 1

Then bit 6 is kept 1

and other bits ie 0-5 are kept low

Now bit 6 and 7 are high

(6) DAC gives out

$$0.5 + 0.25 \rightarrow 0.75\,V$$

bit 7 bit 6

(7) Now the comparator changes its output ie goes high

if the input is > 0.75V

it goes low if V_{in} < 0.75 V

Then bit 7, 6 are kept as one

bit 7 is kept as 1
bit 6 is put back to zero

Third trial

Here bit 5 is kept as 1
bits 0-4 are kept 0

Now DAC → 7 6 5
0.5 0.25 0.125
→ 0.875

0.75
.12
0.875

Now the comparator goes high if it is < 0.875

like this

8 trials are made

In each trial if the comparator goes up that bit is put back otherwise it is kept as as zero

one

By this at the end of eight trial the converted digital output is ready.

It needs only eight trials to get the conversion completed

For example

If $1 \mu s$ is required for one conversion

Then only $8 \mu c$ is required to complete the conversion

So it is very fast

Important

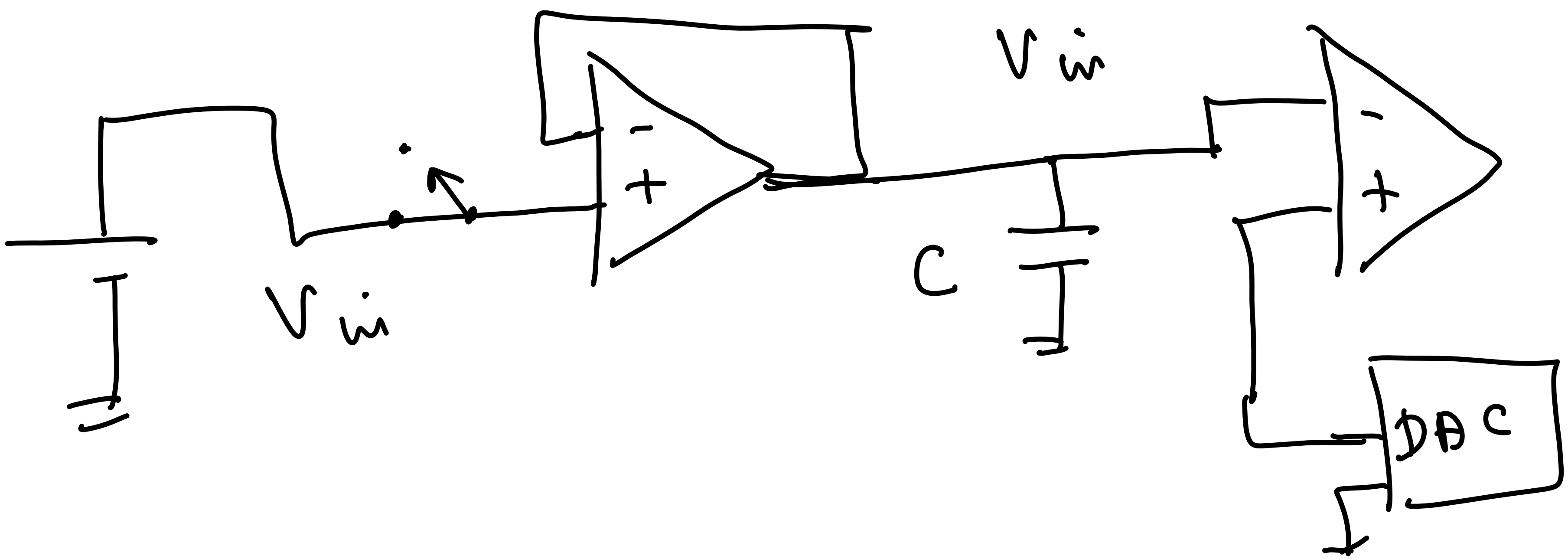
This is valid ie

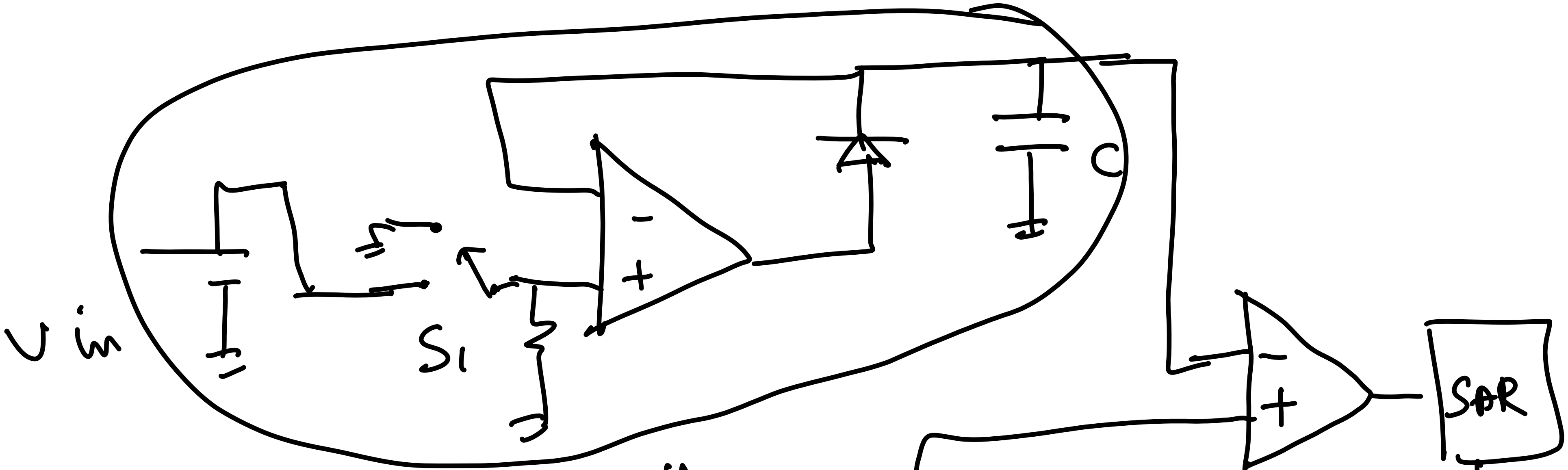
Converted output is valid only if the input voltage is constant during the conversion time

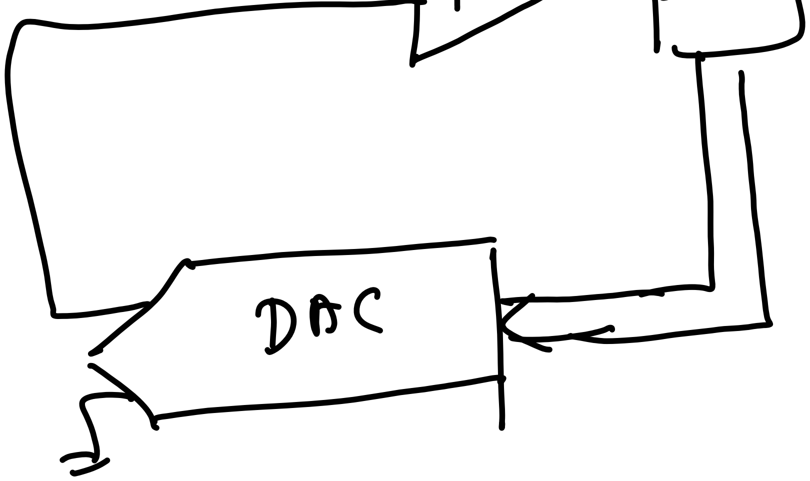
If the V_{in} changes during $(t \times 8)$ period then result is wrong.

How to keep the
input vol constant?

Latch the V_{in} using Sample
and hold circuit







1. S_1 is connected V_{in}
   After small time
2. $S_1 \rightarrow$ GND
   Vol in C remains
   constant
3. Start the SAR conversion

This way of acquiring the V_{in} to the ADC is called Sample and hold amplifier

Sample and hold amplifiers are commercially available

Total conversion time

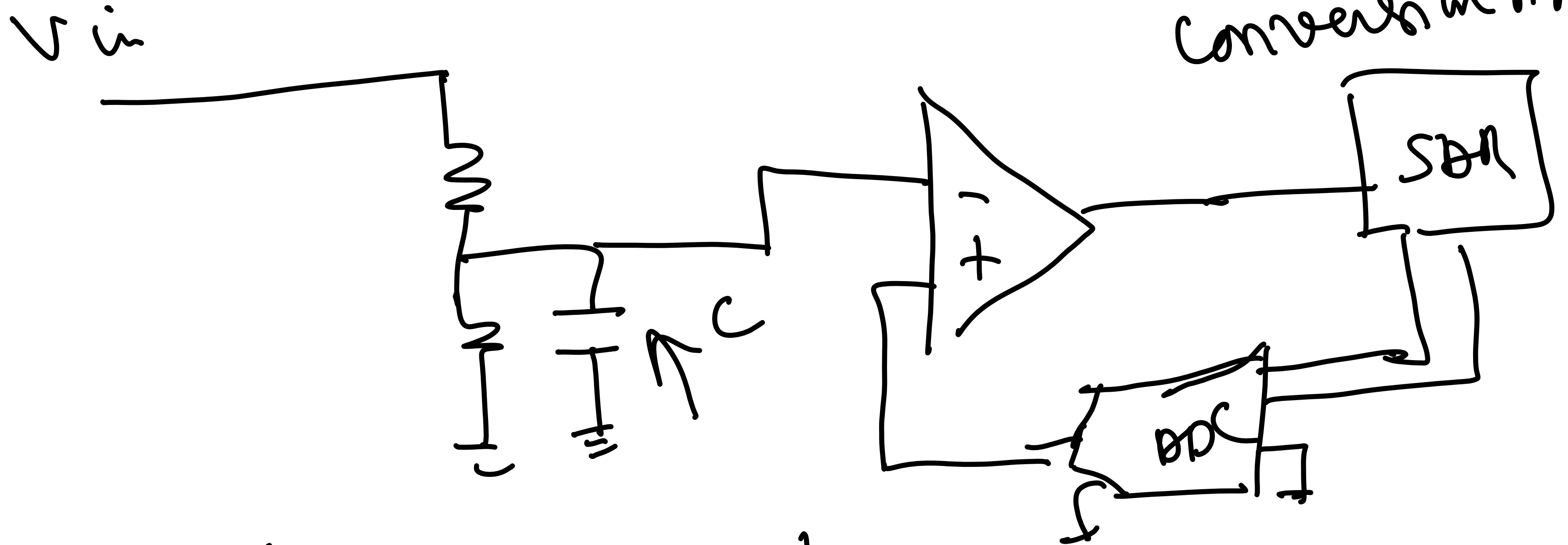
$$T_{ac} + T_{con} = T_{total}$$

If Sample and hold Amp is
(SHA)

not used

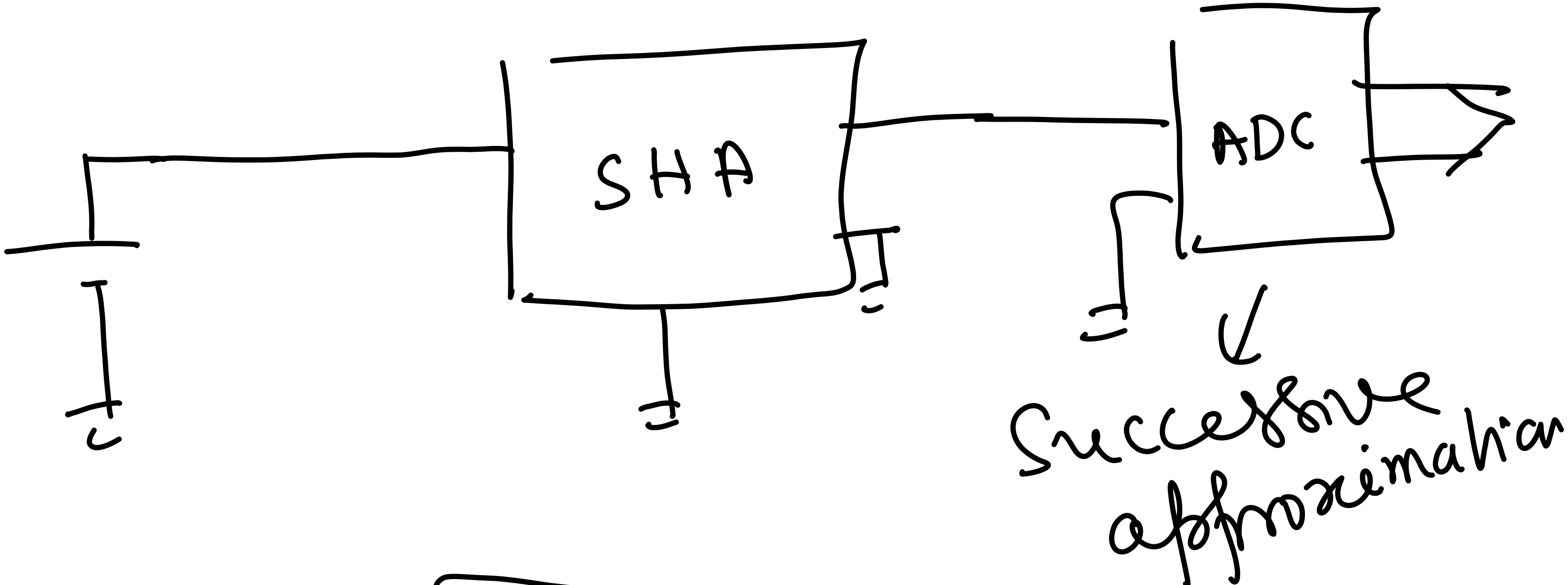
Then $\delta V_{in} \leq \frac{1}{2}$ LSB

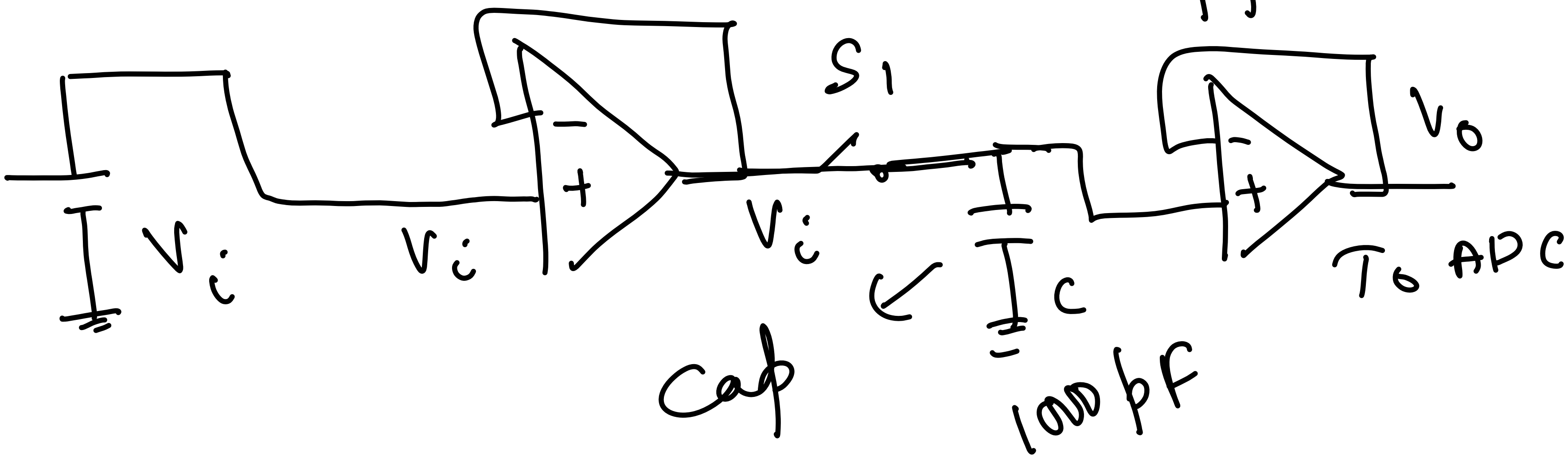
(during the conversion time)



if C is very large so that $\Delta V_{in} \leq \frac{1}{2}$ LSB

Lecture No: 37





When S_1 is closed

The cap charges up to V_i

Since R_o of the op amp is very low

The charging is very fast

Normally it is the settling time of the op amp determines the charging time

is $<$ 1MS

Then Switch of S_1

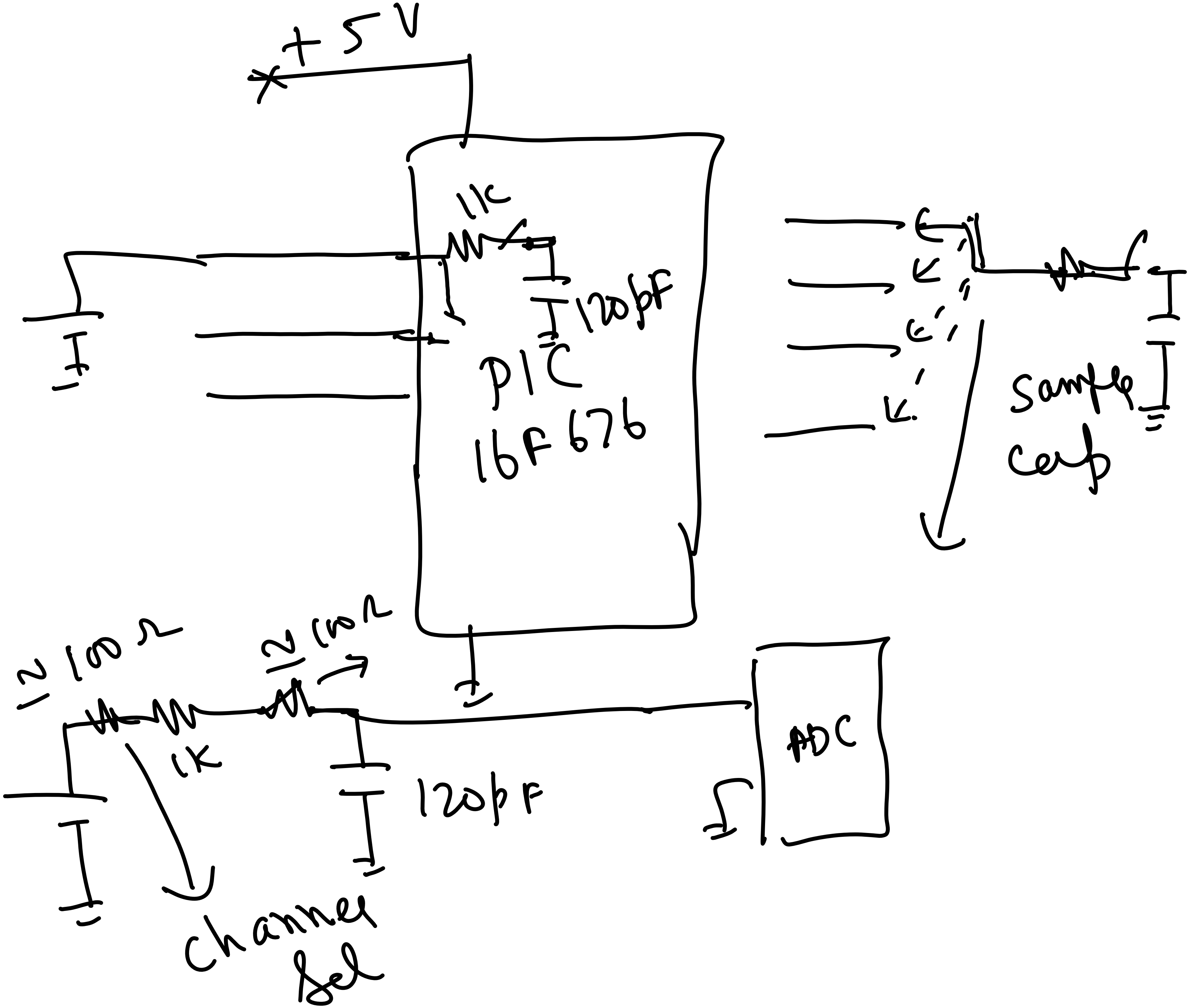
Now ADC start command can be given.

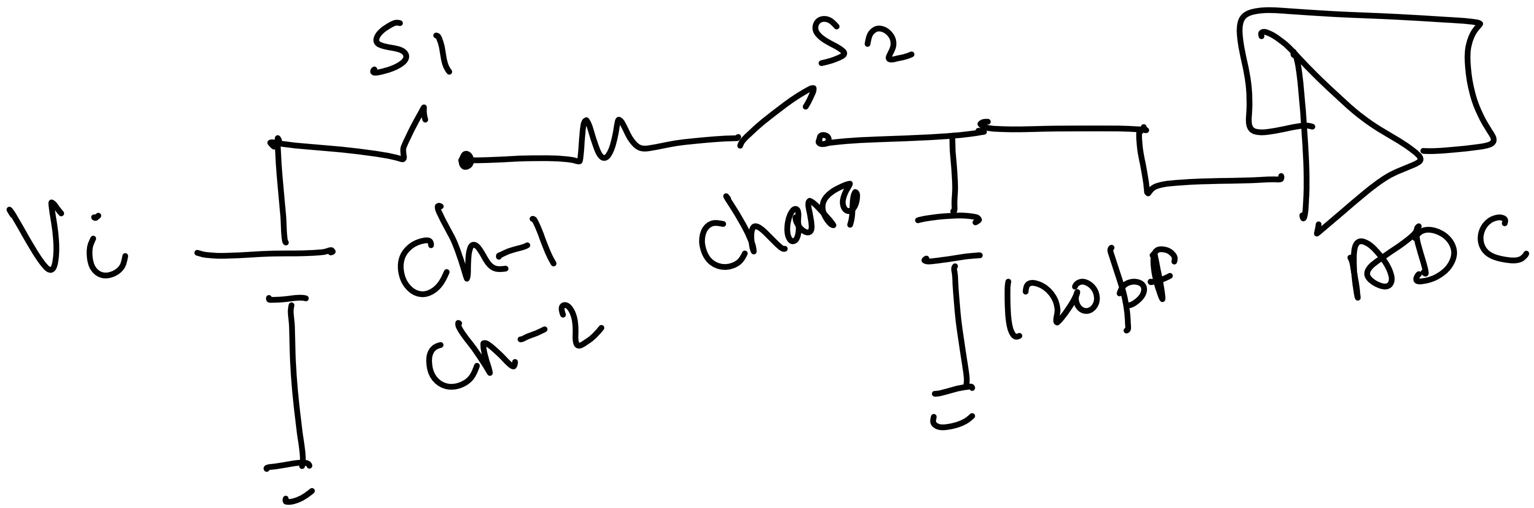
ADC takes normally few μs for conversion

During conversion the voltage in the capacitor should not decrease

PIC MC

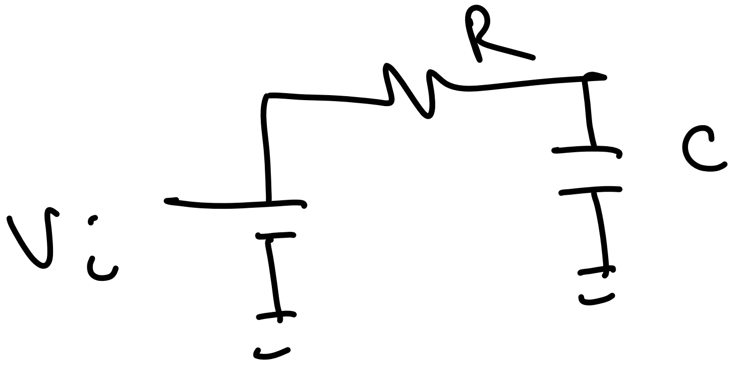
1. In build 10 bit ADC
2. Multiple Analog input channels
3. In build SHA


+5V
1k
120pF
PIC
16F676
sample
cap
~100Ω
~100Ω
1K
120pF
ADC
Channel
Sel



1. S_1 and S_2 are ON
2. Allow the cap to charge up to $1/2$ LSB

$$V = V_i e^{t/RC}$$



How much time I have to wait for charging up to $(V_i - \frac{1}{2} LSB)$

$$V = V_i \, e^{t/RC}$$

$$\frac{V}{V_i} = e^{t/RC}$$

$$\frac{256 \times 4}{1024} = \qquad \frac{2^{10} - \frac{1}{2} LSB}{2^{10}} = e^{t/RC}$$

$$\ln \left(\frac{1024 - \frac{1}{2}}{1024} \right) = \frac{t}{RC}$$

$$\ln\left(1 - \frac{1}{2048}\right) = t/RC$$

$$\ln\left(\frac{1}{2048}\right) = t/RC$$

$$RC \ln\left(\frac{1}{2048}\right) = t$$

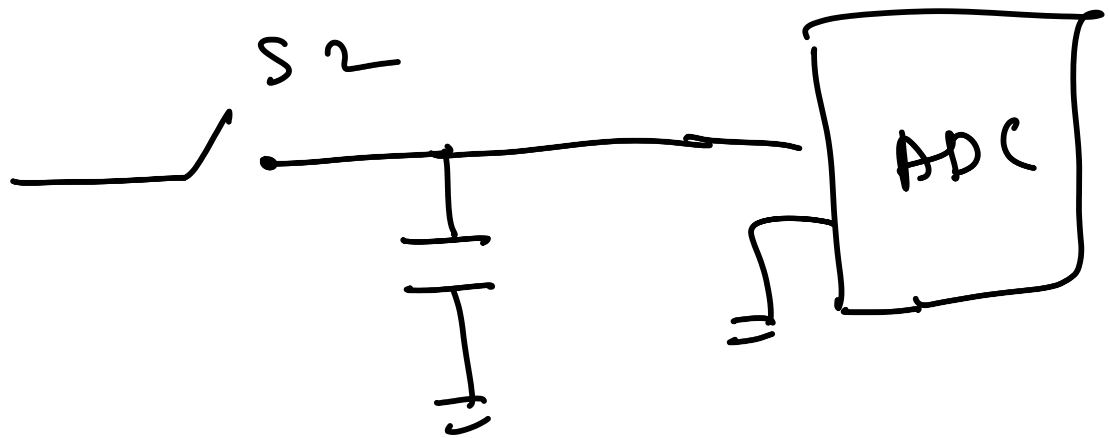
For $R = 1k$

$C = 120\,pF$

$t = 20\,\mu s$

(3) After waiting for t secs Start Conversion Command is given to ADC

(4) At this stage S_2 is Switched off



In PIC MC

For 10 bit ADC it needs 10+1 cycle time for conversion.

The conversion is done from MSB onwards LSB

1 1 1 1 1 1 1 1 1 1

2nd 3rd 11th

Minimum cycle time is $2\mu s$

For 10 bit complete conversion

wait for $11 \times 2 = 22\mu s$

SHA time + conversion time

$22\mu s$

$20\mu s$

$\longrightarrow 42\mu s$

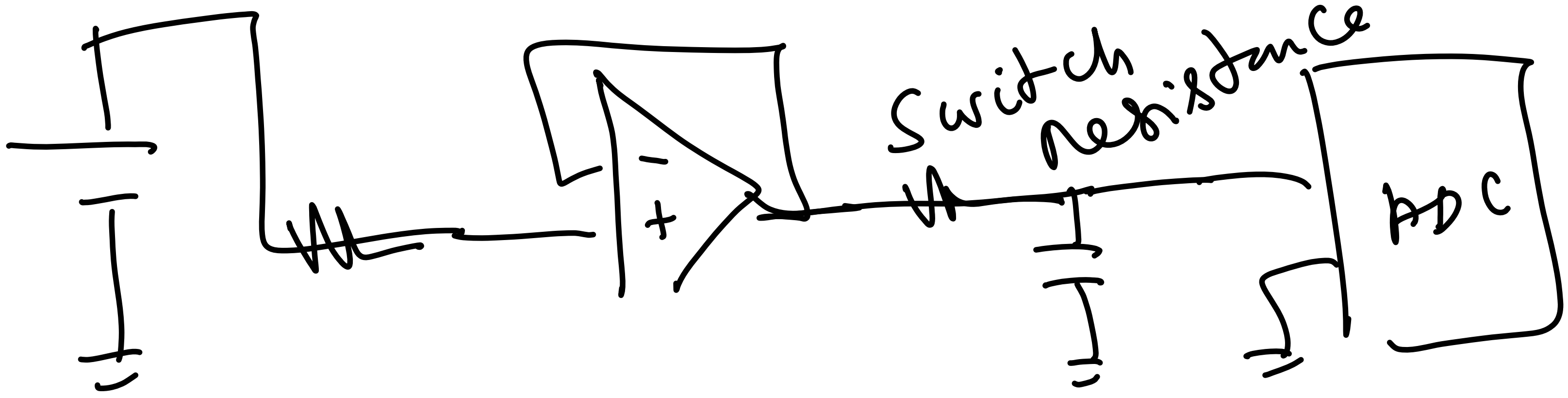
In actual case

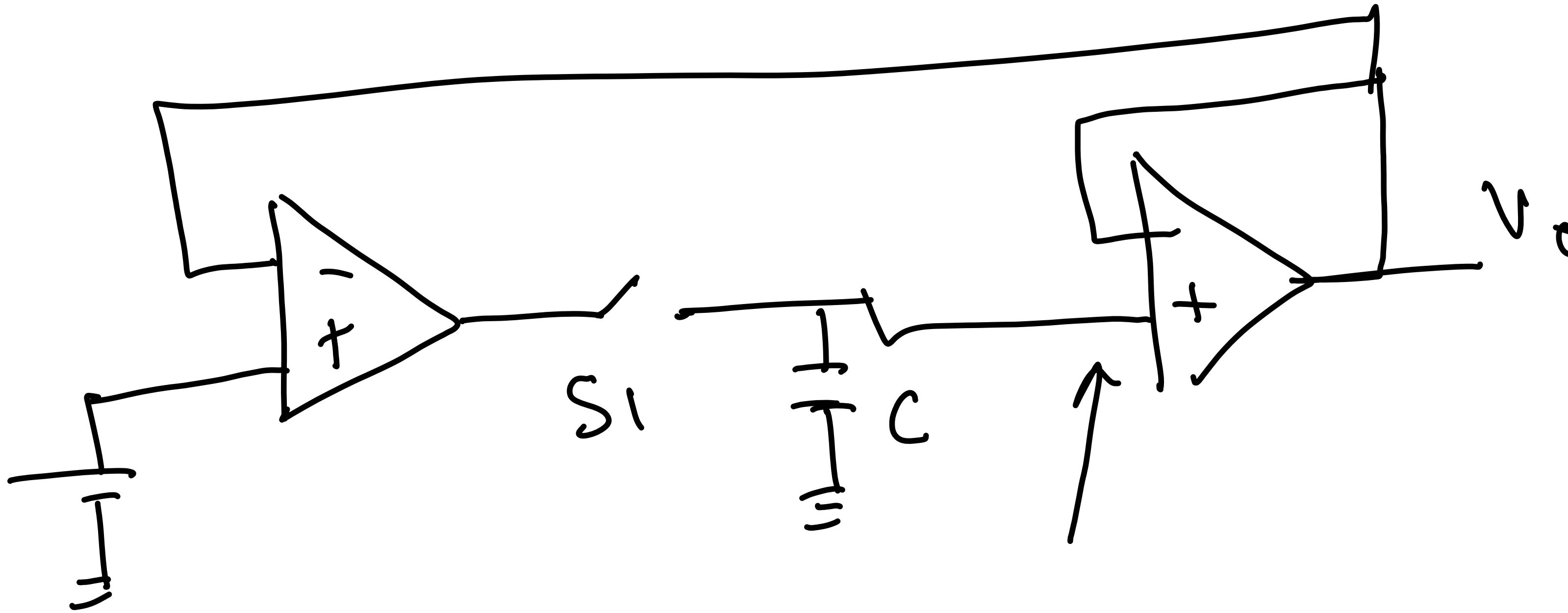
we need to configuration of channel

All the "settling" mismatches can take up to 10 μs

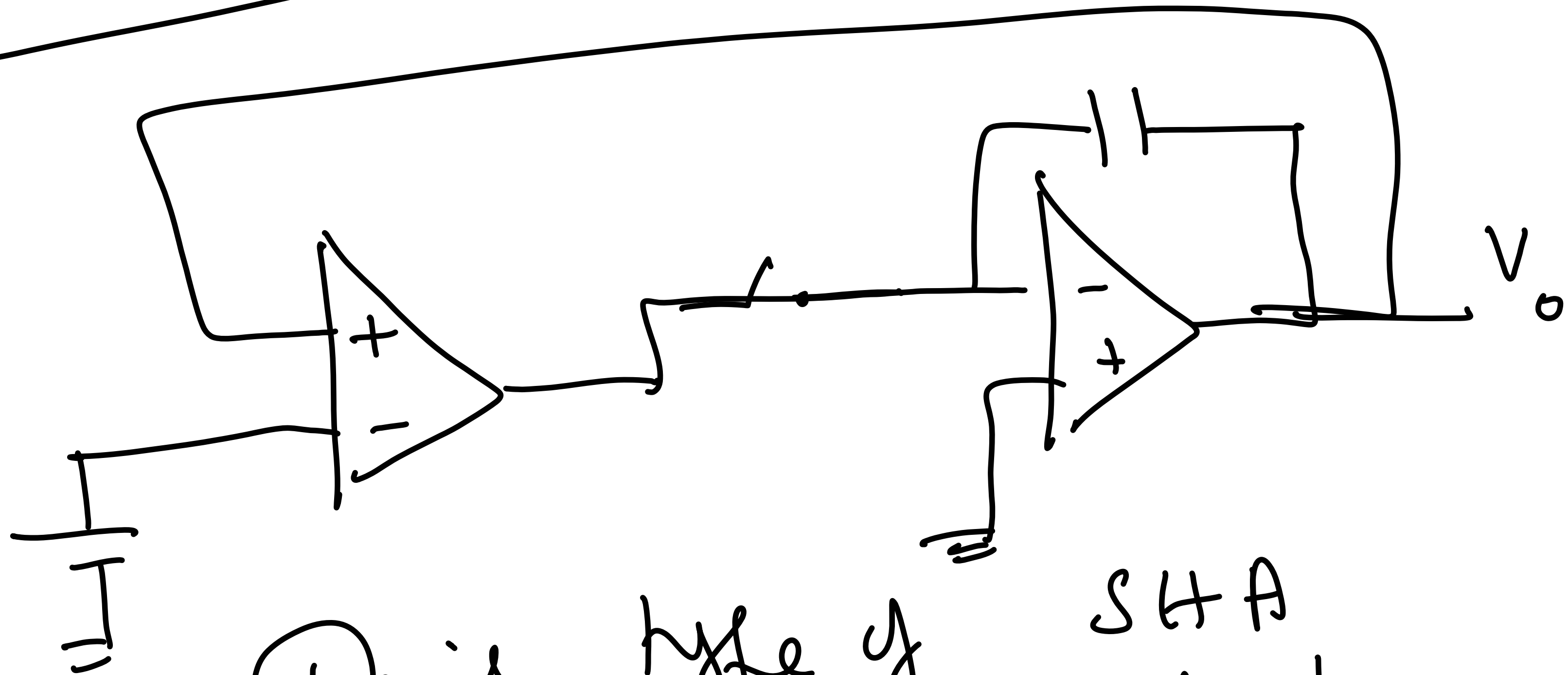
1. How to reduce the sampling time?

2. How to reduce the conversion time of ADC


Switch
resistance
ADC

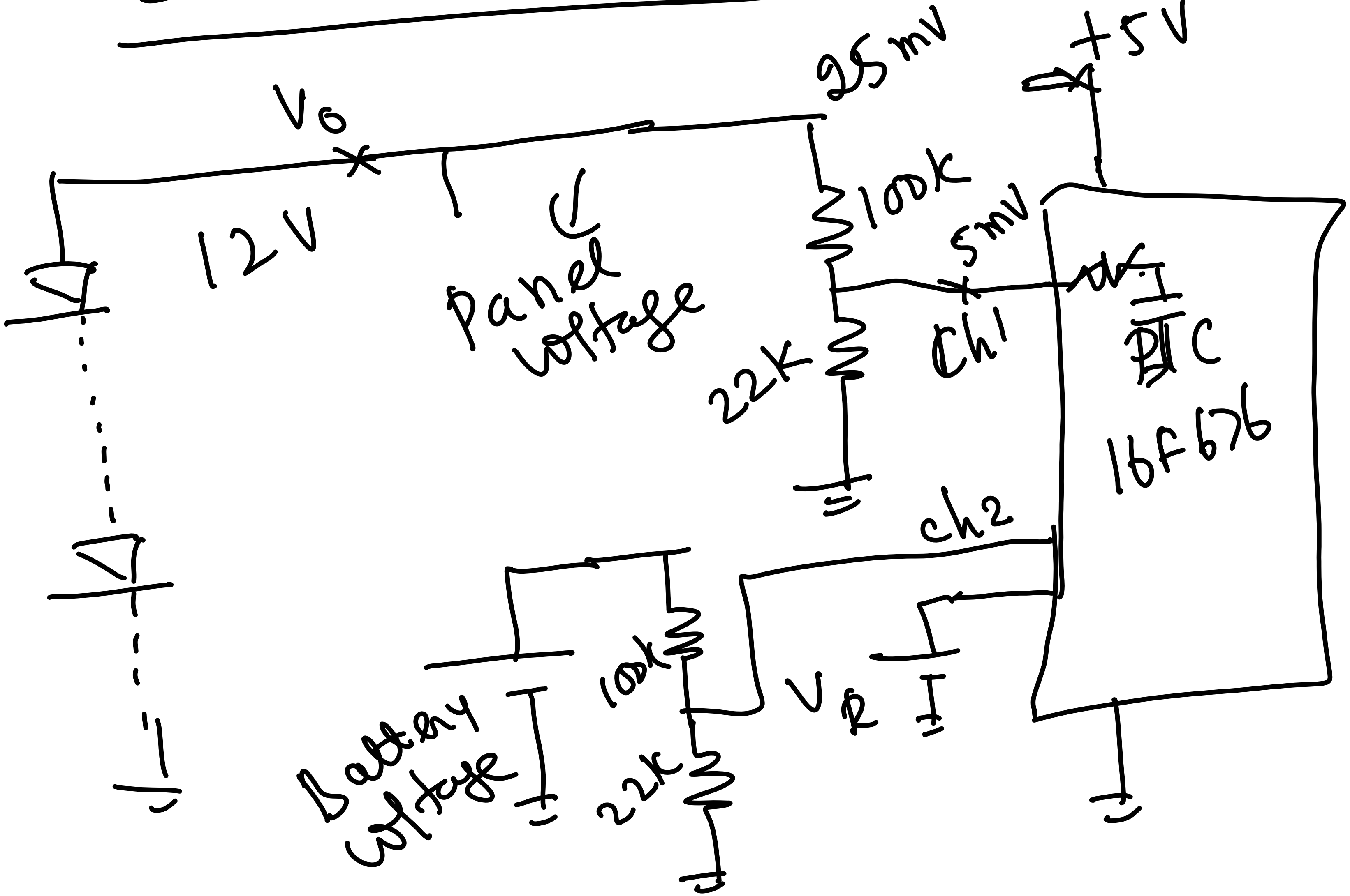

S1
C
V_0

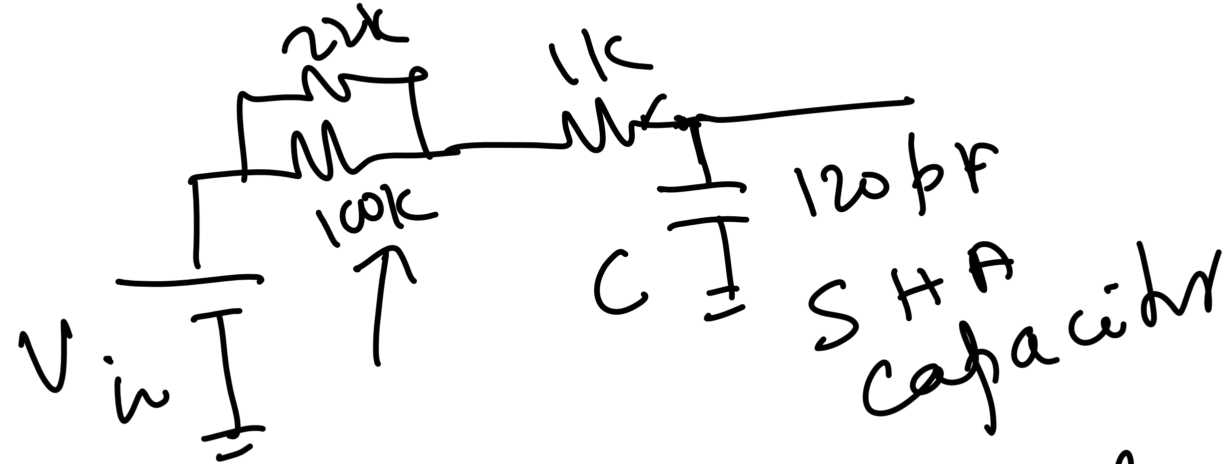
II nd type



This type of SHA will not ~~any~~ introduce additional charge to the cap

Solar cell based Street lighting

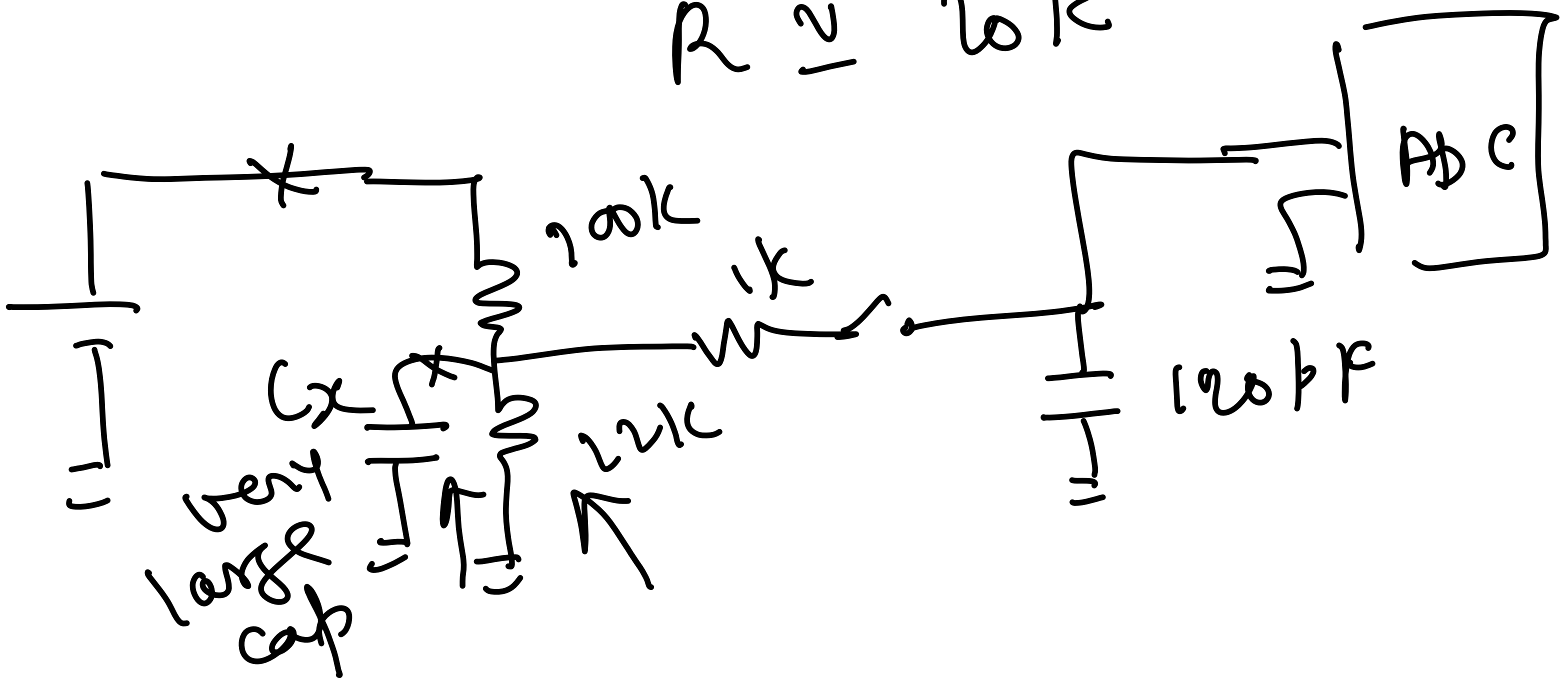


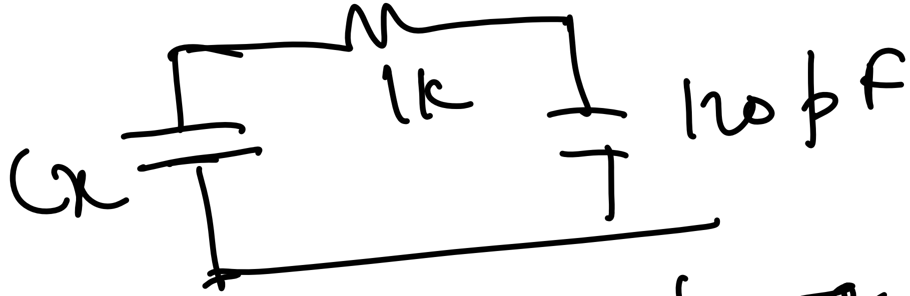


We have to wait for

$$t = \ln\left(\frac{1}{2048}\right) \times RC$$

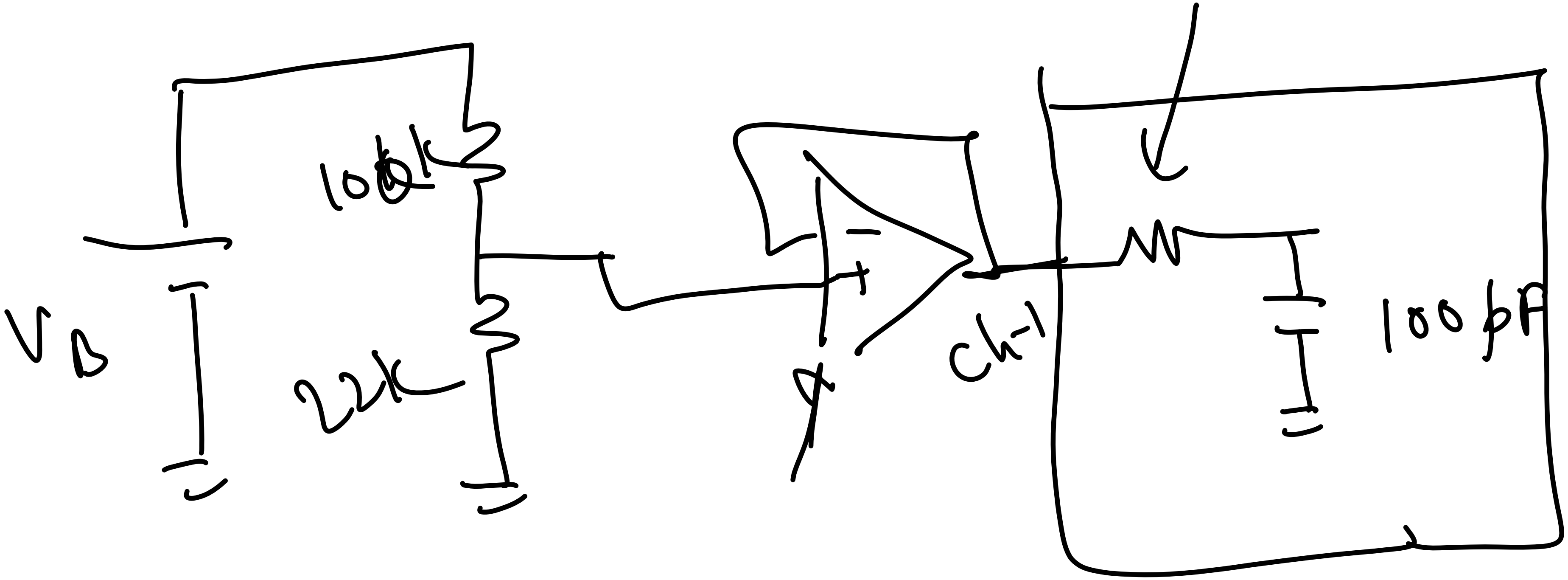
$R \simeq 20k$



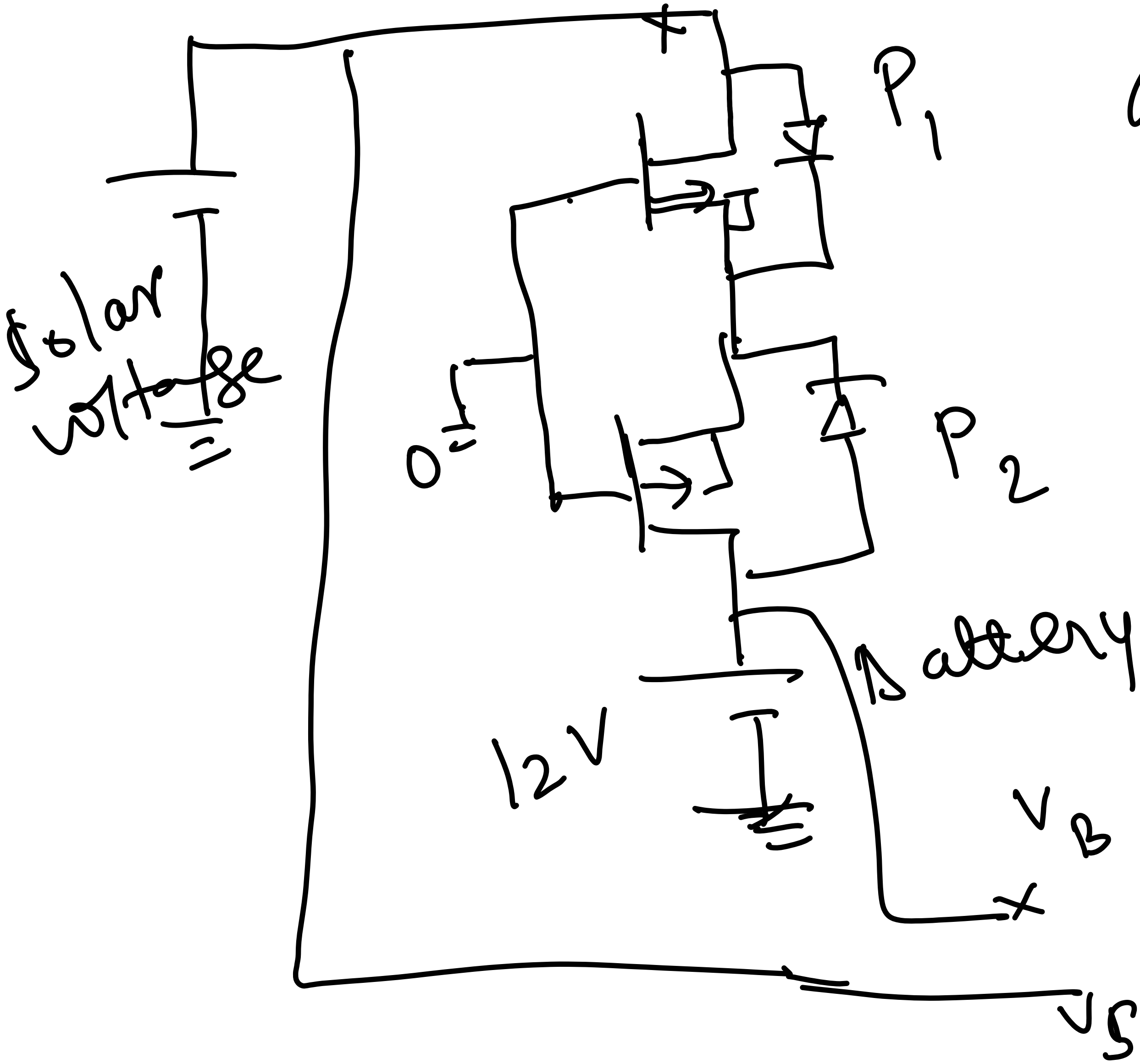


This method charges the hold cap very rapidly.

But it is not fast enough to respond to the battery voltage change



100k
22k
V_B
Ch-1
100 pF



Gates are grounded
Both P_1 and P_2 are on

If gates are kept at 12V
Then P_1 and P_2 are OFF

If $V_S > V_D$

Put gates to zero

then charging of the battery takes place

However

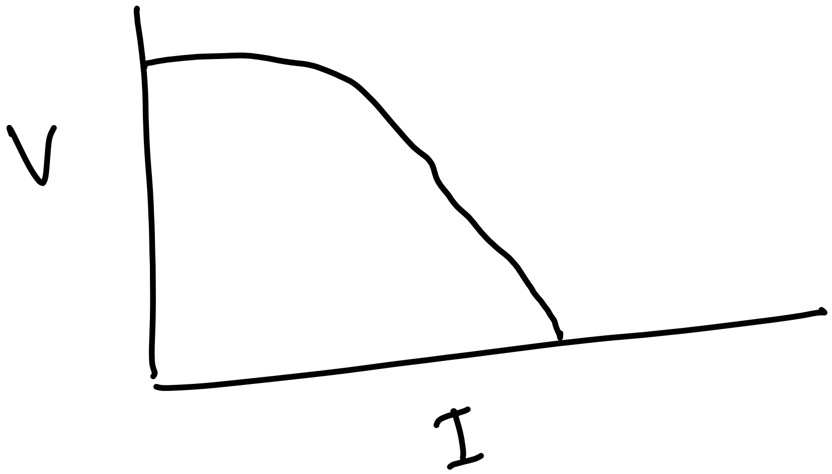
If V_s comes down suddenly due to cloud or any thing else

Then battery will dis charge to the panel

This ~~pri~~ must be stoped immediatly

ie if $V_B > V_S$

P_1 and P_2 must be switched off quickly



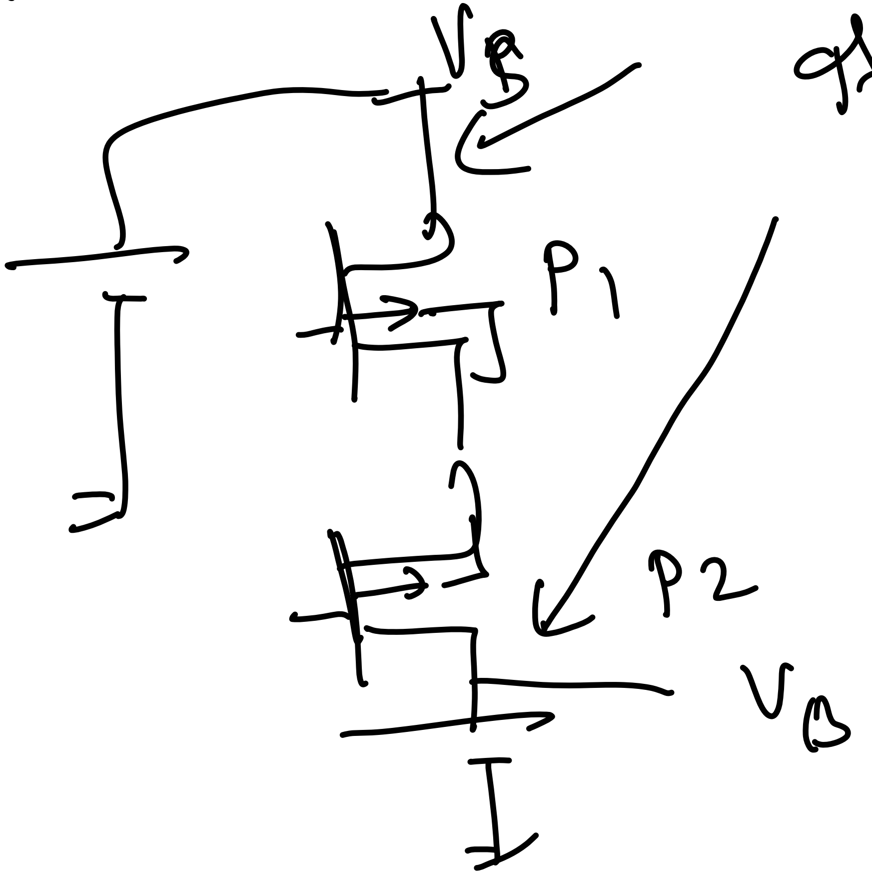
For example
in this case $V_R = 5V$
If is a 10 bit converter

$$1\,LSB = \frac{5}{2^{10}} = \frac{5}{1024} = 5mV$$

Considering the voltage divider
the minimum detectable voltage
at the panel is

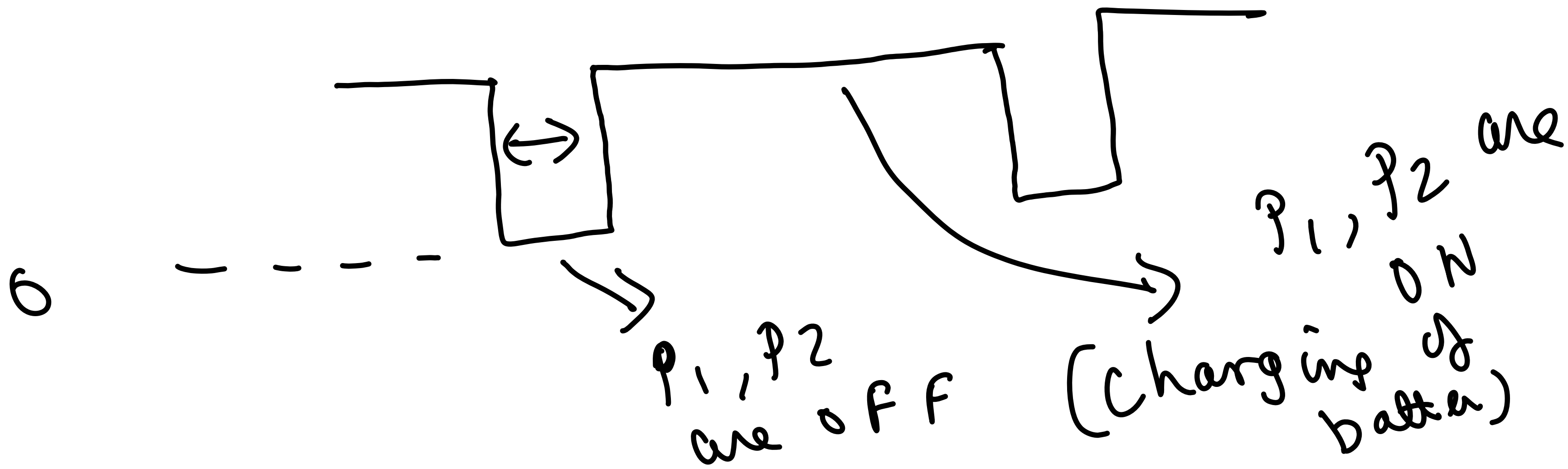
$$5mV \times \frac{122k}{22k} = 30mV$$

Minimum vol resol at the panel is 30 mV



If it is less than 30 mV

1. Switch off P_1 and P_2
2. Measure quickly V_S and V_B
3. If $V_S > V_B$ switch on P_1 P_2
4. Keep this ON state for about 3 to 5 ms
5. Repeat from step 1



During P_1, P_2 energy is not drawn from the solar cell ie we are not utilising the cell fully.

So the OFF time should be very short compared to the ON time

ie ADC must be very fast

why not longer ON time?

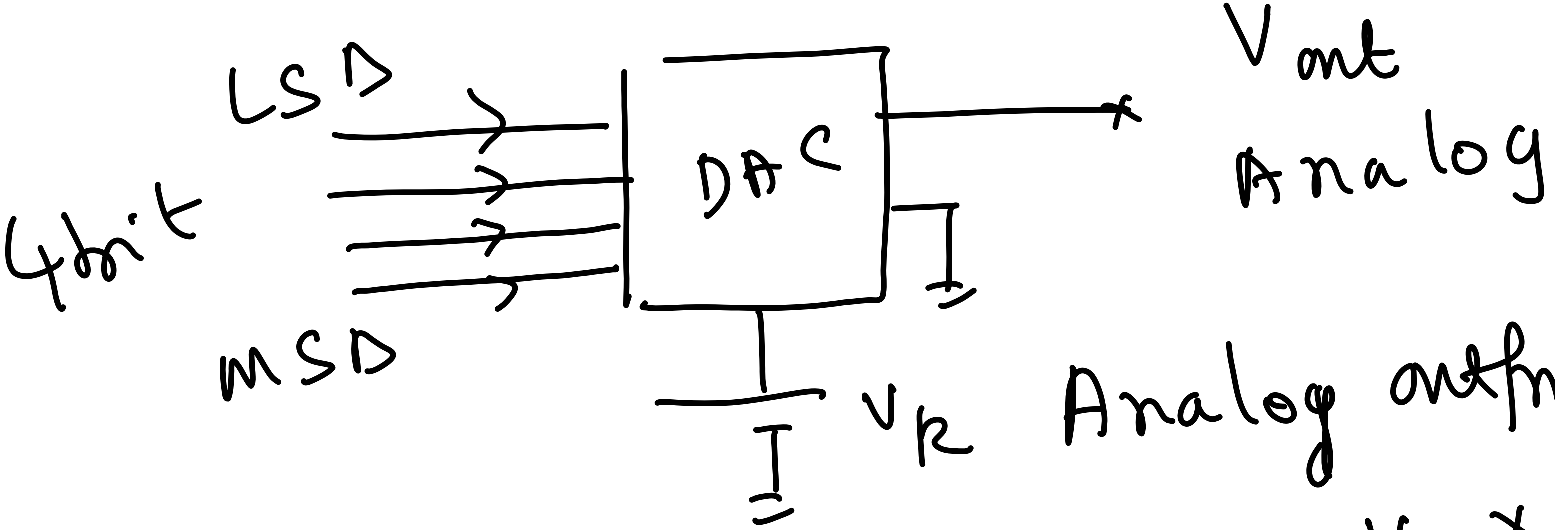
This is bad because if $V_B > V_S$ during ON time battery will be discharging to the panel.

Diode is not used to block the reverse flow. ~~because~~ the diode will eat away 0.4V

Lecture No: 38

Digital to Analog Converters

DAC



Analog output

$$= V_R \times \frac{(\text{Binary value})}{2^n}$$

$V_R = 1 \quad V$

$n = 4 \quad$ (4 bit converter)

0001

$$V_0 = \frac{1 \times (BV)}{2^4} = \frac{1 \times 1}{16} = \frac{1}{16} V$$

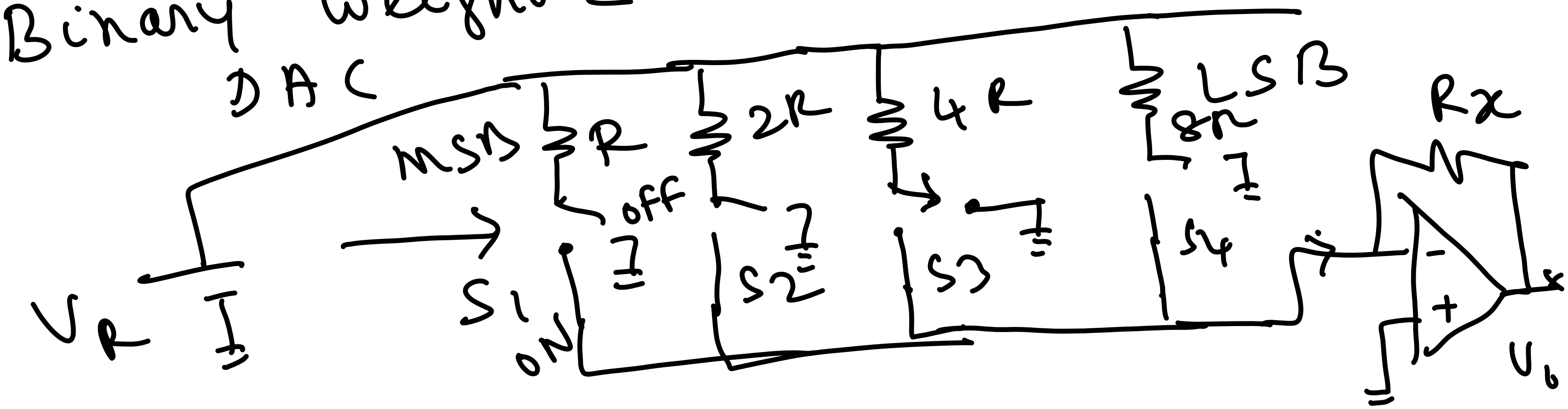
For 1111 $\qquad V_0 = \frac{1 \times 15}{16} V$

Two types of DAC

① Binary weighted DAC

② R - 2R based DAC

Binary weighted DAC



MSD $\rightarrow$ 1 and All other bits are zero

Input $\rightarrow$ 1 0 0 0 then

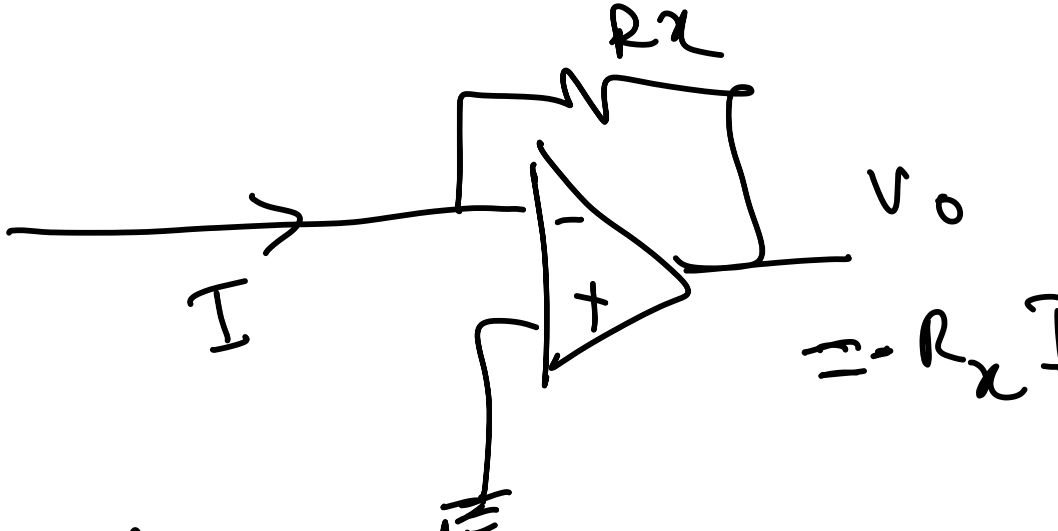
S_1 ON $\longrightarrow$ connected to op amp

$S_2, S_3, S_4 \longrightarrow$ OFF connected to gnd

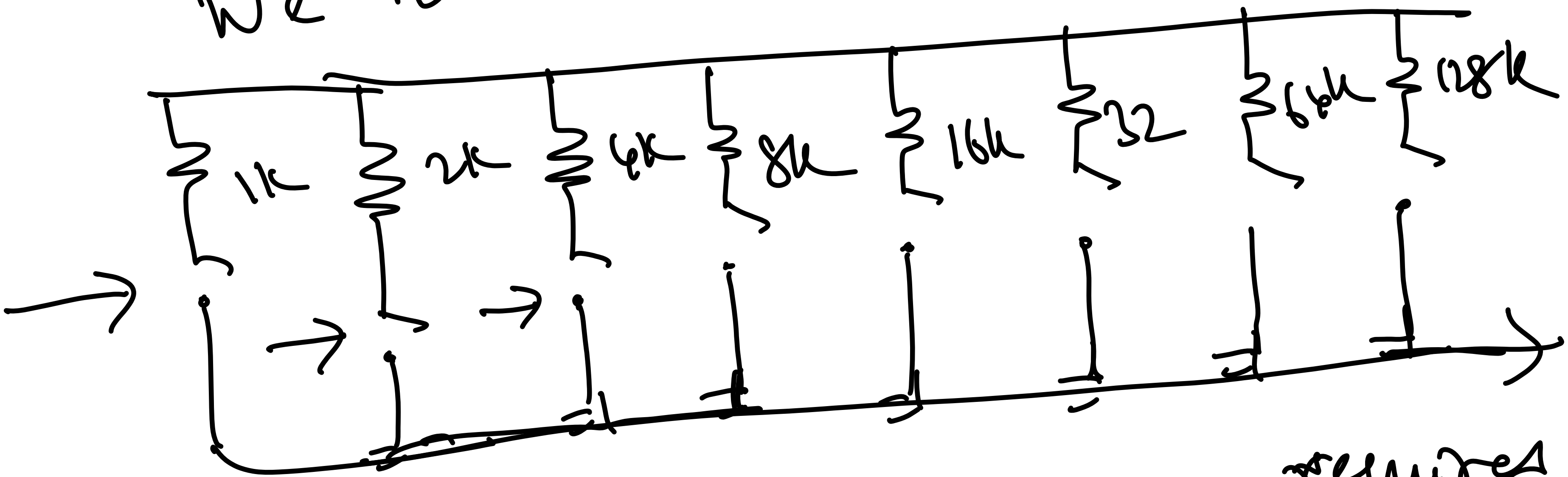
Then current I to the inverting input of op amp

$$= \frac{V_R}{R} + 0 + 0 + 0$$

For digital input 1111 then $I = \frac{V_R}{R} + \frac{V_R}{2R} + \frac{V_R}{4R} + \frac{V_R}{8R}$



For 8 bit Converter
we need 8 resistors



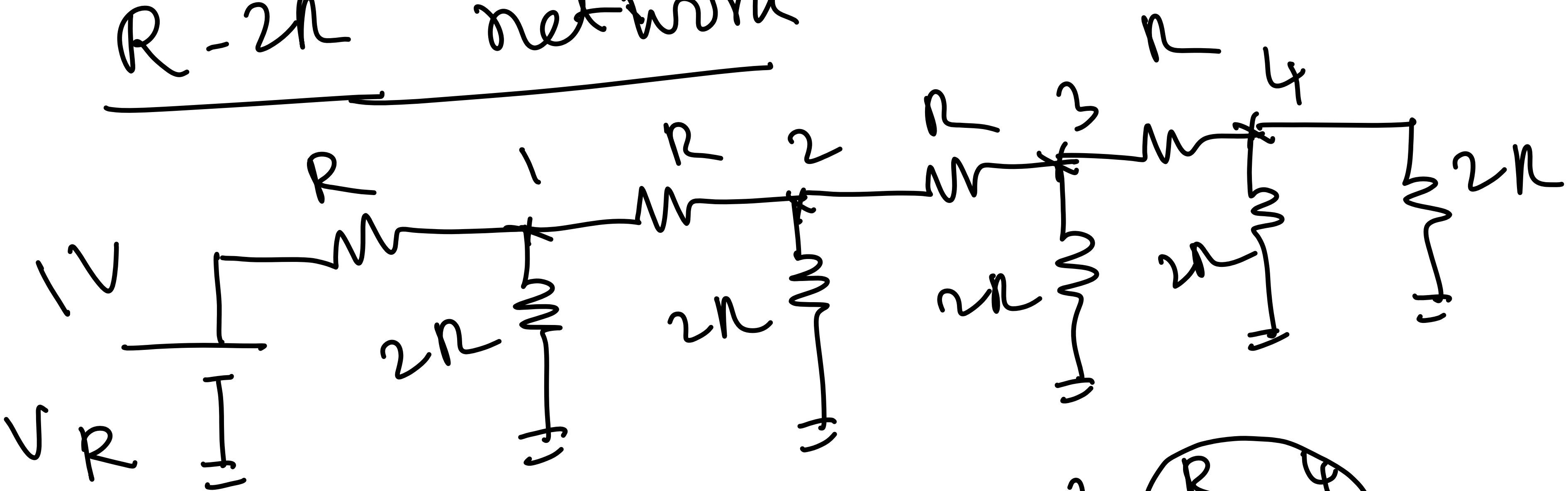
$\frac{1}{2^8}$ Accuracy required
for all the 8 resistors

In binary weighted resistor based DAC

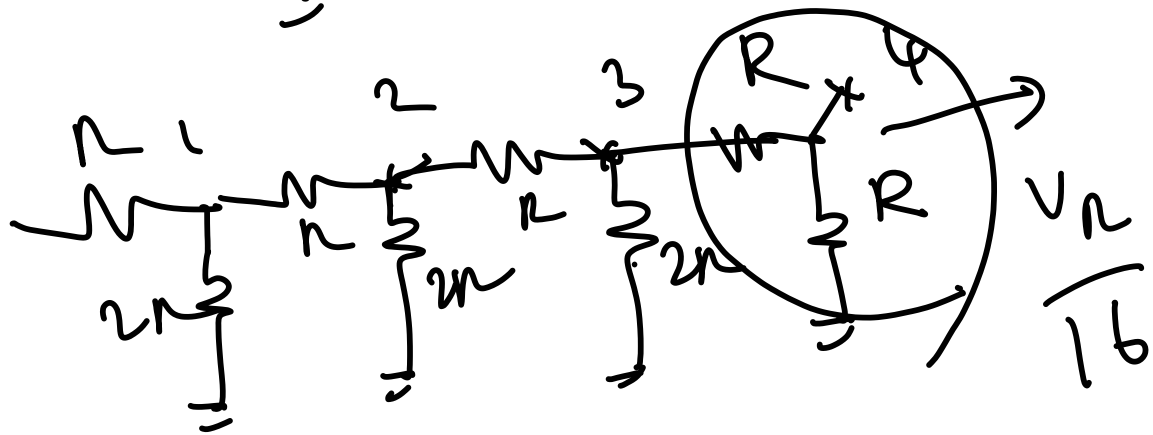
Accuracy requirement for resistors are difficult to achieve particularly for high bit DACS

R-2R network based DAC

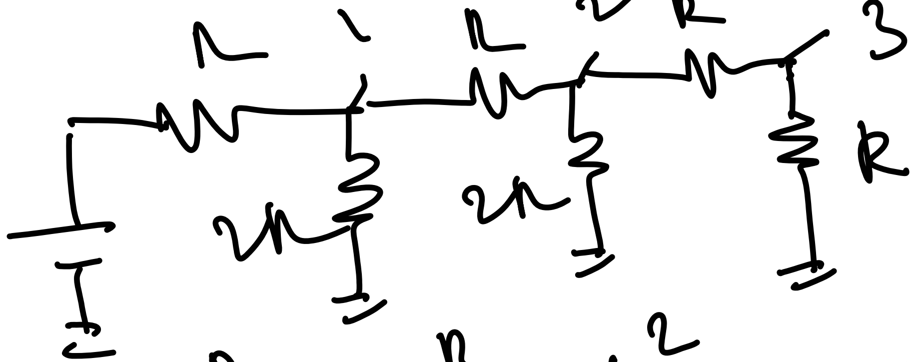
R-2R network



Node-4

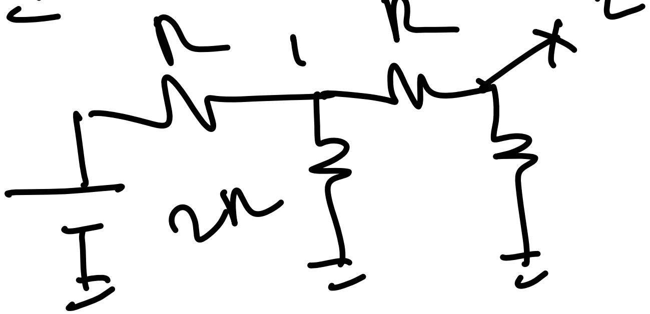


3



$\frac{V_R}{8}$

2

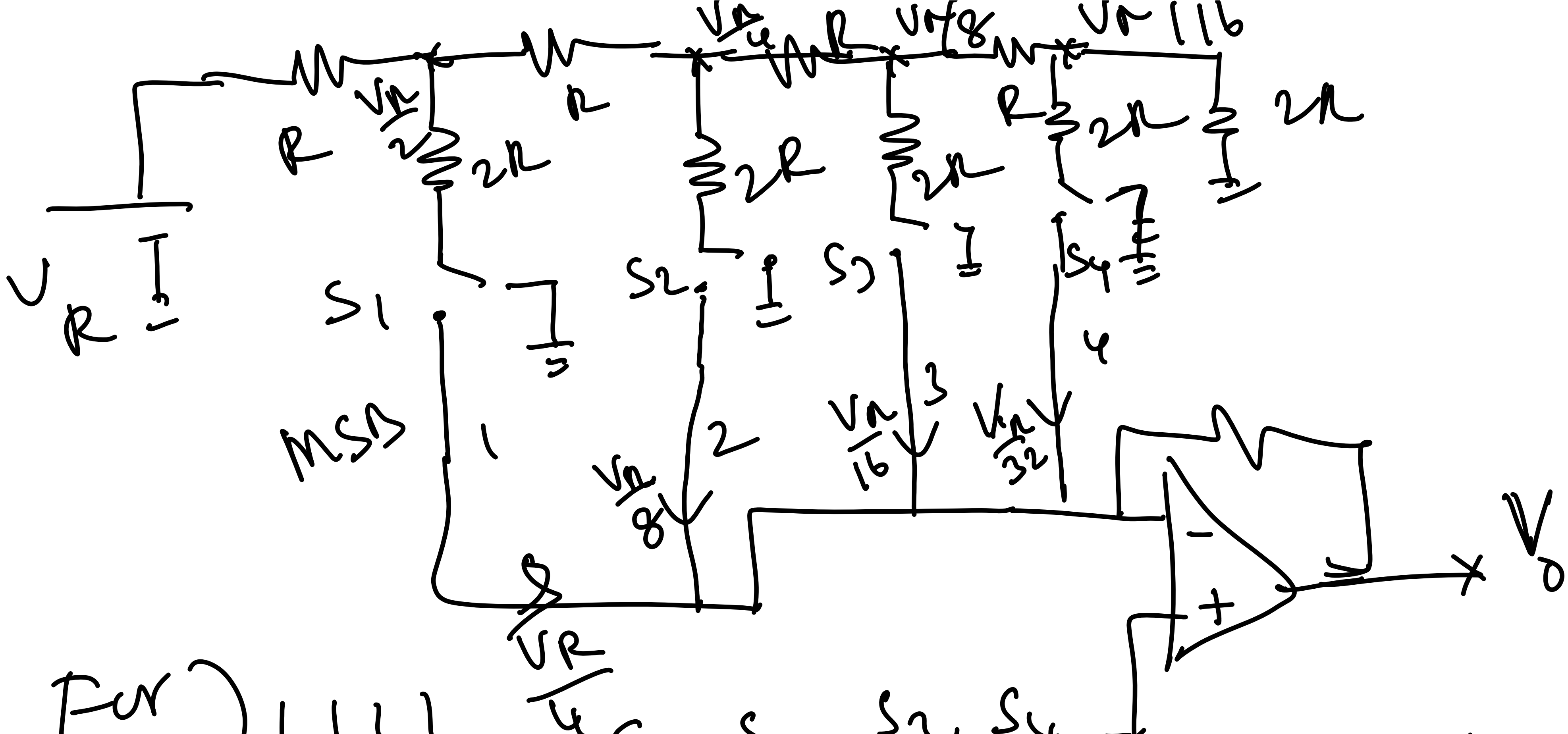


$0.25 = \frac{V_R}{4}$

1

$0.5V = \frac{V_R}{2}$

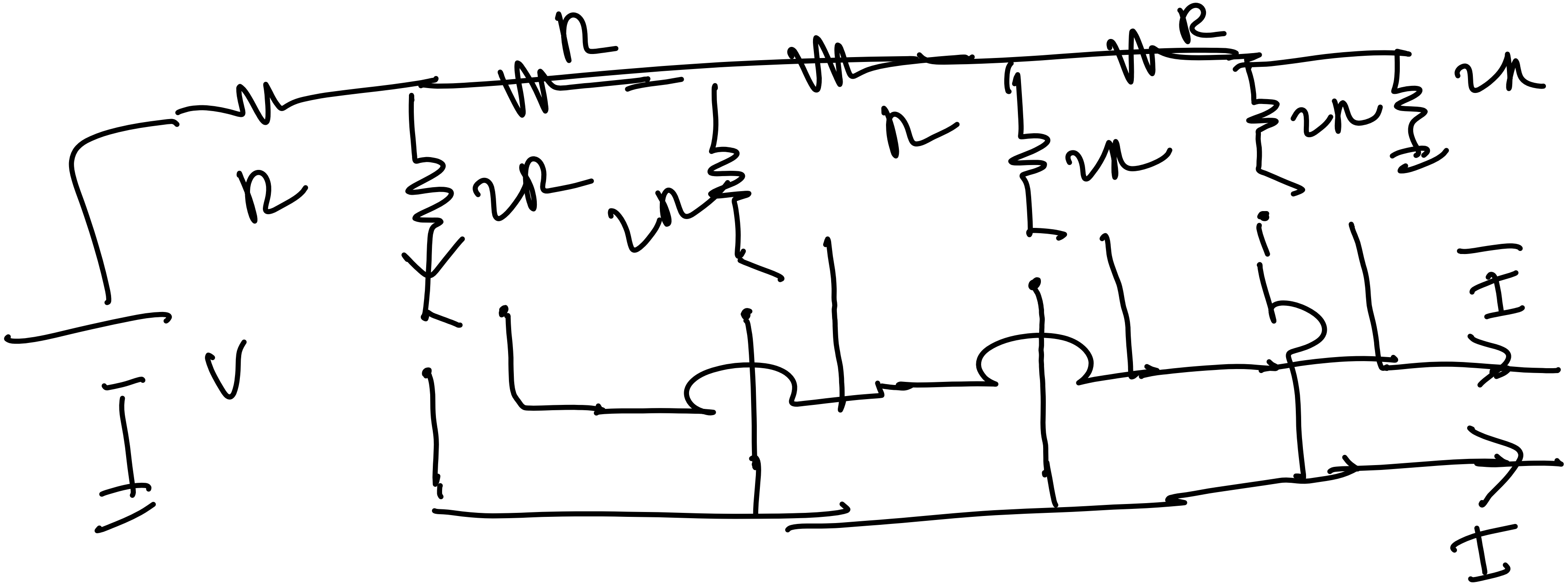
One can get binary weighted voltage using only two different resistors (R and 2R)



For digital input } 1111 S_1, S_2, S_3, S_4 are delivering current to inverting input of op amp

For digital input } 0 0 0 0 $\longrightarrow$ $S_1, S_2, S_3\ S_4$ all grounded i.e. zero current to op amp

For digital input 1
current is delivered to
op amp by the corresponding
resistor
For digital input 0
current from the corresponding
resistor is grounded



For 1111

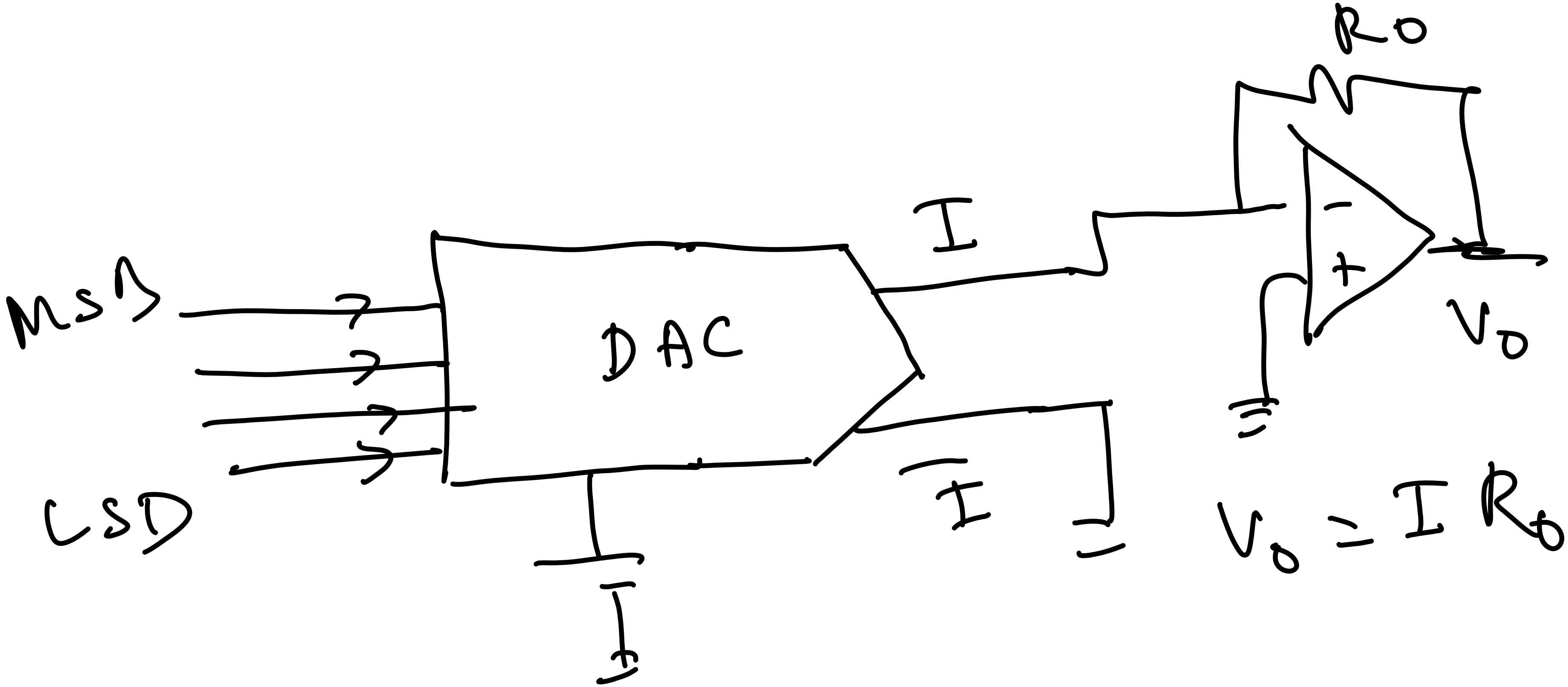
$$I = \frac{V}{4R} + \frac{V}{8R} + \frac{V}{16R} + \frac{V}{32R}$$

$$\bar{I} = 0$$

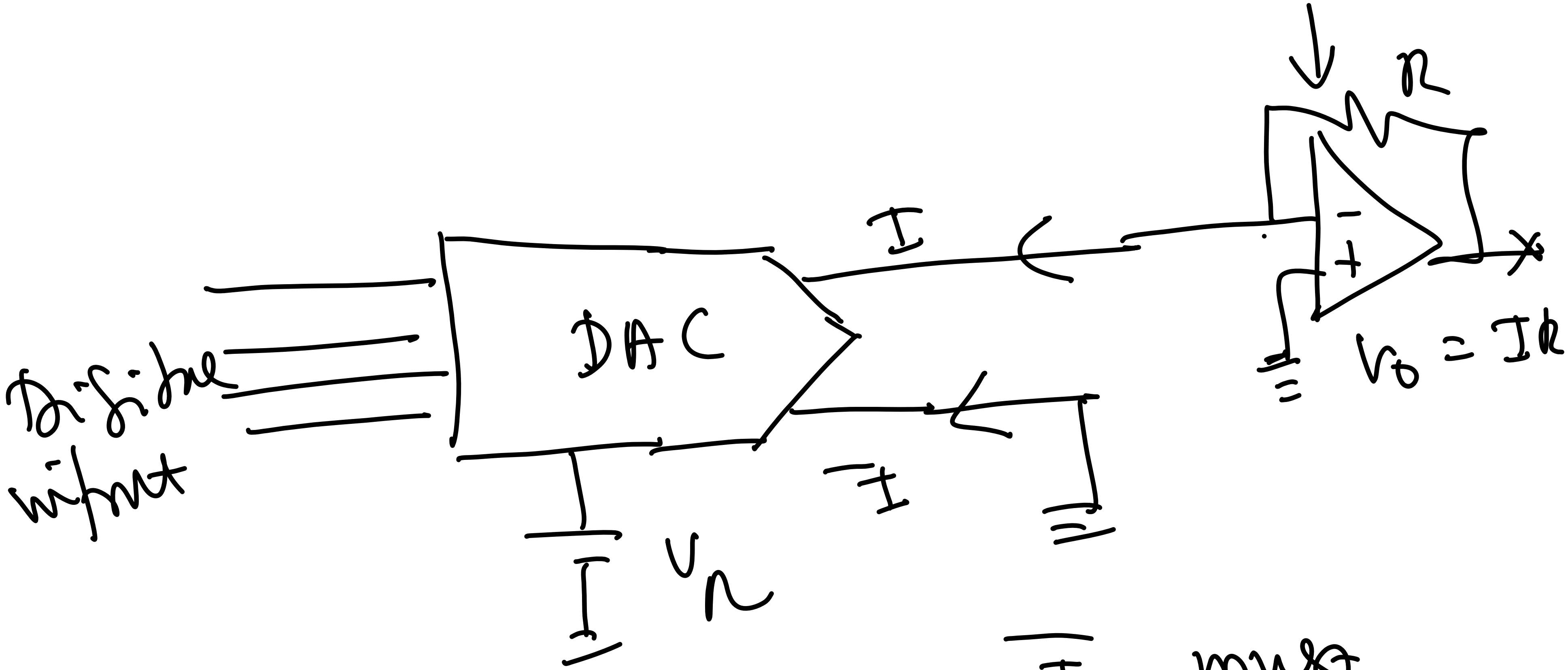
For 0000

$$\underline{I} = 0$$

$$\overline{I} = \frac{V}{4R} + \frac{V}{8R} + \frac{V}{16R} + \frac{V}{32R}$$

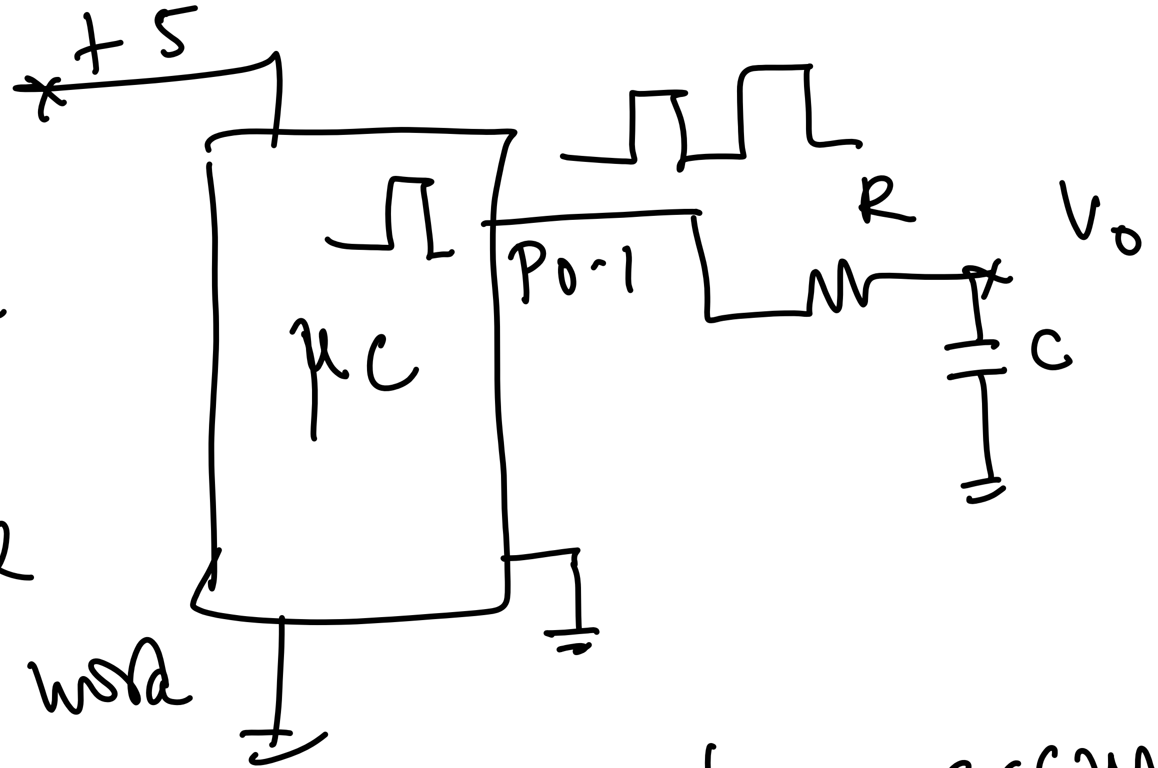


$$V_o = I R_o$$



Both I and $\bar{I}$ must see always ground potential

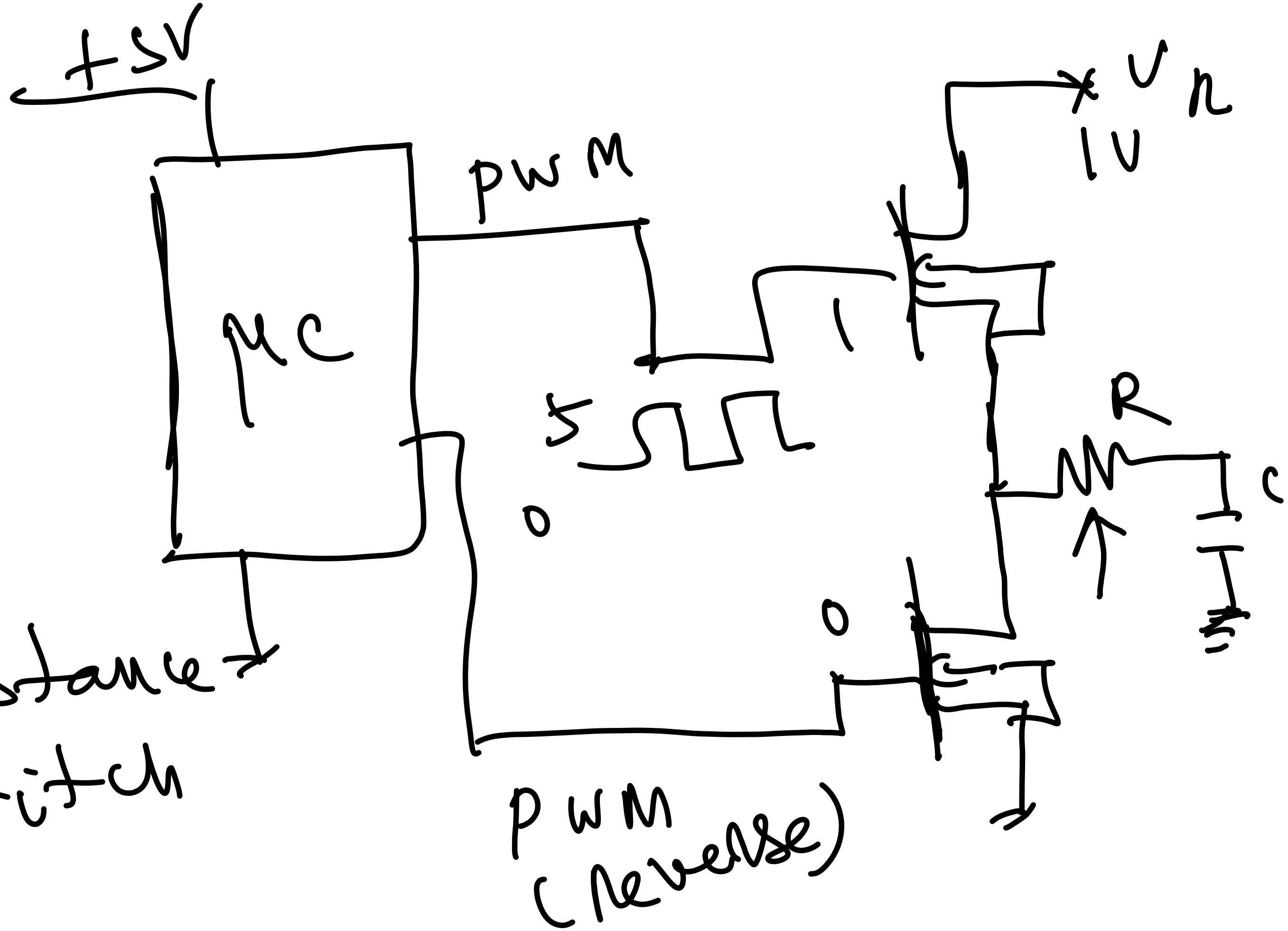
PWM based DAC



Pulse width must be proportional to digital word

The output may not be accurate because

1. Supply vol error
2. Switches may not be accurate (voltage loss in the switches may be on)



*: ON resistance of the switch must be very small compared to R

*: It is very slow DAC

However it is very simple cost effective DAC for μC systems

ADC

1. Dual Slope ADC
2. Successive approximation ADC
3. Flash ADC

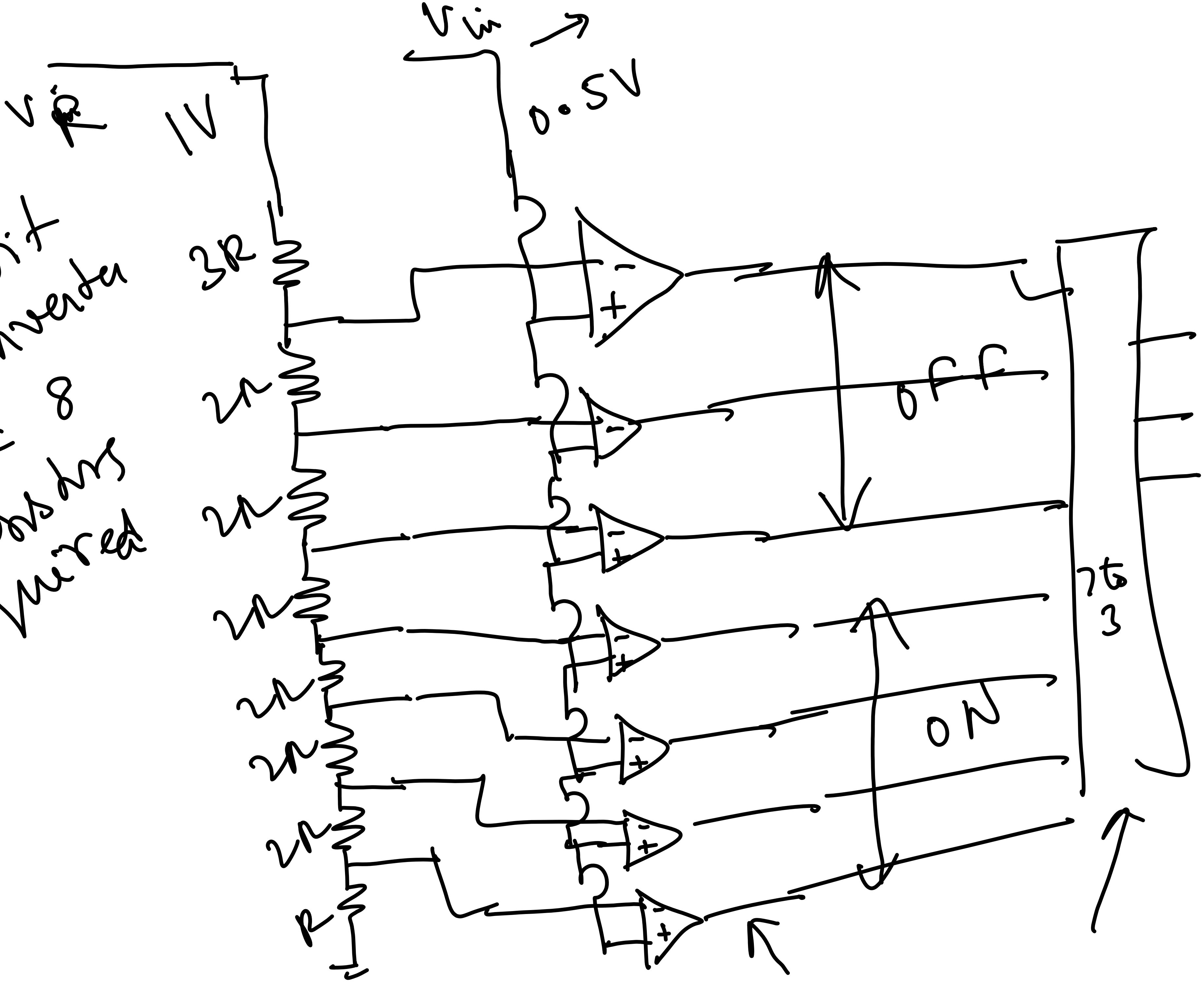
Dual Slope ADC $\longrightarrow$ 5 to 10 Convers per sec

SA/ADC $\longrightarrow$ 10 to 20 μ ~~Sec~~ Seconds

Flash ADC

3 bit Converter

$2^3 = 8$ resistors required



Conversion time is very small

Comparator response time + logic conversion time

<u>Advantage</u> → less conversion time

<u>Disadvantage</u> → large hardware

1. 2^n matched resistors
2. 2^n comparators

(3) Huge Conversion logic

For 8 bit Converter

(1) 255 resistors — matches

(2) 256 Comparators

For 16 bit Converter

(1) 2^{16} → 64 000 resistors

(2) 64 000 Comparators

(3) Conversion logic to convert 64 000 (2^{16}) → 16 bit output

$0.4 / 16$

For 1V → V_{ref}

16 bit Converter required

tolerance error of $= \frac{1}{2^{16}} \approx 16 \mu V$

Use two 8 bit Flash ADC

to get one 16 bit Flash ADC

This is called Subranging

ADC

Lecture - 39

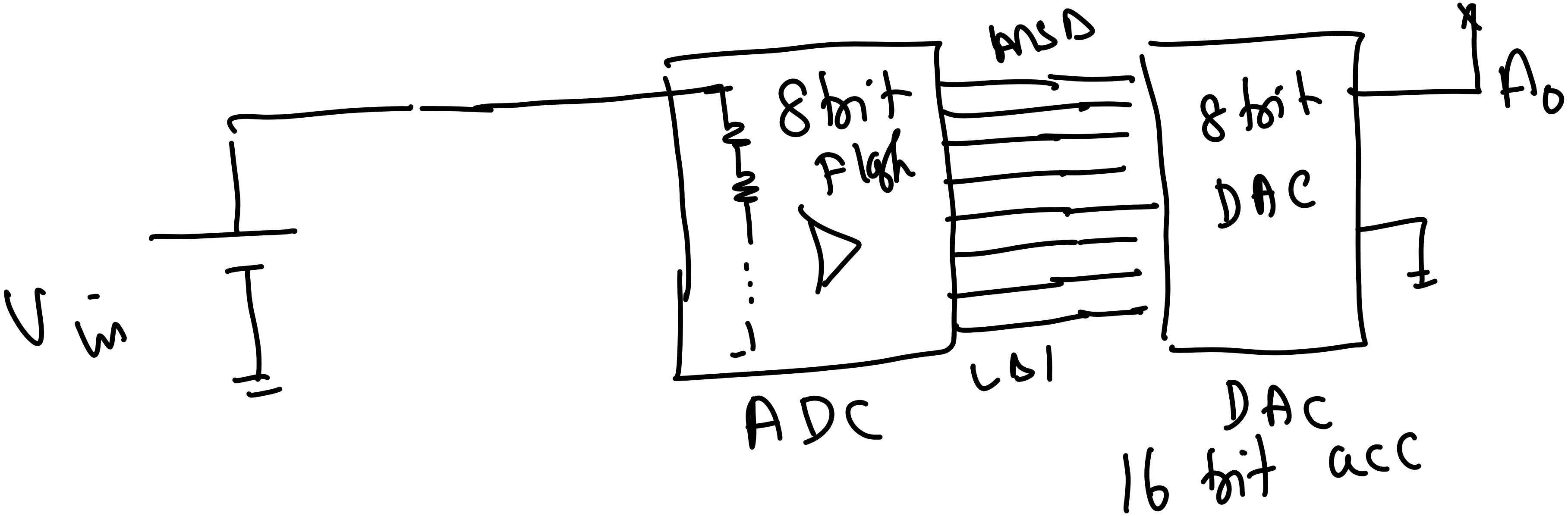
8 bit Flash ADC

1. 256 Comparators required
2. 256 Divider resistors

16 bit Flash ADC

1. 2^{16} Comparators
2. 2^{16} Divider resistors

Sub ranging ADC is used for 12 bit or 16 bit Flash ADC



8 bit DAC 8 bit acc → error will be 1 LSB of 8bit

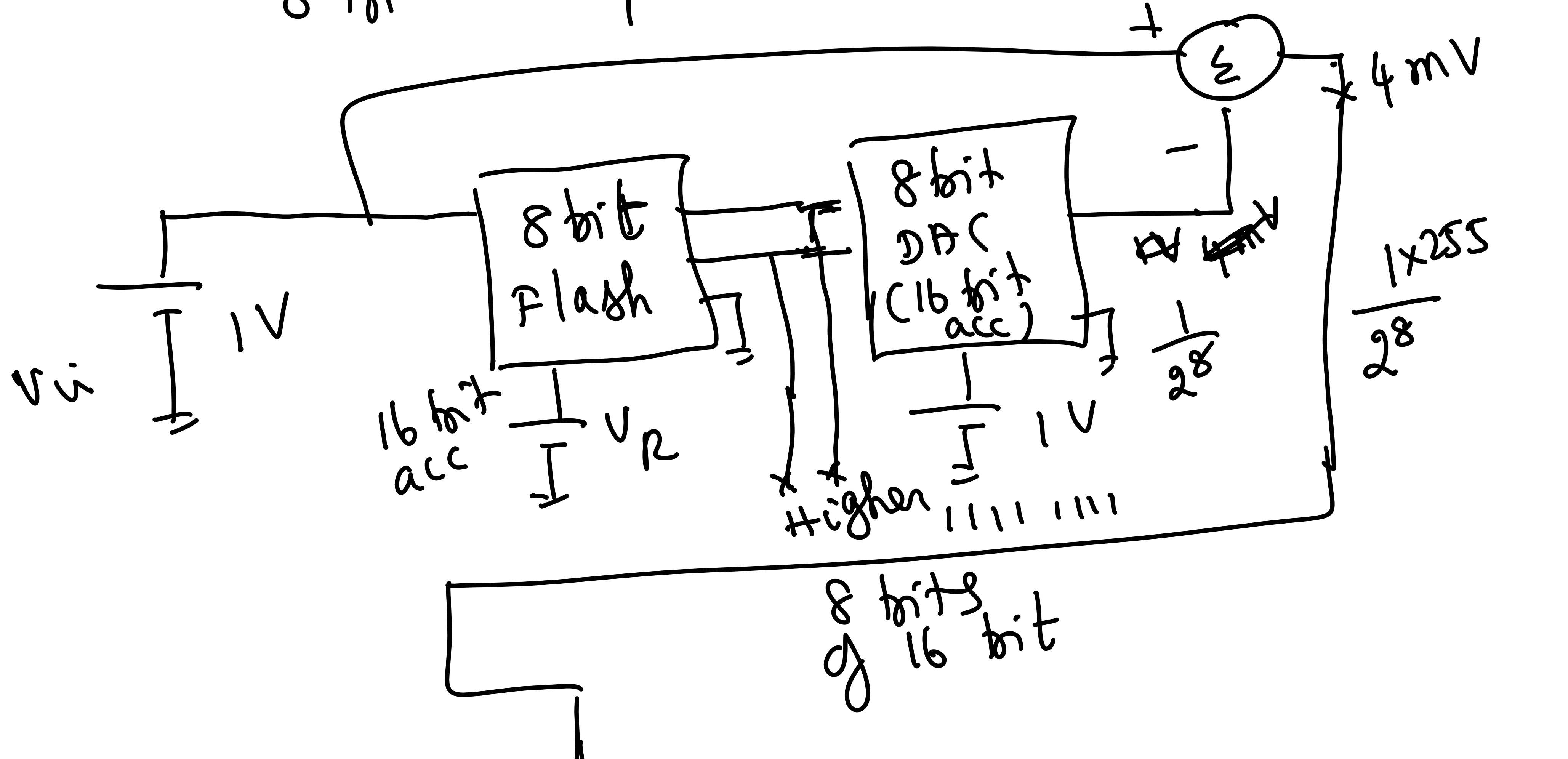
8 bit DAC 16 bit acc → 1 LSB of 16 bit

For 1 volt ref

8 bit DAC / 8 bit acc $\longrightarrow \frac{1}{256} \simeq 4\,mV$

For 1 volt ref

8 bit DAC / 16 bit acc — $\frac{1}{2^{16}} = \frac{1}{64000} \simeq 16\mu$

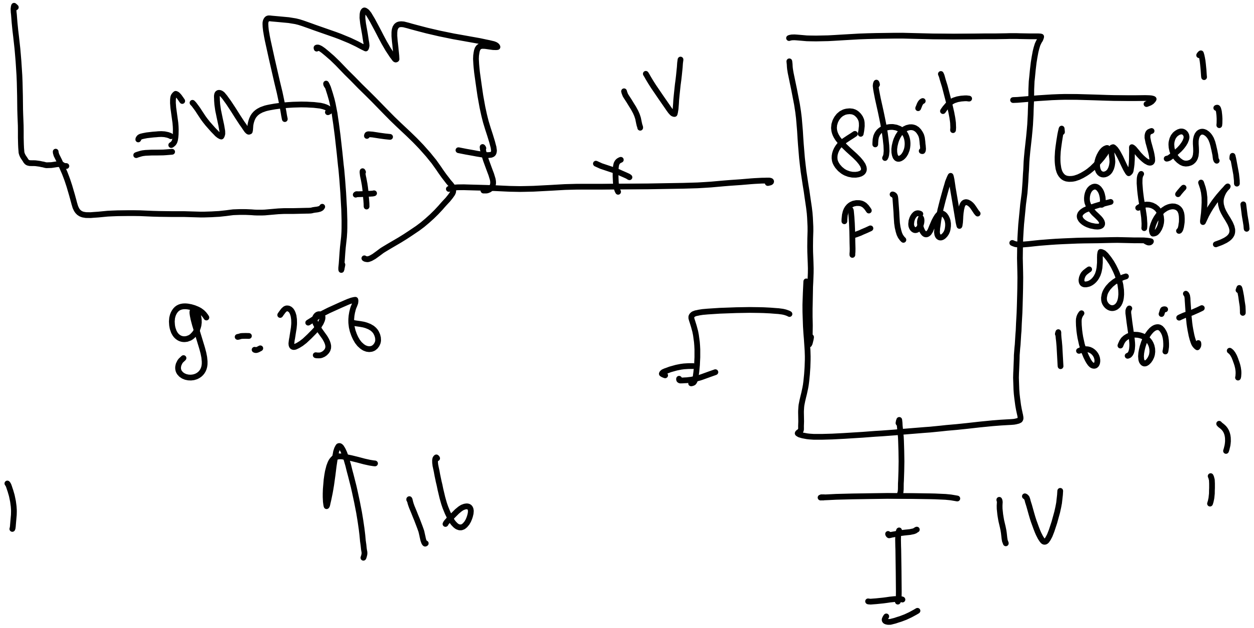


$V_{in} = 1V$

Digital

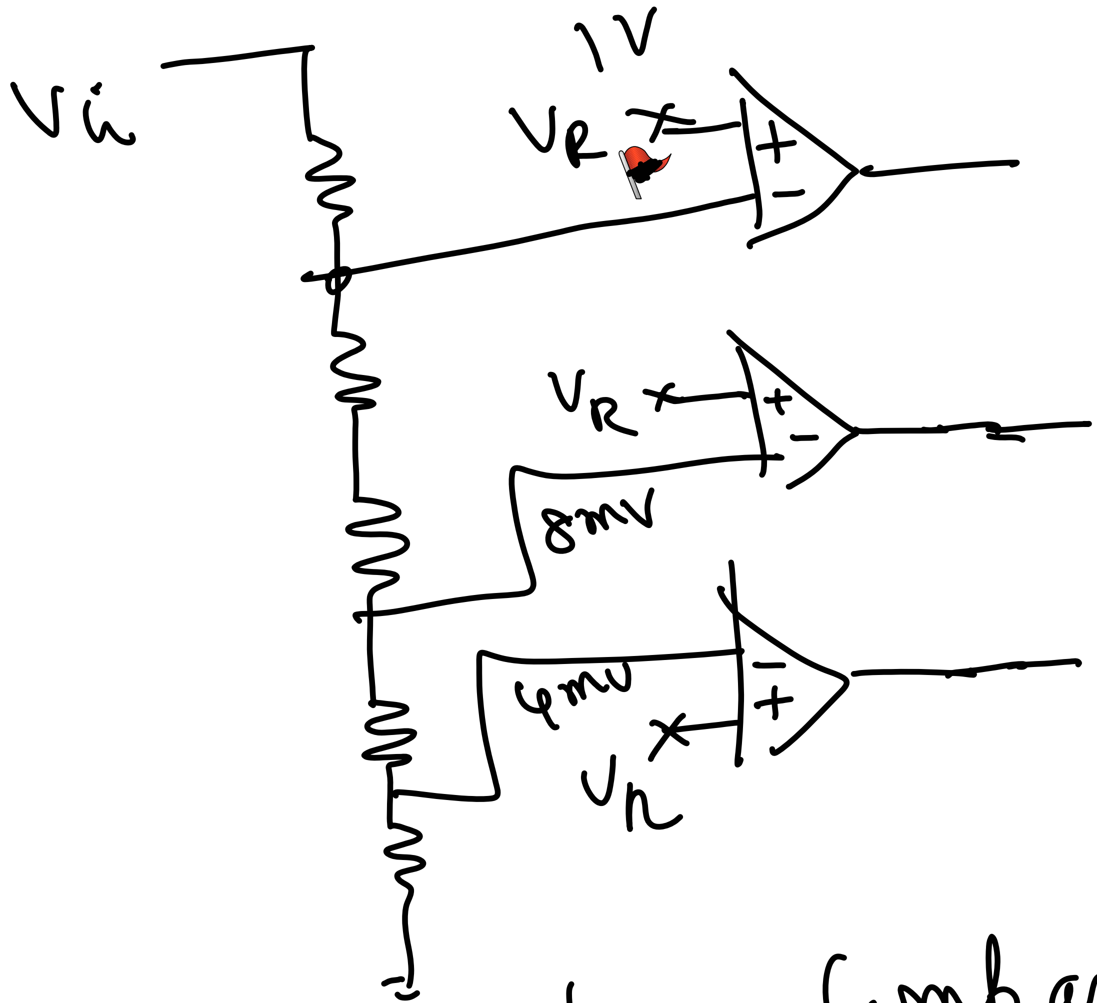
$V_o = 1111\ 1111$

$1111\ 1111$

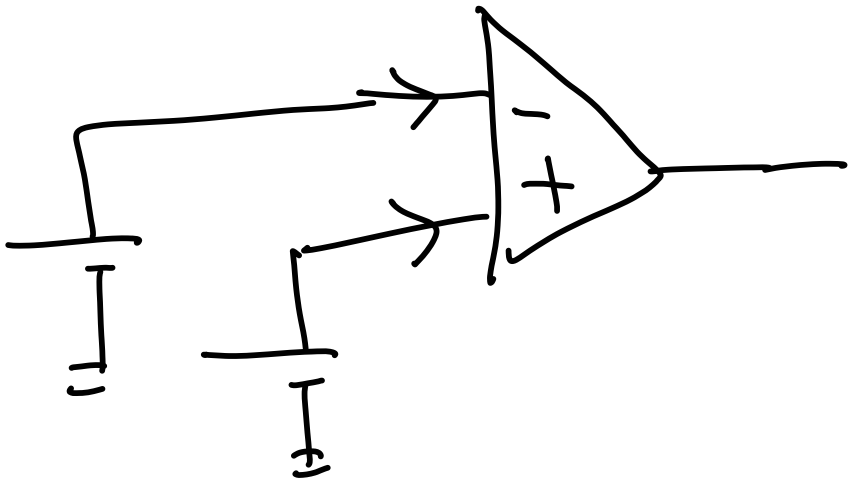


Adv → Very short Con time

Dis → Costly

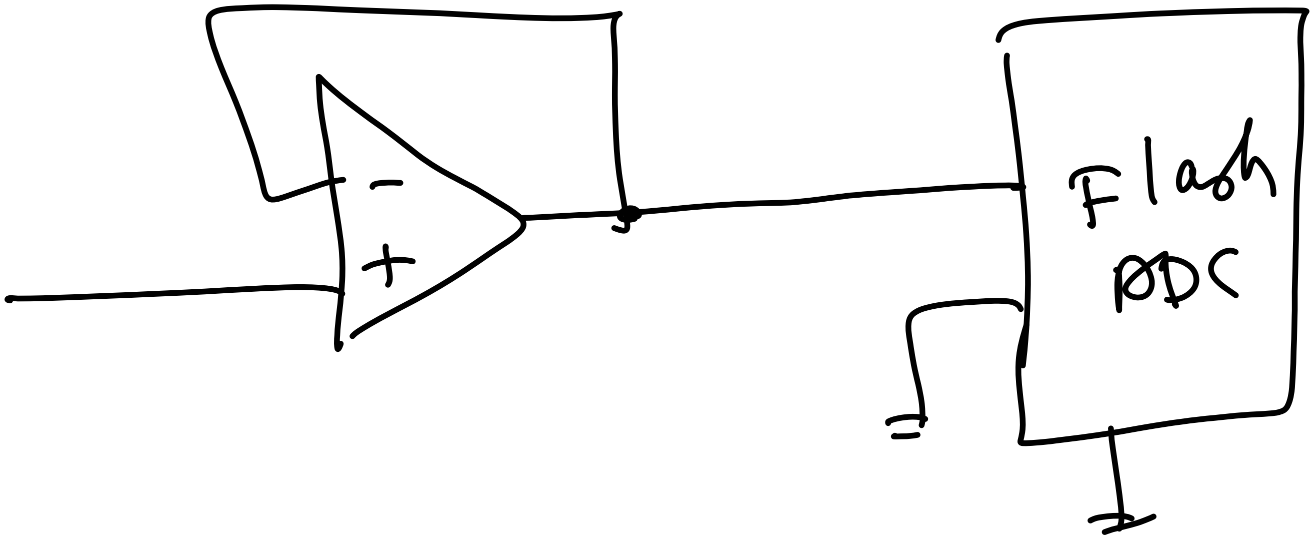


Always all the Comparators are in Saturation mode only



In saturation mode the input bias current is higher

Drive Flash ADC with very low output resistance source



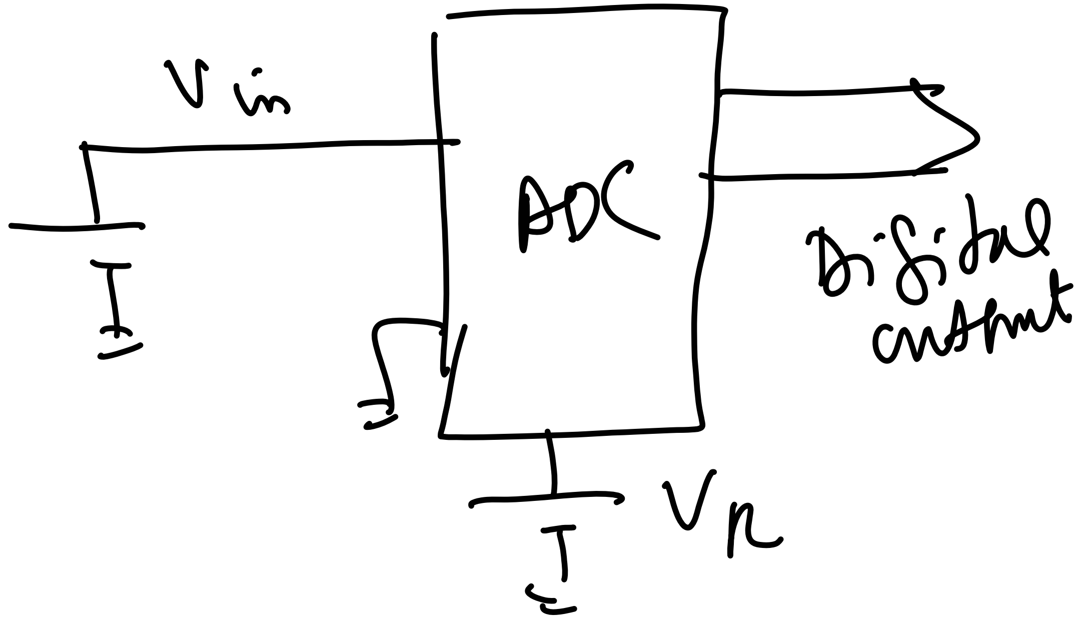
Always drive the Flash ADC with low output resistance

8–10 ① Dual slope ADC → display

1 μs ② Successive approx ADC

50 ns ← ③ Flash ADC
100 ns

Errors



Resolution = $\frac{1}{2^n}$

For 8 bit = $\frac{1}{256}$

① Reference vol drift

② OFF set vol drift

For $V_{in} = 0$ $V_0 \neq 0$

③ gain drift

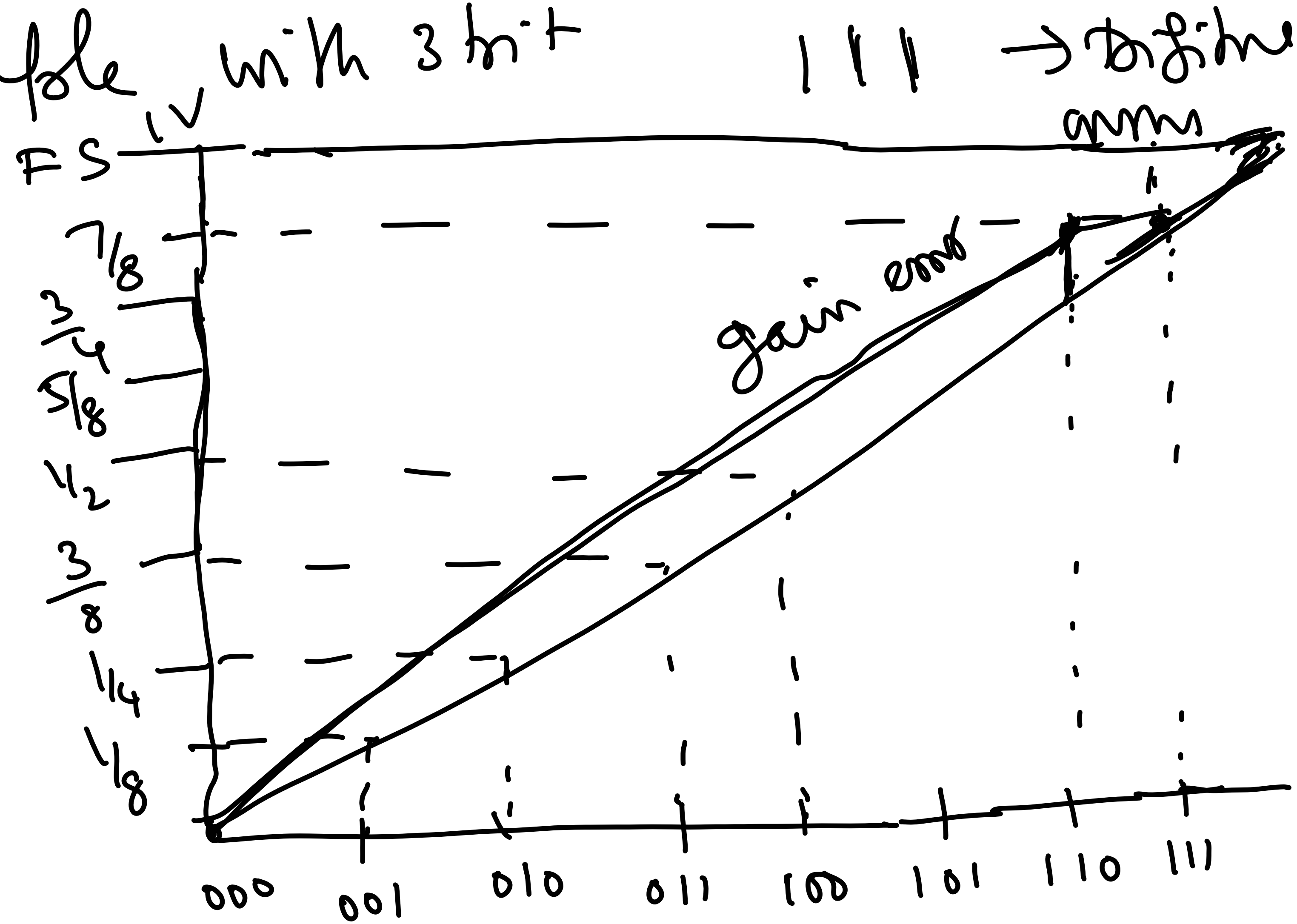
Temp related errors

The static errors (not related to temp)

1. gain error
2. offset voltage error
3. Linearity error
4. Differential linearity error

Example with 3 bit

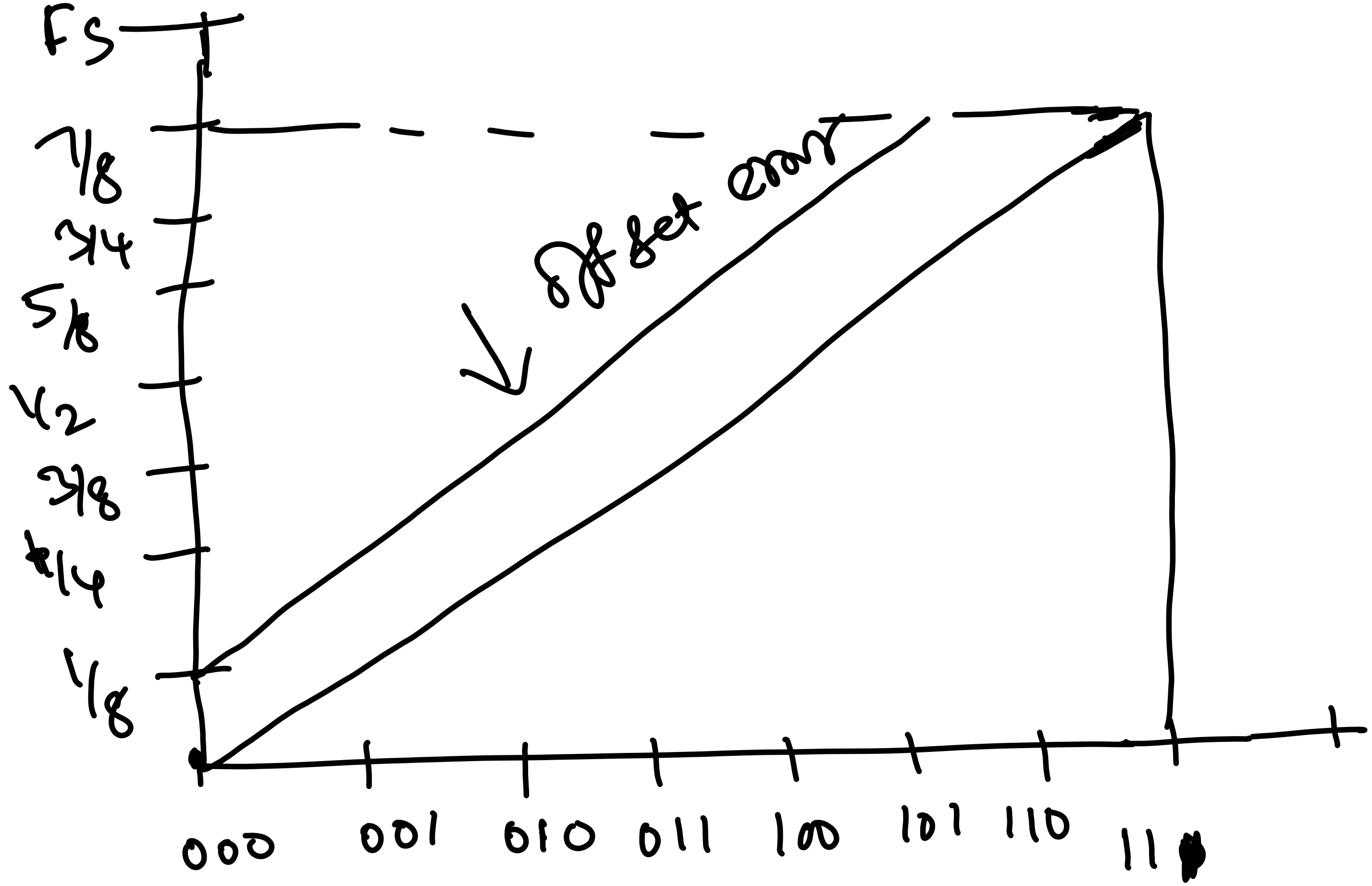
$V_R = 1V$



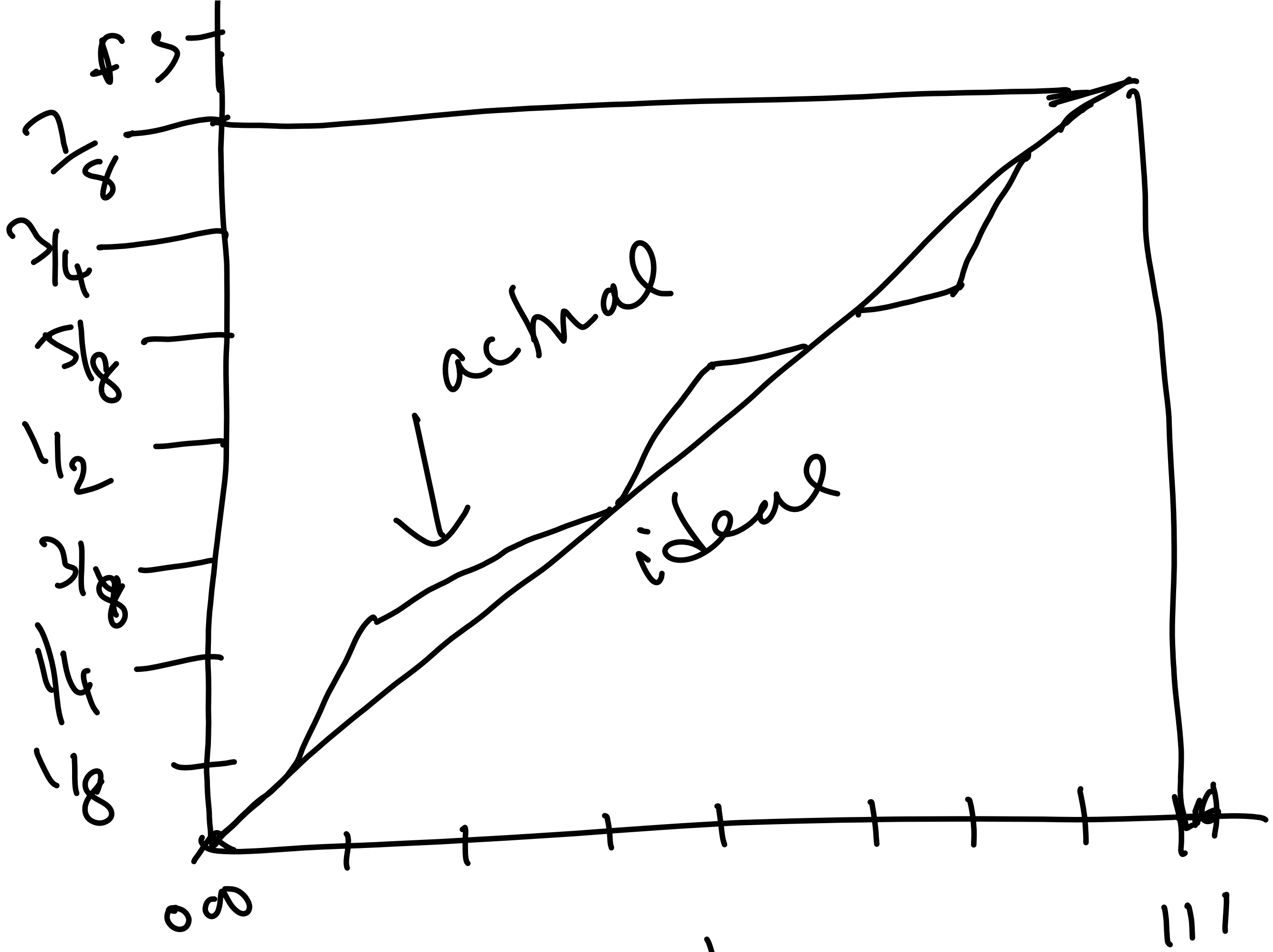
$\frac{7}{8}$ V/t at 111 → ideal

$\frac{7}{8}$ V/ at 110 → real value

OFFset error



Linearity error

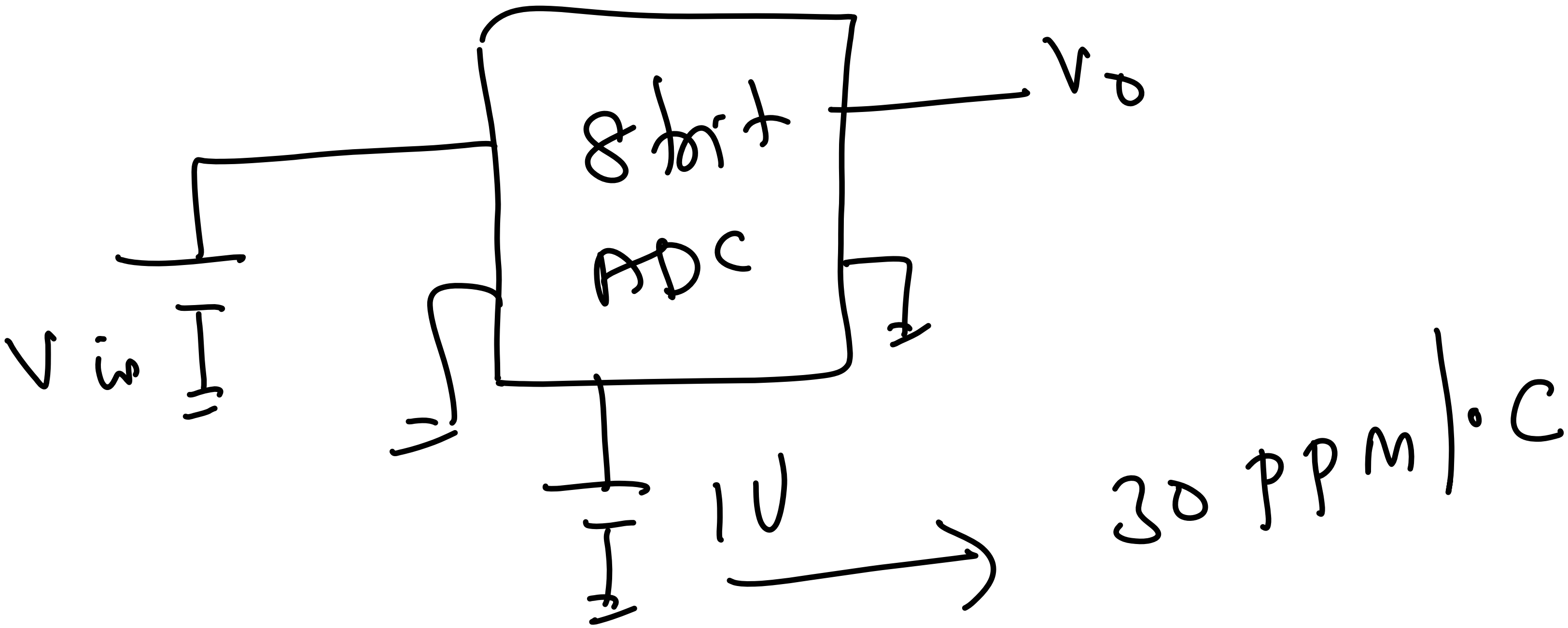


Differential linearity error

$$\left.\begin{array}{l} 110 \rightarrow \frac{3}{4} \text{ volt} \\ 111 \rightarrow \frac{7}{8} \text{ volt} \end{array}\right\} 1\,LSB$$

For differential non linearity the difference $\simeq$ 1 LSB

These 4 errors are correctable but temp related errors are not correctable


8 bit
ADC
Vo
Vin
1V
30 PPM/°C

For $\Delta T = 100^{\circ}C$

The V_R drift

$$= \frac{30 \times 100 \times 1}{10^6} = \frac{3000}{10^6} = 3\,mV$$

For 8 bit converter

$$1\,LSB = \frac{1}{2^8} \simeq 4\,mV$$

For 16 bit converter

$$1\,LSB = \frac{1}{2^{16}} \simeq 16\,\mu V$$

For 16 bit converter

total error $= \frac{3000}{16}$ LSBs

$\simeq$ 200 LSB

For 1 ppm/°C drift zener

the total drift $\simeq \frac{100}{10^6} \simeq 10^{-4}$ V

$= 100\,\mu V$

$\simeq$ 6 LSB

II OFFset drift error

offset w/ drift at the input of ADC $\rightarrow$ 1 to 5 µV/°C

For example

For $5\,\mu V/^\circ C$ drift case

For $\Delta T = 100^\circ C$

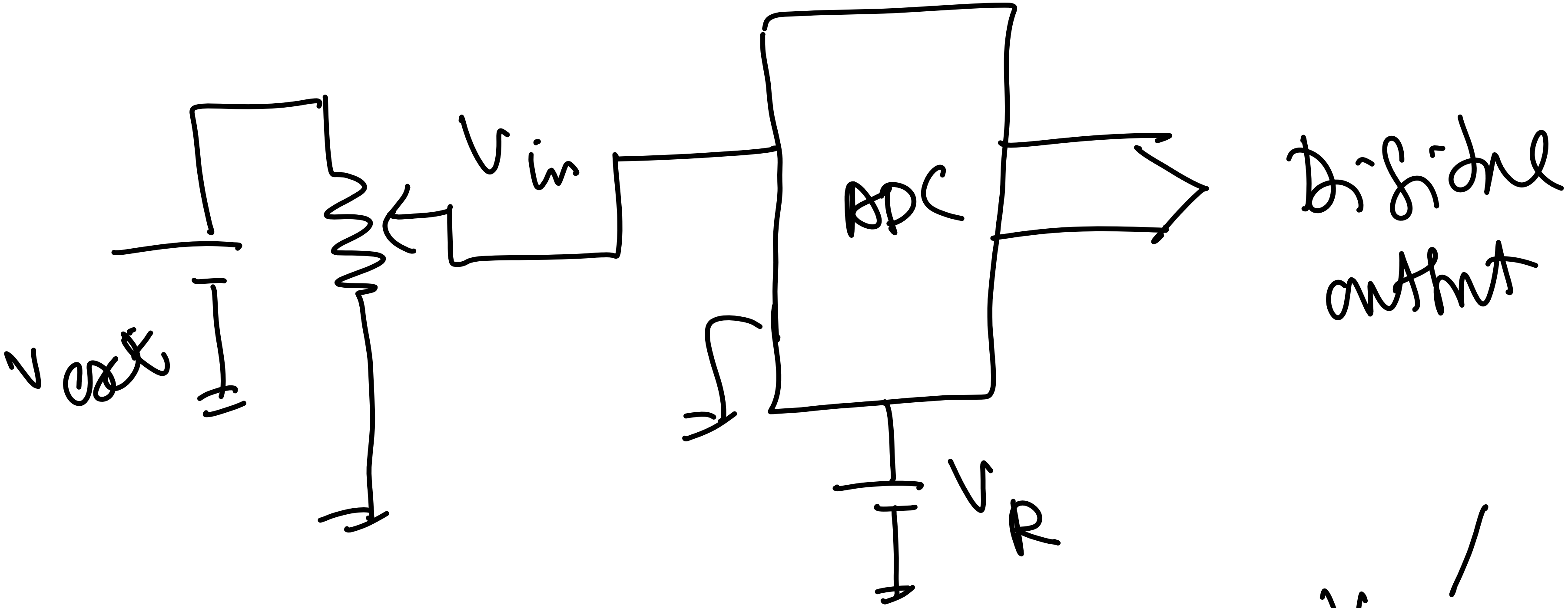
Then total drift $= 500\,\mu V$

For 16 bit converter

this drift $= \dfrac{500}{16} = 32$ LSB

1. For 8 bit converters
Resolution and accuracy will be more or less same
i.e. it is possible to get 8 bit accuracy against ambient temp variation
2. For 16 bit converters
it is difficult to get 16 bit accuracy.

Ratiometric conversion



$$\frac{\text{Digital output}}{\text{Full Scale Digital output}} = \frac{V_{in}}{V_R}$$

If V_R changes

then for given V_{in}

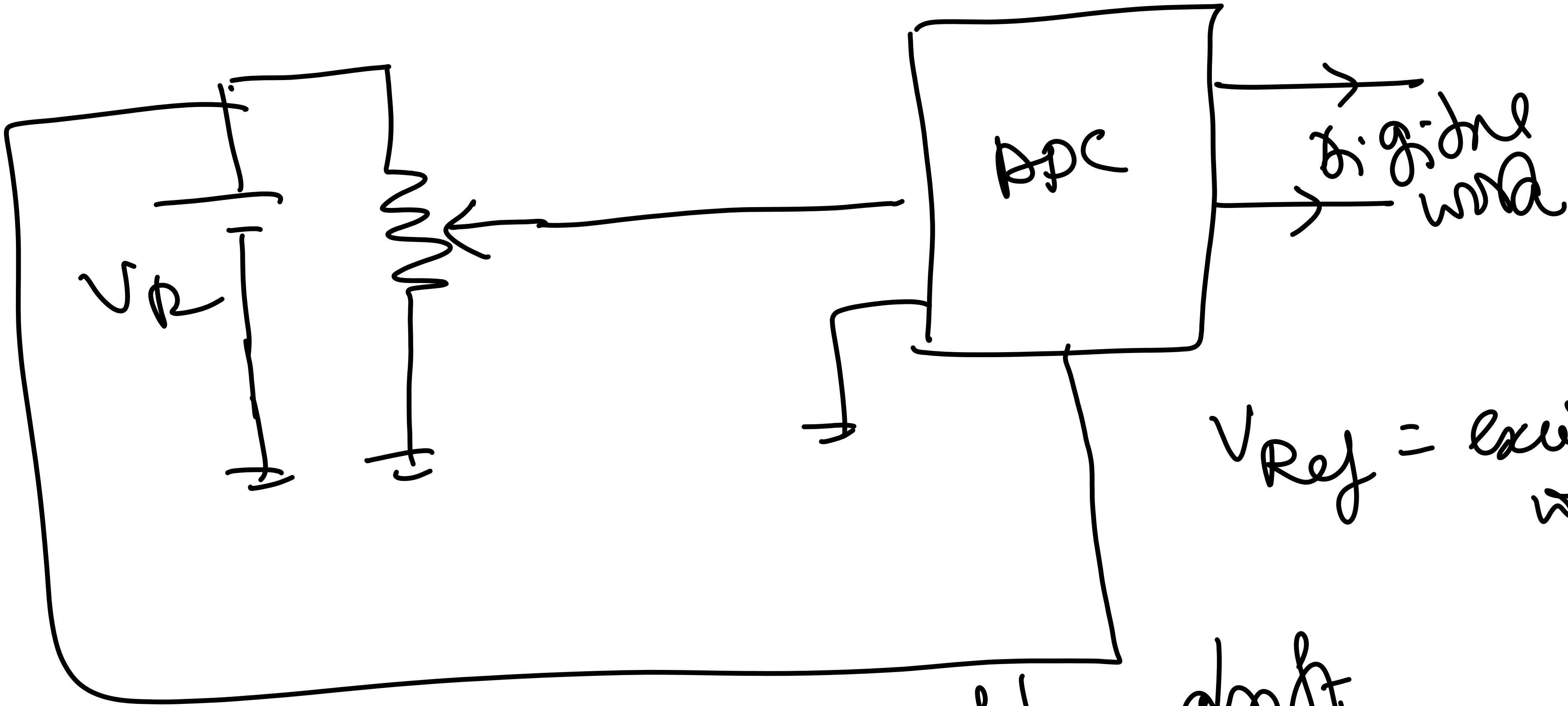
digital output is changing

V_{in} depends on the excitation

vol.

Make V_R as excitation voltage

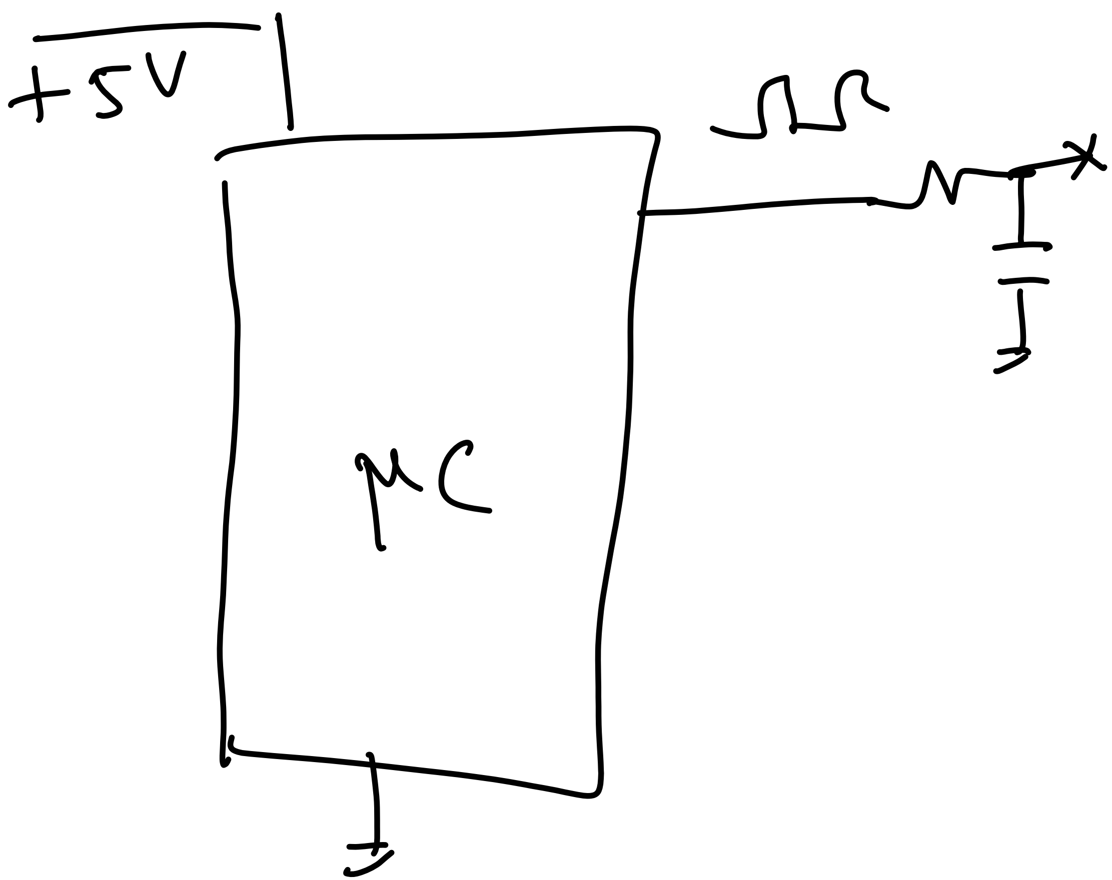
then no drift due to V_R



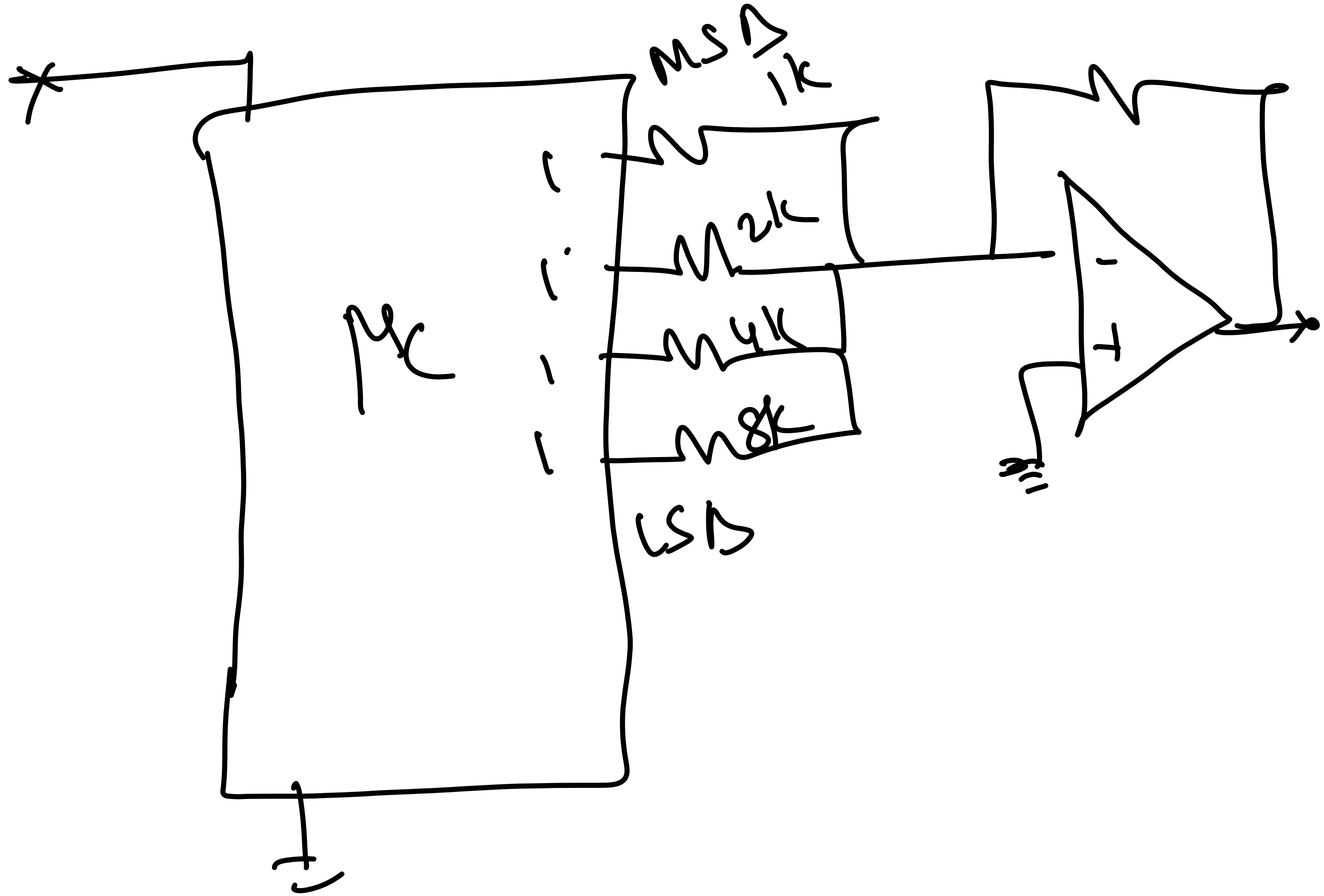
V_{Ref} = excitation wt

So reference voltage drift will have no effect on the output

DAC using μC



Binary weighted DAC using μC



Lecture no: 40

Σ - D ADC

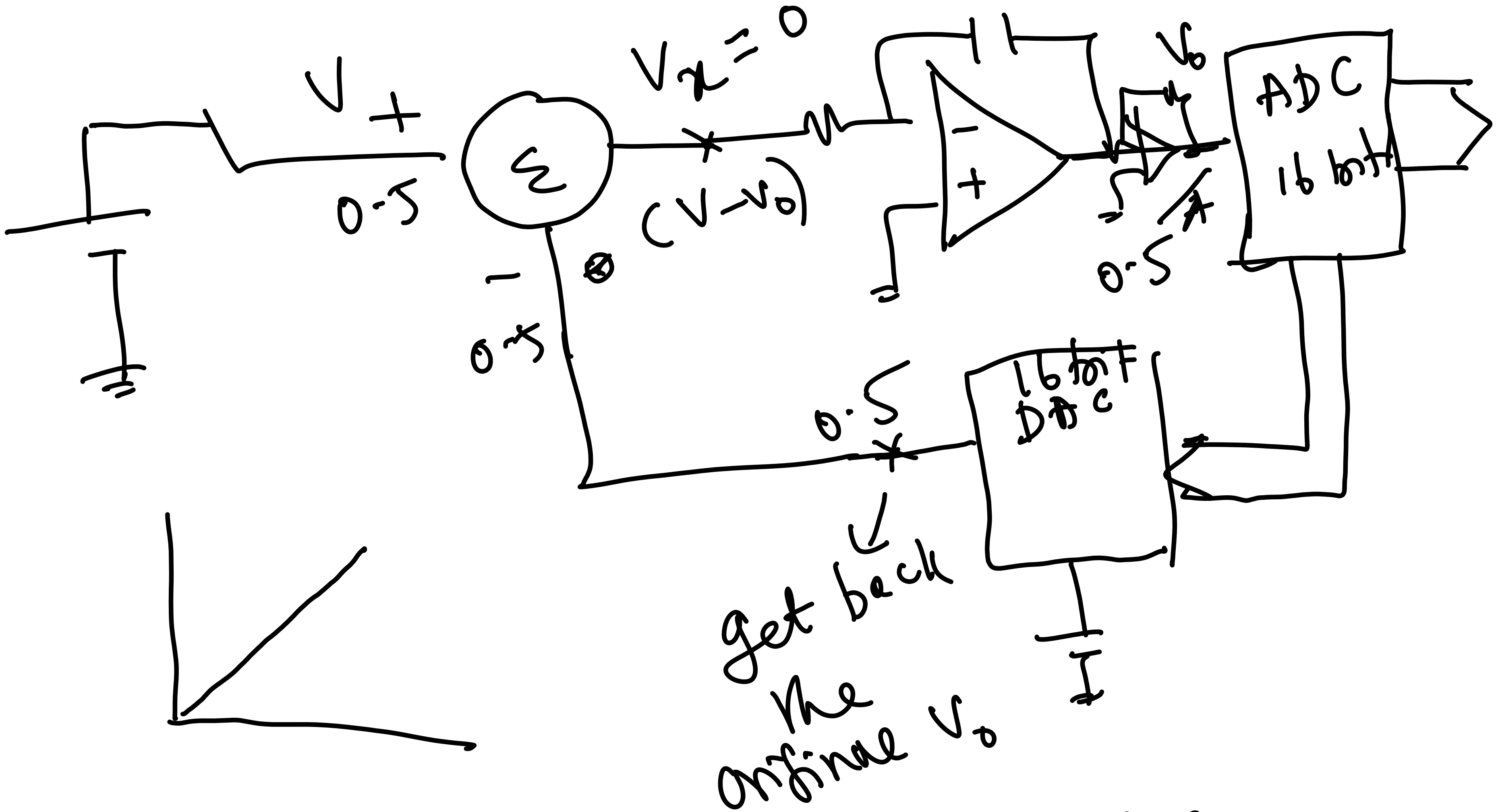
Dual slope - noise rejection
good
but slow

Successive app — 2 μs / bit
1 μs / bit

getting 24 bit converter
is difficult ~~be~~ cause we
need 24 bit DAC

When we need
say 24 bit or 32 bit Resolution
but satisfied with 16 bit
accuracy then we can look for
$\Sigma - \Delta$ ADC

$\Sigma - \Delta$ ADC



Assuming ADC and DAC reference is 1V

Assume that $V_{in} = 0.5V$

Then if $V_0 = 0.5$ at some time

Then ADC output

$= 1000\ 0000\ 0000\ 0000$

$=$

DAC output $= 0.5V$

This makes adder output $= 0$

So integrator will stop charging.

So V_0 remains constant

ADC and DAC are also give fixed output

If $V_{in} \neq 0.5$

$V_{in} \simeq 0.5 + 10\,\mu V$

For 16 bit converter

$1\,LSB = \frac{1}{2^{16}} = \frac{1}{64000}$ V

$= \frac{1}{64} \times 10^{-3} = \frac{1000}{64}\,\mu V$

$\simeq 16\,\mu V$

$V_{in} = 0.5V \longrightarrow$ i.e ADC input

is 0.5V then

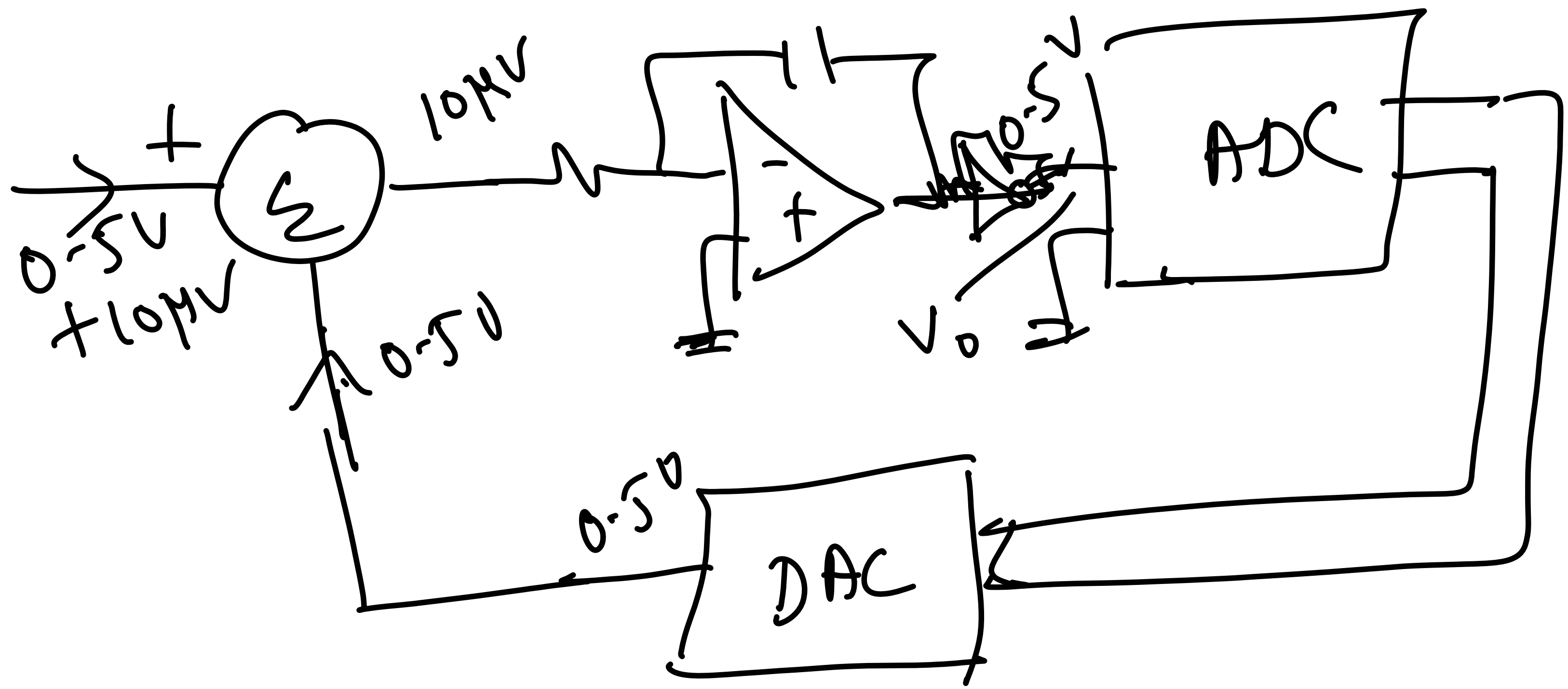
its output will be 1 000 0 000
0000 0000

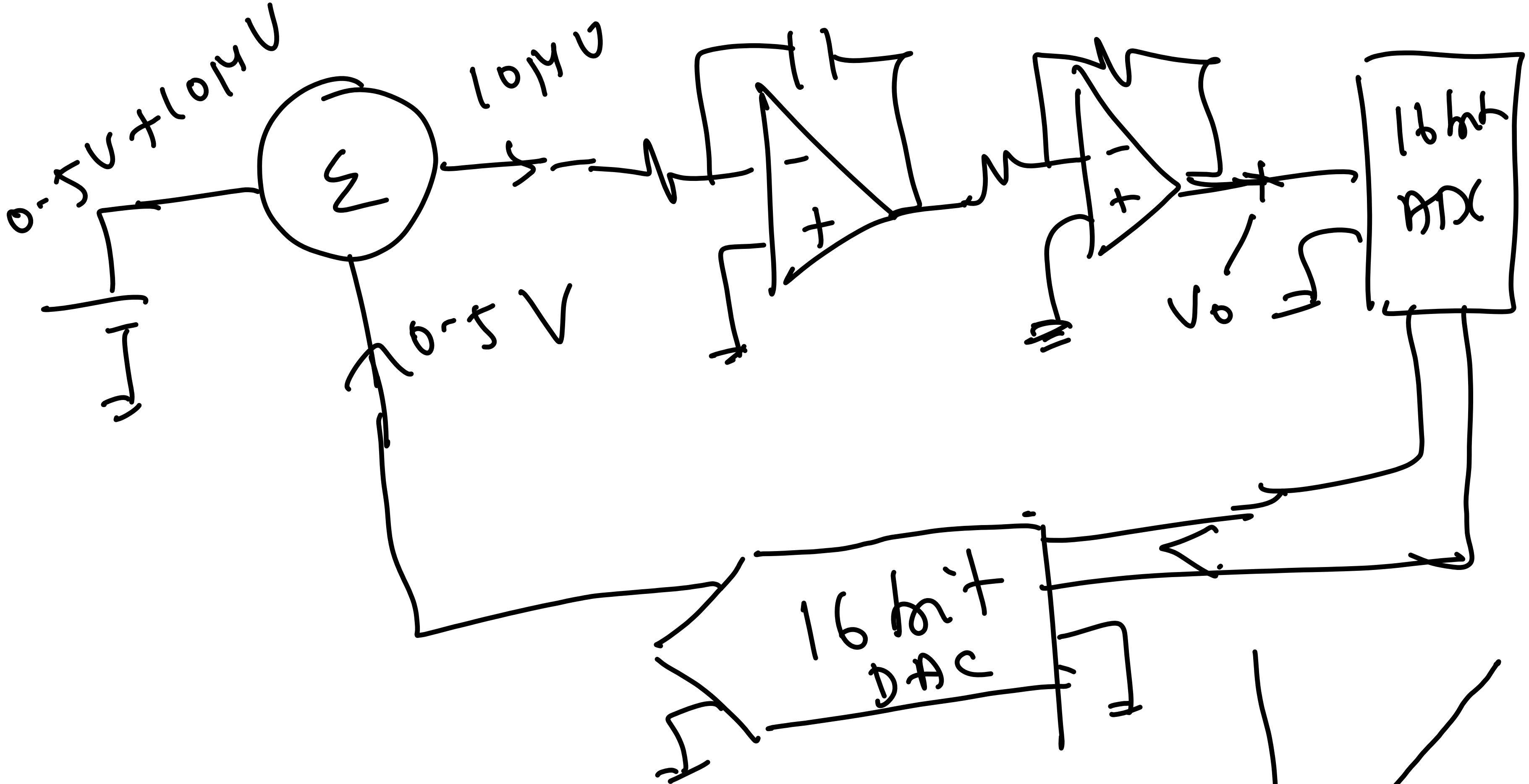
$V_0 = 0\text{-}5 + 10\ \mu V$

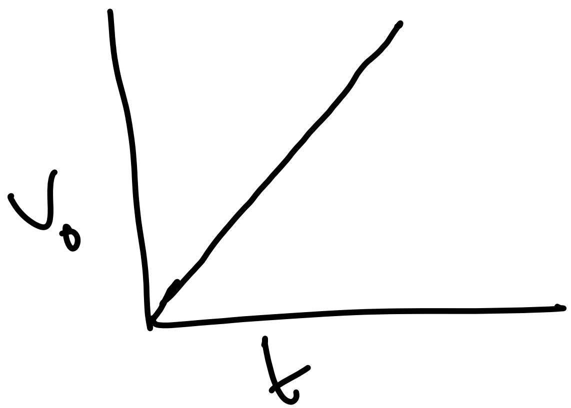
Then

ADC = 1000 0000 ~~000~~ ~~00~~G

DAC = 0-5V



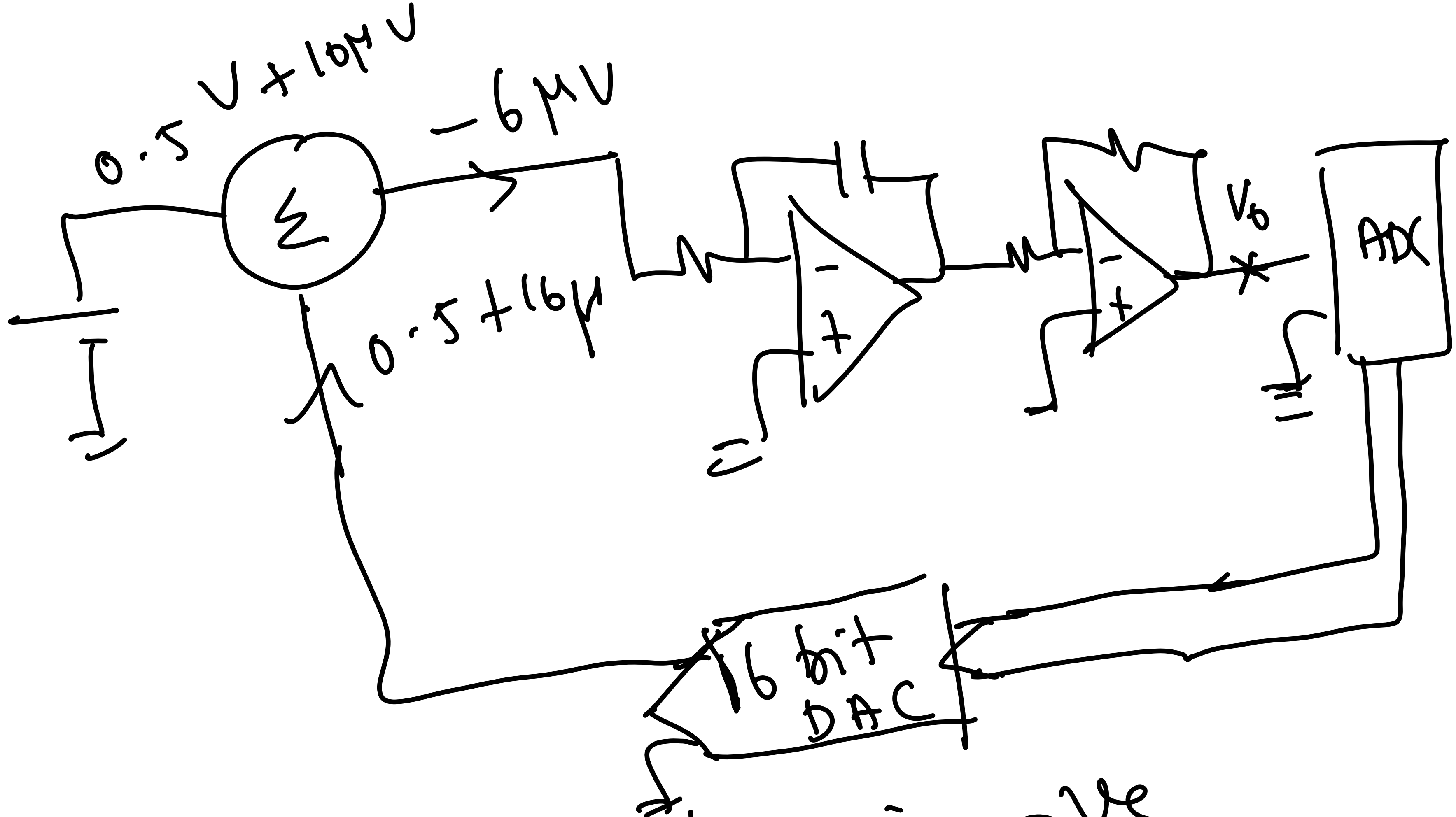




When v_o becomes $0.5V + 16\mu V$

ADC $\longrightarrow$ 1000 0000 0000 0001 =

DAC $\longrightarrow$ $0.5 + 16\mu V$



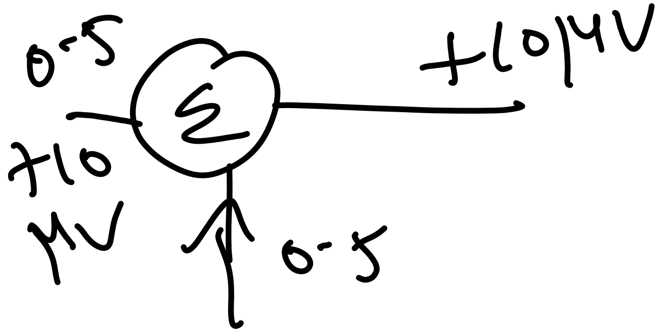
Integrator input is −ve
So integrator started discharging
So Vo Started decreasing

When V_o becomes $0.5\,V$

Then ADC → 1000 0000 0000 0000

DAC → $0.5\,V$

Then



That is ~~it~~ will make integrator to charge again. So V_o will start increasing

ADC → 1000 0000 0000 0001

DAC → $0.5 + 16\,\mu V$

ie ADC → 1000 0000 0000 0001
0000
0001
⋮

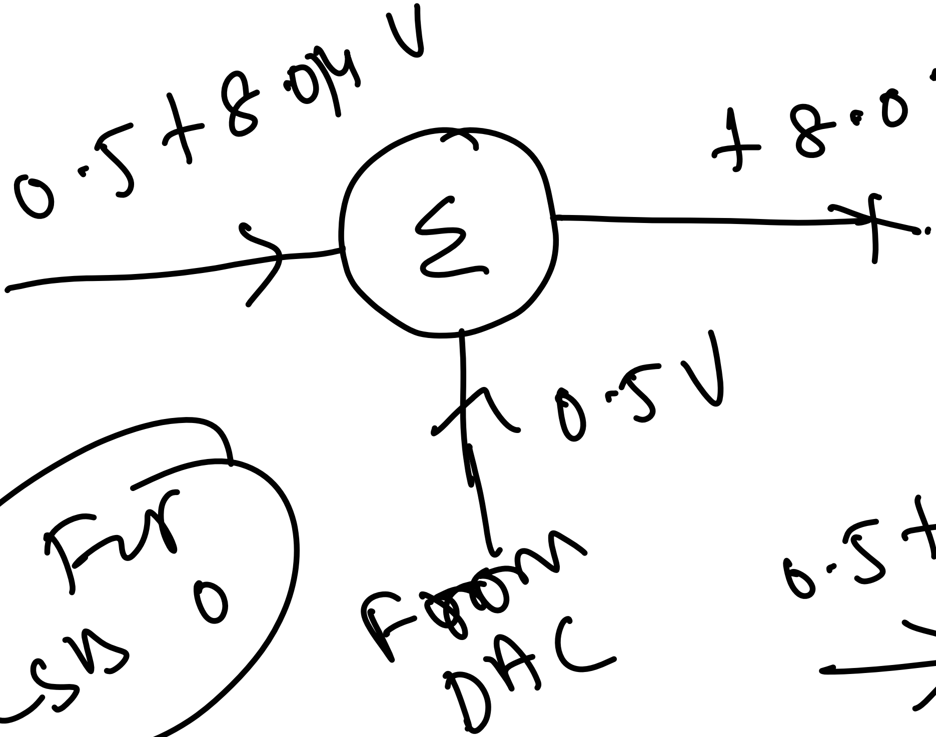
What is ADC output in time domain?

For 100 samples how many LSB's will be 1? and how many LSB's will be 0?

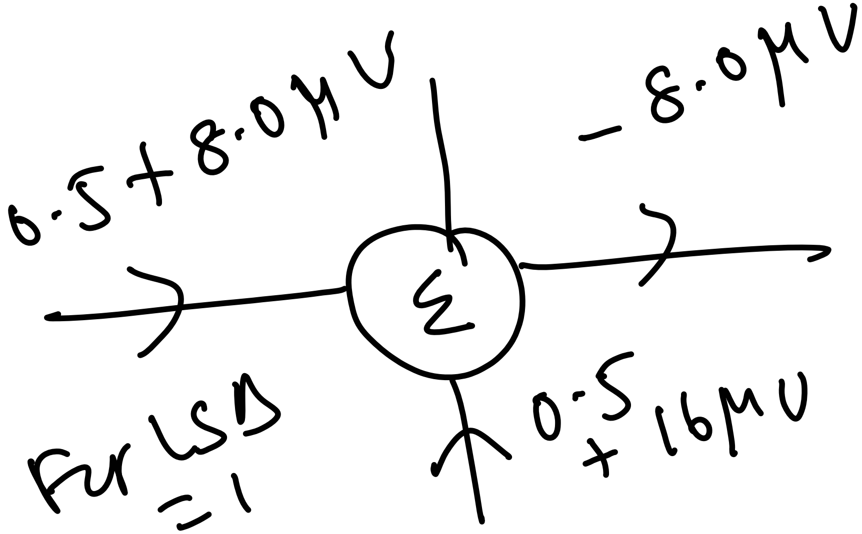
If the input is 0.5V Then All 100 LSB's will be zero

If the input is
0.5V + 8μV

Then 50 LSBs 1 } array
50 LSB's 0 } 100

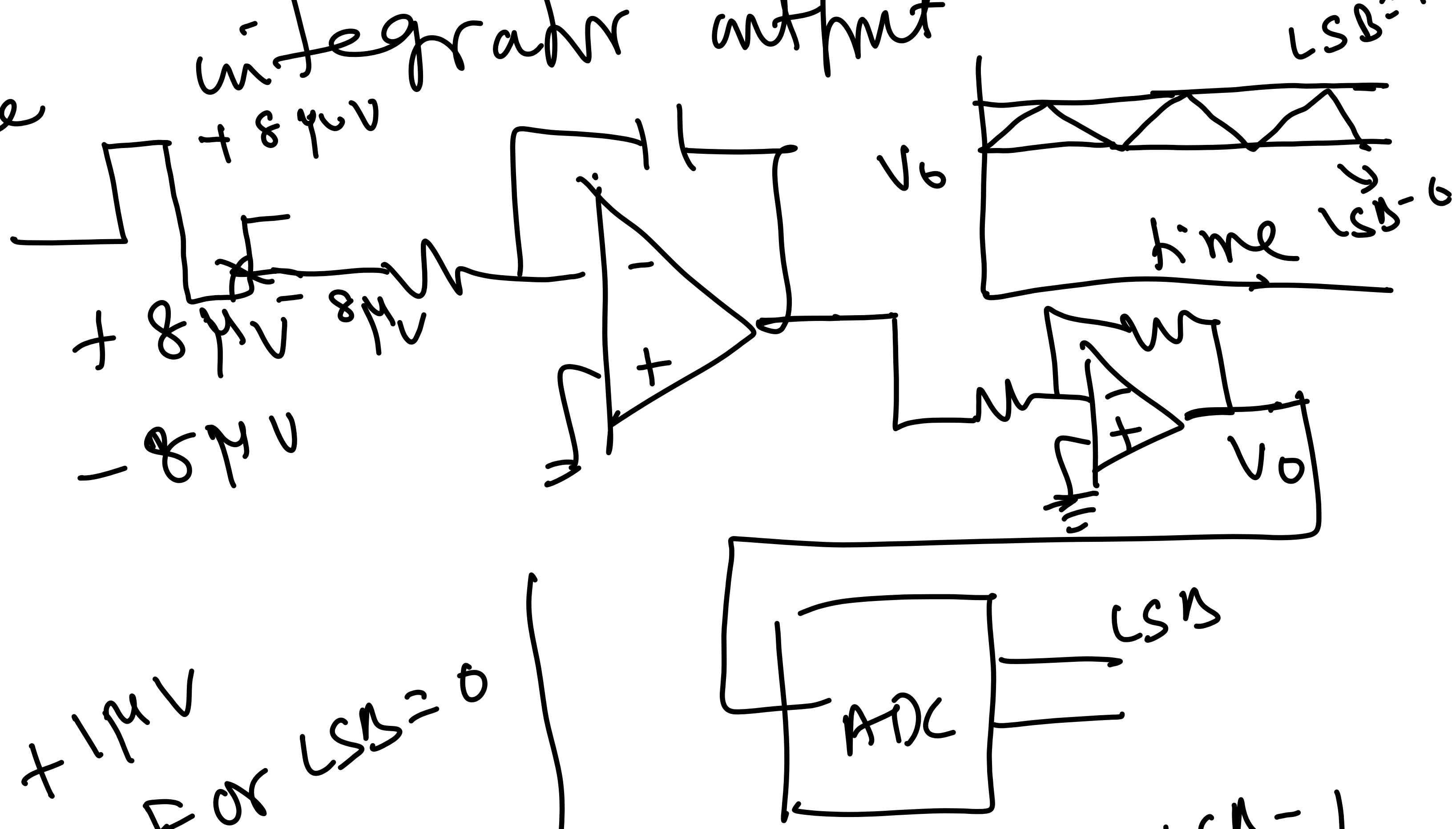


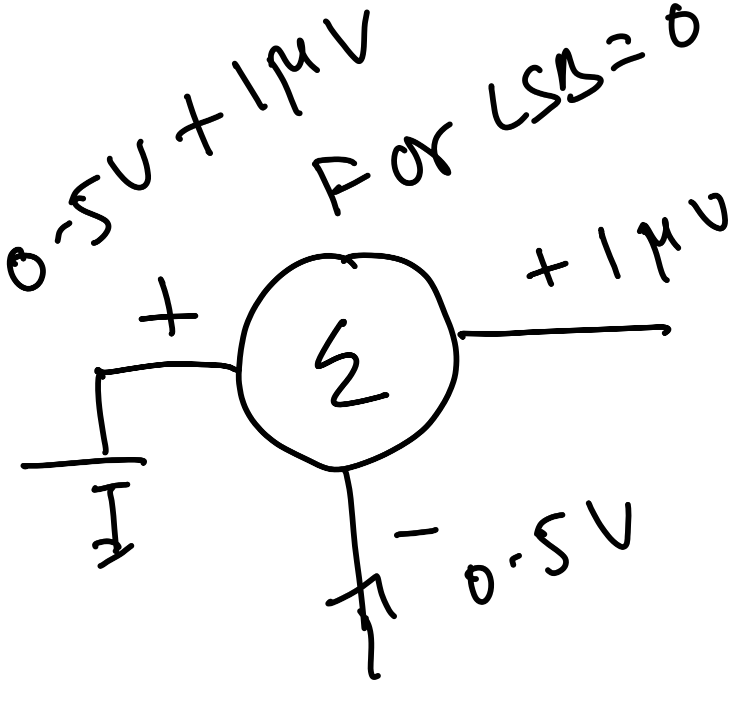
For LSB 0

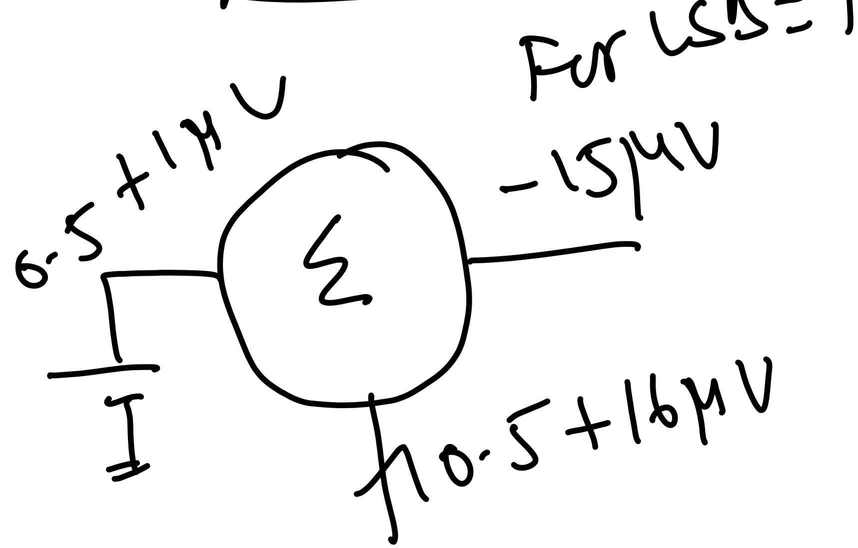


For LSB =1

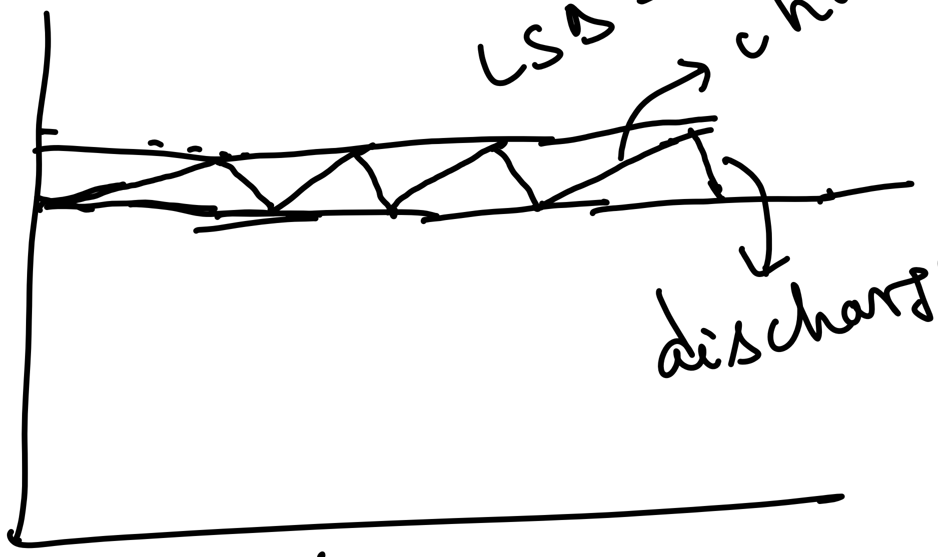
ie integrator output





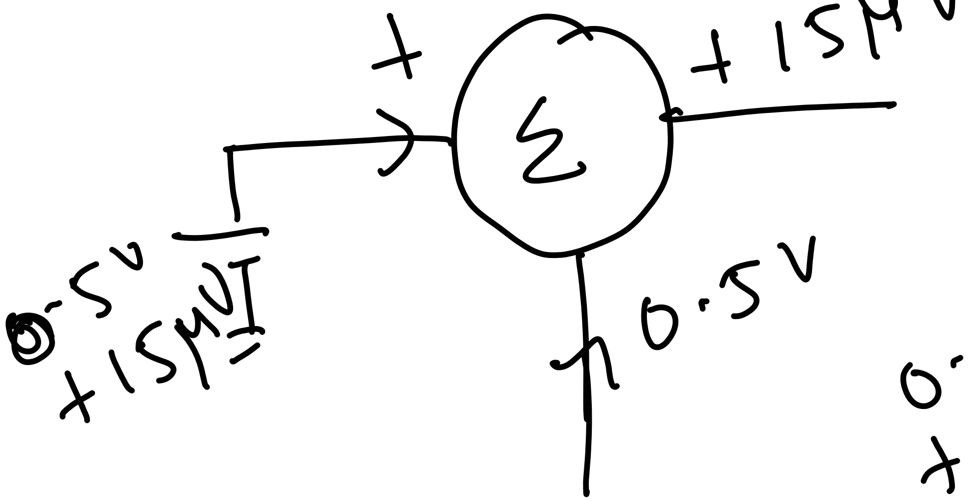


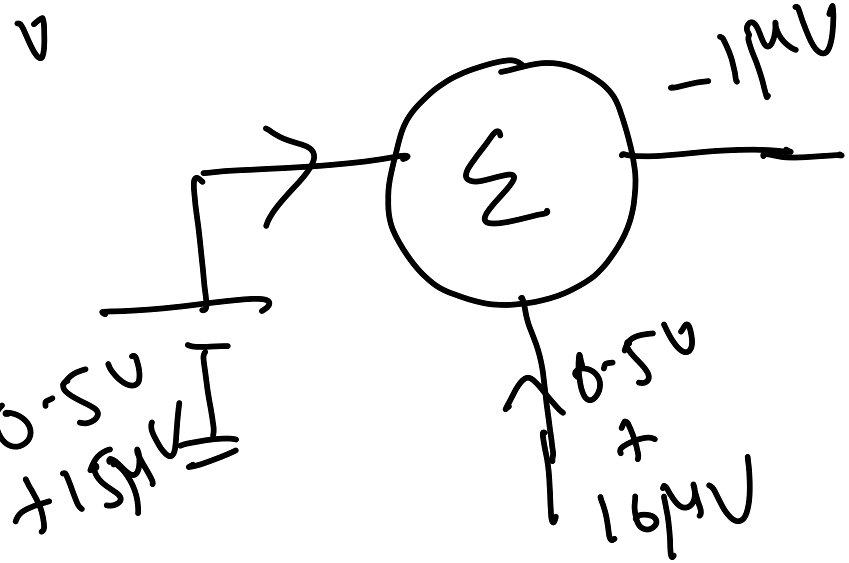
For this Int
discharge is faster than
charging



LSB 1
will be for
less no of time
and LSB 0 will be obtained for more no. of
times

For $V_{in} = 0.5 + 15\mu V$





LSB = 0 LSB = 1

Now charging is fast and discharging is slow

For more no. of times
LSB is 1
less no. of times LSB = 0

Ratio between
LSB = 1 and LSB = 0
is same as ratio between
$\frac{\text{fraction } V_{in}}{16}$ and 16/4V

For example

for $V_{in} = 0.5 + 1\mu V$

Then the fraction is $\frac{1}{16}$

For $V_{in} = 0.5 + 5\mu V$

the fraction is $\frac{5}{16}$

The ADC LSB = 1 and LSB = 0

ratio also will be as as

$\frac{1}{16}$ and 5/16

over sample the ADC and find out n and $n+1$ at the output

$$\frac{n}{n+1} = \frac{1}{16} \quad \text{for } 0.5 + 1\,\mu V$$

$$\frac{n}{n+1} = \frac{5}{16} \quad \text{for } 0.5 + 5\,\mu V$$

$$\frac{n}{n+1} = \frac{8}{16} \quad \text{for } 0.5 + 8\,\mu V$$

By this way ADC resolution can be increased by over sampling using DSP over sampling and the required calculations are carried out.

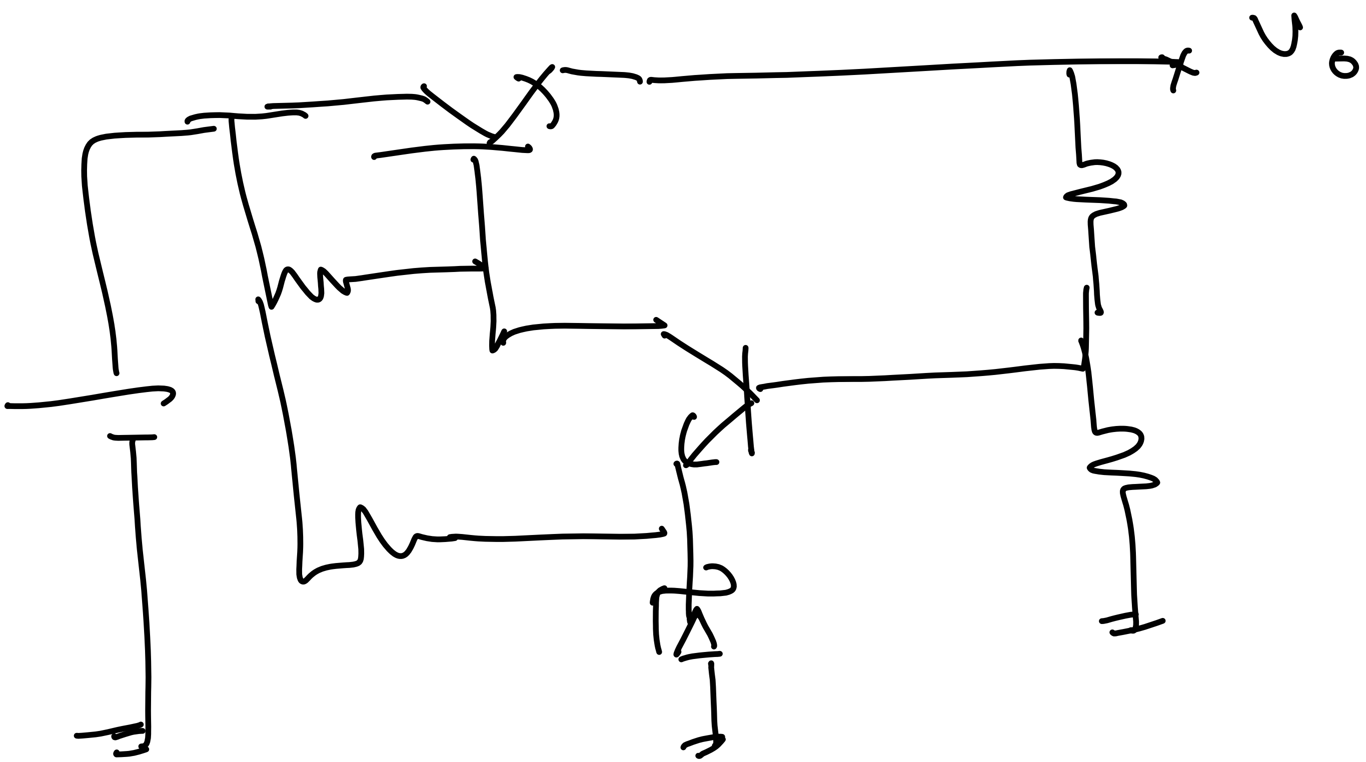
So one can get resolution more than 16 bit using 16 bit ADC and DAC but accuracy is limited to 16 bit only.

① Transistor applications

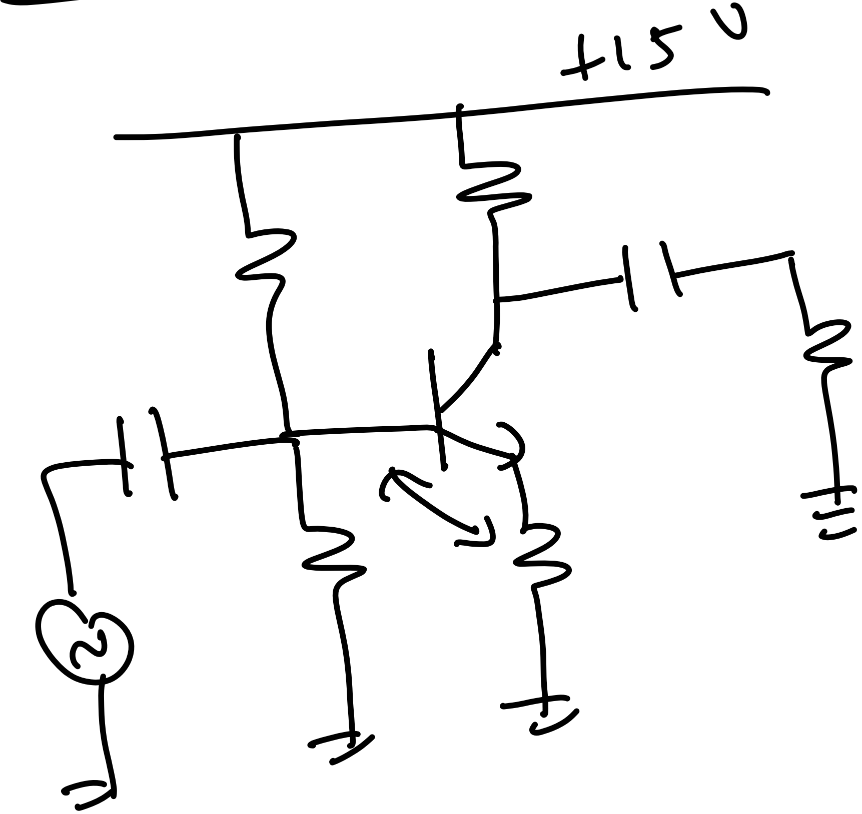
↳ as a switch → $\beta \gg 1$, → $\beta < 1$

as a Latch

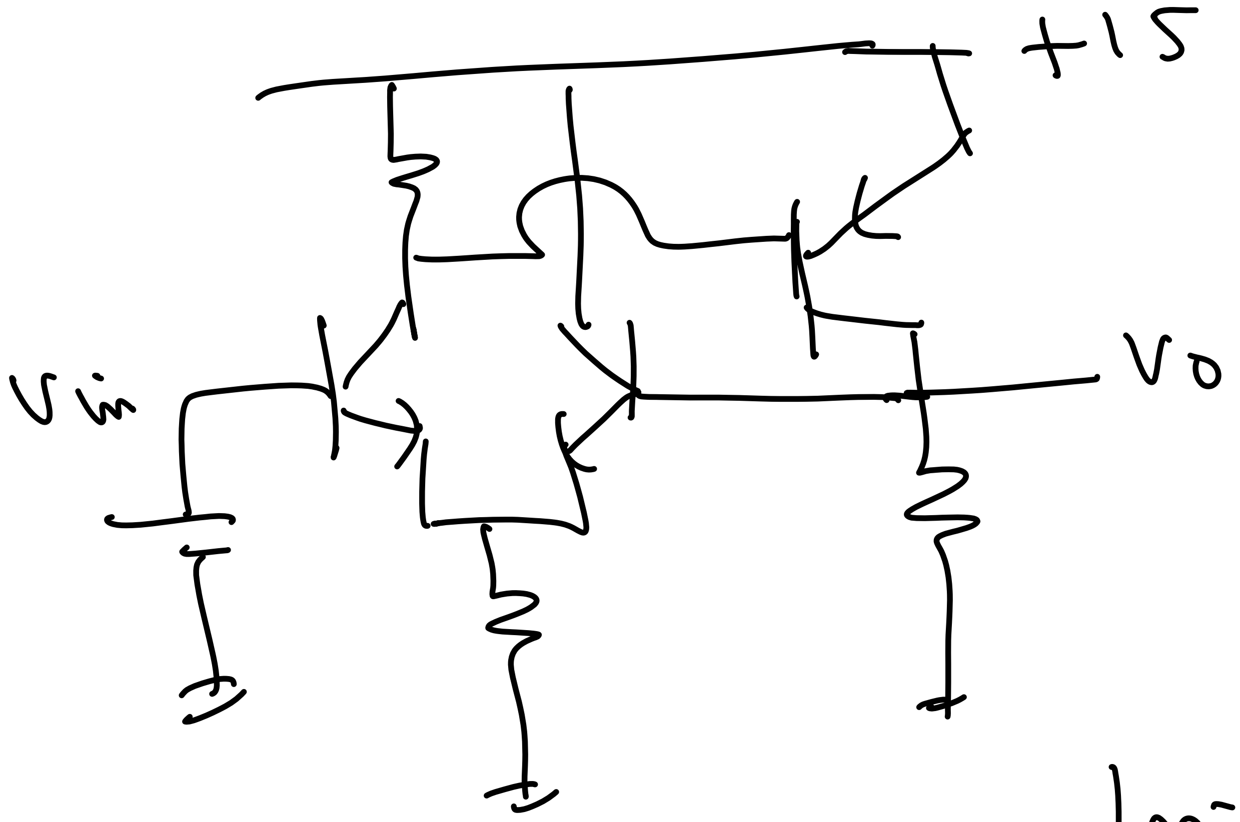
to boost the current



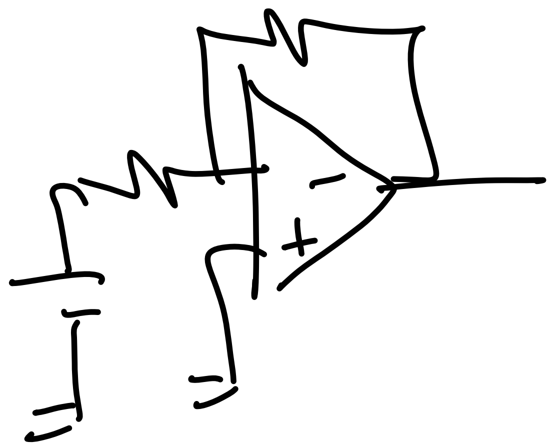
Operational amplifiers



V_{BE} drift created problem for DC amplification

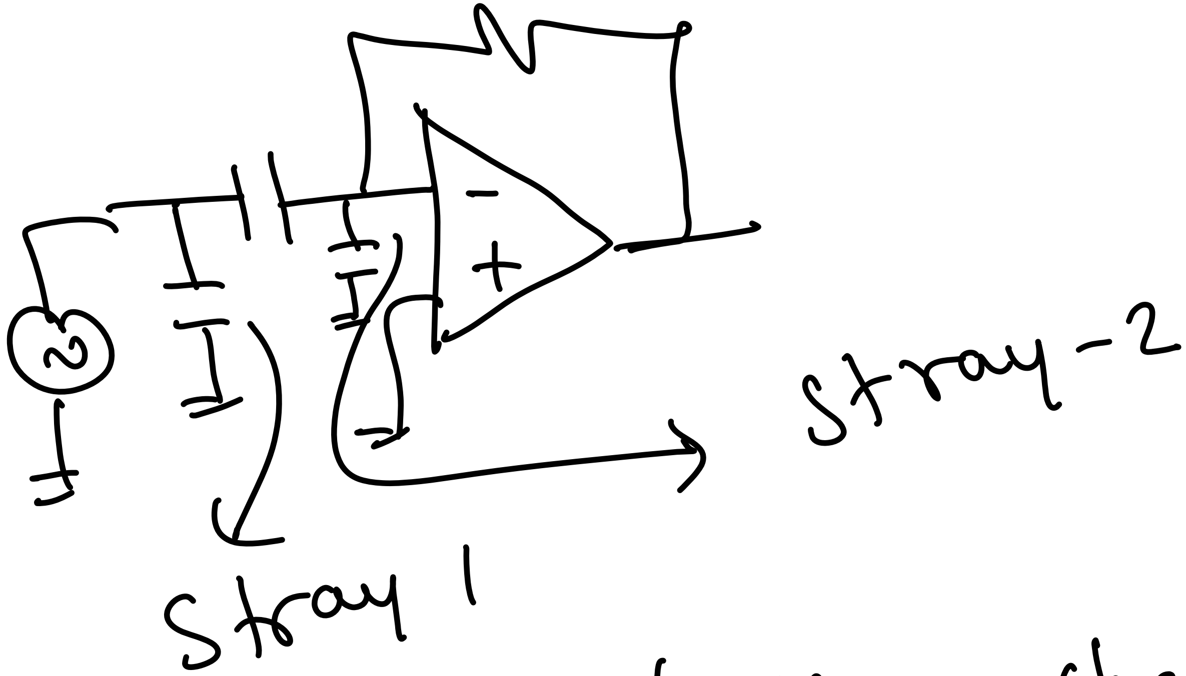


This V_o is not drifting due to temp!!!



1. Voltage regulator using op amp
2. Constant current source
3. 4 - 20 mA current transmitter
4. Transformer
5. LVDT signal conditioning
6. Capacitive transducers

7 Ratio transformer bridge applications



8 Solar based battery charging

Error budgeting

1. Offset voltage drift
2. Input resistance
3. Output resistance
4. Bias current errors
5. CMRR related errors

Making an error budget for the circuit is an important task for the analog circuit designer.

ADC / DAC

1. Dual slope ADC
   Dual slope ADC with MC
   errors involved in these converters also discussed
2. Successive app
3. Flash ADC
4. Σ – Δ ADC

1. Error budgeting
2. Circuit design skill
3. Various techniques involved.

— x — x — x —